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THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

Winter 1998

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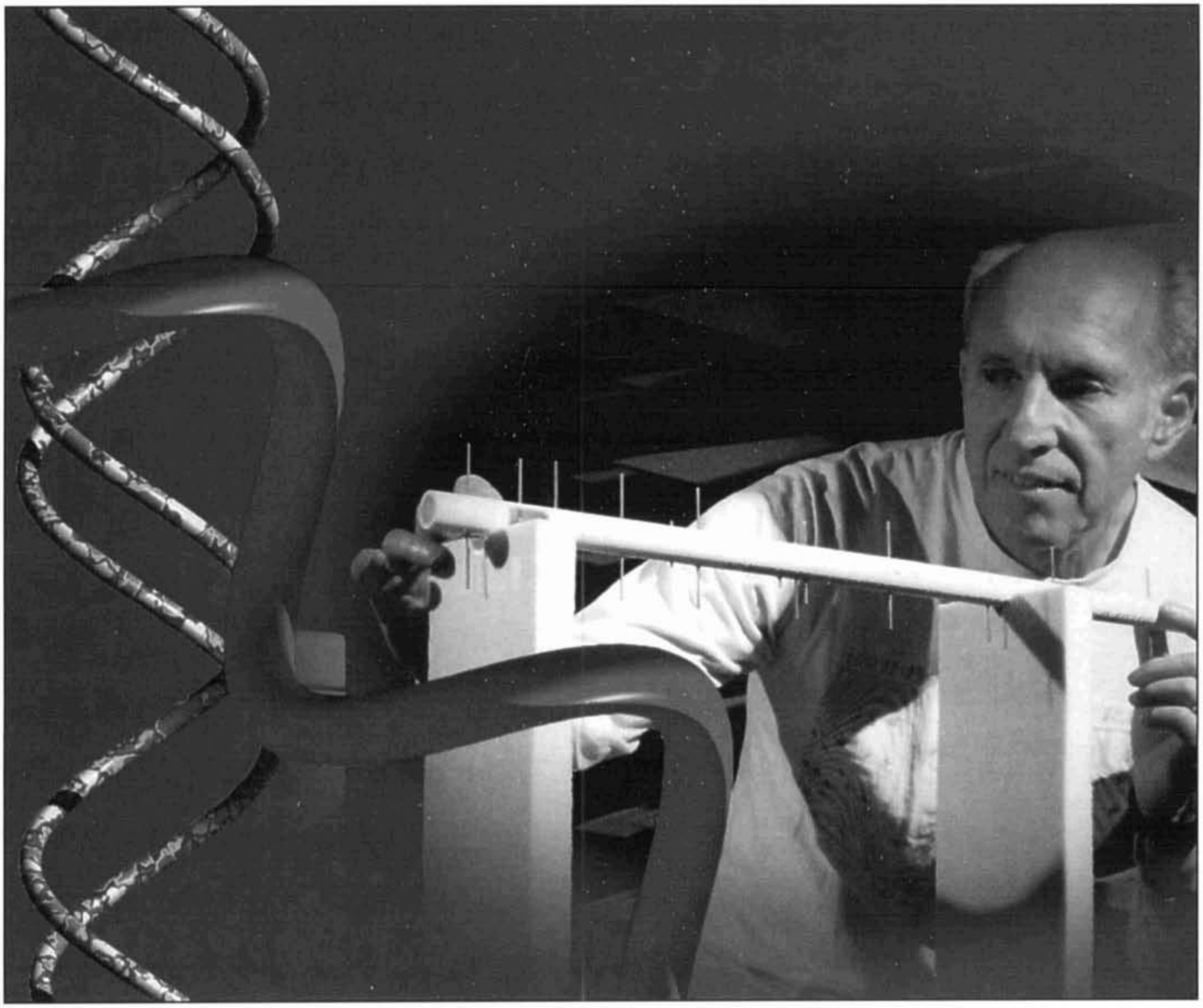

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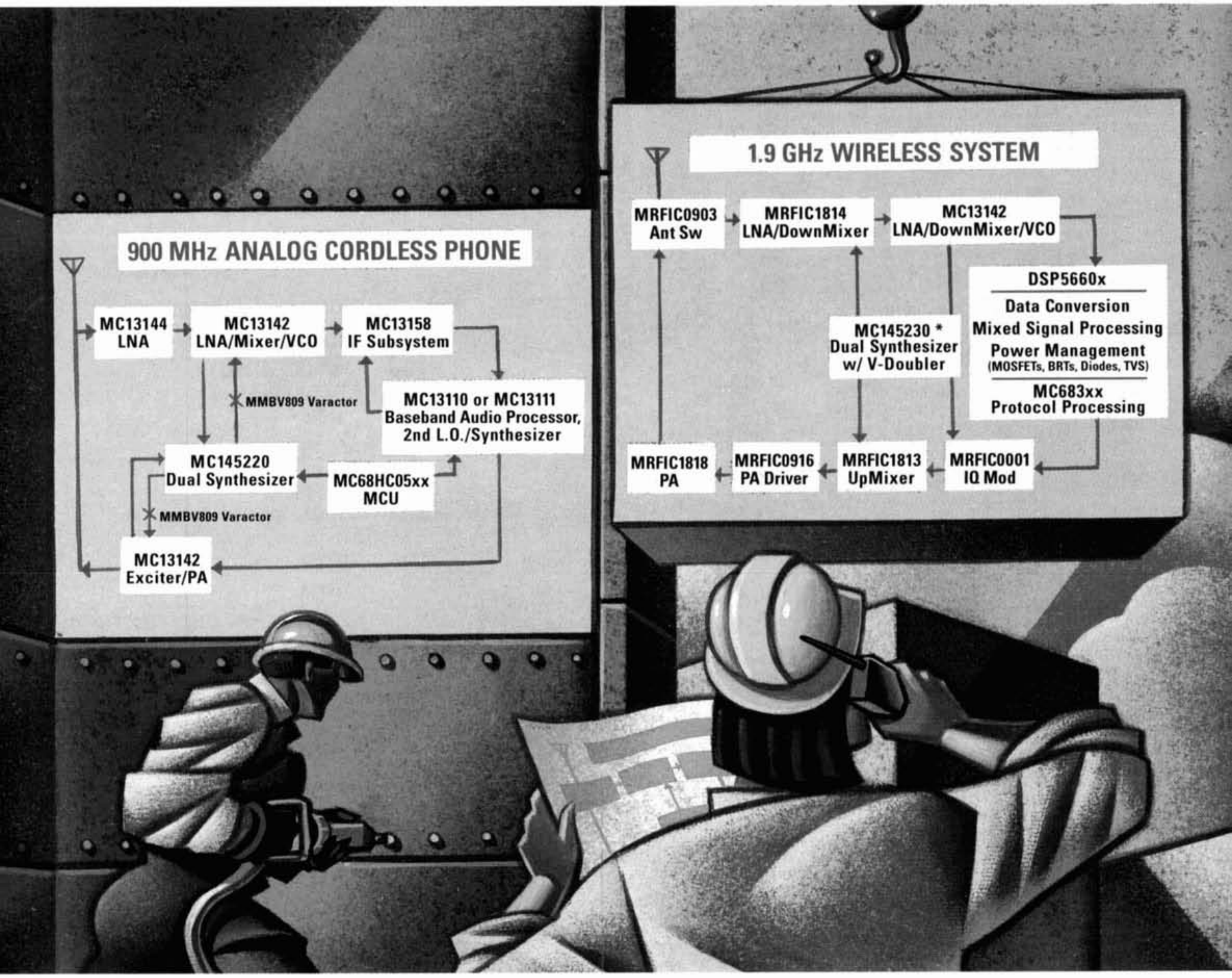

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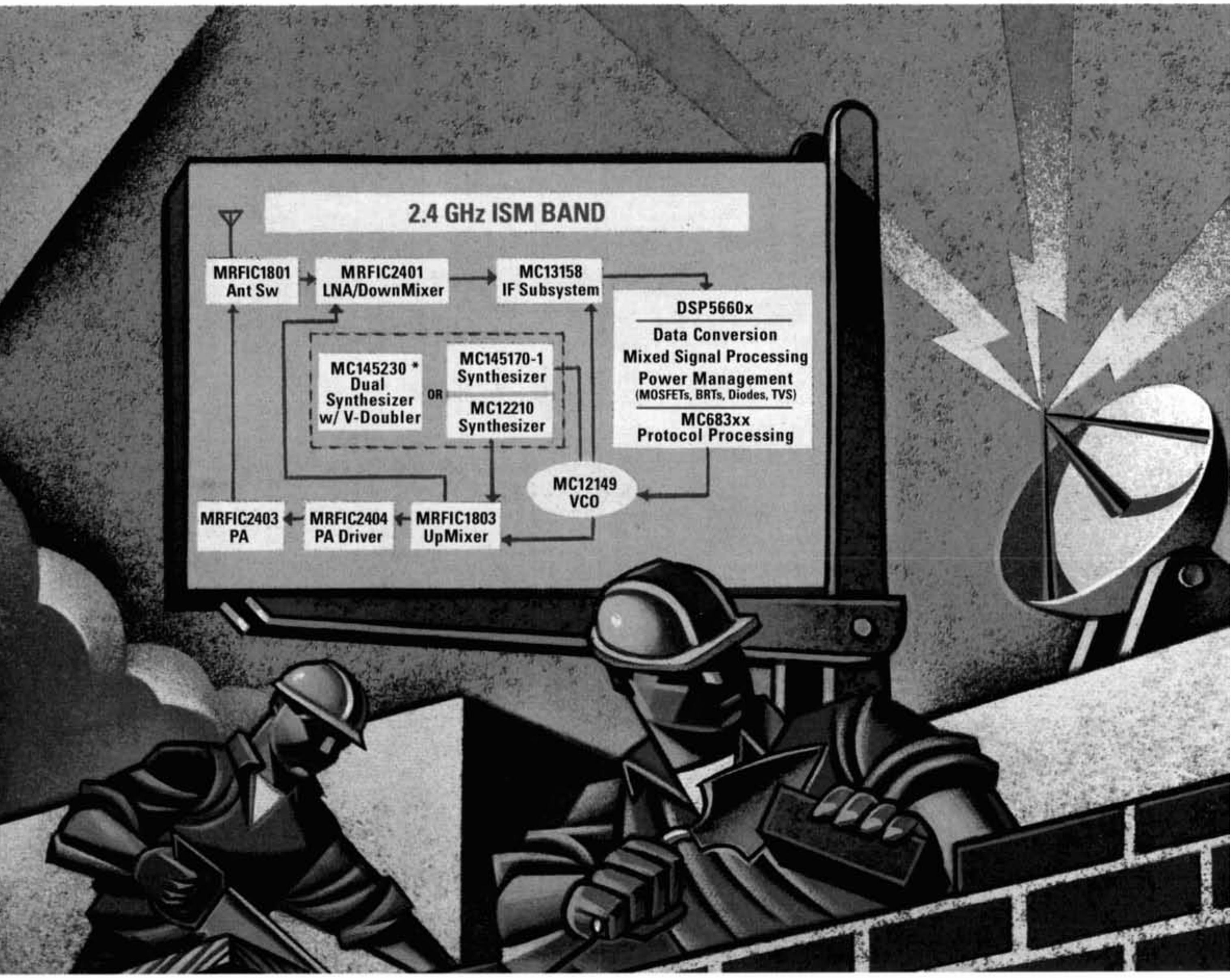

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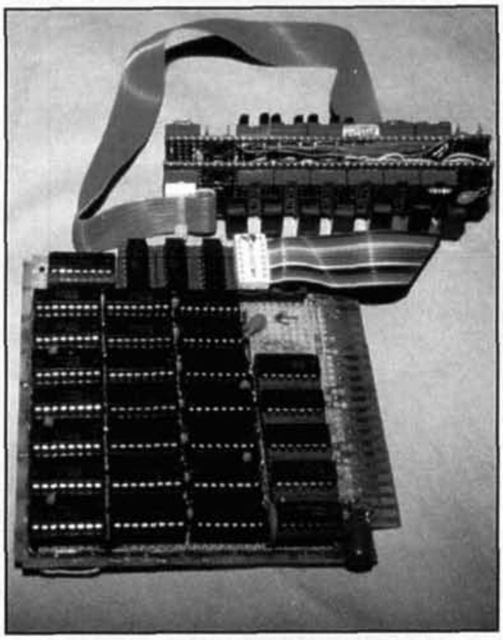

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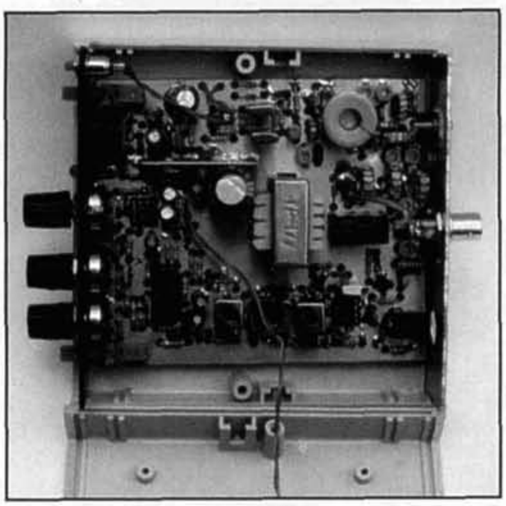

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On the Cover: In "Yagi Antenna Design Using a Genetic Algorithm" (page 11), authors Altshuler, Linden, and Wing discuss a revolutionary process for creating Yagi antennas based on genetics. Cover artist Bryan Bergeron, NU1N, imagines Mr. Altshuler constructing such an antenna.

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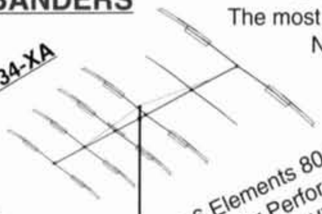

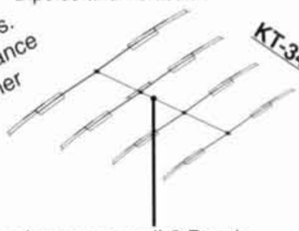

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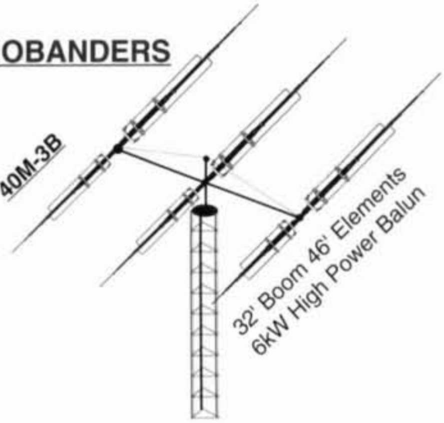

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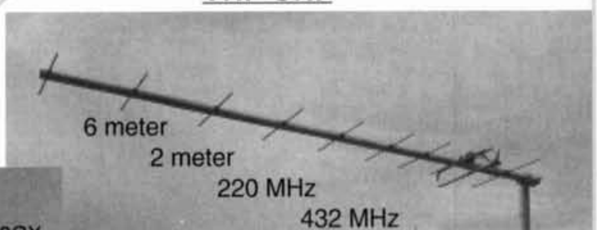

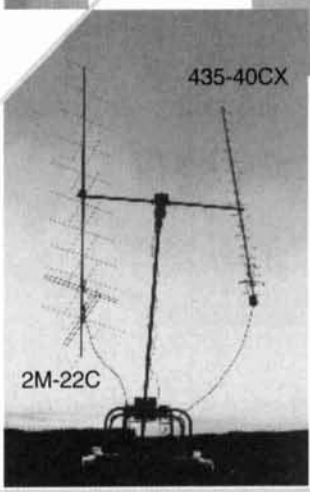

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EDITORIAL

Looking both ways

Ideas seem to ebb and flow in the ham population with the passage of time. In the 10+ years that I've been editing amateur radio publications, there are certain threads that weave themselves into the fabric of the amateur literature and then break off, only to reappear months or years later in another form. As we've progressed from tubes to transistors to integrated circuits to surface mount technology, it seems we've continued to move forward while looking back to our roots.

Advances combined with history

Even though *Communications Quarterly*'s editorial purpose is to look at advances in communications technology in general and amateur radio technology in particular, the occasional historical article has always generated more enthusiasm than negativity. One of the reasons you give for liking these articles on past technologies is that they provide insight into techniques that worked before and may work again in another form.

The changeable world of antennas

Probably the one amateur radio technology that's the subject of the most discussion and modification is the antenna. The majority of articles we receive here at *Communications Quarterly* are about antennas. We'll go through spates of dipoles and Yagis, and Zepps, and Vees, and then we'll start all over again. Every once in a while, we venture into new territory (remember the fractal?) and then return to the antennas with which we are most comfortable. This issue's antenna selection covers Yagis, vertically polarized ground-plane antenna systems, early EME arrays, and an alternative to millimeter-wave antennas.

Of course the ebb and flow of interest in antennas isn't confined to antenna types. We also talk about how we build them. At first, antennas were built using cut-and-try methods. Sure we'd pour over lists of equations in an attempt to determine the best element or wire lengths for the best gain or optimal SWR; but often the building process came down to a little snip here, and a longer element there, and a balun, or a trap, or a different matching system. Cut and try was the only technology we had.

These days, we have the luxury of using antenna optimization programs. Hallelujah! These software gems, now offered for sale in the pages of all the amateur radio publications, have made the design and optimization of antennas a snap. Simply plug your parameters into the program of your choice, push a key, and you'll have an antenna that's custom designed for your needs. Still, the old cut-and-try practice continues to fit into the antenna equation as we take what we've learned from our computer program and put it to work in the real world—which may not always reflect the perfect scenario.

Thought about AM lately?

And what about AM? Yeah, AM. Recently, Rick Littlefield, K1BQT, our regular "Quarterly Devices" columnist and "Tech Notes" author became intrigued with 6-meter AM. Rick noticed that there had been somewhat of a revival of interest in antique radios and also in "this old band" and wondered if anyone was out there listening. At the same time, Senior Technical Editor Peter Bertini, K1ZJH (author of *Popular Communications* "Radio Connections" column, which deals with finding and restoring old radios), was having much the same kind of epiphany. They spent months batting around the idea of a 6-meter AM rig. Finally, Rick sat down and designed the little transceiver offered in this issue's "Tech Notes."

Now even though AM may be old and outmoded, a lot of people like it because it *sounds good*—considerably better, in fact, than the restricted and compressed side-

(Continued on page 101)

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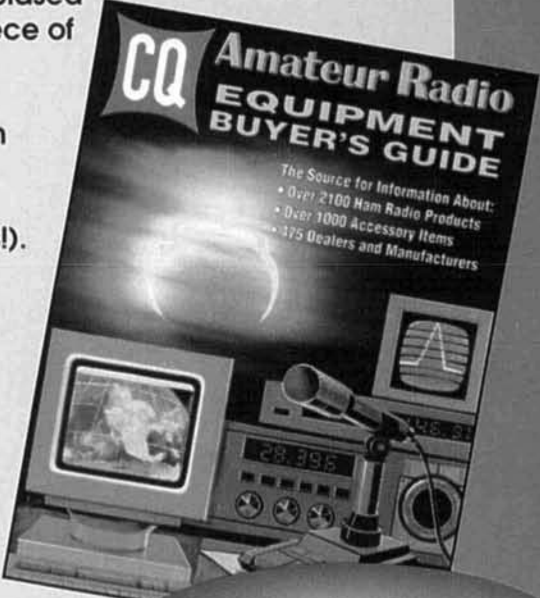

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Comments on Kirchhoff's Laws

Dear Editor:

The article ["Kirchhoff's Laws," Fall 1997] looked very nice as presented in the magazine. There were some problems that cropped up, however. In **Figure 2**, the non-intersections marked by x's are not easily seen. In **Figure 4**, the point labels C and D are interchanged. These items were brought to my attention by readers who wrote to me.

Another reader indicated that I should have shown that, in the cube problem, there is a short way to obtain the solution. He is quite correct. However, while I was aware of several possible shortcuts, due to the fact that all the resistors are equal, I avoided mentioning this so that the general method would be the focus. Using the laws, the circuit may be solved for any values of resistance, whether alike or not. In this problem, I should have stated that I did not want to treat this as a special case (which, of course, it is).

Finally, it was nice to know that others were interested in a topic I found interesting myself. Thanks for putting my article in your impressive magazine.

Jay Jeffrey, WV8R
Beachwood, Ohio

Dear Editor:

Thanks so much for a fine periodical. I appreciated the variety in the Fall issue. The Tesla* story was most enjoyable, as was the article by Jay Jeffrey, WV8R, explaining the application of Kirchhoff's Laws in solving the "classic cube problem." He stated that the solution cannot be achieved by simple series and parallel resistor computations. The problem does however lend itself to a solution that uses wye-to-delta and delta-to-wye transformations rather than current flow simultaneous equations.

Wye and delta conversations are easy-to-remember tools for calculating solutions to many kinds of lattice and bridge circuits. The amount and the complexity of calculations required is often less than if Kirchhoff's simultaneous equations are applied. But we must acknowledge that these shortcut delta/wye conversion formulas are derived from Kirchhoff's work.

The resistance of any leg of a wye is computed as the product of the resistances of the two adjacent delta sides, divided by the sum of the resistances of all three delta sides. Conversely, the conductance (1/R) of one delta side equals the product of the conductances of the two adjacent wye resistors, divided by the sum of the conductances of the wye resistors. An example, easy to remember, consists of three 1-ohm resistors in the form of a "Y." An equivalent delta consists of three 3-ohm resistors. (If the resistors have various values, the arithmetic rules are the same.)

In the cube of **Figure 1** in Jay's article, one can imagine a vertical plane of symmetry passing through the end terminals and resistors R3 and R5. Then only one half of the cube circuit needs to be analyzed for a partial solution. Later, the identical resistance of the second half is paralleled with the resistance computed for the first half for a complete solution of the problem. This method requires that resistors R3 and R5 that are common to both sides of the divided cube each be replaced by a pair of 2-ohm resistors in parallel, one for each side of the split cube. (When the half-cube resistances are later combined, R3 and R5 will assume their original values.)

Any cube corner that does not contain the end terminals is a wye that may be converted to a delta junction at will, thus eliminating a cube corner from the calculations. The computation of series and parallel resistances then becomes possible. Similarly, a delta that defies easy analysis can be converted to a wye to ease the addition of resistances.

After the delta/wye conversions are completed and the net value of resistance between the terminals calculated, that net value is divided by two to represent the paralleling of halves of the cube. The answer of 5/6 ohm agrees with the value stated by Jeffery in his article.

Bruce L. Meyer, WØHZR
Bloomington, Minnesota

Dear Editor:

I have just received and read several interesting articles in the Fall 1997 issue. As usual, there are always a number of fine articles that catch my interest. Impedance matching always catches my interest and so does Nikola Tesla.

I am writing, however, in response to the nice tutorial article by Jay Jeffery, WV8R, "Kirchhoff's Laws: The Classic Cube Problem." I first encountered the resistor cube

*Those who enjoyed the article "Nikola Tesla" in the Fall 1997 issue can check out John Wagner, W8AHB's Web site at <http//www.concentric.net~jwagner>. There you'll find more information about Tesla's life and Wagner's campaign to boost Tesla's historical profile.

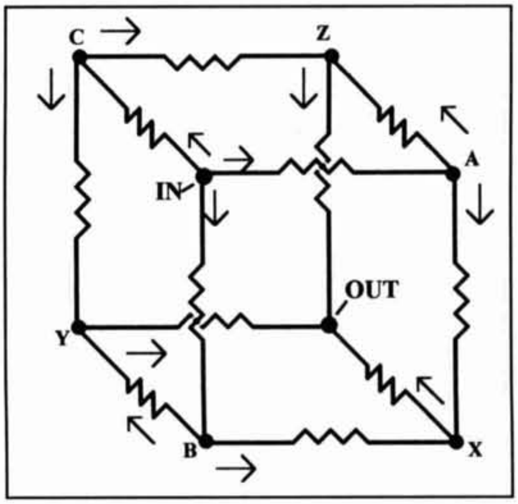

Figure 1. The Classic Cube viewed slightly off the symmetry diagonal.

about 40 years ago and have been giving it to others as an experience-stretching problem ever since. It is usually cast, as in WV8R's article, with each resistor having the same value, usually 1 ohm.

I was disappointed, however, to note that Mr. Jeffery did not point out that, with all resistors having the same value, the problem reduces to a special case that is easily solved by noting that the problem has (electrically) three-fold rotational symmetry about the input-to-output diagonal. This means that the current must divide equally three ways from the input connection and that the three currents to the output are also equally divided. The interconnecting resistors must, again from symmetry, each carry one-half of the one-third or one-sixth of the total current. The voltage drop, following any path, is $1 \bullet I/3 + 1 \bullet I/6 + 1 \bullet I/3 = (5/6) \bullet I$ and the resistance across the diagonal connections is 5/6 of all the (all equal) resistor value. These observations show that Mr. Jeffery's answer was correct, but he seemed to have some lingering doubt. His experimental result, with not quite equal resistors, is to be expected.

The resistors need not be all equal to still have symmetry in the problem. As long as all resistors connected to the input node have the same value (say R_{in}) and all connected to the output have the same value (say R_{out}) and all the remaining interconnect resistors equal have equal values (R_{ic}), then the problem still has three-fold symmetry and the currents must divide as before. The total resistance is $R_{in}/3 + R_{ic}/6 + R_{out}/3$

The general problem has all of the resistors different. Kirchhoff's equations must be used. Although I have run into the problem many

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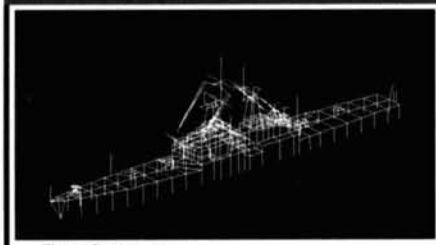

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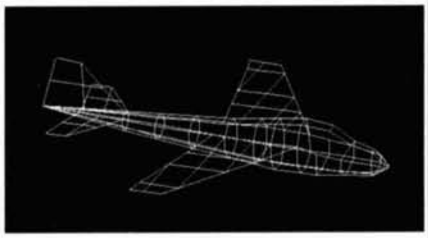

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times, it has always been posed in a symmetrical example, as if the reach to the extra insight was to be invoked and learned. When I give the problem to others, I always point out that having all equal resistors is a special case with an easy solution if a special viewpoint is also invoked.

Bob Plummer, W3RP
Robesonia, Pennsylvania

Dear Editor:

I enjoyed the article in the Fall 1997 issue on Kirchhoff's Laws as applied to the "famous" resistor cube problem. While this may be a demonstration of Kirchhoff's Laws, it is certainly the hard way to get the answer.

We observe:

1) Three resistors lead from one corner into the array of 12 resistors.

2) From symmetry, the voltage drop across all three must be the same.

3) Therefore, we can tie the ends of these resistors together without changing any currents. Let's pretend we did. We have three 1-ohm resistors in parallel = 1/3 ohm.

4) Do the same for the opposite corner, get another 1/3 ohm.

5) The remaining six resistors are all in parallel and connect the two "let's pretend" junctions together. These total another 1/6 ohm.

The grand total is 5/6 ohm.

Brooks Shera, W5OJM
Santa Fe, New Mexico

"Salt Water Taffy" and some comments on the conjugate match

Dear Editor:

Joe Carr's piece "Is Salt Water Taffy Being Distributed?" (Fall 1997) is true for any study or thesis. A case in point is the tuned amplifier/conjugate match controversy. As a fellow said, "Half of what you're telling me is pure crap. The problem is I don't know which half!" So, question we must. But what is the basis for my conclusions? Am I knowledgeable enough to draw a conclusion?

You must have been hard up to fill your page allotment to carry the disorganized and rambling piece "Source Impedance and HF Tuned Amplifiers and the Conjugate Match." I don't think a simple editing job could salvage it, but editorial comments and questions might have led the authors to tighten it up. It might be like blind men describing an elephant; it depends on the point of view.

I think the authors should have limited the discussion to a few general questions: 1. What is a conjugate match and why is it important? 2. What is the source impedance of a tuned amplifier and why is it important? 3. What is the source impedance of a tuned amplifier and why is it important? How is the source impedance of various class tuned amplifier calculated and how is it measured? These questions were answered years ago when I was in school. The authors didn't add anything new to the classic analysis. Beating up on Bruene was unnecessary. A fourth but separate question could have been: How does a π matching network transform a complex load to a conjugate match for some complex source?

It's easy for me to second guess.

Parker R. Cope, W2GOM/7
Prescott Valley, Arizona

And more on the conjugate match...

Dear Editor:

I was advised to check out the article "Source Impedance of HF Tuned Power Amplifiers and the Conjugate Match" in the Fall 1997 issue of *Communications Quarterly*. It arrived today and that was the first thing I read.

I am bowled over. What a masterful presentation and a wonderful demonstration of measurement techniques. Certainly not everyone can, or will, follow the thread of their arguments and I don't feel I've fully digested it yet. Wonderful to get to read a magazine twice and get more out of it the second time.

Thanks for making the space available for their work. Good choice.

Bill Carver, W7AAZ
Twin Falls, Idaho

Dear Editor:

The following comments are offered with respect to this thoughtful and carefully written article ("Source Impedance of HF Tuned Power Amplifiers and the Conjugate Match," Fall 1997) with which I nevertheless disagree on certain issues:

1. There is an almost exclusive emphasis on driving and loading for maximum power output and to demonstrate that a 50-ohm resistance measurement at the PA output can be achieved. The problem is that always, especially in an AM or SSB transmitter, the actual settings are for a compromise between efficiency and linearity. These settings can be found graphically from the constant-current curves for the tube. The value of the measured resistance, using the methods described, can change considerably over the course of the speech modulation cycle. The article mentions this briefly, but it is not given the attention that it deserves. If this resistance is large at low modulation amplitudes, but takes a sudden significant drop on modulation peaks, the guaranteed result is degraded

intermodulation products. It is a sure sign of flat-topping. For this reason, the preoccupation with tune and load for a 50-ohm value can be deceptive and counterproductive. If the PA is loaded beyond the optimum point that I mentioned, the output power can easily increase as much as a dB and the average tube resistance, as measured by the methods described, can decrease considerably. ALC and TGC circuits are intended to keep the PA within its proper range.

2. One way to get a preliminary understanding of a problem is to look at the simplest possible model. Consider a perfectly linear class A amplifier with no L or C in the circuit, only an ideal 1:1 output transformer as shown in **Figure 1**. Refer to **Figure 2** and consider (for now) only the load resistance RL1, represented by load-line AB. The plate voltage swings between ep1min and ep1max, and the plate current swings between ip1max and ip1min. Also, the tube's DC resistance is RDC = VBB/IBB. VBB and IBB define a DC operating point Q that corresponds to a certain fixed value of grid bias VGG. The slanted VGG lines are all parallel to each other. For a sine wave variation of grid voltage between VGG1 and VGG2, this bias point and the value of RL determine the sine waves of plate voltage and plate current and, therefore, the AC power output.

Consider two kinds of measurements that we can make at the output of this amplifier:

• Turn off the AC grid drive signal and measure the output resistance of the tube with an impedance bridge. This measures the dynamic plate resistance rp (the output resistance of the tube). In **Figure 2**, it is the slope of one particular plate volt-ampere curve at one particular fixed value of grid bias VGG. That is,

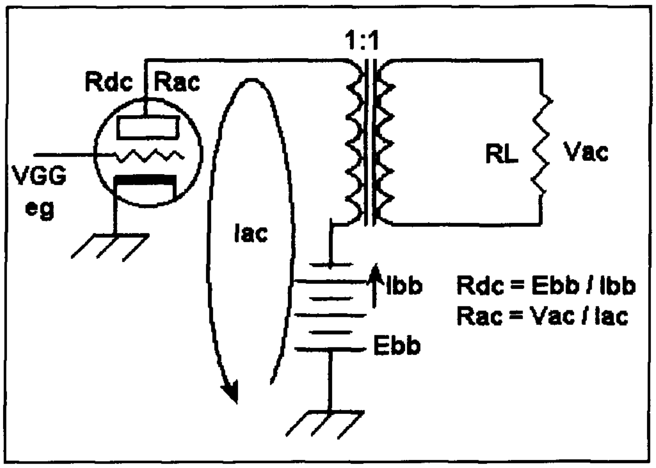


Figure 1. Linear class A example.

$$r_p(e_g) = \left.\frac{\Delta e_p}{\Delta i_p}\right|_{e_g = V_{GG}}$$

where lower case implies instantaneous values. For simplicity, assume there is no kind of negative feedback. A very small amount of AC power, provided by the impedance bridge, is dissipated in the tube.

• With the grid AC signal generator running, measure the AC output resistance of the perfectly linear class A amplifier using a test generator at the output as described in the article. The vacuum tube is, in this situation, basically a variable resistor r_T connected between B+ and ground. To see this, note that the plate voltage and the plate current are 180 degrees out of phase. That is, when the plate voltage is large, the plate current is small (high value of resis-

(Continued on page 102)

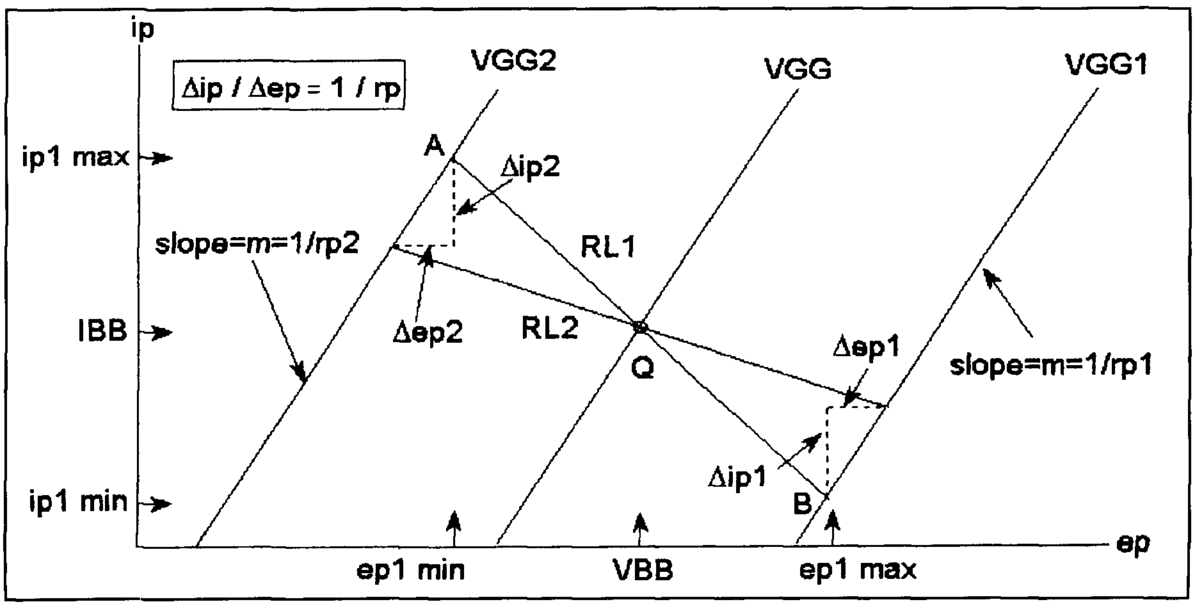
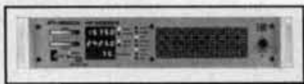


Figure 2. Measuring plate resistance.

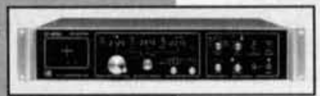
PRODUCT LINEUP

MILITARY AND COMMERCIAL PRODUCTS



ST8000A **- Ruggedized HF Modem, MIL connectors, high MTBF, touch panel controls**

ST8000 **- HF Modem, Tuneable Mark/Space tones, memories, space tuning indicator, up to 1200 baud FSK**



ARQ1000B **- Error Correction Terminal, CCIR-476/625 TOR ARQ/FEC**

HFCS1000 **- High Frequency Communications Simulator, Training device simulates HF receiver operation, .5 to 30 MHz**

LP1210 **- Ten Channel ruggedized Loop Power Supply, 10 channels, MIL-188 to neutral loop, 100K hrs MTBF**

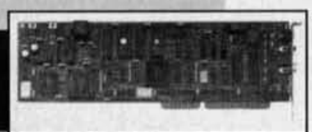
LP1200A **- Loop Power Supply, 1 channel, polar or neutral, 20/60 ma, RS-232/MIL-188, motor control**

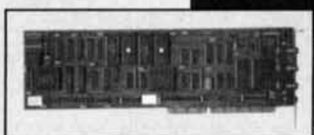
DS3486 **- Radio Data Communications Terminal, Rack mount, 486, VGA, Floppy Drive, 420MB Hard Drive**

DSP PRODUCTS

DSP4100 **- HF DSP Modem, Stand-alone DSP modem for CLOVER-II, CLOVER-2000, TOR, RTTY, and ASCII, + 12VDC**

PCI4000+ **- HF DSP Modem, PC plug-in card, Operates CLOVER-II, CLOVER-2000, TOR, RTTY, and ASCII**





P38 **- HF DSP Modem, PC Plug-in card, Designed with the Amateur in mind, Operates CLOVER-II, AMTOR, RTTY, and ASCII**

FAX4100 **- FAX-OVER-RADIO Interface, Interfaces a G3 FAX machine to the DSP4100/CLOVER-2000 Modem**

LI4100 **- Line Interface for FAX4100, Share G3 FAX machine between phone lines and FAX4100**

CLOVER-2000 **- Voice Bandwidth CLOVER software, for PCI4000+ and DSP4100, includes TOR, RTTY, and ASCII**

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YAGI ANTENNA DESIGN USING A GENETIC ALGORITHM

A new revolutionary design process

Looking for a new way to design Yagi antennas? This article describes a revolutionary process for designing Yagi antennas using a genetic algorithm (GA). We specify the desired electromagnetic properties of the antenna, and the algorithm produces a configuration that most closely approaches the design goals, allowing one to custom design a Yagi. To illustrate this new method, we chose to maximize the gain of four Yagi antennas with boom lengths ranging from 3.6 to 6.1 wavelengths at a frequency of 432 MHz; VSWR was of secondary importance, and back and sidelobes were not included in the optimization. We compared these designs with those using the best design techniques currently available.

The gains, radiation patterns, and VSWRs of both conventional and genetic Yagi antennas were computed using the Numerical Electromagnetics Code (NEC). Antennas having a boom length of 5.16 wavelengths were also fabricated and tested. The genetic algorithm produced configurations that were quite different from typical Yagis. The conventional Yagi has directors that start out with lengths that decrease gradually and spacings that increase gradually along the array. The genetic Yagis had elements with lengths and spacings that did not show any systematic change along the antenna. The genetic Yagis had computed gains that ranged from 0.4 to 1.1 dB higher than those of conventional Yagis at the design frequency of 432 MHz. The measured gain of the genetic Yagi with a boom length of 5.16 wavelengths was 0.8 dB higher than the corresponding conventional Yagi. This algorithm provides the antenna designer with a powerful technique for synthesizing Yagi antenna configurations with properties that have heretofore been unattainable.

Introduction

The Yagi antenna evolved as a special configuration of an endfire array. It is a traveling-wave antenna with a surface wave that propagates along the array with a phase velocity slightly less than that of free space. First proposed by Professor H. Yagi and his student, S. Uda, in the late 1920s, it consists of a single driven element and a number of parasitic elements made up of a reflector and a set of directors. The Yagi has been exhaustively investi-

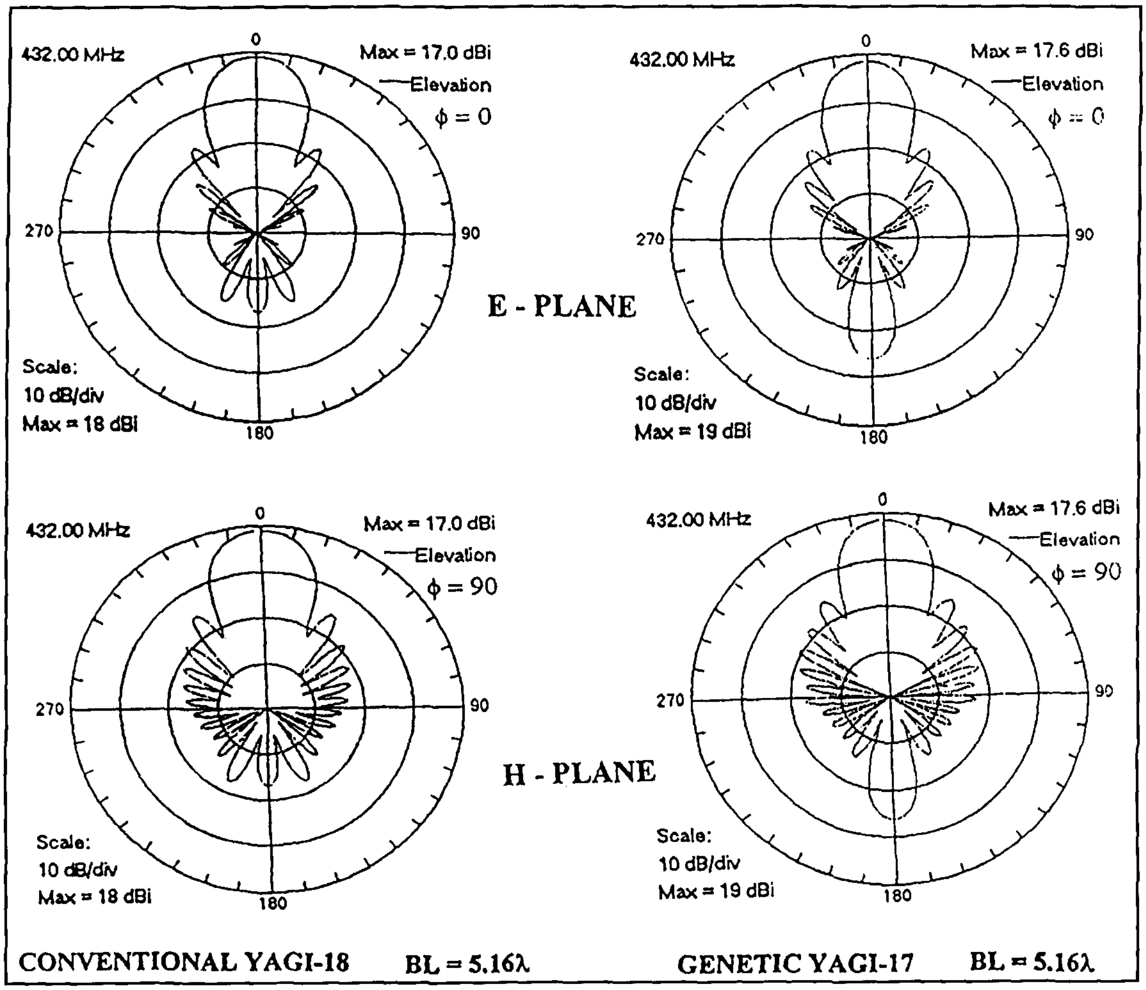


Figure 1. Computed E- and H-plane patterns of 5.16-wavelength boom length conventional and genetic Yagis at 432 MHz.

gated, both theoretically and experimentally, for many years. The Yagi configuration has not been amenable to practical analysis because it is an array of elements of different lengths with non-uniform spacing and cannot be treated using conventional array theory. Most analyses have been restricted to relatively short arrays; progress throughout the years for longer arrays has been achieved mostly experimentally and computationally.

The Yagi is lightweight and inexpensive. It has been widely used for many high gain and narrow frequency band applications, particularly by the amateur radio community. Over the years, the performance of Yagi antennas has been slow to improve. It is believed that maximum gain is achieved by controlling the phase velocity of the surface wave; the Yagi structure must be designed so the surface wave is properly retarded. This has been accomplished with some success by logarithmically tapering the elements—the director spacings are increased gradually while the lengths are decreased gradually until they approach constant values at a distance of about 3 or 4 wavelengths from the driven element. Minor changes in the antenna configuration have produced only a small improvement in performance.

Here, we describe a new process for designing Yagi antennas using a genetic algorithm (GA).[1–4] Our approach is one of synthesis: we specify the desired electromagnetic properties of the antenna, and the algorithm produces a configuration that most closely approaches these design goals. This method allows one to custom design a Yagi. The gains, radiation patterns, and VSWRs of four Yagi antennas having boom lengths ranging from 3.6 to 6.1 wavelengths are computed using the Numerical Electromagnetics Code (NEC)[5] with the gains achievable using currently available design techniques.[6,7] We then fabricated and tested a conventional Yagi and a genetic Yagi, each having the same boom length.

Approach

The GA begins with a large population of potential Yagi configurations. These configura-

	CONVENTIONAL YAGI						GENETIC YAGI							
BL	3.60λ		5.16λ		6.10λ		3.60λ		4.88λ		5.16λ		6.10λ	
El.	L cm	S cm	L cm	S cm	L cm	S cm	L cm	S cm	L cm	S cm	L cm	S cm	L cm	S cm
1	34.5		33.7		34.6		33.1		33.1		33.8		33.6	
2	32.8	13.0	32.5	12.7	34.0	10.4	34.6	17.5	30.4	13.4	31.5	13.2	28.8	9.3
3	30.8	5.5	30.0	5.7	32.1	4.2	31.5	10.1	30.9	9.7	30.3	17.3	31.5	2.1
4	30.4	12.5	29.8	11.1	31.1	7.8	30.4	20.5	29.8	18.2	29.5	24.2	30.4	15.5
5	30.0	15.0	29.5	14.9	30.5	10.8	17.9	10.1	11.4	14.4	27.9	21.6	29.3	25.6
6	29.6	17.5	29.1	17.5	30.1	13.4	22.2	12.6	29.3	7.8	27.6	12.6	11.9	3.0
7	29.3	19.5	28.9	19.4	29.7	15.6	29.3	4.6	14.1	24.8	27.2	18.3	20.6	5.9
8	29.3	21.0	28.7	20.8	29.5	17.6	28.8	32.0	28.2	4.9	27.2	13.7	28.8	19.8
9	28.9	22.0	28.6	21.9	29.3	19.2	28.8	29.0	20.6	3.2	28.0	23.3	28.2	31.3
10	28.9	23.0	28.4	22.7	29.1	20.6	26.0	26.0	27.7	26.5	25.8	21.2	28.8	31.3
11	28.9	24.0	28.3	24.0	28.9	21.8	27.1	6.1	27.7	34.3	27.3	15.0	28.8	22.2
12	28.5	25.0	28.1	24.9	28.8	22.8	28.8	28.0	28.2	24.8	27.7	27.4	12.5	15.0
13	28.5	26.0	27.9	25.7	28.6	23.7	28.8	29.5	25.5	38.1	27.6	29.7	27.1	25.5
14	28.1	26.0	27.8	26.4	28.5	24.3	29.3	28.0	27.7	37.1	27.4	33.2	27.7	29.9
15			27.5	26.7	28.4	25.1			27.7	25.8	27.7	29.4	27.7	28.0
16			27.3	27.4	28.3	25.6			28.2	32.4	27.9	32.4	22.2	26.5
17			27.1	27.6	28.1	26.1			29.3	22.9	28.6	25.9	26.0	16.5
18			27.0	27.8	28.0	26.4							27.1	27.0
19					27.9	26.8							19.5	11.2
20					27.8	27.1							27.7	20.3
21					27.7	27.4							28.2	30.9
22					27.6	27.5							28.8	24.0

Table 1. Dimensions of Yagi antennas.

tions are determined by the constraints of the problem and the method of encoding all configuration information (e.g., start and endpoints and wire sizes) into a binary string of 1s and 0s called a binary chromosome or a set of numbers called a real-valued chromosome. For this investigation, we tested both types of chromosomes. The GA randomly selects a small set, or a sample population of Yagi configurations, and evaluates the performance of each member of the population using NEC. The outcome is compared with the design goal, which is represented by a cost function, and returns a single number to the GA that is a measure of its fitness. As in the evolutionary process of "survival of the fittest," high-quality strings (chromosomes) mate and produce offspring, while poor-quality strings perish. Offspring are created by combining randomly selected parts of two chosen parent strings. With succeeding "generations," the quality of the strings improves continually and, usually after a few thousand antenna simulations, an optimal solution is ultimately obtained.

The GA method of antenna design is analogous to that of breeding race horses—the "horses" are antenna designs, and the "race track" is a simulation to determine performance. "Champions" will sire many offspring, while those who do not perform well will not. This method is resistant to becoming trapped in local maxima, which allows it to work well for antenna design problems.

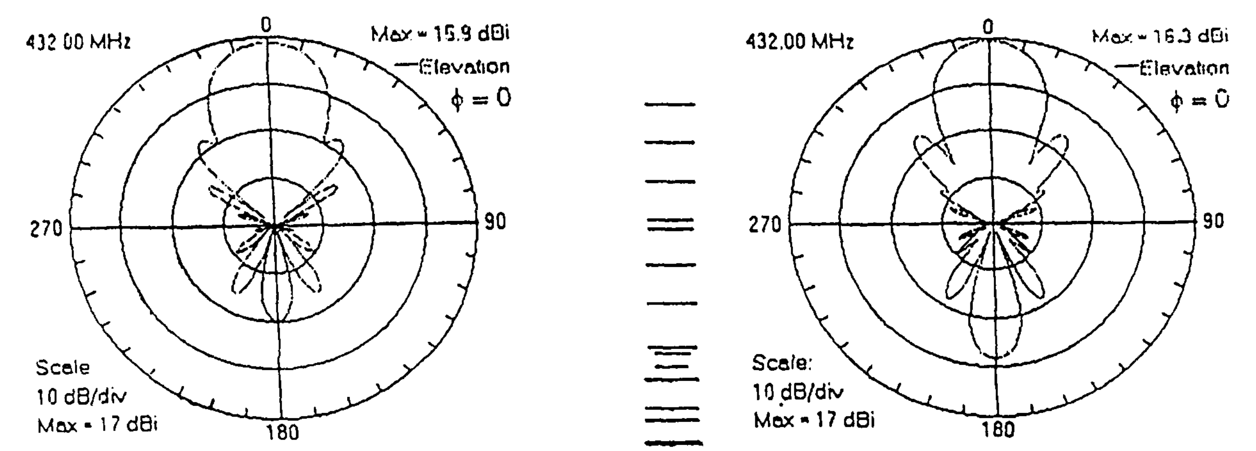
The antenna design problem we created generally employs a steady-state GA (i.e., a portion of the current population carries over to the next generation). In our implementation of this steady-state GA, the chromosomes that are carried over are used to generate the offspring that refill the rest of the population (as opposed to

CONVENTIONAL YAGI					GENETIC YAGI				
N	BL λ	FREQ MHz	GAIN dBi	VSWR	N	BL λ	FREQ MHz	GAIN dBi	VSWR
14	3.60	424	15.5	1.41	14	3.60	424	15.4	1.88
		428	15.8	1.11			428	16.0	1.80
		432	15.9	1.23			432	16.3	1.09
		436	15.7	1.60			436	16.1	9.5
		440	15.5	1.85			440	9.4	39
					17	4.88	420	15.8	5.10
							424	16.4	4.50
							428	17.0	3.70
							432	17.4	1.55
							436	17.0	13.0
18	5.16	424	16.5	1.37	17	5.16	420	16.4	2.55
		428	16.8	1.13			424	16.9	2.60
		432	17.0	1.03			428	17.3	2.35
		436	17.1	1.27			432	17.6	2.80
		440	17.1	1.43			436	16.5	16.0
							440	6.4	18.0
22	6.10	424	17.8	1.34	22	6.10	416	16.4	23
		428	17.8	2.28			420	17.1	19
		432	17.8	2.92			424	17.9	12
		436	17.5	2.81			428	18.5	9.3
		440	17.5	5.20			432	18.9	1.51

Table 2. Summary of Yagi computations.

using all chromosomes to generate the next set of children before removing the poor performers from the population). Since each NEC simulation takes several seconds, it is important to use a GA that will converge as quickly as possible and still arrive at a good result.

Using the steady-state GA with the restriction that the parents be from the top percentage of the population allows it to converge on a solution much faster than the more conventional, simple GA described in **Reference 2**. A virtual "weighted roulette wheel" is filled according to each chromosome's fitness: the fitter a chromosome, the larger its share of the wheel. For each new position to be filled in the population, the wheel is "spun" and the first parent is chosen. The wheel is "spun" again and the second parent is chosen, unless the wheel points to the same member of the population, in which case it is spun until a different member is



CONVENTIONAL YAGI-14
BL = 3.60 λ

GENETIC YAGI-14
BL = 3.60 λ

Figure 2. Computed E-plane patterns of 3.60-wavelength boom length and genetic Yagis at 432 MHz.

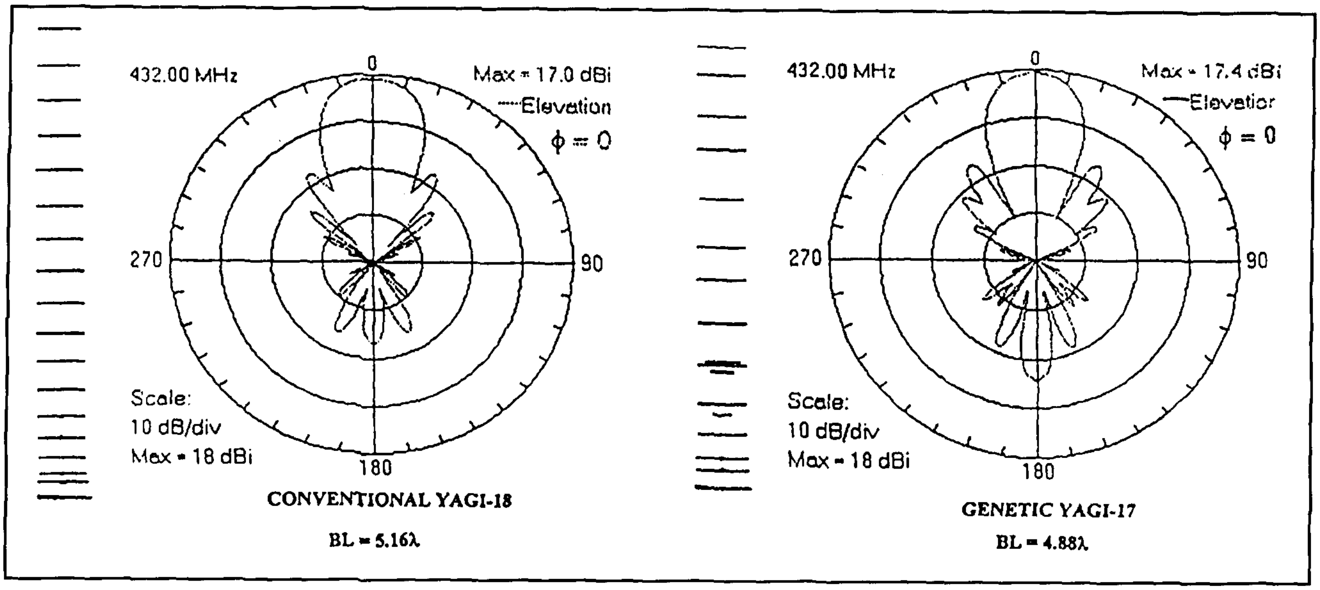


Figure 3. Computed E-plane patterns of 5.16-wavelength boom conventional and 4.88-wavelength genetic Yagi at 432 MHz.

selected. A single crossover site is chosen at random, and a child is produced from the first part of the first parent's chromosome, up to the cross-over point, and the second part of the second parent's chromosome, and vice-versa for a second child.

This process is repeated until the population is full again. Mutations are included in the algorithm to allow the GA to search for solutions outside the existing gene pool. We use a gaussian mutator to mutate the genes of the children. This mutator operator pulls the mutated gene from a gaussian distribution centered around the original gene. We give this distribution a standard deviation of 10 percent of the gene range (thus restricting the mutated gene to within ±30 percent of the gene range of the original gene about 99 percent of the time). The new population members are evaluated using the NEC and the generation process is begun anew. The program also remembers evaluations it has computed, so a chromosome that has been previously evaluated will be copied, eliminating the need for duplicate calculations of the same chromosomes. This feature greatly decreases the time required to run the GA. We generally allow this process to proceed for 70 to 90 such generations, although there are many other criteria that can be used to halt the GA.

The genetic Yagis are optimized for gain and VSWR at a frequency of 432 MHz. Sidelobes, backlobes, and bandwidth are not included in the optimization. The cost function is:

$$F = -G + C_1 \times (VSWR) \qquad (1)$$

where G is the endfire gain and C_1 is 1 when the VSWR is greater than 3.0 and 0.1 when the VSWR is less than 3.0. The objective is to minimize F. The encoding of the algorithm is done as follows:

length of element #1 (reflector)
length of element #2 (driven element)
spacing between elements #1 and #2
length of element #3 (first director)
spacing between elements #2 and #3
length of element #4 (second director)
and etc.

Each element has a range of lengths up to 0.75 wavelength with minimum spacings of 0.05 wavelength. The possible wire diameters range from 2 to 6 millimeters with 1 millimeter increments. A sample population of 175 Yagi configurations is randomly selected from the total population. A steady-state GA is used with 30 percent of the population saved from generation to generation. We used real-valued chromosomes to optimize the Yagi with a boom length of 5.16 wavelengths; binomial chromosomes were used for the other Yagis.

The algorithm using the real-valued chromosomes appeared to converge more quickly. We found optimal configurations for 14, 17, 18, and 22-element Yagis having corresponding boom lengths of 3.60, 4.88, 5.16, and 6.10 wavelengths.

After optimal configurations were produced by the algorithm, we selected conventional Yagis having corresponding boom lengths for comparison, with the exception of one having a boom length of 4.88 wavelengths. This particular Yagi was optimized to have 18 elements and a boom length of 5.16 wavelengths; however, the end element of this genetic Yagi was very small and had a very low current. We found that

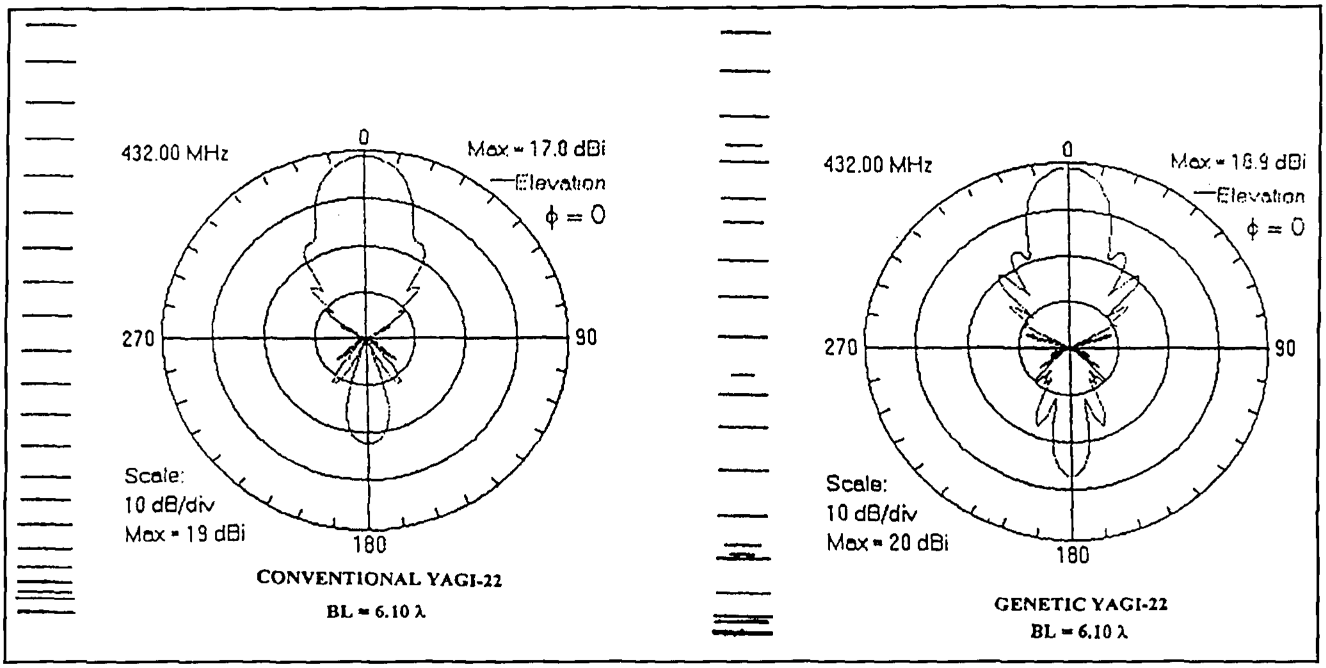


Figure 4. Computed E-plane patterns of 6.10-wavelength boom length conventional and genetic Yagis at 432 MHz.

removing it from the antenna did not change any of the characteristics.

We fabricated the conventional and genetic Yagis having a boom length of 5.16 wavelength and measured their gains, E- and H-plane patterns, and VSWRs. Because the antenna pattern range did not operate satisfactorily at frequencies below about 800 MHz, and because the full-scale antennas would have been quite large, we decided to work with a fourth scale model having a center frequency of 1728 MHz. The conventional Yagi elements were made of 1.2 millimeter (3/64 inch) copper rod, while the genetic Yagi used 1.6 millimeter (1/16 inch) rod. These elements were inserted into 1/2-inch PVC pipe. It is estimated that the Yagis were built to an accuracy of about ±0.5 millimeter. The measurements were made over the frequency range from 1650 to 1750 MHz, which corresponded to the full-scale range of about 412 to 438 MHz.

Results

Computations. The genetic algorithm produced configurations quite different from a typical Yagi. The director lengths and spacings for the conventional and genetic Yagis are shown in **Table 1**. Note that the conventional Yagis have directors with lengths that decrease gradually and spacings that increase gradually along the array. For the genetic Yagis, neither the lengths nor the spacings showed any systematic change along the antenna. They appeared for all practical purposes to be random.

The gains and VSWRs were computed for both designs using NEC. Although bandwidth was not included in the optimization of the genetic Yagi, we did compute the frequency dependence of the antennas for completeness. In **Table 2**, we show the computed gains and VSWRs for both types of antennas near the design frequency of 432 MHz. The conventional Yagis have nearly uniform gain and relatively low VSWRs over most of the band. The genetic Yagis have a higher gain at 432 MHz, but the gain drops off sharply near 440 MHz, particularly for the 3.6 and 5.16 wavelength antennas. Also, the VSWRs are low at 432 MHz, but tend to increase sharply at frequencies other than 432 MHz.

The computed E- and H-plane radiation patterns are shown in **Figure 1** for both types of antennas with a boom length of 5.16 wavelengths at the design frequency of 432 MHz. The sidelobes and backlobes for the conventional Yagi are about 15 and 22 dB down; those for genetic Yagi are down 16 and 13 dB, respectively. The E-plane patterns for the other Yagis are shown in **Figures 2**, **3**, and **4**. Once again, the sidelobes for the conventional and genetic Yagis are about the same; however, the backlobes for the conventional Yagis are lower.

Measurements. The VSWRs of the conventional and genetic fourth-scale Yagis, each having a boom length of 5.16 wavelength, were measured with a Hewlett-Packard 8510

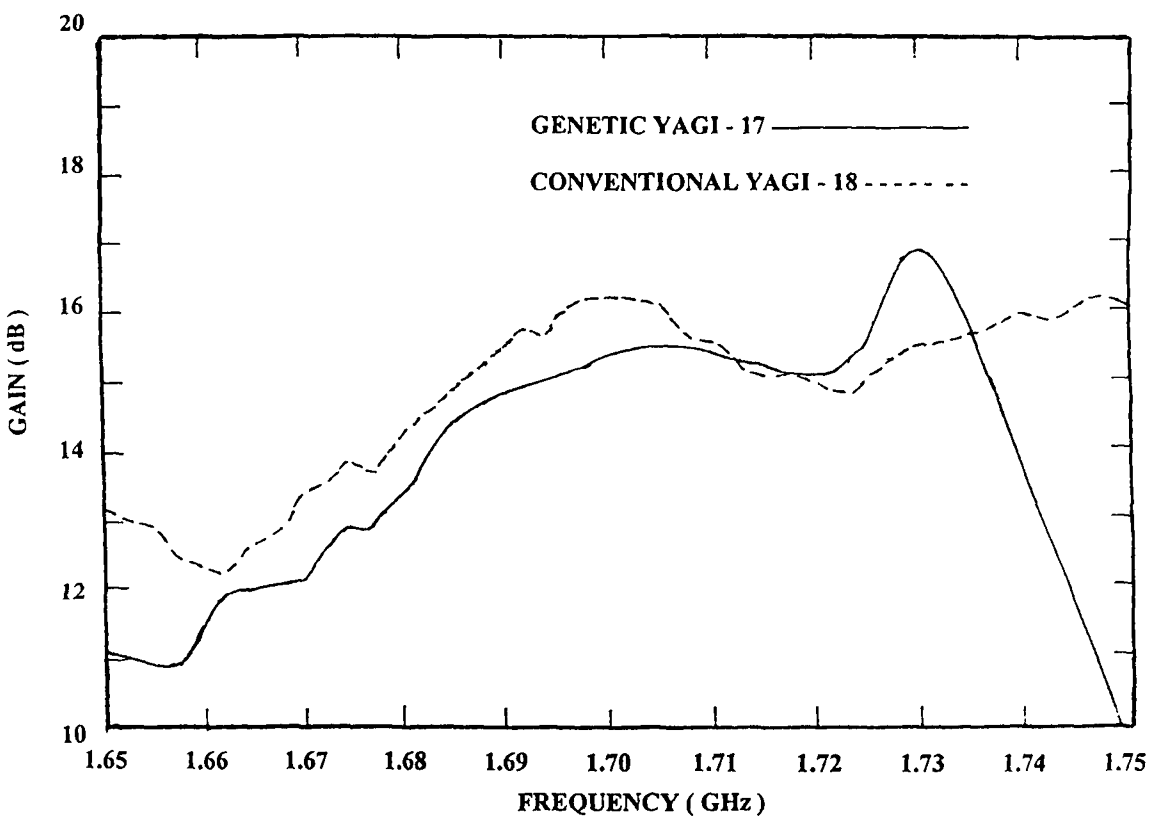
Network Analyzer over the frequency range from 1650 to 1750 MHz. The gain and radiation pattern were measured on a 2500-foot far field range using a Flam & Russell Model 959 Automated Measurement Workstation with a Hewlett Packard 8530A receiver. The transmitter was a Hewlett Packard 8340B synthesized source used in conjunction with a 4-foot parabolic dish that provided an ample signal level and dynamic range. A Scientific Atlanta model 12-1.1 standard gain horn was used as a reference antenna.

There are two main sources of error in the measurement of absolute gain: the first is the uncertainty of the true gain of the standard gain horn; this has a 1 σ value of ±0.17 dB. The second source of error arises from reflections and multipath from the terrain. We observed fluctuations of approximately ±0.3 dB as we moved the antenna back and forth along a track. Thus, there is an uncertainty of about ±0.5 dB in the measurement of the absolute gain and about ±0.3 dB in the relative gain. In addition, it is necessary to correct for the mismatches that may arise from the transmission line to both the standard gain horn and the Yagi antennas. There are two options that can be used for this correction: a matching device (e.g., stub tuner) can be inserted to reduce the mismatch losses, or the input VSWR can be measured and the corresponding loss in gain can then be calculated as follows:

$$\text{Gain loss (dB)} = 10 \log_{10} \left[1 - \left(\frac{\text{VSWR-1}}{\text{VSWR + 1}} \right)^2 \right] \qquad (2)$$

We used a combination of both methods. Because the Flam & Russell system automatically swept over the frequency range, it was not practical to match the antennas at all frequencies. The standard gain horn had VSWRs below 1.2, so the corresponding losses were less than 0.04 dB. The conventional Yagi had a maximum VSWR of 2.1, so corrections of up to about 0.6 dB were needed. For the genetic Yagi, the VSWRs were between 2.5 and 3.0 over most of the band, but they increased sharply above 1730 MHz. At three frequencies in the vicinity where the Yagis exhibited maximum gain, we used a double stub tuner to minimize the mismatch loss.

In **Figure 5**, we plot the corrected gains for the conventional and genetic Yagis as a func-



5. Measured gains of 5.16 λ boom length conventional and genetic Yagis from 1.65-1.75 GHz.

Figure 5. Measured gains of 5.16-wavelength conventional and genetic Yagi from 1.65-1.75 GHz.

tion of frequency. We note that the conventional Yagi has a maximum gain of 16.2 dB at 1700 MHz—slightly lower at 1728 MHz, the scale frequency that corresponds to the full scale frequency of 432 MHz. The genetic Yagi had a maximum gain of 17.0 dB at 1730 MHz. As expected, the gain of the conventional Yagi was reasonably flat over most of the band, whereas the gain of the genetic Yagi decreased sharply at the higher frequencies. The measured gains were somewhat lower than the computed gains; as mentioned previously, this can be attributed to the uncertainty in the gain of the Standard Gain Horn. However, the relative gains of the Yagi antennas are in good agreement with the computations.

Conclusions

It has been shown computationally and verified experimentally that by using a genetic algorithm in conjunction with NEC it is possible to synthesize, at a single frequency, a Yagi antenna design that has a significantly higher gain than that of a state-of-the-art Yagi. For this example, we chose to focus on maximum gain and did not weight the VSWR very heavily. We could also have included bandwidth and sidelobe and backlobe levels in the cost function; of course, we expect that we would have sacrificed some gain. We have previously designed, using a genetic algorithm, a Yagi-type antenna that had sidelobes and backlobes greater than 25 dB down over a 13.6 percent bandwidth.[8] Also, other types of wire antennas have been successfully designed using this process.[9,10] This algorithm provides the antenna designer with a powerful technique for synthesizing antenna configurations which have properties that have heretofore been unattainable.

Acknowledgments

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PRODUCT INFORMATION

Hand-Portable Power with Muscle

Cutting Edge Enterprises has added the Powerport 149 to its line of rechargeable, portable power supplies. The unit is compact (4 x 4.5 x 6 inches) and lightweight (9 pounds). It provides 140 watts of 115 volts AC (surges to 200 watts) and up to 20 amps of 12 volts DC power. The optional padded bag, with detachable accessory pouches, protects the unit from impact and makes transport easy.

This rechargeable power supply contains a 12-volt, 9-amp gel-cell battery. Powerport can be charged via a vehicle's cigarette lighter without the engine running. It is also equipped with a fully automatic wall charger, which can be left plugged in without fear of overcharging.

Powerport will run and charge many devices such as handheld radios, test equipment, emergency lighting, handheld GPS receivers, laptop computers. For additional information on Powerport's capabilities contact Cutting Edge Enterprises, 1803 Mission Street, Suite #546, Santa Cruz, California 95060; phone: (800) 206-0115; e-mail: <cutedgent@aol.com>.

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A SINGLE CONVERSION FM RECEIVER

For wireless data communication on the 902 to 928-MHz ISM band

This paper describes an FM receiver designed for short range wireless data links in the "Part 15" 902 to 928-MHz band. Applications include data links between PC or notebook computers and PC and peripheral devices. The receiver is a single-conversion superheterodyne incorporating a low-cost silicon monolithic IC chip set. The receiver uses surface mount components, small size pc board layout, and few external components, making it suitable for wireless local area networks (LANs) and portable applications. Techniques to enhance the receiver sensitivity are analyzed and developed, and the results are verified in a demonstration receiver.

Introduction

Wireless data applications in the 902 to 928-MHz ISM band are dominated by spread-spectrum applications that are costly, very sophisticated, and use excessive power (**Figure 1**). An alternative is to use an approved lower power transmitter and receiver to achieve short-to-medium range performance.

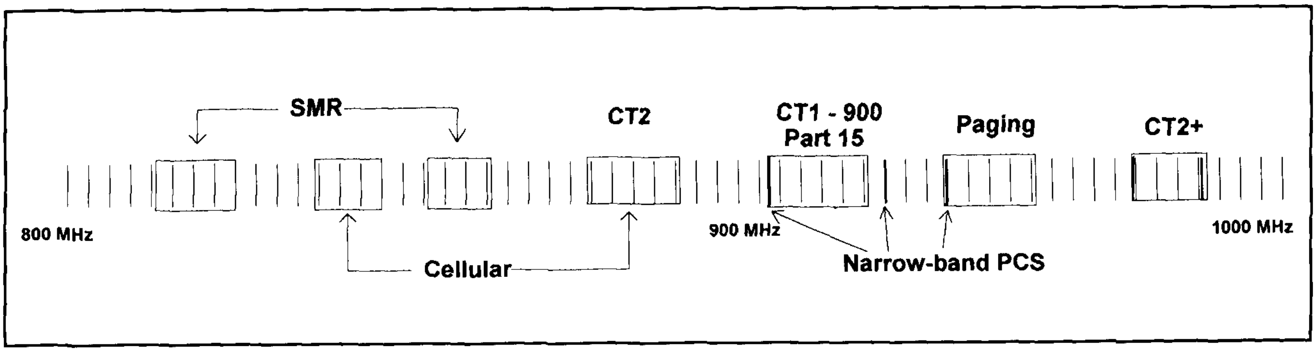


Figure 1. 900-MHz frequency spectrum.

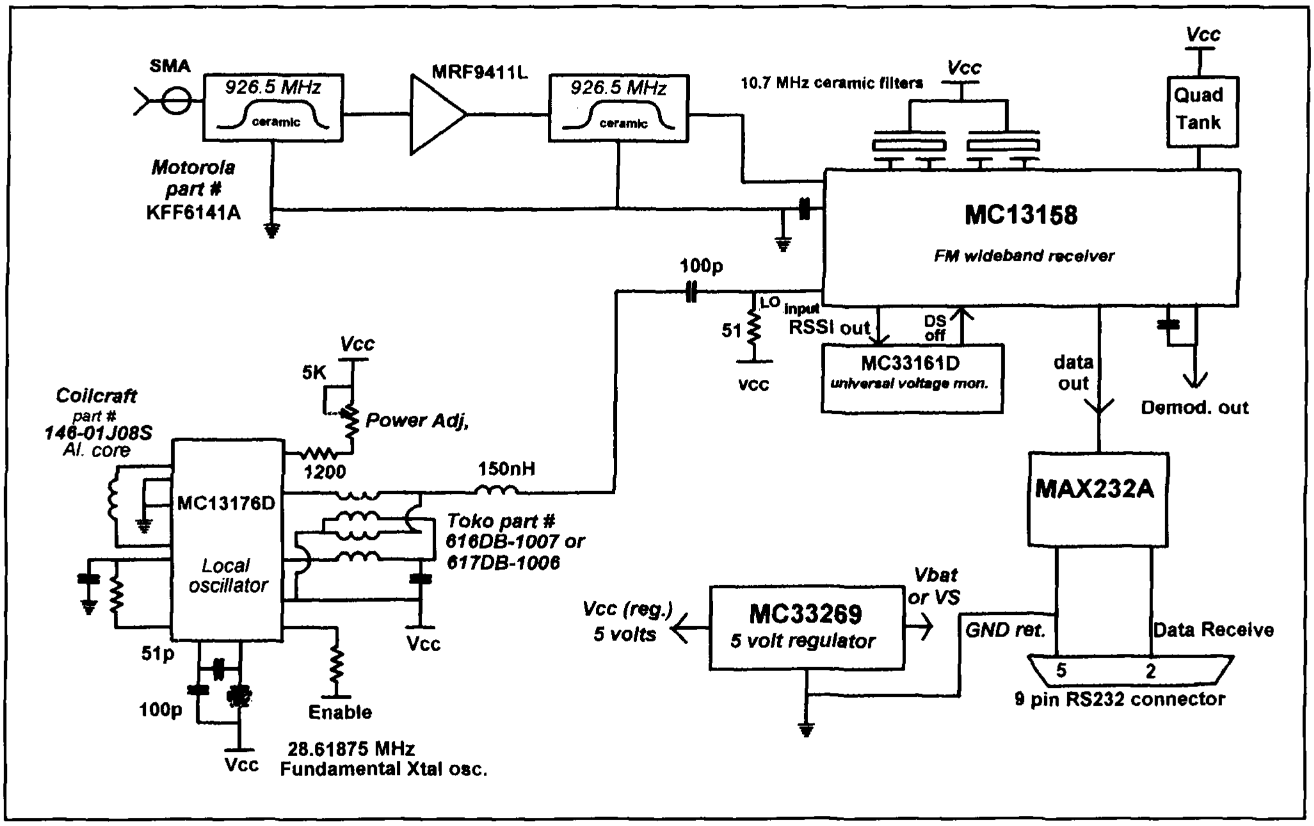


Figure 2. 926.5-MHz receiver block diagram.

A low-power transceiver in a typical office space provides 150 to 250 feet of operating range between notebook, personal computers, and peripherals like printers and fax terminals. This operating range may be reduced by commonplace obstructions, such as walls, cubicle partitions, filing cabinets, and workbenches. Interference from other RF sources like pagers and cellular phones and other manmade sources such as florescent lighting, electrical equipment, and computers may also diminish system performance. The solutions presented here are simple and cost effective, while providing an FM receiver that has excellent immunity to noise and interference.

System constraints on size, cost, and power consumption often dictate smaller and fewer parts: both passive and ICs. Surface-mount components are required over larger leaded components where pc board space is minimized, such as in PCMCIA and portable applications. In the following discussion, a small, low-cost receiver using very few ICs and external components is described.

System block diagram

Figure 2 is the block diagram of the 926.5-MHz wideband FM receiver. It is comprised of an antenna, RF ceramic filters (KFF6141A), low-noise amplifier (MRF9411L),[1] single-conversion wideband FM receiver (MC13158FTB),[2] local oscillator (MC13176D),[3] RS232/TLL converter (MAX232A[4] or MC145407D[5]), and cable and connector to interface with the computer. The system is powered from batteries or an external "bump on the wall" power source that may be regulated on the circuit board at 5 volts DC (MC33269D-5.0,[6] low dropout three terminal regulator IC).

Design concepts

FCC regulations. An operating range of typically 200 feet may be obtained under CFR Title 47, Part 15 (FCC Rules and Regulation governing unlicensed operations).[7] Paragraph 15.249 defines operation in the 902 to 928-MHz ISM band using low-power non-spread spectrum applications where the maximum transmitter field strength at a distance of 3 meters is 50 mV per meter. Harmonics of the intentional radiator must be less than -50 dBc or 200 microvolts per meter at 3 meters, whichever is smaller. Paragraph 15.109 states that unintentional radiators at a distance of 3 meters (216 to 960 MHz) also must be no greater than 200 μV per meter field strength. In a superheterodyne receiver, the first local oscillator (LO) is the likely culprit as an unintentional radiator. The LO must be attenuated at the antenna and often is shielded to contain its radiation energy. **Caution**: Manufacturers should consult with the FCC before proceeding with a new product to assure that the product design is in compliance with rules and regulation outlined in "Part 15."

Range equations relating to receive power, transmit power, and path loss. The specification of 50 mV/meter may be related to the maximum transmitter power, P_t, by the following rationale. The simplest case is to consider the radiated energy from an isotropic source: an ideal antenna that radiates energy with uniform intensity in all directions. The power flux per unit area at a distance, d, from the loss-free isotropic antenna radiating a power, $P_t(W)$, is given by:

$$P_a = P_t/4\pi d^2 \qquad (1)$$

The field strength, E, in volts/meter, at a point where the power density is P_a in watts/square meter is given by:

$$E = (120\pi P_a)^{1/2} \qquad (2)$$

The effective aperture area of the receiving antenna is related to the gain of the antenna by the expression:

$$A_e = G\lambda^2/4\pi$$

Thus, for a loss-free isotropic antenna G = 1:

$$A_e = \lambda^2/4\pi \qquad (3)$$

The power available from a loss-free isotropic antenna is determined from **Equations 1**, **2**, and **3** above:

$$E = (480\pi^2 P_r/\lambda^2)^{1/2}$$

The above equation may be solved for P_r and rearranged as:

$$P_r = (E^2/120\pi)(\lambda^2/4\pi) \qquad (4)$$

The electric field produced by a transmitter radiating power, $P_t(W)$, at a distance, d(m), in free space can be derived from **Equations 1** and **2** and is given by:

$$E = (30P_t/d^2)^{1/2} \qquad (5)$$

Substitute **Equation 5** in **Equation 4** above and solve for Pr/P_t, the path loss in dB:

$$P_r/P_t = (C^2/16\pi^2)/(d^*f)^2$$

where:

$C = 3.0 \times 10^8$ meters/second, thus:

$$P_r/P_t = 10\ \mathrm{Log}[5.7\times10^{14}/(d^*f)^2] \qquad (6)$$

Solving **Equation 6** for P_t, transmitter power in dBm:

$$P_t = P_r{}^*10\ \mathrm{Log}[(d^*f)^2/5.7 \times 10^{14}] \qquad (7)$$

Solve **Equation 4** for P_r and solve **Equation 7** for P_t in dBm,

where:

E = 50 mV/meter
d = 3 meters
f = 926.5 MHz

yields:

$$P_t = 5.53 \times 10^{-5}\mathrm{mW} + 41.3\ \mathrm{dB}$$

Since 1 mW = 0 dBm, then

$$P_r = [10\ \mathrm{Log}\ 5.53 + 10\ \mathrm{Log}(10^{-5})]\mathrm{dBm}$$
$$P_r = -42.5\ \mathrm{dBm}$$

$$P_t = -42.5\ \mathrm{dBm} + 41.3\ \mathrm{dB}$$
$$P_t = -1.27\ \mathrm{dBm}$$

The calculation above describes a radio system in which only free space conditions and loss-free isotropic antennas are considered. In the real world, antenna gain is a significant factor. If directive antennas are used in place of isotropic antennas, the free space path attenuation equation becomes:

$$10\ \mathrm{Log}(P_t/P_r)_{\mathrm{directive}} = 10\ \mathrm{Log}(P_t/P_r)_{\mathrm{isotropic}} + (G_t + G_r + L_d)_{\mathrm{dB}} \qquad (8)$$

where:

G_t and G_r are the free space antenna power gains with respect to isotropic for the transmitting and receiving antenna, respectively.

L_d is the polarization coupling between antennas. L_d equals 0 dB when the transmitting and receiving antennas have the same polarization.

The above equations form the basis of propagation, but they do not consider real world factors such as the presence of obstructions found in an office or home environment. In an actual radio system, these losses need to be added to the free space equations shown above.[8,9]

Noise analysis of overall receiver system

Cascaded noise stages. The system noise analysis block diagram in **Figure 3** shows the cascaded noise states contributing to the system

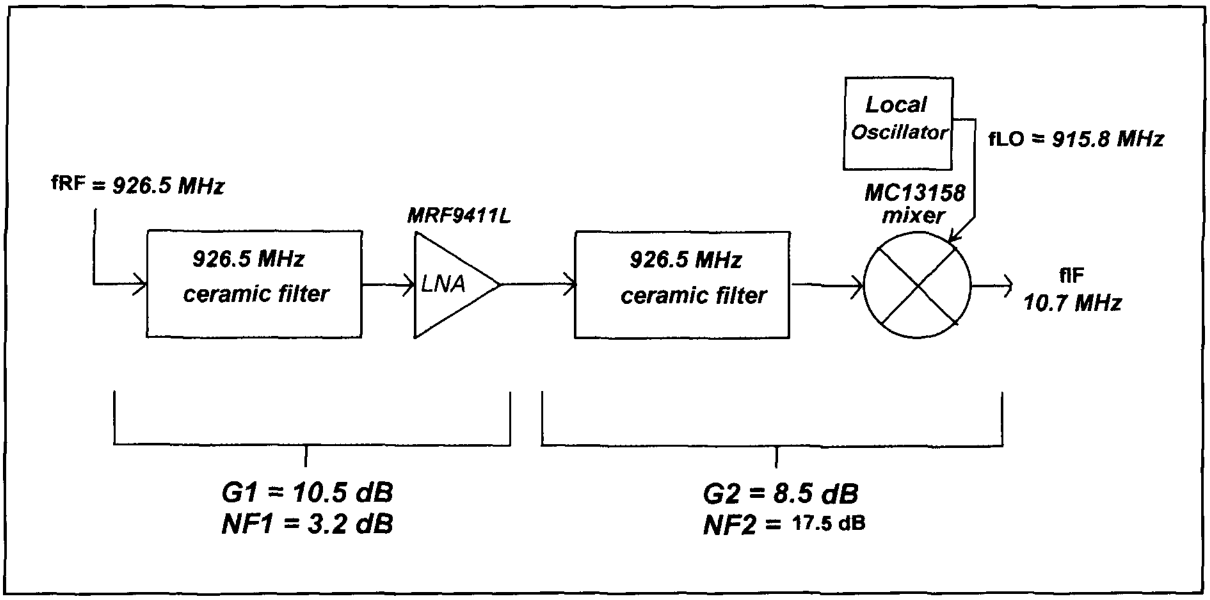


Figure 3. System noise analysis.

noise. It represents the MC13158, the low-noise amplifier using the MRF9411 transistor and RF ceramic filters, and the interface to the antenna. The noise figure of the MC13158 mixer is approximately 16 dB at 926.5 MHz, which is a typical value in a double-balanced mixer. The LNA has a noise figure of typically 1.7 dB and a gain associated with the noise figure of 12 dB. Insertion loss of the RF ceramic filters is typically 1.5 dB.

In the cascaded noise analysis, the simplified system noise equation is:

$$F_{system} = F1 + [(F2 - 1)/G1]$$

where:

F1 = the noise factor of the preamp and mixer
G1 = the gain of the preamp
F2 = the noise factor of the 2nd RF ceramic filter and mixer

Note that the proceeding terms are defined as linear relationships and are related to the log form for gain and noise figure by the following:

$F = \log^{-1}[(NF \text{ in dB})/10]$ and similarly
$G = \log^{-1}[\text{Gain in dB})/10]$

Calculating in terms of gain and noise factor yields the following:

$F1 = 5.05$; $G1 = 10.2$
$F2 = 12.4$

thus, substituting in the equation for noise factor:

$F_{system} = 6.17$; $NF_{system} = 7.9$ dB

The system noise figure is strongly dependent on the gain and the noise figure of the first stage (preselector filter and LNA). In addition to improving the noise figure, the LNA pro-

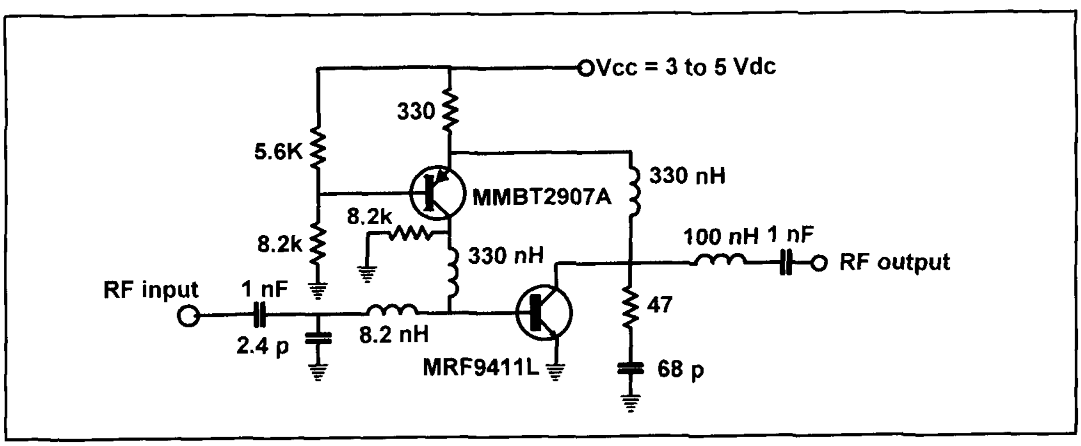
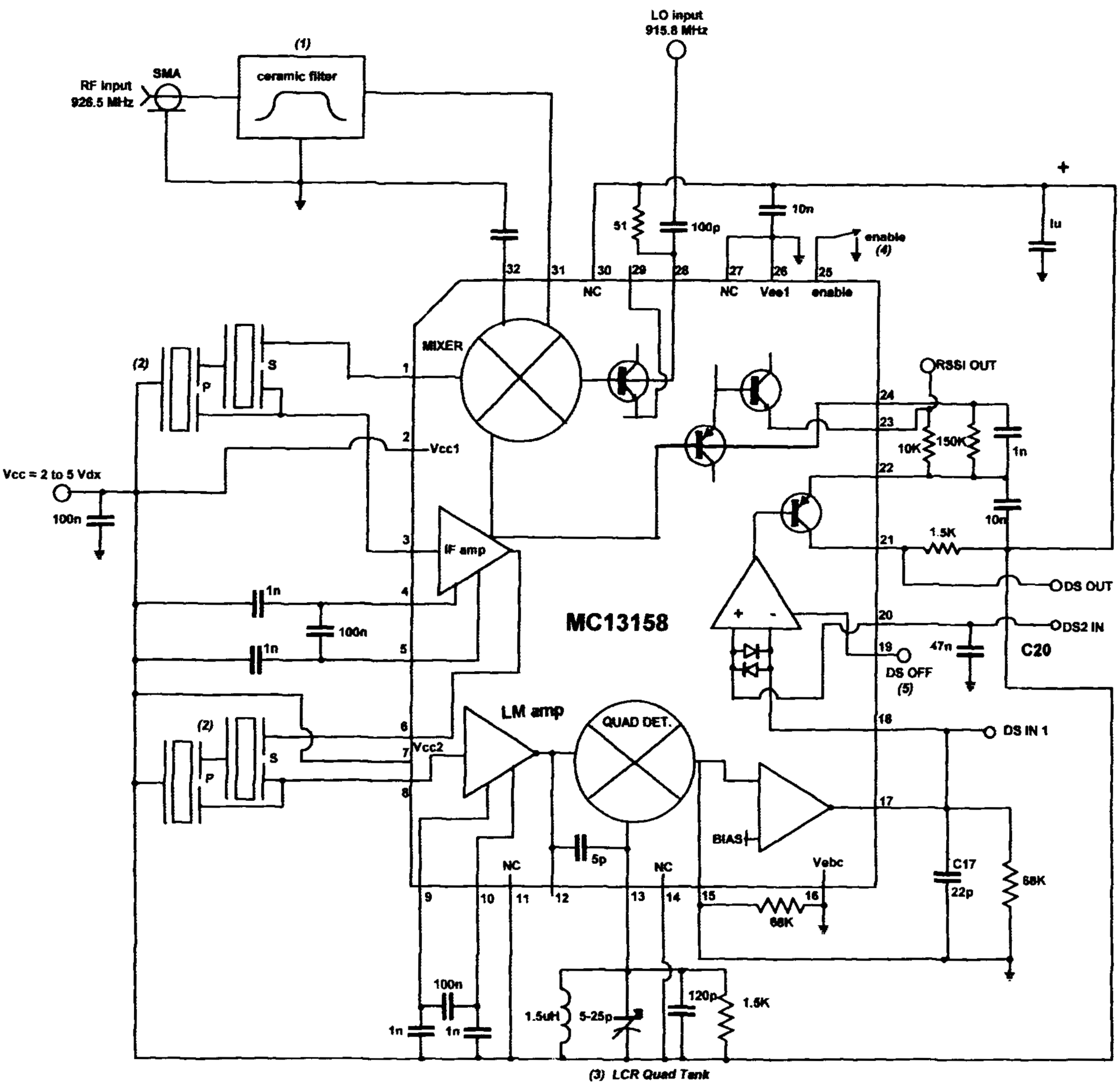


Figure 4. 900-MHz low noise amplifier.



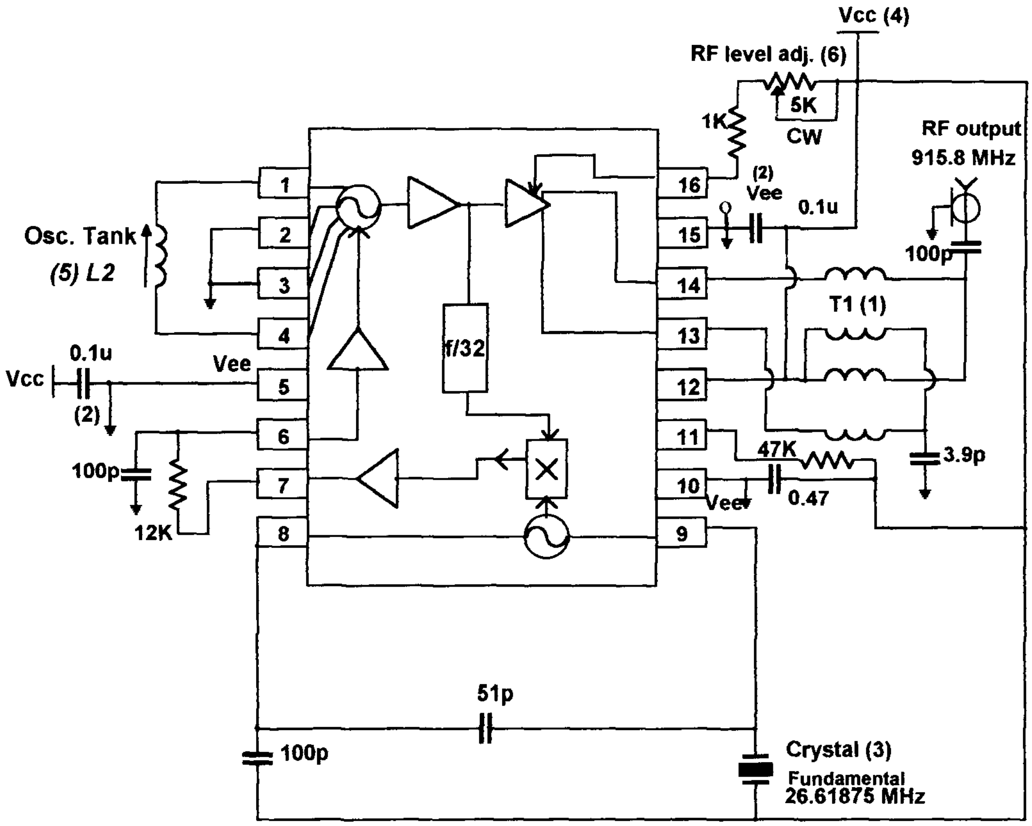
Notes:

1. RF ceramic filter at 926.5 MHz; Motorola Part #KFF614A
2. 10.7-MHz IF ceramic filters; Murata Part #KMFC545 (S, Series and P, Parallel Pairs).
3. The quadrature tank components are: L = 1.5 µH surface mount inductor, R = 1.5 k chip resistor, and C = 120 pF chip capacitor and 5-25 pF variable capacitor.
4. To enable the IC, pin 25 is taken to VEE.
5. To shut data slicer off, pin 19 is taken to VCC.

Figure 5. MC13158 wideband FM receiver.

vides reverse isolation from the LO to the antenna. Thus, enhancing the LNA performance overcomes the high noise figure of the MC13158 mixer.

Modulation method. Significant noise improvement in the receiver is achieved due to the high modulation index of the wideband FM modulation at the transmitter. The modulation index is the ratio of the maximum frequency deviation to the maximum modulating frequency. The output signal-to-noise ratio, S_0/N_0 in the receiver is increased over the receiver input



Notes:

1. Toko BALUN transformer, T1 is Toko Part #617DB-1006 or 616DB-1007.
2. Pins 5, 10, and 15 are grounds and are connected to VEE which is the component side ground plane. These pins must be decoupled to VCC; decoupling capacitors should be placed as close as possible to the pins.
3. The crystal is a parallel resonant, fundamental mode, specified at 50 PPM tolerance and a 32-pF load calibration capacitance.
4. VCC is regulated at 3.3 volts DC, with a low dropout regulator, MC33269. VBAT or Vsupply is a four-cell battery. The circuit may be operated without the regular or from a Lithium battery.
5. L2 is a 7-mm shielded variable inductor or Coilcraft Part #146-01J08S with an aluminum slug Part #51-129-169.
6. The output power is set by IMOD.

Figure 6. MC13176D 915.8-MHz local oscillator.

signal-to-noise ratio S_i/N_i, as defined in the following equation:[9]

$$S_o/N_o = 3\beta^2(S_i/N_i)$$

For $\beta = 4$, the improvement in output signal-to-noise is a ratio of 48, which is 16.8 dB.

IF filtering. IF filters following the mixer provide selectivity and reduction of noise in the high-gain IF stages; the split IF allows for additional filtering to further reduce the IF noise contribution. The IF filter bandwidth must be wide enough to pass the occupied bandwidth of the wideband modulated IF frequency.

Post demodulation filtering. Further noise improvement is accomplished with de-emphasis filtering of the recovered waveform. The de-emphasis filter is a low-pass or bandpass filter that attenuates the unwanted high-frequency components of the recovered signal, as well as the higher frequency components of the inherent "white" noise; as a result, the output S/N ratio is increased.

Design Considerations

Antenna design. System constraints often dictate the antenna type, physical size, and mounting configuration. In a wireless local area network, a simple omni-directional antenna may be the best choice to communicate with any terminal within the network. Thus, a vertical antenna is used. The antenna efficiency is compromised in its physical implementation in which the antenna is mounted as a 5/8-wavelength vertical with ground strips on the plastic cover of the system housing.

This antenna has a radiating resistance of 30 ohms because the 1/4-wavelength strips (3.03 inches in length) are in the perpendicular plane to the antenna.[10] The antenna is interfaced to the LNA with a 50-ohm coaxial cable. The length of the coaxial cable may be adjusted to obtain the optimal noise impedance to the input of the LNA. This is accomplished due to mismatch of source impedances of the antenna and the LNA. Varying the length of the coaxial cable rotates the phase angle of the source impedance presented to the LNA.

Low noise amplifier (LNA) design. A preselector filter is used before and after the LNA as an image rejection filter. The filter is a Motorola Ceramic Division component (Part #KFF6141A),[11] which has special bandpass characteristics like 1.5 dB (typical) insertion loss at 926.5 MHz, and a 3-dB bandpass at 926.5 ±1 MHz. The filter is unique with a notch of -32 dBc at 905.5 MHz. This notch is used to attenuate the image frequency at 905.1 MHz. With the image at 905.1 MHz, the local oscillator is set at 915.8 MHz; thus, the limiting IF is 10.7 MHz.

The LNA is optimized for gain and noise figure at 3 to 5 volts DC. The typical gain is 12 dB with 1.7 dB noise figure. The circuit uses an active bias network comprised of an MMBT2907 or equivalent; it maintains constant collector current over temperature. The MRF9411 circuit topology is modeled after an example in an article by Nagaraj Dixit in the March 1994 issue of *RF Design*.[12] In order to stabilize the circuit at all frequencies while retaining low noise performance, output shunt resistive loading is used. **Figure 4** is a schematic of the LNA.

Epoxy glass was chosen as the pc board material because of cost considerations. Printed circuit board layout size dictated the use of lumped components. The passive components are surface mount; inductors are Coilcraft 0805 series having unloaded Q>50.[13]

MC13158 receiver design considerations

The MC13158 is a wideband IF subsystem designed for high-performance data or analog applications. Excellent high frequency performance is achieved with low cost through Motorola's MOSIAC 1.5 RF bipolar process. The IC is comprised of 1) a double-balanced four-quadrant multiplier that has sufficient performance through 900 MHz, 2) a high-gain IF amplifier that is split to accommodate two low-

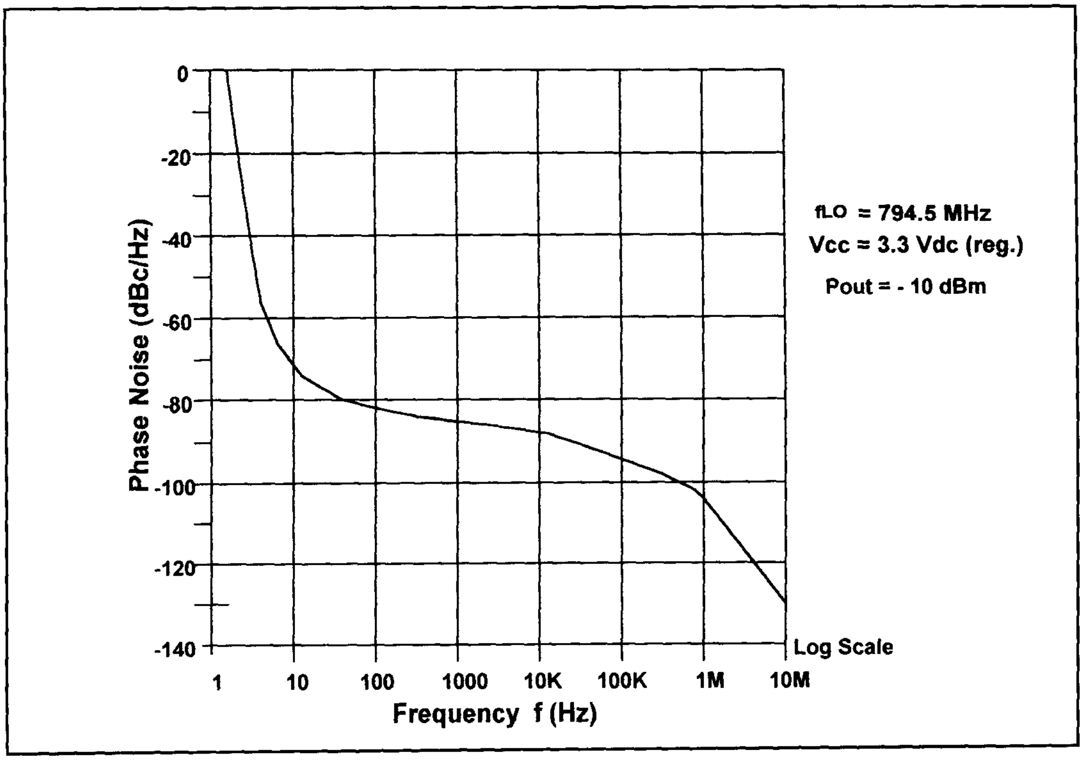


Figure 7. Typical "UHF/900-MHz" phase noise performance.

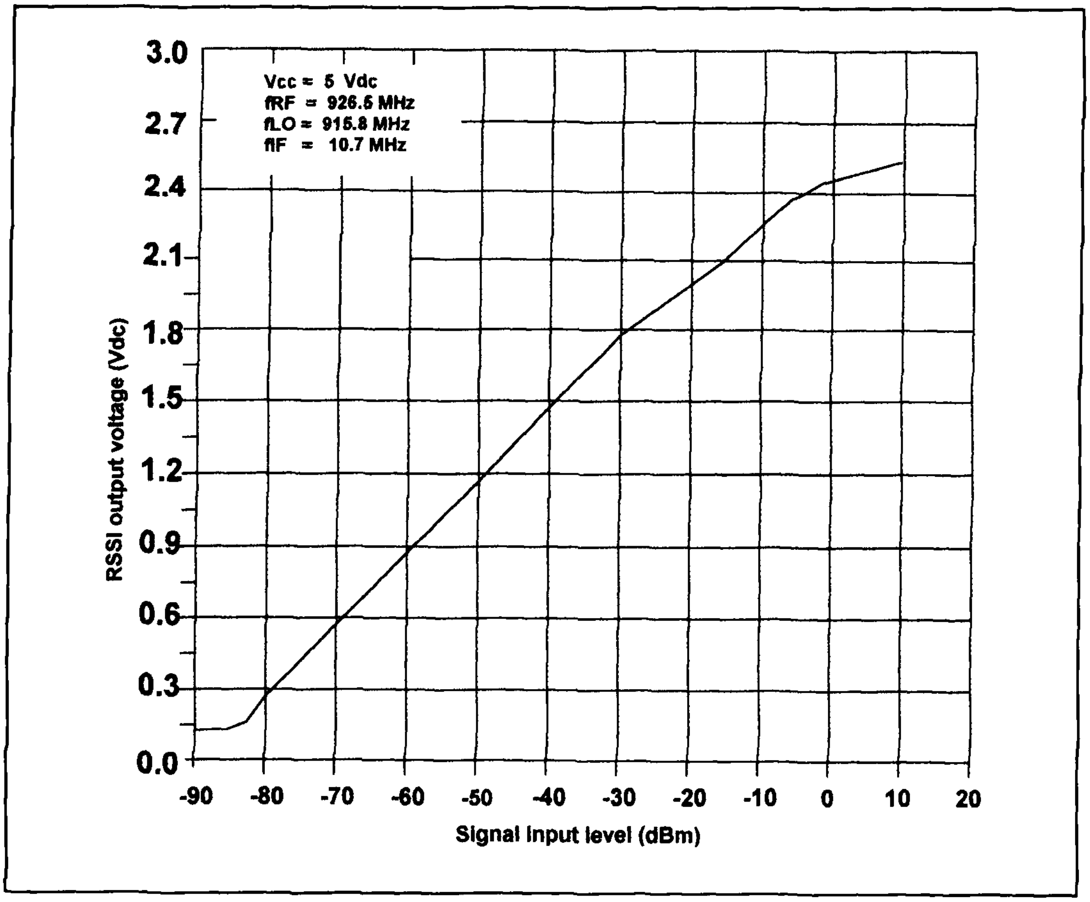


Figure 8. RSSI output versus RF input.

cost cascaded filters, 3) an RSSI output that is derived by summing the output from both IF sections, 4) a precision data shaper or slicer that has an "off" function to shut the data slicer off to save current, and 5) an enable control to power down the IC for power management in battery operated or TDMA applications (**Figure 5**). The following sections discuss each portion of the MC13158 circuit design.

Downconverter

The MC13158 mixer gain at 926.5 MHz is approximately 10 dB—reduced over its performance in the VHF band. However, the gain is sufficient to maintain noise figure and improves the intermod performance of the system. The lower gain at 926.5 MHz allows for addition of an LNA without overly impacting the third order intercept point. The mixer requires a local oscillator with -10 dBm to -5 dBm of drive; the MC13176 is selected for this function.

Local Oscillator

The MC13176D is the local oscillator at 915.8 MHz (**Figure 6**). This IC is comprised of a Colpitts crystal reference oscillator, phase detector, current controlled oscillator (CCO), and divide-by-32 prescaler forming a versatile PLL system in a low-cost 16 pin SOIC package. The phase noise is -90 dBc/Hz at 1 kHz to 100 kHz[14] (**Figure 7**).

IF/limiter, filtering, and RSSI

The choice of IF filters is important in rejection of nearby RF sources. The MC13158 data sheet discusses the IF filter configurations, such as an LCR and series-parallel, dual ceramic filters. The dual ceramic filters (Murata Part #KFMC545)[15] provide very good bandpass filtering with 3 dB bandwidth of 650 kHz with a center frequency at 10.7 MHz. If wider bandpass filtering is required, LCR filtering may be used as shown in the application example in the data sheet.

The occupied IF bandwidth is specified by Carson's Rule:

$$3 \text{ dB BW} = 2 (\text{fmod} + \text{Idev})$$

In the design example, a maximum data rate of 115.2 kbps is used. So, for a data rate of 58 kHz and a β modulation index of 4 (fdev = 232 kHz), the occupied bandwidth is:

3 dB BW = 580 kHz

Thus, the 650-kHz dual ceramic filters are a good choice.

The RSSI dynamic range is a linear 70 dB and may be used as a carrier detect to determine when to enable the data slicer (**Figure 8**). The data slicer may be shut off with no or too little RF input by monitoring the RSSI output level with the MC33161D[16] universal voltage monitor IC. This unique IC may be programmed to display no signal, low signal, and high output signal levels with LED indicators. When the RSSI output is less than the reference level, the MC33161 output goes high, shutting off the data slicer. When input signal strength is sufficient to recover the incoming data, the data slicer is enabled.

Quadrature detector tank circuit

The wideband performance of the detector is controlled by the loaded Q of the LC tank circuit. The detector 3-dB bandwidth defined as a function of the Q of the LC tank circuit is given by:

$Q/f_{IF}/3$ dB BW

where:

$f_{IF} = 10.7$ MHz
3 dB BW = 800 kHz

thus the Q = 13.4.

The loaded Q of the LC tank circuit is further defined as:

$Q = R_T/X_L$

where:

R_T is the equivalent shunt resistance across the LC tank
X_L is the reactance of the quadrature inductor at the IF frequency ($X_L = 2\pi f_L$).

Similarly, as shown in the design example in the MC13158 data sheet, the following external components of the quadrature tank circuit are calculated for quadrature bandwidth of 800 kHz:

Rext = 15 k; L = 1.5 μH and C = 139 pF

The LC quadrature circuit is implemented using a 1.5 μH surface mount inductor, a chip capacitor of 120 pF in parallel with a 5 to 25 pF variable capacitor.

Post detection filtering

As explained in the design concept section, the post detection filter response improves the output S/N of the receiver by filtering the recovered signal and attenuating the unwanted higher frequency components. The detector output is buffered, allowing for filtering and gain selection, as well as setting the average DC level of the recovered waveform. C17, the low-pass filter shunt capacitance at pin 17 is chosen much larger than the parasitic capacitance present at pin 17; a value of greater than 18 pF is sufficient. The single section low-pass RC filter response is given by:

$f_L = 0.159/R_{17}C_{17}$

Solve for R_{17} where:

$f_L = 100$ kHz; $C_{17} = 22$ pF
$R_{17} = 72.2$ k

Choose standard value $R_{17} = 68$ k.

Data slicer

The MC13158 data slicer is a comparator designed to square up the recovered signal from the quadrature detector. The coupling capacitor at the input of the data slicer is chosen to maintain a time constant long enough to hold the charge on the capacitor for the longest string of bits at the same polarity. For a data rate at 58 kHz, a bit stream of 15 bits at the same polarity would equate to an apparent data rate of approximately 3.87 kHz or 7.7 kbps. The time constant would be approximately 258 μsec. The following expression equates the time constant, t, to the external components:

$t = 2\pi(R_{17})(C_{20})/100$

Solve for C_{20}:

$C_{20} = 15.9(t)/R_{17}$

where:

$R_{17} = 82$ k
$t = 258$ μsec
$C_{20} = 50$ nF

The closest standard value is 47 nF.

Performance results

Range study. Results have shown error-free data recovery in a dense office environment at

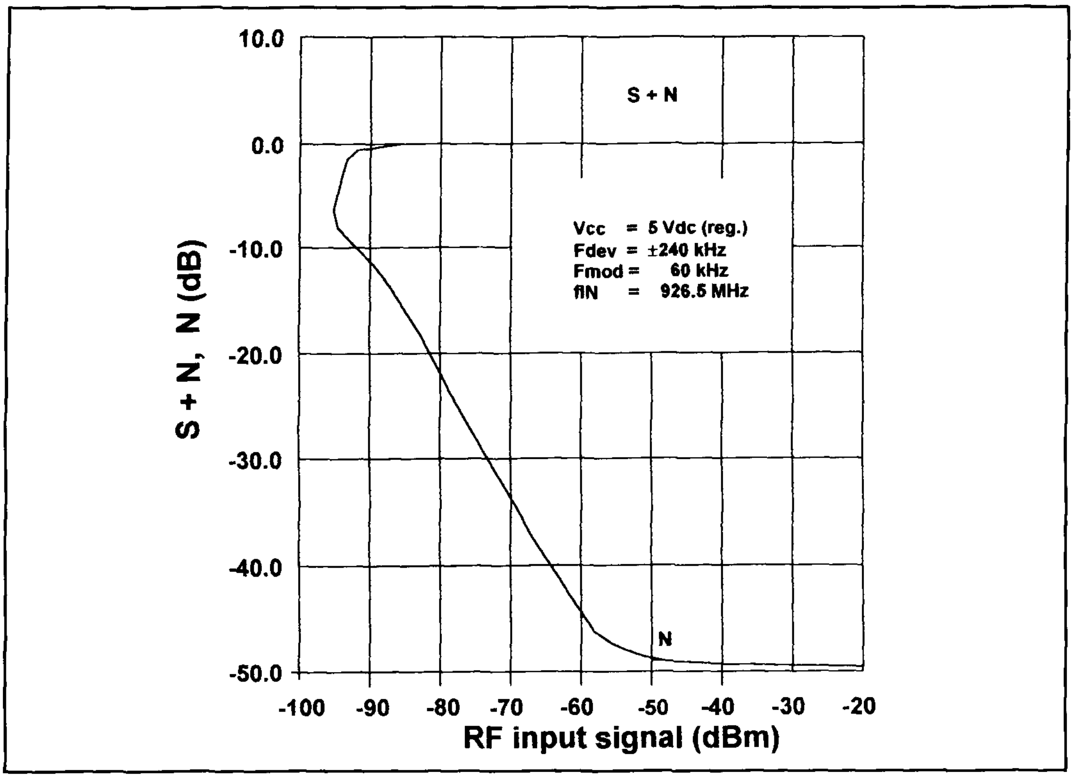


Figure 9. System SINAD performance.

a 150-foot range. Quantitative data such as Bit Error Rate (BER) and image frequency rejection performance were not taken prior to submittal of this paper.

SINAD performance. The SINAD performance curve shown in **Figure 9** was measured using a sinusoidal waveform at fmod = 60 kHz and fdev = ±240 kHz yields a 12 dB sensitivity of -90 dBm.

Unintentional radiation levels. The receiver is compliant with Paragraph 15.109; unintentional radiation levels are less than -60 dBm (200 µV/meter).

Summary

The system demo, in which data is sent between notebook computers, verifies that the system sends data over the range of 150 feet—typically without any bit errors. The receiver design meets the constraints of a low-cost implementation using the fewest ICs, small size, and having excellent medium operating range performance. ■

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ELEVATED RADIAL WIRE SYSTEMS FOR VERTICALLY POLARIZED GROUND-PLANE TYPE ANTENNAS

Part 1—Monopoles

Resonant quarter wavelength radials (electrical length at a quarter wavelength) can be used with practical elevated ground-plane type antennas and to simulate "connection" to ground for numerical modeling programs such as NEC-2, which does not allow a wire to touch lossy ground.

Introduction

From the earliest days of radio, the merits of elevated counterpoise and radial systems have been recognized because of the way in which current densities in the ground are more or less uniformly distributed over the area of the insulated counterpoise (which is, in effect, a large capacitance ground). **Figure 1** shows wires running radially outwards on insulated supports without connection being made to earth plates or ground stakes at the outer ends of the system. An alternative view of the way an elevated radial system works is that it allows the collection of electromagnetic energy in the form of displacement currents, rather than conduction currents flowing through lossy earth. Certainly,

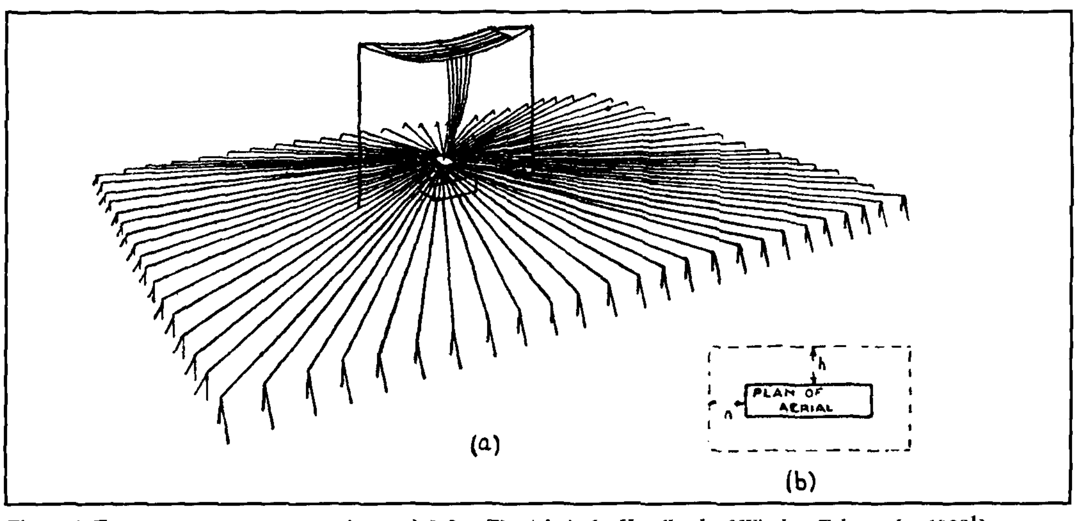


Figure 1. T-antenna over a counterpoise earth [after *The Admiralty Handbook of Wireless Telegraphy*, 1938[1]].

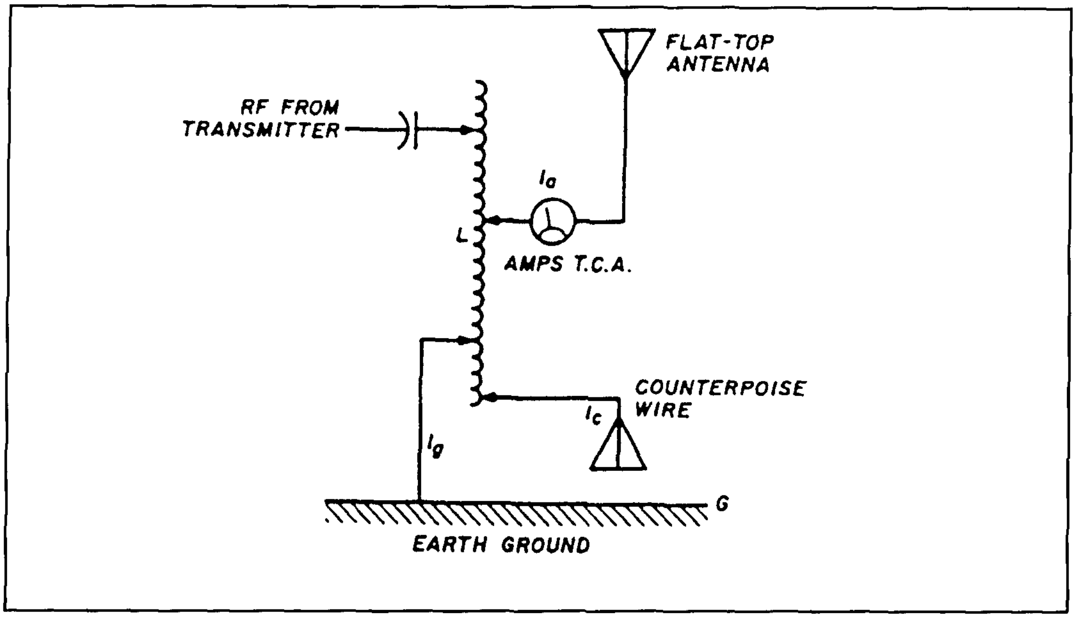


Figure 2. 2BML's "flat top" antenna system of 1921.

when only a few elevated radials are used, we cannot consider the counterpoise forming a large capacitance ground.

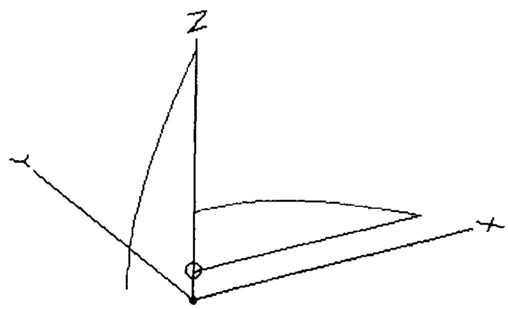
H.H. (Bev) Beverage, 2BML (now a silent key), in a 1921 publication by RCA described an aerial and counterpoise system suggested by Alexanderson and using a coupled ground wire. Details of 2BML's antenna system are shown in **Figure 2**. Having permission to operate above 200 meters, Beverage chose to tune his antenna system to 280 meters (1070 kHz). With a fair ground, his measurements showed a system resistance of around 70 ohms and 0.5 ampere antenna current. But with the elevated counterpoise attached, the system resistance dropped to 10 ohms. The ground lead tap on the inductor was adjusted to cancel the capacitive reactance of the elevated counterpoise. With both the earth and counterpoise connected, a system resistance of 4 ohms and an antenna current of 8 amperes was obtained. With 8 amperes of RF current going into the antenna, the counterpoise current (Ic) was about 6 amperes, and the earth ground current (Ig) was about 2 amperes.

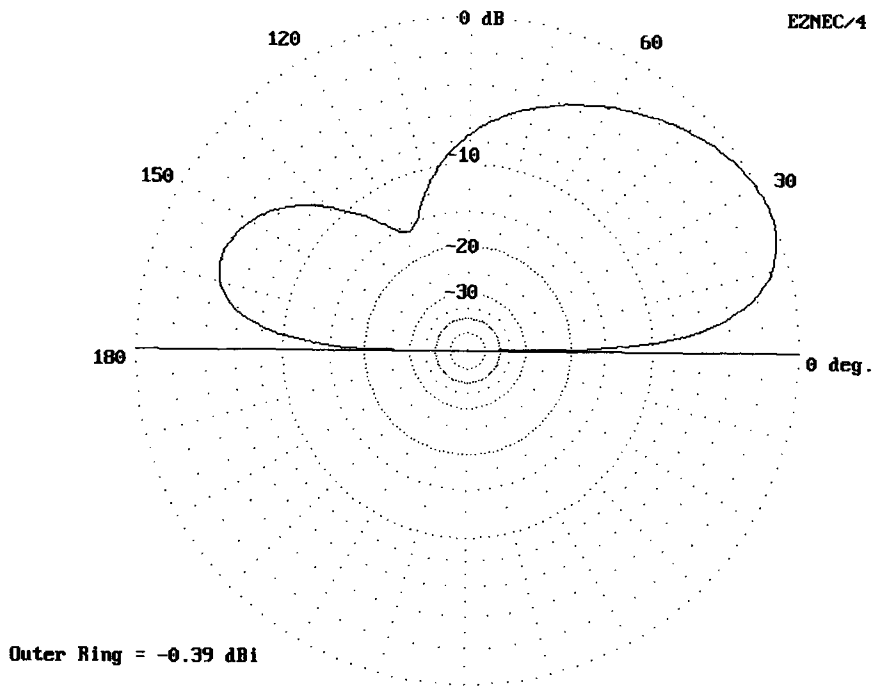
It was stated that most amateurs, and many broadcast stations (KGO was one of them) were already using the counterpoise at that time. In December 1921, the amateur radio station of The Radio Club of America, 1BCG, also employed a counterpoise system containing 30 wires each 73.5 meters long (0.31 λ for an operating wavelength of 230 meters).[2]

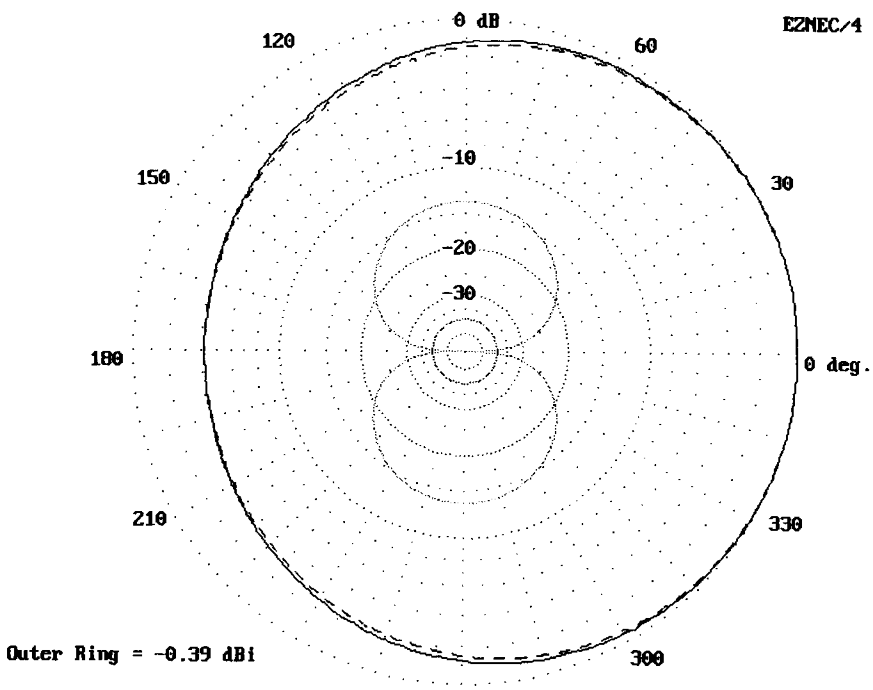
The development of such ground systems was, by necessity, empirical in nature, as sophisticated instrumentation and standardized antenna testing procedures were not available in the early days of radio. Unfortunately, after about 1937, the use of elevated radial systems became an almost forgotten art, because of research by Brown et. al.[3] in favor of the buried radial system. These authors carried out a very extensive measurement program to determine the impedance and radiation efficiency of a vertical monopole as a function of ground system parameters. A result of their work was their recommendation that 120 radial wires, each one about half a wavelength long, should be used to maximize radiation efficiency; 0.4 λ is a length quoted in many text books as optimum. This is a curious result (a half a wavelength?) in light of what we know today.[4]

Elevated counterpoise types of ground systems were extensively studied through experiments by Doty, Frey, and Mills.[5] The radial systems they used were not unlike that shown in c.f. **Figure 1** (except a perimeter wire joined the ends of the radial wires), i.e. the counterpoise radial wires filled a rectangular area over the ground, rather than being of equal or a resonant length. Christman, KB8I,[6] was perhaps the first to show by numerical modeling that a vertical monopole with as few as four horizontal radial wires could be at a height quite low over real ground before the radiation efficiency of the antenna was significantly degraded. Unfortunately the topic of ground mounted and elevated radials, and antennas with elevated feed, has remained clouded by controversy, e.g., posting on the Internet News Group on Radio Amateur Antenna; and, c.f. Belrose[7,8] and Hawker.[9] More recently Weber[10] has described measurements he made on systems employing elevated radials, which revealed

Figure 3(A). Wire model and current on the wires for a 20.2-meter vertical wire antenna with one resonant radial (length 19.184 meters for a height of 2.5 meters over average ground; (B) vertical radiation pattern; and (C) principle plain azimuthal pattern (resonant frequency 3.75 MHz).







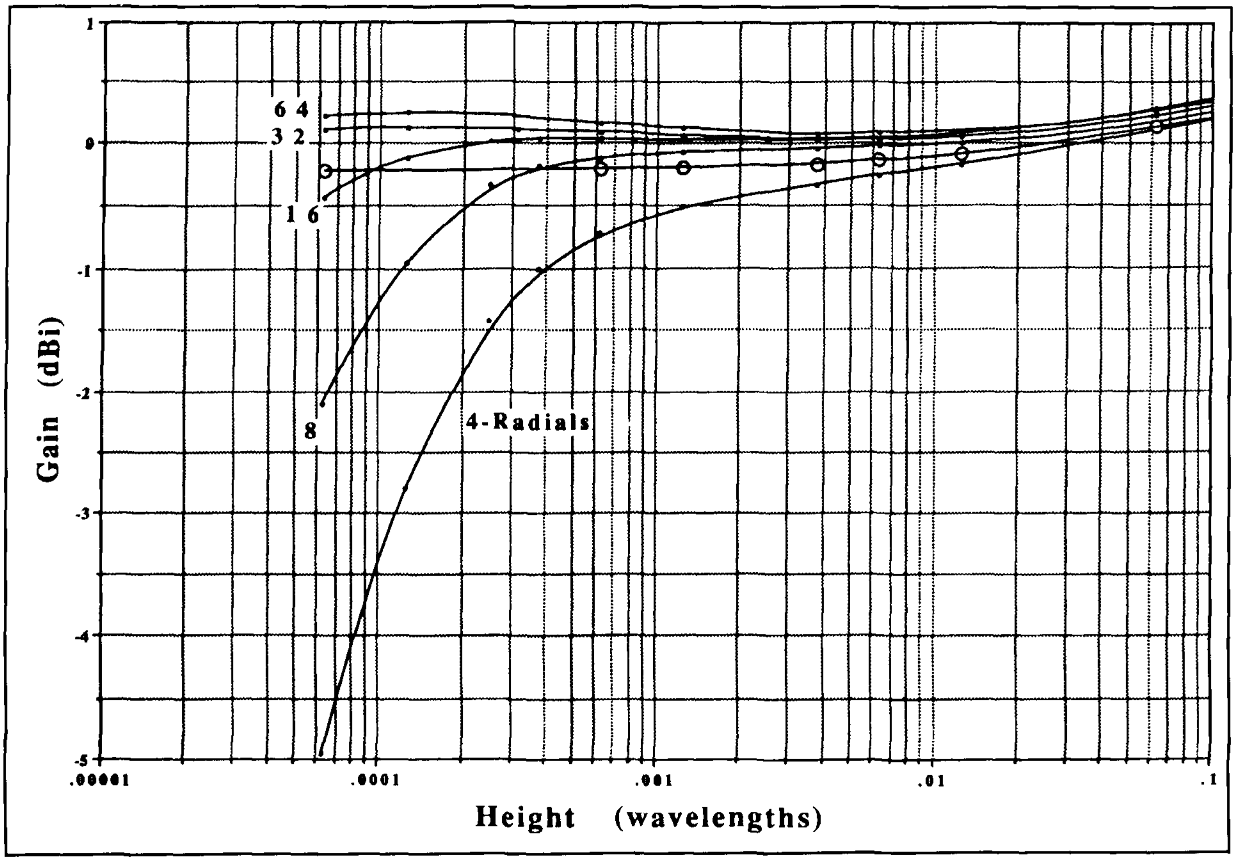


Figure 4. Predicted gain for a 20-meter monopole versus number of 20-meter radials and height in wavelengths of radials over average ground; and for a vertical half-wave dipole for reference, where height is the lower end height of the dipole. Frequency is 3.75 MHz.

unequal radial currents, and he employed numerical modeling to aid his understanding of the subject (see also comment by Severns[11] and **Appendix 1** of the present article).

To continue, if four elevated radials can be used to realize performance for a monopole comparable with many buried radials, then elevated radials can be used for real antennas and to simulate ground "connection" for numerical modeling programs like NEC-2, which does not permit a wire to connect to lossy ground.

This article is an overview of the subject of elevated radials.

Initial Experience

NEC-2 is currently enjoying wide application by radio amateurs using PC based programs such as EZNEC, available from Roy Lewallen, W7EL,[12] and EZNEC/4 pro for those registered for NEC-4. The ability of the NEC codes to accurately treat the air-ground interface (based on the Sommerfeld-integral formulation) is assumed to be validated by the developers of the program[3].

About 20 years ago, while conducting experiments to determine the efficiency of electrically short antennas, I noticed that a vertical antenna with one radial exhibited a marked directional effect. The direction of maximum gain (maximum groundwave field strength) was the direction toward which the radial ran. A front/back ratio of 10 dB was observed for a short center-loaded whip with a quarter wavelength insulated radial lying on the ground.

This discovery lead to the development of a simple transportable antenna, dubbed, by Christman,[6] the "VE2CV Field Day Special" antenna. This antenna was simply a quarter wavelength vertical wire with a tree support (or a pipe mast for 40 meters and down), with one radial elevated above head height (2.5 meters) directed toward the azimuth of principle interest. The directional pattern can be changed by installing more than one radial and selecting the appropriate radial by connect/disconnect relays.

Figures 3A through C show the patterns for such a simple antenna over average ground (ITU-R definition of average ground, viz. $\sigma = 3$ mS/m; $\varepsilon = 13$). The patterns are for a frequency of 3.75 MHz, quarter wavelength vertical (resonant length 20.2 meters), with a single resonant radial (length 19.184 meters) for a radial height of 2.5 meters over average ground. The input impedance according to NEC-4 is a good match

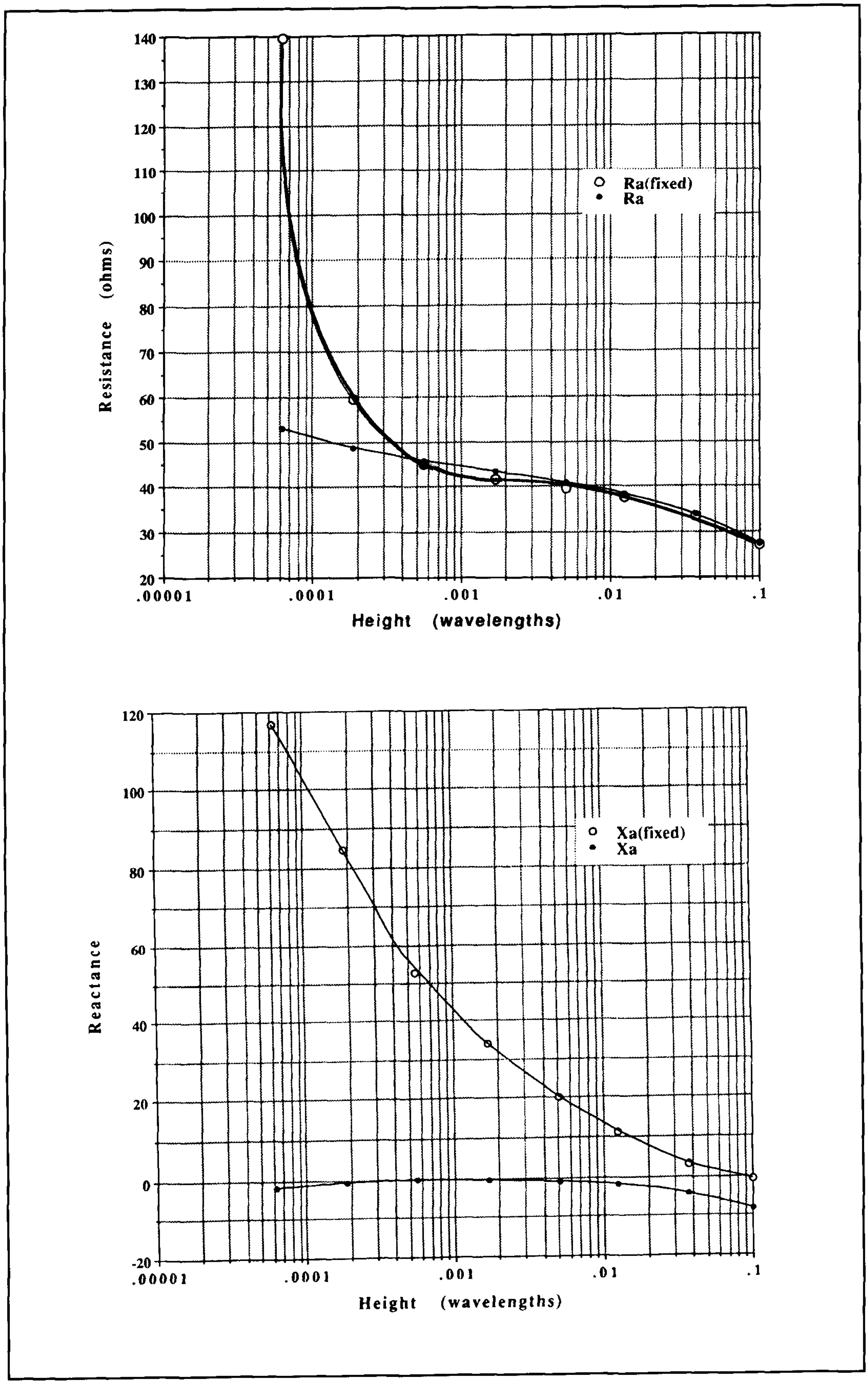


Figure 5. Predicted impedance (resistance and reactance) for a quarter wavelength, 20-meter, 3.75-MHz monopole versus radial height over average ground, for four fixed-length (20-meter) radials and for four resonant radials.

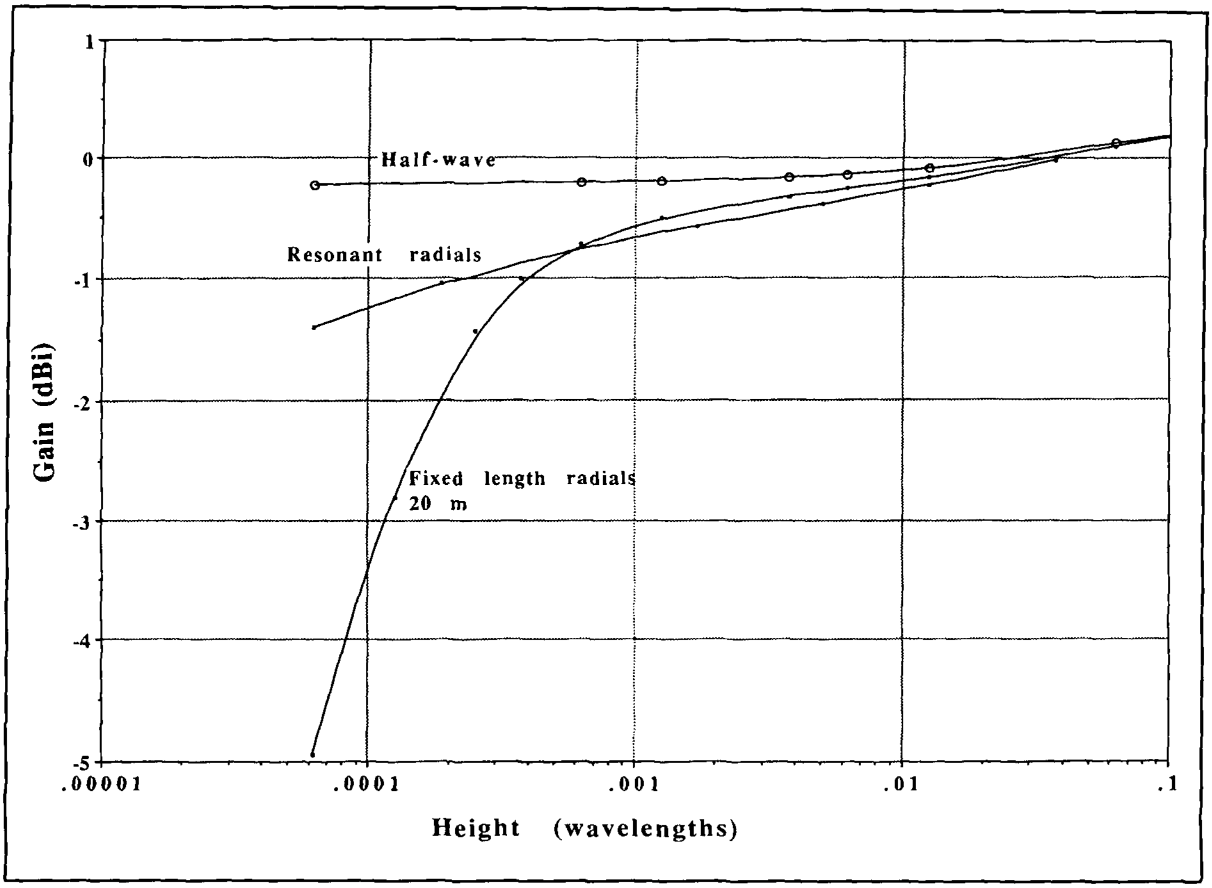


Figure 6. Predicted gain for a quarter wavelength monopole with four quarter wavelength radials and with resonant radials versus height (wavelengths) of radials over average grounds; and for a half-wave dipole where height is the height of the lower end of the dipole. Frequency is 3.75 MHz.

for 50-ohm coax (57 ohms), but a current balun should be used.

Note that **Figure 3C** shows the vertical, horizontal, and total fields—dashed, dotted, and continuous lines—since there has been some discussion about the horizontal field not canceling when there are fewer than three radials (three radials is the minimum number that can be arranged symmetrically about the monopole).

A detailed study

For the case studies to be reported here, I chose to model a 3.75-MHz monopole over a radial wire system where all elements were #10 copper wire, and where all wires, monopole height, and radials were 20 meters long (0.25 λ long for a wavelength of 80 meters).

Tom Rauch, W8JI[13], conducted a series of experiments with a 3.7-MHz quarter wavelength vertical wire antenna using elevated and ground-mounted radials. His measured relative field strengths for the ground mounted radial systems changed as the number of radials changed in a way closely predicted by NEC-4. For four, eight, and 16 radials on the ground, his measured gain differences, referenced to the relative field strength measured for 60 radials on the ground, were -5.5 dB, -2.7 dB, and -1.3 dB.

Since NEC-4 does not allow a wire to touch lossy ground in more than one place (NEC-2 does not allow a wire to touch the ground at all), I modeled a radial on the ground by assuming an insulated radial system with radials 5 millimeters high. With four, eight, 16, 32, and 64 quarter-wavelength radials, EZNEC/4 predicts gains for an antenna over average ground of -4.66 dBi, -2.16 dBi, -0.6 dBi, +0.03 dBi, and +0.23 dBi.

Note: the spacewave gain of our monopole with perfect ground beneath the antenna (equivalent to, say, an infinite number of radials), but with average ground in front of the antenna (MININEC analysis), is 0.18 dBi.

Now let us see how gain changes as a function of height (height of radials) and number of radials (**Figure 4**). For reference, the gain of a half wave dipole is shown (large open circles), where height is the height of the lower end of the dipole. Note that the lowest height, $6.25 \times 10^{-5}\ \lambda$, corresponds to the 5 millimeter height referenced above; and a height above head height (2.5 meters) corresponds to $0.031\ \lambda$ at the reference frequency of 3.75 MHz (wavelength 80 meters).

All experiments that I am aware of have used radials of a fixed length (quarter wavelength) as the height of the radial system was changed. But a radial wire at a low height over finitely conducting ground needs to be increasingly shortened as its height is lowered in order to realize and maintain resonance. The length for the ground mounted radial case study described above was for a fixed length radial (quarter wavelength long). The resonant length of a radial wire 5 millimeters above average ground is $0.138\ \lambda$, and so clearly the radials were not resonant. The impedance at 3.75 MHz, according to NEC-4, for our 20-meter high vertical wire antenna with four radials, 20 meters long, is $Z_a = 130 + j\ 114$ ohms (which is certainly not a resonant monopole), and the gain is -4.66 dBi. But, if four resonant radials, 11.1 meters long (quarter wavelength electrical length for a radial height of 5 millimeters over average ground), are used, the antenna is more or less resonant at $Z_a = 52.3 - j\ 0.9$ ohms, and the gain is -1.45 dBi.

In **Figure 5**, I have plotted the monopole resistance and reactance versus height of the antenna for four fixed-length (20 meters) radials (open circles). For resonant radials (closed circles), length is a function of height and ground conductivity, which for this graph is average ground. Notice in **Figure 5A** the sudden break between the two curves for heights less than $4 \times 10^{-4}\ \lambda$ (4.8 centimeters).

In **Figure 6**, I have plotted the predicted gain for a 20-meter-high vertical wire antenna versus height in wavelengths for four radials a quarter wavelength long (20 meters); for radials of resonant length; and the gain of a vertical half wave dipole, for reference, where height is the lower end height of the dipole. Again, notice the sudden break between the two curves for heights less than $6 \times 10^{-4}\ \lambda$.

Note: the curves on **Figures 4**, **5**, and **6** have been computed by NEC-2 (earlier work by the author[14]), but NEC-4 gives very similar results.

The reason for the marked difference between radials having a physical length of a quarter wavelength and radials having an electrical length of a quarter wavelength can be seen in **Figure 7**, where I show the wire models for these two antenna systems and the current distribution on the wires for a radial height of 5 millimeters. Since a ground system is one side of a GP antenna, its purpose is to provide a low impedance against which the antenna can be driven. Ideally this impedance should have a

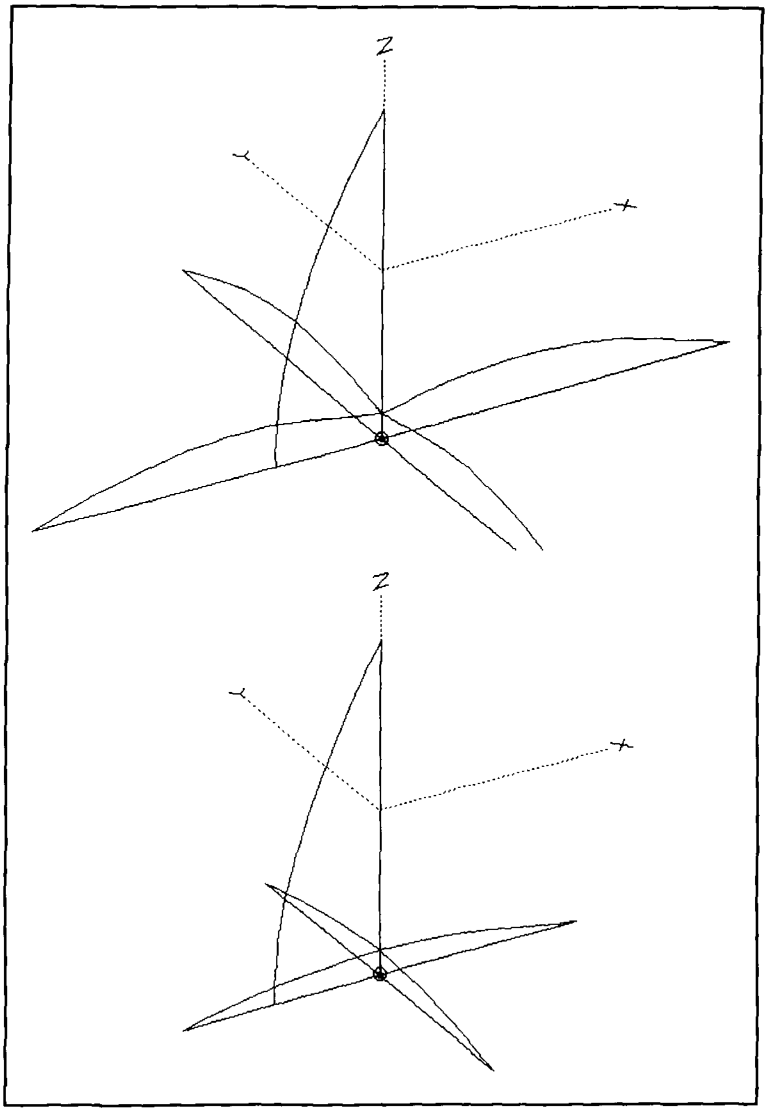


Figure 7. Wire models and current on the wires for a 20-meter vertical wire antenna with four 20-meter long radials; and with four resonant radials (length 11.1 meters) 5 millimeters above average ground.

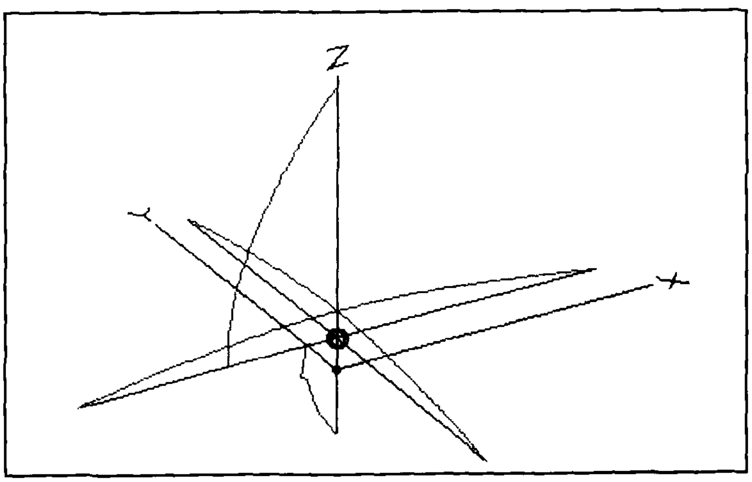


Figure 8. Model and current on the antenna system for a grounded tower (ground is a ground rod) with elevated radials, see text for details.

low resistance and only a small reactance since we want to maximize the current on the antenna. Clearly, the longer radials, shown in **Figure 7A**, do not well fulfill the requirement for a resonant radial. And, since the radials are not a resonant length, the impedance of the monopole is changed. Furthermore, since the current on the source end of the radial wires is not the maximum current on the wires, the radial system couples more current, and hence more power loss, to the ground beneath then does the shorter resonant radial system (see **Figure B**).

Comparison with measurements

There are only a few measured results to compare with predicted performance. Christman[15] measured groundwave field strengths at three distances for an 8-MHz vertical monopole with various ground systems. He made measurements with four quarter wavelength long elevated radials for three heights, 1 meter, 3 meters, and 5 meters, using direct and isolated feed, compared with 120 ground mounted radials. In his experiment, he recorded 18 field strength values, but he had to reject a block of four values since the results (impedance and field strength) were quite inconsistent with the full set of measurements. His measurements, in my view, illustrate the difficulty experienced in accurately measuring small differences in gain between different antenna systems at a different sites on different days. For example, his set of measured field strengths at three sites for a quarter wavelength monopole with four radials at 1 meter and 3 meters, direct feed, compared with 120 radials on the ground, corresponds to relative gain differences, of -4, -2, -1, 0, 0, and +1 dB, for a median difference between 0 and -1 dB.

Beverage[16] (no relation to Bev Beverage) determined the efficiency of an experimental 0.17 λ tower with six 0.25-λ radials, 0.024 λ high, at the operating frequency 1580 kHz, by measuring field intensity along 12 radial directions extending out to a distance of up to 85 kilometers. The measured RMS efficiency was 287 mV/m for 1 kilowatt radiated at 1 kilometer, which is the same measured value as would be expected (FCC files) for a 0.17-λ tower above 120 buried radials.

In spite of this good agreement between theory and experiment, there are some measurements (for a simple monopole antenna) that are not in accord with expectation. Rauch's elevated radial measurements[13] do not agree at all with prediction. He measured a trend in the change of relative gain with change in number of radials (for a radial height of 2.44 meters, or 0.03 λ at 3.7 MHz) that is almost identical with that (earlier reported) for radials on the ground. For four, eight, and 16 elevated radials, referenced to 60 radials on the ground, Rauch measured relative gains of -4.17 dB, -2.28 dB, and -1.08 dB. For four, eight, 16, and 32 quarter wavelength elevated radials, EZNEC predicts spacewave gains of -0.02 dBi, +0.08 dBi, +0.13 dBi, and +0.16 dBi.

Direct feed versus isolated feed

There is no doubt that a coax feeder supplying power to a monopole with elevated radials can have currents induced to flow on the outer surface of the shield, if the antenna is directly fed. Beverage[16] noted that if he disconnected the feedline from the antenna tuning unit input to an 1160-kHz ND tower with elevated radials (radial height 0.018 λ), and installed an RF choke made up of toroidal cores around the coaxial cable, that the antenna's impedance changed by a small amount, and a slight heating of the cores indicated that an RF current path did exist along the outer shield of the coaxial cable. When the antenna was retuned, however, no change in field strength could be detected.

I recommend however that a current balun be used to feed GP-type antenna systems with elevated radials.

Feeding grounded towers

Numerical modeling studies using NEC-2 or NEC-4 are for the case of insulated base towers, and broadcasters who have used elevated radials have used them with insulated base towers. Radio amateurs, however, have used elevated radials in a sort of reversed feed arrangement as a method of feeding a grounded tower (a tower with or without top-loading by a Yagi antenna), c.f. Russell.[17] The radials are attached to a ring centered on the tower at the point of feed, but insulated from the tower. The feed uses coaxial cable with the shield connected to the tower leg and the center conductor to the ring. The legs of the tower should be strapped together at that point. Evaluation by impedance measurement and over-the-air testing support the view that this method of feed works well, and indeed NEC-4 confirms that this method can be used, providing the height of the monopole (height above the feed height) is not greatly different from 0.25 λ (see **Appendix 2**).

I have compared the insulated base feed with the grounded tower arrangement just described for a tower height of 20 meters (a resonant height for a frequency of 3.75 MHz for the case of an insulated base tower, base height 2.5 meters, with four resonant radials of length 19.184 meters). The base impedance according

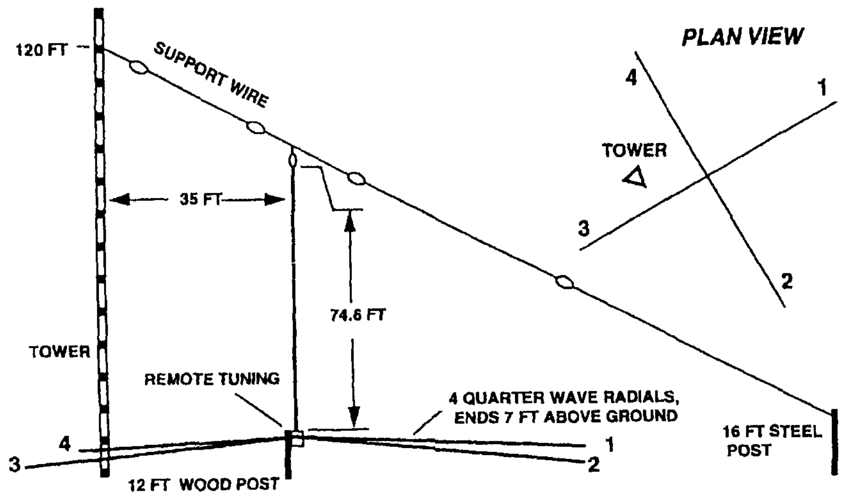
Appendix 1

K5IU's Elevated Radial Vertical Antenna Is Modeled

Dick Weber, K5IU,[10] has used an elevated radial vertical antenna suspended from a sloping wire (broken by insulators) attached to his 140-foot tower at the 120-foot level (**Figure 9A**) for about 13 years. He was quite satisfied, until recently, with the performance of this antenna, but had noted that it seemed to work better in one direction. He considers that this might be due to unequal currents on his radial wire system—unequal because the lengths of the radial wires may not have been identical.

The more likely explanation for any directional effect, in my view, is that currents are induced to flow on the tower, on any Yagi antenna it may support, and on the wire (broken by insulators) used to support his antenna. This effect is also due to directional differences in ground conductivity, a particular concern when working distant stations due to the elongated extent of the Fresnel zone for signal arriving at low elevation angles. The latter effect is a dominant factor at my QTH (see **Reference 18**) and came as a surprise since the direction of least gain was in the direction of a golf course (no buildings, perhaps good ground).

K5IU did not tell us what was on top of his 140-foot tower. Undoubtedly this tower supports a



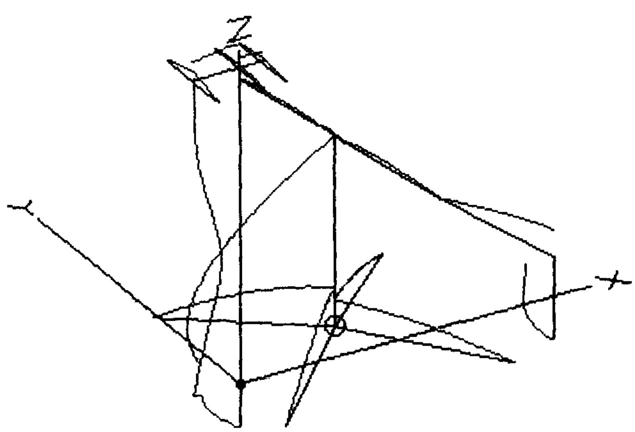


Figure 9 (A). K5IU's elevated radial vertical antenna; and (B) the model for this antenna system, modeled with a 20-meter Yagi on top of the tower. The figure shows the model and currents on the conductors (amplitude only to emphasize magnitude) at a frequency of 4 MHz.

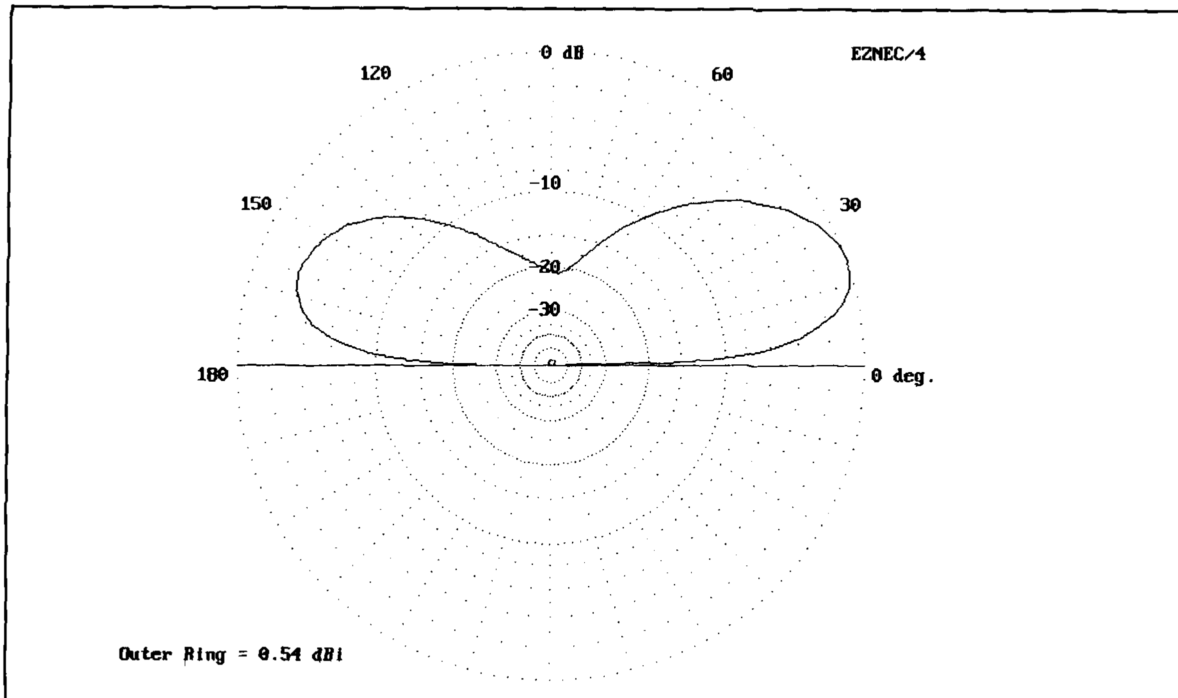


Figure 10. Radiation patterns (see text) at a frequency of 4 MHz for Figure 9(B)'s antenna system.

beam or beam antennas. For purposes of illustrating the point I wish to make, I have assumed (see **Figure 9B**) that his tower supports a wide spaced monoband 20-meter Yagi antenna. **Figure 9(B)** also shows the currents on all conductors (traces superimposed on the antenna diagram). It is clear that, as expected, an appreciable current is carried by all conductors: the grounded tower (in this model I have used a 5-meter ground rod), the 20-meter Yagi, the sections of the support wire, the grounded steel post, and the piece of wire connected to it. The phase and amplitude of these currents are frequency dependent, and so is the resulting radiation pattern.

The vertical plane pattern, for a modeling frequency 4 MHz is shown in **Figure 10(A)**; the azimuth angle is 0°. The principle plane azimuthal pattern is shown on **Figure 10(B)** at an elevation angle of 19°.

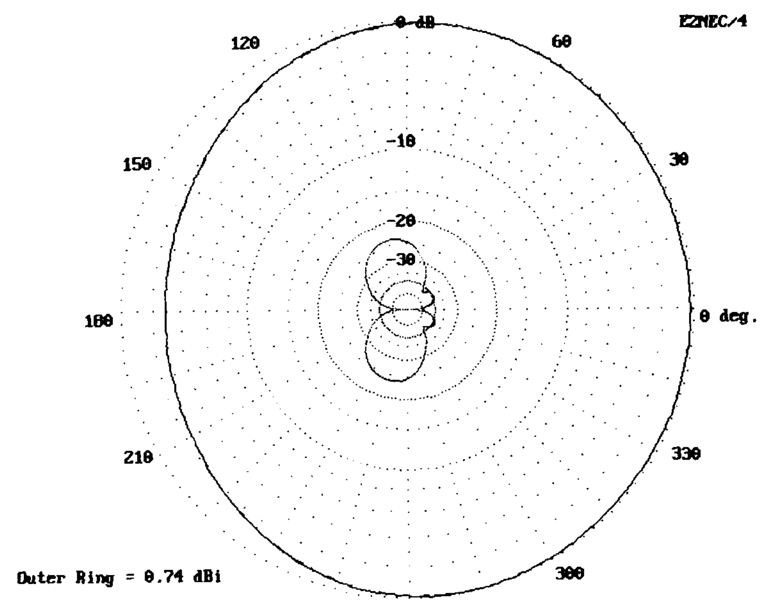
to NEC-4 is 36 - j 0.8 ohms, and the spacewave gain 0.01 dBi. For a grounded tower, "ground" being a 5-meter ground rod (a somewhat impractical length but no problem for modeling), with the four elevated resonant radials as above, the source impedance is 40 + j 15 ohms and the gain is -0.63 dBi. The small decrease in gain is certainly due to current on the part of the tower below the feed, which for 1 ampere $\angle 0°$ into each radial wire is 1.1 amp $\angle 110°$. Note that, for this grounded tower model, the tower height is 22.5 meters since we have kept the height of the tower above the radial wires 20 meters.

Figure 8 shows a wire model for the grounded tower arrangement and the currents on the wires (current amplitude only is shown).

Conclusions

Elevated radial ground systems have many applications, including: 1) realizing achievable gain for ground plane type antennas (monopoles and loops) over real ground with only a few radials; 2) providing versatility, e.g., control of radiation pattern; and 3) modeling to simulate a connection to ground. To model GP type antennas, one would often want to simulate a connection to ground; and, for purposes of modeling, a connection to ground is sometimes desired to simulate the effect of support structures or other grounded towers in the vicinity of the antenna system.

To work well, and to not change the resonant frequency of the antenna system, *the radials must be resonant.* On discovery that resonant radials should be used, in retrospect, this should have been anticipated, it is perhaps a surprise to see how different the characteristics of the antenna can be, compared with using non-resonant radials. This is particularly striking when there is a significant difference



Notice in particular the strong current on the 16-foot steel post in **Figure 9(B)** and on the piece of support wire connected to. While this certainly has an effect on the radiation pattern, we can remove the support wires entirely from our model (assuming a nylon rope), and we'll still find a pattern not unlike that shown in **Figure 10**.

For our purposes, there is no need to model K5IU's antenna system exactly, because I do not have data on the directional pattern, and for whatever model, the effects are very frequency dependent (the direction gain effect is less at 3.5 MHz compared to 4 MHz for this model). The point of this analysis is only to illustrate that one cannot hang a vertical antenna in front of a tower and carry out a numerical modeling analysis to calculate gain and pattern as though tower were not there.

between the electrical and physical length of the radial wire, such as a radial wire at a very low height.

We should note that the case studies discussed here are for ideal antenna systems, i.e. there are no nearby antenna systems or towers which could affect impedances, currents, and radiation patterns, c.f. **Appendix 1**.

A detailed study by the author (experimental and numerical modeling) of full- and half-wavelength ground-plane type 80-meter transmitting loops, where elevated resonant radials are used to simulate connection to ground for half-loops, has recently been published.[18]

In Part 2 of this article, I will discuss case studies that confirm that phased array directional systems can indeed employ elevated resonant radials, and I'll note, in particular, that the currents on the radial wires can be unequal.

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Appendix 2

Predicted Performance for Monopoles with Elevated Radials for Heights Much Different from a Quarter Wavelength

My analysis so far has been concerned with vertical antennas employing elevated radials, but for heights not very different from 1/4 λ. Certainly resonant elevated radials can be used for any antenna height, providing the antenna is of the insulated base driven type (see **Figure 11**).

But if we employ the method of feed described by Russell 17 (see section in the present article entitled "Feeding Grounded Towers"), the height of the tower above the feed height must not be too greatly different from λ/4. Compare the predicted performance (space wave gain for antenna over average ground) for the grounded tower radial feed antenna type with the conventional base insulated base fed antenna (see **Figure 11**)—both antennas employ four resonant radials at a height of 2.5 meters. The graph is for a tower with no top loading, no Yagi antenna on the tower. Height in wavelengths is physical height. Problems with the grounded tower, reduced gain, will occur for lower tower heights if the tower supports a Yagi, since the electrical height of the tower with top loading is increased.

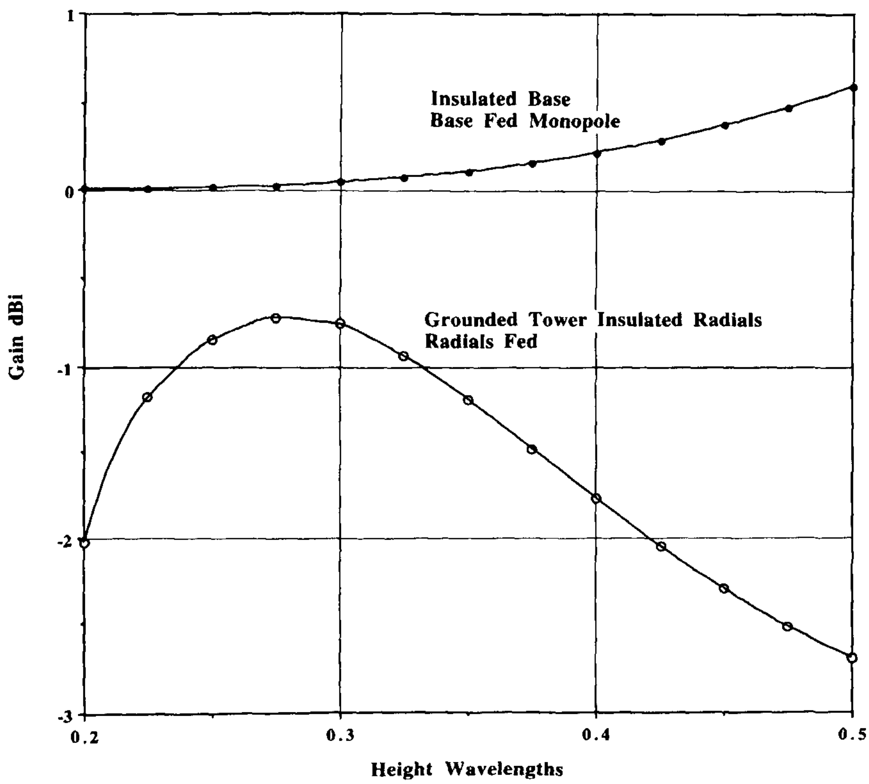

Figure 11. Gain versus height of monopole above feed height (wavelengths), frequency 3.75 MHz, average ground, elevated radials, radial height 2.5 meters. The upper curve is for the conventional method of feed, viz. insulated base tower base feed. The lower curve is for a grounded tower, radials insulated and fed.

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SCIENCE IN THE NEWS

Flexible semiconductors and the Rotman lens at MMV frequencies

The Future of Electronics: Peel-off Semiconductors

The incessant cry for ever more sophisticated electronic devices has scientists and engineers bending over backwards to supply the demand.

In something of a serendipitous discovery, researchers in pursuit of smaller, faster, and less expensive semiconductor chips have stumbled on the design of a semiconducting material that bends over backwards itself: the first single-crystal semiconductor that bends, but does not break.

The new semiconductors, some made with ordinary weather-stripping silicone, are so flexible that they can be peeled off their substrates like address labels, say University of Buffalo (UB) researchers.

"When we think of semiconductors we think of a crystal, something very hard and very fragile," said Hong Luo, an assistant professor of physics at the University of Buffalo who, along with Athos Petrou, another UB physics professor, headed the research. "But these semiconductors bend like rubber."

Characteristics

Despite their pliability, the new components (**Photo A**) retain both their structural integrity and their optical properties, characteristics that make them particularly applicable in the future of photonics—optical computing—where data is transferred by waves of light instead of streams of electrons. The devices are expected to make ideal components in optical waveguides (the optical computing equivalent of

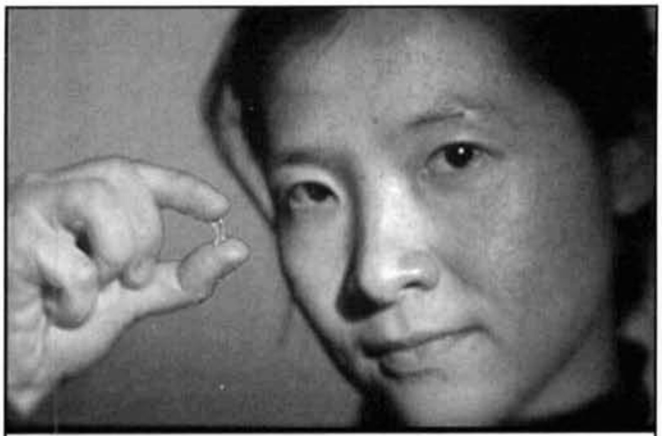
(University at Buffalo/Frank Cesario)

Photo A. The first flexible semiconductor to retain optical and structural properties is shown here by Myung-Hee Na, graduate physics student at the University of Buffalo, who helped with the invention of a device that may revolutionize photonics. (Photo by Frank Cesario/University of Buffalo).

wires), which are required in both fiber optic-based telecommunications and cable television links. The waveguides and semiconductors can now be contained inside the same component.

Their flexibility also makes it possible to transmit optical signals in three-dimensional optical circuits, making their applications far more efficient and allowing for a far more versatile design than is currently possible with two-dimensional light transmission.

A significant discovery

The researchers believe they have discovered something significant. "These semiconductors could help expedite the transition from electronics to optical computers by allowing us to exploit optics in semiconductors much more efficiently than has been possible," said Luo. "We have developed a general technology to be used with all semiconductors." For instance, besides applications in telecommunications, the discovery will likely yield new generations of high-efficiency products, such as electronics and solar cells for the space program, where flexibility will allow components to withstand the vibration and stress of rocket launches. Lightweight, flexible electronics capable of withstanding battlefield commotion will likewise appeal to the military.

According to Luo:

> Our immediate goal is to make flexible LEDs, waveguides, and other opto-electronic devices. Such devices will involve the flexible semiconductors as the 'active' material, with polymers being the mechanical support. We are using different materials so that various waveguides can be covered. At the same time, we are pursuing pure electronic devices along the line of flexible (sometimes called 'plastic') computer chips.

The new semiconductors are flexible because they are deposited on substrates in such ultra-thin layers. This is done using a technique called molecular beam epitaxy (MBE), which produces high-quality single-crystal specimens by depositing thin films on substrates in a carefully controlled ultra-high-vacuum chamber. This technique is particularly suited to the growth of compound (i.e., gallium arsenide) semiconductors. The high vacuum assures the highest levels of purity.

Luo explained:

> Theoretically, if you could make it thin enough, even a diamond could be flexible. But such thin materials are, of course, extremely fragile. They need to be supported by something, which makes this a problem for physics. We have to figure out a way to give mechanical support to this type of semiconductor structure.

Using MBE, Luo and colleague Petrou grew quantum wells—structures so thin they obey the laws of quantum physics, not classical physics. The wells were originally grown out of zinc selenide and zinc cadmium selenide on gallium arsenide, a typical semiconductor substrate. Next, the MBE-grown sample was bonded to the silicone, after which the gallium arsenide was etched away. What remained was a one-micron thick quantum well structure on top of the silicone.

An accidental discovery

The flexibility characteristic of the semiconductor was discovered accidentally by Luo and Petrou's graduate students. Luo said they were trying to glue the semiconductor to another piece of semiconductor, but it didn't glue very well and just came off. They thought it was ruined. The next day, however, when the researchers performed optical testing on the material, to their surprise, all of its optical and structural properties were found to be intact.

Luo said that while other semiconductor work has achieved flexibility using inexpensive polymeric materials, they have not always performed as well as inorganic semiconductors in their ability to emit light or to maintain structural integrity. "The flexible semiconductors we've developed are manmade structures that are fabricated using conventional semiconducting elements," he said. "Such materials possess superior optical properties and can be combined with polymeric materials because both are flexible."

The semiconductors fabricated at UB are only about one centimeter in diameter, but the researchers said industrial facilities should be able to adapt the technology to the construction of samples up to five inches or larger in diameter. Said Luo:

> Our initial objective was to make single crystal semiconductors that can be used for photonic devices. Because the drive for polymers for photonic applications is currently quite strong, the addition of semiconductor/polymer structures would greatly enhance application possibilities. We started out with II-VI semiconductors (ZnSe and ZnTe-based quantum well structures), but have extended that to GaAs II-V and other II-V structures.

For instance, they are now exploring silicon. "This way the technique will be truly universal and will cover needs in photonic applications and general electronic devices," Luo said.

For more information

The discovery is described in *Applied Physics Letters* ("Fabrication of flexible monocrystalline ZnSe-based foils and membranes"), September 9, 1996, page 1608.

Alternative Millimeter Wave Antenna Makes Gains

University researchers have designed and built the prototype of what they believe is a low-cost, high-performance, electronically scanned antenna that could be an alternative to current millimeter wave antenna technologies.

The Georgia Tech prototype is the first Rotman lens (**Photo A**) to operate at millimeter wave (MMW) frequencies in the Ka-band (33 to 37 GHz). Applications for such a device—which has no moving parts, no phase shifters, and can be packaged in plastic—could be found in a range of military and civilian fields, including aircraft landing systems, communications equipment, auto collision-avoidance systems, missile seekers, and tank radars. In these applications, wide scan angles, low sidelobes, low insertion loss, low cost, and rapid inertialess scanning are desired. The Rotman lens may be the answer.

According to Georgia Tech's Ekkehart (Otto) Rausch, senior research scientist:

> MMW components are compact and well suited for integration into missile seeker heads, smart munitions, automobile collision avoidance systems, and synthetic vision systems. Most MMW antennas that operate at frequencies equal to or greater than 35 GHz use either a mechanical scanning approach or phase shifters for electronic steering. Phase shifters that operate at MMW frequencies are costly and introduce considerable RF loss. Mechanically steered antennas are relatively low in cost, but are slow in response, sensitive to shock and vibration, and have moving parts that are subject to wear and failure. Thus, low-cost, high reliability, and electronic scanning are generally incompatible unless the design is based on a MMW Rotman lens.

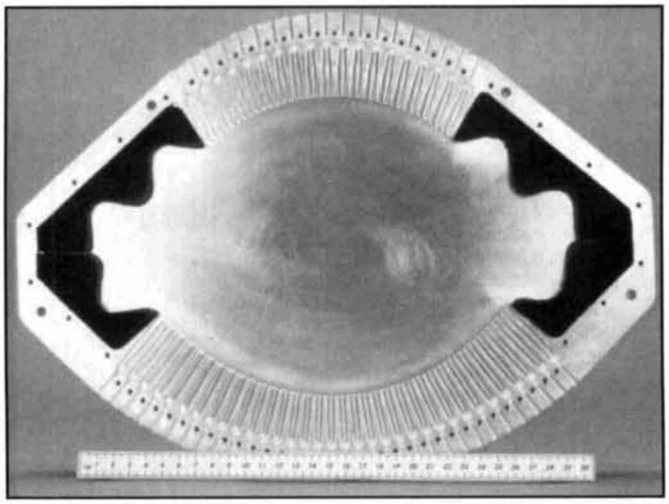

Photo A. Internal view of Rotman lens antenna.

Design particulars

Rotman lens devices get their names from their ability to focus microwave or millimeter wave energy coming from a particular direction by passing this electromagnetic energy through a pair of lens-shaped parallel plates.

As Rausch explains:

> This Rotman lens consists of a parallel plate region with waveguide ports distributed around the periphery of the plate. Beam-forming or focal ports are located on one side of the plates. These ports are fed by a switch array. The array ports are on the opposite side, each connected to an antenna element. Energy, when input into a specific focal port, will emerge from the antenna elements and produce a beam along a particular direction. Switching the input from focal port to focal port will steer the beam electronically in one dimension. The concept may be extended to two dimensions by modifying the Rotman lens equations and generating two-dimensional surfaces for the focal and array contours.

Since the objective of this research was proof-of-concept, the design was restricted to one-dimensional scanning.

Currently, the lens is constructed of aluminum. The potential exists, however, to implement both the lens and the switch array in plastic. Production costs can be kept low by hot-pressing the antenna in plastic, which could then be coated with a conductor, such as gold. The antenna feed horns and switch arrays would be assembled the same way.

Low throughput loss and sidelobe emissions

Other features of the Rotman lens, in addition to low cost, durability, and compact size, are the low throughput loss and sidelobe emissions. Sidelobe power in the prototype can be suppressed by a factor of 1,000 below the energy of the main beam. The lens itself loses less than 2 dB power.

"The antenna we are designing for Phase II of this project will have a gain in excess of 30 dB," said Rausch. "This antenna gain will be achieved with an array of 32 horn antenna elements having a beamwidth of less than four

Photo B. Otto Rausch attaches waveguides to test Rotman lens.

degrees in azimuth and less than eight degrees in elevation."

In the past, Rotman lenses were implemented with microstrip or stripline technology, usually between 6 and 18 GHz. Microstrip, however, is unreliable at high frequencies and is, therefore, not suitable for use in the millimeter wave region. To reduce the losses to an acceptable value (between 1 and 2 dB), waveguides (**Photo B**) and air dielectric must be used between the parallel plates.

Rausch and Andrew F. Peterson, Georgia Tech School of Electrical and Computer Engineering, fabricated the Rotman lens MMW antenna out of a solid block of aluminum using designs that specified tolerances of 0.0005 inches. A fabrication facility in New Jersey carved out the necessary shapes at those tolerances using an electrical discharge method.

"Everything about this lens, from the width of the waveguides to the shape of the absorber foam, matters a great deal," Rausch said. "The surface roughness and even the placement of the screws all have been designed according to strict design principles."

Applications

Potential applications for the antenna include:

• Aircraft landing systems. Heavy fog and other weather extremes can limit landing strip visibility, preventing pilots from seeing the runway. A synthetic vision system based on millimeter wave radar technology could produce heads-up images through the fog, permitting safe landing in spite of difficult visibility.

• Automobile collision-avoidance systems. Plastic MMW radar systems used in automobiles could provide drivers with warnings of approaching vehicles. Plastic construction would keep costs low enough to make the systems economically practical.

• Commercial applications. Rotman lens antennas could be used in short-range, building-to-building wireless communication. Implementation in plastic would lower the investment cost of such systems.

• Synthetic vision for assault vehicles. Operators of ground assault vehicles, such as tanks, also need fog and smoke vision aides. Unfortunately, harsh conditions and vehicle vibration limit the practicality of conventional antennas. A Rotman lens could be integrated into the tank's structure, eliminating the need for an external dish and providing necessary durability and reliability.

• Missile seekers. The Rotman lens antenna's low cost, reliability, and small size could be ideal for use in airborne systems such as missile seekers.

A paper on the Rotman lens was submitted to the 1996 Antenna Applications Symposium. An abstract of the research has been submitted to the 1997 National Radar Conference.

Before the antenna can be successfully implemented in any of the application areas, Rausch anticipates expanding the antenna's operating frequency and adding the capability to scan in two dimensions.

For further information, contact the Georgia Institute of Technology, Research Communications Office, 233 Centennial Research Building, Atlanta, Georgia 33032-0828; Phone: (770) 528-7777. ■

PRODUCT INFORMATION

First Application of NT as Automated File Communications Server

Momentum Systems Limited announced a native Microsoft Windows NT™ version of its Intelligent Network Gateway software. The product will be the first application of an NT Enterprise Server managing and automating file communications traffic for an organization and is designed to run on industry standard microprocessor-based servers in a multiprocessor environment.

Momentum's Intelligent Network Gateway software enables an organization to automate and directly manage file transfers with the back-end processing environment, remote customers, service providers, trading partners, and Value Added Networks using multiple communications protocols. Communications support includes BSC, SNA, ASYNC, and TCP/IP FTP. The software supports individual mailboxes for each endpoint as well as inbound and outbound scheduling.

The native NT version of the Intelligent Network Gateway will be available for shipment in the fourth quarter of 1997. An initial release, executing under control of an NT shell, is available now.

For more information about the Windows NT version of the Intelligent Network Gateway, contact corporate headquarters at (800) 279-1384. For additional company and product information visit Momentum's Web site at: <http://www.momsys.com>.

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FACTS, OPINIONS, THEORIES, HYPOTHESES, AND LAWS: Part 1

"My facts ain't your facts"

The basis for reasoned judgments in any field is facts, properly interpreted. Decisions, opinions, and judgements that are not based on facts are inherently flawed and are to be viewed with suspicion and accepted only cautiously.

In the scientific world, this concept is almost an article of religion, while in the political world it is dredged up only when it suits the purpose. Too often, we hear that facts are elastic. A general opinion seems to exist that facts are relative and can be stretched, twisted, and distorted at will to make whatever point the speaker wishes. The phrase "Your facts ain't my facts" characterizes this belief. Some people claim that facts are true or untrue based on one's ethnicity, gender, or some other attribute.

One of the most irksome relativist tricks is to use a concept stolen from physicist Albert Einstein. His "Theory of Relativity" applies to space and time—the stuff that physicists deal with. Einstein never intended it to be used to support moral relativism or anything other than some ideas physicists held about nature. Yet there persists a view that "...Einstein proved that truth is relative." Einstein did not say that truth is relative. If you wish to believe that truth is somehow relative, you will have to dredge up something more concrete than the misquoted theories of Albert Einstein to prove your point.

Exploring the concept of "facts"

The concept of "facts" needs some exploration because it is so central to good thinking. Questions such as "What is a fact?" need to be answered. We also need to know something about the general reliability of facts. Why do so many people, leaving aside those who wish to deceive, become confused enough to relegate facts to the realm of opinion or relativism?

Webster's Dictionary[1] tells us that a fact is "...the quality of being actual; having objective reality; having actual existence." A fact, then, is something that can be shown to exist in reality which, according to Ruchlis[2] means that it is something that can be sensed through sight, sound, smell, touch, taste, or the instruments that extend these senses into domains that are not normally observable by humans. The term "objective" used above means "...existing independent of mind; belonging to the sensible world and being observable or verifiable."

The concept of verifiability of facts is critical to the reasoning process. Making an assertion does not create a fact. I could assert a claim that a white sheet of paper is purple, but that doesn't make it true. One of the chief failings of debates of serious public issues is that mere assertions—usually ideologically based—are assumed to be facts, and no critical analysis is done to determine otherwise. You are not morally entitled to believe that "my facts ain't your facts" as all facts must be verifiable through objective means that are, at least in theory, open to all people to determine for themselves. You may have a private opinion, or a *private interpretation* of what facts (or collections of facts) might mean, but you are not entitled to assert private facts, open only to your personal discernment, and expect others to accept them uncritically.

It is sometimes the case that a thing is believed to be a fact, but is not. When new data becomes available, the perceived fact is modified and a new view is accepted. Such changes do not in any way take away from the concept that facts are rooted in reality and objectiveness, but rather acknowledge that humans can make mistakes, or suffer only partial understanding. The willingness to correct what are believed to be facts, now found wrong, is a matter of *intellectual integrity and is a root* premise of the scientific method, but not of "my facts ain't your facts."

The "my facts ain't your facts" concept is a consequence of a subjectivist point of view. Something that is subjective is "...relating to or determined by the mind; belonging to reality as perceived rather than as independent of mind; knowledge conditioned by personal mental states; lacking in reality or substance."[1] Facts being objective, then, there cannot be "subjective facts." Anything that appears to be a subjective fact either is not a fact at all; or it is untrue under all circumstances, it is true only under specific conditions, or it merely reflects the imperfect state of human knowledge about the particular thing.

Crisp and fuzzy facts

Facts can be either crisp or fuzzy. A crisp fact is one that has concrete boundaries and is characterized by the statement "A is A and cannot be NOT-A." Aristotle's Law of the Excluded Middle and the Law of Contradiction apply to crisp facts. A fuzzy fact is one that is either vague (unclear) or ambiguous (has different meanings).

Consider temperature, for example. To say that the water temperature is 130 degrees Fahrenheit is a crisp fact (never mind measurement error; that's not a consideration here), but to say "the water is hot (or NOT-hot)" is a fuzzy fact. The "hotness" or "NOT-hotness" attribute is a function of circumstance and definition. For example, 130 degree water could scald a bather within a few seconds (it is definitely "hot"), but it would not cook a hard-boiled egg (it is definitely NOT-hot).

The demarcation between fuzzy facts and alleged subjective facts is often not easy to see. A chief difference, however, is that fuzzy facts are contextual but can be defined according to some reasonable scheme. Subjective facts are determined at whim, or according to the beholder who is informed by an emotional value rather than some reasonable objective criterion.

Factual errors

There seems to be several different types of errors made with respect to facts:

Type 1 Error: Acceptance of non-facts as if they are true It is common for people to uncritically accept as factual those statements that are mere opinions, or assertions, or wishful thinking, or are honestly but erroneously drawn. It is for this reason that rational people insist on verification of alleged facts before they are put to serious use in making decisions or forming positions on serious matters.

Type 2 Error: Acceptance of facts that are partially true as if they are fully true. There are many things that are only partly true, or true under certain circumstances, or are true only at certain times. Such a fact should be presented only with the qualifiers on how true it is, or under what circumstances it is true.

Type 3 Error: Acceptance of changing facts as if they were fixed and immutable. Facts can change, and perceptions of what facts are true also change. It is an error to assume that something that was once true, or considered true, will always be true in the future as well. My house is painted white today, but tomorrow I may paint it red.

Type 4 Error: Confusing fuzzy facts and crisp facts. Humans have a tendency to "round off" things, to make them fit into nice clean categories, and as a result often take fuzzy concepts and force-fit them into boxes with hard, crisp boundaries. For example, we can measure a crisp thing like temperature and we can state fuzzy things like "hot" and "very hot." But our tendency is to make crisp definitions such as: Water between 130 degrees and 150 degrees is "hot," but if it is 150 degrees or higher then it is "very hot." The crisp temperature 150 degrees becomes the hard boundary between fuzzy categories "hot" and "very hot." But can anyone really find a reason to assert that 149 degrees and 150 degrees are so significantly different as to require different classes?

The process of verification is the principal means for guarding against errors. In legitimate rational discourse, verification is a public process that is open to all; any claims to private verification delegitimizes the "fact" and makes it useless for the discussion at hand. The public nature of verification acts as a filter that asserted facts must pass through before they are accepted in the debate.

When using or verifying facts, it is often helpful to look at the system of things in which the fact is claimed to fit. It has been said that anything can be made to look factual if enough adjustments are made to the system containing it.[3] If the adjustments in the surrounding universe seem too many or too major before a thing is accepted as fact, then regard it suspiciously: it may be a relativist's attempt to distribute salt water taffy, or it may be an honest but erroneous means for making wishful thinking come true. It is none the less devastating to rational processes.

Opinion

Opinions are often put forth as if they were verified facts. In the United States, our tradition of personal freedom of speech, enshrined in law by the First Amendment to the United States Constitution, sometimes causes us to confuse fact and opinion.

The word "opinion" can be defined several ways, but the two elements relevant to our present discussion are:

1. A formal expression by an expert of his or her judgment or advice;[1]
2. "A belief or judgment based on incomplete information";[2] or
3. "A belief stronger than impression but less strong than positive knowledge."[1]

The first of these definitions is found in law, medicine, engineering, and other fields of arcane knowledge that is open primarily to experts, and only incidentally to lay people. The physician may examine a patient, consider the results of tests, and then offer an opinion as to the proper course of treatment. The process is inductive and, therefore, suffers the usual Problem of Induction; i.e., the expert cannot state for certain what will happen in the future. The opinion is formed on the basis of past experience with similar sets of facts, and a reasonable inference of the future course of events is made.

Expert opinion is often taken for fact, even though it is often wrong. We can be reasonably certain that expert opinion has far more validity than a lay person's opinion, but that does not make it absolute. Expert opinion is often marshaled by both sides of an argument. Assuming that the differences are honestly held, it becomes apparent that many situations dealt with by experts are too complex or too ambiguous to be certain. The proper attitude towards expert opinion is skeptical, tentative acceptance.

More generally, an opinion is a judgment or position taken when information is incomplete. We may make a tentative judgment based on what evidence is at hand, but keep in mind that it is not knowledge; it is, rather, an opinion and may change.

The proper way to handle opinions is with an open mind that is constantly searching for additional facts to confirm or refute the opinion. Opinions must be subjected to the same scrutiny, debate, and evaluation as any other assertion or position. It is unfortunate, however, that people tend to prefer the former to the latter; i.e., they tend to seek evidence that confirms an opinion and ignore (or not seek) evidence that refutes it. When rational people are confronted with evidence that works against a fondly held opinion, they act to correct their position; that untenable opinion must be modified or abandoned no matter how uncomfortable it makes them feel.

The term "opinionated" does not refer simply to a person who holds opinions—which is everyone functioning above the vegetative state—but rather someone who either irrationally holds to a position that has been shown to be false, or who treats opinions as if they were hard, well-proven facts. The opinionated person is typically not open to reason, at least in the area of the opinion held.

Opinions are soft knowledge, tentative in nature, and always subject to amendment as facts become better known. They are useful, especially when facts are difficult to come by, or when the situation is so complex as to defy analysis. While everyone is indeed legally and politically entitled to their own opinion, that does not mean that all opinions are equally valid, or that we must give the same credibility and honor to all opinions. Your right to hold a silly, unsupported opinion is modulated by my right to laugh at it.

Scientific laws and theories

Collecting facts can be quite useless to the rational enterprise because facts in isolation can be totally meaningless. But if facts are arrayed in some manner that reveals an underlying pattern or order, then advances in knowledge can be made. That's where scientific laws and theories come into play.

Scientific laws

The word *law* is used quite a bit in talking about science, but is it often misunderstood. Many people, when they hear the phrase "scientific law," immediately think of something that is both universal and immutable. But, in science, a law is merely "[an] observed regularity [in nature]...or "[a] statement of order or [the] relation of phenomena that *so far as is known is invariable under the given conditions*." Thus, a scientific law is neither universal nor immutable, but rather is something that is always observed to happen the same way—at least that is the way it was in the past.

The use of the word "law" does not in any way exclude the possibility that someday, under some circumstances, we might find an exceptional observation that forces us to change the law. It sometimes happens that refinements in measuring techniques, improved experimental apparatus, better experimental design, or new observations that were previously unsuspected lead us to reconsider the concepts underlying some specific scientific law. Like the legislative kind, scientific laws are subject to amendment. A good example is Newton's laws.

Any scientific law is merely a *tentative* statement, or intellectual construct, of observed invariability of certain phenomena or their interrelationships. In the course of a single scientific career, you routinely expect to see laws modified or cast aside altogether. I suspect that a Nobel Prize or two is lost from time to time by people looking for the expected, and thereby overlooking or discarding anomalies that don't fit the preconceived mold: Was a peculiar data point "experimental error" or was it Nature whispering a hint of a secret in your ear?

The most common form of law is the empirical law. It is structured to integrate a large number of observations into a single coherent general statement. The general statement will set forth interrelationships, observed patterns, and provide a means for predicting future events of the same type. Casti[3] provides four properties of a good scientific law:

1. It generalizes to classes of events, rather than individual events;
2. It demonstrates a functional relationship between two or more classes or kinds of events;
3. It is supported by a large amount of experimental data, with no disconfirming evidence; and
4. It is applicable to different events; i.e., it is not unique to an individual event or some narrow class of events.

In Part 2, we will look at the word theory and the different ways in which it can be interpreted. Specifically, we will look at what theory means to laypersons versus scientists. ■

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1. *Webster's New Collegiate Dictionary*, Springfield, Massachusetts, G.&C. Merriam Company, 1979.
2. Hy Ruchlis, *Clear Thinking: A Practical Introduction*, New York, Prometheus Books, 1990.
3. John L. Casti, *Paradigms Lost: Images of Man in the Mirror of Science*, New York, William Morrow & Co., Inc., 1989.

PRODUCT INFORMATION

Technical Product Information and Immediate Support

Pentek has announced the expansion of its Website. This expansion includes a variety of technical documents for engineers, an exclusive technical support database called the Knowledgebase, plus detailed product data sheets with specifications and block diagrams, along with access to full product manuals. The Knowledgebase offers answers to frequently asked questions, products, tips and techniques, product compatibility, and other useful information not found in product documentation.

The Knowledgebase is available free of charge to all Pentek customers and is online now at <http://www.pentek.com>.

HP Has Software for EMI Testing and Reporting

Hewlett-Packard Company has introduced software designed to speed the process of testing to meet commercial radiated electromagnetic interference (EMI) regulations. The new software provides an array of features to help electromagnetic capability (EMC) engineers and technicians test radiated emissions from commercial products.

From device prescans to the processing of final emission test results, the HP 85876B software automatically performs a series of predefined measurements, data comparisons, file transfers, and reporting sequences. A report-generation interface allows the user to define custom compliance reports for submittal to regulatory agencies or for archives.

The software runs on Microsoft® Windows® 3.1, Windows 95, and Windows NT® 4.0 environment and requires no additional programming or scripting to operate. Reports can be saved as Rich Text Format (.rtf) files, and graphics of equipment setups and receiver traces can be saved as Windows Metafile (.wmf) files for export to popular word-processing applications. HP EMI receivers supported by the new software include the HP 8542E, 85422E, 8546A, 85462A, 8571/2A, and 8573/4B.

For information, contact Hewlett-Packard Company, Test and Measurement Organization, P.O. Box 50637, Palo Alto, California 94303-9512. You may also visit HP's Web site at: <http://www.hp.com/go/tmnews>.

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YAGI/UDA ANTENNA DESIGN

Part 1: A different approach

The Yagi/Uda antenna, commonly referred to as the Yagi, has been a proven performer for over 70 years. For much of that time, this type of antenna design required builders use the "cut-and-try" method. Often front-to-back (F/B) ratio was the sole parameter measurable by amateurs because of the difficulty in gauging small differences (1 dB or less) in forward gain. Impedance matching was left to the imagination of the builder, and the Gamma match was usually favored because of its simplicity and ability to be tuned for a 1:1 VSWR.

The design approach mentioned above

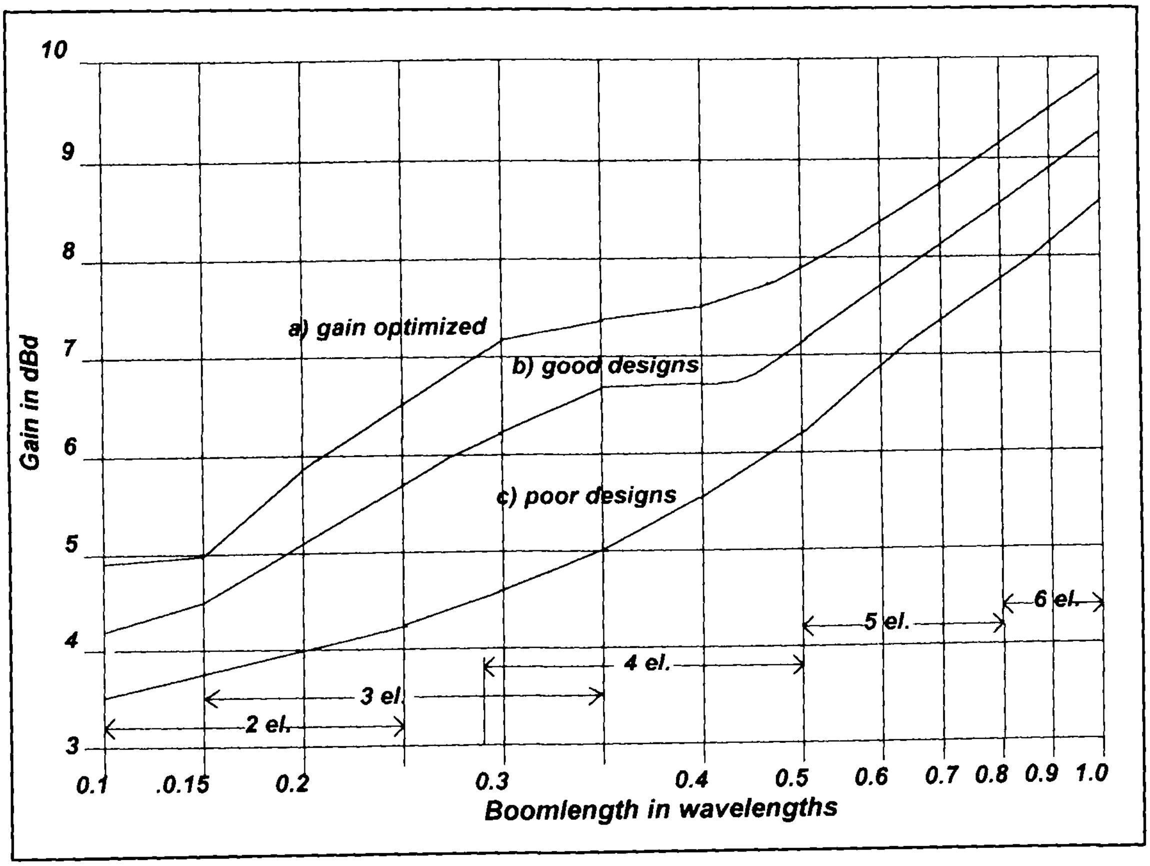

Figure 1. Actual gain attainable and recommended number of elements versus the boomlength for a Yagi antenna. The upper curve is the maximum gain attainable if other parameters are sacrificed; the middle curve shows an optimized Yagi; the lower one represents a poor performance design.

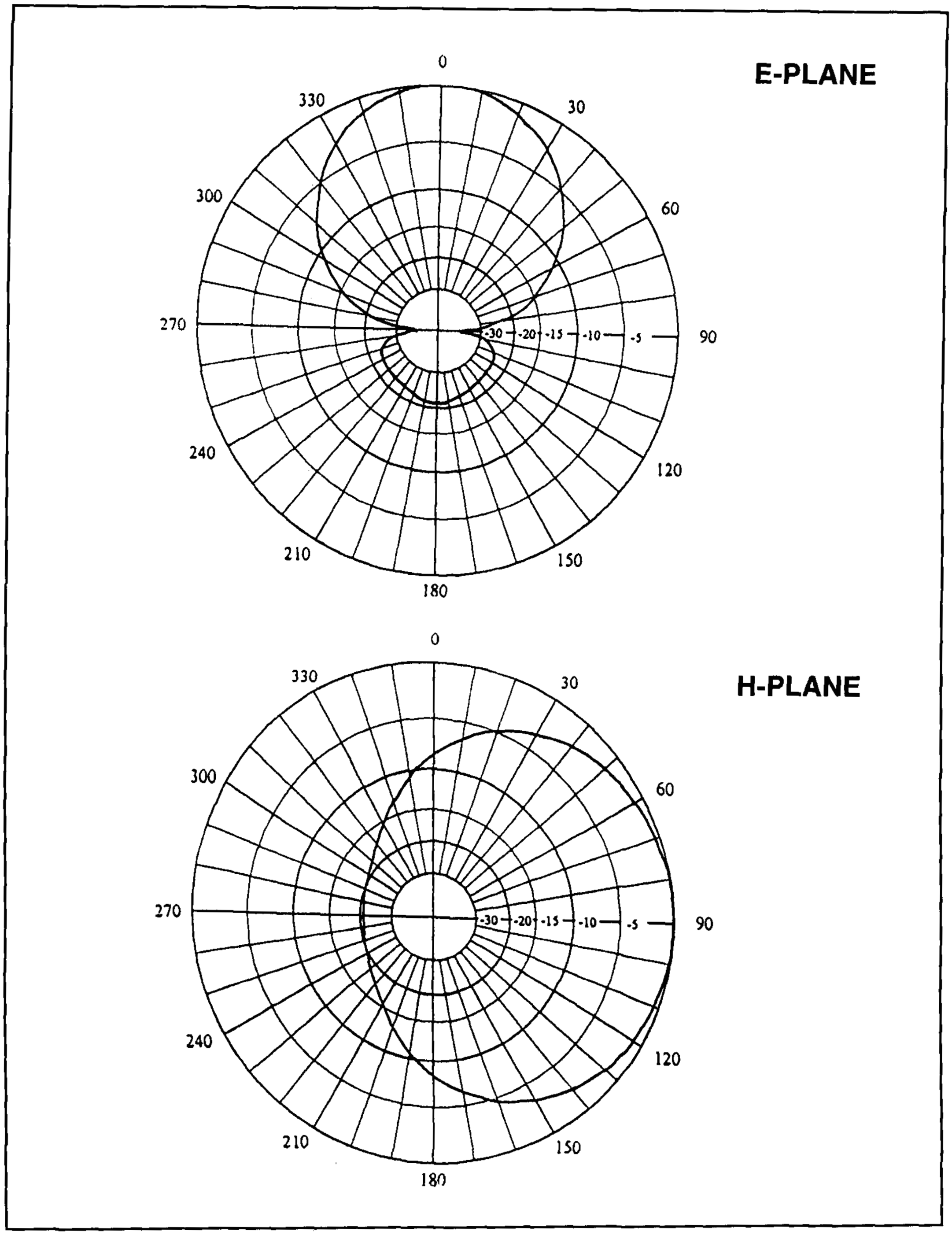

Figure 2. Typical radiation patterns for the simple three-element Yagi described in the text.

worked well because the basic Yagi antenna is very tolerant to wide variations in spacing and element lengths. You might say that a Yagi antenna just "wants to work." Cut-and-try tuning for the best F/B ratio often yields a reasonable forward gain because the two parameters are somewhat interrelated in the overall scheme of things.

Nowadays, most Yagi designs are modeled and optimized on a computer, which has led to a big step forward in performance. In this article, I will discuss computer-optimized Yagis, make suggestions for improvements, and provide an alternative design approach that has worked quite well for me.

Enter computer modeling for antennas

In 1964, I. Larry Morris, while working at Harvard University on his doctoral thesis,[1] produced what I believe was the first useful software program to model Yagi antennas. The program was written in FORTRAN and required a large mainframe computer like the IBM 7094. Morris discovered how to compute gain, pattern, and the impedance of a Yagi.

Morris' program was probably the basis for the early work done by the late Dr. James Lawson, W2PV. His classic set of articles was published in the early 1980s in *Ham Radio* magazine and later summarized in his book, *Yagi Antenna Design*.[2] John Kenney, W1RR, later obtained a copy of the Morris program and modified it to optimize HF Yagi dimensions.[3]

Since the mid-1980s, modeling programs such as NEC (Numerical Electromagnetics Code) and MININEC (a smaller program based on NEC) have been available.[4] Both programs have now been optimized for the personal computer (PC) market and incorporated within many software programs that are available to the general public at reasonable prices. The beauty of these modeling programs is that we no longer have to get dirty hands and aching backs cutting aluminum and testing a new Yagi design in the field until we have an acceptable working antenna. We can now "cut-and-try" a Yagi antenna design for maximum gain, F/B ratio, impedance, or a combination of the three by changing element lengths, diameters, and spacings using a PC keyboard.

For the more sophisticated designer, software programs are now available which, with a properly weighted set of input parameters, will perform the optimization for you. All that's required is that you input a reasonable Yagi design, properly weigh the tradeoffs or Figure of Merit (FOM), the relative importance of each antenna optimizing parameter such as gain, pattern, and impedance and, voila, out comes a design. Now all you have to do is properly interpret the computer results and build the antenna of your dreams!

Pitfall #1: modeling

The new field of modeling antennas on a PC is quite exciting, but the picture I've presented

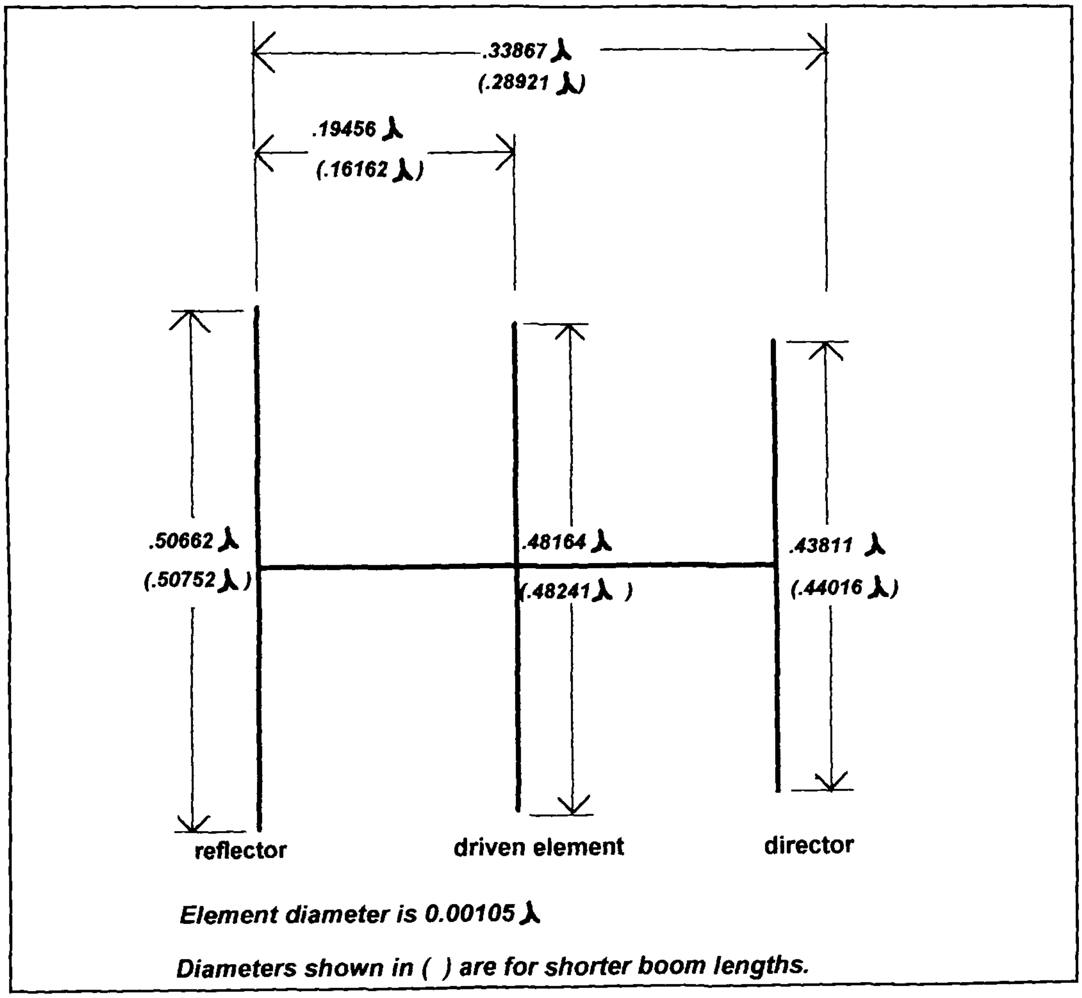

Figure 3. Lengths and spacing of a typical 50-ohm three-element Yagi, as mentioned in the text. Values shown in parentheses are for the shorter boom version. The diameter of all elements is 0.00105 wavelength (0.875 inch at 14.175 MHz).

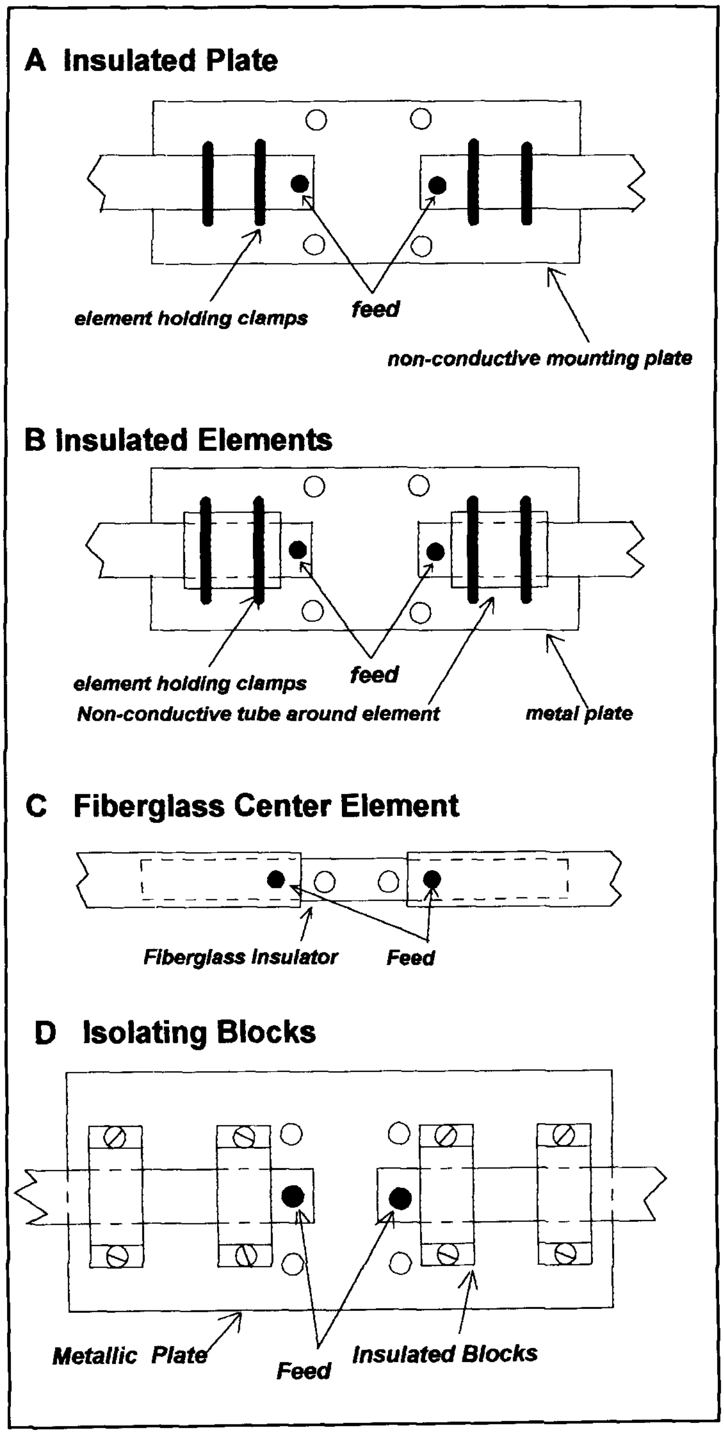

Figure 4. Some recommended center mounting configurations for the driven element on a directly fed Yagi. In all cases, the driven element must be insulated from the boom on both sides.

is an oversimplification. There are numerous pitfalls to avoid that have prevented many builders from attaining the expected results. I'll first examine some of these pitfalls, while offering some insight and suggestions. Then I'll present a different approach to Yagi design that hopefully will eliminate, or at least diminish, some of the present problems.

Modeling and antenna optimization software is a great tool because the human mind is still the force behind the keyboard and is ultimately the judge and builder. If you don't like the results, just make a little tweak here or there and see if the computer can produce an improved design.

Is the computed design real? The accuracy of the programs available to amateurs, primarily NEC- or MININEC-based, is still a topic of discussion.[5,6] The diameter of the wires (or elements) and the number of segments used for each wire are still being debated. The newer modeling programs and higher speed PCs are allowing more flexibility and speed, but there are limitations.

Another major problem with the above mentioned software is that it's human nature to try to pry every last bit of gain and pattern improvement out of an antenna, especially because the results on most computer modeling programs are typically displayed to 1/100 (0.01) of a dB—a mighty small amount of improvement! Also, don't forget to consider all the hours spent tweaking on the computer or the mechanical tolerances that must be considered. Furthermore, due to the methodology of currently available optimizer programs, usually a gradient-based optimizer, the best design possible may be missed or "jumped over" since the computer uses knowledge of the previous design in deciding on how to quickly zero in on a better or final design.

What we don't want to do is skip over more optimum designs. This is why the new technique of genetic algorithms, or GA is being explored by antenna designers as possibly the next frontier in Yagi design.[7] Genetic algorithm programs are not gradient based and acquire a greater population of design candidates.

In fact, the "ultimate" Yagi design may not be the most efficiently performing antenna in the "real" world. Why is this? Often the computer-optimized designs yield a very narrowband design and/or one that has a very low input impedance. There are ways to circumvent narrowband designs by optimizing at several different frequencies, but this isn't foolproof either. Narrowband designs are fine if you only need to operate over a very small portion of the spectrum, but they have drawbacks. More on this shortly.

An example of a high-performance, albeit narrowband Yagi design, is the famous PV4, a HF four-element Yagi on a 0.5735-wavelength boom.[8] It has great gain and F/B ratio over a narrow bandwidth, but the impedance match and F/B ratio drop very rapidly 1 to 1.5 percent above and below the optimum design frequency.

Front-to-back ratio is another area where problems occur. Always remember that F/B ratio is the difference in dBs between the front of the antenna (usually 0 degrees) and the point exactly 180 degrees from the front. I've seen antennas with a 30- to 40-dB null exactly at 180 degrees, somewhat analogous to the classic cardiod pattern, but at 160 and 200 degrees the signal was only down 20 dB!

John Kenney, W1RR, suggests a better performance criteria: the front-to-rear (F/R) ratio.[3] The F/R is the difference between the front of the antenna and the highest sidelobe (in the rear hemisphere between 90 and 270 degrees).

Always remember that the modeling programs have their limitations. When designing a Yagi, try to balance the different parameters. Then change the number of segments. I prefer eight to 10 segments per half-wave element, but others like up to 30. Watch the F/R versus the F/B ratio.

Also don't try to tweak the last bit of gain out of a Yagi at the expense of decreasing bandwidth. Occasionally re-tweak an element length, a spacing, and/or the FOM weights and let the optimizer hunt for a possibly better design. Finally, if you have the resources, use more than one modeling program and compare the results.

Pitfall #2: element tapering and scaling

In the early days of HF Yagis, elements were often made from similar diameters of plumber's pipe and the designs were dubbed "plumber's delights." Aluminum tubing with different diameters wasn't readily available, so many HF Yagis were built using only one or two diameters of tubing. Simple formulas were established and used as if they were gospel.[9] Then the so-called "New England" taper became popular as standard aluminum tubing became available in the U.S. in 1/8-inch (0.125 inch) diameter increments with 0.058-inch wall thickness. This made telescoping a bit easier.

However, the earlier designs seldom considered that tapering different diameters might affect the overall "electrical length of the elements." Hence the performance, whether observed or not, was often compromised. I believe this had a great influence on why many Yagi users switched to quad antennas in the early 1960s, when tapered Yagi designs often fell short of expectations.

In 1961, I visited the home of the late Roger Mace, W6RW. Roger had a full-size, three-element 40-meter Yagi perched in the Hollywood Hills overlooking the city of Los Angeles. Roger's potent 40-meter DX signal was legendary. A full-size 40-meter Yagi is quite large; elements in the vicinity of 70 feet (21.3 meters) long are required. For his elements, Roger used many different diameters of tubing, as in the New England taper. By using the cut-and-try method, he knew he had a pretty good design, but the elements were over three feet longer than the standard formulas used by everyone else. He pointed this out to me and said that he was dumbfounded. I had no answer then, but often pondered the question.

It wasn't until 1980 that Lawson noted that when elements are tapered, a length-correction factor must be applied.[2] A program for tapering is shown in **Reference 8**. However, Lawson's approach, while a step forward, was an algorithm based on empirical data to satisfy the antenna modeling programs of the day. It is no longer considered accurate enough for modern high-performance Yagis.

Dr. Dave Leeson, W6QHS, studied this problem and, with a modification of the work of Schelkunoff,[10] derived a more accurate element scaling algorithm.[11] This isn't a problem with modern modeling programs because you can tweak the element lengths to restore a design to the expected performance.

Scaling a good design from one frequency to another is also a problem. If you obey the laws of scaling, all dimensions, including element diameters and spacing, must follow the scaling factor. For example, a 14-MHz Yagi scaled to 28 MHz should have element diameters exactly one half the original diameter. However, this is not always practical as the tubing sizes may not be the ones desired or available in your stock supply. The procedures stated above under element tapering still apply. Furthermore, the diameter of the element also affects the ultimate bandwidth because fatter diameter elements inherently yield wider bandwidths than thinner elements.

Tom Ring, WA2PHW, and Brian Beezley, K6STI, came up with an even better scaling and tapering solution. *Remember that when you scale or taper an element, you are essentially trying to come up with an element that has the same electrical impedance at the new frequency.* What Ring and Beezley did was to adjust the element lengths after scaling and tapering for the same electrical impedance using the YO (Yagi Optimizer) program.[12]

Suffice it to say that scaling and tapering of elements does introduce a possible area for inaccuracies. Furthermore, there's still a controversy over which modeling program is the most accurate, especially with tapered element diameters. While NEC-based programs in general are considered the most accurate, they don't work well with multiple diameter tapered elements. Likewise, MININEC-based programs handle tapered elements well, but are considered less accurate overall.

Pitfall #3: other element correction factors

Although I seldom hear it mentioned, the physical configuration at the end of the element is important. More specifically, the electrical length of a tubing element is different from that of a rod element, both having the same diameter and length, especially at VHF and UHF.

At VHF/UHF the end effects or fringing is altered further by the shape of the tip of the element. The electrical length of a tube is different from that of a rod. Furthermore, a rounded or hemispherically shaped tip has a different electrical length than a shear-cut rod element. This was pointed out by Mailloux.[13]

Mailloux recommends adding slightly less than one radius of the element diameter of tubular and shear-cut tipped elements to compensate for the end effect. Steve Powlishen, K1FO, also noticed this effect and recommends adding a 0.031-inch (0.8 mm) 45-degree chamfer to the ends of the rod elements.[14]

Furthermore, the caps often used on the tips of Yagi elements add an "end effect." I first noticed this in 1979 with the black end caps—the type most often used in the antenna industry. More recently, with the help of a Hewlett Packard 8753C Network Analyzer, I was able to verify the effect more accurately at 50 and 72 MHz.

When end caps are used, the end of each element should be "shortened" by the thickness of the end cap material, typically 0.062 to 0.125 inch (1.5 to 3 mm). This is more pronounced on 6 and 2 meters (and above) than on HF. In addition, if an end cap later breaks or falls off during use, a typical result of a year or so of antenna weathering, the VSWR and the Yagi performance is altered.

Weather-related effects must also be considered. Rain, water on the elements, wind, ice buildup, and even breakage of an element must be taken into account. Generally speaking, larger diameter elements aren't detuned as much as smaller diameter elements when ice and water are present, but larger diameter elements do have greater wind load. Also, if an element breaks or a tip falls off, wider bandwidth Yagi designs will suffer less degradation than narrowband designs.

Rain and water on Yagi elements, especially at VHF and above, will detune the antenna. Many years ago, I experienced increased VSWR on several 2-meter EME schedules during rainstorms. Every 10 to 15 minutes, I would go outside and vigorously shake the Yagi array to remove the raindrops and restore performance! Afterwards, I discovered that some manufacturers leave a thin film of oil on their elements which, for the first year or so of use, causes rain droplets to form and cling to the elements rather than dispersing. When I figured out the cause of the problem, I thoroughly rubbed down the element surface with a solvent (acetone) and the problem diminished. Likewise, chamfering the ends of the elements, as I mentioned above, will cut down on this detuning effect.

Therefore, even with a sophisticated modeling programs, and so on, we still experience factors that may be out of our control. This is exacerbated with higher frequency and narrower band designs. Hopefully, some of the recommendations mentioned above will be useful.

Pitfall #4: element-to-boom correction factors

Along with element tapering and scaling, the effects of the element attachment to a Yagi boom must be considered. In many HF Yagis, the directors and reflectors are attached to a metallic boom with a flat plate, a shaped bracket, or a "U" bolt. Therefore, depending on the element mounting method and its size, the diameter of the element, the boom diameter, and the height of the element above (or through) the boom, the electrical length of the element is shortened.

The modeling programs available today don't consider element-to-boom corrections. At best, some suggest approximations that are made based on an equivalent diameter for the portion of an element mounted to a plate or a bracket. Element-to-boom correction factors were discussed and some equations were presented in Lawson's book.[15] Leeson pointed out some errors in Lawson's assumptions and derived a new set of more accurate equations for element correction factors.[16]

Suffice it to say that the element mounting method affects element tuning: the smaller the clamp or plate, and the further the element is mounted from the boom, the less the affect on the electrical length of the element. Some high-performance antenna designers have even resorted to using insulated plates.

Element correction and other VHF/UHF Yagis facts and fallacies on this subject were discussed in **Reference 17**. Many VHF/UHF Yagis use through-the-boom element mounting. They may use either ohmic or insulated bushings. It was pointed out in **Reference 17** that insulated through-the-boom elements only require about half the correction of ohmic contact methods.

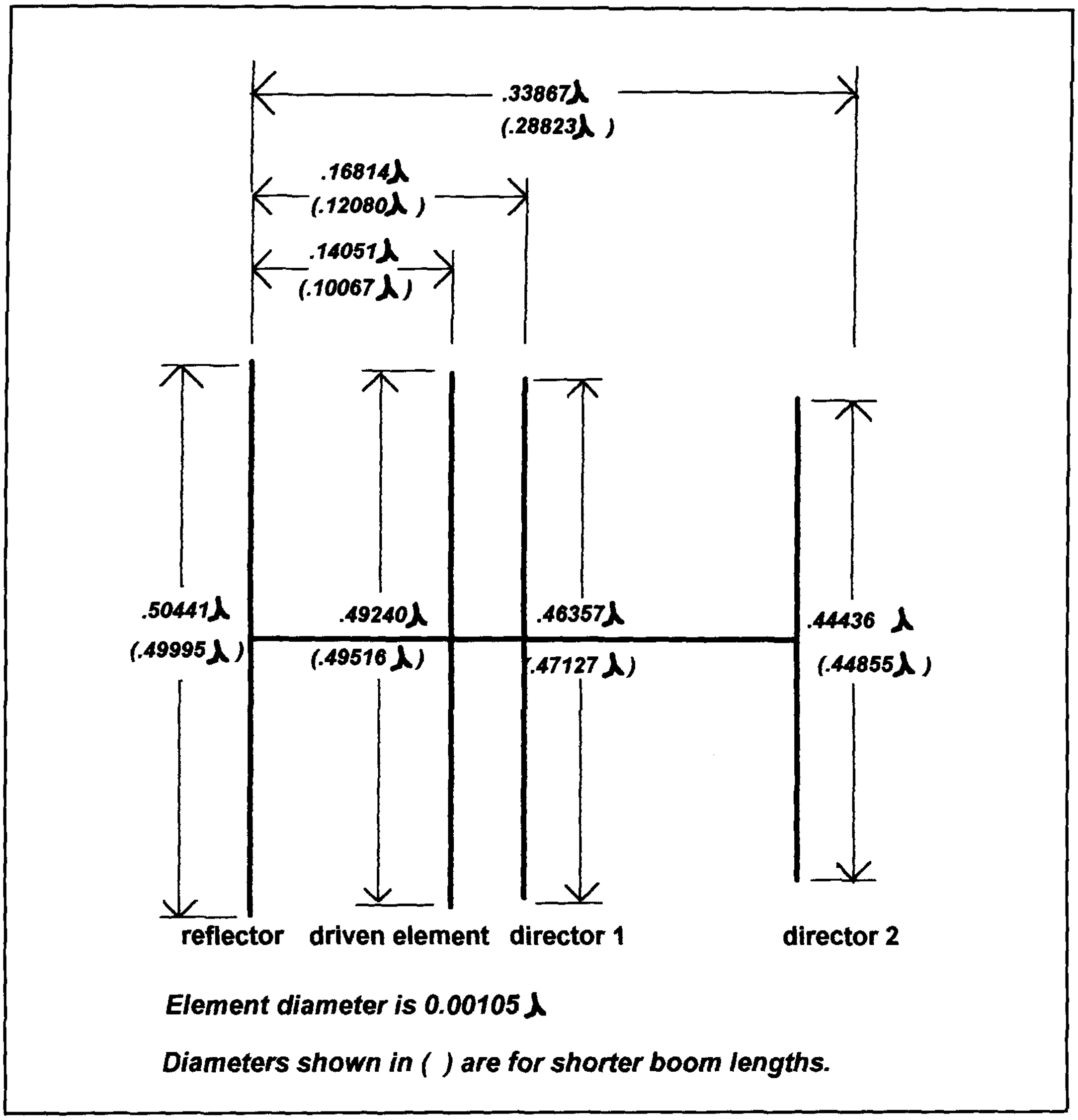

Figure 5. Lengths and spacings for a 50-ohm four-element improved Yagi. Values shown in parentheses are for the shorter boom version. The diameter of all elements is 0.00105 wavelength (0.875 inch at 14.175 MHz).

Shortening the element usually has less affect on Yagi performance than does lengthening an element because performance on most Yagis decreases more quickly on the high side of the design frequency. Finally, remember that the narrower the bandwidth of the Yagi, the greater the possibility of performance degradations due to corrections needed for the element-mounting configuration.

Pitfall #5: impedance matching

Impedance matching a Yagi antenna is probably the biggest headache for antenna designers and builders. *I believe that one of the reasons why many amateurs purchase Yagi antennas rather than build them is that they don't want to deal with all the hassles of impedance matching*. Only hearty souls, such as dyed-in-the-wool DXers and contesters who want specific performance, are willing to put in the effort required to build their own Yagis.

Over the years, many different types of impedance matching networks have been used. In the early days, balanced feeds such as the Delta or "Y," "T," and folded dipole were popular.[9,18,19] However, to be effective, balanced feeds also require a balun, usually the 4:1 half-wave type, thus adding weight, loss, and wind load to the antenna.

Next came the Gamma match. It was popularized by Katashi Nose, KH6IJ.[20] However, even he said: "The conclusion is that this is easier said than done [adjusting the Gamma match until the proper match is obtained] and that some authors, including myself, have been prone to underestimate the work involved."

This led to the Omega match, which shortens the Gamma rod but adds another capacitor to assist in the matching of the antenna.

Later, Nose studied the unbalanced geometry of the Gamma match and concluded that the driven element should be asymmetrically mounted by making each side of the driven element a different length.[21] Despite his work, I have never seen anyone use an asymmetrically mounted driven element in a Yagi antenna. Even if such a system were used, some would think the unbalance was unacceptable mechanically and aesthetically!

Gamma matches also depend on a ground return via the shield on the coax. Fortunately, or unfortunately, ground is usually many feet below the antenna! Hence the ground return of the impedance-matching network found on a boom is not there, and unbalanced currents often flow on the boom as well as on the outside of the coax shield.

Another problem with the Gamma match is that it is difficult to waterproof. Many protection schemes, such as a coaxial capacitor,[22] have been proposed; but, sooner or later, water seems to get inside the tube and destroy the match and VSWR.

In 1962, the "Hairpin" match, sometimes dubbed the Beta match, was proposed.[23] It's a good way to feed the antenna in a balanced manner. However, it is sometimes difficult to break the driven element in the center and tuning of the hairpin is still required. Furthermore, to be successful, this type of match requires a 1:1 balun, which was not in common use in the early 1960s. Furthermore, the optimum time to match an antenna is when it's permanently mounted, often high on a tower. If the matching network isn't accessible from the mast (as it seldom is!), it may be very difficult to obtain a good match. I have vivid memories from my younger years of tuning Omega-matched Yagis with pulleys and strings while perched 50 feet in the air!

Note that the impedance-matching networks and traps presently being used on Yagis often limit the ultimate operating bandwidth of an antenna design. It's not uncommon to see an antenna with a bandwidth of only 50 to 60 percent of what was expected from the computer-modeled bandwidth.

Many of the new computer-optimized Yagis have very low drive-element impedances, typically 12 to 20 ohms, which are often very reactive. From measurements I've conducted on VHF, these low-impedance Yagi designs usually have narrower bandwidth and often are less efficient in terms of actual measured gain (on an antenna range) compared to Yagis with driven-element impedances of 20 ohms and higher. This is probably the result of increased current in the driven and first director elements and matching network, etc. Therefore, I recommend higher-impedance, wider-bandwidth Yagis because the lower the driven-element impedance, the greater the difficulty in obtaining an impedance match. Also, the impedance-matching techniques presently in use aren't necessarily the easiest to use or optimize.

Pitfall #6: gain and number of elements in a Yagi antenna

Still another misunderstanding of the Yagi antenna is its amount of gain. There are all kinds of myths floating around, but the principle one is that the gain is a function of the number of elements employed.

This is only partially correct. *The gain of a Yagi is primarily a function of the length of the boom* (in wavelengths). The number of elements and their length and location on the boom is a secondary function that mainly affects bandwidth, impedance, and pattern of the Yagi antenna. This subject was thoroughly documented in **Reference 24**.

But how is this true? The answer is that each *Yagi design needs a "sufficient number of elements" to support the boomlength and parameters required.* While I'm in favor of using the fewest number of elements to create a design, there are some problems in this approach. *Adding an extra element to a Yagi will sometimes improve bandwidth and pattern and make it less critical to design and build.*

Table 1 may be used as a Yagi design guide and shows the recommended number of elements for various boomlengths (in wavelengths). **Figure 1** is a graph showing gain that can be attained in a short (up to 1 wavelength) Yagi antenna. Note that there's a slight plateau between 0.35 and 0.45 wavelength booms. The optimum designs are usually those that are close to the middle curve; poorer designs are often near the lower curve. Higher gain near the upper curve is sometimes possible, but pattern and bandwidth are usually sacrificed. *I recommend that you aim for the middle curve on the graph, which will usually yield the best compromise between gain, pattern, and bandwidth.*

How to avoid some of the pitfalls

By now most of you are probably asking (if you had the patience to stick it out so far), where is this heading? In a nutshell, I'm proposing Yagi designs with higher-impedance driven elements typically 50 ohm with low reactance and direct-balanced center feed.

I'm also in favor of broadband designs.

Furthermore, I want to show you how to take much of the hassle out of impedance matching and designing Yagi antennas. Then I'll describe a higher performance Yagi using these techniques. I hope to generate interest in these techniques so there will be a greater effort to improve the state of the art and to build homebrew antennas.

As I pointed out earlier, many of the new computer-optimized Yagis have low driven-element impedance. There's no doubt that these antennas have a very clean pattern on the computer model, but can that performance be obtained in real life? I think not. Many of these designs only have the peak performance over a narrow bandwidth. In addition, narrowband Yagis require more attention to the element correction factors and the impedance match. Then there's the problem of proximity to other antennas—especially if used in a Christmas tree stack. Narrowband Yagis are very easily detuned by surrounding antenna.

The truth of the matter is that *you can design and build a good 50-ohm computer-optimized Yagi with broadband characteristics*, albeit with slightly less gain and F/B ratio, that doesn't require impedance-matching networks. The design isn't new; in fact, many of the older triband trap Yagis fall into this category.

Typical Gain Vs. Boomlength

Boomlength (in wavelength)	**Gain Range (dBd)**	**Number of elements**
0.1–0.25	3.5–4.5	2
0.15–0.35	5–6.5	3
0.3–0.5	6–7.5	4
0.5–0.8	7–8.5	5
0.8–1.0	8.5–9	6

Table 1. The typical gain versus boomlength as well as the recommended number of elements well-designed short (<1.0) Yagi antenna.

A simple Yagi to build

I've been using several different three-element Yagi antennas for some time now. They typically have a boomlength between 0.25 and 0.35 wavelength, a gain of about 5 dBd (dB over a dipole), and F/B ratios of 20 to 25 dB. No impedance matching is required because they are already 50 ohms and can be fed directly with a 1:1 balun. A typical radiation pattern is shown in **Figure 2**, and **Figure 3** shows typical lengths of such a Yagi in wavelengths. I have normalized the element diameter to 0.000105 wavelength to be near the values in Lawson.[2] I presume you all have calculators, and probably some type of antenna modeling software, so you can scale the antenna to any desired center frequency.

The performance of this antenna is reasonably good, and, unless you have rather sophisticated measuring equipment, I doubt that you could see the difference of 1 dB less gain than that of a highly optimized antenna. The bandwidth is very large, typically 5 percent for a VSWR of 1.5:1 maximum. *I make no claim that this is an optimized design because everyone has there own parameter priorities.* Furthermore, there's no need to use a 0.34 wavelength boom. A shorter boomlength (0.29 wavelength) also works well with only slightly less gain. I have used this shorter boom on 6 meters because it only requires a 68-inch (1.73 meter) boom. The element lengths for this antenna are shown in parentheses in **Figure 3**.

These Yagis use a balanced feed in the center, so some provisions must be made to isolate each side of the driven element. I have provided a few examples in **Figure 4**. Provisions should be made to tune the overall length of the driven element so any stray tuning effects will be tuned out later. This can easily be done with a smaller diameter tube at the end, held in place with a worm clamp.

The only other requirement is a simple 1:1 current balun. I recommend the ferrite type. You can use either the ferrite toroid model, the first real simple current balun I suggested in 1978,[25] or the ferrite bead type recommended by Walter Maxwell, W2DU.[26] If you use the W2DU type, all that's required is 25 to 100 short ferrite beads over an RG-58 (for low power), or for high power, RG-303 Teflon™ coax. Several manufacturers sell this type of current balun for HF and VHF.

Matching is a snap. All you have to do is adjust the overall length of the driven element for a good impedance match. Remember that adjustments to the length of a driven element of a Yagi antenna, within reason, have no effect on gain or pattern: they only affect the impedance match. If you followed the dimensions shown in **Figure 3**, a close to 1:1 VSWR should be easily attained. Just make the tips of the elements adjustable. Check the VSWR at your favorite frequency and adjust each side of the driven element the same amount until a perfect match is attained.

Another advantage of this type of Yagi design is that the computed VSWR can be quickly compared to the measured VSWR. If you have one of those new VSWR analyzers from Autek, AEA, MFJ, etc., matching is very

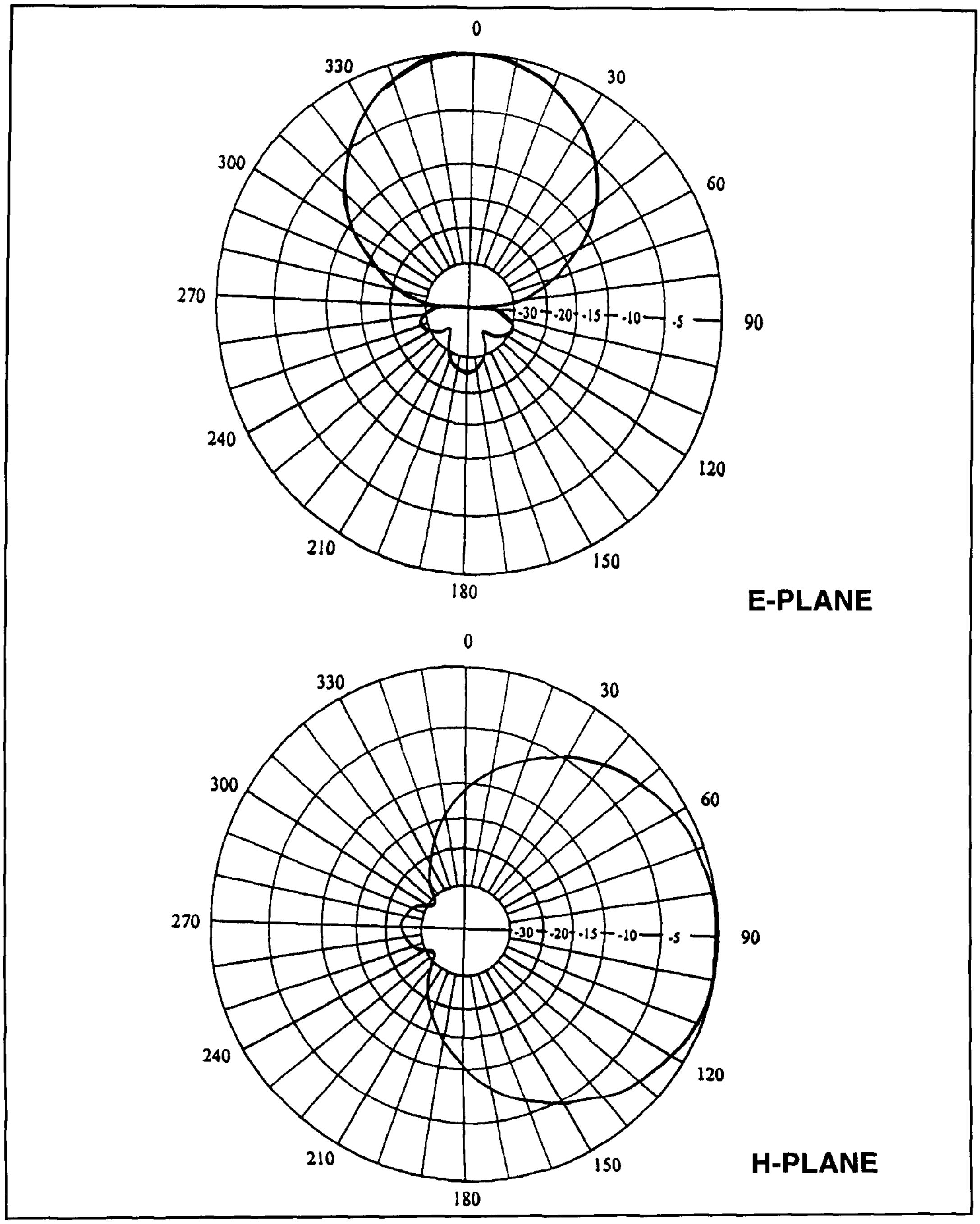

Figure 6. Typical radiation patterns for the simple four-element Yagi described in the text.

simple: just measure the VSWR over the frequency range and compare the results to the computed model. If you match the results very closely, especially at the upper and lower frequency end where VSWR is increasing rapidly, you have done all the right things and all your element corrections are good. If not, you may need a small correction.

Finally, I do not represent these three-element Yagis as fully optimized designs. You may want to spend the time (I don't) to optimize them further on a computer and possibly pick up additional performance while maintaining the 50-ohm direct feed. Save cutting aluminum until you get the performance you need!

A higher performance Yagi

Some of you may be wondering if the above approach is all I have to offer. Actually, there is an even better method that I've been using for many years, and I have finally put together several working models. *This approach improves the gain and the bandwidth at the expense of one more element on the same boomlength.* Over the last few years, I have built more than 200 different Yagi models with four to 31 elements from HF through UHF using this principle.

It all started in 1974 with my quest for a simpler impedance-matching method. I experimented with an extra director between the driven element on an NBS five-element Yagi.[27] At that time, I didn't have a way to optimize the results. I later discussed this technique with Stan Jaffin, WB3BGU, who had just finished an antenna modeling program. He found it quite useful to improve existing Yagis and wrote about it in **Reference 28**.

However, I still wanted this design to be more universal, and then the idea of making it part of a 50-ohm impedance match came to mind. My first attempt was a 20-meter Yagi design. I did the preliminary computer modeling, but the results were mixed depending on which modeling program was used. I never had the time to make a full-scale model.

A few years ago, I needed a Yagi with more gain but didn't have the liberty of a longer boom. I set about to optimize this design and it came out quite well. To my surprise, I was able to obtain a typical gain of 6 dBd with F/B ratios in the mid to upper 20s near design frequency. *A gain improvement of almost 1.0 dB on the same boom is impressive, and with a 50-ohm match to boot!* Since then, I've built working Yagis at 50, 72, and 144 MHz. All have matched the computed results very closely in VSWR, F/B ratio, and forward gain, all verified on an antenna range with a Hewlett Packard 8753C network analyzer.

A layout of the four-element Yagi antenna, scaled as above, is shown in **Figure 5**. As you can see, it has the same boomlength, 0.34 wavelengths, and it too can be shortened to 0.29 wavelength (values in parentheses) with very little performance degradation. You will also notice that the antenna is mechanically well balanced—not "tail heavy" like many of the newer optimized HF designs. In fact, the feedpoint is near the boom so adjustments are easy to perform.

Figure 6 is a typical radiation pattern for the antenna. You'll notice that it's clean and the gain is almost as high (close to 6 dBd) as some of the new computer-optimized, lower-impedance Yagis. The bandwidth is very good, typically 6 to 9 percent for a 1.5:1 VSWR. The mechanical construction details mentioned above under the three-element Yagi still apply.

Again, this four-element Yagi design isn't fully optimized, but the model shown is real and can be used as the basis for a more fully optimized design. In fact, with only 100 watts in the shack, I have easily worked African, European, and Caribbean stations on 6-meters during the summer of 1997 using the four-element 0.29 wavelength Yagi design shown in **Figure 5**.

Summary

I've discussed some of the major pitfalls in Yagi design, especially those applying to computer modeled and computer-optimized designs. I also made some suggestions to improve results and to avoid errors.

Then, I provided some tips on impedance matching. The suggestions given should, if nothing else, improve the results when designing and building the new generation of computer-optimized Yagis. The many references cited should help those who need more information to fill in the gaps.

Next, I proposed a three-element antenna that's easy to match and has good performance. Although, it may not be as high-performance as the best models (about 1 dB low in gain), it's nevertheless simple and straightforward. You may even find ways to improve its performance while keeping the simple 50-ohm direct feed.

Finally, I proposed a new design: a four-element Yagi, that has many of the attributes of the computer-optimized versions. It's easy to match, has high gain and low VSWR over a very wide frequency range, and is the same boomlength as the three-element Yagis. These four-element designs are not fully optimized, so it's possible you may improve them slightly.

I hope this material will cause some Yagi antenna designers to rethink the kind of "go for smoke" approach that's now in vogue. The fact that these proposed designs are so broadband makes them easy to reproduce with less tolerance problems, such as the pitfalls mentioned earlier. With the 50-ohm balanced center-feed system, matching is significantly simplified and improved, and performance is easily verified with only a VSWR check. I hope that you will also try these designs and enjoy the great performance that is possible without all the pitfalls of the narrowband computer-optimized Yagis.

Part 2 will discuss longer boom models with higher gain using these same principles as well as more mechanical information. Stay tuned! ■

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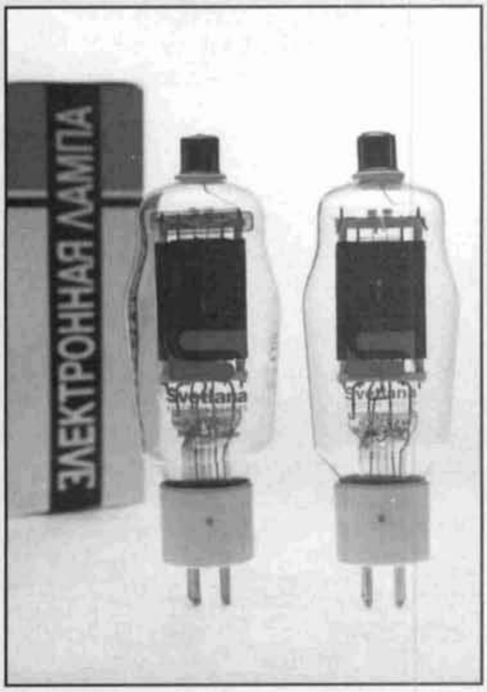
PRODUCT INFORMATION

Svetlana 812A

Svetlana has introduced a modern version of the 1939 RCA 812A. This medium-mu triode is a companion to Svetlana's high-mu 811A.

Like the 811A, the Svetlana 812A is a power triode intended for use in class AB, class B, and class C RF audio amplifiers. Both Svetlana tubes feature a low-loss ceramic base and a bonded-ceramic plate cap thermal insulator for high-power RF transmitting tube capability. The envelopes are made from hard glass intended for high-temperature operation of transmitting tubes. Svetlana provides a ceramic socket for both types (the SK4A, as well as a ceramic plate cap connector, the PC1A).



For more information contact Svetlana Electron Devices, Inc. at 3000 Alpine Road, Portola Valley, California 94028; phone: (650) 233-0429; fax: (650) 233-0439.

Circuit Design Software Features Mixed Analog/Digital Simulation

MicroCode Engineering, Inc. has announced the availability of CircuitMaker® Version 5. The program, a Windows-based schematic capture and simulation tool, now features fast, accurate mixed-simulation previously available only in high-cost EDA software. Along with the expanded simulation capability, the new release also features a larger device library of over 4,000 devices, easier SPICE model import, and no limit to the number of pins for an individual device.

CircuitMaker is an EDA software tool that integrates schematic capture and simulation in one program. Professional schematic capabilities include a built-in symbol editor, a macro feature for hierarchical devices, and SmartWires® automatic wire routing. Designers can export CircuitMaker schematics as PCB netlists for use in MicroCode's TraxMaker® or other printed circuit board layout products.

CircuitMaker Version 5 operates on the Windows 3.1x, 95, and NT platforms. Single-user copies are priced at $299, and special upgrades are available for current registered users. Multi-user site licenses for corporate networks and labs are also available.

To learn more about CircuitMaker Version 5 contact MicroCode Engineering, Inc., 927 West Center Street, Orem, Utah 84057-5203; phone: (801) 226-4470; fax: (801) 226-6532; Website: <http://www.microcode.com>.

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QUARTERLY DEVICES

Almost All Digital Electronics' digital frequency display kits

About 25 years ago, I undertook the building of a synthesized HF receiver. As Iraq's Sadam Hussein would describe it, this receiver was going to be the "Mother of all Receivers." The digital display required the counting of both the VFO and HFO and used an elaborate preload scheme to accommodate the IF frequencies. The design involved about 45 power-hungry TTL devices—a virtual raft of SN74192s and supporting chips (**Photo A**). A second board, with almost as many TTL devices, controlled the VCO synthesizer. The digital components used more power than the analog portions of the receiver. Most of the IC sockets were laboriously hand wired on a Vector Board that had a special ground plane foil.

Six months later, the digital portion of the radio was finished. The schematics overran two poster sized drawings, and the end was nowhere in sight. Pieces of that project are still scattered around my workshop. The receiver was never finished—the design was obsolete before I had time to complete the project.

Had the company Almost All Digital Electronics (AADE) been around back then, things might have been different! AADE has reduced that raft of ICs to a simple and inexpensive PIC processor-based kit that accommodates the simplest direct -conversion QRP transceiver, or complex HF, VHF, UHF, or SHF (up to 8 GHz with the appropriate prescaler) heterodyne mixing schemes.

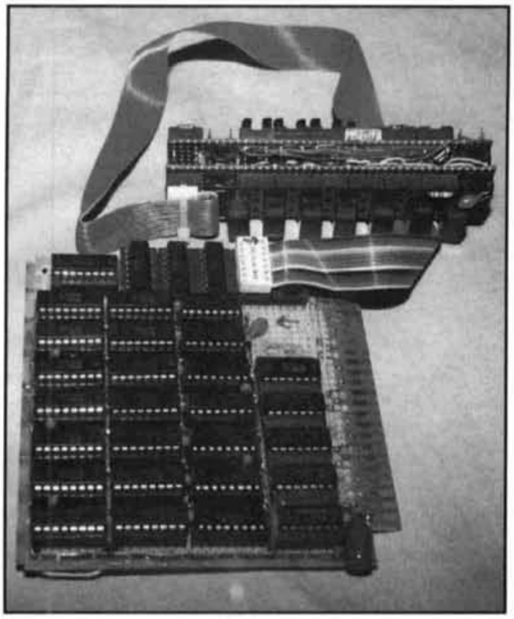

Photo A. My early attempt at a digital display required nearly 50 TTL devices. Today, it's replaced by a single PIC processor and LCD display.

General overview

The heart and soul of the AADE Digital Frequency Display (DFD) is a preprogrammed PIC16C71 microcontroller that features built-in A/D converters. Those with PIC programmers, who wish to scratch build any of the AADE display units, can download the programming code free of charge from the AADE Web site. The basic models allow prescaled inputs, ranging from 1 to 256. An onboard 15-turn trimpot permits an input multiplication of 1 to 256 via one of the A/D inputs. This corrects, in the software, the computed input frequency. The use of A/D processor inputs with fixed or variable resistors to set the operating parameters was a rather odd concept to me, at first. But, after seeing how well it works, I'm amazed at the elegant simplicity of these chips.

The DFDs can be used as inexpensive bench counters; the presets are set to zero for this application. The internal timebase is a non-tem-

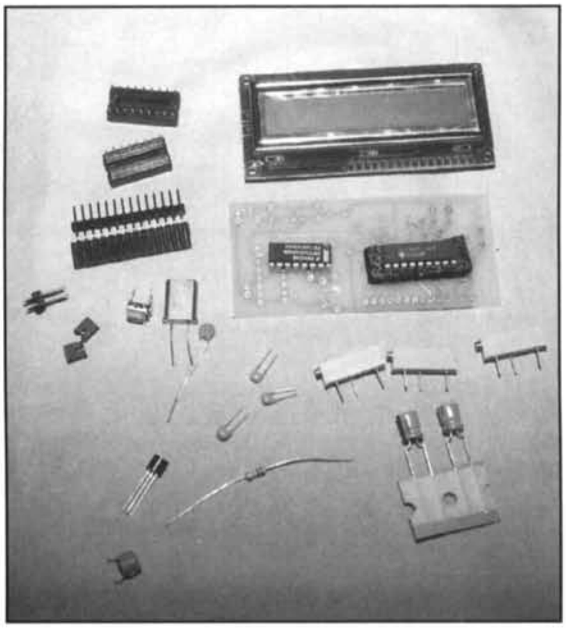

Photo B. The DFD1 kit.

perature compensated 16-MHz crystal. Calibration is performed by warping the crystal frequency so the display agrees with an external standard feeding the LO input. The exact frequency is based on CPU timing cycles; attempting to set the crystal to exactly 16 MHz will give inaccurate readings. These displays are intended for low-cost applications where the ultimate in display resolution or lab accuracy is not required. The seven-digit display limitation may bother some folks, but it fairly represents the limitations of a simple timebase standard and software-controlled timing intervals.

These little displays are great for modernizing vintage analog-display transceivers, receivers, or small QRP rigs. The different models handle premixing schemes, up or down conversions, or tunable IFs with equal ease. I intend to use one with my vintage tube Hallicrafter SX-28 receiver, and also in my

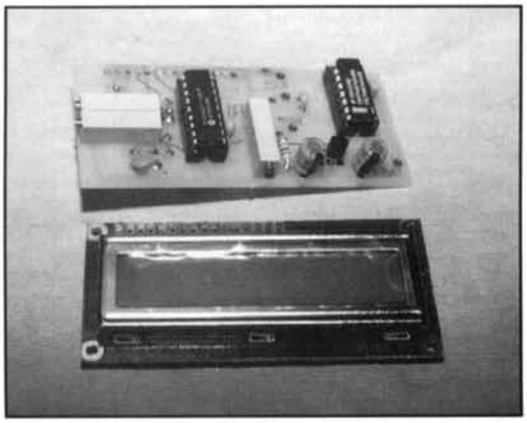

Photo C. The assembled DFD1.

MFJ 9406 6-meter and MFJ 9420 20-meter QRP SSB transceivers. The DFD requires LO signals between -22 dBm and + 18dBm (50 mV and 5 volts p-p). For most models, the maximum input, unless prescaled, is 32 MHz. This is a limitation of the 4046 LO buffer. CMOS level signals up to 60 MHz may be directly connected to the PIC processor.

The display consists of a 16-digit LCD panel. The AADE digital frequency display shows seven digits for a maximum 100-Hz display resolution for most models. A typical display presentation might be <14.2245 MHz LSB>. The LCD panel and its associated display driver chip are a preassembled OEM package. The LCD display assembly piggybacks to the main counter board.

The operating mode is displayed and is selected by a resistor value presented to an A/D PIC input. This may be a fixed value resistor for single mode radios, or a resistor value selected by an extension of the mode switch. Backlit LCD panels are optional. Typical power requirements are 8 to 18 volts DC at 20 mA. When used on a vintage tube radio, these voltages are easily supplied by a simple voltage doubler on the filament supply. The display

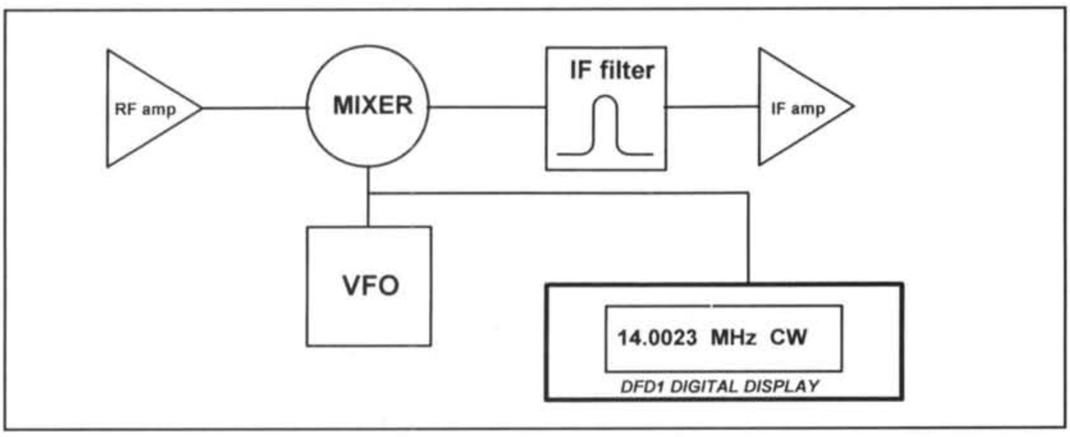

Figure 1. DFD1 connections to a single-conversion receiver or transmitter. Only the LO signal is sampled.

roughly measures 3-1/4 inches wide by 1-5/16 inches high. Optional mounting enclosures and bezels are available.

Model DFD1

Model DFD1 (**Photos B** and **C**) permits the use of IF offsets from 0 to 16 MHz in 1-kHz steps. The offset is jumper-selectable for either plus or minus offsets, allowing for high or low side injection. The offset polarity is set via a jumper, the IF frequency is set via a 15-turn trimpot. The maximum LO input frequency is from 0 to 45 MHz.

The DFD1 permits up to a plus or minus 12.5-kHz correction for BFO shifts for each mode. Mode displays for this model are AM, FM, CW, USB, LSB, FSK, FAX, or <blank>. The DFD1 is well suited for single or direct-conversion designs (**Figure 1**). The simplicity of these devices is amazing! Only two ICs—the PIC and the 4046 level converter—are used on the main processor board (**Figure 2**).

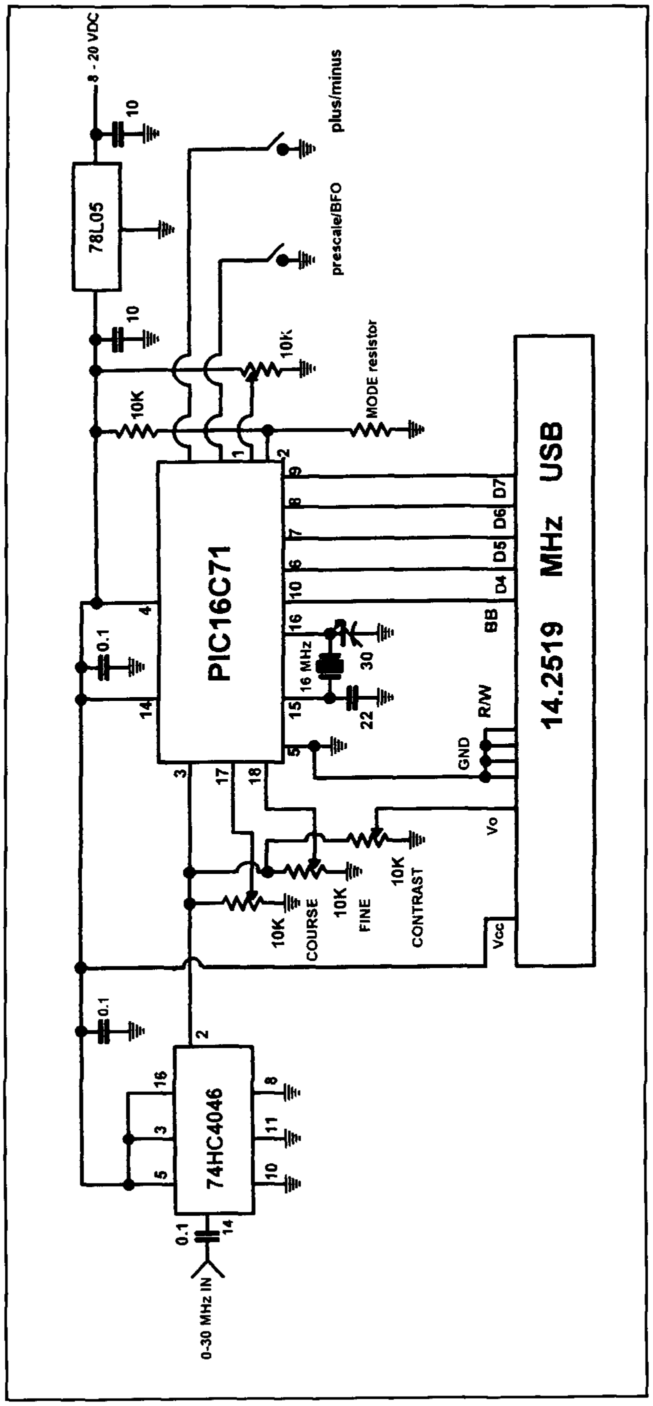

Figure 2. Schematic of the DFD1 processor board.

Model DFD2

Most of the features found on the DFD1 are also included in the DFD2. The DFD2 is ideal for tunable IF designs. Besides the ±0 to 16-MHz IF offset, the DFD2 samples signals from the HFO, BFO, and LO (**Figure 3**); each of these may be between 0 to 45 MHz. Custom chips are currently offered for the Collins S line, Yaesu FT-101, and the Kenwood TS-520. The automatic display modes for this model are AM, CW, USB, and LSB.

Model DFD3

The DFD3 is a variation of the DFD2, requiring only one direct connection to the VFO (**Figure 4**). The IF offset and HFO frequencies are custom programmed. Small variations in HFO and BFO calibration would be automatically corrected for in the DFD2 during the count sequences. The DFD3 solves the problem by permitting each band to be individually calibrated and allowing those values to be stored in EEPROM to correct for variations. This is a one-time procedure. Currently, several versions of Collins, the Yaesu FT-101 and FRG-7, and the Kenwood TS-520 are supported, with more to come.

Model DFD4

The DFD4 models cover the VHF and UHF spectrum. The IF offset is adjustable in 1-kHz steps from ±0 to 2 GHz. Prescalers are on board. Model DFD4A has an LO range of 0 to 0.5 GHz, DFD4B from 0 to 1.5 GHz, and DFD4C from 0 to 3 GHz.

Experienced builders should be able to assemble and have any of the DFD kits up and running in under an hour. Assembly instructions are cursory and consist of the schematic, photos, and a pictorial drawing. The usual precautions for working with CMOS devices

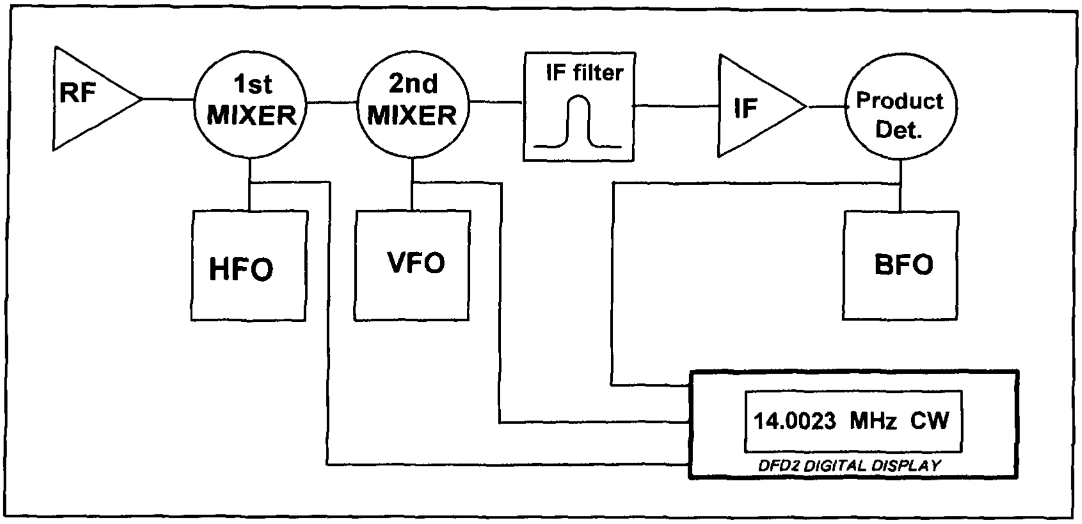

Figure 3. DFD2 connections to tunable IF in a receiver or transmitter. BFO, HFO, and VFO are sampled.

should be taken. These little PIC processors are truly amazing devices! Another AADE product, the model IIB L/C meter uses a PIC processor. I suspect AADE hasn't scratched even the surface for finding applications for these chips!

Ordering information

The DFD1, DFD2, and DFD3 kits cost \$49.95 plus \$1.50 shipping and handling. Assembled units are \$69.95. The backlit LCD module is an additional \$10. The DFD4 kits are \$59.95, or \$79.95 assembled. The custom extruded aluminum housing and bezel measure 4-3/8 x 2-1/2 x 4-3/4 inches (WHD). The kit price is \$15.95 plus \$4 shipping and handling.

If you'd like to order one of these kits, or you want more information, contact Almost All Digital Electronics, 1412 Elm St. SE, Auburn, Washington 98092. You can also reach them by phone at (253) 351-9316, by fax at (253) 931-1940, or on the Internet at <neil@aade.com>. Be sure to check out AADE's Web site, too. It's: <http://www.aade.com>. ■

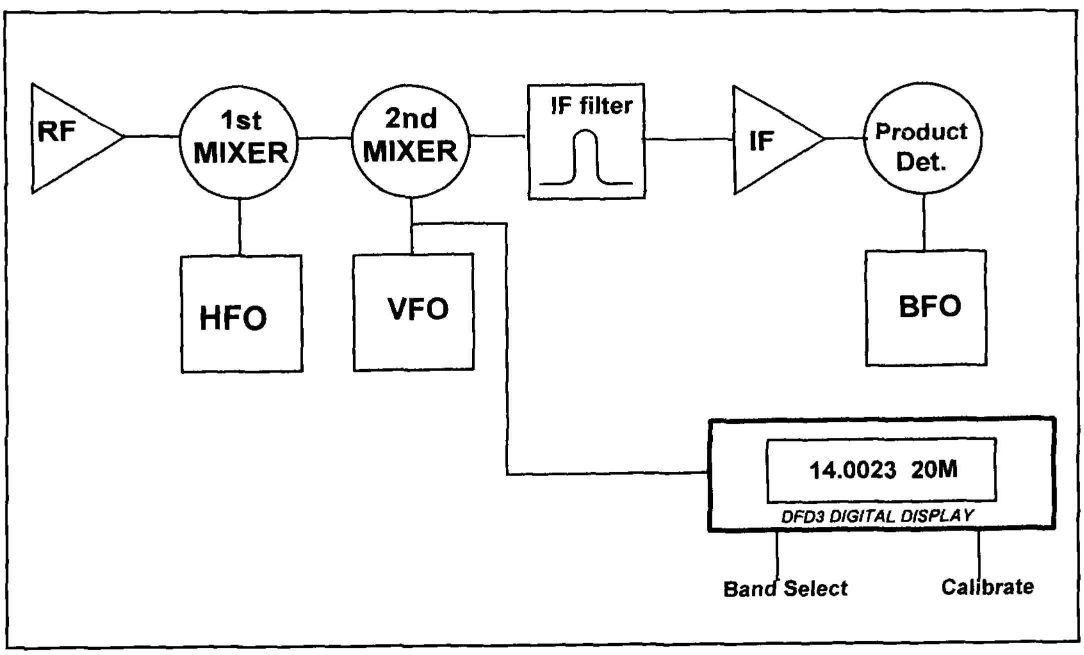

Figure 4. DFD3 requires only a single connection to the VFO of a tunable IF system. Custom values for several radio models are factory programmed.

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RADIO COMMUNICATION VIA THE MOON

The early days of EME

Radio reflections from the moon were first detected (in the United States and Hungary) in 1946 using VHF radars constructed from military radar equipment. These and subsequent observations at HF (in Australia) and VHF (in England) revealed that the echoes were subject to both rapid and slow fading. By 1954, it had been established by means of experiments at Jodrell Bank, England, that the slower fading was caused by the rotation of the plane of polarization of the radio waves in the Earth's ionosphere, or the so-called Faraday effect. The rapid fading was believed to be caused by interference between the many scattering centers on the surface of the moon, whose relative distance from the Earth is constantly changing owing to its libration (spin rate).

Studies of this fading at Jodrell Bank in 1956 confirmed this theory and revealed that most of the power in the reflected signals arose from scatterers lying near the center of the visible disk. The range extent of these returns was less than 1 ms; that is, much less than the full radar depth of the moon (approximately 10 ms). Accordingly, it was recognized that radio waves modulated by speech, or music, could be reflected from the moon and remain reasonably intelligible. Subsequent experiments, supported by the Pye Company, explored this form of communication as a possible alternative to overseas HF broadcasting.

In 1957, it was disclosed that the U.S. Navy had independently reached the same conclu-

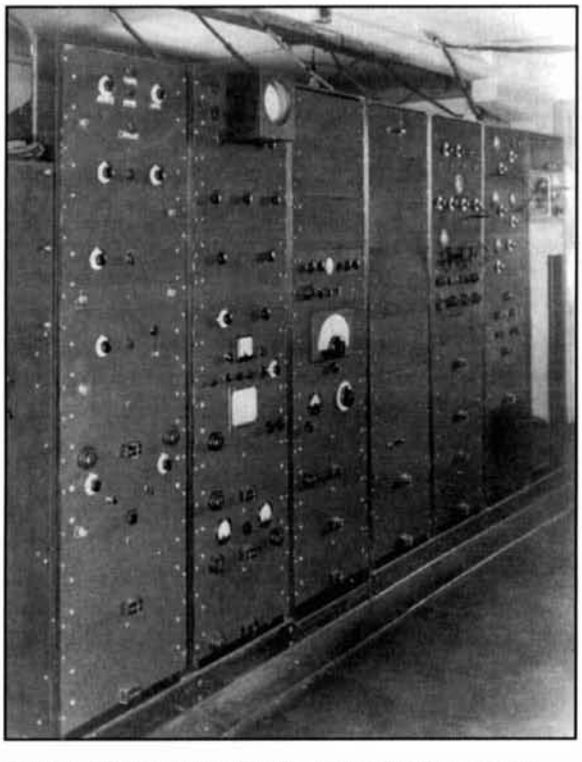


Photo A. The receiver and exciter states of the moon radar employed at Jodrell Bank. The receiver had IF frequencies of 10 MHz, 1 MHz, 100 kHz, and 1 kHz, permitting the echoes to be heard on a loudspeaker. They were also photographed using a camera housed in the case at the far left. The exciter (at the far right) employed frequency multiplying states, but this proved to be a poor approach as it exhibited unwanted chirp when pulsed. It would have been better if built as a reverse superhet.

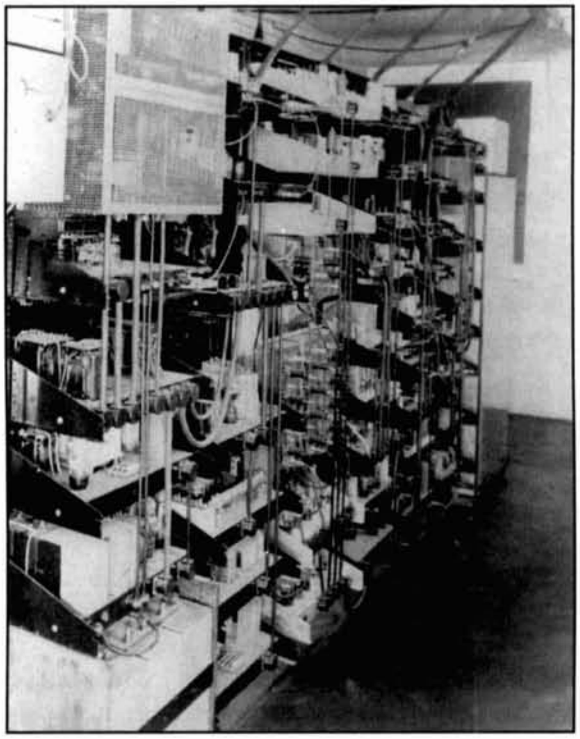


Photo B. The back of the receiver and exciter equipment employed in the moon radar at Jodrell Bank. Two racks of exciter equipment are nearest the camera with a pair of triodes in push-pull serving as the driver stage behind the screen. Power supplies are at the bottom of each rack. Vacuum tubes were '50s vintage B7G, B9A, or octal based.

sions through experiments in which short pulses reflected from the moon were examined for their range spread. The Navy had initially classified these findings. Subsequently, the Navy began construction of a 600-foot diameter, fully steerable radio telescope to observe the moon and thereby eavesdrop on Soviet military radio nets by monitoring any reflected signals. This project was never completed; the cost was grossly underestimated, and the advent of satellites gave the U.S. a more reliable way of eavesdropping. For similar reasons, the use of the moon as a passive reflector for overseas broadcasting never developed. Geostationary communications satellites provided a far better means of achieving this service.

Introduction

From the vantage point of the 1990s it seems incredible, but, for a brief period around the late 1950s, there was some serious consideration of using the moon as a passive reflector of radio waves for transoceanic communications. We now live in a "global village" created primarily through television and its worldwide distribution via communications satellites. It's easy to forget that this technology came into being only recently. Prior to 1926, there were no voice circuits across the Atlantic, only telegraph cables. During World War II, Churchill and Roosevelt conversed over radiotelephone circuits, which were not at all secure. Coaxial cables capable of supporting voice communication were laid across the Atlantic after the war; but, as late as 1965, when the first INTELSAT communications satellite was launched, there were only 300 such circuits. It's not surprising that, before that time, there was interest in other ways of spanning the oceans. These included reflecting signals off large balloons placed in orbit around the Earth, or copper wires and, for a brief while, the moon.

My involvement in these matters began when I attended the University of Manchester's radio astronomy research station at Jodrell Bank in 1954 for graduate work. There, I was assigned the tasks of studying the moon by radar to ascertain its scattering properties and of using the moon to reflect radio waves back to Earth to try to determine how many electrons were in the Earth's ionosphere. I will review only the first of these efforts here.

Prior work

The exploitation of radar during World War II led immediately to two attempts to reflect radio waves off the moon using an apparatus largely constructed from surplus wartime radar equipment. Mr. Z. Bay in Hungary succeeded in detecting echoes, but was obliged to integrate many of these echoes (by applying the receiver output voltages to a set of electrolytic cells which released gas) to establish their presence. DeWitt and Stoldola[1] in the United States used a more powerful radar that allowed them to see individual pulse returns on the display. They found considerable variation in the amplitudes from pulse to pulse, and on some occasions the echoes were absent altogether, though the radar appeared to be operating properly.

These results spurred Grieg et al.[2] to examine the possibility that the moon could be used as a passive reflector in a radio relay circuit between continents. Two concerns arise in such a scheme. First, the radio system must be capable of overcoming the large loss of intensity in traversing the approximately 384,400 kilometers to the moon and back. Owing to the spherical expansion of the waves from any antenna, the flux density falls with the square of the distance R from the source. Thus, in the two-way journey to the moon and back there is a reduction of a factor $\sigma/(4_{\pi}/R^2)^2$, where σ is the scattering cross section of the moon. Even if the moon were a perfect reflector, this loss would be $2.78 \times 10^{-24} m^2$ or -235.6 dB. In actuality, the moon reflects only about 7 percent of any

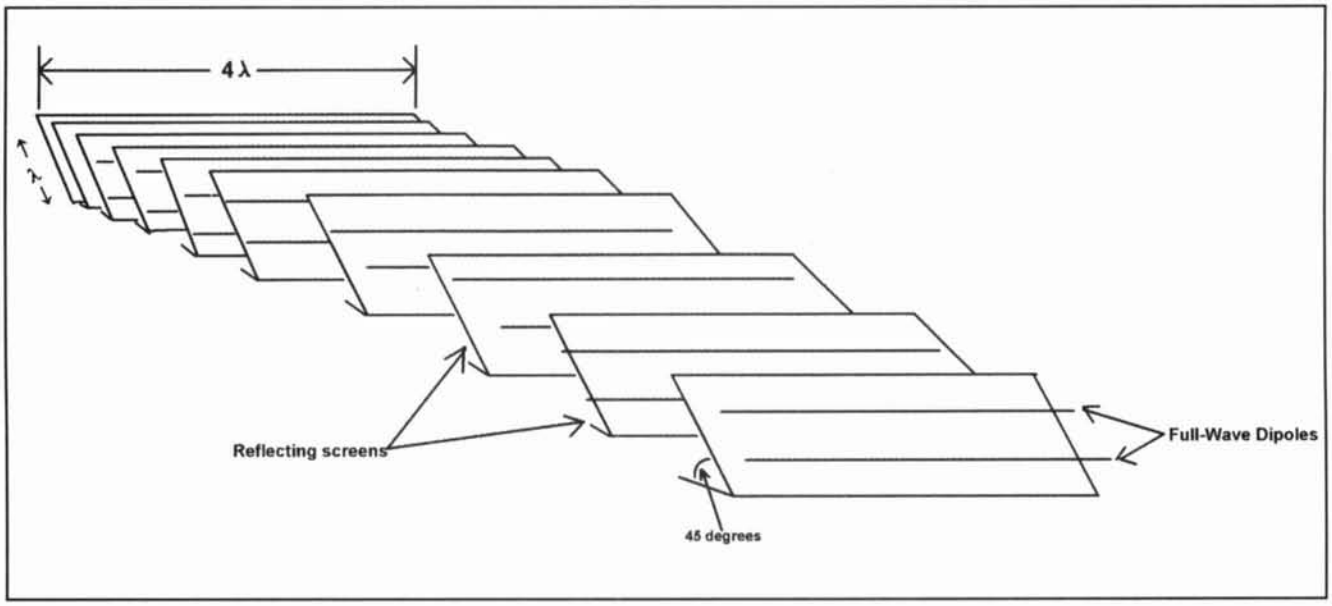


Figure 1. The antenna system used at Jodrell Bank during 1953–57 to study the moon at 120 MHz.

incident meter-length radio waves, so the overall loss is closer to -247 dB. Therefore, a fairly powerful transmitting station is required with a directive antenna capable of beaming the energy toward the Moon, and the receiving station must have a similar antenna. Waves shorter than about 5 meters are required to reliably penetrate the Earth's ionosphere, and, at the time Grieg et al.[2] wrote their paper, transmitters suitable for the kind of service they contemplated could be built at high power only for wavelengths longer than several centimeters. Accordingly, these considerations bounded the region of the radio wave spectrum in which such a relay service might be contemplated.

The second issue that Grieg et al.[2] considered is the way in which the moon might scatter the incident wave. They speculated that some of the fading observed by DeWitt and Stodola may have been caused by the presence of multiple scatterers on the lunar surface. These would contribute reflections that were sometimes constructive and at other times destructive. The critical issue was the extent in range of these scatterers.

When viewed optically from the Earth at full moon, the lunar surface appears uniformly bright. That is, the limbs are about equally bright as the center, despite the fact that sunlight is there incident at a grazing angle. Were the moon to scatter radio waves in the same fashion, the echoes would be returned with a spread in delay of 11.6 ms. This would cause any modulation on radio waves with a frequency of greater than about 100 Hz to be destroyed. There the matter was left until further (secret) work was undertaken in the United States and my work began at Jodrell Bank in 1954.

Progress was made, however, in understanding the nature of deep echo fading. Kerr and Shain[3] in Australia performed experiments using a shortwave broadcast transmitter and were able to distinguish between a short period

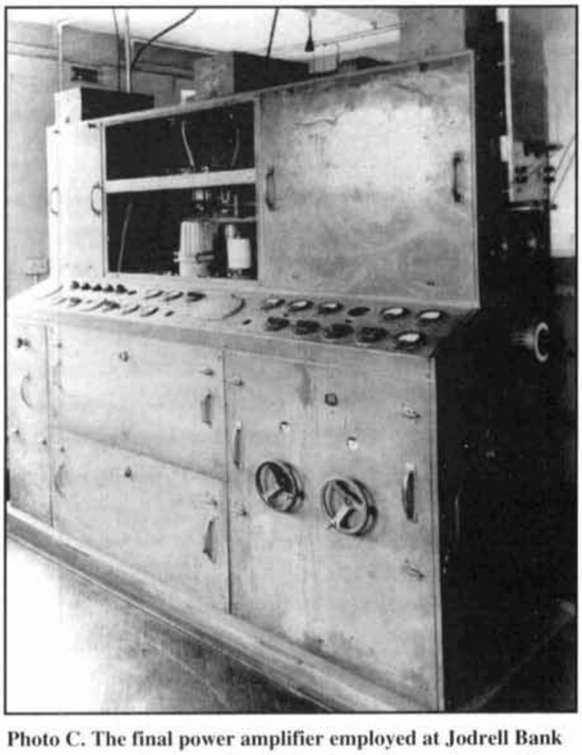


Photo C. The final power amplifier employed at Jodrell Bank in the moon radar experiments (replacing an earlier one that only provided 3 kW output). This was built into the case of a World War II radar set. Shown with the side cover removed is the Class C push-pull amplifier. This employed a pair of QY4-250s with tuned anode and grid lines. The tubes were mounted upside down so a fan could direct a flow of air, through the glass chimney, onto the anode seals. The plate voltage applied was 10,000 volts.

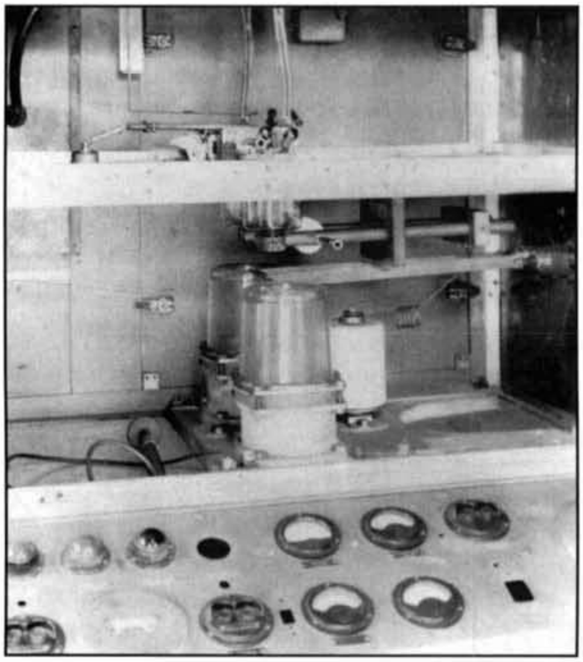


Photo D. Close-up of the final power amplifier at Jodrell Bank. This was capable of transmitting 10 kW pulses of 30 ms duration. Resting on the glass tubes that direct air onto the anode seals is the output cooling loop, which was connected to the open wire transmission lines that fed the antenna. Power was measured by operating the transmitter at a high pulse repetition frequency with short pulses while it was coupled to a strip lamp. Main power, also supplied to the strip lamp via the variac (shown at the end of the cabinet), was lowered to keep the brightness constant (and the impedance of the lamp constant) as the transmitter power was increased.

fading of the echoes (presumed to be of lunar origin) and slower overall large changes in the strength of the echoes, which they suggested was of terrestrial ionospheric origin. This was confirmed later at Jodrell Bank by Murray and Hargreaves,[4] who recognized that plane-polarized radio waves traversing the ionosphere have their plane of polarization rotated depending upon the number electrons lying along the path. This is known as the Faraday effect. The amount of rotation is doubled on the return path. Thus, in the course of the day, as the ionosphere builds up or decays, there will be periods when there is 90 degrees difference between the polarization of the waves and the receiving antenna (see also Browne et al.[5]); the echoes will then be unobservable.

The Jodrell Bank experiments

Radar reflections from the moon were successfully obtained at Jodrell Bank by Murray and Hargreaves[4] using a radar they constructed that operated at 120 MHz at a power of about 3 kW for pulse lengths of 30 ms (see **Photos A** through **D**). The antenna was an array of dipoles (**Figure 1**) that could be phased to alter the elevation of the beam. The array consisted of 10 elements, each of which comprised a reflecting screen tilted back at 45 degrees and placed in line behind the previous one. Each screen was 1 wavelength wide and 4 wavelengths long and was illuminated by two rows of full-wave dipoles. The physical aperture was 250 m_2, and the overall efficiency was between 60 and 70 percent. The antenna allowed the moon to be seen at transit (i.e., due south) for about an hour each day for about two weeks each month, when the moon is highest.

The echoes obtained with this equipment were photographed from an A-scope display. In **Photos E** and **F**, which illustrate both the slow- and long-period fading, the transmitter was pulsed every 1.8 seconds. Measurements of the correlation between the echoes, through the calculation of the noise-corrected autocorrelation coefficient, gave values in the range 0.2 and 0.4, indicating a fading period on the order of 1 second or less. The amplitude distribution was found to match the Rayleigh Law (see **Figure 2**), indicating that a large number of scatterers contributed to the returns (i.e., the lunar surface did not offer one dominant reflecting region).

The correlation between the echoes exhibited small night-to-night variations, which appeared to depend upon the apparent libration of the moon as seen from Earth. **Figures 3A** through **C** illustrate the causes of this libration. Diurnal libration (having a value at transit of approximately 12 cos ø x 10^{-7}rad/s, where ø is the latitude of the terrestrial observer) is the largest of these; but at different times in the lunar month, the libration in longitude (having a maximum value of 4 x 10^{-7}rad/s) can add or subtract.

This apparent spin of the moon can be thought of as giving rise to a Doppler broadening of the echoes. A strip of the lunar surface along the apparent instantaneous spin axis reflects signals without imparting any Doppler shift (see **Figure 4**); but strips on the approaching hemisphere are Doppler-shifted to higher frequency, while the reverse is true for the receding hemisphere. The maximum Doppler shift, f_0, for reflection from the limbs in the Jodrell Bank experiments was given by 1.4 x $10^6 L_T$Hz, where L_T is the libration rate and is on the order of ±2 Hz.

Nowadays, one could measure this Doppler broadening by performing a phase-coherent analysis of the returns in a digital computer, but in 1955 we had no digital computer and had to approach the problem by measuring the echo autocorrelation function—recognizing that this is the Fourier transform of the echo power spectrum. (Strictly, this is true for power-law detectors and is modified in the cast of linear detectors; see Lawson and Uhlenbeck.[6])

To explore the correlation of the echoes over intervals shorter than the normal 1.8-second

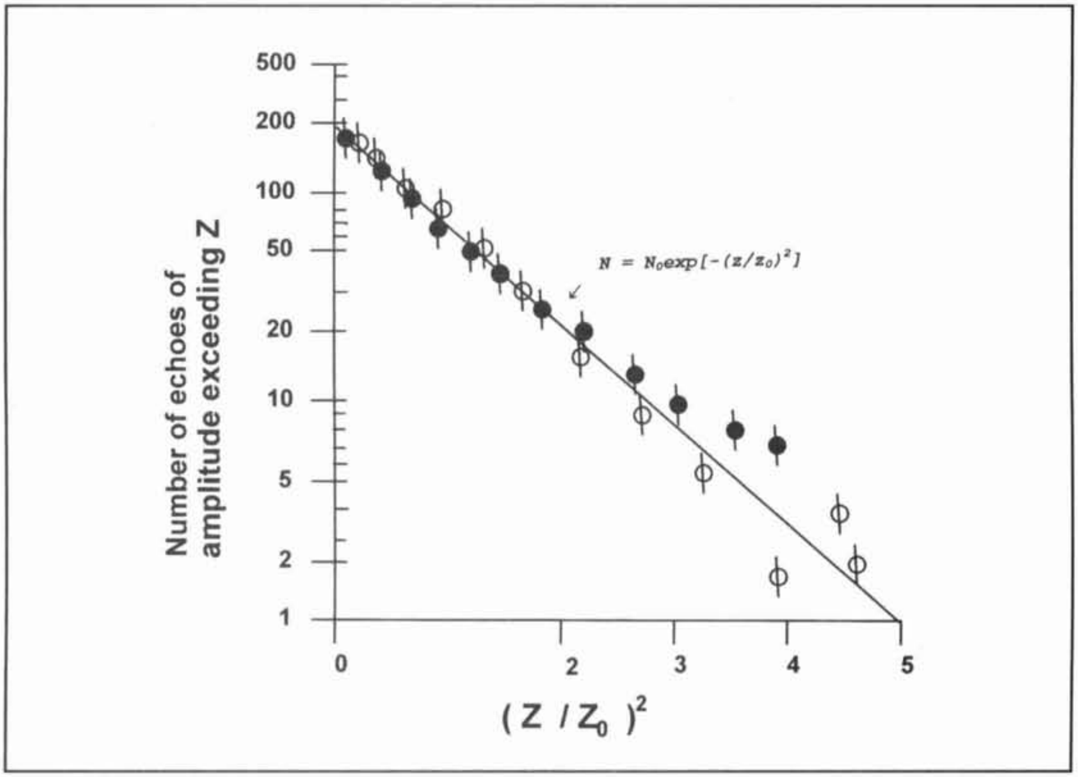
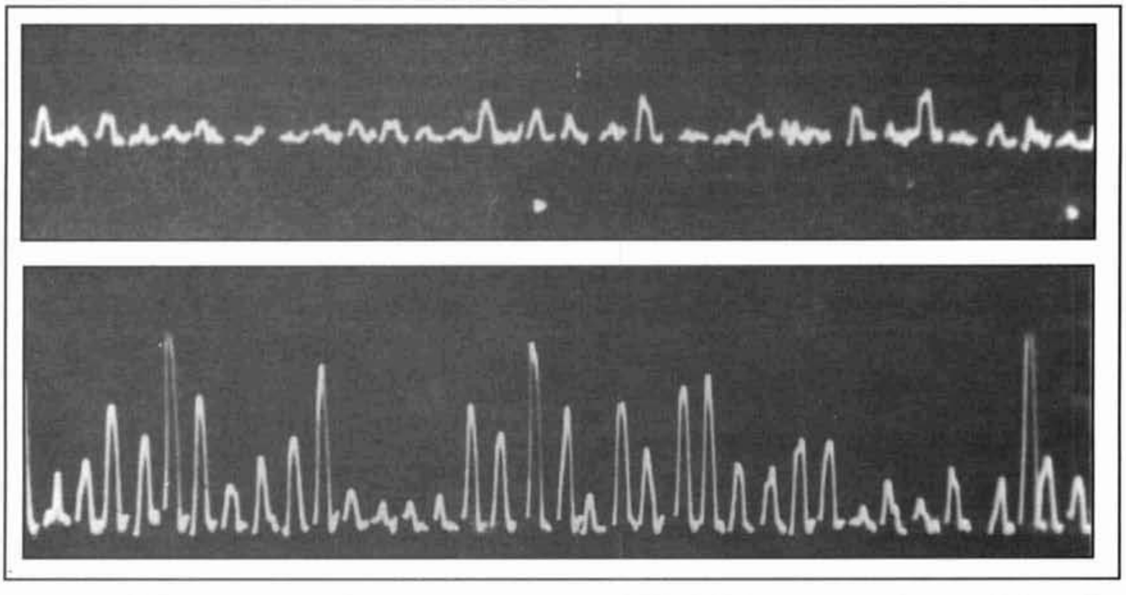


Figure 2. Plot of the number of echoes having an amplitude greater or less than the mean, z_0, showing that the distribution fits the Raleigh Law.

repetition interval, the radar was modified to transmit pairs of 20-ms pulses at intervals of 1/4, 1/2, 3/4, 1, 1-1/4, and 1-1/2 seconds every 4 seconds. **Figure 5** shows the autocorrelation function measured using these pulse-pairs plotted against the product of the maximum Doppler shift, f_o, and the pulse separation, t.

A sensible single curve is obtained showing that the fading varies with the libration rate, L_T, and is therefore of lunar origin.

Figure 6 shows the resulting power spectrum for the echoes obtained by fitting a Gaussian function ($\exp.[-1.3 f_o t)^2]$) to the observed autocorrelation function. **Figure 6** also shows the



Photos E and F. Examples of echoes from the moon observed at Jodrell Bank. Pulses were sent at intervals of 1.8 seconds, and the pulse-to-pulse variability is caused by multiple reflections from the lunar surface. The overall difference between A and B is caused by the Faraday effect in the Earth's atmosphere.

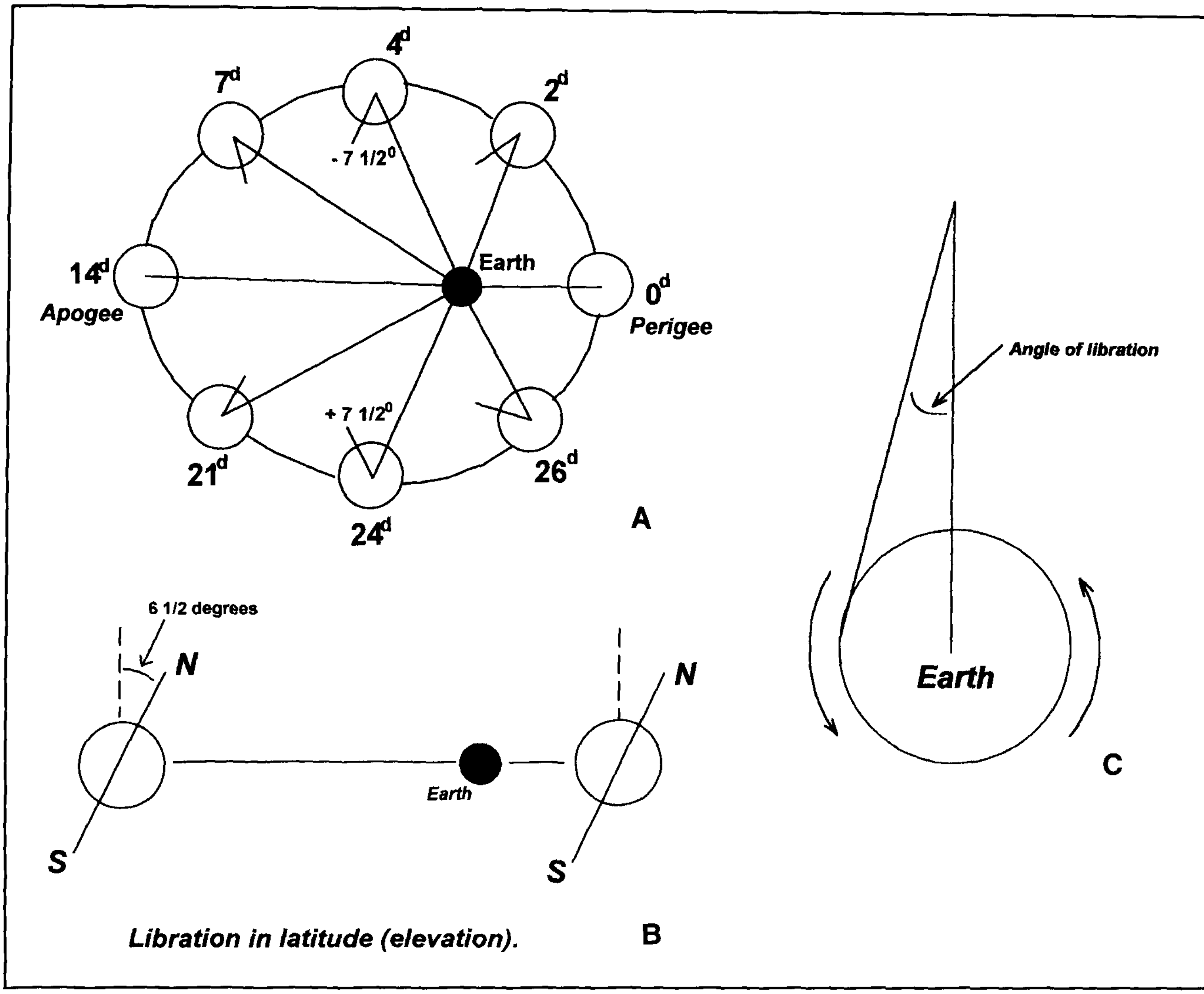


Figure 3. Causes of lunar libration: (A) Libration in longitude is caused by the elliptical orbit of the moon, which prevents the moon from presenting exactly the same face to the Earth; (B) Libration in latitude caused by the tilt of the lunar spin axis with respect to the plane of its orbit; and (C) Diurnal libration caused by the motion of a terrestrial observer.

expected Doppler spectra were the moon to scatter uniformly brightly (Lommel-Seeliger*) or according to the Lambert Law (reflected power varies as the cosine of the angle of incidence). It's evident that the bulk of returns are from a region of the lunar disk having a diameter on the order of 1/5 that of the moon. Evidently, the regions most nearly normal to the ray path return most of the echo; hence, it can be concluded that on the scale of the wavelength employed (2.5 meters), the moon appears relatively "smooth." Subse-quently, more precise experiments and improved theory permitted the mean surface slope to be determined from radar reflection studies.

* This is the scattering law that the moon's surface obeys at optical wavelengths, and it corresponds to the case where the backscatter is independent of the angle of incidence. The Lambert Law is one in which the backscatter intensity falls with the cosine of the angle of incidence.

For intervals on the order of 10 times the wavelength λ, this is about 5 degrees for $\lambda = 2.5$ m.[7]

To confirm the conclusion that most of the echo was from the center of the lunar disk, experiments were performed with pulses only 2 ms in length (i.e., shorter than the 11.6-ms range depth of the moon). The echoes showed no measurable range delay broadening when contrasted with photographs of the same transmitter pulses leaked into the receiver.[8]

The results described here were presented at an international conference (URSI General Assembly, Boulder, Colorado) in 1957 and provoked the release of then classified results obtained by the U.S. Navy.[9] In this work, a very powerful (1-MW) radar operating at 198 MHz was employed, together with a large parabolic reflector. Using pulses of only 12 μs in length, Trexler[9] reached the same conclusion I did—namely that the bulk of the observable return was from the center of the lunar disk and had a limited range extent (on the order of 100 to 200 μs [see **Photo G**]). The implications of these findings were that amplitude-modulated radio waves reflected by the Moon would not be so distorted that all of the intelligence would be removed.

Sequel

At Jodrell Bank, we conducted further experiments to examine the bandwidth of a moon-relay circuit using the newly completed 250-foot diameter radio telescope. These experiments were performed at 162.4 MHz with equipment kindly loaned by the Pye Company of Cambridge.

The idea was to see if the moon could be used for broadcasting overseas more reliably than HF. **Figure 8** shows how average intensity of a single audio tone fell off as the modulating tone frequency was increased.[10] These experiments were performed using amplitude-modulated (i.e., double sideband) signals and no simple theory is available to relate the results to the impulse function for the moon. Had single sideband been employed instead, the correlation between the carrier and the tone sideband would be expected to fall off as the Fourier transform of the impulse response.[10] The average intensity should then have fallen off less rapidly than shown in **Figure 9** because only one sideband would have been recovered at the receiver and the possibility of the two sidebands interfering (i.e., having different amplitudes and phases) is removed.

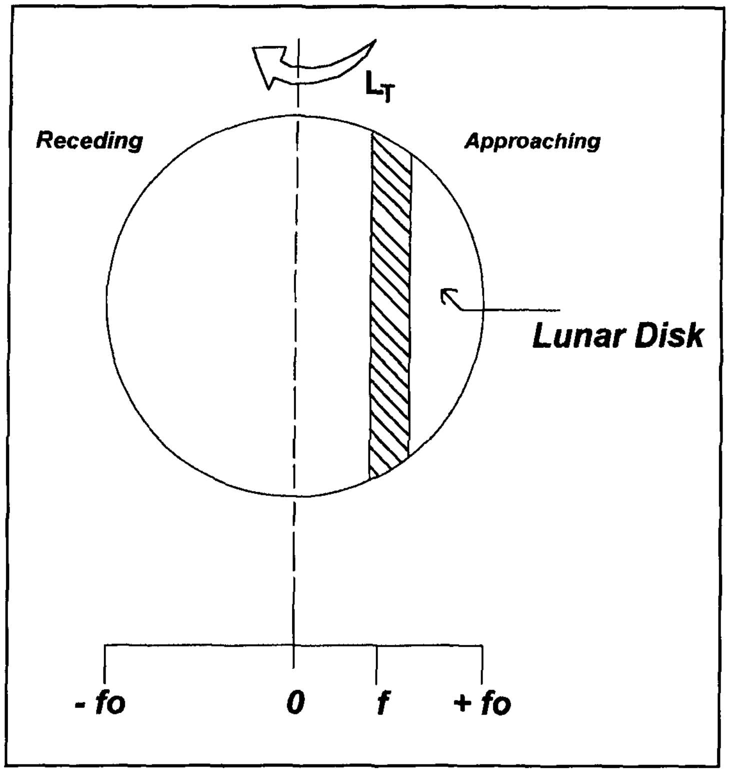


Figure 4. Doppler broadening of lunar reflections caused by the instantaneous apparent spin (at rate L_T). Strips on the lunar disk parallel to the spin axis contribute the same Doppler shift, f.

These experiments showed that speech-modulated signals were recognizable after reflection

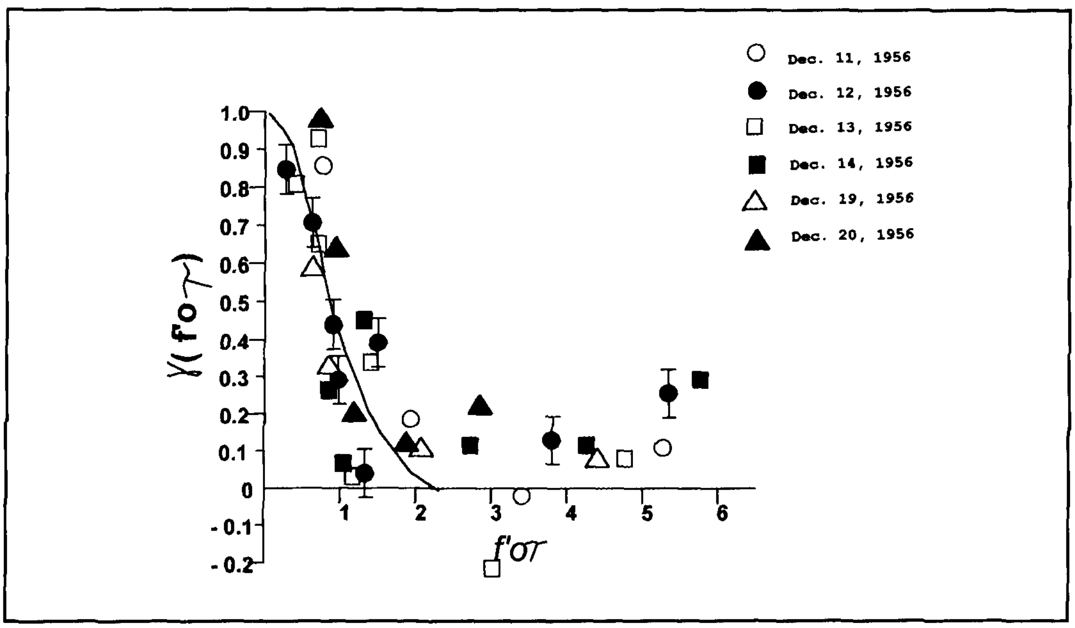


Figure 5. Correlation between pairs of echoes observed at Jodrell Bank as a function of their spacing τ and the maximum Doppler shift, f_0 (see Figure 4). The solid curve is a fitted Gaussian function.

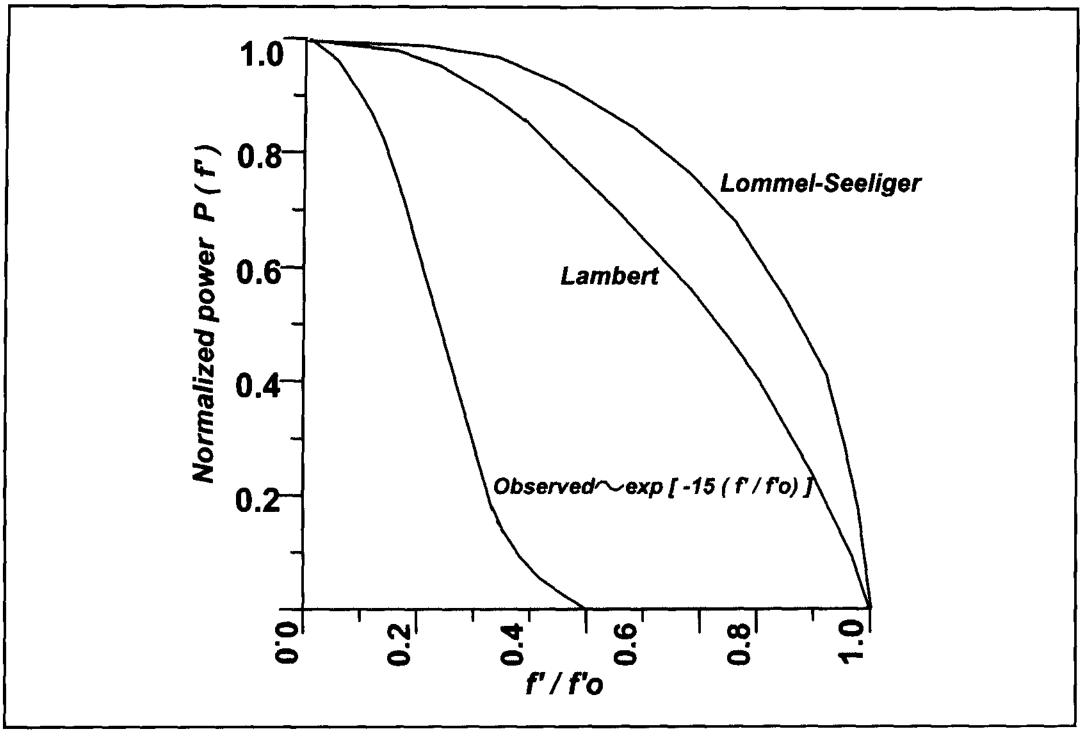


Figure 6. The Doppler broadening of lunar echoes derived from the results shown in Figure 5 compared with a uniformly bright lunar surface (Lommel-Seeliger) and one obeying the Lambert Scattering Law.

from the moon, but that the fading distorted music and made it unsatisfactory. Evidently, a feasible relay system could be constructed if: circularly polarized waves were employed (to overcome the Faraday rotation); suppressed-carrier signal sideband were used with modulation frequencies limited to <3–4 kHz; and a large signal-to-noise ratio, S/N, was provided so the effects of the rapid fading could be minimized using automatic gain control (AGC). The advent of communications satellites, however, obviated the need for any such scheme. Radio amateurs around the world still use the moon to make contact with one another (using Morse

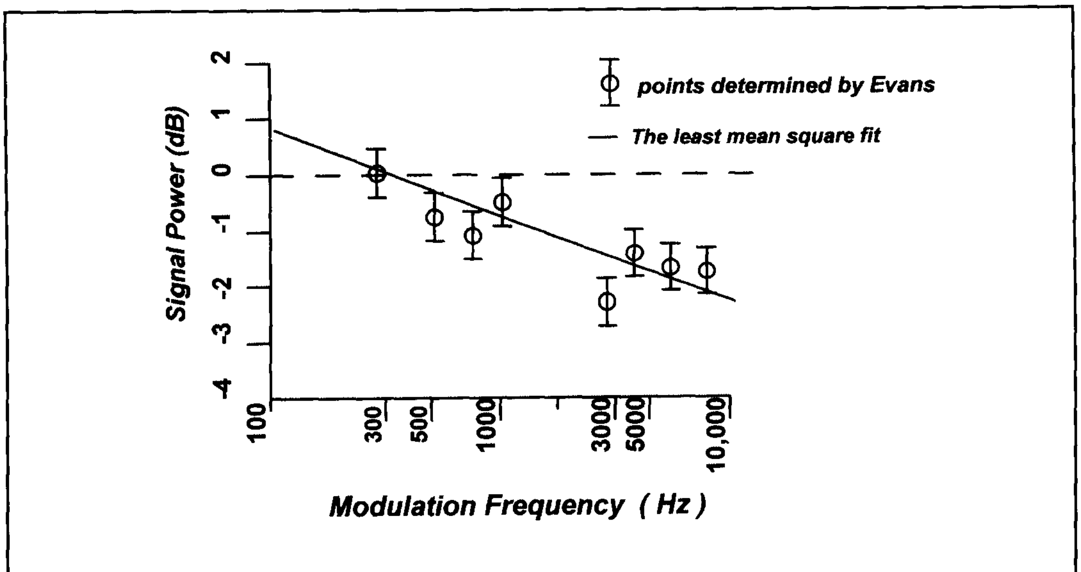


Figure 7. Results obtained at Jodrell Bank showing the average intensity of an amplitude-modulated signal reflected from the moon as a function of the modulation frequency. These results demonstrated that the bandwidth of a moon-relay circuit is adequate for speech.

code). They call this mode "earth-moon-earth" (EME), and operate at frequencies between 50 MHz and 12 GHz.

In the United States, a different idea was conceived for exploiting the radio reflection properties of the moon. Since it was evident that intelligible signals could be received after lunar reflection, it was recognized one could "eavesdrop" on radio transmissions in other countries if a large enough antenna were built and pointed continuously at the moon. Given that considerable military traffic in the USSR was then being passed over radio circuits, the U.S. Navy decided to build a 600-foot diameter radio telescope to monitor this traffic by lunar reflection. Construction of this telescope was begun at Sugar Grove, West Virginia, in about 1958, but was abandoned in 1963. It fell victim to rising costs and the recognition that satellites could be used to collect this information more reliability.

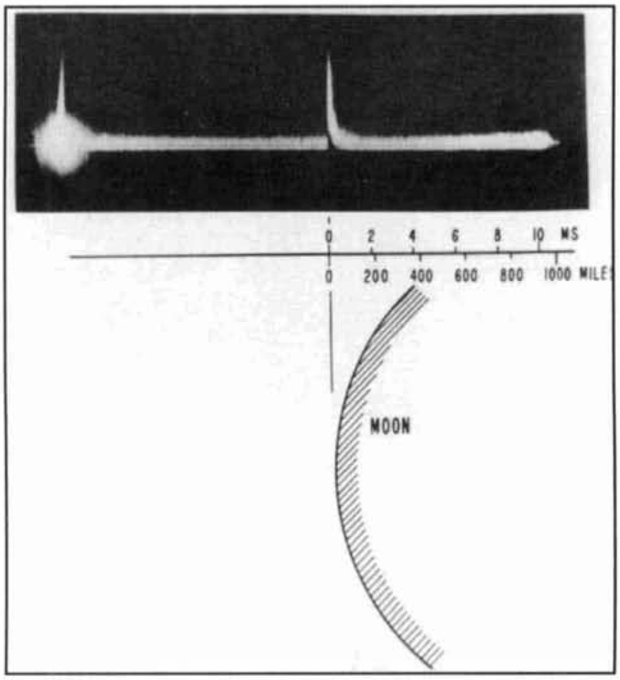


Photo G. Reflections observed from the moon (shown to scale here) by Trexler[9] employing a powerful radar using 12-µs pulses. The short range extent confirms that the reflections are largely from the center of this visible disk.

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10. J.V. Evans, "The Bandwidth of a Moon Communication Circuit," *British Journal of Applied Physics*, 12, 1961, pages 406–490.

PRODUCT INFORMATION

New Serial Communications Catalog

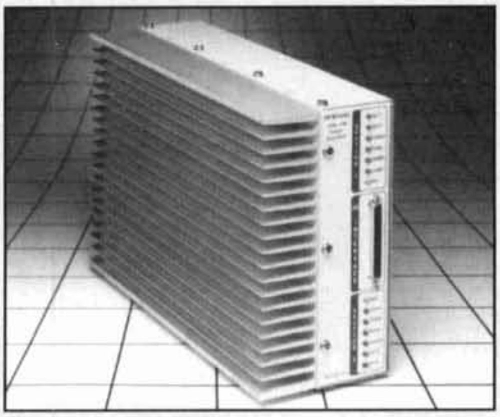
B&B Electronics offers a 48-page full-color catalog that contains 34 new products, including expanded sections of data acquisition equipment and products for fiber optic communications installation and testing. For a free copy, contact B&B Electronics at B&B electronics Mfg. Co., 707 Dayton Road, Ottawa, Illinois 61350; phone: (815) 433-5100l; fax (815) 434-7094; e-mail <sales@bb-elec.com>.

New RF Amplifier for Broadcast Applications

RF Gain, Ltd. has announced the availability of a new, compact, 1000-watt broadband amplifier for FM broadcast applications. The MFM100 features a rugged, fully heat-sinked connectorized assembly and all on-board circuitry. The device's production circuit design helps prevent failure and increase product life.

Operable between 87.5 and 108 MHz and featuring a high gain of 17 dB minimum, the MFM1000 is ideal for FM transmitters or as a driver for high-power tubes in other broadcast applications. It can also be combined for multi-kW transmission.

The MFM1000 is available from stock from RF and microwave component distributor Richardson Electronics. Check their Website at <http://www.rell.com> for more information.



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The HBR-Twenty

By M.A. Chapman, KI6BP

3625-7 Vista Oceana
Oceanside, CA 92057

Summary: An article describing a 20-meter, high-performance CW receiver. The receiver circuitry is built on several PC boards for installation in an attractive cabinet with a digital display.

Winter 1998 issue, pages 74-80.

Please contact the author for additional information.

PRODUCT INFORMATION

Frequency Synthesized VHF FM Exciters and Receivers

Hamtronics has a new line of VHF FM transmitters and receivers. The T301 exciter and R301 receiver provide NBFM and FSK operation on 144 to 148 MHz (and 148 to 174 MHz for export and government services). Features include dip switch frequency selection, low noise synthesizer for repeater service, commercial TCXO for tight frequency accuracy in a wide range of environmental conditions, and fast delivery with no wait for channel crystals.

The T301 exciter uses direct FM modulation, which allows FSK transmission of data up to 9600 baud. Power output is 2 to 3 watts and is rated for continuous duty in demanding applications, such as repeater service. The R301 has the same sensitivity, selectivity, and squelch found in other Hamtronics® receivers.

Both the T301 and the R301 are available either in kit form or factory wired and tested. The T310 exciter is $109 for the kit, or $189 wired and tested. The R301 receiver is $139 for the kit, or $209 wired and tested. Kits use a crystal (supplied) to generate the reference frequency, and TCXO is optional at $40. Factory-built units include at TCXO as standard equipment.

For details, contact Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535; phone: (716) 392-9430; fax (716) 392-9420; e-mail <jv@hamtronics.com>.

Low-Cost Wideband Microwave Synthesize

RF Prototype Systems offers a microwave synthesizer that is useful as a local oscillator for satellite upconverters or low-cost test sources. It has an output bandwidth range of 2600 to 4750 MHz. Output power is typically +12 dBm and accepts a 10-MHz reference input. The

phase noise is typically -115 dBc/Hz at 1-MHz offset from the carrier, with a parallel interface. Lower phase noise options are available.

For more information, contact RF Prototype Systems, 9400 Activity Road, Suite J, San Diego, California 92126 for pricing and availability. You may also call (619) 689-9715 or fax (619) 689-9733. Those outside of California may call toll-free at (800) 874-8037.

Antique Electronic Supply Has New Catalog and Website Address

Antique Electronic Supply has released its 1998 catalog. The company is the leading supplier of most new old stock (N.O.S.) vacuum tubes and currently manufactured vacuum tubes. They also supply many hard-to-find parts and supplies for the audio, guitar, amateur radio, and antique radio restoration markets. Also available are books and literature for tube gear enthusiasts.

The catalog has been expanded to 60 pages with lots of new tubes, parts, supplies, and books. Those who are not already customers can call (602) 820-5411 or fax (800) 706-6789 for a free copy, or you may visit Antique Electronic Supply's new Website at: <www.tubesandmore.com>.

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Quarterly Computing—Using Electronic Workbench to Analyze RF Circuits

By M.A. Chapman, KI6BP

3625-7 Vista Oceana
Oceanside, CA 92057

Summary: A description of the features of *Electronics Workbench* software (EWB). By using EWB, you can design RF circuits quickly and avoid problems when building prototypes.

Winter 1998 issue, pages 81-86.

Please contact the author for additional information.

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PREDICTIONS FOR SOLAR CYCLE 23

A graphical solution based on historical evidence

The sunspot cycle is one of the greatest mysteries of modern science. Currently, no satisfactory theory exists for the origin of sunspots. Therefore methods for predicting the behavior of future solar cycle activity levels are based on data from previous cycles

The first cycle

Somewhere, there is a 25,000-year-old antler displaying long to short markings. This antler is the earliest-known evidence that our ancestors noted the waning and waxing of the Moon. In the beginning, all time was Moon time.

Hot and cold cycles

Eight thousand years ago, along the Russian steppes, our ancestors discovered the hot and cold cycle of 365 days. Now, for the first time in history, we no longer had to survive as hunters. Because we had learned to measure time in relation to the angle of the Sun to the southern horizon, we could become farmers.

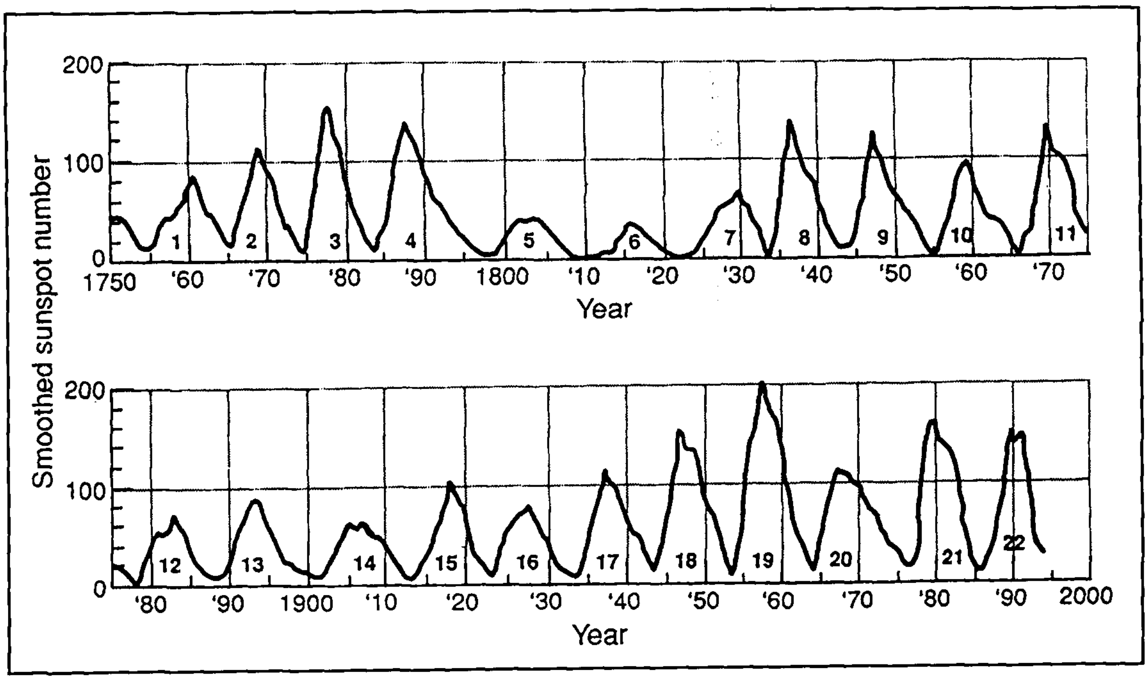


Figure 1. Variations in the sunspot cycle from the 1700s to the present.

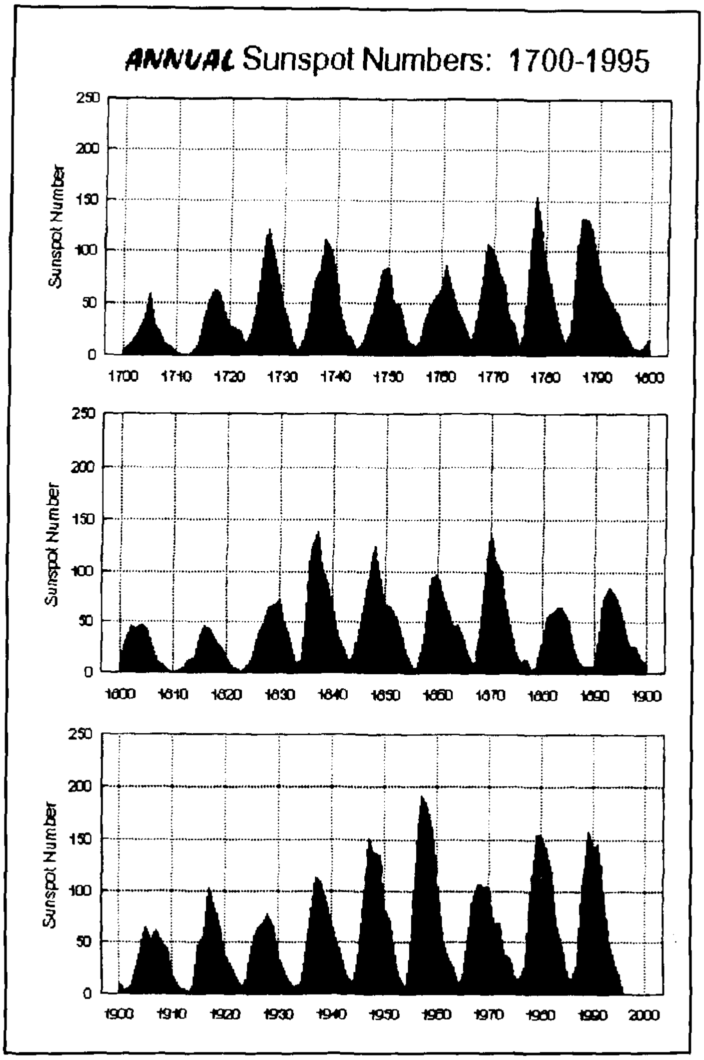


Figure 2. Annual sunspot numbers: 1700–1995.

All time was now Sun time. Our very existence, as well as the waxing and waning of history and the rising and falling of civilization, has been due to the rising and falling energy of the Sun.

We think of today's temperatures as being normal, but they are not. During the first 90,000 years of the past 100,000, the Earth's temperature was 5 to 7 degrees centigrade cooler—a glacial climate. For the last 10,000 years, there have been at least five cycles of warming and cooling.

The Maunder solar minimum

During the 1600s, the Earth became colder and the output of the Sun waned. It had all happened before, but this time was different. Our ancestors numbered in the millions, and they burned all the wood they could find to stay warm. In desperation, they began burning rocks or, as we would say today, coal. These were hard times, and a century would pass before the sunspots would come again and the Earth would warm.

Late in the 1700s, the sunspots sputtered back to life. It seemed they ran in 11-year cycles. However, these cycles sometimes varied with higher and higher peaks followed by lower and lower peaks; then the pattern would begin again (**Figures 1** and **2**). One would think that tracking such regular cycles of ups and downs would be easy, but this hasn't proved true, as no predictions have been consistently correct. It seems the cycles never repeat themselves exactly, making such predictions difficult.

Hampering the efforts of those who would predict the sunspot cycles is a lack of knowledge regarding how the Sun works. What we do know is based on fundamental physics. This much we have determined: If there is, in one place, a large amount of hydrogen, gravity from the mass will cause this hydrogen to move into a smaller and smaller space. This, in turn, will create higher and higher temperatures. The process will continue until fusion occurs and hydrogen becomes helium plus energy. This energy creates a back pressure that inevitably builds until it equals the central or inward **pressure**, creating a stable and ongoing process. Gravity and fusion have "turned on" the Sun and all the other suns we call stars.

This is truly wonderful knowledge, but we have no details. For instance, how does this energy propagate from the center to the surface? Why does the Sun have a magnetic field? Why does the magnetic field give rise to sunspots and why do we have more ultraviolet radiation when we have more sunspots? Why do these sunspots come in cycles? Why do we experience centuries of no sunspots followed by centuries of varying sunspots?

The Sun as a dynamo

The Paris Observatory began solar observations in 1661. Observation records show very few sunspots for the following 75 years. The few sunspots that were reported were usually located in the Sun's southern hemisphere. They were also traveling much more slowly across the Sun's surface than today's sunspots do. Today's sunspots are evenly distributed between the Sun's two hemispheres.

A British amateur astronomer, Richard Carrington, found that some sunspots near the solar equator rotated faster than those at mid-latitudes. The rotation is 26 days at the equator and about 28 days at a latitude of 45 degrees. By 1911, George Hale, at the Mount Wilson Observatory in Los Angles, observed sunspots came in pairs that resembled the two poles of a magnet oriented east-west on the Sun.

Sunspot Comparisons over 12 Cycles

Cycle	Sunspots	Low-High	Ratio	=1.414
10	98	low	140/98	=1.428
11	140	high		
12	75	low	88/75	=1.173
13	88	high		
14	63	low	105/63	=1.666
15	105	high		
16	78	low	119/78	=1.525
17	119	high		
18	152	low	201/152	=1.322
19	201	high		
20	111	low	165/111	=1.486
21	165	high		
22	158	low	220/158	=1.414
23	220	high		
23	205	minimum	205/158	=1.3
23	237	maximum	237/158	=1.5

Table 1. Comparison of the relationship between the amplitude of the mated pairs.

At the start of a new solar cycle, the first sunspot appears at a high latitude—about 40 degrees—in both hemispheres. They are created closer and closer to the equator as the 11-year cycle continues. As the old cycle dies away along the equator, new sunspots appear at high latitudes, but with opposite magnetic poles compared to the old cycle. This means that it isn't really an 11-year cycle, but rather a 21-year cycle.

A graphical solution

For a graphical solution, we must pair two 11-year cycles to make one longer cycle of 21 years; but which two? We must look for a commonality of patterns that form matched pairs. Through my own studies, I have determined that these pairs are Cycles 10 and 11, 12 and 13, 14 and 15, 16 and 17, 18 and 19, 20 and 21, and 22 and 23. Cycles 1 through 9 don't follow the same pairing pattern, and there may have been some sort of phase and noise changes during those cycles.

By examining these cycle pairs (**Figure 3**), a mathematical commonality emerges. We have a minimum of four rotating vectors: one 11 years long, one 21 years long; one approximately 40 years long, and one approximately 80 years long. For the first two vectors, one is twice the amplitude of the other. This means if

Monthly Sunspot Numbers

Year	Jan	Feb	Mar	Apr	May	Jun	Jul	Aug	Sep	Oct	Nov	Dec
1996											8	9
1997	9	10	12	13	13	15	17	19	21	25	28	34
1998	40	45	52	61	69	78	86	94	104	114	124	132
1999	142	151	155	159	161	165	169	175	177	179	186	191
2000	193	196	199	202	205	206	207	208	207	206	203	196
2001	192	190	190	189	187	186	183	183	181	180	174	170
2002	166	161	156	151	146	142	138	133	130	127	126	122

Table 2. Lower limits of calculations for Cycle 23.

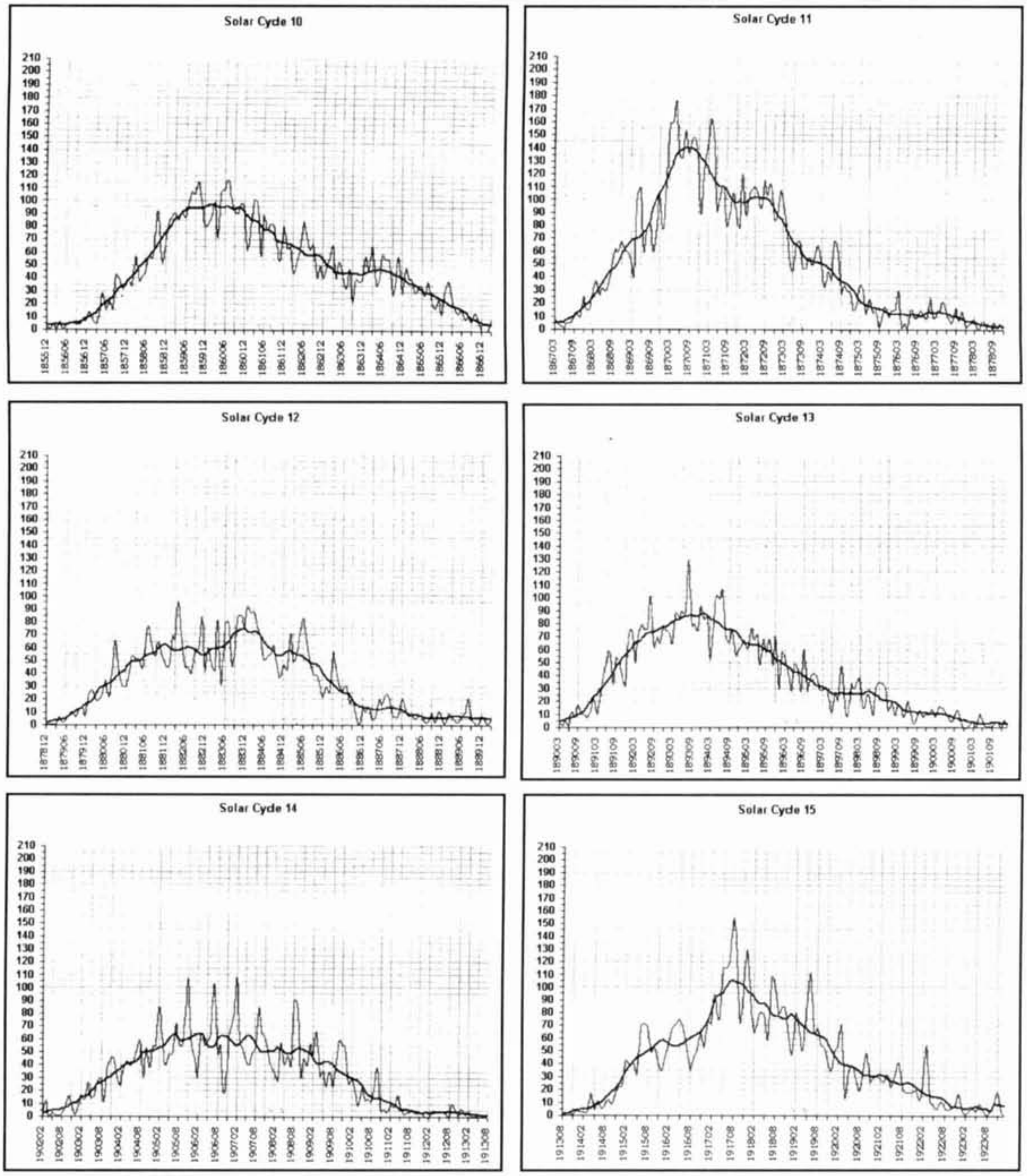


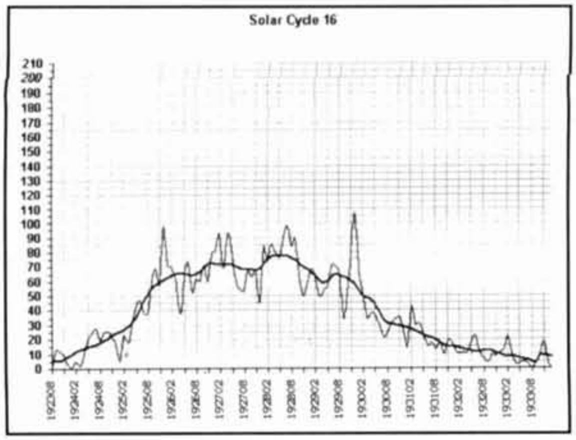
Figure 3. Pairing patterns in solar cycles. Data source: Sunspot Index Center, Brussels, Belgium.

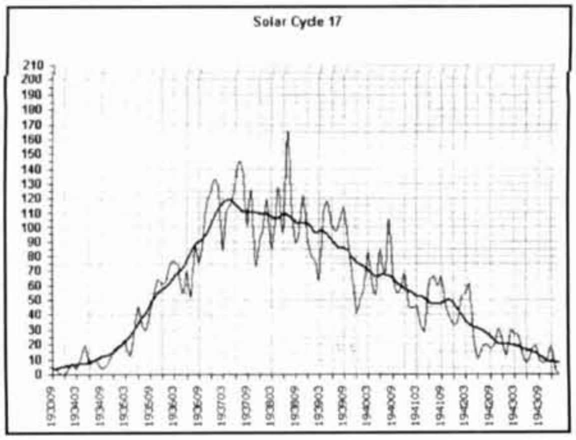
the pattern continues, each pair will contain a relationship with the square root of 2. Thus, the even cycle will be followed by an odd cycle whose amplitude is the square root of 2 larger. **Table 1** provides a comparison of the relationship between the mated pairs' amplitude.

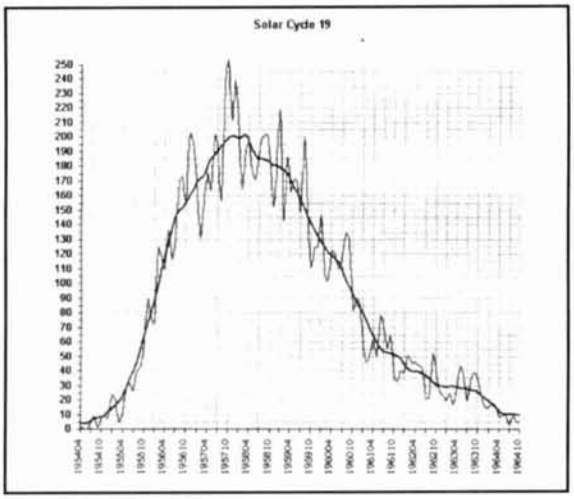
What this means is that we can predict the amplitude of the odd cycles, if we know the maximum amplitude of the even cycles. Cycle 22 is now history, with a maximum amplitude of 158. This means that Cycle 23 should have a maximum amplitude between 1.3 and 1.5 larger than Cycle 22, with a nominal value of 1.4 larger than Cycle 22.

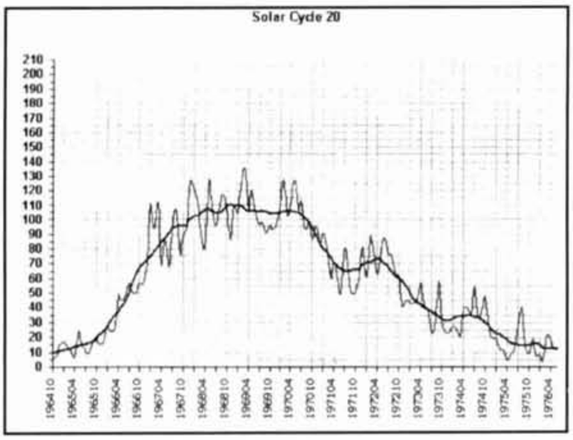
Cycle 23, the greatest in recorded history?

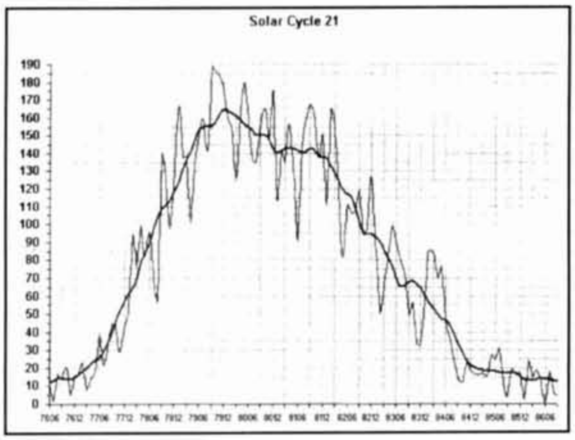
I believe the next sunspot cycle will be the highest in all recorded history! Based on my calculations, Cycle 23's maximum value will











fluctuate from a low of 205 to a maximum of 240, with a nominal value of 220.

Shaping Cycle 23 now that we have the peak value

After you've studied the matched pairs long enough, you'll observe another pattern. The even cycles have lower and broader shapes compared to their odd mates. Many of the even cycles are so broad that they seem to have *two* peaks. This means Cycle 23 (**Figure 4**) should have a shape like that of Cycle 19 because its matched pair, Cycle 18, is similar to Cycle 22. By transferring the shape of Cycle 19 into the time for Cycle 23, and transferring its peak up from 201 to 205, we have a good approximation for the lower limits of our calculations for Cycle 23 (**Table 2**).

This means that hams will experience the best radio propagation in the history of wireless

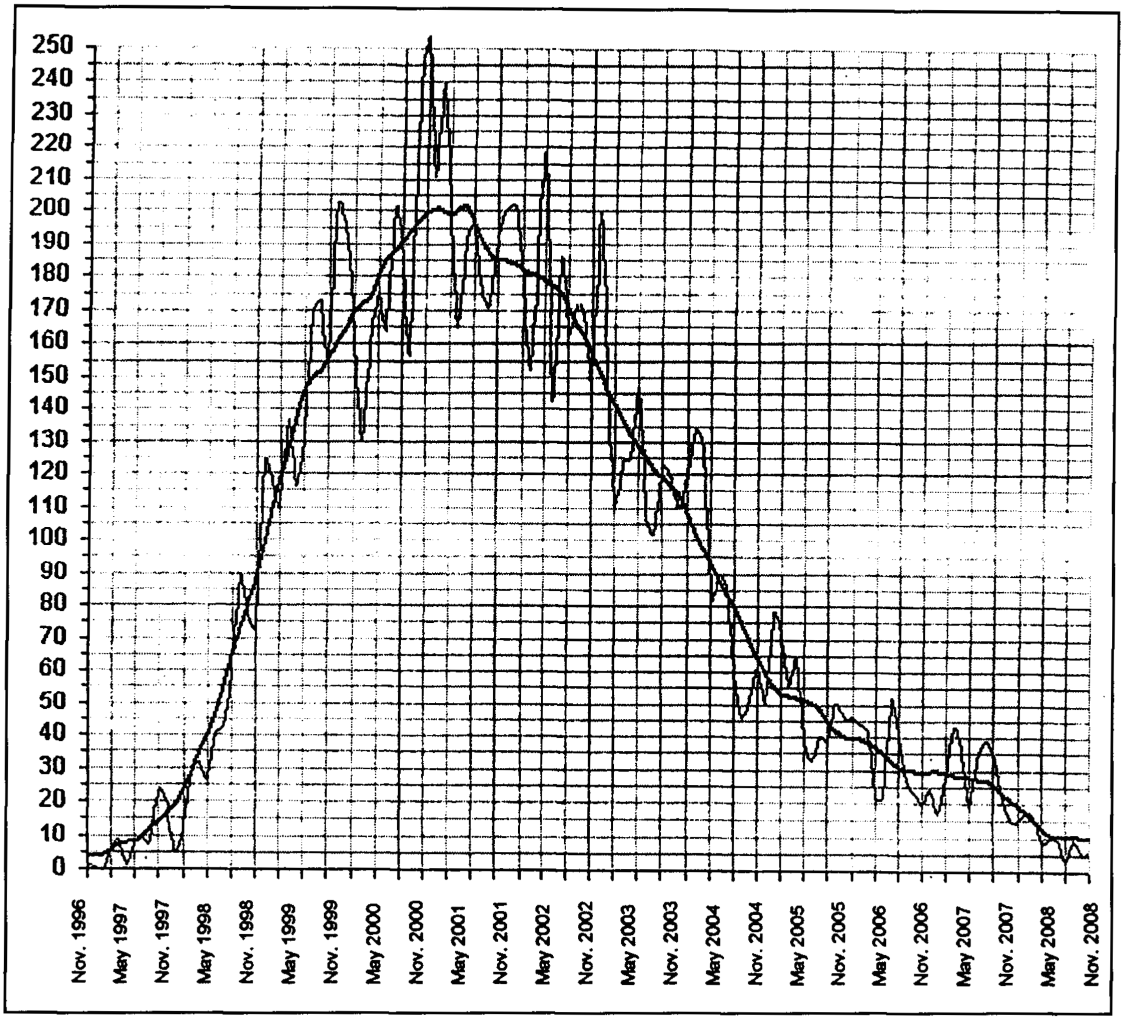


Figure 4. Shaping Solar Cycle 23, which began in November 1996 with a smoothed sunspot number of 8.

communications. Twenty meters, normally the best worldwide DX band, will be replaced by 15 and 10 meters in 1998 and 1999. Even these high HF bands will have to move over, and, for the first time, 6 meters will become the best DX band in 2000 and 2001.

NOAA data

Recently, the NOAA Space Environment Center reported that a panel of solar experts recently fixed the probable date for the recent solar minimum around October 1996, with a possible range from May through December. Panelists believe that Cycle 23 should be a large cycle, although not as large as Cycle 19 in the late 1950s. The experts expect Cycle 23 to peak around March 2000, with a range from June 1999 to January 2001.

Why are these NOAA findings significant? Earlier on, the same group of experts predicted that Cycle 23 would be uninspiring—a real bomb. However, as the maximum of the cycle approaches, these same experts are having to revise their predictions ever upward, leaving the door open for this to be the *highest sunspot cycle EVER*!

Cycle 24 and beyond

At present, there's no really detailed model of how the Sun works. As a result, neither I nor anyone else can say what Cycle 24 will be like. However, once we know its peak, we'll be able to calculate the maximum value for Cycle 25.

A return to winter

During the last 10,000 years, we have known spring, then summer, with only five cool periods. We have gone from hunter-gatherers to farmers, to urbanites. Now, as we face the beginning of the 21st century, summer seems to be coming on warmer than ever; but how long do we have? Winter will surely come. ■

Edited by Peter Bertini, K1ZJH
Senior Technical Editor

Designing a "modern" 6-meter AM rig is something that Rick and I have been tossing around for several months. I felt there would be interest in such a product, based on the number of hams resurrecting old Poycomms, Layfayettes, and "Gooney Birds" (Gonsets). Rick was a bit more skeptical.

Before I could even set pen to paper, or even warm the iron, Rick had a finished model debugged, running and on the air. Perhaps just as well...I am sure his mastery of the QRP arts exceeds mine. His little set rivals, or surpasses, the performance of those early tube clunkers.

After several weeks of operating, Rick finds no shortage of folds to work locally or during opening. An added bonus: Rick reports that 6-meter AM QSOs seem to be more leisurely and enjoyable that the routine hit-and-run "59 FM11 73" contacts offered on SSB.

Perhaps ragchewing is not a lost art. Six meters is a vast band normally begging for activity. Anything—including vintage AM—that gets more folks building projects and on the air and having fun is fair game in our book!

—de K1ZJH

Build the *Nor'easter* 6-meter AM Transceiver

This little radio brings back the nostalgia of AM.

Rick Littlefield, K1BQT

There's been a resurgence of interest in "ancient modulation" lately, due in part to growing legions of antique radio and AM-broadcast equipment collectors. Like CW, amplitude modulation remains a useful and enjoyable mode for those who appreciate its unique qualities. If you enjoy the dulcet tones only AM can deliver, here's a "back-to-the-future" VHF project mixing contemporary design with a touch of nostalgia.

Description

The *Nor'easter* is a kitchen-table construction project that yields a complete VHF-AM transceiver with all the familiar appointments found on old-style VHF rigs. These include a tunable receiver, adjustable squelch, AGC, transmit channel selector, spot switch, built-in speaker, and PTT switching. Although inspired by '60s thinking, construction is rooted in the '90s, with all parts mounted on a small double-sided pc board and the finished project housed in an attractive molded case (**Photo A**). A sensitive single-conversion superhet receiver pulls in weak signals, and the high-level AM-modulated transmitter delivers 5 watts to the antenna (that's 10 watts *input* by Heathkit standards).

Photo A. The finished project housed in an attractive molded case.

Circuit Design

The receiver is a straightforward design using popular homebrew parts. A selective two-section bandpass filter rejects out-of-band signals and low-noise preamp Q1 boosts incoming signals by over 20 dB. Mixer U1 functions as both an active DBM and as a varactor-tuned VFO running at 39 MHz to down-convert 6-meter signals to the 10.7-MHz IF. The receiver's message-channel passband is established by cascaded crystal filters FL1, FL2 at around 15 kHz. Preamplifier Q2 boosts mixer output and overrides insertion loss of the tandem filters. IF amplifier U2 delivers 45 dB additional gain with an AGC range approaching 60 dB. AM detector D1 recovers audio plus a DC level for the AGC system. U3 is a four-section opamp providing AF preamplification, AGC drive, and a comparitor-driven squelch. AF-output stage U4 is a stock LM386 set up for minimum gain.

Giving credit where it's due, the receiver configuration was inspired by a Ramsey Electronics aircraft monitor I built from a kit about 10 years ago. Some redesign yielded a 50-MHz superhet with improved sensitivity and significantly sharper selectivity.

On the transmit side, third-overtone crystal oscillator Q3 is diode-switched to select

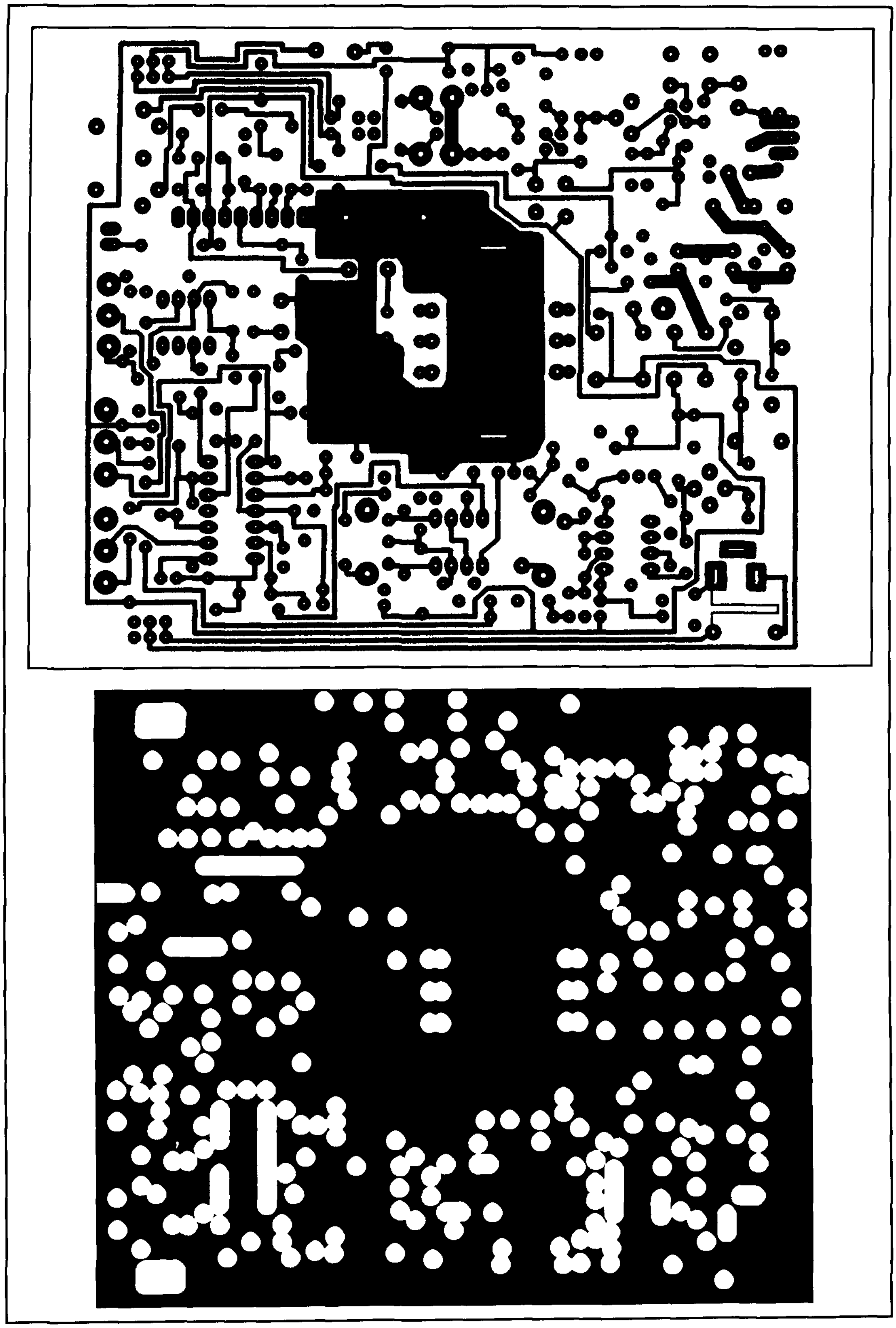


Figure 1. Printed circuit board art; (A) trace and (B) component side.

between two transmit channels (50.25 and 50.4 MHz). Q3 may be activated in receive mode to "spot" transmit channels with the radio's tunable receiver. Driver Q4 and PA Q5 operate in class C, and both stages are modulated by U5.

U5 is a Phillips 4.5-watt audio preamp/power-amp chip. Modulation transformer T2 was parted out from a dead Uniden CB-radio, but other CB modulation transformers may be used. RF output from Q5 is matched by an L-network

Parts List

Capacitors (C)

1	2.7-pF disc ceramic
1	10-pF disc or monolithic
1	15-pF disc or monolithic
1	33-pF disc or monolithic
2	47-pF disc or monolithic
7	60-pF MuRata 50-V trimcap or equiv.
2	68-pF NPO disc or NPO monolithic
2	75-pF 100-V s.m. or monolithic (68-pF if 75-pF unavailable)
1	100-pF NPO disc or NPO monolithic
2	150-pF 100-V s.m. or monolithic (do not substitute ceramics in filter)
1	220-pF disc or monolithic
1	0.001-μF disc
1	0.0022-μF ceramic or mylar
15	0.01-μF disc
12	0.1-μF disc
4	1-μF electrolytic
1	10-μF electrolytic
1	22-μF electrolytic
4	100-μF electrolytic
1	470-μF electrolytic
1	1500- or 2200-μF electrolytic

Diodes (D)

2	1N5235B (6.8 V) zener
1	1N5239B (9.0 V) zener
1	MV2104 varactor
6	1N914 or 1N4148
1	1N34
1	1N4001

Filters (FL)

2	Mouser 520-107-15B, four-section 25-kHz spacing 10.7-MHz crystal filter.

Jacks (J)

1	BNC female chassis mounting (antenna)
1	2.1 x 5.5 power jack, pc mtg, Mouser 163-5004 or equiv.
1	5-pin DIN, female pc mtg, Mouser 161-0504 or equiv.

Relay (K)

1	DPDT sealed relay, 12-volt coil, Mouser 526-R40-11D2-12 or equiv.

Resistors (R): 1/4-watt unless noted

1	10
1	15
2	22
2	100
1	220
1	470
6	1K
3	2.2K
1	4.7K
1	5.6K

Resistors (Cont.)

1	8.2K
6	10K
1	22K
2	33K
5	47K
3	100K
1	330K
2	1M

Chokes (RFC)

1	0.22 μH molded
2	2.7 μH molded
1	10 μH molded
1	22 μH molded
1	100 μH molded

Coils (L)

2	7 turns #22, 0.25" ID x 0.4" (L1, L2)
1	Coilcraft 143-10J12S 10-1/2 turns shielded, 0.42 μH (L3)
2	4 turns #28 bifilar on T25-6 (L4, L5)
1	5 turns #22 on T37-6 (L6)
4	9 turns #22 on T37-12 (L7–L10)

Switches (Sw)

1	SPST miniature, momentary contact, RS 275-1571
2	DPDT miniature, push on-push off, Mouser 612MTH22

Transformers (T)

1	10-mm 10.7-MHz IF transformer, Toko 421F122 or equiv.
1	Modulation Transformer (Uniden TF-177 or equiv.)

Transistors (Q)

1	MRF901	TA 900-5584
1	2N3904	TA 900-5456
1	2N3906	TA 900-5457
1	2N2222A	TA 900-5428
1	2N5109	TA 900-5451
1	2SC2166	RF Parts

Integrated Circuits (U)

1	NE602, SA602, or NE612
1	MC1350P
1	LM324 or equiv. quad op-amp
1	LM386
1	TDA1015 (Philips)

Crystals (Y)

1	50.25-MHz 3rd-overtone GP (ICM 471270)
1	50.40-MHz 3rd-overtone GP (ICM 471270)

Case

1	RadioShack #270-214

TA = Tech America

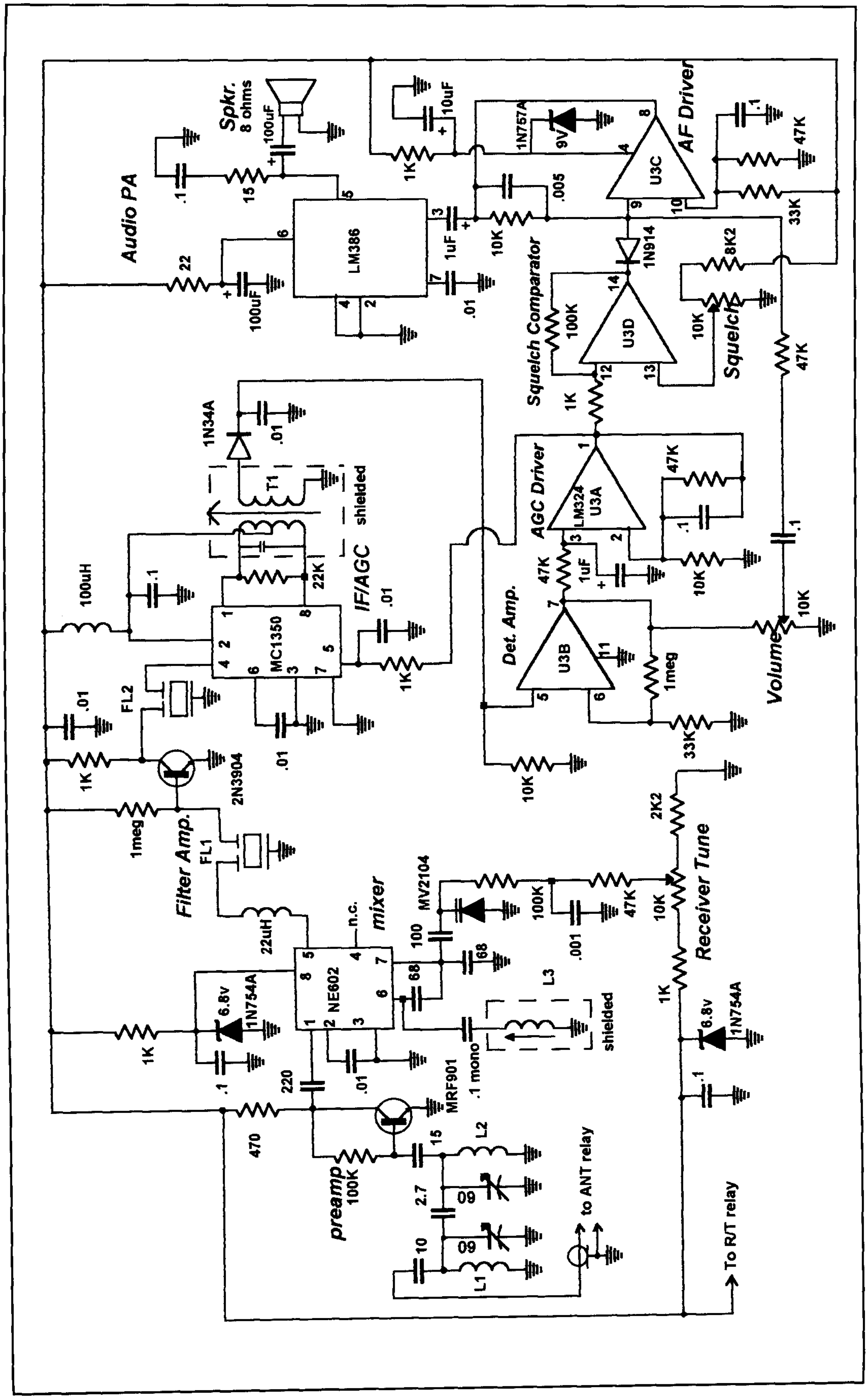


Figure 2. Schematic diagram, receiver section.

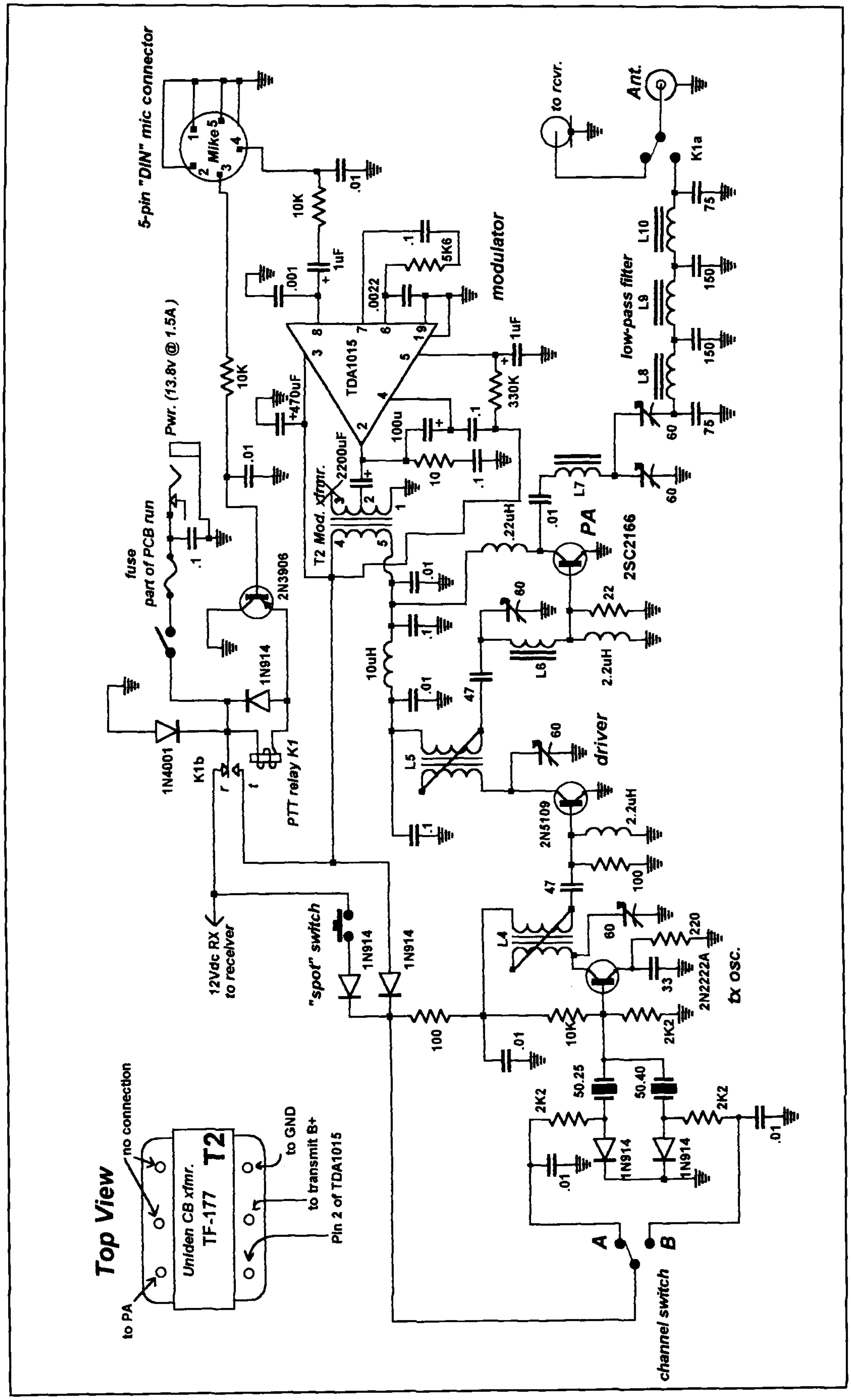


Figure 3. Schematic diagram, transmitter and control.

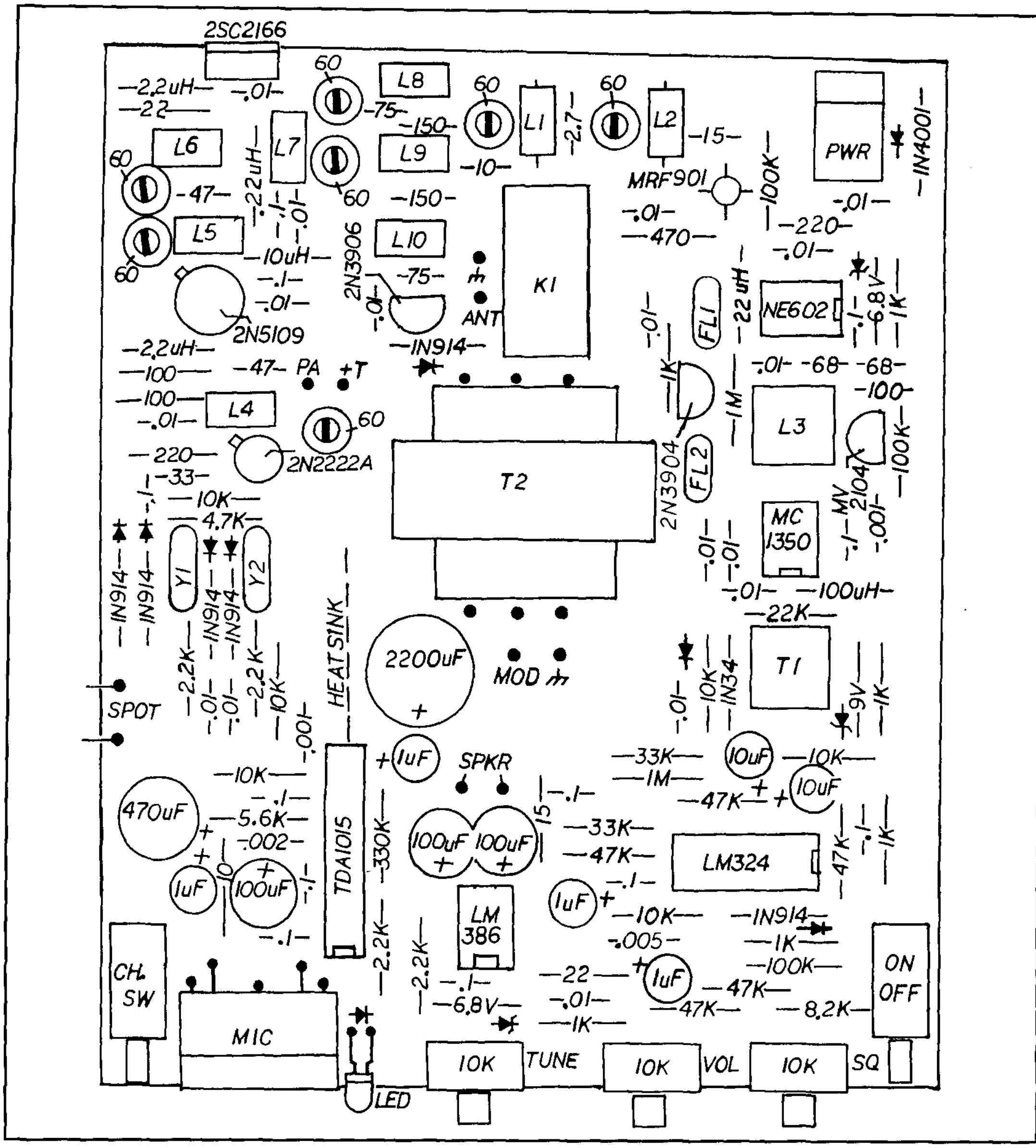


Figure 4. Parts placement diagram.

into a three-section, 50-ohm lowpass filter that reduces spurs and harmonics by -60 dBc or greater. DPDT relay K1 switches the antenna line and provides +T and +R voltages for various transceiver circuits. Switching transistor Q3 activates K1 when the microphone PTT line makes ground contact.

Construction

Virtually all parts, including switches and controls, are mounted on the radio's 3.9 by 4.6-inch double-sided pc board. Printed circuit art is provided in **Figures 1A** and **1B**, and boards are available from FAR Circuits.* To populate the board, follow schematic diagram (**Figures 2** and **3**) and parts-placement diagram (**Figure 4**). I suggest mounting L3, T1, and the ICs first, before installing resistors and capacitors. Install handwound coils and pots last, as these are more easily damaged by handling. Because the board isn't plated through, be sure to solder *all* grounded connections on both sides of the board, stripping the flash off ceramic bypass leads to minimize lead length.

Although a shielded cabinet is always more desirable for radio projects, the *Nor'easter* was laid out to fit into a RadioShack plastic modem case. You must replace the box's rear plastic panel with an aluminum duplicate to provide an efficient heatsink surface for PA transistor Q9. The front and rear panels are secured to the pc board by controls, components, and solder lugs. Panels slot-load into the clamshell-style case

*Far Circuits, 18N640 Field Court, Dundee, Illinois 60118. Boards are available for $13 each plus $1.50 shipping for two boards.

and support the pc board inside without need for internal mounting screws. **Figure 5** provides front and back panel templates, along with a pattern for the modulator's 1/32-inch thick copper heatsink. **Photo B** shows the final assembly.

Part substitutions are generally okay, but the layout is tight and only small components should be used. Avoid substituting disc-ceramic capacitors in the transmitter's output filter—multilayer or miniature silver-mica caps work much better and result in less power loss. I used four-section 10.7-MHz crystal filters, but lower-cost, two-section 10.7-MHz crystal filters may be substituted with some sacrifice in strong-signal selectivity. (Both types are available from Mouser Electronics at $6.60 and $3.41 each, respectively.) Although more sophisticated matching of the crystal filter I/O ports might improve passband shaping slightly, the circuit shown yields acceptable AM performance.

To mount the TDA1015 IC heatsink, attach it to the IC for positioning and tack-solder to the pc board. To complete the installation, unscrew the heatsink from the IC to prevent overheating, and run a bead of solder down one side to thermally bond the heatsink to the pc board. The *Nor'easter*'s modulation transformer came from a defunk Uniden CB, but other types will work, too. I easily obtained several pull-outs through fellow hams who operate radio service shops. Look for a later-model CB that uses a monolithic audio amplifier in the modulator—this will provide the best match for the TDA-1015. If a push-pull type is all you can locate, feed the center tap rather than the entire primary on the modulator side. Not all transformers have the same pinout, so jumper connections are provided on the pc board to facilitate various lead configurations. Connections are shown for the Uniden TF-177.

Tune-up Procedure

Begin receiver tune-up with ballpark alignment of the receiver oscillator. Set the receiver tuning pot fully counterclockwise and lightly couple a counter to pin 7 of U1. Adjust L3 for 39.3 MHz to obtain a receive frequency of 50.0 MHz. The VFO tuning range will be about 600 kHz with the parts specified. Because the counter will probably load down the oscillator tank circuit slightly, you'll need to use a signal generator for final receiver calibration. Once the VFO is tuned, open the squelch and increase volume for an audible background hiss, adjusting T1 for maximum noise in the speaker. Now, connect a weak 50.25-MHz AM-modulated signal source to the antenna jack and tune C2, C4, and T1 for maximum sensitivity. When properly adjusted, the receiver should detect a 0.3-µV carrier and recover a 1-µV AM signal at 12-dB SINAD.

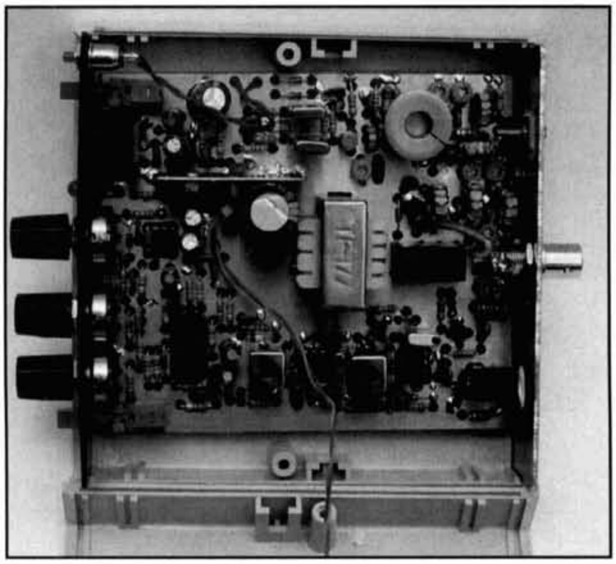


Photo B. The final assembly.

To tune the transmit oscillator, first press the spot switch and peak the oscillator trimcap for maximum signal (use a scope, RF voltmeter, or external receiver). Next, connect the radio to a 50-ohm dummy load through a RF wattmeter. Key the mic and carefully adjust the driver and PA trimcaps for maximum output. When tuned, confirm that each trim cap exhibits *two equal peaks* with each 360-degree revolution, indicating that resonance falls within the trimmer's range. If needed, spread or compress turns on the associated inductor to achieve this response.

Finally, sample the transmitter signal and view the modulation pattern on a 50-MHz scope. Touch up driver and PA tuning with modulation applied, adjusting for maximum p-p voltage. This will optimize the driver and PA for highest peak envelope power (PEP) and improve modulation depth. When done, you should obtain 100 percent modulation on peaks without driving the modulator into saturation, and there should be no visible downward modulation on the RF wattmeter.

Operation

In keeping with good VHF engineering practice, your antenna should exhibit a VSWR of 2:1 or less—the lower, the better. Six meters shares spectrum with a lot of home-entertainment media, and the *Nor'easter*'s built-in three-section low-pass filter suppresses FM-BCI and TVI better when properly terminated into a low VSWR load.

Almost any microphone should work; the input impedance of the TM1015 op-amp preamplifier stage is fairly high (over 20K). When driven with an AF signal generator, modulation response measured flat past 8 kHz, with gradual LF rolloff below 150 Hz, due mostly to the natural cut-off frequency of the small CB

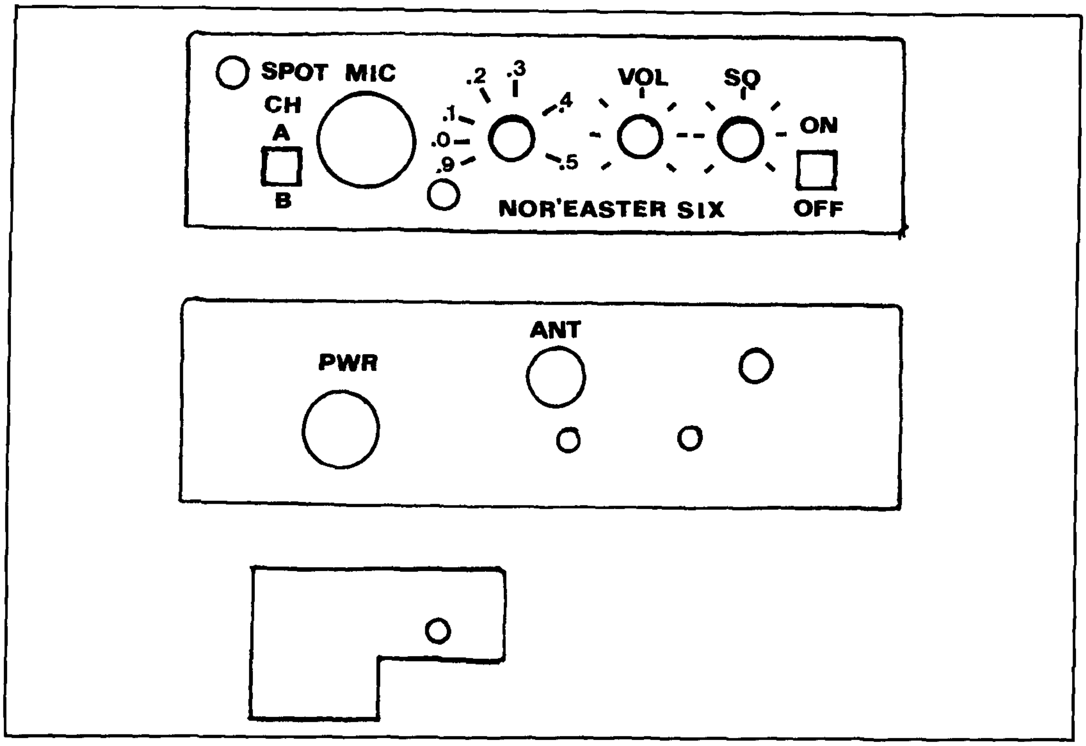


Figure 5. Panel and heatsink templates.

modulation transformer. With the right microphone, the *Nor'easter* delivers near-broadcast quality audio.

There's no mic-gain control, so it's possible to overdrive the speech amp if you crowd the microphone or use a cartridge with unusually high output. Overdriving typically results in over-modulation of the transmitter and/or in distortion of the audio waveform due to clipping in the modulator's output stage. If splatter and distortion are evident, increase the value of the 5.6-k coupling resistor between the preamp and modulator stages (pins 7 and 6) to reduce mic gain.

In receive mode, the relatively wide IF filters provide bandwidth characteristics typical of many vintage 6-meter receivers. This permits easy tuning with the radio's small front-panel knob, and you won't have to chase crystal-controlled stations operating a couple kHz off frequency or retune stations because of thermal drift from the 39-MHz VFO. On the negative side, some weak-signal sensitivity is undoubtedly sacrificed due to increased noise energy in the IF passband. However, AM signals under 1 μV should be readable under quiet band conditions and receiver noise should be well under the 50-MHz atmospheric noise floor present at most locations.

Cascaded filters provide relatively steep skirts to prevent strong off-channel SSB or packet signals from interfering with the AM channels. These filters also facilitate locating transmit channels with the spot switch. To spot channels accurately, turn the squelch off and the volume up; there's no BFO, so you'll need to hear the detector background noise. Once you are centered on the signal, you can reduce volume and set the squelch.

Conclusion

While ironing out the bugs in my prototype, it occurred to me there may not *be* any AM activity left on six! Like the Maytag repairman's telephone, I began to envision the *Nor'easter* sitting idle, week after week, without so much as a crackle to break the squelch. Fortunately, when I connected the antenna, the radio came to life almost immediately with a gathering of New Hampshire locals who meet daily on 50.25 MHz. In addition, over the next couple weeks I managed several QSOs on 50.4 MHz, including a number of single-hop *E*s contacts during two band openings. I also heard local activity on 50.51, but didn't have a transmit crystal for that channel. Six-meter AM has never been a hot-bed of activity in New England, but it surely isn't dead.

As for the future of 6-meter AM, the sunspot cycle is finally rising and some experts predict a rapid ascent peaking well over 200. If these optimistic forecasts are correct, little radios like the *Nor'easter* could rack up some fantastic worldwide frequent-flyer mileage (deja 1958) in the months and years to come! ■

EDITORIAL *(from page 4)*

band and retransmitted FM we use to relieve congestion. However, in some segments of some of our bands, congestion is anything but a problem. So much so, that we may even loose our allocations. We stand on the threshold of tremendous technological changes. As we look to the future and look to the past, it's a good time to reflect both upon what we had and what we might like to have (such as good-sounding audio) in the years to come. Once again, looking both ways.

A correction

Dean Shutt, AL7CR, noticed an error in Reference 18 in the article "Source Impedance of HF Tuned Power Amplifiers and the Conjugate Match," (Fall 1997, page 25). The date of publication for that reference should have been listed as December 1994, not December 1995. Thanks, Dean!

A new column

As you browse through the pages, I'd like you to take a look at our new column "Science News." You'll find it on page 41. It's written by Douglas Page, a science writer from Redondo Beach, California, who will offer short, technically oriented news features on the latest developments in the communications and electronics industries, along with anything else he thinks we might find interesting. in this issue, Doug reports on peel-off semiconductors and a low-cost electronically scanned millimeter wave antenna. Welcome aboard, Doug!

Catch me online

As I mentioned last issue, you can reach me online at <ka1stc@aol.com>. Not only are you assured (usually) a timely answer to your query, but I also really appreciate all your comments (both positive or negative). Reader input provides vital information regarding the direction the *Quarterly* should take in future issues, and it's important to work together to make *Communications Quarterly* the best-read amateur radio journal around.

Terry Littlefield, KA1STC
Editor

PRODUCT INFORMATION

Getting Inside Wire Insulation

Save troubleshooting time by taking wire measurements without having to find the ends. Jensen Tools has a device designed to penetrate insulation of any wire from 30 AWG to 1/2 inch in diameter. Adjustable pressure control prevents damage to even the finest wire wrap wires. After the needle is removed, the insulation "heals" the tiny hole made by the insertion. The 8-inch probe length gets into hard-to-reach places.

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AD Has Bonus Issues of "Ask the Applications Engineer"

In celebration of the 30th anniversary of *Analog Dialogue*, Analog Devices, Inc. has bundled with the 1997 spring/summer edition a 60-page bonus issue called "Ask the Applications Engineer." An *Analog Dialogue* standard feature since 1988, "Ask the Applications Engineer," answers frequently asked questions as documented by Analog Devices' applications engineers.

Copies of publications or product information are accessible through Analog Devices' Web site at <http://www.analog.com/publications/magazines/dialogue/dialog.html> or by calling the HELP line at (800) 262-5643 in North America and (617)-937-1428, elsewhere.

The PowerPort PowerSafe

The PowerPort PowerSafe from Cutting Edge Enterprises provides everything you need for a 75- to 200-amp uninterrupted power supply. There are three DC models designed for light, medium, or heavy use. All come with a heavy duty, vented battery enclosure suitable for use in the home, triple port automotive cigarette outlets for DC use, and fully automatic charger. The deluxe model provides 500 watts peak AC power (300 watts continuous). Dimensions are 18 x 9.5 x 10.5 inches.

Prices for the DC models range from $66 to $168. The deluxe AC/DC model costs $230. For extended field operations, Cutting Edge offers a full line of accessories including: solar cells, lights, DC extension cords, etc.

For more information contact Cutting Edge Enterprises, 1803 Mission Street, Suite #546, Santa Cruz, California 95060. Phone: (800) 206-0115. E-mail: <cutedgent@aol.com>.

tance). When the plate voltage is small, the plate current is large (low value of resistance), and the instantaneous resistance of the DC circuit is:

$$r_T = \frac{E_{BB} \bullet (1 - k \sin(\Theta))}{I_{BB} \bullet (1 + k \sin(\Theta))} \quad ; 0 \le k < 1$$

This is a nonlinear function of Θ; nevertheless, it causes a sine wave of AC voltage on the plate and a sine wave of current through this linear tube. The AC resistance R_{AC} is the ratio of AC plate voltage to AC plate current.

The AC resistance R_{AC} of the tube is also seen by referring to line AB in **Figure 2**:

$$R_{AC} = \frac{e_{p1\max} - e_{p1\min}}{i_{p1\max} - i_{p1\min}} = R_L$$

The interesting thing about this equation, based on the geometry of **Figure 2**, is that the AC sine wave voltage across the tube (a.k.a. variable resistor) and the load are identical (the tube and load are in parallel for AC voltage). Also, the AC sine wave current in the tube and the load are the same (the tube and load are in series for AC current). Therefore, by Ohm's law R_L, the AC resistance of the load and R_{AC}, the AC resistance of the tube, for the sine wave signal are identical. If we change R_L we also change RAC an identical amount. The tube is essentially a load resistor R_{AC} for the external test generator. This is what is being measured in this situation. And a measurement of R_{AC} is equivalent to a measurement of R_L. If the test generator resistance is 50 ohms at the secondary of a stepdown transformer, then the SWR (with respect to 50 ohms) that the test generator sees, looking into the amplifier output, is always 1:1 for any value of the turns ratio of the transformer, as long as the tube is perfectly linear. This is a significant observation for the ideal class A amplifier and also for the article being discussed.

• The difference between these two methods is that, in the first case, plate voltage changes cause the change in plate current (this is rp) and in the second case the grid variations dominate the plate current changes (this is gm).

3. We now consider the following question for this linear amplifier. Having determined RAC, the tube's AC resistance when it is used as a "load" for the test generator that is connected to the output, is this resistance also the "generator" source resistance when the tube is operating in its normal mode as an "amplifier?" The answer is no. In this mode, the tube's AC plate current is the constant value gm • eg and the plate resistance rp (which is in parallel with

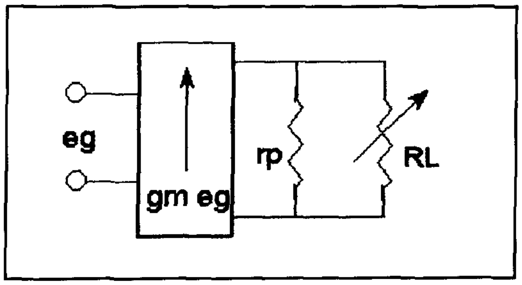

Figure 3. Linear generator and variable load resistance.

RL) is a dynamic (non-dissipative) AC resistance from plate to ground (see **Figure 3**). Under linear small-signal (that is, non-saturating and non-limiting) operating conditions, maximum power output occurs (if only RL is varied) when RL = rp. In **Figure 2** this means that the slope 1/RL1 would be the same as 1/rp, the slope of the constant VGG line, except that it is the negative of this slope. At this point the load power, which is the product:

$$P_{out} = \frac{1}{8} \bullet (e_{p\max} - e_{p\min}) \bullet (i_{p\max} - i_{p\min})$$

is a maximum. All vacuum tube text books and all of their mathematical analyses regarding small-signal class A amplifiers throughout history agree on this issue. Since maximum power out occurs when RL = rp then, by definition, rp is the generator source resistance. And, to repeat, this is not what the test generator at the output is measuring when the grid is being driven.

4. One thing that needs to be considered is that if an actual real-world tube has a certain VBB, and if the tube dissipation has an allowable maximum value and if the maximum linear value of instantaneous plate current i(MAX) is limited by the design of the tube, then the load resistance R_L for the absolute maximum linear power output is determined by VBB and i(MAX). If R_L is smaller than this value, the limitation on i(MAX) will produce a reduction in peak output power $i(MAX)^2 R_L$. This absolute max power output has nothing to do with internal dynamic resistance rp. A pentode (rp = 100,000 ohms) and a triode (rp = 100 ohms) have virtually the same capability, the same AC resistance and the same limitations, except possibly for minor differences due to the presence of the additional grid(s). Conjugate match cannot just simply be "declared" in this situation.

5. We have considered a counter-example to the conjugate match proposal in an ideal linear class A amplifier. If the theory is true, then it must work for all examples and at all frequencies. The class A amplifier is not a "special

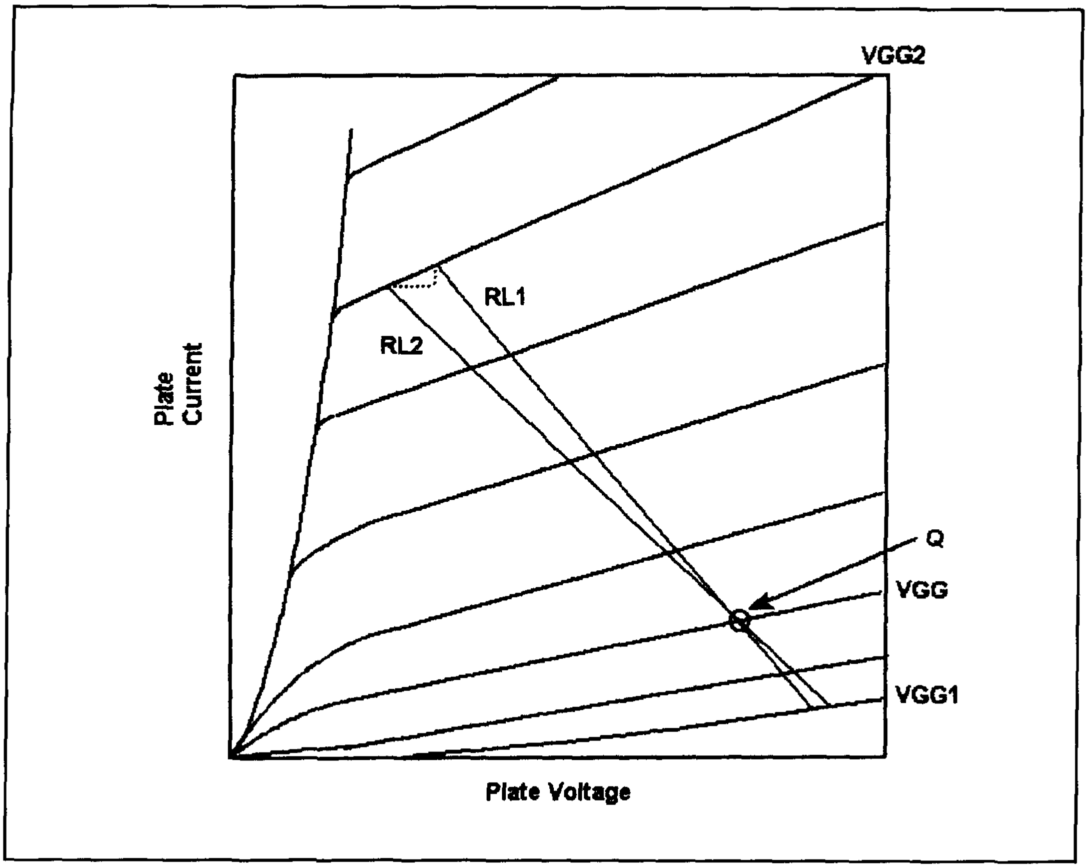

Figure 4. Typical power tube plate curves in class AB operation.

case" that can be excluded. And when considered as a load to an external test generator the tube's AC resistance RAC is one thing, but when considered as a generator its dynamic output resistance rp can be (and usually is) quite different.

6. There is another way to measure the dynamic output resistance of the linear class A amplifier. Refer again to **Figure 2**. If the peak-to-peak grid drive signal is held constant between VGG1 and VGG2, and if the load resistance (load line) is changed slightly, the geometry of **Figure 2** shows how the rp of the tube is determined. The slopes of the VGG1 and VGG2 lines should be the same, so that rp = rp1 = rp2. This method was described in **Reference 8**, written by myself, where it was mentioned in conjunction with the linear class A amplifier. It can also be used, with caveats, in amplifiers other than class A. **Figure 4** is the plate curves for a typical power tube operating in class AB. In this case rp varies from one end of the load line to the other. The *small* change of RL produces a one-cycle *average* value of rp(ave) for the particular load line and grid excursion that is shown. But, unlike **Figure 2**, rp(ave) depends greatly on the operating point Q, load RL and drive level (VGG1 and VGG2). Non-trivial changes in these require a new rp(ave) measurement. And if the amplifier is driven and loaded in a manner that is not correct, from an SSB linearity standpoint, it is possible to get a wide variety of answers. Another thing is that the selectivity of a tuned plate circuit means that only the fundamental component of the tube's dynamic output resistance is being measured (the non-linear tube resistance has harmonic components). This leads to complications that are beyond the scope of this letter.

7. To test some of the main conclusions of the previous sections, I measured my ancient but well-maintained SB400, which uses parallel 6146s (no negative feedback), using the resistance variation method. I used a two-tone test at 100 W PEP at 3.8 MHz. A spectrum analyzer and an oscilloscope looked at the two-tone RF signal to make sure there was zero flat-topping. The resistive load was switched between 51.0 and 46.15 ohms. RF voltage was measured with my Boonton 92A analog meter with a 100:1 voltage divider probe. At 100 W PEP, the third order intermod products were 30 dB below PEP. I then reduced the power to 50 W PEP and then 25 W PEP. The denominator of the equation (**Equation 4** in Belrose) involves the difference of two similar numbers, so I repeat-

ed everything several times to get some confidence in the readings. The average results were as follows:

100 W PEP 147 ohms
50 W PEP 460 ohms
25 W PEP 325 ohms

When the amplifier was loaded to maximum possible output, single tone class C with some grid current, the reading dropped to about 60 ohms and the power output increased about 0.5 dB. These numbers are approximate because of instrumentation limitations, but I think they are not off by more than 25 percent, based on repeatability. My conclusion was that the output resistance varies substantially with PEP and that for a speech waveform the average output resistance is not 50 ohms. An exception would be the KWM-2 and other types that use negative feedback circuitry.

8. The article describes how maximum output occurs when the VSWR, looking back into the amplifier with a test generator, is 1:1. I won't be able to analyze this method in sufficient detail to say exactly what is happening, but I am sure that injecting a test signal into the output does not actually measure the dynamic output resistance, for the reasons that I have mentioned in considerable detail. I do think that if the tubes are driven and loaded to their extremities it is possible to get a 50-ohm measurement due to the nonlinearities involved. That occurred in the tests that I reported in Part 7 of this letter to the editor. For example, at plate voltage min (plate current max), the tube resistance can reach some very low values. But I do not agree that this *proves* an unequivocal "conjugate" match between the PA dynamic output resistance and a 50-ohm load, as that term is usually understood.

9. I believe that the concern about PA conjugate match is for the most part unnecessary. Under some conditions, the value of the output resistance may be a significant interface parameter, for example in lowpass filter design or possibly antenna tuner design or for interfacing with another amplifier. It has no influence on load VSWR. In most situations its actual value is seldom considered to be important, except in certain special systems. I believe the best way to measure dynamic output resistance of class AB, B, and C amplifiers is the resistance change method mentioned in Part 6 of this letter. I do not think that injecting a reverse test signal into the output port is the right way.

William E. Sabin, WØIYH
Cedar Rapids, Iowa

JFET discussion missed important point

Dear Editor:

Parker Cope's discussion on JFET biasing ("JFET Simplified," Fall 1997) missed an important point: the operating point selection for linear design. His article leaves open the question of where to operate the JFET for reasons other than the best gain. As a rule, I operate the JFET at about $0.6 I_{dss}$. If you look at the V_{gs} versus I_d curve, you can see this linear region. For good linear design, set I_d near $0.6 I_{dss}$. To find I_{dss} if it is not available, use Parker's suggested test setup.

If you plug the value $I_d = 0.6 I_{dss}$ into Parker's equations, out skips a lot of analysis simplifications.

M.A. (Mac) Chapman, KI6BP
Oceanside, California

Reader calls Fall 1997 issue outstanding

Dear Editor:

I have been reading *Communications Quarterly* for quite a few years rather faithfully, and it has helped me grow technically.

The current issue is outstanding; it is deep, it has breadth within your charter, it is lively and interesting. At no time does it reach out in technical jargon that would confuse the reader who wishes to learn, but does not have a focused education. I really want to applaud your effort, and the achievement that this issue represents. You make a valuable contribution, and the magazine is well worth the price.

Also, I just wanted to send a happy message. My wishes and thanks for doing a good job.

Arthur Westneat, W1AM
Durham, New Hampshire

Praise for K1ZJH's 80-dB log amp for spectrum analyzers

Hi Peter:

I just wanted to let you know how much I have used and enjoyed your article on the 80-dB log amp in the Fall 1996 *Communications Quarterly*. I did a Web search on your callsign and found you here. I ordered the pc board from FAR Circuits and have built it up. I have only used the mixer/filter portion of it so far because the AD606 from Newark is over $50. I have connected a Phillips TDA1576 chip to the 10.7-MHz output and used the RSSI pin for the log output. This chip has over 80 dB of dynamic range, but I'm sure it is not as accurate as the AD606. It works okay for now.

Your tip on using a TEK x/y monitor is a great idea. I picked up a TEK 608 at the West Palm Beach hamfest, it is far superior to using my Leader oscilloscope! I have been playing with homebrew spectrum analyzers for several years now and your article has taken what was a toy and made it a very useful piece of test equipment.

Thanks again for your great article! If you have any more tips regarding improvement of homebrew SAs, I would sure be interested. Also, if you have any suggestions on where to get the AD606 at a reasonable price, I would be interested. Keep the excellent articles coming!

Tim Heffield, N4IFP
Sunrise, Florida

On traveling wave dipoles

Dear Editor:

I enjoyed reading Richard Formato's article on traveling wave dipoles (*Communications Quarterly*, Summer 1997). An antenna which has dissipative elements added to improve its efficiency is sure to get some attention.

Often the same phenomenon can be viewed from several perspectives. In this case, the data begs to be examined as a colinear resonant dipole, whose segments are isolated by resistors. The small resistors near the feedpoint, have a lesser effect and thus combine several segments into one, while the larger resistors nearer the end serve to detach the elements further out into parasitic resonators. This would explain the periodic shape of the graphs with respect to frequency.

On a more practical note, while the topic makes interesting reading. I haven't come up with a practical use for it. The range of frequencies where the gain is high and the SWR low seems to start at two or three times the original dipole resonant frequency. I suspect that the materials required to make a 22-meter long, 10-centimeter thick dipole with dozens of segments and resistors could make a fairly good antenna using other topologies in equal or less space. Its broadbanded nature would not be

ANTENNA BOOKS

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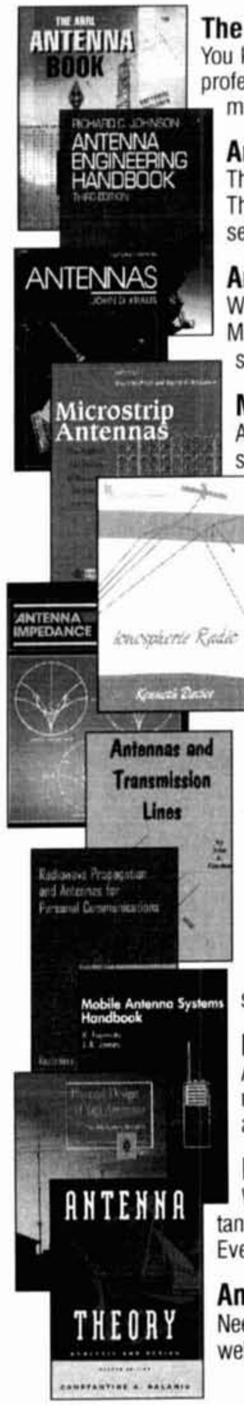

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CRESTONE TECHNICAL BOOKS

We accept VISA, Master Card, American Express useful in very many transmitting applications, and, for HF receiving, its electronics can generally easily make up for significant antenna gain deficiencies. A scaled-down version might be useful in some extremely wideband spread spectrum applications at UHF.

The author did whet my appetite by saying that reactive elements could be added as well. I would have liked to see the results of doing so, although I can understand the complexity of adding dozens of resistors and capacitors to the design.

Wilton Helm, WT6C
via Compuserve

A small correction

Dear Editor:

I got the article ("Diplexers," Fall 1997, page 19); it looks great! The only problem was in **Figure 3**. The scale on the right side of the graph for the normalized imaginary part of Yin is missing. It may confuse some readers, but most should figure out what the scale should be from the reading. Other than that, it really looks good.

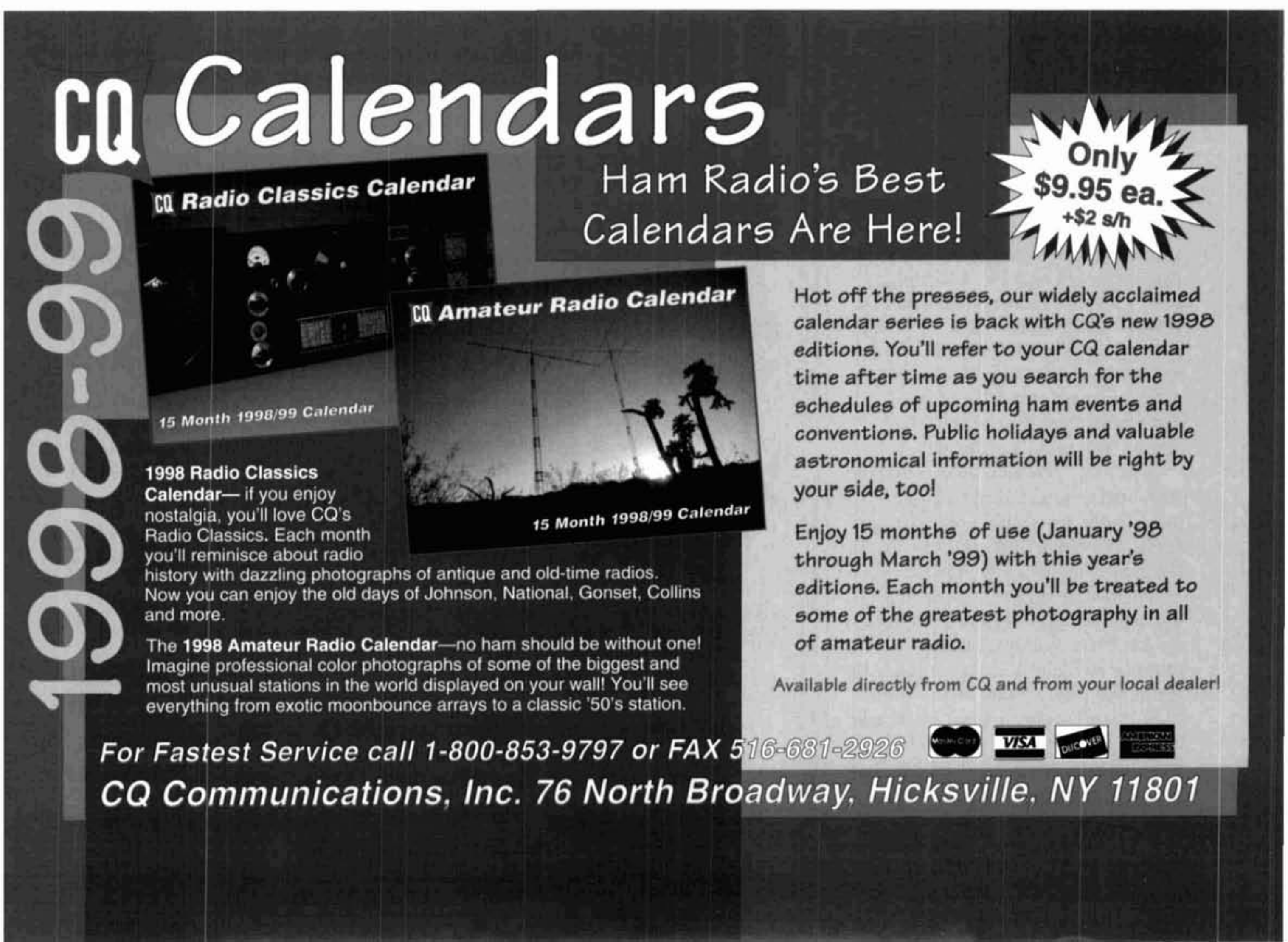
Tom Cefalo, WA1SPI
Winchester, Massachusetts

Last minute arrivals

Dear Editor:

The article on the conjugate match in the Fall 1997 issue was interesting and I learned some things from it. It also started me thinking. Almost all Class B RF power amplifiers these days are used to amplify SSB signals. The power output is varying from zero to maximum. Source resistance varies also. Thus, it would seem that a conjugate match is obtained only at some specific power output.

Modern solid-state transceivers do not have tank circuits; they have push-pull output stages with broadband transformers. They are thus the same as Class B audio amplifiers, about which much has been written over the past 65 years. I reviewed the sections on Class B in the few texts I still have available to me. They all gave formu-



las for the proper load impedance, but only at maximum power output.

However, in the original article by Loy Barton of RCA in the November 1931 issue of *QST*, he gives the formula for load resistance at maximum power, but adds that it loses accuracy at lower powers.

In my 60 years of hamming, I have built a number of Class B and Class C amplifiers and always got the power output and efficiency I expected and never wondered whether or not I had a conjugate match.

Harry R. Hyder, W7IV
Tempe, Arizona

Dear Editor:

Stuff and nonsense! I tried to keep quiet, but cannot. I'm disappointed with the Fall edition of *Communications Quarterly*. Sixteen pages of misconceptions and misinformation ("Source Impedance of HF Tuned Power Amplifiers and the Conjugate Match")! You and the authors should read "Is Salt Water Taffy Being Distributed?" in the Fall issue and "Junk Science" in the summer issue. These articles point up most of the fallacies perpetrated by the article in question.

Circular logic (no matter how large the circle) cannot be used the prove anything. Saying that A is true because of B and B is true because of A doesn't get us anywhere. The construct of a "lossless" resistance is an interesting confusion factor as is "average impedance." If the article is correct, then a 1-amp load on the power utility 120-volt line "proves" that the source impedance is 120 ohms! Indeed! The implication that somehow a tuned amplifier is different is then totally discounted by tuning the amplifier to resonance where there is left only a generator, a source resistance, and a load resistance.

The experimental data is mostly invalid, as there are too many dependent variables. For example, when the load resistance is changed on a π network, the impedance radio also changes (the Q changes). To "hold things constant," it is necessary to change network values.

While purporting to provide an understanding of the basics and to dispel misunderstandings, the article actually does just the opposite.

Gerald S. Bower, AB6B

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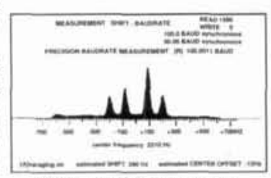

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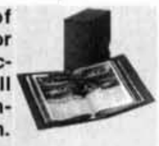

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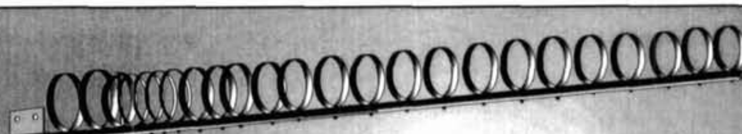

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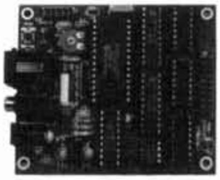

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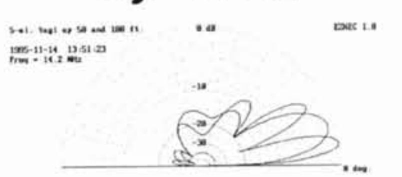

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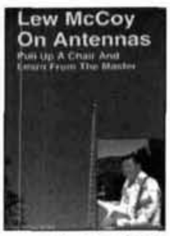

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4 Functions in One
MX-9300
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Features
- One instrument w/ four test leads and measuring systems.
  - 1.3GHz Frequency Counter
  - 2MHz Sweep Function Generator
  - Digital Multimeter
  - Digital Triple Power Supply
- 0-30V @ 3A, 15V @ 1A, 5V @2A

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- 40,000 Count
- High Accuracy
- Tektronix Quality
- 3 Year Warranty

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DMM 914	$235
DMM 916	$275

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Model 4040
- 0.2Hz to 20MHz
- AM & FM Modulation
- Burst Operation
- External Frequency Counter to 30MHz
- Linear and Log Sweep

$399

10MHz - Model 4017	$309
5MHz - Model 4011	$239

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96B		$1695
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28 tools plus a DMM contained in a large flexible tool case with a handle ideal for everyone on the go!
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Fluke Multimeters

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Model LCR-1810
- Capacitance 0.1pf to 20µF
- Inductance 1µH to 20H
- Resistance .01Ω to 2000MΩ
- Temperature -20°C to 750°C
- DC Volts 0-20V
- Frequency up to 15MHz
- Diode/Audible Continuity Test
- Signal Output Function
- 3 1/2 Digit Display

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Model M-1700
$39.95
11 functions including freq to 20MHz, cap to 20µF. Meets UL-1244 safety specs

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Handheld Universal Counter

Model F-2850
10Hz - 2.8GHz
Dual Input
$149

NEW
F-2800
1MHz - 2.8GHz
$99

Features:
- Full range 10Hz - 2.8GHz (F-2850)
- 16 segment RF signal strength bargraph 1MHz - 2.8GHz (F-2800)
- Dual 50Ω/1MΩ inputs. (F-2850 only)
- 16 segment RF signal strength bargraph
- 10 digital LCD display with backlight. (F-2850 only)
- 6 hour NiCd battery operation.
- High speed 250MHz direct count for high resolution.
- Multifunctions: Frequency, Period and Auto Trigger & Hold. (F-2850).

Specifications
Input Sensitivity (Typical)

Amplifier	1MΩ (F-2850 only)	50Ω
Impedance	1MΩ (F-2850 only)	50Ω, VSWR<2:1
Range	10Hz - 50MHz (F-2850 only)	1MHz - 2.8GHz
Sensitivity	<10mV @ 10Hz - 10MHz <20mV @ 10MHz - 50MHz (F-2850 only)	<1.5mV @ 100MHz <5mV @ 250MHz <5mV @ 1GHz <100mV @ 2.4GHz
Maximum Input	100Vrms (F-2850 only)	15dBm

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SCANCAT'S WINDOWS FEATURES
- Unattended Logging of frequencies
- Scan Create Disk Files.
- Spectrum Analysis to Screen OR Printer.
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- UNIQUE database management system with moveable columns. Even SPLIT columns into doubles or triples for easy viewing of ALL important data on one screen.
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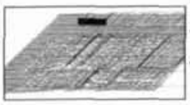

- Demographic search for frequency co-ordination and 2-way Usage Analysis.
- Detailed logging to ASCII type files with DATE, TIME, Sig Str, Air Time.
- Exclusive "MACRO" control by frequency of Dwell, Hang, Resume. Sig. Treshhold and even 6 separate programmable, audible alarms.
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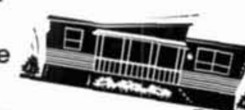

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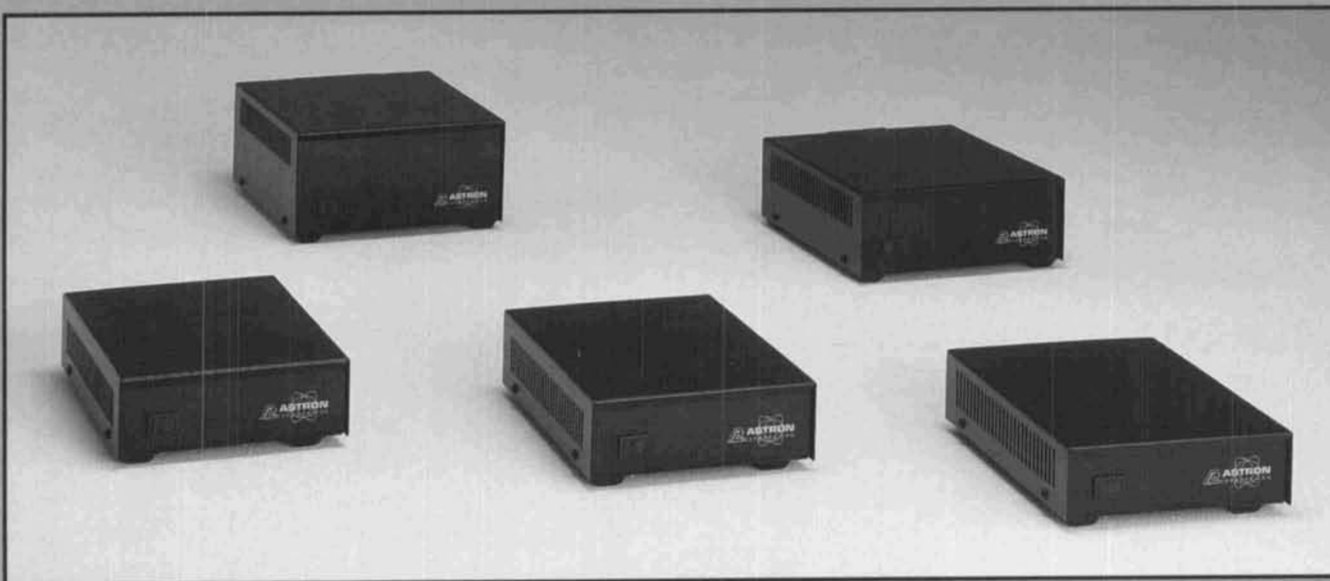

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OUTPUT VOLTAGE: 13.8 VDC

MODEL	CONT. AMP	ICS	SIZE	WT.(LBS)
SS-10	7	10	1 1/8 x 6 x 9	3.2
SS-12	10	12	1 3/8 x 6 x 9	3.4
SS-18	15	18	1 3/8 x 6 x 9	3.6
SS-25	20	25	2 7/8 x 7 x 9 3/8	4.2
SS-30	25	30	3 3/4 x 7 x 9 5/8	5

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R11 TEST RECEIVER

30MHz - 2GHz

Handheld Receiver

Optoelectronics is pleased to introduce the all new R11 Nearfield FM Test Receiver. Capable of sweeping 30MHz - 2GHz in less than one second, the R11 can lock onto a 5 watt UHF signal as far away as 500 feet in less than one second, demodulate the signal through its built - in speaker, and display the general band the frequency is transmitting in on its LED indicator. The R11 Test Receiver presents all new performance, features, and capabilities.

U.S. Patent No. 5,471,402

Volume & Squelch Control Knobs

Instruction Indicators: LED's will illuminate which mode the R11 is configured for.

CI-V and Headphone jacks: CI-V jack allows for connection to the Scout for Reaction Tune. The Headphone jack connection also allows for external speaker.

Frequency Band Indication: Displays what band the received frequency is transmitting on.

Built - in Speaker : Instantly demodulate any receiver frequency between 30MHz - 2GHz (Cellular Blocked).

Hold / Mute Button: The Hold button allows the R11 to stay locked on the received signal.

Power

Lockout / Lockouts on-off: The R11 allows for 1000 user activated lockouts.

Shift / Off: The Shift button controls all of the R11's secondary functions.

Skip / Clear Lockouts: Press the Skip button to continue sweeping. Clear Lockouts will empty the lockout memory.

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