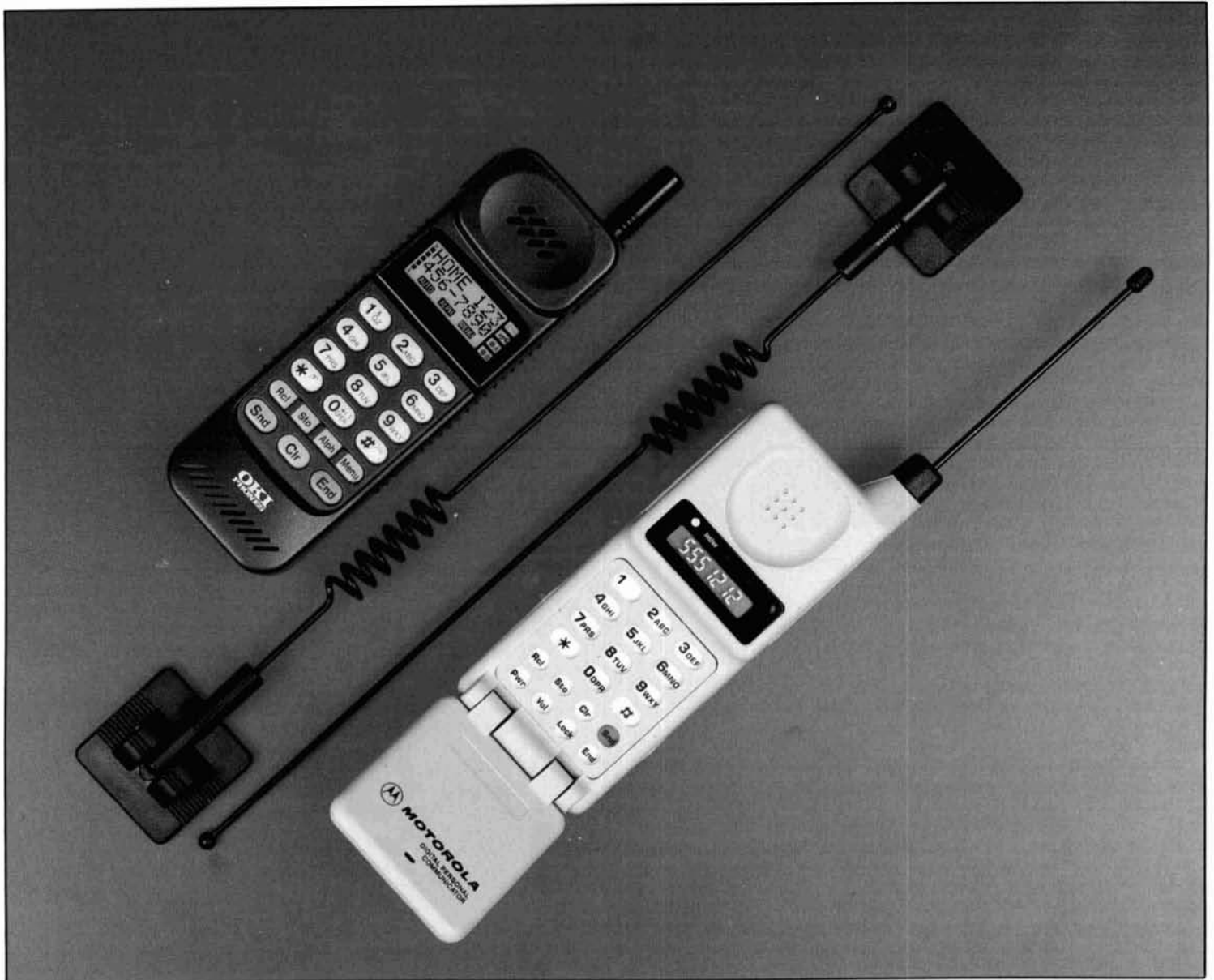# COMMUNICATIONS QUARTERLY

THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

- The Cellular Connection
- An Introduction to Fiber Optic-Basic Local Area Networks
- Micro-Computer-Based Instrumentation Systems
- UHF Voltage Variable Phase Shifter
- Build This High Power S-Band Step-Recovery Diode Tripler
- The Titan 425 Reliability Enhancements
- The 160-Meter Semi Vertical: A Case History
- Propagation Losses Between Randomly Located Antennas at UHF

6 Reasons w
build your sys

IC-725 HF Transceiver   PS-55 Power Supply   IC-2KL Linear Amplifier   IC-475 UHF Transceiver
                                                                                      IC-275 VHF Transceiver

*The IC-725 system above is just one example of how you can build your system.*

"Don't just buy a radio. Build a system." Experienced ham operators have been giving this advice for a long time. As you build your station, you don't want "stand alone" rigs that cannot integrate with the rest of your equipment. You can avoid serious disappointments in the future by comparing compatibility, performance, reliability and service *before* you purchase each component of your system.

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If you have ever made a weak signal QSO, had it fade out, then dialed the same frequency on an ICOM transceiver and found a strong, clear signal... you're already aware of ICOM's superior performance.

We have devoted three decades of research and design to ensure that our transceivers consistently meet or exceed your critical expectations. Our engineers have achieved a balance of dynamic range, selectivity and sensitivity thought unobtainable a few years ago. Their development of

the ICOM DDS System, with an extremely low carrier-to-noise ratio, has resulted in loud and clear transmissions and the ultimate in HF receiver performance.

PLL CARRIER-TO-NOISE
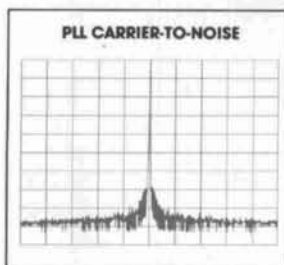
2 DESIGNED FOR COMPATIBILITY

It has been said many times, "Once you learn to operate one ICOM radio, every other ICOM is a snap to pick up." This is especially helpful as you upgrade or when you operate two or more transceivers simultaneously.

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This IC-765 system is another classic combination of ICOM components.

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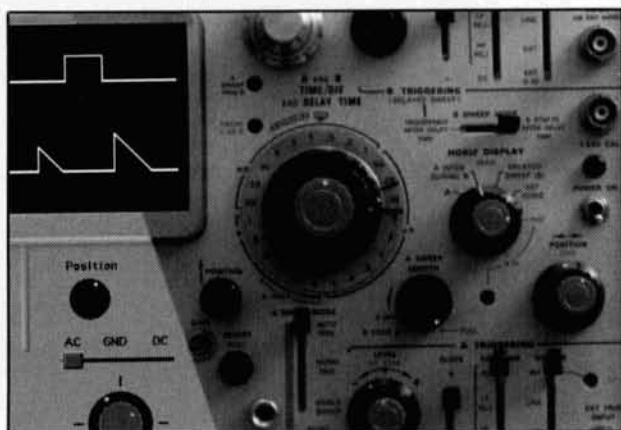
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COMMUNICATIONS QUARTERLY

THE JOURNAL OF
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Winter 1992

Cover photo: Nicholas Thomas, N1KCZ. Cellular phone and antennas courtesy of Cellular One, Boston, Massachusetts.

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EDITORIAL

The "New" Age of Technology

About twenty years ago, my friend Mark decided to build a computer. I was skeptical (Mind you, at this time, most computers I knew of filled entire rooms with banks of tape machines, racks of back-up tapes, CPUs, and three-foot high disk drives loaded with disks the size of dinner plates.), but Mark was determined. Not knowing much about his subject, or how to go about his task, he scoured the available literature for clues. He ordered an IMSAI 8080 front panel, memory chips, sockets, piles of plastic bags full of assorted parts, and set to work. He put together the front panel, built a motherboard, connected everything with a tangle of ribbon cables, and mounted the finished product in a clear Plexiglas® case. His monitor was an old TV set he had lying around. The computer had a cassette interface to allow Mark to store his data on tape—no hard drives here! In the beginning, Mark didn't even have a keyboard, so he entered all his data (which was considerable) via rocker switches on the front panel. This entry process took hours and hours. And, it had to be done all at once because he had little resident memory in the computer. Often an equal amount of debug time was required to correct errors which crept in during late-night data-entry sessions. But Mark persevered, refined, reworked, and upgraded his "baby" until he had quite a system up and running. And that home-built computer got him his first job in the computer industry! Since that time, Mark has moved on to commercially built PCs and hard drives, but he's never forgotten the lessons he learned building that first computer.

New technologies have always given wings to the creative fantasies of scientists, design engineers, entrepreneurs, and basement inventors. What seemed so new and innovative when Mark set to work on his dream, is commonplace and even archaic today. However, the same companies that were in the forefront of solid-state technology then—like INTEL, Motorola, NEC, Signetics, and Texas Instruments, to name a few—are still pushing the envelope of research and development today. Many experimenters have visions of combining prod-

ucts from the DSP, fiber optic, and cellular arenas with the latest RF engineering techniques and surface-mount technologies in ways never before imagined. The applications for commercial products and home construction projects are mind boggling! Conditions are ripe for a new breed of inventors. We at *Communications Quarterly* want to give the inventors and experimenters among you the help and encouragement you need to latch on to up-and-coming technologies and run with them.

In this issue, you'll find the first installment of "Quarterly Devices." It's written by another "basement inventor" friend of mine—Rick Littlefield, K1BQT. Rick has spent many years designing and building amateur radios and test equipment. Because he has always enjoyed incorporating new products into his projects (and because he said it would be great if "CommQuart" ran such a column), we've asked Rick to talk to today's leading R & D wizards and ferret out all the information he can about the latest technical "building blocks." He'll also give you hints on how to use these devices in your next construction project. And because the exchange of knowledge can enhance the process of experimentation, Rick hopes you will share what you've learned with him.

Although the technology may change, the urge to create remains the same. Many of us have the desire to do something differently, do something better—just **do** something. No matter what you build, how you build, whether you're a basement inventor or a design engineer, the time is ripe for experimentation. There are new technologies, with their attendant hardware and software, sprouting up every day. With all this abundance at our fingertips, it's important to work together as a community of thinkers and builders so we can discover the best applications for the latest technologies and build our dream projects. So, look around; check out the possibilities. Enter the "new" age of technology, and let us know what you find out.

Terry Northup, KA1STC
Editor

LETTERS

Article piques reader's interest
I am a member of the Board of Directors of the QRP Amateur Radio Club International, which is one of many non-profit groups around the world of hams interested in home-brewing, building, and operating with low-power equipment for the enjoyment of it. (The generally accepted definition of QRP is 5 watts output or less.)

In your Summer 1991 issue, you had an article by Cornell Drentea, WB3JZO, titled "Upgrading the FT-ONE Transceiver." This included a drawing on page 53 of a precision air dielectric trimmer capacitor. I have acquired many of these over the past few years from surplus equipment, but have not yet played with them. My interest has been piqued by the article and I plan to do some experiments with these capacitors.

By the way, although all your issues have been excellent, I found this one to be particularly good since it contained an unprecedented number of articles (nine) of immediate and lasting interest to me, and *all* of the remainder *will* be studied and learned from later. (Although I stand by the comment I made in my last letter about the slick paper and doodling space, I do have to admit that I used the space for making a number of comments, observations, and calculations.) Incidentally, I'm sure someone has already pointed out that the drawings for **Figure 2** and **Figure 4** on pages 51 and 52 were reversed, although the captions agreed with the text.

Michael A. Czuhajewski, WA8MCQ
Severn, Maryland

Some interesting observations
Regarding the article by Cornell Drentea, WB3JZO, on "Improving Receiver Performance" in the Fall 1991 issue. I can't say I've ever had a problem of the sort he describes, but your readers may be interested to know that at HF, the 1N4000 series rectifiers make extremely good RF PIN-type switches! They have a reverse capacity of about 8 pF, it is true, which can be a limit, but at a few cents each, they're sure a lot cheaper than most PINs. I've used one very successfully under test at 100 watts at 1.8 MHz, with a bias current of about 400 mA. The reverse recovery time is about 8 to 10 microseconds. I've been using one for about 2 years now for switching on the input of my linear.

John True's article on "Shunt Feeding Grounded Verticals" was very interesting. After a lot of experimentation, I gave up on the

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• LOW PROFILE POWER SUPPLY						
SL-11A	•	•	7	11	2¾ x 7⅝ x 9¾	11

RS-L SERIES

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• POWER SUPPLIES WITH BUILT IN CIGARETTE LIGHTER RECEPTACLE				
RS-4L	3	4	3½ x 6⅛ x 7¼	6
RS-5L	4	5	3½ x 6⅛ x 7¼	7

RM SERIES   MODEL RM-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• 19" RACK MOUNT POWER SUPPLIES				
RM-12A	9	12	5¼ × 19 × 8¼	16
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50
RM-60A	50	55	7 × 19 × 12½	60
• Separate Volt and Amp Meters				
RM-12M	9	12	5¼ × 19 × 8¼	16
RM-35M	25	35	5¼ × 19 × 12½	38
RM-50M	37	50	5¼ × 19 × 12½	50
RM-60M	50	55	7 × 19 × 12½	60

RS-A SERIES

MODEL RS-7A

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

RS-M SERIES

MODEL RS-35M

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

VS-M AND VRM-M SERIES

MODEL VS-35M

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

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

RS-S SERIES

MODEL RS-12S

• Built in speaker

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

Tom McMullen, *W1SL*
4109 Waterway Drive
Lake Worth, Florida 33461

THE CELLULAR CONNECTION

An in-depth look at cellular communications

I n the midst of many highly visible high-tech revolutions, one that is perhaps second only to computers is making a giant impact on our way of life. The invader is a system, rather than an object, and it stands out as a success story that grew from much careful research, market assessment, and far-reaching planning.

The system is called "cellular" in popular jargon, but the proper terminology is "Cellular Mobile Telephone System." This mouthful is usually restricted to the various textbooks and papers published in the academic and engineering worlds. For those who know what the system is all about, "cellular phone" or "cellular system" is sufficient, but to most users it is simply a "car phone."

How did we get here?

A brief look back will help us see what created the need, and why the sweeping change became the success it is today.

As recently as the early 1970s, there were only three mobile telephone systems: one in the vicinity of 40 MHz called Mobile Telephone System (MTS), one near 150 MHz called Improved Mobile Telephone System (IMTS), and another IMTS near 450 MHz. MTS at 40 MHz and IMTS at 150 MHz each offered 11 channels for mobile callers to use; IMTS at 450 MHz provided 12 channels.

It doesn't take much imagination to see that any reasonably large metropolitan area would have enough subscribers (mobile users) to overload such a system in short order. The situation was made even more intolerable by the allocation of only a few channels to any one city or area. New York City, for example, was allocated six channels at 150 MHz and another six at 450 MHz. Their subscriber base was less than 600 customers, with just under 4000 on a waiting list!

When so few people must support a large system, the price is inevitably high, and the frustration levels among users and potential users becomes equally elevated. In the case of New York City, for example, among the thousands of commuters going to and from the city daily, only 12 could make a call at any given time. Without a way to expand the user base, there was no feasible way to either lower the price or increase profitability. The limited amount of spectrum available in the VHF region precluded simply adding more channels.

An additional inhibiting factor was inefficiency in spectrum usage. Each mobile telephone occupies two frequencies, one for transmitting and one for receiving, in order to have full duplex operation.

Clearly, some creative thinking and planning was needed to develop a system which would satisfy the needs of a growing segment of the public (read also as business community), and provide an alternative that would make efficient use of the spectrum and expand to meet present and future mobile communications requirements.

The Cellular Mobile Telephone system grew out of this planning and testing.

Finding room in the spectrum

At the time the planners were looking for spectrum space for an expanded mobile telephone system, the higher channels in the
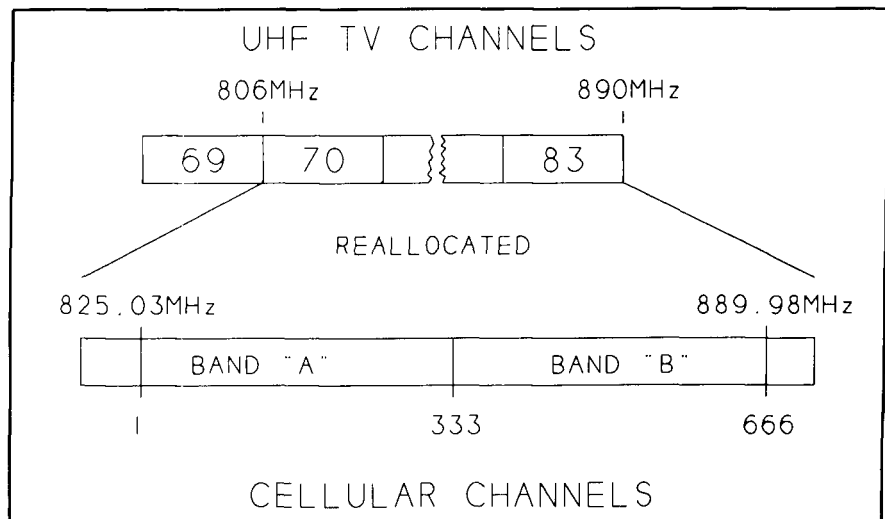
Figure 1. The frequencies used by cellular channels were once part of the UHF Television assignment, particularly those intended for Educational TV. The cellular channels range from 825 through 890 MHz, with a 45-MHz separation between transmit and receive frequencies.

UHF television range were mostly unused. Some were assigned to Educational Television. This limited use was largely a result of poor coverage and, to some extent, was due to the difficulty in developing high power (50 or more kilowatts) at these frequencies. Also, television receivers of the day were notoriously poor performers at the upper end of the UHF TV allocation. Another factor was the proliferation of cable television that began providing quality signal coverage to a growing market.

These combined circumstances made that chunk of the UHF spectrum attractive for other purposes, and early theories and propagation tests showed that a mobile telephone system could be made to work in the 800-MHz range.

This all happened during the mid 1970s, when 800-MHz was pushing the limitations of the technology of the day. Fortunately, the system showed enough promise to spur development of medium-power RF transmitting equipment that could perform efficiently at 800 MHz. Those of us who were experimenting with Amateur 420 and 1200-MHz communications at that time relished a chance to look at the specification sheets of new transmitting tubes. We noted, but didn't comprehend the reasoning for, operating parameters that peaked at 900 to 1000 MHz.

Other technology was developing smaller, lower-power mobile solid-state equipment that had receiver noise figures impressively better than most tube-type equipment. Another fast-growing resource was large-scale integration (LSI) of complex circuits that improved speed, reliability, and effi-

ciency and also made our compact mobile and portable units possible. The market and the technology were coming together to provide a service that most of us didn't yet know we needed.

In 1979, the Illinois Bell Telephone Co. was authorized to develop a cellular mobile telephone system in the Chicago area. Another group (ARTS—American Radio Telephone Service, Inc.) was to operate a system in the Baltimore-Washington D.C. area. These were the test beds of cellular-telephone development, and they demonstrated that such systems were technically feasible and could be affordable.

Early FCC plans allotted only one licensee for each market. However, studies showed that the public would be better served by having two systems serving each area. The Commission reconsidered its initial plans, creating what is now called Band "A" for non-wire-line operators, and Band "B" for wire-line operators.

Wire-line operators are those who are also in the business of providing "land-line" telephone service; in other words, "telephone companies." (Research into the beginnings of cellular technology has uncovered information that the original "B" allocation was for the "Bell" licensees. The "A" allocations were for the "alternate" licensee.) This early distinction has blurred now, and some areas are served by two "wire-line" entities. This can happen if, say, a local telephone company obtains the "B" license, and the licensee of the "A" service later sells the business to some non-local telephone company that wishes to expand its coverage area.

Band A frequencies are 825.03 through 870.99, corresponding to channels 1 through 333. Band B occupies 880.02 through 889.98, corresponding to channels 334 through 666 (see **Figure 1**). Each channel contains two frequencies, one for cell-site transmit and one for cell-site receive, with a 45-MHz separation between the two. A cell can be assigned up to 16 channels for its use, one of which must be a control channel. (See the sidebar for instructions on developing your own spreadsheet that shows the channels and frequencies for both A and B bands.) Since the original allocation of frequencies, an additional 166 channels have been provided. Some channels have been added below the current Channel 1; the remainder were added above Channel 666. At this writing, most systems are still using the standard 333 channels for Band A and 333 for Band B.

Developing a coverage area.

The methodology of developing a coverage area and designing the components to provide service in that area can be very complex. Let's look at some of the issues.

One consideration is the size of the cell that's needed. This, of course, depends upon the amount of traffic in the proposed area. Surveys can be made of vehicular traffic on major highways with particular attention paid to peak traffic hours. Based on some assumptions (which have been refined as experience is gained) about the number of vehicles that have telephones and how many of those will want to place a call during any typical rush hour, designers can predict how many cells and how many frequencies will be required to provide the desired level of service.

The overriding theme in developing a system for a given area is customer satisfaction, which translates into the number of calls that can be completed versus the number of calls attempted. The difference between the two is referred to as "call blockage."

In planning, the magic number seems to be a two percent call blockage. Any percentage greater than that (except during very unusual circumstances) would require a new look at the design parameters. One of the remedies of poor performance is cell dividing or cell overlays. More on this later.

Now let's get back to traffic calculation. To grab some numbers out of a hat, if an area has 50,000 cars per hour on its major highways, and if half of them have car phones, then there are 25,000 potential callers. If 80 percent of these users make a call,

the traffic load is 20,000 calls per hour. This provides a base number for a large area that is to be covered by several cells.

Most of the calls made will be of short duration (depending, of course, on the charge-per-minute for each call). An average number that has been widely used in traffic loading calculations is approximately 1.75 minutes per call. A 1991 survey by the CTIA, however, shows that an average call length is now 2.37 minutes. When plugged into the equations, this number will indicate a higher probability of call blockage. This could accelerate cell splitting and cell overlays to accommodate the users.

Let's look at what one of these cells might involve. If our hypothetical cell has a vehicular traffic load that produces a potential 2000 calls per hour, and the average call is 1.75 minutes in duration, then the probable load is:

$$(2000 \times 1.75)/60 = 58.333 \text{ erlangs}$$

Since we want to provide enough capacity to allow for that two-percent call-blocking cushion, the required number of channels for that traffic load is 70.

How did we arrive at this number, and what in the world is an erlang?

Erlang is a term widely used in congestion theory, or in the case of telephone systems, trunking theory. It's an international, dimensionless unit of occupancy (or traffic intensity); a ratio of the time a facility (such as a telephone system) is occupied to the time the facility is available.

The formula used to arrive at the number of cells is the Erlang "B" formula. (Here, B refers to the assumption that a certain percentage of the blocked calls will hang up. This is known as Blocked-Calls-Cleared.) The formula is:

$$P = \frac{\dfrac{A^n}{n!}}{\displaystyle\sum_{x=0}^{n} \dfrac{A^x}{x!}}$$

Where:

A = traffic density (erlangs)

P = probability that calls will be lost (blocked)

n = number of channels

x = variable representing the number of busy channels

Not being one to reinvent the wheel, I used some conveniently published tables.[1] Knowing the erlangs and the two-percent blockage limit, 69 channels (or trunks) will

Create a Spreadsheet That Shows Channels and Frequencies for the A and B Bands

```
B2: ((B1*0.03)+870)                                        READY
```

	A	B	C	D	E	F	G
1	Channel (C)	333	332	331	330	329	328
2	Trans., MHz	879.99	879.96	879.93	879.9	879.87	879.84
3	Rcv., MHz	834.99	834.96	834.93	834.9	834.87	834.84
4							
5	Channel	312	311	310	309	308	307
6	Trans., MHz	879.36	879.33	879.3	879.27	879.24	879.21
7	Rcv., MHz	834.36	834.33	834.3	834.27	834.24	834.21
8							
9	Channel	291	290	289	288	287	286
10	Trans., MHz	878.73	878.7	878.67	878.64	878.61	878.58
11	Rcv., MHz	833.73	833.7	833.67	833.64	833.61	833.58
12							
13	Channel	270	269	268	267	266	265
14	Trans., MHz	878.1	878.07	878.04	878.01	877.98	877.95
15	Rcv., MHz	833.1	833.07	833.04	833.01	832.98	832.95
16							
17	Channel	249	248	247	246	245	244
18	Trans., MHz	877.47	877.44	877.41	877.38	877.35	877.32
19	Rcv., MHz	832.47	832.44	832.41	832.38	832.35	832.32
20							

```
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```

If you would like to create your very own chart showing the channel and frequency assignments for cellular systems, here's how you can do it in Lotus 1-2-3®. If you use another spreadsheet, such as EXCEL® or Quattro Pro®, you will have to modify the commands and procedures to fit.

To create a spreadsheet like the example shown, start by putting the labels in column A. Start at A1 with the word Channel, A2 with Trans. (MHz), and A3 with Recv., (MHz). Place a horizontal line in A4 by typing in \ - (back-slash and hyphen).

Go to B1 and enter 333
Go to B2 and enter ((B1*.03) + 870)

Go to B3 and enter ((B1*.03) + 825)
copy A4 to B4

You should now have the first channel listing in the spreadsheet. If it worked correctly, B column should show channel 333, Transmit frequency 879.99, and receive frequency 834.99, with a horizontal line in B4, under the last frequency. These frequencies are those used by the Cell Site; the mobile unit transmit and receive frequencies will be the reverse.

Go to C1 and enter (B1-1)
Copy the range B2..B3 to C2.
Now copy the range C1..C4 to a range D1..V1

work with 58.2 erlangs; rounding up to 70 channels will provide a slight cushion. Because each cell can be assigned no more than 16 channels, it will require 70/16 = 4.375, or 5 cells to handle our theoretical 2000 calls per hour.

When a cellular coverage area is initially designed, some of the cells will be working close to full capacity and others, in outlying areas, will be lightly loaded. This disparity in loading is handled by assigning the full complement of 16 channels to the heavily loaded cells, and fewer channels to those that don't need them.

Cell spacing

One of the keys to a successful system is frequency reuse. This means that the same set of channel frequencies can be used in another cell if the separation between cells is great enough that interference either doesn't occur or isn't severe enough to cause a degradation in service.

The separation (more properly, frequency-reuse distance) calculations are based on a ratio between D (Distance between cell sites or cell centers) and R (Radius of the cell). The formula is:

$$D = (\sqrt{3K}) R$$

Where:

D is the center-to-center distance.

K is the number of cells in the basic group.

R is the radius of a single cell.

When the cells in an area are of uniform size, the number of cells that can fit in distance D is easy to envision. **Figure 2** shows a cell configuration with a D/R of 4.6 and K = 7. Notice the method used to find the

The numbers in V column should be 313, 879.39, 834.39. If not, go back and check the formula in C1, C2, and C3, or check your "copy" procedure.

Copy your labels in the range A1..A4 to A5.

Go to B5 and enter (V1-1).

Copy the range B2..B4 to B5.

Go to C5 and enter (B5-1).

Copy the range B6..B8 to C6.

Copy the range C5..C8 to the range D5..V8.

The numbers in B5..B7 should be 312, 879.36, 834.36. If not, go back and check formulas.

The numbers in V5..V7 should be 292, 878.76 and 833.76.

Hang in there; it gets easier from here on.

Copy the range A5..V8 to A9.

Copy the range A9..V12 to A13.

Continue copying the ranges in this manner through range A61.

The numbers in B61..B64 should be 18, 870.54, and 825.54. Cells T61, U61, V61 will have a negative number (-1). These last three addresses are invalid. Erase range T61..V63.

You should see channel 1 in address S61, with frequencies 870.03 and 825.03 below it.

Save the spreadsheet with a filename.

This chart represents the "A" Band referred to in the article. The highest numbered channel in each column is the control channel. If you like, you can go back to A1 and modify that entry to read "Channel (C)" to indicate these are the control channels.

You can create a spreadsheet for the "B"

Band by modifying the formulas in a few places as follows:

Save the spreadsheet again but give it a *different* filename.

Change the entry in B1 to 334.

Change the formula in C1 to (B1 + 1).

Copy C1 to a range D1..V1.

Change the formula in B5 to (V1 + 1).

Change the formula in C5 to (B5 + 1).

Copy C5 to a range D5..V5.

Copy the range B5..V5 to B9.

Continue to copy the ranges down the spreadsheet until you have changed all the subtraction formulas to addition. Note that this version of the spreadsheet adds one number to each channel as you progress. The highest channel number is 666, and should apear in address Sxx. The frequencies under channel 666 should be 889.98 and 844.98. As before, addresses T61 through V63 should be erased if they are not empty.

In this band, the lowest channel in each column is the control channel, so you can type in the label "Channel (C)" in A1.

If you want to put a header at the top of each chart, insert a new row at A1 and type in the label "BAND A" or "BAND B" or whatever you desire.

I used SIDEWAYS® to print my chart without splitting it across several pages. By proper setting of line spacing and type size, it can be printed on the 11-inch width of a standard printer. Wide-carriage machines make it easier, of course.

location of the next cell that reuses a set of channels. Using frequency group 1 as a sample, count outward (from a cell face) two cells, then turn left 45 degrees and count one cell. That cell is at the proper distance to reuse frequency group 1. For a structure with K = 12, the procedure is to count outward two cells and left two cells. Of course, other ratios can be used. Smaller separation requires reduced power and highly directional antenna patterns to obtain protection. Larger separation requires that a greater number of channels be available to cover the area.

As an area grows, the demand for service increases, and will eventually reach the point where the start-up cells can't handle the demand. Two solutions are possible here: cell splitting and cell overlay.

Cell splitting involves creating several smaller cells within the area formerly used by one cell. If the original cell had, say, a 5-mile radius, several cells of 1-mile radius can be put into place—each with as many channels as necessary to handle the load. This increases the problem of interference prevention, so the new cells must have careful planning for both frequency assignment and frequency-reuse spacing from previous cells.

Cell overlay consists of placing another cellular coverage area on top of the established one. The new cell has its own assigned channels. Its coverage area doesn't have to match that of the parent cell exactly, but can be smaller. Often, this means installing a second set of transmitters, antennas, receivers, combiners, etc., at the same site as the original. Careful channel selection is needed to prevent intermod and ad-
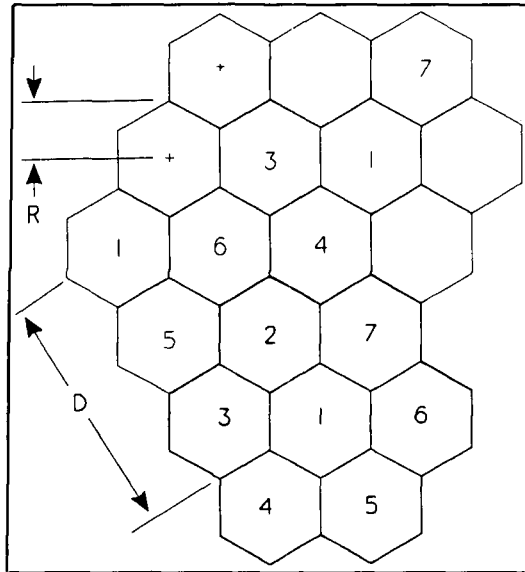
Figure 2. One of the secrets of cellular operation is the frequency reuse system. Planners can calculate spacing between cells that use the same frequencies to minimize interference while maximizing service. A working formula obtains a cell-center-to-center distance from a ratio of D to R.

jacent-channel interference.

With either cell splitting or cell overlay, the MTSO capabilities must be upgraded to handle the greater number of call originations, hand-offs, and other tasks that the increased traffic load will require. Also not to be overlooked is the need for extra land-line trunks to handle the calls.

That there will be a need for upgraded capacity in the current cellular system is strongly suggested by the fact that the number of subscribers nationwide is growing at the rate of 183,000 per month.[2]

Signal coverage

The previous discussion about providing enough channels to handle the expected traffic load is only one part of the planning process. Covering the cell area with transmit and receive capability also requires careful consideration. Fortunately, these properties are reciprocal; that is, a condition that effects the transmitter coverage will impact the receive coverage in a like manner (discounting, for simplicity, any effect man-made noise has on the quality of signal received). This allows a system designer more freedom to tailor the radiated signal to the desired area.

An outer-contour received-signal-strength limit of 39 dBμV (dB in μV/meter), or -93 dBm, has been established by the FCC. Experience has shown this figure to be generous; a limit of 32 dBuV (-100 dBm) is

practical in many systems. Accepting these numbers as the desired end result, it's easy to use the transmitter output power (at the cell site), combiner and transmission line losses, and antenna gain to determine the ERP (effective radiated power). From this point, propagation path losses can be factored in to obtain the signal level at the edge of the cell.

The system

A cellular system is composed of three main blocks in its functional description: the Mobile Telephone Switching Office (MTSO), the cellular site or station, and the mobile (or portable) unit (see **Figure 3**).

The MTSO

The real brains in the outfit reside in the MTSO. This office is linked to all of the cell sites within its jurisdiction via land-line or microwave links. It also has access to the normal telephone line "trunks" to local standard telephone central offices and to other MTSO sites for coordination of "roaming" users. When you delve into the tasks that the MTSO must handle, it's easy to appreciate how complex the software must be, and how reliable the hardware is, to keep the system operating smoothly.

What the software does

One of the most complex parts of the cellular system is the software that keeps it all working. In processing a call originated by a mobile unit, the MTSO program must:
• Verify the electronic serial number (ESN) of the unit to see if it is legitimate (not stolen).
• Verify that the caller's account is not in arrears.
• Determine if the caller is a "roamer," (a unit operating in other than its home system) and if so, where the billing for the call is to be sent.
• Notify the cell site of the results of these verifications.
• Notify the cell site what voice channel has been assigned to the caller.
• Arrange for connections to the land-line "trunk" system to route the call.
• Monitor other cells in its jurisdiction to see if the caller has a stronger signal in another nearby cell.
• Make connections between the mobile unit and the telephone being called as soon as the number has been identified.
• Coordinate access and handoffs among all cell sites in its jurisdiction.
• Maintain liaison with other MTSOs for
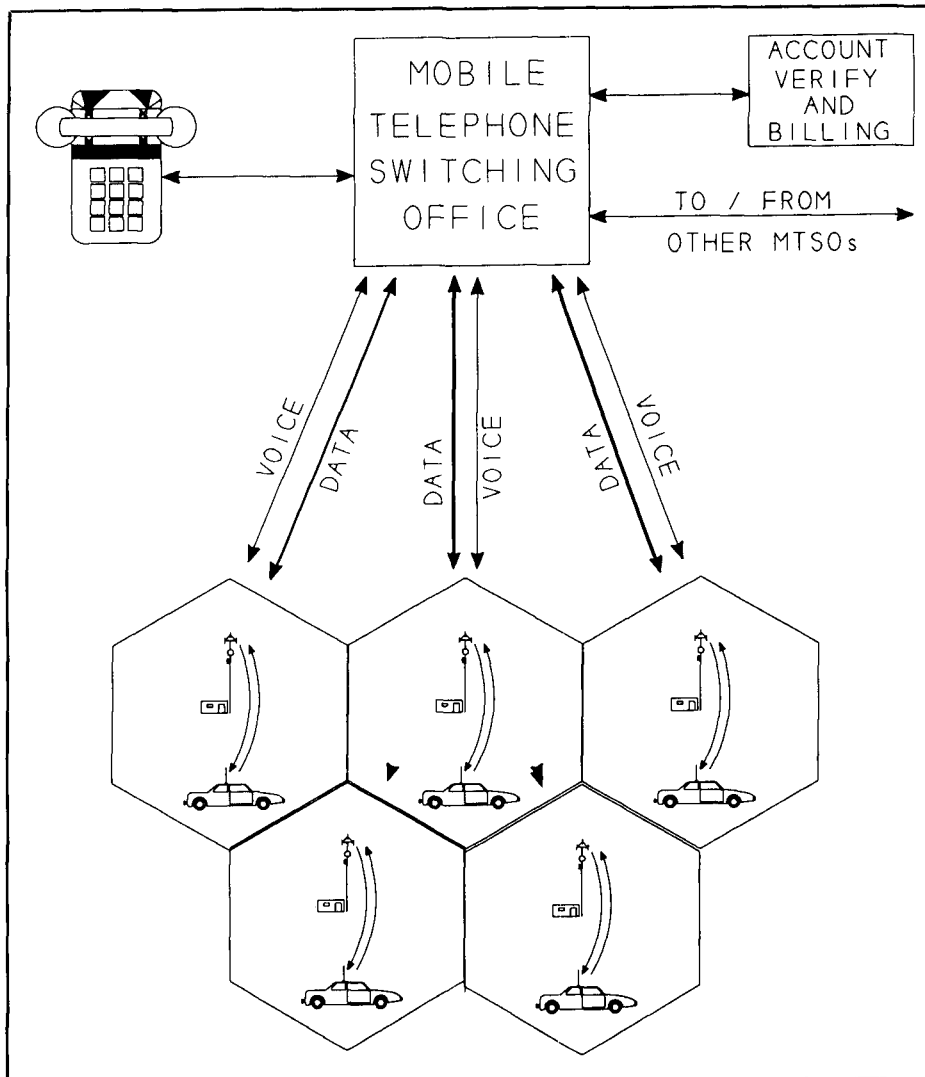
Figure 3. A cellular system requires three main functional entities: the Mobile Telephone Switching Office (MTSO); the Cell Site, which has the antennas, transmitters, and receivers; and the mobile unit. Data links between the MTSO and the cell site and other MTSOs can be via land-line trunks or microwave links.

possible handoffs from other cellular groups.

The task list for routing a call originated by a land-line phone to a mobile is even longer.

The cell site

The schematic representation of a cell site most often used by engineers and planners is a hexagon, shown in **Figure 4A**. This theoretical "coverage area" works out well for planning purposes, but real life isn't that precise. A circular representation is sometimes used, as in **Figure 4B**, but, again, antenna coverage patterns like this are obtainable only in the wide-open spaces where no structures, foliage, or terrain can distort them. The actual coverage area of a cell is likely to be an irregular shape molded by ter-

rain, buildings, wires or metal structures, and other geographical features **(Figure 4C)**.

The cell site contains the "base station" transmitter and receiver, antenna(s), an RF combiner, power supply and back-up supply (usually a bank of rechargeable cells), and a computer with software to handle procedural messages to and from the mobile units. The cell site also provides an interface with the MTSO, either through land-line data and voice trunks or via microwave links.

The base-station has one transmitter for each channel it has been assigned, up to 16 channels (15 forward voice channels and one control channel). Transmitter output power is adjustable over a range of approximately 3 to 25 watts. The FCC limits effective radiated power in a Metropolitan Statistical Area (MSA) to 100 watts; in a Rural
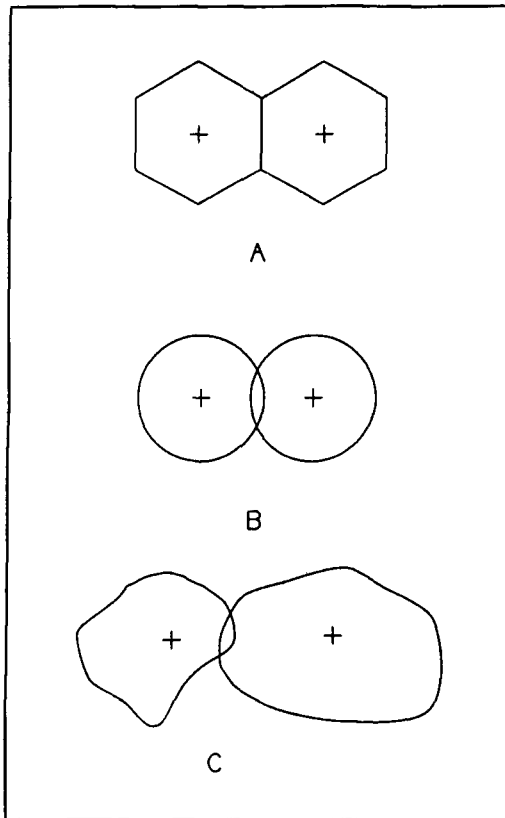
Figure 4. A theoretical cell is represented by a hexagon for planning purposes, A. This allows the planners to ignore the overlap area in a theoretically circular coverage area, B. Real-life coverage areas can take on almost any shape, depending upon terrain, obstructions, and antenna patterns, C.
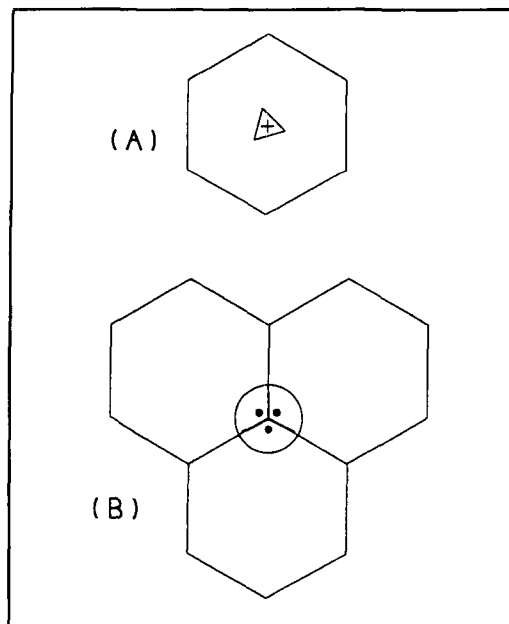
Figure 5. When the cell site is located at the center of its cell, the cell is said to be "center excited," A. When the cell site is at a corner, as in B, it is called "edge excited." This type of cell uses directional antennas to cover three cells from one equipment site.

Service Area (RSA), the ERP limit is 500 watts.

The output from each transmitter is fed to a 16-channel combiner that feeds a single transmission line to the antenna. If a cell has less than 16 frequencies allocated to it, the input ports on the unused combiners must be terminated with a dummy load to maintain isolation between channels.

The receivers, one for each return voice channel, are fed by two antennas and a combiner. The antennas, with combiner circuitry and filters, provide space diversity reception to minimize fading of the received signal.

Antennas and signal coverage

When the cellular system was first envisioned, the antenna model that was most often used assumed an omnidirectional pattern in the horizontal plane. Cells using this scheme are said to be "center excited" (Figure 5A). This model still applies in many areas, particularly suburban and rural conditions. In its simplest form, a tower at the cell site will have a triangular structure at its top. An omnidirectional antenna is mounted at each corner of the triangle (see Photo A). The usual configuration uses two of these antennas for receiving and one for transmitting. The more complex systems add more antennas for both transmitting and receiving as required. These omnidirectional antennas are made up of vertically stacked radiating elements that provide 6 to 9 dBd gain, which also concentrates more of the radiated energy along the horizontal plane.

As systems "mature" (the traffic load becomes higher), they require innovative solutions to the coverage-versus-loading-versus-frequency reuse problems. Many cells now use directional antennas to tailor their coverage more closely to a desired area and to protect against co-channel interference. These cells are called "edge excited" (Figure 5B), and their directional characteristics can be tailored to fit 120-degree sectors (a natural for a corner of a hexagonal cell), or 60-degrees for more critical interference-reduction needs.

Another method that provides good coverage in a cell and minimizes interference with other cells is the use of tilted-beam-pattern antennas. By properly designing the downward tilt of the main horizontal lobe of the antenna pattern, coverage in the desired area can be improved, and sometimes a null can be created to protect distant cells (see Figure 6). These patterns are often referred to as "umbrella" patterns. Stacked
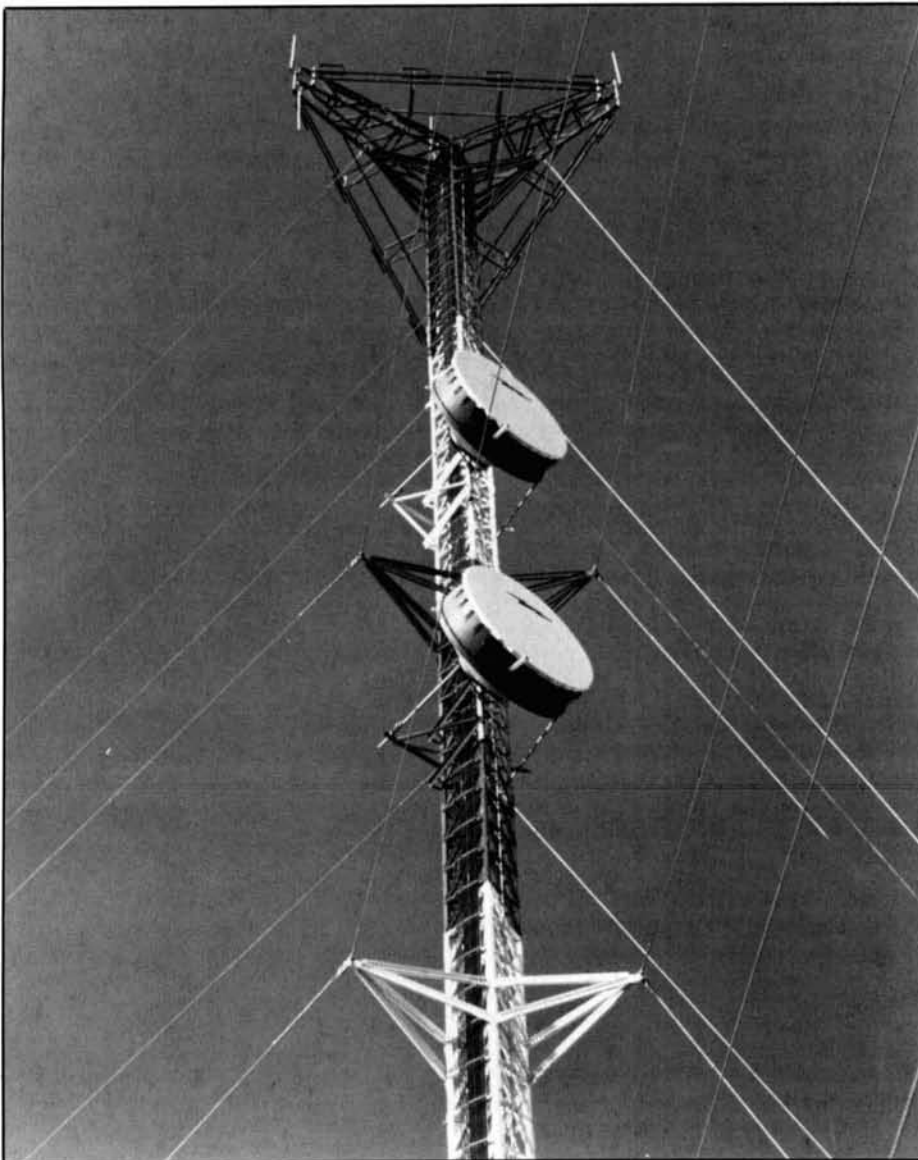
Photo A. Omni-directional cell sites have antennas mounted on triangular platforms atop a tower. Two receiving antennas and one transmitting antenna are used for a normal cell. As traffic loading increases, more equipment and antennas can be added. Microwave links provide data paths between the cell site and the MTSO. (Photo courtesy of Andrew Corporation.)

"discone" antennas can provide this type of pattern, as can properly phased stacked dipoles.

Most of the designs for cellular systems are based on a coverage pattern that is generally symmetrical, with several cells nestled within the boundary of a large cell-like area. There are cases, however, where the coverage must be tailored to shapes other than ideal. For example, a long stretch of heavily traveled highway may require cell sites with directional antennas to provide a pattern that blankets the highway while ignoring the surrounding rural area.

Studies were made of 800-MHz coverage and propagation early in the development of cellular mobile telephone structure. Many of these studies involved sending mobile units into the streets of a metroplolitan area to collect data on such things as depth of fading when the vehicle was moving toward or away from the transmitter, or when it was moving across the cell at a somewhat constant radius from the transmitter. The effect of buildings, wires, other vehicles, precipitation, and vehicle speed were also tabulated and studied to develop the criteria for a viable cell system.[3,4]

One thing that came out of these studies was the realization that, in general, the signal-strength decrease with distance fits a curve that drops 40 dB per decade. That is,
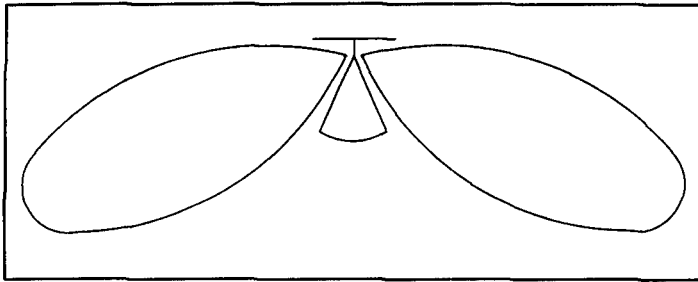
Figure 6. Shaped (tilted) antenna patterns can help prevent interference to distant cells. An "umbrella" pattern can be produced by spaced, stacked discone antennas among other types.

if a signal strength reading is taken at a point one mile from the transmitter, it will be 40 dB lower at a point 10 miles away, and 40 dB lower than that at the next 10-mile point. Also, the signal will fit within a standard deviation of 8 dB centered on this curve (see **Figure 7**).

The information derived from these studies became important not only in predicting and tailoring the signal coverage area, but also in developing "hand-off" criteria that enables an MTSO to prepare for and execute a hand off when a mobile unit approaches the boundary of a cell. I'll talk more about this procedure later.

The quality of the received signal in the mobile unit can be effected by many things in the coverage area. Buildings, trees, towers, and hilly terrain all produce some degree of flutter or fading on the signal, as **Figure 8A** illustrates. Even in rural areas, with flat terrain and no obstacles, the signal isn't steady. The phase combining of the direct signal and the signal reflected from ground causes peaks and dips **(Figure 8B)**. When the signal is strong, as it is near the
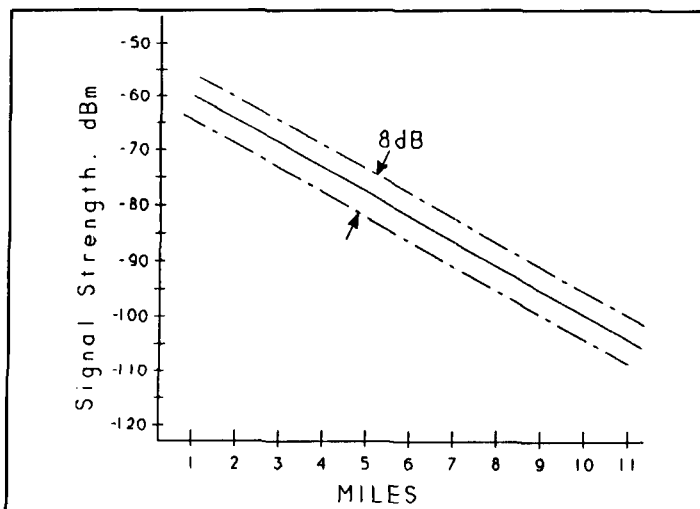
Figure 7. Signal strength decreases within a predictable margin as distance from the cell site increases.

cell site, these dips aren't severe enough to be of any concern. When the mobile unit is approaching the limits of the cell, however, they can alert the cell site to prepare for a hand-off.

Mobile units

Mobile units are, in some ways, similar to the mobile communications units we have long been using for other services—although modern technology has created small, stylish, and multi-functional transceivers. The basic "box" contains the bulk of the transceiver circuitry, along with a duplexer or combiner to allow both reception and transmission with a common antenna and feed line. This unit can be mounted under the dash, in the trunk, or at any other convenient location in the vehicle. The handset is similar to a modern handset for a home telephone. It has the familiar microphone and earphone, and an array of buttons to permit the user to dial a number, "start" the calling sequence, call a number from memory, and other useful features and functions. Some newer models have the capability of storing a phone number referenced to a name. You simply enter "John" on the keypad and it recalls the correct number from memory. In case of duplicate names, you are presented with the names and numbers so you can make a choice. These handsets have a display, usually liquid-crystal (LCD), to show the number you have selected prior to starting the call. The calling sequence is described in greater detail later.

Demand is increasing for an innovation called Voice-Activated Dialing. Using voice-recognition techniques, a microphone in the vehicle can pick up the spoken commands and dial a number while the driver concentrates on the task of driving—an important safety feature. The same microphone, plus a speaker, can then be used to carry on the conversation after the call connection has been made. (And you thought Star Trek was fiction!)

Mobile antennas

Mobile antennas generally fall into one of three types: roof-mounted whips, roof-mounted gain antennas, or glass-mounted gain antennas.

The whip antenna hasn't found a lot of favor with most users, perhaps because a 1/4-wave whip has loss compared to a dipole or because the radiation pattern above a ground shows a high lobe-angle that, in effect, looks where most of the signal isn't.
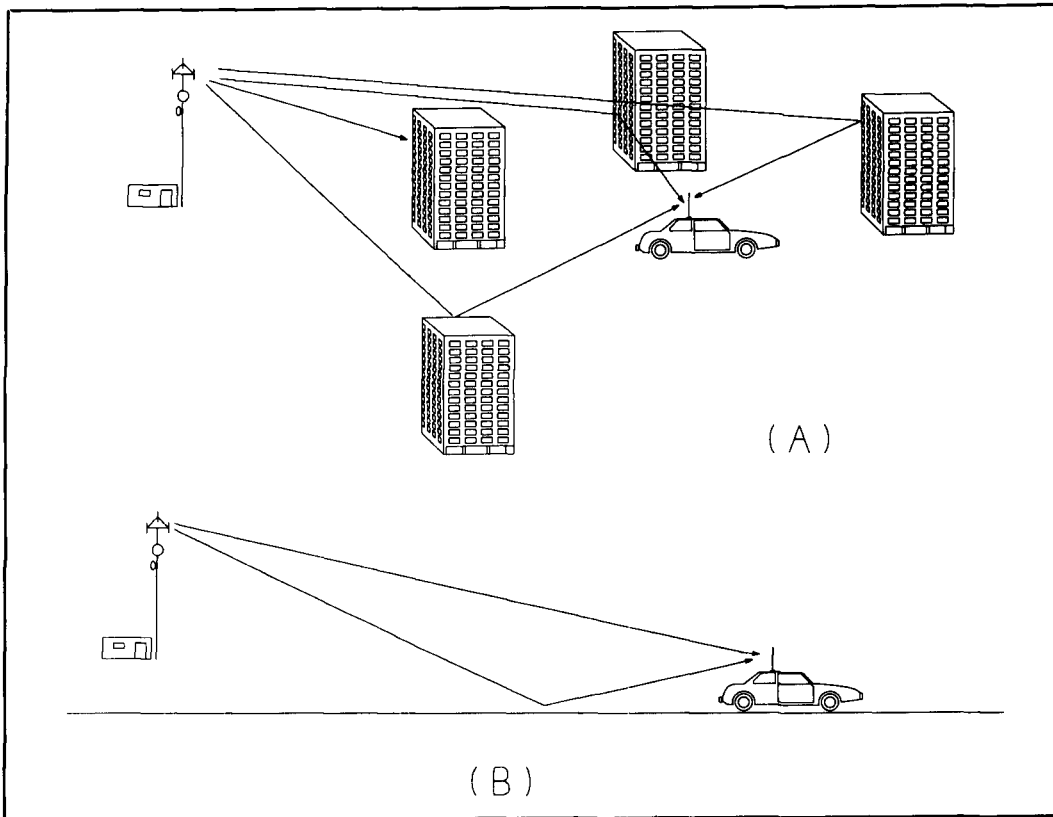
Figure 8. A mobile unit in an urban environment experiences signal flutter caused by reflections from buildings and other structures, A. Even when the signal path is line-of-sight, reflections from ground cause periodic fluctuations in received signal strength, B. When the overall signal is weak, as near a cell boundary, these fluctuations cause noise bursts that can alert the cell site and MTSO to get ready for a hand-off to the next cell.

The 1/2 or 5/8-wave antenna that is so useful at 150 and 450 MHz has a more effective radiation pattern, but becomes difficult to feed when translated to the approximate 6.5-inch length (1/2 wave at 850 MHz) for the cellular bands.

A roof-mounted gain antenna can be created by placing a phasing coil between two radiating elements. This type works well, but it requires that a mounting hole be drilled in the vehicle—something that isn't always popular with users.

The type of antenna most often seen (70 percent of the vehicles have them) is the glass-mounted gain antenna, or "pigtail." The nickname comes from the phasing coil near the center of the antenna. This antenna mounts easily on the glass of a windshield or rear window, and is coupled to the transceiver through the glass by means of a pad on the outside and a coupler on the inside, as shown in **Figure 9**.

Portables

Portable units have many of the same features that mobile units have, but are self-contained handsets with built-in transmitter/receiver units, a rechargeable battery, and an antenna. They operate at power levels of 0.3 to 0.6 watts (300 to 600 mW), which somewhat limits their range within a cell. The most common type looks like a deeper version of a telephone handset with an antenna on the top (see **Photo B**). A newcomer to the scene is the MicroTAC Lite™ by Motorola, Inc., shown in **Photo C**. It's truly a lightweight, checking in at 7.7 oz. It will provide eight hours standby time or 45 minutes of "talk time" on its internal battery. The unit folds into a package small enough to fit in a suitcoat pocket or purse. This is one of the few portables that is dual-mode compatible—it works with both AMPS and NAMPS standards.

NAMPS (pronounced en amps) is short for Narrow-band Advanced Mobile Phone System—a recent addition to the cellular world from Motorola. It uses closer channel spacing (10 kHz versus 30 kHz for AMPS), and 5-kHz deviation instead of 12 kHz.

Portable units certainly provide "take-it-with-you" personal communications, but they also suffer some problems unique to their power level and portability. More on this later.
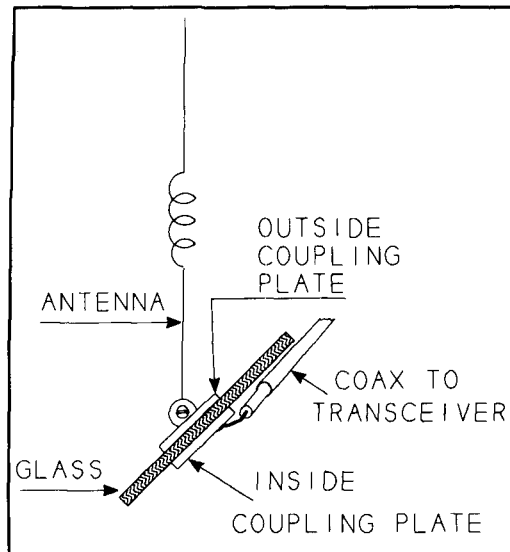
Figure 9. A glass-mounted antenna is the most popular type in use. RF energy is coupled through the glass by plates inside and outside, with no need to drill holes. The plates are held on by special adhesive. The coil near the center of the whip provides phasing that effectively creates stacked vertical elements for signal gain and a radiation pattern with a lowered elevation angle. This type is often referred to as a "pigtail" antenna.

There are also "transportable" units. These are packaged to allow easy transfer from one vehicle to another, or from a vehicle to a home, business, or boat, for example. Some obtain power from the vehicular battery (or a power supply in an office), and are designed to easily plug into a vehicle supply by a "cigarette-lighter" type connector. Another type of transportable is self-contained; that is, it has its own battery supply built in, and needs to be plugged into an external supply only when the batteries are being recharged. Both of these types provide approximately the same power level (3 watts) as a mobile unit, thus providing more range than the low-powered "portable" types.

Making a call

Whenever the mobile unit is on, the receiver monitors a control channel of the local cell. All cells transmit status information continuously, allowing the receiver to scan for the best signal and lock onto it. The control channel provides "overhead" information such as its identification, and a traffic-load indicator. Thus the mobile receiver not only knows if the signal path is good, but also if there is a channel available to make a call.

The caller enters the number to be called via a key pad on the receiver handset (or

enters a memory number, if the unit is so equipped). After the number is entered, the user presses the "SEND" key to start the call. The receiver scans again to confirm that it has the best control-channel signal, and seizes that channel. The transceiver then sends a service request to the cell site indicating that it wants to make a call. The cell site equipment forwards that information to the MTSO, which then selects a voice channel and returns the information to the cell. The cell site then forwards a channel-designation message to the mobile unit. The cell site also sets up the voice channel and transmits an "SAT" (Supervisory Audio Tone) tone on that channel.

The receiver tunes to the assigned voice channel and detects the SAT signal, which it then verifies and retransmits to the cell site. The cell site detects and verifies the SAT, and notifies the MTSO by giving it an "off-hook" condition. (An "off-hook" condition is what your local telephone office equipment detects when you pick up your home telephone handset.

Let's take a moment to look at the SAT and why it is so important. There are three SATSs available: 5890, 6000, and 6030 Hz. Each cell is assigned only one of these tones. When it transmits the SAT to a mobile, it expects to see the same tone returned by the mobile. It verifies this to be sure that the mobile hasn't inadvertently switched to another cell because of a momentary propagation anomaly. Proper identification and routing is important to the subscriber. For example, if a mobile initiated a call while near the boundary between two service areas (i.e., two area codes), and inadvertently captured a cell site on the "wrong" side of that boundary, what was intended to be a mobile-to-local call could end up being a mobile-to-other area call, and the billing would reflect long-distance rates rather than local. This whole process takes a only few microseconds and is completely transparent to the user.

The transceiver has also sent the telephone number information through the cell site to the MTSO, and the MTSO completes the call through its network—either landline or mobile-to-mobile. All this happens in a few milliseconds, of course, and the caller is blissfully unaware of the small miracle being worked by the various hardware and software components in the transceiver, cell site, and the MTSO.

If the strongest control channel indicates that it has no available voice channel for the call, the receiver scans for the next best and locks on—checking the status information for channel availability. If all control chan-

nels indicate "busy," the receiver informs the caller of that fact with a busy signal.

The "hand-off"

One of the most carefully executed operations in cellular telephone systems is the hand-off procedure. It usually takes place so quickly that the user seldom notices the slight gap (200 ms) in a word when it takes place.

The cell-site equipment is constantly monitoring the signal strength of the signal from the mobile unit. It stores and updates information on the strength of the RF signal, as well as the amount of voice degradation due to noise.

In a mobile unit that's moving across a cell at a reasonably constant radius from the site, the signal strength is somewhat steady, with occasional noise "hits" because of reflections caused by obstructions, foliage, terrain, and so on.

Another interesting procedure takes place here. If a mobile is moving toward the cell site, the signal strength will increase, and if it exceeds a threshold level, the cell site will send a digital message to the mobile telling it to reduce the transmitter output power. This is done to prevent receiver overload and/or adjacent-channel intermod.

However, when the mobile is moving away from the cell site, the signal strength decreases in proportion to the distance and speed of the moving vehicle. Superimposed on that signal-strength "envelope" (shown in **Figure 7**) are closely spaced peaks and valleys indicating obstruction and reflection-induced fading, and phase-fading caused when the distance between the mobile and the cell site changes in terms of wavelength. At a frequency of 850 MHz, a wavelength is approximately 13 inches, and a vehicle in motion intercepts peaks and valleys (analogous to standing waves) that produce a characteristic phase flutter directly related to the vehicle speed. A vehicle moving at 20 miles per hour will experience signal fades or dips at a rate of 20 per second; a speed of 15 miles per hour will produce a rate of 15 per second.

In a cellular system, the goal is to make a hand-off before it becomes annoying to the user. When the number and depth of these dips, which cause noise "hits," reaches a predetermined value, and the signal strength curve generally fits the 10-dB/decade envelope, the cell site assumes that the mobile is moving toward the boundary of its service area, and the cell and MTSO requests signal-strength information from the surrounding cells. Each cell checks and reports back to the requesting cell, which then tells

Photo B. A portable hand-set type of transceiver like this one provides communications almost anywhere you go. (Photo courtesy of Cellular One.)

the cell with the best signal to "prepare" to receive a hand-off. If the selected cell can handle the call, it will reserve a "slot" for the mobile, and informs the requesting cell of the new channel assignment. It also transmits a SAT signal on its control channel. When the new cell sends a hand-off order to the old cell, the old cell formats a message and sends it to the mobile in a burst of data over the voice channel. The transceiver responds by returning a burst of tone, turning off the transmitter for a few milliseconds, and changing to the new channel.

The old cell then turns off its transmitter and places an "on-hook" condition on the trunk to the MTSO. The MTSO switches the call to the new cell's voice trunk. Meanwhile, the transceiver has received, verified, and retransmitted the SAT. When the new cell site detects the SAT, it places an "off-hook" condition on the trunk to the MTSO. The MTSO looks at the on-hook from the old cell, the off-hook from the new cell, and concludes that the hand-off was successfully completed.

The priorities established by the MTSO are such that preference is given to a hand-off rather than to a call being initiated. The theory is that it's more disturbing to a user
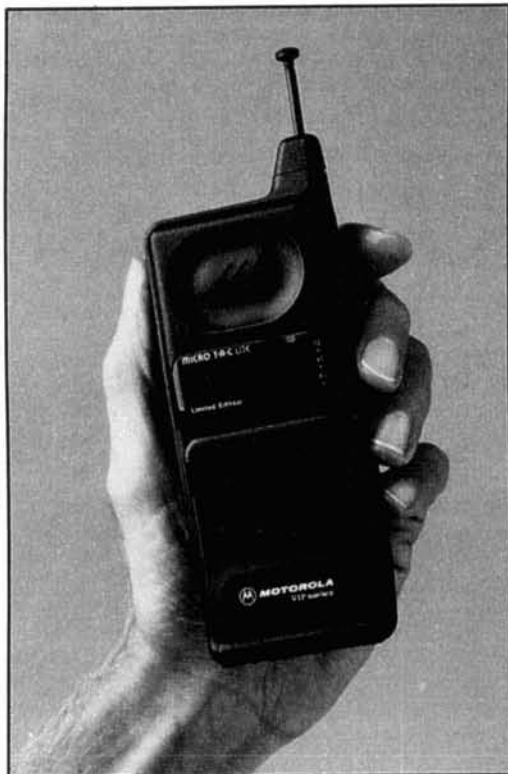
Photo C. Compact, light-weight units such as this one allow you to carry your telephone in a purse or jacket pocket. It weighs only 7.7 oz., but still provides 45 minutes of "talk time" or 8 hours standby on its internal battery. (Photo courtesy of Motorola, Inc.)

for a call in process to be dropped than it is for a user to hear a busy signal when attempting to make a call.

Hang up the phone!

This part of the procedure is the simplest of all; the user simply hangs up the handset when finished. The mobile unit isn't quite done, however. It must inform the cell site and the MTSO that the call is completed. It does this by sending a 10-kHz tone for 1.8 seconds, and then leaves the channel. The receiver will return to its scan-and-lock routine for monitoring the control channel. The MTSO will place an ON-HOOK condition on the land-line connection and transfer call-duration data to the billing section.

Calling a mobile from a land-line telephone

The sequence of events that takes place when you call a mobile unit from your home or office requires a different procedure as far as the MTSO and the cell site are concerned. We assume, of course, that the mobile unit is turned on, in which case

it is locked onto and monitoring a control channel. The MTSO receives an incoming call from the "land-line," and sends a "page" message to the cell sites to locate the mobile. The cell site then sends the "page" via the control channel. The mobile detects and identifies the page as one directed to it, and then checks to be sure it has the best control channel. It then seizes the control channel and returns a service request. The cell site relays this service request to the MTSO, which then selects a voice channel. This selection is passed back to the cell site, which, in turn, sends the channel designation to the mobile via the control channel. The cell site also activates the voice channel and transmits the SAT. Meanwhile, the mobile has tuned to the assigned voice channel and it then detects, verifies, and retransmits the SAT. This returned SAT prompts the cell site to place an "off-hook" condition on the trunk to the MTSO.

But wait, we're not done yet! Up until now, the mobile user hasn't heard a thing; the mobile set has been doing all this quickly and quietly.

The MTSO, when it sees the off-hook, sends an "alert" order to the cell site, which relays it to the mobile via a packet "burst" on the voice channel. Now, the mobile telephone rings to tell the user that someone is calling. The mobile also sends a 10-kHz tone to the cell site, which triggers the cell to send an "on-hook" signal back to the MTSO. The MTSO senses this signal and generates a "ring" tone which the land-line caller can hear. When the mobile user (call recipient) picks up the handset (answers the phone), the mobile set stops sending the 10-kHz tone, which causes the cell site to send "off-hook" to the MTSO. The MTSO then stops the "ring," and completes the connection to the land-line caller.

All this sounds complex, but all these steps are necessary to assure that each part of the system is doing its job and that the call is completed as dialed. Besides, the whole sequence takes place in a few milliseconds, and neither the land-line caller nor the mobile user is aware that anything is much different from a call from one house to another via the normal land-line "twisted pair."

Problems for portables

Portable (hand-held) cellular telephones have become a status symbol of the "90s." So much so, in fact, that a very popular item in gift shops is a "fake" handset that looks very much like the real thing. It even emits a series of tones and beeps to heighten the illusion.

The utility of the real hand-held unit is so great that many customers are using them inside their vehicles with some degree of success.

The RF field from the cell site is obviously not uniform inside a partially-shielded space such as an automobile, thus the variation of signal strength is likely to be more rapid and the depth of fading more intense. Additionally, windows with dense "tinting" attenuate the signal.

One way to alleviate these problems is to mount an external antenna, usually a "pigtail" type, on the glass of a windshield or rear window. This antenna is then coupled to a "distribution" system inside the car that improves the signal path to and from the portable. Reports indicate that sometimes this system works, but many times it has no noticeable effect. It could possibly increase the phase interference when the direct signal through the windows is at odds with that induced from the "on-glass" antenna system.

A similar problem exists when a hand-held is used within a building. Because most building materials attenuate VHF and UHF signals rather severely, successful use of a portable is often limited to the vicinity of window areas. Tall buildings, however, are a special case. The upper floors, being more directly in the main lobe from the cell-site antenna, would receive significantly greater signal coverage than the floors near ground level. Again, highly tinted windows, especially the reflective, mirror-like type, can drastically attenuate the signal.

Passive repeaters have been tried with some success. These use an antenna with considerable directivity and gain on the roof of the building, with the output of the antenna coupled to a distribution system on the lower floors. The distribution system can be a form of "leaky" coaxial transmission line, allowing radiation to fill the inner offices with a weak but usable signal.

Active repeaters do work, but are limited to a few channels for simplicity.

Another solution is to place a miniature cell inside the building, using the inherent attenuation of the building material to prevent interference to the outside world. By carefully tailoring the power output and the signal distribution system, most of the signal can be contained inside the building, with only a small "leakage" area outside. Thus, service is provided inside the building without impacting the neighborhood cell sites and their frequency plan.

This would, of course, require a large enough population of in-house users to justify the cost of setting up the miniature cell.

Things to come

The cellular industry is a very fast-growing one, and several groups are looking at ways to provide more service to more people. One idea being explored is the concept of "micro-cells." This system would work in a manner similar to the current cellular network, but the cells would be limited to very low power, and cover only a few hundred feet. By saturating a small area, like a shopping mall, airport, or an office building, with low-powered signals, they hope to provide a market for smaller, low-powered hand-held units that can be manufactured inexpensively. A system of this type is in the early stages of market expansion as this is being written. Called CT2, it currently involves a small "home" base station for use in a home or office, and a self-contained handset. This same handset can be used through "public" base stations located in malls, airports, etc.

With a home base station, CT2 can originate and receive calls the same way that the 49-MHz cordless telephone in place today does, except that the transmit and receive frequencies are in the 800 to 900-MHz range. This provides far less interference and much greater privacy than that offered by the current 49-MHz system. As configured today, the CT2 system offers only outgoing-call capability through the "public" base stations in malls and other public places. Hand-off and incoming-call capabilities are not yet in place. The next generation, however, could include some type of "paging" system to tell you to call back, similar to the pocket pagers in wide use today.

Whether this system will ever provide the full range of services that standard cellular systems do isn't clear. Technologically, it's possible; the problem is mostly regulatory, but some industry committees are studying the issue.

World-wide personal communications could be in our future as well. Motorola has proposed a system it calls IRIDIUM®, which involves a network of satellites in low orbit to provide, initially, paging capability to reach anyone anywhere in the world. The next step could be to expand the one-way capability to two-way, and configure the system to work in conjunction with the cellular structure now in place.

Digital communications

In an effort to increase efficiency and provide service to more subscribers, the industry is looking into digital techniques. Many modulation schemes, such as Quadra-

ture Phase Shift Keying (QPSK), Binary and Differential Binary Phase-Shift Keying (BPSK and DBPSK), Multiphase-Shift Keying (MPSK), and variations on these, are being investigated for efficiency, voice quality, and noise immunity. Also under consideration is a method of sampling that will allow two or more voice channels to use the same carrier at the same time. This is referred to as Code Division Multiple Access (CDMA), and is a type of spread-spectrum system.

Time Division Multiple Access (TDMA) is another digital encoding method scheduled to undergo field tests in 1991 and 1992. This system divides a time unit, like a second, into slices, then assigns a part of a conversation to each slice. For example, if each second is divided into 200-millisecond slices, and each slice carries a piece of a conversation, then five conversations can be conducted at the same time. The receiver will separate these "slices" of recovered audio and process them into a replica of the original. With a high enough sampling rate (thin slices) and well-designed filters to reshape the recovered audio, the user doesn't notice any degradation in quality. The system currently being field tested is designed to permit three conversations to take place at the same time, thus increasing the trafficload capacity of a channel. As the system is refined, the number of conversations per channel will increase, greatly alleviating the demand for spectrum.

Roamers

Some changes of great interest to travelers are taking place as this is being written. One involves making it easier for a "roamer" to be reached anywhere in the United States. It's more of a software issue than anything else. Both wire-line and non-wire-line licensees are implementing systems to find a mobile that's not in its home district. This can be done to a limited extent today, if the user "pre-informs" his home system of his approximate itinerary. When the home system doesn't receive an answer from the mobile unit, it switches the caller to another number that can play back a recorded message giving information on how to reach the mobile subscriber.

The improved system will allow a roamer to check in to a nearby system during the trip. The local system can then use the mobile's NAM (Number Assignment Module) information to inform the roamer's home system of the unit's current location. This information can be used by the home sys-

tem to forward calls to the roamer's current location.

The NAM mentioned above is a memory area in the mobile (or portable) unit that stores the electronic serial number of the equipment plus its assigned access number. It also allows a mobile unit to be registered to more than one number. This information is sent to the cell site when the mobile unit sends a service request or is interrogated by the cell.

As I wrote this, I noticed an announcement that McCaw Cellular Communications has linked their systems together into a North American Cellular Network with the express purpose of helping users to be reached easily when operating away from their home base. They are now working on a system to provide an interface between dissimilar equipment. Eventually, they hope to link with other systems. McCaw Cellular Communications is the parent company of Cellular One, and others.

Another system called RoamAmerica has been provding a unit-registering and call-forwarding service tailored to roamers in some parts of the country.

The CTIA has recently formed a subsidiary called CIBERNET CORP. to help roamers by coordinating inter-carrier transactions. This allows clearinghouses to submit roamer billing information to one entity. Their CIBER Record contains information on billing practices and serves as a guide for transactions.

Forbidden territory

Modern cellular systems permit you to go almost anywhere and take your communications with you, but there is one thing that is absolutely not permitted. YOU MAY NOT USE A CELLULAR TELEPHONE WHILE ON AN AIRCRAFT IN FLIGHT!

The reason is very simple—wide-area interference.

Cellular's success is based on frequency reuse with well-spaced separation, and the coverage area of the cells is tailored to prevent interference between cells that use the same frequency. With a cell of, say, 5 miles in diameter, the next cell that reuses the frequency would be 7 cells away, or 35 miles. An aircraft at 8,000 to 10,000 feet altitude will have a coverage area of nearly 180 miles, and the radio horizon expands as altitude increases. Every cell within range would pick up the signal and try to process it, causing massive interference. The FCC is quite blunt about phoning from an aircraft: "Just don't do it!"
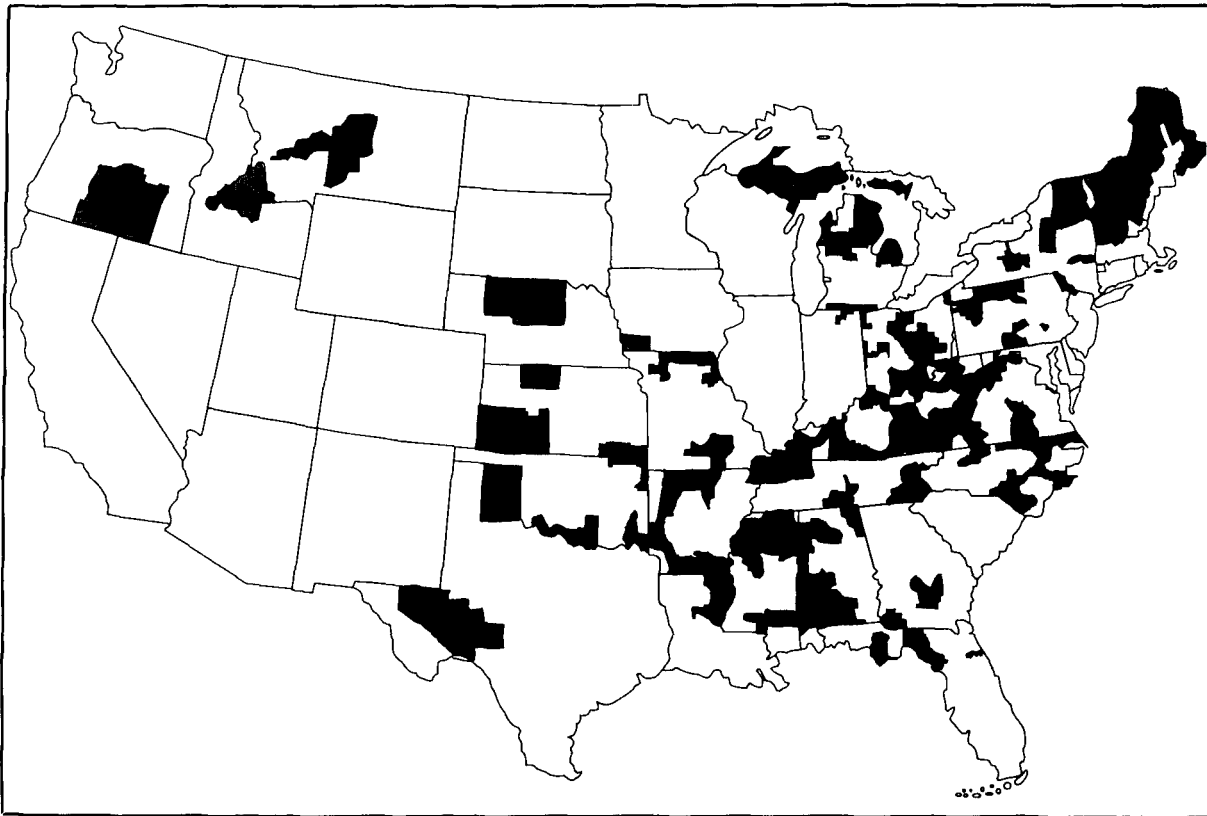
Figure 10. *Cellular communications is available in most of the United States, as shown by the white areas in this map. Grey areas represent Rural Service Areas (RSAs) that are not yet in service, although all are licensed at this time.*

There are other means of making telephone calls on most modern passenger jets, so your best bet is to leave your portable cellular telephone packed until you arrive at your destination.

Privacy

One thing cellular users should always keep in mind is that anyone can listen in. This has been true since the days of the MTS and IMTS systems and is also true of our current 49-MHz cordless telephones. As technology moves upward in frequency, scanner-equipment manufacturers match the expansion, so there is no "private area" where a call is safe from the curious.

There are telephones that have voice-scrambling capabilities, but at a premium price. GTE Mobilnet has been working on a method that will allow a user to press a button to initiate scrambling on either mobile-to-landline or mobile-to-mobile calls. This system does require special circuitry in the mobile unit.

One hope for the future is that the digital techniques described earlier will be able to provide some measure of security. Until then, use caution in your cellular communications.

Where are the cells?

While doing research for this article, I asked the service manager of the local Bell-South Mobility office, "How far can I travel and still remain in contact...say, from Miami to....?" His answer was, "How about Montreal?"

This is a good indication of the development of cellular telecommunications today. The white areas on the map, **Figure 10,** show that most of the United States is covered by either MSA (Metropolitan Statistical Area) or by Rural Service Area (RSA) systems. The grey areas on the map represent RSAs that were not covered as of mid 1991, but as of this writing, the FCC has granted licenses for these remaining RSA pockets (all the MSAs were already licensed). There are 306 MSAs and 428 RSAs in the United States. The entire map should be "white" within a few months after you read this.*

Hawaii, Alaska, most of Puerto Rico, and significant portions of Mexico and Canada near their U.S. borders are also served by either MSA or RSA systems.

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*The map in **Figure 10** is based on a large wall map obtained from the CTIA in Washington, D.C.

Many parts of the world are developing cellular systems, some of them following their own standards, and others following the AMPS standards used in this country. AMPS (Advanced Mobile Phone Service) is the name used by AT&T and Bell Laboratories when it developed its cellular technology.[5] This standard has been modified slightly as the industry has grown, but it still serves as a yardstick to assure that cellular equipment will work in all service areas.

Among the nations using the AMPS standard for their cellular systems are: Canada, Mexico, Australia, Argentina, Bahamas, Bolivia, Brazil, Brunei, Cayman Islands, Chile, Columbia, Costa Rica, Dominican Republic, Egypt, Guatemala, Hong Kong, Hungary, Indonesia, Israel, South Korea, Malaysia, Philippines, New Zealand, Pakistan, Paraguay, Peru, Samoa, Singapore, Taiwan, Thailand, Uruguay, Venezuela, Virgin Islands, and Zaire.**

Many countries of the European community are using either their own standards or one called GSM (Group Speciale Mobile). The frequency bands in use are not at all uniform; some are still in the 150-MHz area, others at 450 MHz, and a few are using 850 MHz.

Conclusion

The cellular telecommunications industry is large (6.4 million subscribers in mid 1991), complex, and growing rapidly. Technologists are developing many new ways to make the service more useful and more attractive to an ever-widening public. By the time you read this, many new services will be available; the industry is definitely not static. One thing that stands out in my research for this article: Designers of cellular systems haven't invented any new techniques, but rather have applied old familiar ones in new and fascinating ways. One example is the use of computer technology to look at a received signal and determine when a mobile unit will soon be out of range, thereby initiating the hand-off procedure. The pieces were already there; a new industry provided the impetus to assemble them into a service called cellular.

Acknowledgements

When doing research for an article like this, one inevitably meets a variety of helpful people who are glad to fill in any knowledge gaps. I would like to express my appreciation to some who were particularly helpful:

Sue Labak of Andrew Corporation; Jacqueline Stewart of the West Palm Beach office of Cellular One; Greg Bonner of BellSouth Mobility in West Palm Beach; David Pinsky of Motorola's Cellular Support Group, Schaumburg, Illinois; Greg Saeli of Motorola's Wireless Enterprize Systems; and Michelle Solomon of the Cellular Telecommunications Industry Association.

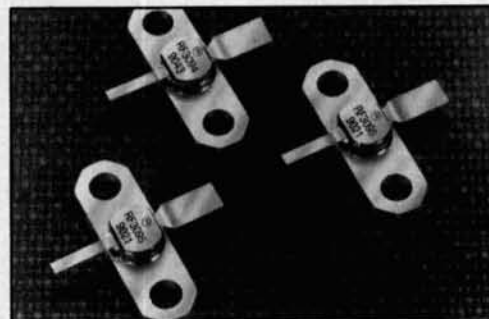The map in **Figure 10** was drawn using Corel Draw® and Microsoft Windows®. ∎

**Foreign-system information derived from the 1991 annual report of the CTIA.

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AN INTRODUCTION TO FIBER OPTIC-BASED LOCAL AREA NETWORKS

The use of fiber optics in computer networks

The fiber optic tutorial in our summer 1991 issue touched on the basics of fiber optic technology. Here's an application that combines fiber optics with computer-based communications. Ed.

T his article is intended to serve as an introduction to computer networks. It concerns the Waterways Experiment Station (WES) local area network that connects over 1000 personal computers, 40 VAX computers, several CDC mainframes, and a CRAY Y/MP supercomputer. The fiber optic-based network accommodates TCP/IP, DECNET, and Novell software communication protocols. The fiber optic ethernet allows for speeds up to 10 million bits per second (Mpbs). A recently installed Fiber Distributed Data Interface (FDDI) network handles aggregate traffic up to 100 Mbps. Remote access is handled by indial and outdial modem pools connected to terminal servers, and by 56 Kbit/second routers connecting the WES network to the Internet.

Fiber optic ethernet

The 10-Mbps ethernet was installed in 1989. **Photo A** shows an aerial view of WES. Twenty-one buildings are connected by the fiber-optic backbone. Only two strands, out of a bundle of 24, of fiber optic cable are needed for the hop from building to building, because the fiber backbone is designed in a ring rather than a star topology. The fiber is graded index multimode fiber optic cable (62.5/125 micron with an aperture of 0.29). Optical windows of transmission in the fiber cable occur at wavelengths of 850 and 1300 nanometers. The computers inside each building are connected by copper wire, usually 50-ohm RG-58 cable. They are then bridged to the fiber optic backbone at a single point within the building by an inexpensive electrical-to-optical converter, or transceiver. Though the fiber optic network spans over 8 miles, any computer can access any other computer as quickly and conveniently as if it were accessing data from its own local hard drive.

Basic ethernet operation

Ethernet is basically a simple bus system (see **Figure 1**). A computer with data to transmit, will first listen for 9.6 microseconds of silence before attempting to transmit the packet onto the bus. If two computers try to transmit simultaneously, a data collision occurs. The transceivers on the network inform the computers involved that a collision has occurred. Once a collision is confirmed, both computers back off
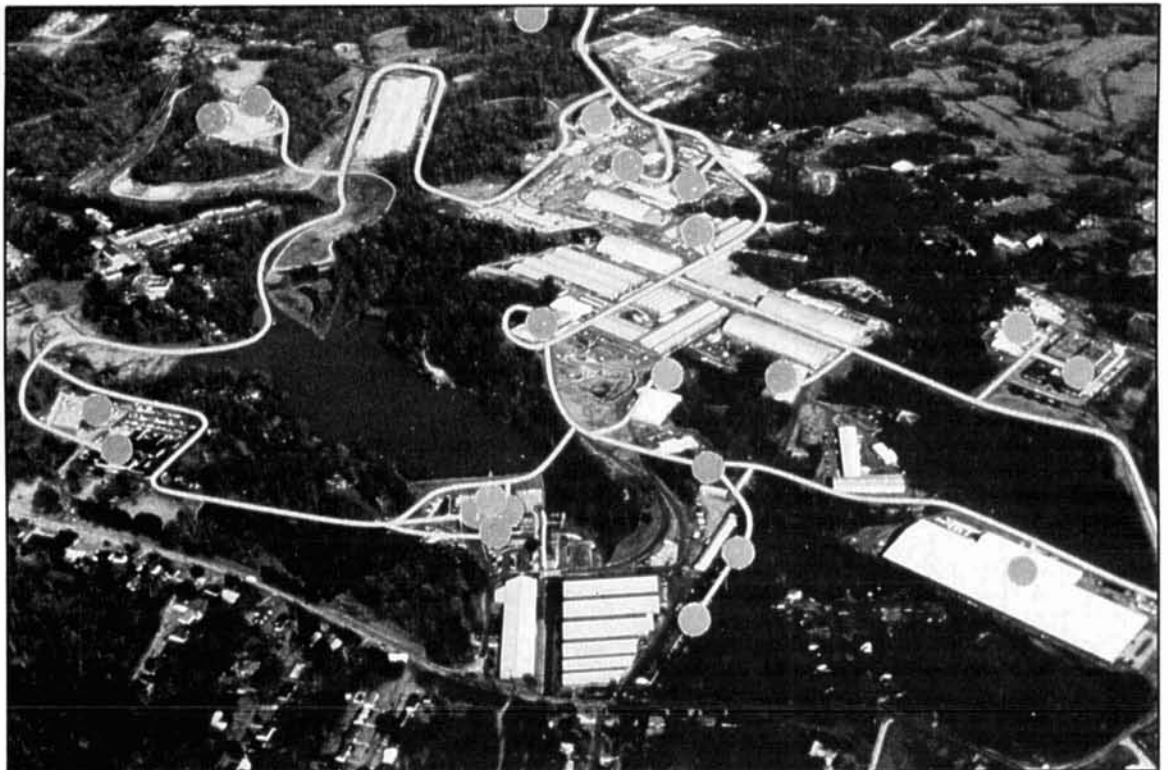
Photo A. Aerial photograph of the Waterways Experiment Station. The fiber optic route is shown by the lines connecting the buildings. The circles represent buildings that are connected to the fiber backbone.

for a random amount of time before attempting to retransmit. The computer that transmits first has control of the local ethernet network (bus) for the duration of the packet.

Photo B shows the proper use of a T-connector to allow a computer to tap into a 50-ohm RG-58 based ethernet bus. This method lets the network preserve a 50-ohm characteristic impedance at each node. Although it would be tempting to install the T-connector in the ceiling and run an RG-58 drop cable down to the computer, the drop

cable would cause a characteristic impedance mismatch that would result in the complete failure of the ethernet.

Network information is encapsulated in packets as it flows from computer to computer. The structure of the ethernet packet is shown in **Figure 2**. The bits in the packet are Manchester encoded at 10 million bits per second. The first 64 bits are called the preamble and synchronize the circuitry in the receiving computer. The next six bytes contain the ethernet hardware address of the sending station, followed by another six
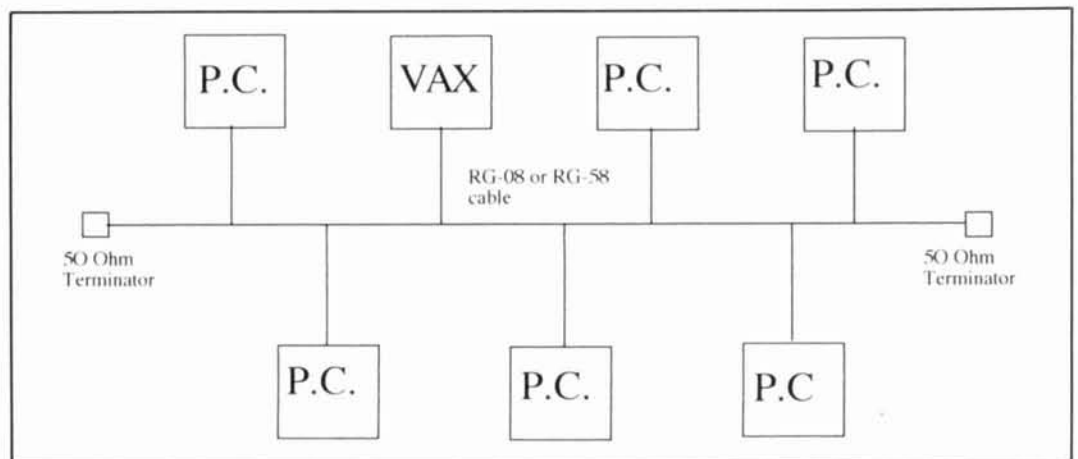
Figure 1. Basic ethernet system configuration.
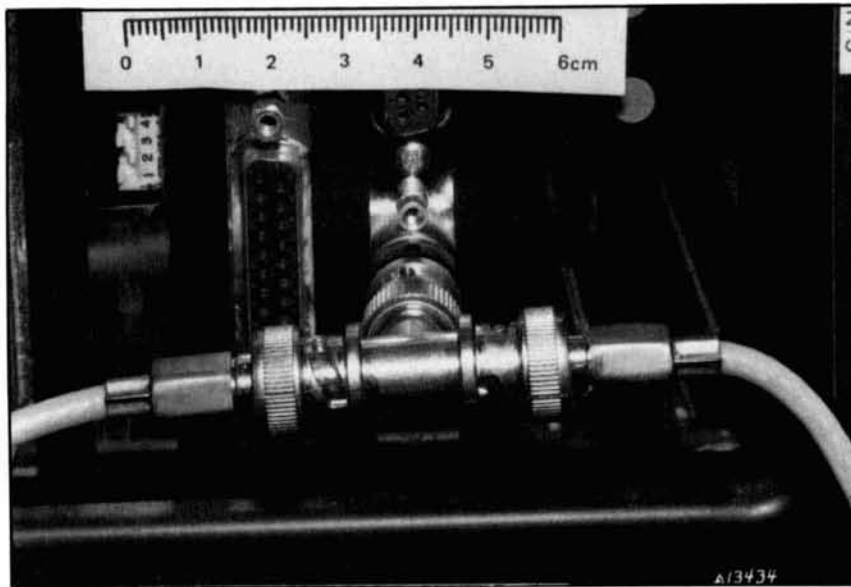
Photo B. For proper connection to an ethernet, the T connector must be attached directly to the back of the PC.

64 BIT PREAMBLE	6 BYTE DESTINATION ADDRESS FIELD	6 BYTE SOURCE ADDRESS FIELD	2 BYTE TYPE FIELD	46 - 1518 BYTE DATA FIELD & HIGH LEVEL PROTOCOL FIELD	4 BYTE ERROR CHECK

Figure 2. The source and destination address refer to the actual hardware address of the ethernet board inside the computer.

bytes for the ethernet address of the receiving station. A two byte type field follows. The data field contains the actual data, or message. This data field also carries its own protocol that determines if the packet is a UNIX TCP/IP packet, DECNET packet, or Novell IPX packet. The high level protocol in the data field contains its own format for information about the network addresses of the transmitting and the receiving stations. This is independent of the ethernet hardware addresses in the source and destination frames of the packet.

Bridging data onto the fiber-optic backbone

Figure 3 shows a typical ethernet within a building, bridged onto the fiber optic ethernet backbone. All computers inside the building receive, or listen to, all packets. However, the destination ethernet address in the packet must correspond to the actual

hardware ethernet address of the computer, or the computer will disregard it. A special device, called a spanning tree bridge, listens to all packets within a building. If the destination ethernet address of a packet specifies a computer in another building, the bridge passes that packet to the electrical-to-optical transceiver that transmits it on the fiber optic backbone. Bridges in other buildings listen to all packets on the fiber backbone. These bridges will only allow those packets with a destination address that corresponds to some computer within a building to enter that building.

The spanning tree bridges are fully automatic. They consist of port 1, port 2, and an off-on switch, and cannot be programmed. The bridges learn about the location of computers on the network by tabulating source and destination ethernet addresses of packets that appear on port 1 and port 2. Each bridge automatically exchanges its information with all other bridges in the
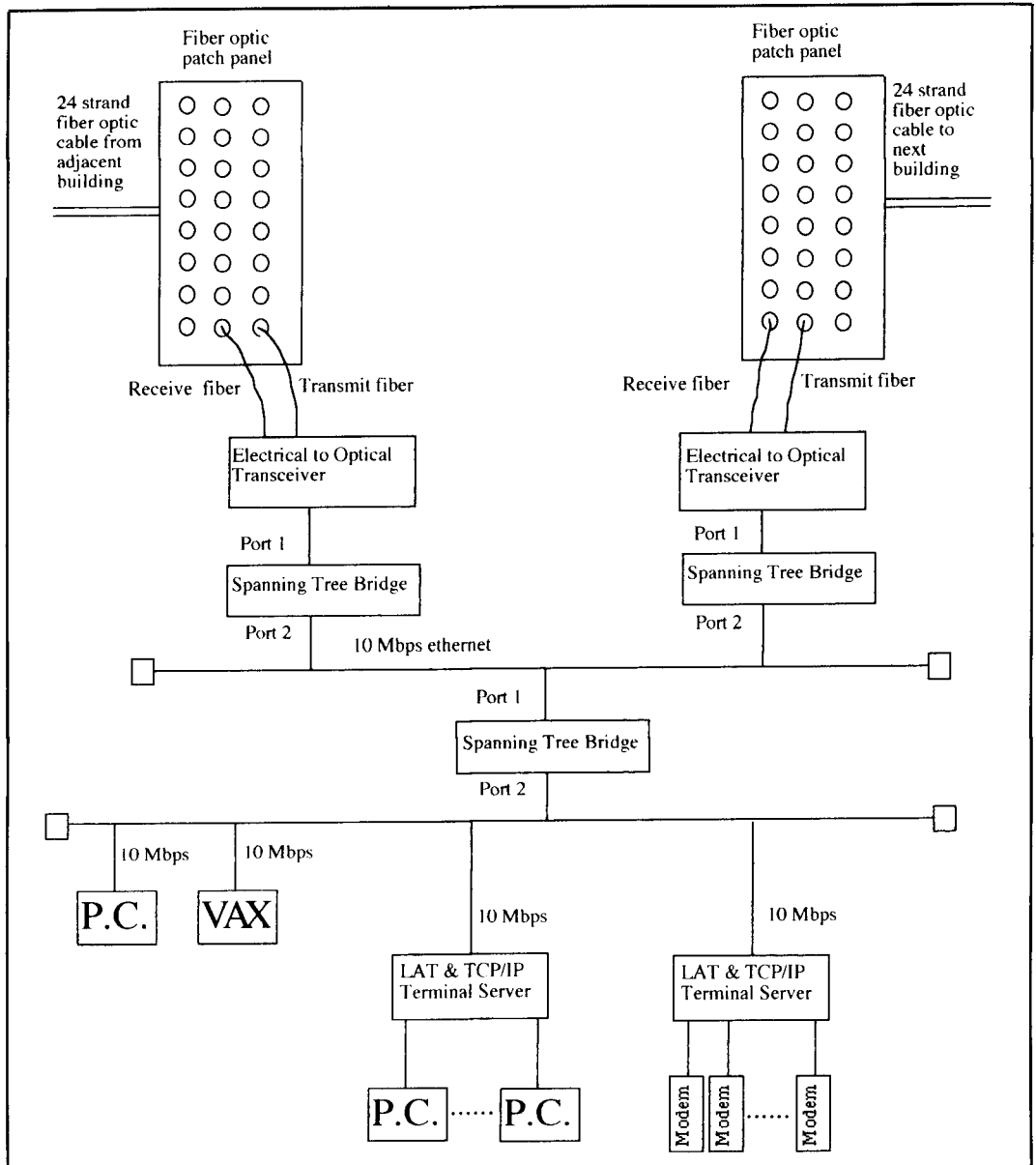
Figure 3. Conceptual diagram of a building ethernet.

network, until every bridge knows in which building each computer is located. The bridges then pass, or block, packets based only on the ethernet source and destination address, ignoring any address information that may be contained in the high-level protocol (data field). The ethernet network therefore achieves protocol independence by routing packets based on the ethernet address, rather than the address given by the high-level protocol inside the data field.

Photo A shows that the backbone is composed of two rings. Ethernet isn't compatible with rings because, in theory, a packet could flow around a ring forever. Therefore, the bridges choose one segment in each ring that will be disabled. They will use this disabled segment as a backup in case

any other segment is destroyed, or fails. If a segment failure occurs, the bridges will automatically activate the redundant segment. The switchover happens so fast that a user transferring a file won't even notice a glitch. The bridge then notifies the monitoring station of the fault.

High level software protocols

TCP/IP, DECNET, and Novell are software protocols that operate on the ethernet hardware. A personal computer operating with TCP/IP software can only communicate over the network to another computer running TCP/IP. The same is true for DECNET, Novell, and other high level protocols.

TCP/IP stands for Transmission Control Protocol/Internet Protocol and was originally a Department of Defense standard for networking computers. TCP/IP software is available for every computer made and is also built into the UNIX operating system. At WES, many personal computers on the network run TCP/IP under MS-DOS to communicate with UNIX mainframes. TCP/IP also allows the user access to the Internet, which provides connections to several hundreds of thousands of computers around the world.

Primarily, TCP/IP lets the user place his PC in terminal emulation mode and log onto a remote mainframe. File Transfer Protocol (FTP) is part of the TCP/IP standard and allows users to transfer files between computers. Electronic mail, via TCP/IP, can be sent to any of the millions of network users on the Internet, without requiring a User ID on the destination mainframe. There are many excellent freeware TCP/IP software packages for the personal computer.

DECNET stands for Digital Equipment Corporation Network and consists of many protocols including LAT (Local Area Transfer), CTERM, and NFT. DECNET has undergone several recent name changes, including PCSA and Pathways. I've used the name DECNET here in a generic sense to show that it's a product of Digital Equipment Corporation.

Personal computers must use DECNET software to communicate with VAX computers running DECNET. Both machines must use the same DECNET protocol; a PC running only DECNET LAT can't communicate with a VAX running only DECNET CTERM. The DECNET software allows remote terminal emulation, file transfer, and virtual MS-DOS style hard disk emulation of a VAX computer.

Novell is useful primarily for PC to PC communications, through a PC fileserver. In MS-DOS, drives A through E are designated as they are always designated—as local drives. Under Novell, drive F would switch to the Novell fileserver that could be configured to have several Gbytes of storage. The F drive can be partitioned so part of it is shared between all Novell users on the network, and parts of it contain private areas for the network users.

Network printing is very convenient with Novell. Utilities allow any printer connected to any PC to serve double duty as a personal printer and as a network printer. Some utilities will support up to 40 distributed network printers through a single fileserver. A terminate and stay resident pro-
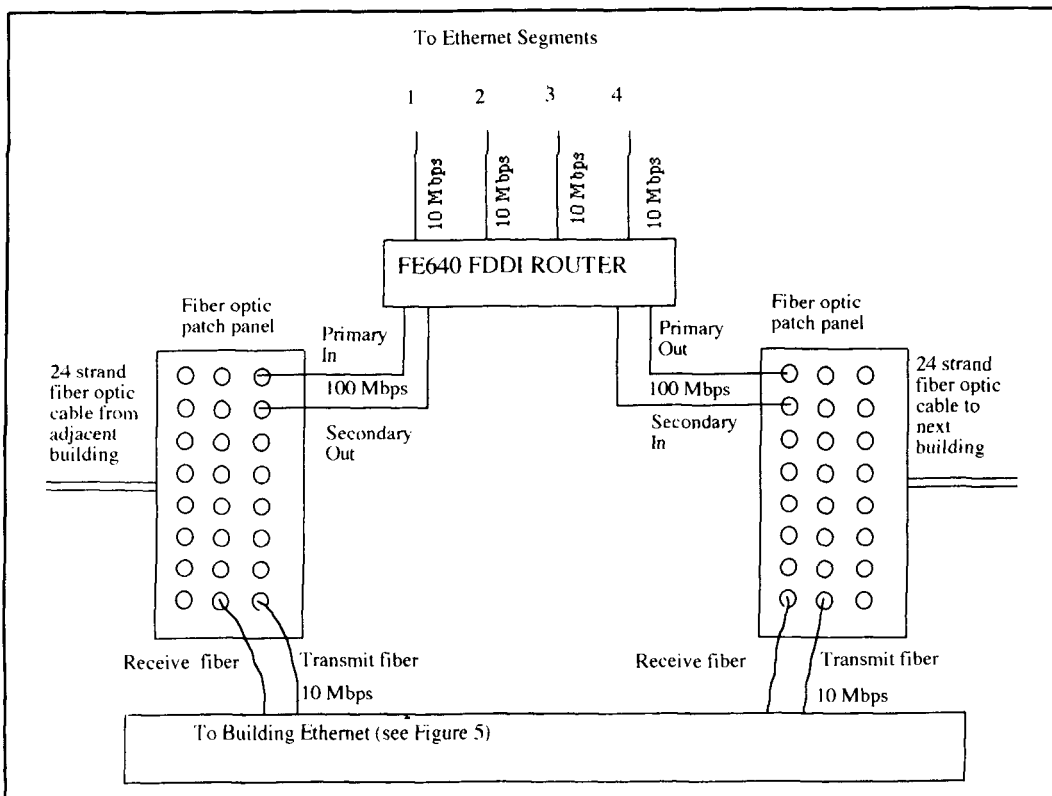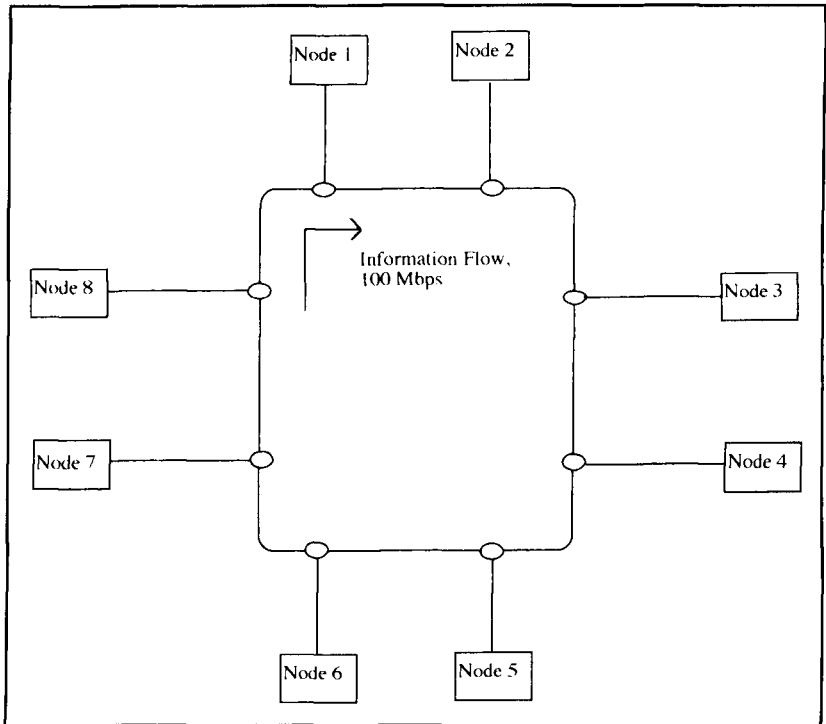

Figure 4. FDDI interface for building networks.

Figure 5. General FDDI token ring topology. Each attached node represents a FE640 interface in a separate building.

gram inside one user's PC lets the printer attached to his PC also receive files over the network. This operation is completely transparent to the user, who can continue to use the printer as if it were solely a stand-alone system.

A number of third-party vendors supply utilities that work on Novell networks. Some of the more powerful utilities allow users to view screens on other PCs, access hard disks on other PCs, and even reboot PCs remotely. Although the Novell IPX is incompatible with TCP/IP, there's a growing tendency in newer versions of Novell software to combine IPX with TCP/IP. Novell Version 3.11 allows TCP/IP tunnelling for IPX packets, so Novell networks can be interconnected across the Internet.

Given an existing ethernet, it's a minimal investment to add a single Novell fileserver that can provide all network users with these additional features. Although IPX drivers are required on each PC to access the Novell fileserver, there's no cost associated with the software for the client PCs, and a single fileserver will support up to 250 users.

Gateways

Gateways perform software protocol conversion. A mainframe computer may run both DECNET and TCP/IP. Users with only TCP/IP software may communicate with DECNET services by first connecting to the gateway computer, then by initiating communication from the gateway computer to the final DECNET destination.

Routers and brouters

Routers differ from bridges in that a router makes a decision to transmit or block a packet based on the address information given in the high-level data field. A router ignores the ethernet source and destination address when making routing decisions. Routers are protocol dependent; a TCP/IP router can't route an IPX packet. Devices that route TCP/IP packets and bridge other protocols are sometimes referred to as brouters.

Terminal servers

PCs connected asynchronously through their serial ports communicate at 300, 1200, 2400, 9600, or 19,200 bits per second. No special communication software or hardware is required in the PC for a terminal server link, because the TCP/IP and DECNET protocols are built into the terminal servers. The PC may be direct connected to the terminal server, or may dial into a modem pool connected to the terminal server. A single modem pool on the terminal

server can be configured to support both in-dial and outdial traffic, while retaining autobaud over the standard asynchronous range of 300 to 9600 bps.

The Fiber Distributed Data Interface network (FDDI)

The FDDI network at WES is a separate, but parallel, backbone. It forms a ring that connects nine buildings. A typical building configuration is shown in **Figure 4**. Users in each building may choose to attach either to the fiber optic ethernet backbone or the fiber optic FDDI backbone. The FDDI backbone is connected to the ethernet backbone through a SUN 4/280 computer acting as a TCP/IP router between these two networks.

The ring configuration allows the backbone to absorb a break in a segment of the fiber optic cable and still continue to operate. A network monitor is alerted while the FDDI router seals the break by wrapping Primary In to Secondary Out. Users won't be affected if there's a single break; however, multiple breaks will cause the FDDI network to split into separate disjunct networks.

Most of the previous discussion of ethernet also applies to FDDI because the FDDI NSC FE640 router fans out into four ethernet segments. The FD640 router can route TCP/IP, DECNET CTERM (for SET HOST operations), and DECNET NFT (for file transfers and VMS mail). The IPX protocol and DECNET LAT protocol aren't currently being bridged over the FDDI network, although this capability will be available with future upgrades of the FDDI hardware.

The FDDI backbone is a token ring network. The nodes in **Figure 5** would represent the FE640 routers. A single token circulates from node to node. If a FE640 needs to transmit data to another FE640, it attaches a packet to the token. The receiving FE640 sets the acknowledgement bit on the token, but doesn't strip off the packet. When the packet returns to the originating FE640, it reads the acknowledgement bit. If the acknowledgment bit is set, it resets the acknowledgement bit and strips off the packet.

Summary

The National Center for Supercomputer Applications produces a public domain TCP/IP software package for personal computers. This software can be downloaded using anonymous FTP over the Internet to host 128.174.20.10. The NCSA package comes with C source code and compiled code. It has drivers that support the simultaneous use of Novell IPX and TCP/IP on the network without having to reboot the PC to switch between these two protocols.

Both the NCSA TCP/IP and a LAN tutorial diskette are available from the author for a $25 shipping and handling fee per program. The LAN tutorial is a collection of documents that discuss the theory behind FDDI, MIL1553, Hyperchannel, and ethernet networks. A simple network modeling tool is provided for ethernet simulations. ∎

PRODUCT INFORMATION

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MICRO-COMPUTER-BASED INSTRUMENTATION SYSTEMS

An introduction to microcomputer-based instrumentation, with an emphasis on virtual instrumentation technology
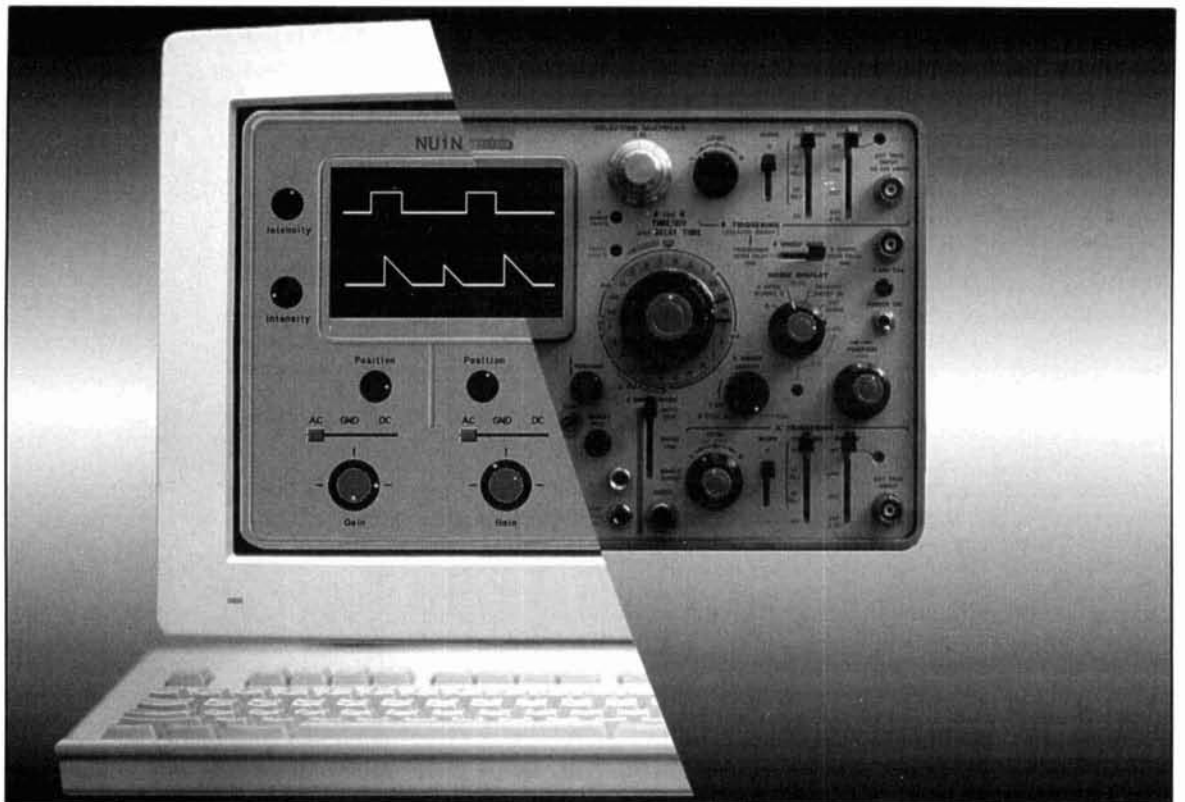
For over a decade, the personal computer (PC) has been the driving force behind tremendous gains made in text processing, desktop publishing, database management, and digital communications (PACKET radio and microprocessor-controlled transceivers, for example). PCs, when combined with affordable and powerful digital signal-processing (DSP) hardware, including inexpensive analog-to-digital (A/D) converters, now allow amateurs to generate, record, and manipulate signals to an extent once possible only in a few highly sophisticated laboratories.

Advances in instrumentation software that provide graphic user interfaces, along with the introduction of affordable and powerful plug-in data acquisition boards, have created an entirely new class of instrumentation. These systems, commonly referred to as "virtual instruments," provide a functional equivalent to that of stand-alone instruments, but are defined entirely in software. Virtual instruments now give superior performance at reduced cost, and are an attractive alternative to conventional stand-alone instruments.

Introduction

Instrumentation facilitates the advancement of science and technology by providing a means of quantifying our world. Whether they are determining the spectral purity of a signal, resistivity of a component, or periodicity of a waveform, amateurs involved in the design, testing, or repair of communications equipment share the need to record, transform, and analyze data once it has been acquired. Although many methods for obtaining this data are relatively simple (testing a resistor out of circuit, for instance), data acquisition can be quite a challenge when one is attempting to diagnose a modern microprocessor-controlled digital transceiver. Clearly, the selection of the most appropriate instrument for the task at hand is crucial.

Fortunately, the same microcomputer technology that makes the modern, ultra-compact transceiver possible can also be used to help us understand the complexities of its operation. At the most basic level, data acquired traditionally with multiple, independent instruments (a storage oscilloscope, spectrum analyzer, and digital multimeter, or DMM, etc.), can be manipulated and analyzed in a generic spreadsheet or a problem-specific program executed on a PC. While this level of computer technolo-

gy may represent a step forward when compared to the paper-and-pencil approach, it is nonetheless far from optimum. For one thing, this method requires that the raw data first be entered by keyboard into a computer before processing can begin. Manual entry is not only time consuming, but also a potential source of errors. One obvious solution to this dilemma has been to bypass as many manual operations as possible by connecting the data-collecting instrument directly to a computer.

Personal computers and peripherals

Early attempts at computer-based instrumentation relied on large, expensive, and difficult-to-use computers, connected to discrete data-collecting devices through ad hoc interfaces.[1,2] Today, more and more data-acquisition systems are taking advantage of the flexibility, local-processing power, storage capabilities, and uniform-user interfaces provided by the compact and affordable PC. Although there's considerable overlap in the function and design of modern microcomputer-based instrumentation, on a physical level most systems can be classified into three categories:[3] stand-alone instruments connected to controlling PCs by RS-232 or IEEE-488 interfaces; plug-in analog input/output (I/O) cards; and, less commonly, bus-extension systems (see **Figure 1**). Bus-extension systems, which rely on a bus-extension card linking the PC to a box containing various data-acquisition cards, are less likely to be used by amateurs because of their cost and limited availability.

Computer-aware stand-alone instruments

One approach to harnessing the power of a desktop microcomputer is to create stand-alone instruments with internal microprocessors and RS-232 or IEEE-488 interfaces for *optional* communications with a PC. The optional interface may simply provide the ability to download data from the instrument to a PC, or may support complex two-way communications that allow the PC to specify instrument range, sampling rate, and so forth, in addition to data downloading.

The current trend toward local preprocessing of the raw data as much as possible, before sending the results to a PC, minimizes the time and central processing unit

(CPU) resources spent on communications and computation.[4] For example, a computer-aware instrument, capable of using its own internal microprocessor to calculate the average or RMS voltage of a rapidly fluctuating signal, is much more efficient than one that must send all of the raw data to a PC for processing. In the first case, only a single value must be downloaded to the PC; the second could involve thousands of data points. Fortunately, instrumentation manufacturers have followed the general trend in the electronics industry toward increased use of on-board microprocessors.

IEEE-488 and RS-232 standards

One really cannot discuss interfacing an instrument to a PC without considering the nature of the interface. The most popular protocols used for communicating between stand-alone instruments and PCs are the IEEE-488 and RS-232 standards. Strong points of the inexpensive and ubiquitous serial RS-232 link, a standard feature on most PCs, include the ability to transmit information over very long distances; for example, via telephone lines from remote locations. On the other hand, an RS-232 link

suffers from limited bandwidth, lacks true practical standardization, provides a very noisy interface, tends to lose data, and is capable of connecting only two devices.[5] The maximum bandwidth of only about 10,000 bits/second can be increased with the addition of buffer memory and direct memory access (DMA), but this tends to be expensive. Several companies (Personal Computing Tools is one) have addressed the RS-232 "nonstandard" by offering smart cables that can be configured to provide a variety of handshaking and control-line assignments.

Parallel IEEE-488 interface

The parallel IEEE-488 interface, a modification of Hewlett-Packard's General-Purpose Interface Bus (GPIB), overcomes many of the limitations inherent in the slower, serial RS-232 standard. It's not only resistant to noise, but can support up to 15 peripherals simultaneously, with a bandwidth approaching 350 kHz.[5] Virtually all laboratory instruments manufactured after 1980 have provision for an IEEE-488 interface. Limitations of this otherwise excellent interface include a maximum system cable length of 20 meters, and the need to install
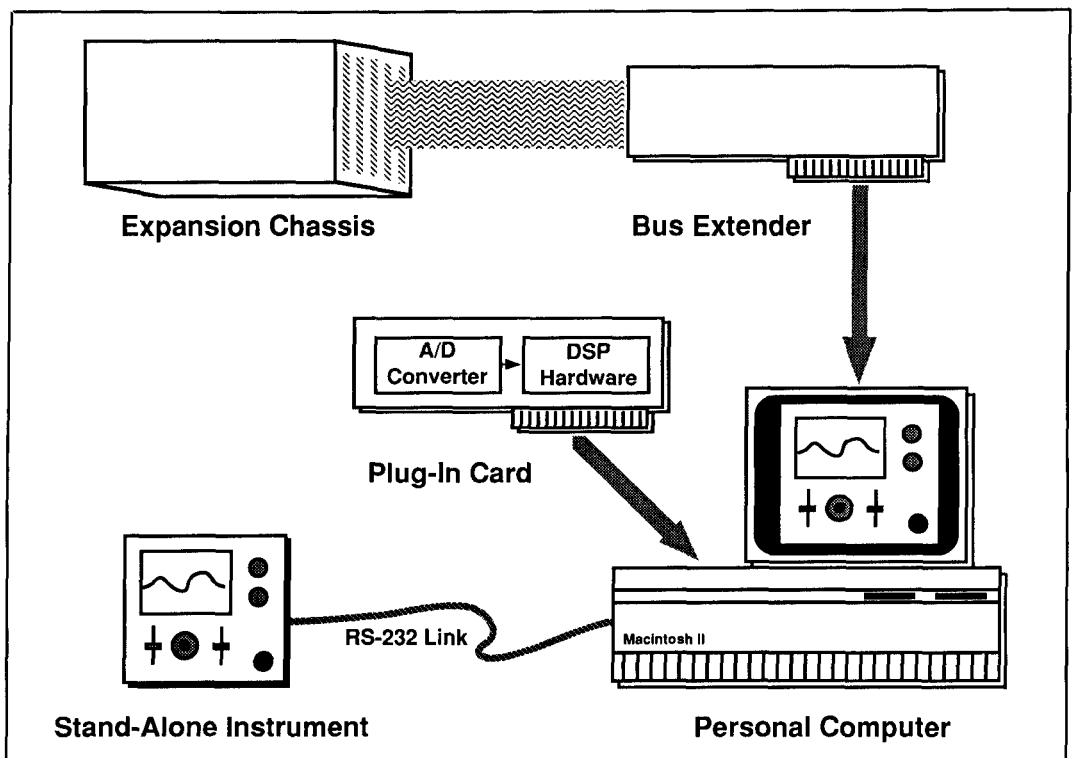

Figure 1. Microcomputer-based instrumentation generally follows one of three basic designs: bus-extension systems (top); plug-in instrumentation cards (center); and stand-alone isntruments connected to controlling PCs by RS-232 or IEEE-488 links.

NUIN SSB Signal Test

value = 2.489628

value = 2.4896289

Threshold

volts

```
Waveform :  0
Rate     :  10017.391304 samples/sec
Length   :  4264 samples
Min      :  2.271836 volts
Max      :  2.721250 volts
Mean     :  2.498126 volts
```

Close

2.59968

2.49136

2.38304

2.27472

0.000   0.106   0.212   0.319
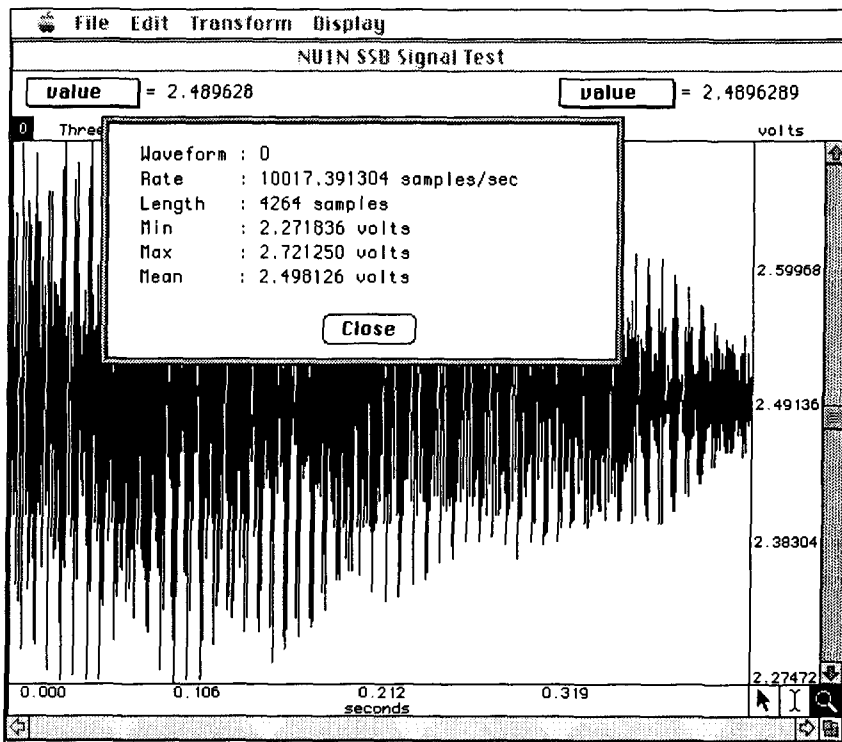   seconds

Figure 2. One of the advantages of virtual instrumentation is the facility for automated statistical analysis. In this example of virtual display, summary statistics have been automatically calculated and displayed for the SSB signal under test.

supplementary hardware on some PCs to make them IEEE-488 compliant.

Generally more expensive, stand-alone data-acquisition systems provide increased flexibility (in part because of their reliance on standard PC-instrumentation interface protocols) and potentially greater performance (for a price) than plug-in cards. For those who require the portability, flexibility, or performance of a stand-alone instrument, there are many units from which to choose. Keithley Instruments, Inc., produces stand-alone data acquisition systems, with software and IEEE-488 controller boards, for the Apple Macintosh,® IBM-PC,® and PS/2® machines. National Instruments offers a sophisticated array of IEEE-488 products for both the PC and the Macintosh.

Stand-alone instruments for Macintosh, PS/2 and IBM-PC compatibles

MetraByte Corporation, Data Translation, Inc., and Omega Engineering distribute a wide range of stand-alone instruments for the Macintosh, PS/2, and IBM-PC compatibles. Consider requesting catalogs

from these distributors. They not only provide information about systems from a variety of manufacturers, but include good technical introductions to data acquisition technology.

Fluke 45 DMM

The Fluke 45 DMM (John Fluke Manufacturing Company) is representative of contemporary, computer-controlled DMMs. Through the use of an RS-232 (or optional IEEE-488) interface, data acquired by this stand-alone system can be stored and manipulated by any computer with a serial RS-232 port. The high-end Hewlett-Packard 3567A stand-alone spectrum analyzer is sold as a module that is connected to the serial port of a 286/386-based IBM-PC. Rapid Systems, Inc., offers the R-370 spectrum analyzer and digital oscilloscope combination. Another inexpensive stand-alone system, Solus, from TMI, Inc., supports 36 channels. It comes with a graphics library of routines for any IBM clone with an RS-232 port.

Plug-in instrumentation cards

PC-based virtual instruments rely on a design that differs radically from that of

conventional instrumentation. Instead of using the power of an on-board microprocessor to enhance stand-alone instrument performance and flexibility, PC-based virtual instruments rely on plug-in data acquisition and processing cards that extend the function of PCs to include those features normally found in stand-alone instruments. A single plug-in card can be used to emulate a variety of instruments, from a storage oscilloscope to a DMM, by simply executing a different program on the PC.

Plug-in cards and the virtual instruments they support make it possible to acquire data, transform it to the appropriate format, and analyze the results in a Lotus 1-2-3® spreadsheet—all without the need to manipulate the raw data manually. In addition to ruling out the need for the error-prone transcription process, plug-in instrumentation cards offer a number of additional advantages: low cost and good availability, excellent performance, readily available software, familiarity, flexibility, and the ease with which data can be shared.

Example using a PC-based counter

To illustrate the potential benefit of PC-based virtual instrumentation, consider the following problem: The stability of a proto-type oscillator must be monitored accurately for an extended period of time. One solution would be to use a conventional frequency meter and log paper to gather the data. Once generated, the user would have to either enter the data from the log paper into a computer, or analyze the data with pen and paper. Alternatively, a PC-based counter could be used to automatically acquire and store frequency data, as well as compute summary statistics (**Figure 2**). For instance, the virtual counter might measure oscillator frequency every 100 ms to assess short-term stability and every 10 seconds to assess long-term stability, and store the data in ASCII files. Statistics compiled automatically might include maximum and minimum frequencies during the measurement period, mean frequency over the period, and total drift.

Advantages of virtual instruments

For most amateurs, a major factor in instrument selection has always been price. Fortunately, the cost of adding a plug-in virtual instrument to your PC system is typically one-third to one-half the cost of acquiring a stand-alone unit with comparable capabilities.[6] For example, spectrum analyzers on a plug-in card for a PC are available
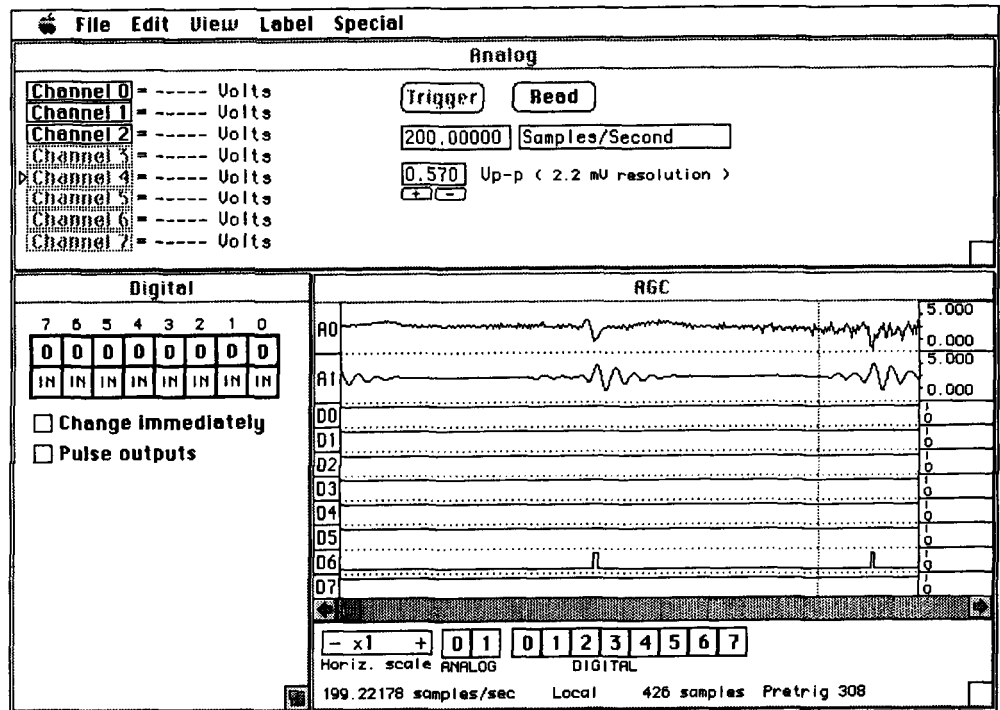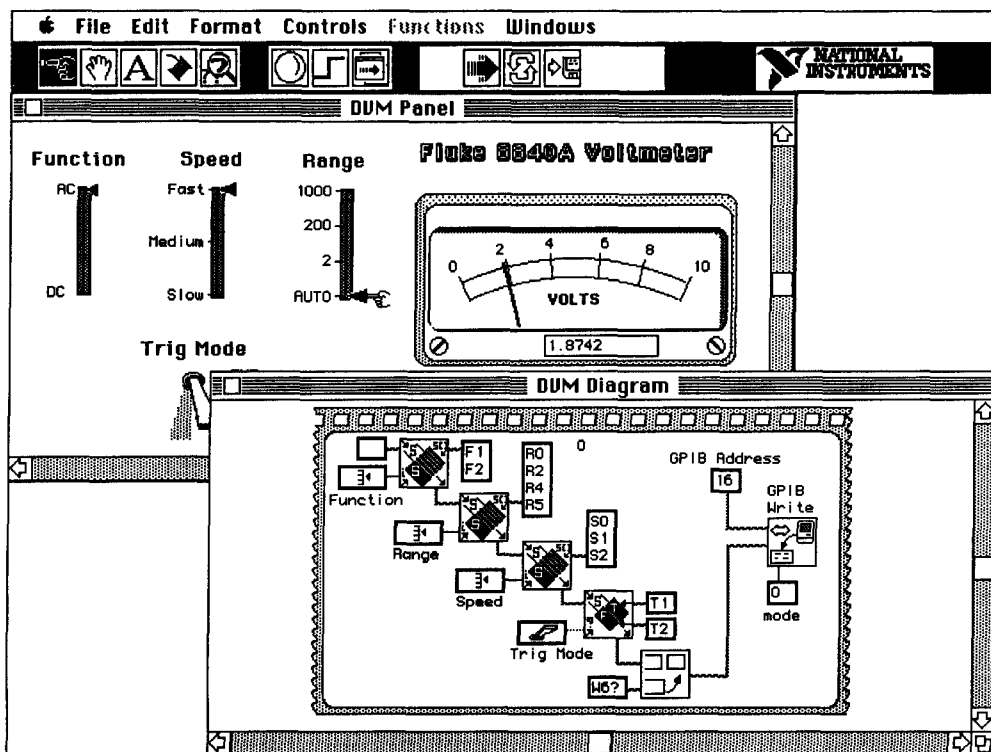

Figure 3. A virtual instrument interface controlling a plug-in card with multiple-analog and digital-input channels. Simple and intuitive controls allow the user to specify active input channels, resolution, and amplitude range.

Figure 4. National Instruments LabView virtual instrument environment configured to emulate a Fluke digital voltmeter (DVM). Note the dual digital analog display provided by the virtual front panel. Also note the iconic programming language used to define the Fluke's operational parameters (foreground). National Instrument's LabView for the Macintosh is a multi-layered, graphic-programming and display environment.

for less than $1000; stand-alone instruments cost many times that.

When you purchase a dedicated stand-alone instrument, you pay for a case, power supply, front panel with controls and displays, printed-circuit boards, and the labor to assemble it. With a virtual instrument, the first three items are included in the price of your PC. (This is especially significant, given that the most costly single item in a conventional spectrum analyzer or oscilloscope is the analog cathode-ray tube, or CRT. Virtual spectrum analyzers and oscilloscopes rely on the PC monitor for data display.) The attraction of cost-savings associated with virtual instrumentation is reflected in the electronics instrumentation industry as a whole. Plug-in boards now account for 60 percent of the world-wide data acquisition market.[7]

User interface

Affordable software for data analysis and signal acquisition/generation is not only easy to obtain, but user interfaces have improved, providing greater ease of use. From the user's perspective, a virtual instrument is often easier to master than a comparable

stand-alone system. In many cases this is due in part to the user's familiarity with using a PC in other work. Another factor is the ability to modify the user interface so that it matches both the task at hand and the user's experience. For example, instead of forcing a user to learn a different, complex interface for each instrument, software-defined instrumentation interfaces can be designed to provide only those functions actually required; distracting controls that aren't needed can be temporarily hidden from view (Figure 3). Virtual instruments can also provide digital and analog displays of values simultaneously—without changes in hardware. Some people assimilate analog readouts faster than their digital equivalents, and virtual instruments can provide both (Figure 4).

Enhancing productivity

In addition to providing a consistent, easy-to-learn user interface, virtual instruments can make use of the underlying PC platform to enhance productivity. By providing downloading routines that perform complex measuring sequences without user intervention, a PC can enhance the diagnostic process. Similarly, the operation of a vir-

tual instrument can be simplified by group-
ing several machine functions into a single,
software-definable function. Perhaps the
greatest boon to productivity is the pro-
vision for efficient data storage and retriev-
al. For instance, the library of normal
waveforms for a particular piece of elec-
tronic equipment can be stored in a PC and
recalled as reference waveforms to be used
when measurements are being made on a
defective piece of equipment. And finally,
since PCs are easily networked, the data
gathered by a virtual instrument at one lo-
cation can be readily shared with PC users
on the network.

Availability of plug-in cards

Plug-in instrumentation cards are avail-
able for a variety of applications and price
ranges for both the PC and the Macintosh.
There are easily over a hundred plug-in data
acquisition cards available for the IBM-PC
and clones, and about half that number for
the Macintosh II® . (See a recent issue of
Byte magazine, or a catalog from a distribu-
tor of data-acquisition boards, like Omega
Engineering, Inc., for descriptions of the
plug-in boards on the market.) ADAC Cor-
poration offers a variety of low-cost

multifunction boards for the PC and PS/2,
as well as external signal conditioning and
expansion modules. Strawberry Computer
Inc. markets a large assortment of plug-in
cards and stand-alone data acquisition sys-
tems for both Macintosh and PS/2 ma-
chines. Rapid Systems, Inc. offers plug-in
cards and software that can transform an
IBM-PC or clone into a frequency counter,
oscilloscope, or spectrum analyzer. GW In-
struments, Inc., which specializes in Macin-
tosh-based data acquisition hardware, of-
fers both plug-in cards and peripherals.
Their MacADIOS series includes a compre-
hensive line of data acquisition and control
hardware for the Macintosh.

Disadvantages of plug-in cards

Paradoxically, the feature that provides
plug-in instrumentation with its greatest as-
set—an intimate link with a PC—is also its
greatest liability. Stand-alone instruments,
when equipped with a standard communica-
tions interface, can be used with practically
any PC. Plug-in cards will function proper-
ly with only one computer bus architecture;
for example, MicroChannel, NuBus, or AT
bus systems. Upgrading your hardware plat-
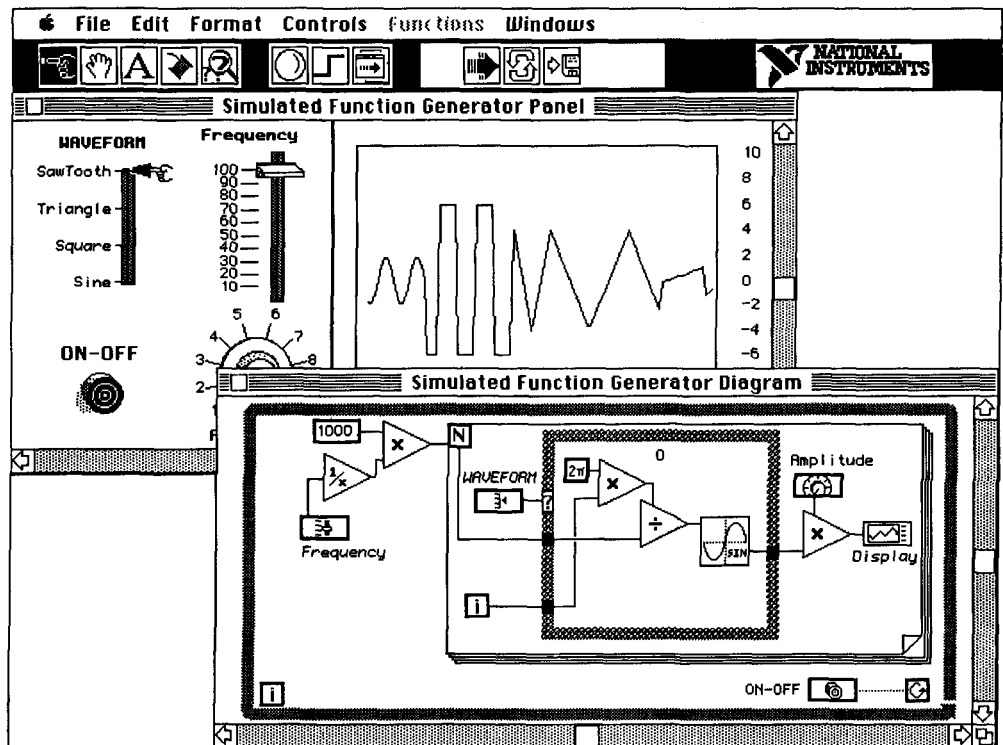form, say, from an AT to a PS/2 machine,

**Figure 5. A virtual function generator defined in LabView. Note the real-time strip chart and the familiar slider
and rotary controls on the virtual front panel, which free the user from interacting with the instrument through a
command line or menu-driven interface. The graphic programming environment that defines the function of the
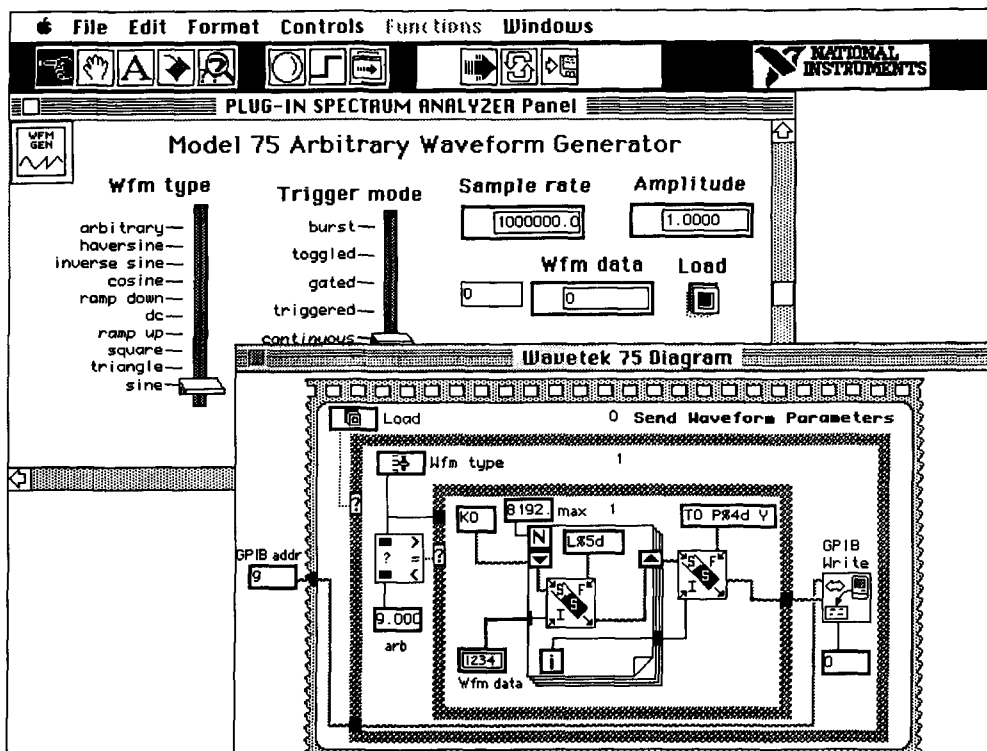function generator appears in the front, lower right panel.**

Figure 6. A virtual WaveTek 75 function generator, defined in LabView. The software definition of the instrument appears in the foreground.

will require that you also purchase a new plug-in card. You also assume responsiblity for system integration if you buy hardware from one vendor and software from another. The use of hardware from multiple vendors may result in conflicts with other boards or peripherals. Luckily, most of these problems can be alleviated if you buy from a single vendor.

Instrumentation software

Regardless of whether a microcomputer-aware instrument is a stand-alone or based on a plug-in card, the controlling PC's software is as critical to operation as the hardware. For example, the speed of data acquisition is more often limited by software than by hardware.[5] Not only does the software support data acquisition and provide data reduction through data transformation and manipulation, but it also defines the user interface. The user interface, provided by software running on a PC, shields the user from the physical device. The virtual instrument defined in software may ultimately control stand-alone instruments, or enable a plug-in card to simulate a physical device.

Instrumentation hardware can generally be controlled by any traditional programming language from ASSEMBLER to BA-SIC. (For a sample BASIC data acquisition routine for controlling an IEEE-488 link, see the article by Clune.[5]) However, the larger manufacturers have developed easier-to-use languages in the interest of attracting nonprogrammers to their products. One company, Analog Connection Workbench (Strawberry-Tree Computers), allows programming using icons on the Macintosh.

LabView instrument programming

Perhaps the most impressive virtual instrument programming environment available is National Instrument's LabView (Laboratory Virtual Instruments Engineering Worksheet) for the Macintosh (see Figures 5 through 7). Programming in LabView is accomplished by drawing links between icons and linking the resulting structure to a virtual front panel, which may be composed of switches, meters, and sliders, depending on the application. National Instruments also offers a similar package for the IBM-PC and PS/2 machines—the LabWindow. Although the programming environment isn't as graphically elaborate as that provided by LabView, LabWindow can be used to implement impressive virtual instrument interfaces.
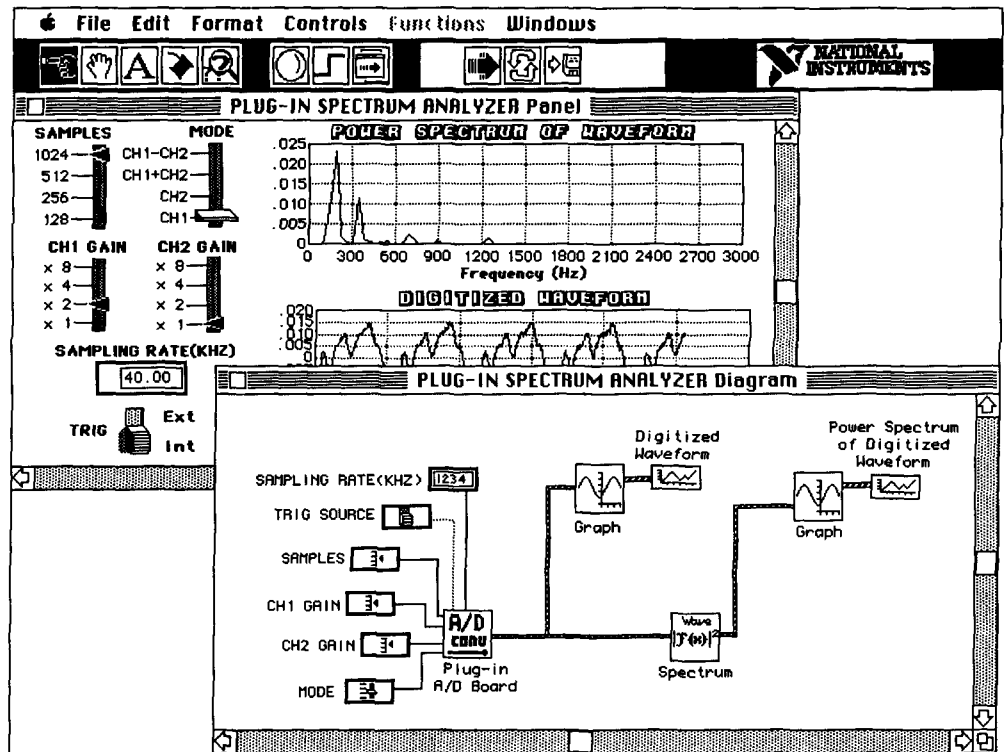
Figure 7. A virtual spectrum analyzer defined in LabView. Note intuitive controls on the virtual front panel and the programming environment in the foreground.

GW Instruments

GW Instruments, as well as National Instruments and Strawberry-Tree Computers, offers a graphically oriented, easily customized data-acquisition front end. GW's Mac-ADIOS Manager and Labtech Notebook, while not as "iconized" as the products from the latter two companies, provide a programming environment that's sufficiently "Mac-like" to allow any nonprogrammer to quickly and easily design a custom front-end for a data-acquisition system. The Mac-Instruments software can also be used to transform a Macintosh into a dual-channel oscilloscope, single-channel chart recorder, or scrolling-strip chart recorder. GW Instruments also offers a hardware/software package dedicated to speech analysis.

MacSpeech Lab and AcqKnowledge systems

MacSpeech Lab, for the Macintosh, provides spectral analysis and can display sound-amplitude envelopes. In effect a high-end version of Farralon Computing's Mac Recorder, MacSpeech Lab can capture and play sounds with a sample rate of up to 80 kHz. AcqKnowledge (Biopac Systems) digital signal processing software for the Macin-

tosh provides a variety of functions, including spectral analysis, filtering, and transformation. It works with the company's dedicated hardware in real time, as well as on previously recorded data in ASCII format.

The future

The future of plug-in instrumentation cards seems assured. Plug-in card manufacturers continue to offer, with each generation, cards with decreasing price to performance ratios, increased bandwidths, higher resolution, and increased dynamic range. These cards are readily available at low cost partly because of the competition within the industry; competition is fierce, and companies continue to offer plug-in cards that offer greater speed and accuracy at reasonable prices. The steadily decreasing cost of memory devices will likely continue to foster the development of inexpensive plug-in instrumentation.

Plug-in instrumentation will probably continue to benefit from the increased ease of use and flexibility provided by the alternative use interfaces made possible by the PC. Although graphic displays, controlled by a mouse or keyboard, have become standard in virtual-instrument interface designs, user interfaces that rely on speech recogni-

tion (SR) have the greatest potential to change the way we work. Speech is the standard medium for interpersonal communications. SR interfaces offer several potential benefits when used instead of, or as a supplement to, graphical user interfaces. Perhaps the most obvious advantage of an SR interface would be to provide instrumentation control when the user's hands are unavailable; for example, when probing inside a modern transceiver.

While instruments designed around plug-in cards seem to be gaining popularity, there are also promising developments in the stand-alone instrumentation field. For example, practically all stand-alone instrument manufacturers are using on-board microprocessors to maximize the function of their dedicated instruments.

Given these parallel developments in the instrumentation field, it seems likely that we will eventually see some type of hybrid PC/instrumentation system that melds the two technologies into one. The HP 8920A communications test set may provide a glimpse of what we can expect in future desktop test equipment. The 8920A relies on an embedded PC to provide the function of over 20 conventional instruments in a single PC-sized chassis. What's even more amazing is that all of the instrumentation features are controllable by user-defined BASIC programs. As a representative of the future in stand-alone instrumentation, consider the hand-held Fluke 97 ScopeMeter® . At only four pounds, the computerized instrument combines a 50-MHz dual-channel digital-storage oscilloscope, a full-function DMM, a signal generator, and component tester, together with ports for an RS-232 link and a printer.* Whatever the shape of future instrumentation, one thing is for certain—the VOM and other noncomputer-aware instruments are destined to suffer the fate of the once-popular slide rule. ■

*For more information, see Product Information on page 32 of the Fall 1991 issue of Communications Quarterly.

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Vendors

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Marlboro, Massachusetts 01752
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Hewlett-Packard Company
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Cupertino, California 95014
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SPECIAL COLUMN
THE SOLAR SPECTRUM

Solar activity was moderate during the latter part of the fall. One particularly long-lived sunspot group in the Sun's southern hemisphere returned to the visible disk for third and fourth appearances. The third transit of this group was especially active in the production of high-energy solar flares.

One of these, a X6.1/3B* flare, was the most energetic event to erupt since the series of extraordinary powerful flares (X12+) that occurred in June. This spot group

*A description of the X-ray flare classification is given in "Outlining June's Strong Solar Flare Activity," in the Summer 1991 issue of *Communications Quarterly*.

(Photo A) grew to encompass an area of about 7800 million square kilometers in October, making it one of the largest clusters to appear during this solar cycle. (For comparison, the surface area of the Earth is a little over 510 million square kilometers.)

For many of us, however, the highlight of the period was the spectacular aurora which occurred on the 8th and 9th of November. This brilliant light show was seen in many parts of the world, from geographical sites as low as latitude 30 degrees. Virtually all of the ensuing reports describe the aurora as an extremely active display of rapidly changing shapes and colors which covered all, or a good part of the night sky. A
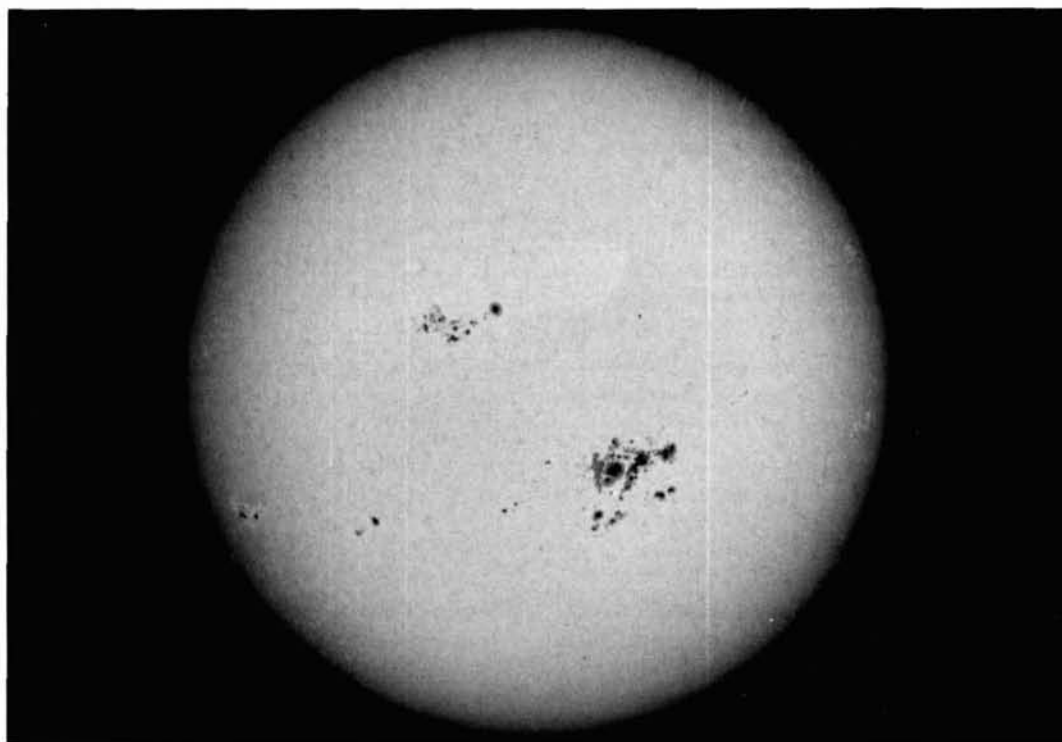
Photo A. The very large and long-lived sunspot group which dominated the solar disk in late October is shown in this fine photograph by Dr. Jean Dragesco, la Riviere, France.
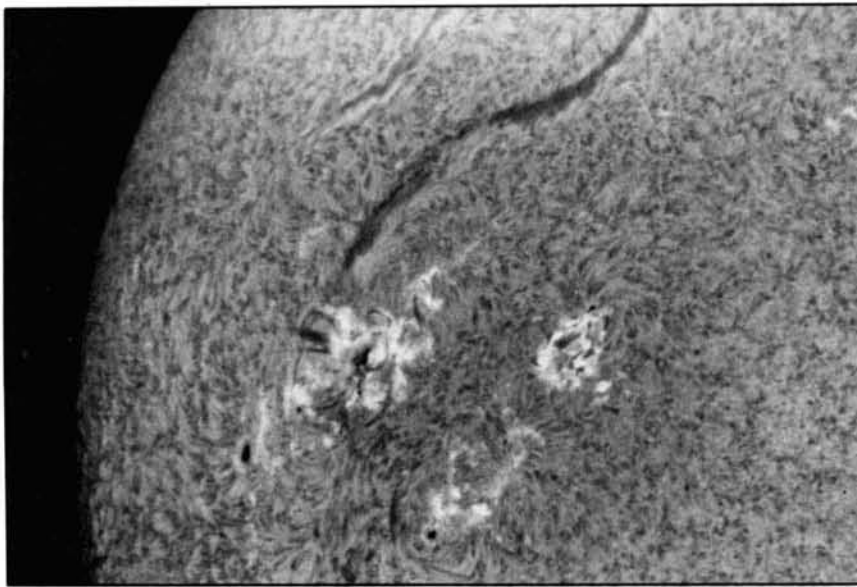
Photo B. A long filament on the disk, photographed by Dr. Dragesco at a wavelength of 6563 angstroms (Hα).

number of experienced observers characterized the phenomenon as the most vivid in recent years.

As might be expected, not all of this activity was visual. Several United States locations reported a sudden signal degradation for some commercial short-wave stations in Europe, which may be linked to poor signal propagation over the Pole. By 22:30 UT on the 8th, a severe geomagnetic storm had rendered propagation paths at mid-latitudes and above to very poor or useless, while low latitudes experienced poor conditions. Circumstances began to improve by early the following day and returned to normal on the 10th.

On the other hand, radio amateurs and others at mid-latitude locations provided graphic accounts of a considerable enhancement of VHF and 2-meter DX which may well have resulted from auroral propagation. Several cited radio effects as their first indication that a strong aurora was in progress.

Disappearing solar filaments

The source of this event and the intense magnetic storm which caused it to occur is intriguing. Most of us first think of powerful flares on the Sun when these conditions are encountered, or perhaps point to high-speed streams of solar wind plasma spawned by well-positioned coronal holes as a culprit.

However, for more than a decade we have known about a third important origin of geomagnetic storms—the disappearing

solar filament, or *disparition brusque*. Indeed, just such a source is a strong suspect in this instance. Solar filaments (**Photo B**) are really prominences (**Photo C**) which are seen in the red light of atomic hydrogen (Hα) projected against the solar disk, rather than at the limb. They are among the more stable of all solar phenomena, often lasting for several months before seeming to vanish overnight.

These dark, relatively dense features generally disappear by eruption, when the mass of gas is almost invariably expelled upward. Strangely enough, this upward motion is no problem for the Sun which undoubtedly has difficulty holding the gases down! Such events are thought to be responsible for a large fraction of coronal mass ejections, which spew forth an estimated 10 billion metric tons of solar material each day.

Although the frequency of disappearing filaments typically rises and falls with the solar cycle (**Figure 1**[1]), they also occur when the Sun is relatively quiet. One of the more infamous of these events—one which had considerable terrestrial implications—occurred during August 1978. Since the solar cycle maximum was well over a year in the future and the Sun was fairly quiet, the magnetic disturbance which resulted came as somewhat of a surprise, and inflicted widespread damage to communication, electric power, and telephone systems.

Filaments vary considerably in size, ranging from features within sunspot groups only a few heliographic degrees in length (which don't concern us here), to large structures over 60 degrees long that general-
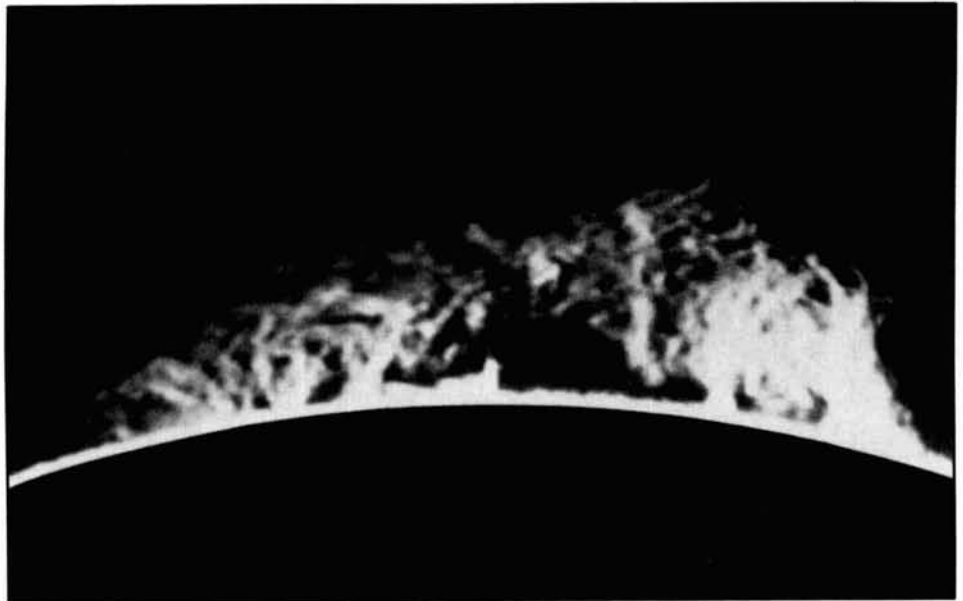
Photo C. A large prominence on the Sun's limb.

ly extend over quiet or previously spotted areas. Filaments act as a marker of sorts, in that they always lie above the lines of magnetic inversion which separate regions of opposite magnetic polarity in the photosphere.

While disappearing filaments have long been suspected as a source of geomagnetic disturbances, early attempts to correlate these data were inconclusive. Shortly after the end of World War II, when the nature of recurrent magnetic storms began to be unraveled, attempts were made to tie the mere presence of filaments on the disk to these storms, but they were not successful.

Research during the years that followed resulted in generally conflicting or unreliable conclusions. Eventually, in the early part of the last decade, studies by Joselyn and McIntosh[2] and by McNamara and Wright,[3] along with observations of coronal transients by instruments aboard the Sky-Lab spacecraft, finally provided evidence of a verifiable connection between these occurrences.

In their research, Joselyn and McIntosh estimated a travel time from disappearance to geomagnetic storm (defined by a daily planetary geomagnetic Ap index $\geq$ 30 of from 3 to 6 days, with a majority clustered
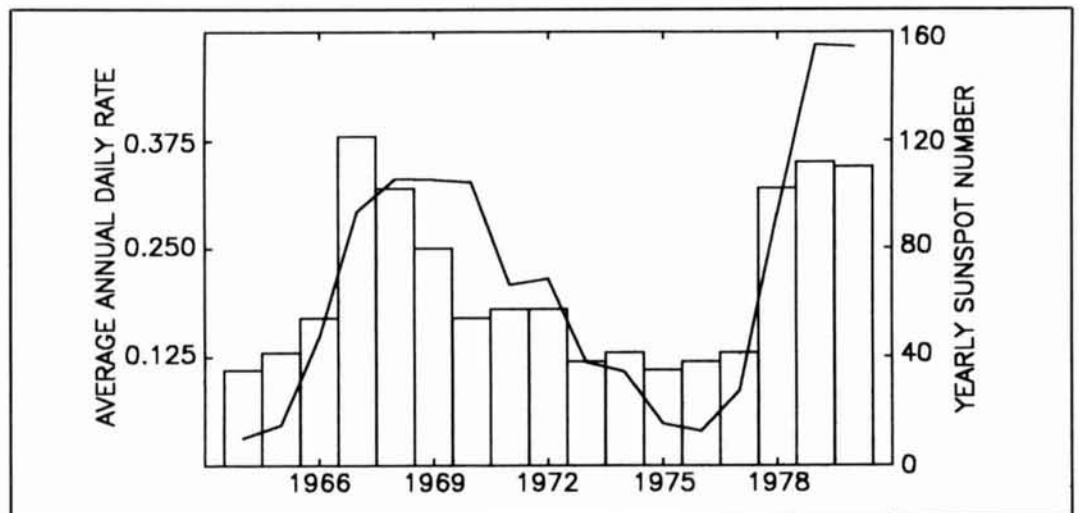
Figure 1. The estimated daily rate of disappearance for filaments on the Sun's visible hemisphere (bars)[1] and the yearly mean sunspot number. Only filaments with areas greater than 10 degrees which originated between heliographic longitudes E45 and W45 are included in the figure.

between 4 and 5 days. In addition, they found that the filament should have previously been "quiescent," or not associated with an active sunspot cluster. Their study of the sources of magnetic storms between 1976 and 1979 disclosed the relationship shown in **Figure 2**.

Because filaments have great vertical as well as horizontal size, McNamara and Wright determined a distinct relationship between the rate of geomagnetic storm production and a disappearing filament's area. Moreover, their research could discern no correlation between small active-area filaments and such storms. Therefore, it appears that only the disappearance of larger filaments is likely to be followed by significant effects on Ap.

These parameters are reasonably well fulfilled for the severe magnetic storm which took place early in November. Its source is assumed to be a 24 degree long, dark filament that disappeared from the Sun's southeast quadrant early on the 6th.[4] Furthermore, flare and/or coronal hole activity prior to this time doesn't appear to justify a response of the type or intensity that occurred.

Outlook

December 1991 marked the 64th month of solar cycle 22. Activity generally declined during the preceding few months, although most indices were up during October due
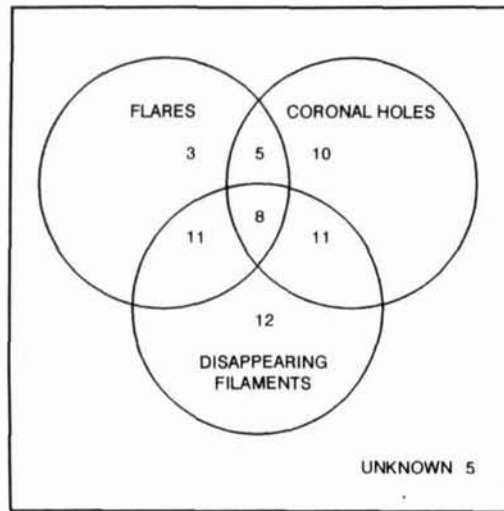
Figure 2. Solar origin of 65 geomagnetic storms (Ap $\geq$ 30 which occurred between 1976.5 and 1979.5.[2]

primarily to activity in the large spot-group shown in **Photo A**. As of December, several prediction models called for another period of enhanced activity to occur in late January or February, but these conditions were not expected to affect the July 1989 presumed maximum of this cycle. ■

REFERENCES
1. Report UAG-199, National Geophysical Data Center, Boulder, Colorado.
2. *Journal of Geophysical Research*, 88, Number A6, 1981.
3. *Nature*, 299, 7 October 1982.
4. *Preliminary Report and Forecast of Solar Geophysical Data*, Number 846, 1991.

PRODUCT INFORMATION

DC Power Contactor

Kilovac's new power contactor will switch a 150-amp, 300-volt DC load and a 600-amp overload condition, yet weighs less than 25 ounces. The model PD150 contactor is designed to meet the needs of distributed power applications, electric vehicle battery disconnect needs, and rapid transit power distribution applications. Because of the vacuum environment inside the contactor, the PD150 is able to carry current with low contact resistance and switch loads at up to 1,000 volts DC.

The contactor is approximately 2 inches in diameter and 4.6 inches tall and requires no heat sink in normal operation. The PD150 dissipates only 35 watts at full load in the "on" state. The contact configuration is SPST-NO; the contactor is bidirectional, so it will function with current flowing in either direction.

For more information contact Kilovac Corporation, P.O. Box 4422, Santa Barbara, California 93140.

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UHF VOLTAGE VARIABLE PHASE SHIFTER

Cascading multiple sections of loaded-line phase shifters

V oltage variable phase shifters are commonly used in communications systems to alter the phase of a transmitted signal as a function of an input voltage. Typical applications include antenna steering networks, phase modulators, and feedforward networks where a phase shift of over 120 degrees is required.

Continuously variable phase shifters realized using varactor-loaded quadrature hybrids are well understood.[1] However, it can be difficult to realize a hybrid above 500 MHz using a ferrite twisted pair approach, and Lange couplers become large at frequencies below 2 GHz. An interesting approach has been developed using the loaded-line phase shifter concept.[2] This article describes a technique for cascading multiple sections of loaded-line phase shifters to achieve phase shifts in excess of 120 degrees.

Single-section, loaded-line phase shifter

The single-section, loaded-line phase shifter is shown in **Figure 1**. It consists of a transmission line with characteristic impedance Zc and electrical length l, loaded on each end by identical varactor diodes. As the bias voltage applied to the diodes is changed, the capacitance change of the diodes shifts the transmission phase of a signal traveling on the line.

For small values of capacitive susceptance B with arbitrary l, the phase shift will be small and the input match to the line not significantly degraded. As B is increased, the input reflection coefficient increases and the line becomes mismatched. However, combinations of B and l exist that allow for greater phase shift range while maintaining
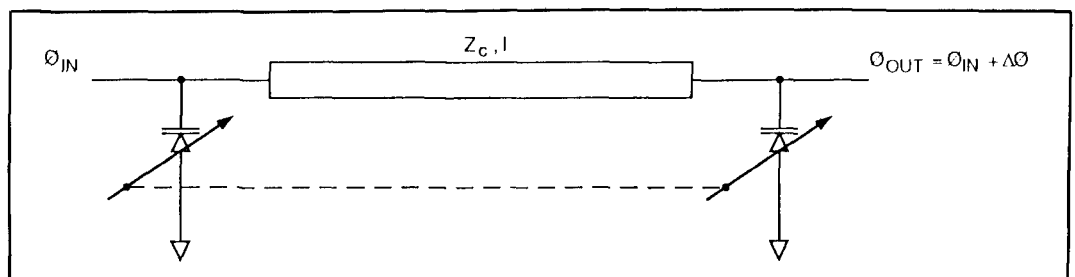
Figure 1. Single-section loaded-line phase shifter.
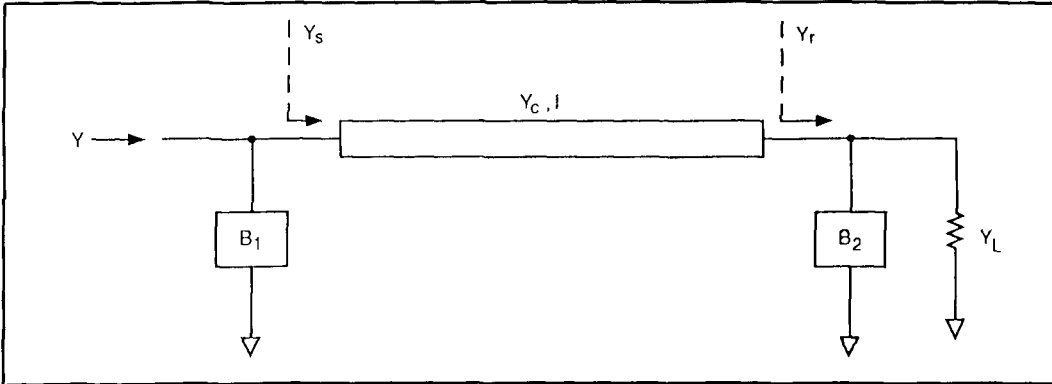
Figure 2. Model for impedance calculations.

a good match. Several matched loaded-line phase shifters may be cascaded to increase phase shift range.

Theoretical analysis

Referring to **Figure 2**, the input admittance Y to a single-section loaded-line phase shifter may be written as:

$$Y = B_1 + Y_s \qquad (1)$$

where Ys represents the driving point admittance of the transmission line. Using the general equation for the lossless uniform transmission line, **Equation 1** can be rewritten as:

$$Y = B_1 + Y_c \ \frac{Y_r \cos \beta l + j \ Y_c \sin \beta l}{Y_c \cos \beta l + j \ Y_r \sin \beta l} \qquad (2)$$

where Yr represents the admittance combination of B2 and the next stage.

Solving **Equation 2** and plotting return loss versus line length l for various B results in the family of curves in **Figure 3**. Note that near l = 65 degrees a wider range of susceptance values maintain a good match at the line input, and therefore a wider phase shift may be obtained.

I created a Touchstone file to analyze four cascaded sections at 875 MHz. My analysis shows that phase shifts of 40 degrees per section can be expected while maintaining an overall VSWR of less than 1.3:1. Insertion loss is primarily a function of transmission line loss and varactor Q.

Practical considerations

Maximizing phase shift range

Unfortunately, varactor diodes have some minimum capacitance (at maximum bias voltage), typically in the 2-pF range. This will limit the matched phase shift range. However, placing an inductor in
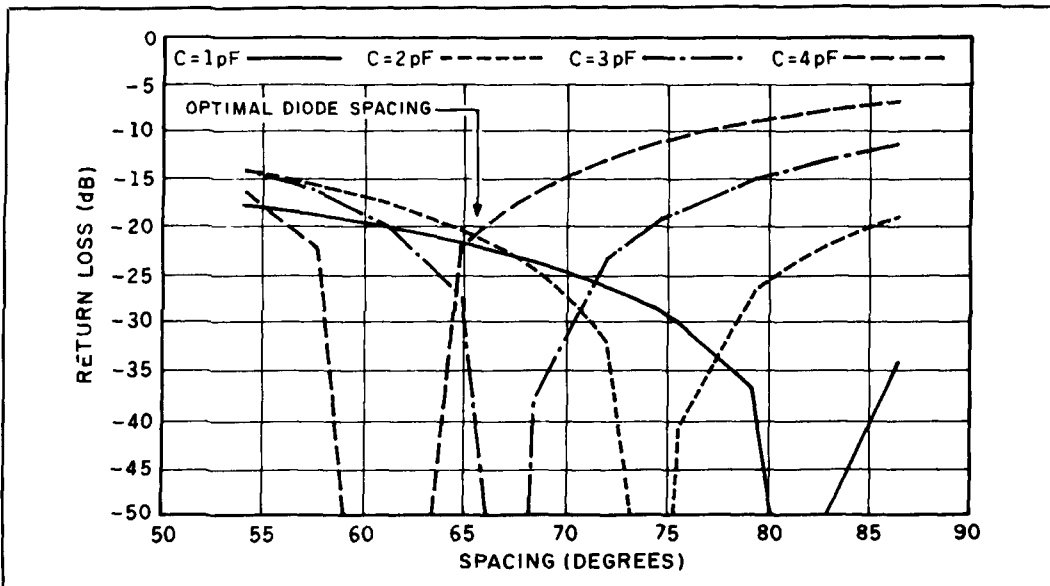
Figure 3. Return loss versus diode spacing.
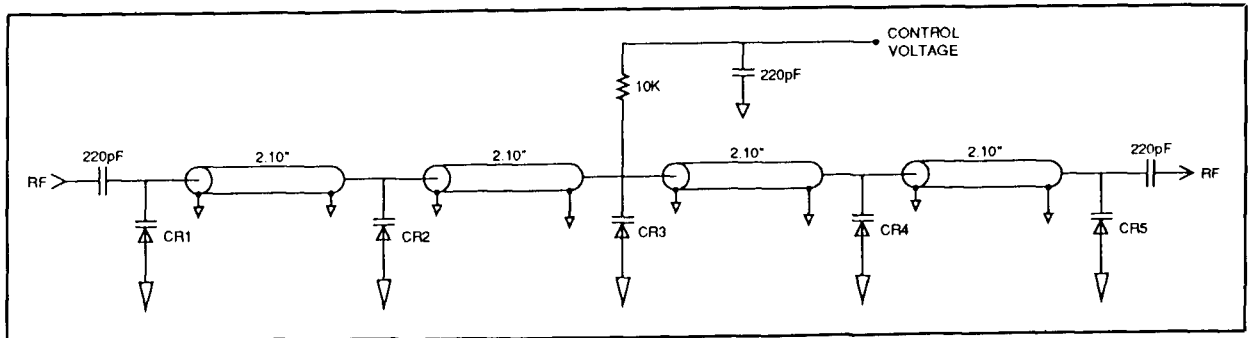
Figure 4. Electrical schematic diagram of load-line phase-shifter prototype.
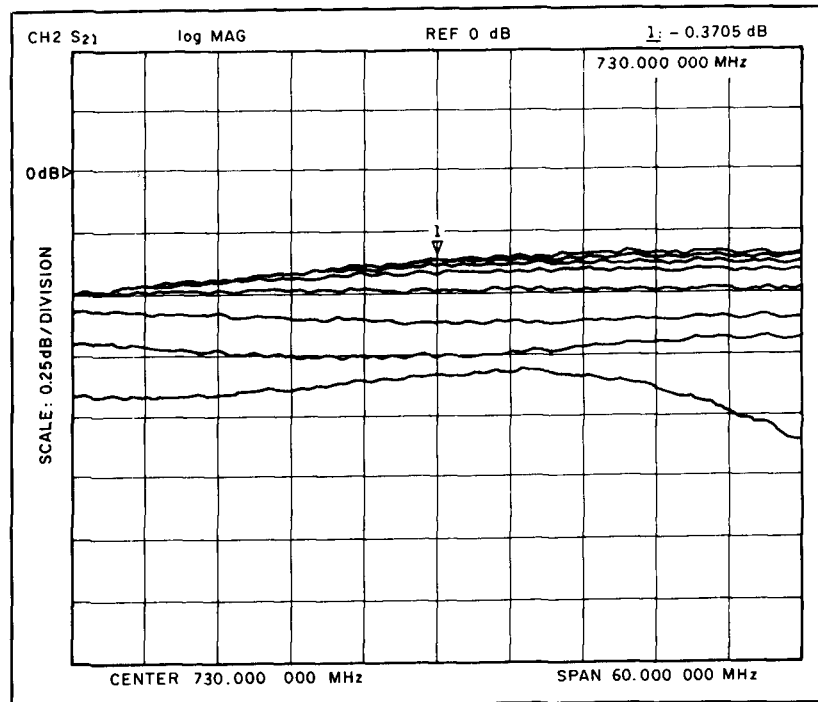
Figure 5. Insertion loss versus phase shift.

parallel with the varactor whose inductive susceptance is equal to that of the capacitive susceptance of the diode at minimum capacity "buys back" that tuning range.

Minimizing insertion loss

Loss, due to the finite Q of the inductor, will increase slightly. The finite Q of the varactor (about 150) will add loss too. Great care should be taken in selecting a low-loss transmission media.

Power handling capability

As the power level on the line increases, the varactor diode can be modulated by the RF signal, producing incidental phase modulation and intermod distortion. This undesirable characteristic is the limiting factor when considering power handling capability. To decrease a varactor's sensitivity to

RF voltages, a −8 volt back bias can be applied at the anode. Control voltages from 0 to +22 volts applied to the cathode will then swing the varactor capacity across the required 2 to 6 pF range.

Phase-control bandwidth

Varactor diodes are inherently speedy devices due to the fact that they operate under reverse bias. No charge must be "swept out" of the intrinsic region of the junction when control bias is modulated. Therefore, the phase-control bandwidth is usually limited by the time constant of the bias resistor and capacitor.

Experimental results

To verify the concept, I constructed a 730-MHz prototype using 2.10-inch sections
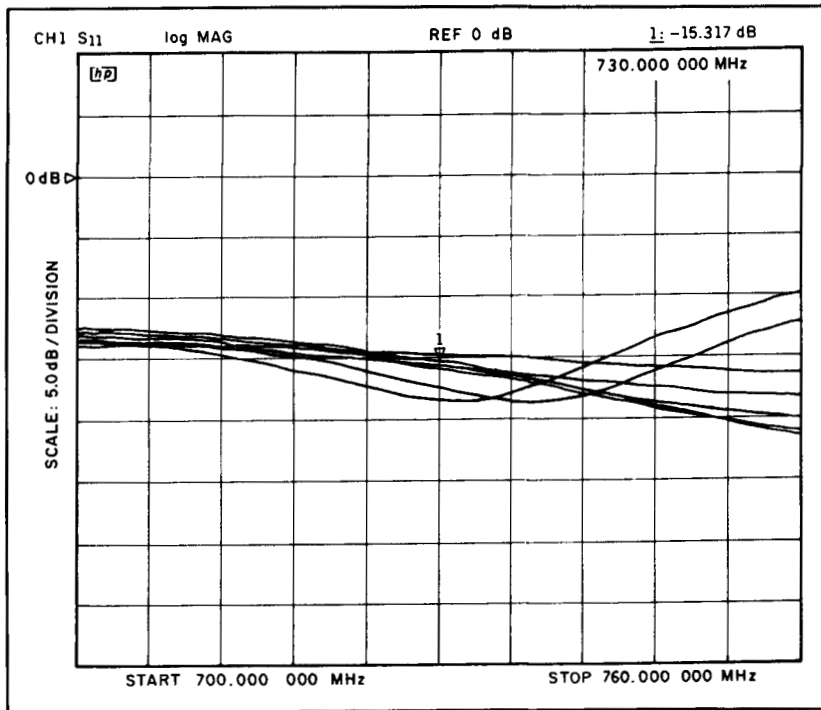
Figure 6. Return loss versus phase shift.

of 0.47-inch diameter semi-rigid 50-ohm coax for the transmission media. At each line junction, I soldered an MMBV105G varactor in an SOT-23 plastic package from the center conductor to ground. I applied bias through a 10-k chip resistor to the center conductor, and used 220-pF coupling capacitors to block the bias at the RF inputs. I soldered the coax to a rigid pc board and attached SMA connectors to the edges (see **Figure 4**). The insertion loss, return loss, and phase-shift frequency responses versus control voltage are shown in **Figures 5, 6,** and **7**. The insertion loss was held to
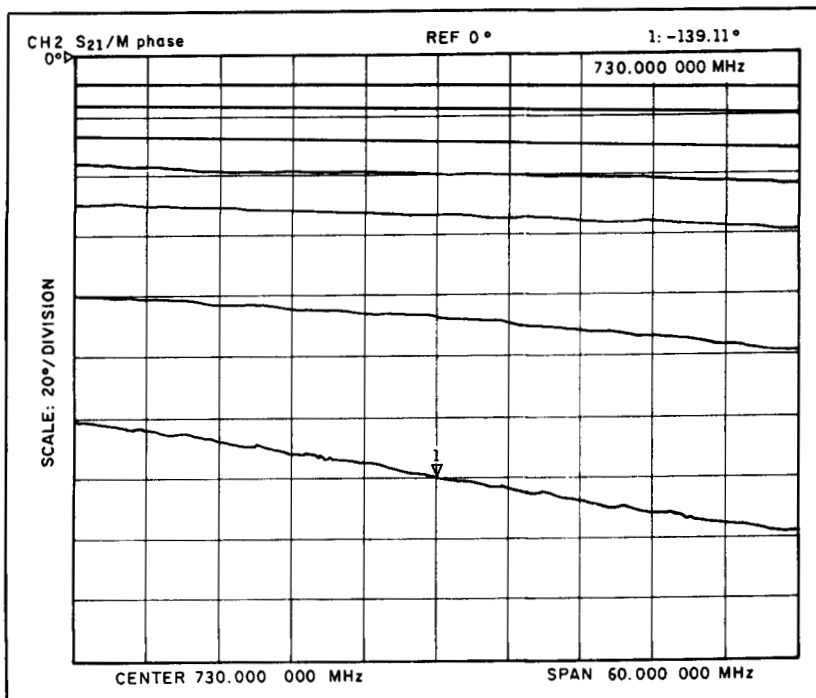
Figure 7. Phase shift versus control voltage.
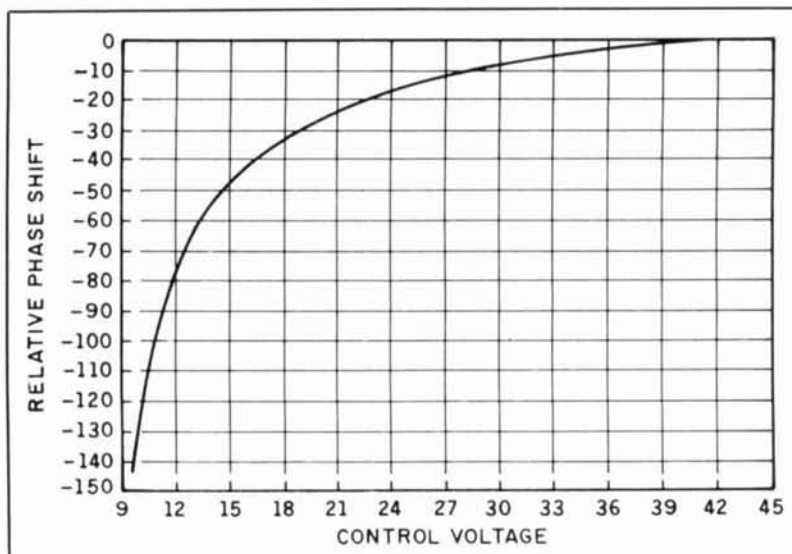
Figure 8. Phase shift versus control voltage for the loaded-line phase shifter.

less than 1.5 dB over a 60-MHz bandwidth centered at 730 MHz (8 percent BW). The VSWR was typically better than 1.5:1, with a worst case of 2.0:1. Minimum total phase shift over the bandwidth was greater than 120 degrees. A plot of phase shift versus control voltage is shown in **Figure 8**. As for applications, there are many. Uses include phase modulators and 0 to 180 signal cancellation networks. This circuit was originally built for a cellular telephone customer

who was using it in an "amplifier feed-forward network." The circuit topology lends itself well to microstrip construction techniques, and I expect that electrical performance would improve considerably using this approach. ∎

REFERENCES
1. G.A. Latheef, "A Variable Phase Shifter and BPSK Modulator," *RF Design*, May 1987.
2. H.A. Atwater, "Circuit Design of the Loaded-Line Phase Shifter," *IEEE Transcripts on Microwave Theory Technology*, Volume MTT-33, No. 7, July 1985.

PRODUCT INFORMATION

New HP Digital Multimeter

Hewlett-Packard Company has introduced the HP 34401A digital multimeter (DMM). Designed for system and bench applications, the new DMM can be used as a stand-alone or computer-controlled instrument. It includes features of a bench digital multimeter like dB, null, diode test, and continuity.

The HP 34401A provides up to 1,000 volts with 6-1/2 digits of resolution and basic 24-hour DC accuracy of 0.0015 percent. It also has limit test, minimum/maximum testing results, 512 reading storage, automatic reading hold, autoranging, and display updating. A menu feature allows the user to modify advanced functions.

The new DMM can be incorporated into existing systems without software changes. It accepts standard commands for programmable instruments (SPCI), HP 3478A, and Fluke 8840A commands. HP-IB and RS-232C interfaces are standard.

An optional pouch, designed to strap onto the instrument, can accommodate the power cord, interface cables, test probes, user and service manuals, and other accessories.

For details contact: Hewlett-Packard Company Inquiries, 19310 Pruneridge Avenue, Cupertino, California 95014.

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BUILD THIS HIGH POWER S-BAND STEP-RECOVERY DIODE TRIPLER

Rediscover the useful SRD

I n the Spring 1991 issue of *Communications Quarterly*, I described a high-level SSB up-converter for use at 3456 MHz.[1] This converter has the principal advantage of being able to develop relatively high levels of sideband power directly at 3456 MHz. Low level mixers and cascades of S-band amplifiers aren't needed. Because of its simplicity, many hard-to-find and often expensive facilitating microwave devices, like coaxial relays, circulators, and S-band filters, aren't needed.

The need for a relatively high-level oscillator signal for the up-converter might be considered one disadvantage of this approach. I needed about 1/2 watt of 3312 MHz CW power. Linearity wasn't a consideration. Thus, this requirement seemed to be reasonable.

Because of the proliferation of various microwave power transistor devices in recent years, you may have forgotten how useful step-recovery diodes (SRDs) can be for high-level and high-efficiency multiplication. I built a varactor tripler patterned after those used a few decades ago to multiply 144 MHz to 432 MHz,[2] or 432 to 1296 MHz. The balance of this article focuses on construction and test of a 3312-MHz SRD tripler.

General description

The tripler uses a 1N5152 power varactor diode,[3] although either a 1N5151 or a 1N5153 diode can be substituted. I've included an idler circuit tuned to the second harmonic of 1104 MHz. The idler recirculates 2208 MHz, achieving approximately 50 percent efficiency. A half watt of output power is obtained with about a watt of 1104 MHz power, which satisfies the LO requirements for my 3456-MHz SSB up-converter. Note, however, that the 1N5152 is capable of handling over 10 watts of drive, so for other applications it has the potential for much higher power output.

While fabrication of the tripler, using only hand tools, may at first seem formidable, the most critical requirement is that the sheet-metal enclosure be rugged. I used 0.025-inch thick brass sheet stock purchased at a hobby shop and took special pains to simplify the cutting and bending procedure for the sheet-metal parts to make it easy to duplicate.

Making the enclosure

Figure 1 shows the frame—a U-shaped piece of brass. The frame forms the top, front, and back faces. All other parts attach to the frame, including the input and output sides, and two partitions, all made from 0.25-inch brass (see **Figures 2** through **5**).

While a lot of holes must be drilled in the frame, you need make only two bends once the blank has been cut and drilled. The other sheet-metal parts require three bends, but are otherwise less complicated.
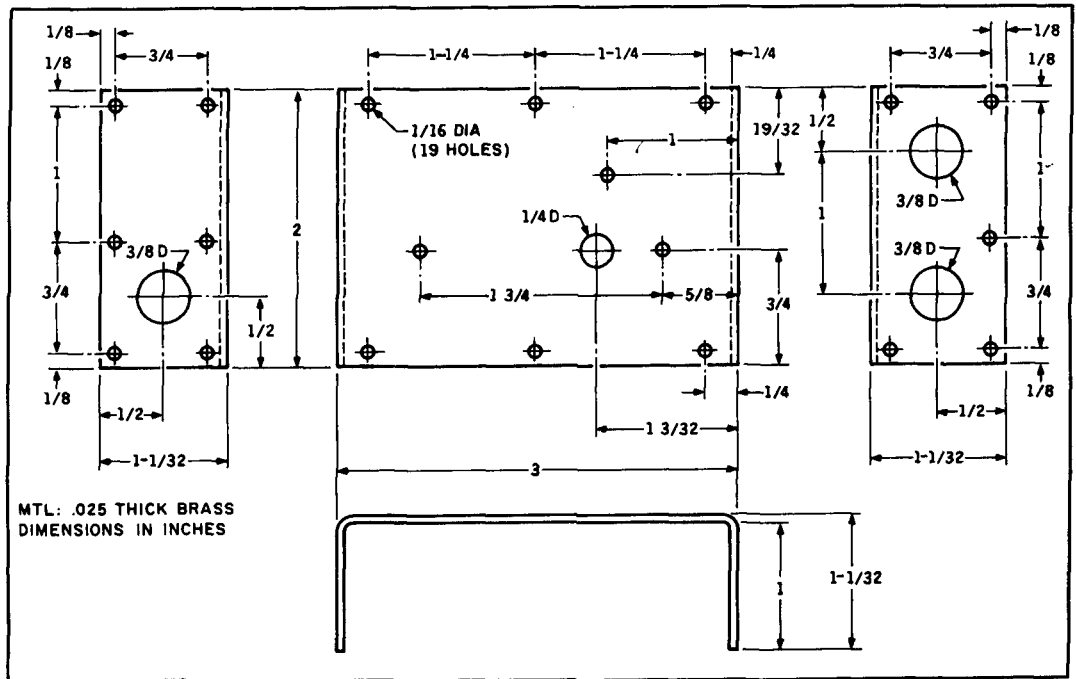
Figure 1. The brass U-shaped frame.

Figure 6 shows the blank form of the center partition before bending. The dashed lines are the bend lines. Note, however, that because the bend tends to increase overall length by approximately the thickness of the metal, the actual bend is made 1/32 of an inch closer to the center of the piece to compensate for this effect.

Because few amateur builders have access to a brake (let alone a shear), a bench vise with two lengths of 1 × 1-inch steel angle bracket clamped into it will do the job. Cut one 2-7/8 inch and one 7/8 inch length of angle bracket, using a hacksaw and plenty of lubricant. (This may be the most challenging part of this project.) Put the brackets into the vise and clamp the blank between them, leaving a 1/32-inch space above the bend line to account for the thickness of the material. Take a 1/2 × 1/2 × 12-inch steel bar and a

hammer and bend the blank into the shape of the brackets. After bending the frame, check to ensure that the overall length is exactly three inches. Both the bar and the angle stock are available from most builder's supply centers.

The frame includes nineteen 1/16-inch diameter holes for attaching the side plates and dividers. Note that the dividers don't initially include mating attachment holes. Instead, after the flaps are bent, each piece is held in position while you drill 1/16-holes, using the frame holes as guides to assure proper hole matching.

Once you've drilled the holes in the sides and partitions, enlarge the 1/16-inch holes in the brackets with a no. 44 drill and thread the mating 1/16-inch holes in the bracket with a 2-56 tap. This allows each piece to be temporarily attached to the frame with 2-56
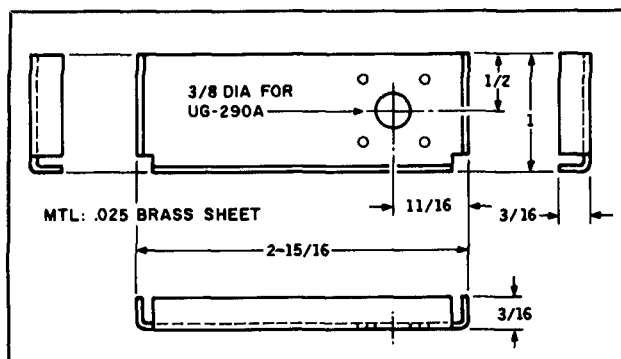
Figure 2. The input side.
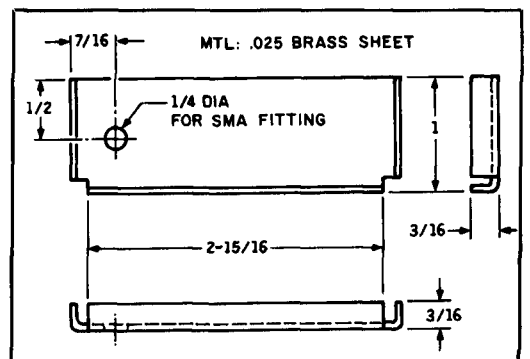
Figure 3. The output side.

stainless screws. These screws hold the assembly together while you sweat-solder the seams with a small propane torch or the heat from the kitchen stove. You'll remove these screws later.

It's easy to solder the parts together if you use a paste flux, like Kester SP-30. The paste causes the solder to flow evenly at each seam. Use the paste carefully because it contains zinc chloride. Read the cautionary note on the container before using.

When each seam is filled with solder, and while the piece is still hot, turn the box upside down and shake out the excess solder. Wear asbestos gloves to protect your hands.

After the box has cooled, remove the 2-56 stainless steel screws. Now clean the box, removing all traces of grease and rosin flux. Extra fine steel wool works well. I wrap a flat file with steel wool and use the combination as a scrubber.

I opted to silver-plate the box after cleaning it up, using the technique described in **Reference 4**. Plating may or may not improve performance, but it makes the finished product look more professional.

Whether or not you plan on plating the enclosure, first make sure the cover makes good contact all around the periphery of the box and along the partitions. Place a 12-inch square piece of emery cloth on a flat surface and, with the open end of the box down, sand the surfaces until there's no evidence of rough edges. I used a piece of 3M coarse (NE-4) crystal bay emery cloth. 3M also makes an aluminum oxide cloth with finer grit that works very well.

Varactor diode assembly

The varactor diode assembly components shown in **Figure 7** are mounted on a 1/4 × 3/4 × 1/16-inch thick fiber glass plate which has three 1/8-inch brass eyelets spaced 1/4 inch apart. First insert a pair of 1/2-inch long 2-56 brass screws in the outer eyelets and solder them in place. These screws will be used later to mount the SRD assembly on the center partition.

Insert a 0.090-inch diameter brass rod or tube in the middle hole and solder it in place exactly as shown in **Figure 7**. Now, solder a 5/16-inch long by 1/8-inch wide tab to the varactor-diode side of the 0.90-inch diameter brass rod or tube. This tab couples the 3312-MHz signal into the output cavity (see **Figure 8**). Solder the 1/8-watt 470-k bias resistor across a pair of eyelets, keeping the leads short. Finally, solder the prongs of the SRD between the other pair of eyelets. Use a hot soldering iron with a small tip, but don't apply heat for more than 6 to 10 seconds. I
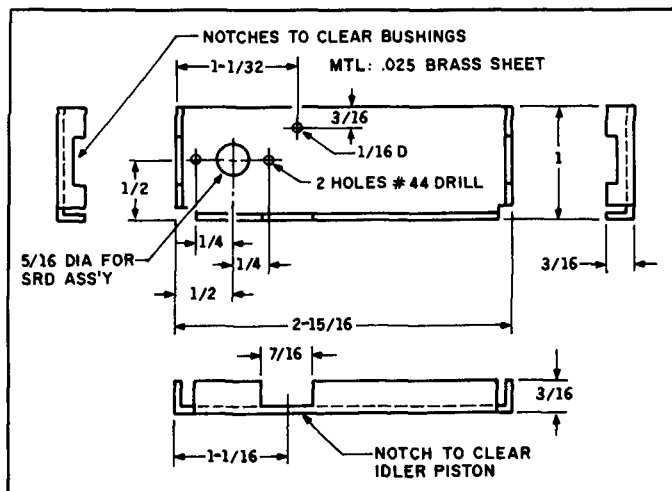
Figure 4. The center divider.
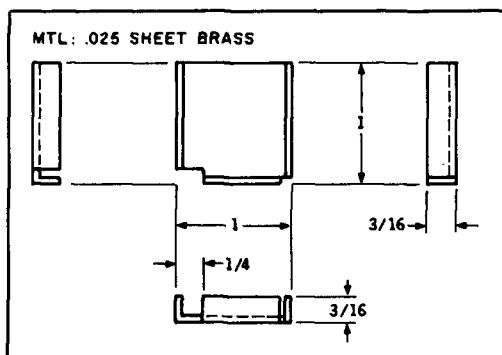
Figure 5. The output divider.

used a Weller TC201 with a no. 7 tip. If you have access to a jeweler's lathe, you may wish to machine small clamps for the SRD so you won't have to solder it into position. However, the direct soldering approach has worked fine for me on the three triplers I've built.

SRD input coupling

Feed the 2-56 screws of the varactor diode assembly through the two no. 44 holes of the center divider and secure them with 2-56 nuts. The SRD should be located closest to the front face of the tripler, as illustrated in **Figure 8**.

Couple into the SRD using a 0.010-inch thick piece of L-shaped tinned copper, as shown in **Figure 9**. Bend the short arm twice (see isometric sketch) and solder it to the trimmer's lug. Solder the long arm to the end of the 0.090-inch diameter coupling rod. The coupler should be spaced about 1/32 of an inch from the center conductor of the input cavity.

The input coupler serves a dual purpose. It capacitively couples 1104 MHz RF into the SRD and, because it resonates a 2208 MHz with the piston trimmer, it simultaneously
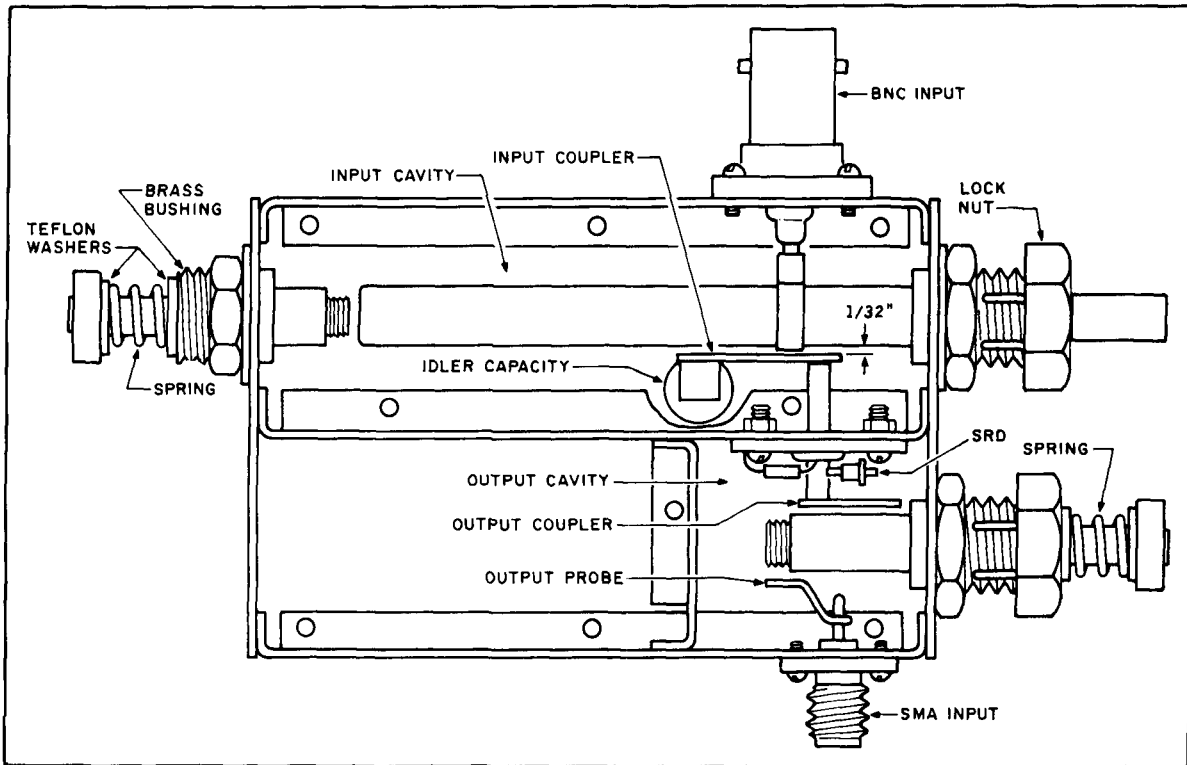
Figure 6. Preparing the center divider for bending.

performs the idler function as well. This simplifies the mechanical construction and tune-up procedure. The idler trimmer capacitor is a 0.6 to 12 pF JMC-5200 piston which extends up into the input cavity about 9/32 of an inch.

Tuning screws

Both the input and output cavities are tuned with spring-loaded 10-32 brass screws. Fit Teflon® washers on the screws to provide smooth tuning. Note that the input cavity conductor is positioned and locked into place with a jam nut, and tuning is done with the short screw at the rear. The RF input is connected directly to the center conductor of the input cavity with a copper strap. The 3/8-inch brass bushings were removed from discarded potentiometers.

Bottom cover

Preliminary tests can be made without a

bottom cover, but only to determine the approximate settings of the three tuning controls. The tripler won't perform satisfactorily until the cover is on. The cover drawing (**Figure 10**) shows attachment holes at appropriate spacings. The cover is held in place by seven 6-32 spade screws mounted on the outside of the enclosure at points of maximum RF current. I attached the double-hole spade screws to the enclosure with 3/16-inch long stainless 2-56 screws through tapped holes.

Output coupling

Both the SRD output coupler and the 3312-MHz output probe are shown in **Figure 8**. Both are made of 0.010-inch thick tinned copper strip, 1/8 inch wide. Drill one end of the probe to fit over the center conductor of the SMA connector. Highest output power is obtained with relatively loose coupling. Note the power output with the cover attached each time you adjust the coupling.

Tripler tune-up

Figure 11 shows the test setup used to tune up the 3312-MHz tripler. Most of my test equipment is surplus, which I keep in good working condition.

The 1-watt 1104-MHz signal derived from the 7768 tube amplifier is fed to the tripler's input for testing. Assuming the tripler delivers 0.5 watts, and the 431C power meter scale is set to 5 dBm, the meter will indicate 2 dBm when the tripler output is 1/2 watt.

Using coaxial patch cords, I connect the

Figure 7. Varactor diode assembly.

Figure 8. Bottom view—assembled tripler.

3312-MHz signal to the spectrum analyzer instead of the power meter to check the relative levels of 1104, 2208, and 3312 MHz, and any spurious signals that may be present. I can also see close-in noise and the noise floor level. The only spurious responses I found were traced to the PLO. These were caused by the 1.078125-MHz reference crystal signal feeding through in the PLO, producing sidebands 25 dB down on each side of the 3312-MHz carrier. While these were 50 dB down at 1104-MHz output, the nonlinearity of the tripler tends to in-

crease them relative to the carrier at 3312 MHz. Thus, it's important to make the PLO itself as clean as possible. The relative amplitudes of these spurious signals will be unchanged at the outputs of the up-converter and power amplifier, because these devices are quite linear.

Start the tune-up procedure by setting the 431C power meter scale to − 10 dBm and move the scale up as tuning progresses, keeping the reading on scale. Adjust the input and output cavities and the idler for maximum power output. Because these controls
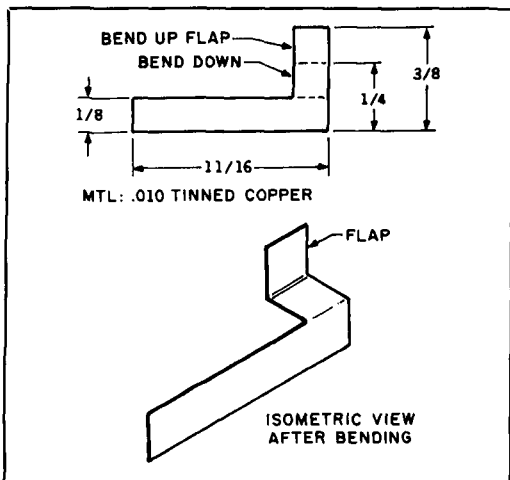
Figure 9. Input coupler.
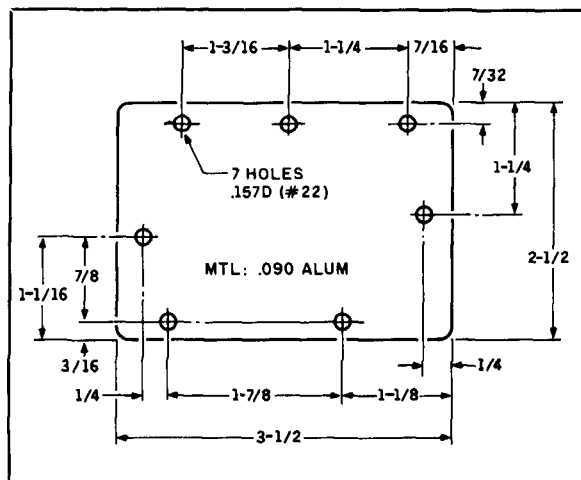
Figure 10. Tripler bottom cover.
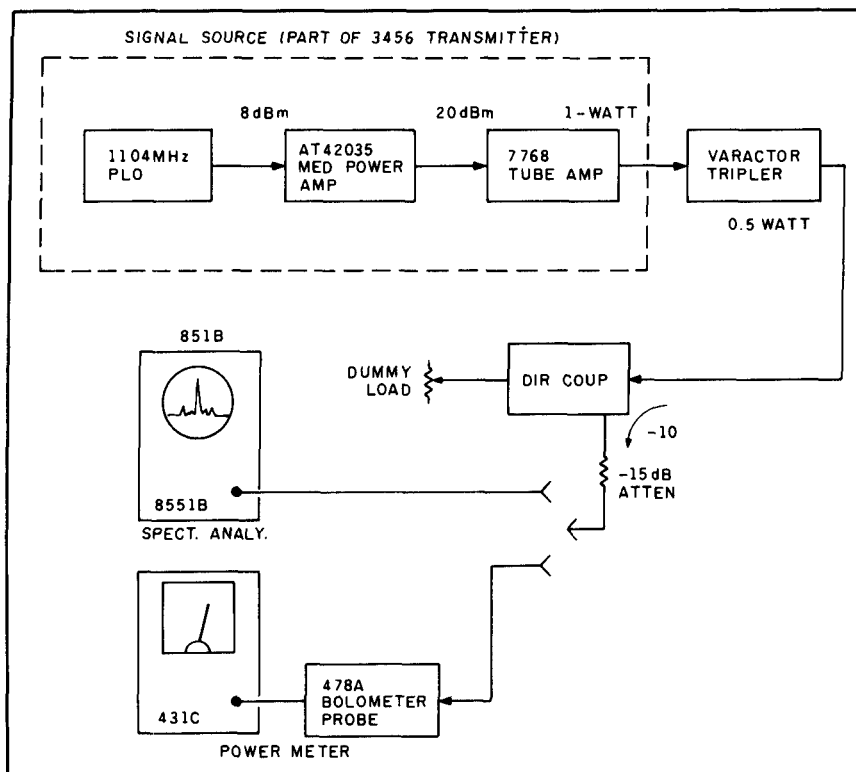
Figure 11. Test set-up.

interact to some degree, readjust all three. Two or three such iterations should suffice. The input cavity tuning may be the most critical.

At this point, tuning should be smooth and relatively broad. The only other adjustment you need to make is to the output coupling probe. Make sure the cover is screwed down tight after each probe adjustment. An ill-fitting cover may produce erratic results.

Final comments

The subject for this article was plucked from bygone days when the varactor was a popular device enabling amateur operation on the 1296 and 2304-MHz bands. Nearly two decades later, when the need arose for a relatively high power local oscillator for a surplus S-band tube up-converter, I found the SRD varactor quite useful.

When contemplating this project, I took inventory of my stock of varactor diodes. To my surprise, I found at least a dozen such devices gathering dust in my "goodie" parts drawer. I wonder how many amateurs like myself have a similar stock of diodes just waiting to be fired up again!

Even if you tend to be "all thumbs" when working with mechanical devices, you should have little difficulty building this SRD power varactor tripler if you follow the instructions provided here.

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1. Norm Foot, WA9HUV, "A High-Level SSB Modulator for S Band," *Communications Quarterly*, Summer 1991, page 71.
2. "A Simple Varactor Tripler for 432 MHz," *The Radio Amateur's Handbook*, Amateur Radio Relay League, 1973, page 217.
3. "Power Varactors Give 5-watts Output at 3 GHz," Motorola Data Sheet AN-176, 1966.
4. Norm Foot, WA9HUV, "Silver Plating Made Easy," *Ham Radio*, February 1977, page 42.

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Cordoba, Argentina

AN EPROM LOADER FOR DIGICOM

Modified game cartridge speeds the loading process

I think the DIGICOM program written by DL8MBT and DL3RDB is one of the best packet programs for the Commodore. While I don't own a C64 or 128, I am seriously considering buying one.

Until now, the only drawback to purchasing the Commodore was that I had to buy a drive in addition to the console, and those Commodore drives are *slow*. Loading the DIGICOM program takes a long time. So when a friend of mine asked me to look into the possibility of modifying a game cartridge into something useful, I grabbed the opportunity to try to put the program (I used version 3.50) into it.

How it works

Some digging into the Commodore documentation revealed the mechanism used in loading the cartridge contents. After some initializing, the startup ROM resident in the console checks to see if it finds a ROM (or EPROM, for that matter) at 8000h (hexadecimal, or 32768 decimal). To assure there is one, a certain code must be stored at the start of the ROM (see **Figure 1**). If that code is present, the code in the ROM is started instead of the resident BASIC interpreter.

It's at this point that two problems arise:
•The space available at this location is only 8 K (8192 bytes), while the program is about 30 K long.
•The DIGICOM program has to be run at its own address, in RAM, at 0801h.

I needed to perform some tricks to make it all work. The game cartridge has a mechanism that breaks the 32-K EPROM in-
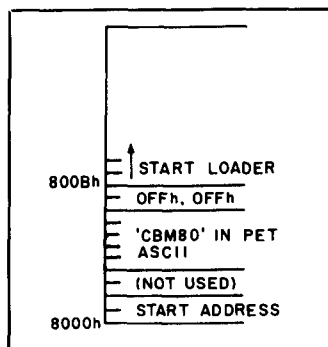
Figure 1. Startup code in ROM.

to four pages of 8 K each. That 8-K space at 8000h to 9FFFh provides a sort of "window" into the PROM, which lets you look at it a windowful at a time. You switch between the pages by writing a number from 0 to 3 into a "latch" which selects the page to be read.

You must put a small program at the beginning of the EPROM that copies 8-K chunks of code into RAM, starting at 0810h. This program switches between pages each time a chunk is copied, then finally jumps to the start address of DIGICOM at 0810h.

It's necessary to perform another trick here. The copying program works fine while copying the first page, but when it switches to the next, the program isn't there anymore! The solution is to copy the copy program into RAM. You can't put it into low RAM (0800h to 7FFFh) because DIGICOM will be loaded at this point. Fortunately, there's more RAM available (4 K of it) at C000h. The complete procedure is shown in **Figure 2**.

8000h — C64/128 STARTS CODE
AT 8000h

STEP 1 STARTUP

C000h

THIS CODE COPIES PART OF
ITSELF TO RAM AT 0C000h
AND JUMPS TO THIS CODE

8000h

STEP 2 COPY LOADER TO C000h

THIS ROUTINE THEN COPIES
REST OF PAGE 0 DOWN TO
LOW RAM (AT 0801h)

PAGE 0

STEP 3 COPY FIRST PART TO RAM

C000h

EPROM PAGE

8000h

PAGE 3

PAGE 2

PAGE 1

PAGE 0

0801h—

STEP 4
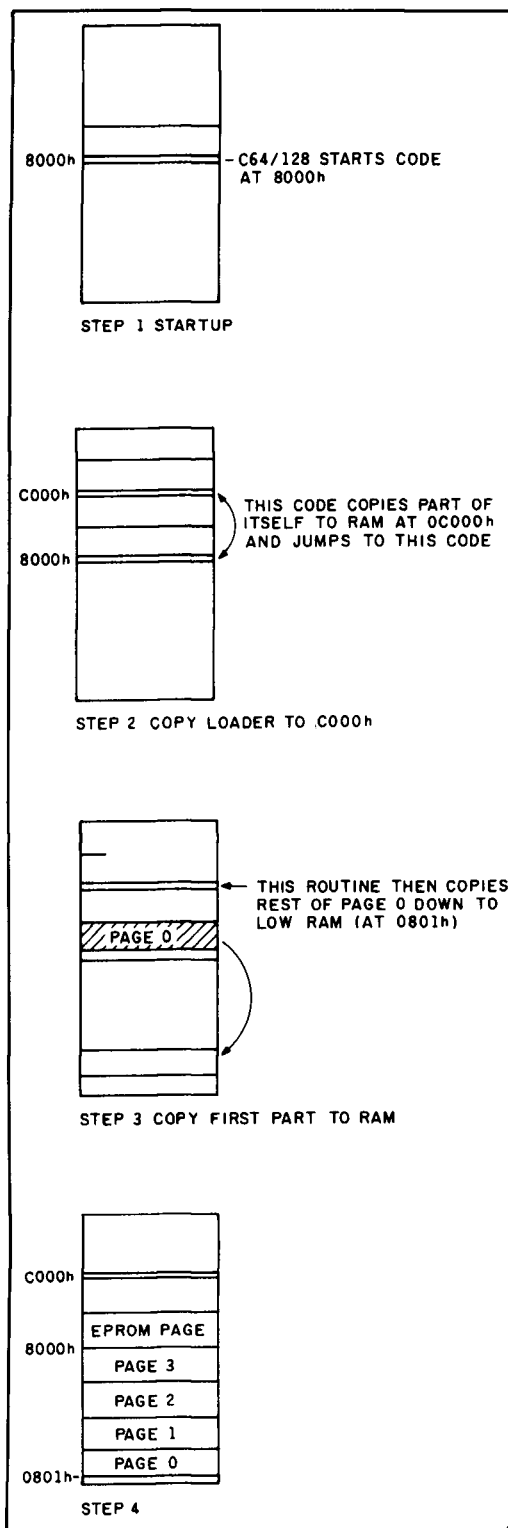
Figure 2. Loading process.

The hardware

It's quite easy to obtain a game cartridge, carefully unsolder the original EPROM, and reprogram it. But you'll need to be very careful, the quality of the PC board in these cartridges is rather poor. *Do check* to see if

the cartridge has a 256-K EPROM; without it there's no hardware to perform the "paging." The circuit of the cartridge I used is shown in **Figure 3**.

I suspect that most of these cartridges have similar hardware in them, so it probably won't be necessary to modify the loader program. The program isn't too long, and you may be able to adapt it if necessary.

Do install a 28-pin socket instead of the EPROM. This will make modification easier if a new version of DIGICOM appears or if other parameters need to be programmed. Check to see if the cartridge has plated-through holes. The one I had didn't have them and I had to solder on both sides. You could also use a wire-wrap socket, leave it a few millimeters above the board, solder it on both sides, and then cut off the excess pin length.

Programming the EPROM

I can't give you exact instructions on programming the EPROM, because our programmers will be different. But I can give you some general guidelines as to how to proceed.

First configure and prepare DIGICOM as you want to have it appear when you switch on the machine. Note that you don't have the option of using the configuration files on disk if you are running the EPROM version with the console only. (If you have a drive connected, it's still possible to call one of the configurations.) Make sure the video mode is properly set in the EPROM when the machine switches on. Set baud rates, your call, and anything else you desire.

Next program DIGICOM in the EPROM starting from 0101h (corresponding to 0801h of the program in the RAM). Transfer everything from 0801h - 7FFFh in RAM to 010H - 78FFh in EPROM. Then type the loader somewhere in RAM and program it starting from 0000h in the EPROM. Don't pass 0100h, or you'll overwrite the already-entered main program. **Figure 4** shows the contents of the EPROM.

The loader program

The complete program is shown in **Listing 1**. If you wish to recompile the listing you may need to make some changes, as this program was written for my own cross assembler on the PC. Differences, if any, will be minor.

In most cases, the hex listing will suffice to make the program work. Using a monitor, just type in all the bytes, recheck them, and program them into the PROM (see **Listing 2**).
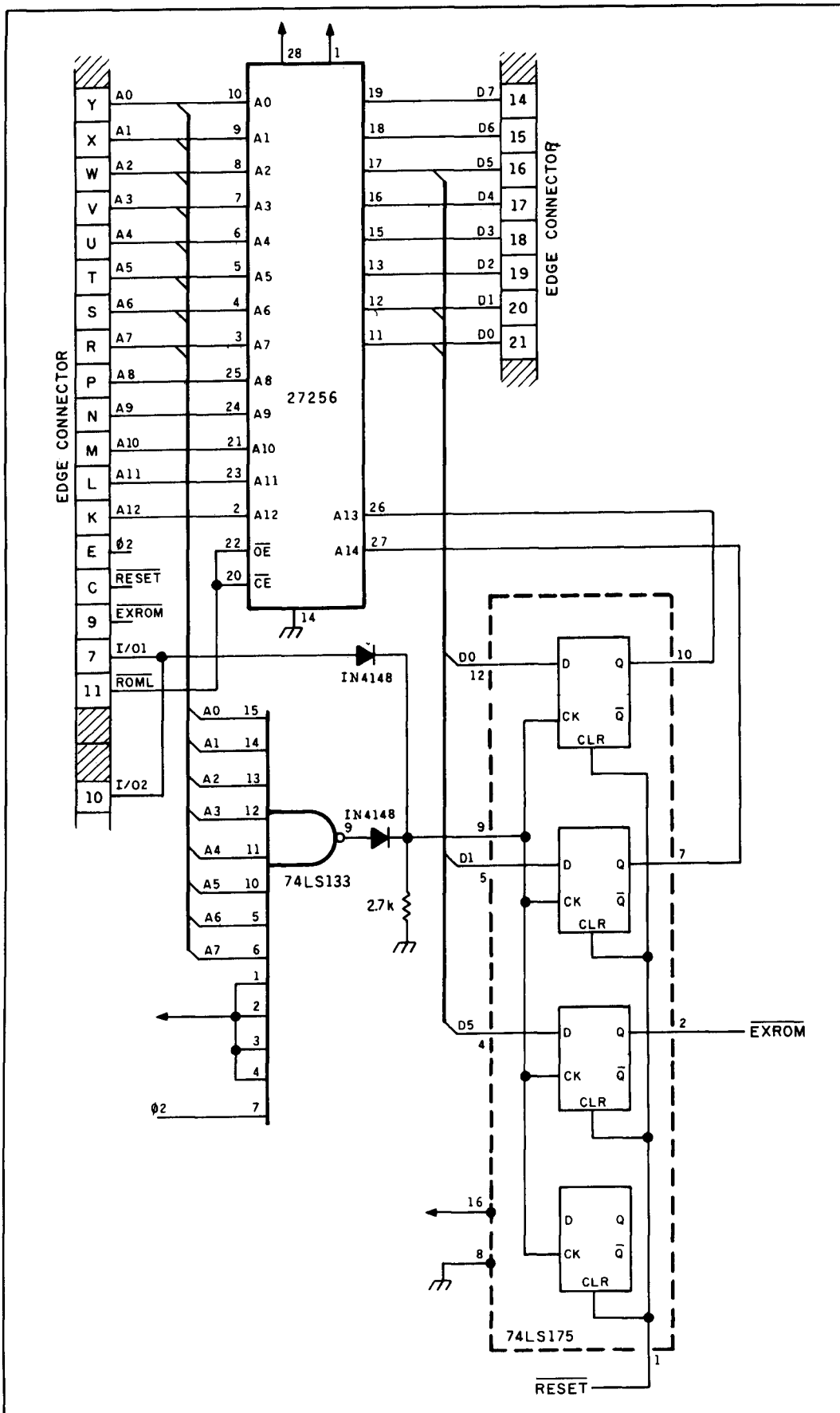
Figure 3. Diagram of the cartridge.

```
                              1              ORG      $8000
                              2
        8000                  3              DW       START
        8002                  4              DB       $BC,$FE,$C3,$C2,$CD,$38,$30
        8009                  5              DB       $FF,$FF
                              6
        800B  8E 16D0         7  START       STX      $D016
        800E  20 A3FD         8              JSR      $FDA3            ; IOINIT
        8011  20 50FD         9              JSR      $FD50            ; RAMTAS
        8014  20 15FD        10              JSR      $FD15            ; RESTOR
        8017  20 18E5        11              JSR      $E518            ; INIT SCREEN, KEYBOARD
        801A  20 53E4        12              JSR      $E453            ; COPY BASIC VECTORS
        801D  20 BFE3        13              JSR      $E3BF            ; INIT BASIC
                             14
                             15  ;------ First step, copy transfer routine to $C000
                             16
        8020  A0 00         17              LDY      #$00
        8022  B9 3080       18  MOVER       LDA      XFER,Y
        8025  99 00C0       19              STA      $C000,Y          ; MOVE BYTE BY BYTE
        8028  C8            20              INY
        8029  C0 38         21              CPY      #ENDR-XFER
        802B  D0 F5         22              BNE      MOVER
                             23
        802D  4C 00C0       24              JMP      $C000            ; START THE XFER ROUTINE
                             25
        8030               26  XFER
                             27
                             28  ;------ This is the transfer routine, now assembled at
                             29
                             30  ;------ ATTENTION: From here on must be relocatable!
                             31
                             32
        8030  A0 01         33              LDY      #$01             ; SOURCE POINTER
        8032  A9 00         34              LDA      #$00
        8034  85 C1         35              STA      $C1
        8036  A9 81         36              LDA      #$81
        8038  85 C2         37              STA      $C2
                             38
        803A  A9 00         39              LDA      #$00             ; DESTINATION POINTER
        803C  85 A3         40              STA      $A3
        803E  A9 08         41              LDA      #$08
        8040  85 A4         42              STA      $A4
                             43
        8042  A9 00         44              LDA      #0               ; PAGE 0
        8044  8D FFDE       45  MPAGE       STA      $DEFF            ; SET PAGE
        8047  48            46              PHA
                             47
        8048  B1 C1         48  MOVE        LDA      [$C1],Y
        804A  91 A3         49              STA      [$A3],Y          ; MOVE ONE BYTE
        804C  C8            50              INY                       ; INC DEST POINTER
        804D  D0 F9         51              BNE      MOVE
        804F  E6 A4         52              INC      $A4
        8051  E6 C2         53              INC      $C2
        8053  A5 C2         54              LDA      $C2
        8055  29 1F         55              AND      #$1F             ; CHECK FOR 8K BOUNDARY
        8057  D0 EF         56              BNE      MOVE
                             57
        8059  A9 80         58              LDA      #$80
        805B  85 C2         59              STA      $C2
        805D  68            60              PLA
        805E  18            61              CLC
        805F  69 01         62              ADC      #1
        8061  C9 04         63              CMP      #4               ; LAST PAGE LOADED?
        8063  D0 DF         64              BNE      MPAGE
                             65
        8065  4C 1008       66              JMP      $0810
                             67
        8068               68  ENDR
                             69
                             70              END
```

Listing 1. The complete program.

```
8000 0B 80 BC FE C3 C2 CD 38 30 FF FF 8E 16 D0 20 A3 FD 20 50 FD 20 15 FD 20
8018 18 E5 20 53 E4 20 BF E3 A0 00 B9 30 80 99 00 C0 C8 C0 38 D0 F5 4C 00 C0
8030 A0 01 A9 00 85 C1 A9 81 85 C2 A9 00 85 A3 A9 08 85 A4 A9 00 8D FF DE 48
8048 B1 C1 91 A3 C8 D0 F9 E6 A4 E6 C2 A5 C2 29 1F D0 EF A9 80 85 C2 68 18 69
8060 01 C9 04 D0 DF 4C 10 08
```

Listing 2A. The hex listing.

```
;1880000B80BCFEC3C2CD3830FFFF8E16D020A3FD2050FD2015FD200D88
;18801818E52053E420BFE3A000B930809900C0C8C038D0F54C00C00CB9
;188030A001A90085C1A98185C2A90085A3A90885A4A9008DFFDE480CCF
;188048B1C191A3C8D0F9E6A4E6C2A5C2291FD0EFA98085C26818691010
;08806001C904D0DF4C100803C9
;0000050005
```

Listing 2B. The hex listing in Commodore format.
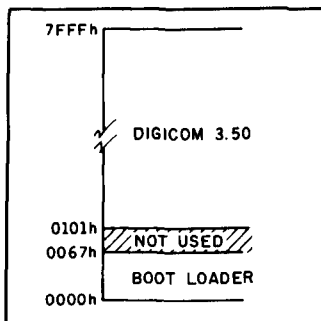
Figure 4. Contents of the EPROM.

Testing

To test the EPROM, plug it into the cartridge, plug the cartridge into the console, switch it on, and wait for the program menu to appear on the display. If the menu doesn't appear in a few seconds, check your hardware. It should be similar to that shown in **Figure 3**. If necessary, make changes to the program.

Conclusion

Having the program ready on power-up makes a world of difference. By using the cartridge, the machine becomes an almost perfect "packet" terminal—and it's inexpensive too!

If you have any problems, I'll be glad to help. Please feel free to write me at the address listed at the beginning of the article. If you'd like a reply, please include a few IRCs to cover the mailing costs.

PRODUCT INFORMATION

BBS for Antenna Patterns

Sinclair Radio Laboratories Inc. has established a bulletin board system (BBS) for antenna patterns. Designed to provide access worldwide, seven days per week, 24 hours a day, the new BBS offers on-line information about Sinclair antenna patterns, available software programs, and new products.

To obtain antenna information using your computer terminal and modem, call (716)874-2914. The service is provided free for the cost of the call, and information can be retrieved at a speed of 300, 1200, 2400 baud.

Users can view information by logging on and calling up files from main menus. To obtain specific information, users download the data onto their system via a password. Callers can obtain a password by selecting the word "induction" from the main menu, or by telephoning (800) 288-2763.

Sinclair Radio Laboratories Inc. is located at 675 Ensminger Road, Tonawonda, New York 14150.

New Sencore Catalog

Sencore Electronics has released its new Full Line Catalog. The 68-page catalog includes the latest in VCR, computer monitor, television, and cable system analysis equipment.

For a free copy call Sencore at (800)SEN-CORE, or writing Sencore Electronics, 3200 Sencore Drive, Sioux Falls, South Dakota 57107.

Rick Littlefield, *K1BQT*
109A McDaniel Shore Drive
Barrington, New Hampshire 03825

QUARTERLY DEVICES

Communications Quarterly's showcase for the latest developments in solid-state technology

Writing the "Quarterly Devices" column is a bit like putting on a talent show for up-and-coming solid-state devices. From the vast number of new products entering the marketplace, only a few will emerge as standard building blocks and take their place along side the venerable MPF-102 JFET, 741 op amp, NE555, and LM380 audio amplifier. The purpose of this column is to search out potential candidates, and present enough data to enable you to try them in a project.

To qualify as a *quarterly device*, a new product must offer a practical design solution for ham radio type projects. For example, it might provide low-IMD mixing, stable frequency control, low-noise amplification, or several useful functions on a single chip. Also, the device should be relatively inexpensive and easy to obtain. True winners will ultimately prove themselves as user-friendly performers in real-world applications like yours. Finding talent of this caliber isn't always easy. If you have a "hot tip" on a new device that fits these guidelines, I welcome your input.

Multi-function FM receiver chips

In this issue, I'll focus on a new FM receiver-on-a-chip from Motorola. The MC13135 dual-conversion narrowband FM IC is undergoing final tests in preparation for release to the market this spring. I'll also touch on another related Motorola device—the MC2833 FM transmitter IC.

History

The MC13135 is the latest entry in a long line of receiver-on-a-chip devices from various manufacturers. Receiver-on-a-chip ICs first appeared in the early 1970s with the introduction of the RCA CA3089. In recent years, Motorola's MC3359 and MC3361 single conversion packages—and the MC3362 dual-conversion IC—have found a comfortable place in the market, along with the Signetics LM604 and LM605. In 1991, Motorola updated the MC3359 and MC3361 with the release of the MC3371 (LC quadrature detection) and MC3372 (ceramic discriminator detection). In doing so, they removed the scan function and added a logarithmic RSSI (receiver signal strength indicator) output.

Those FM chips featuring RSSI capability have proven especially popular in low-cost instrumentation products such as the Smith Probe and Scientific Workshop Spectrum Analyzers. With each successive generation, FM receiver chips seem to improve and offer designers expanded opportunities to develop new construction projects, kits, and commercially manufactured amateur radio products.

The Motorola MC13135 narrowband FM receiver

Functionally, the MC13135 represents a new generation of the popular MC3362 FM receiver IC first introduced in 1987. Like the MC3362, the MC13135 is a complete dual-conversion receiver with local oscillators, first and second mixers, limiting IF amplifier, quadrature discriminator, meter circuitry, and audio amplification on a single chip. Both devices are inexpensive (under $2), come in a 24-pin package, operate
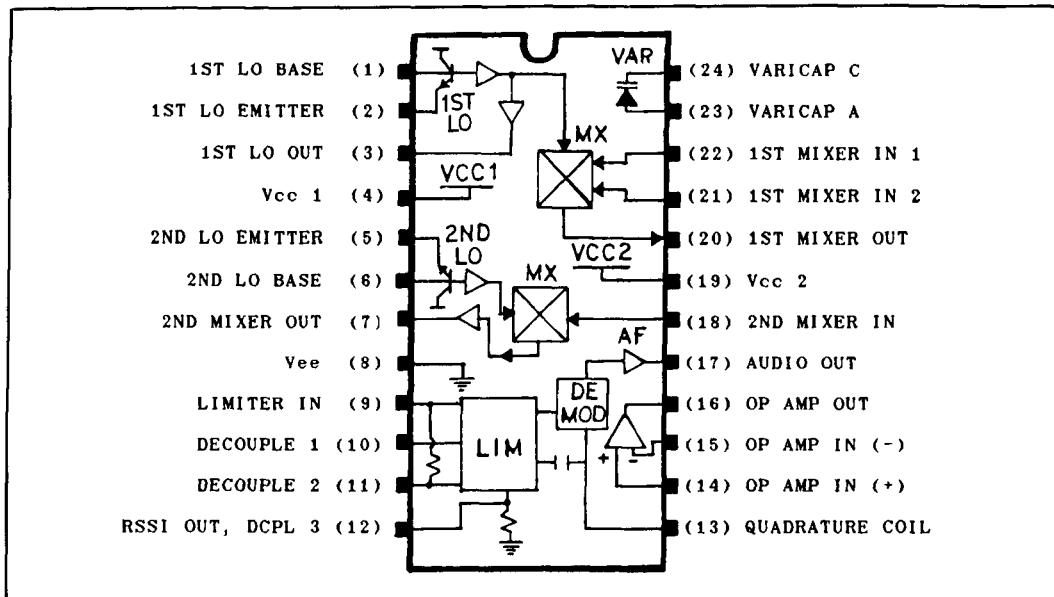
Figure 1. Internal structure of the MC13135 narrowband FM receiver IC.

The pin diagram shows the following connections:

Left side:
- 1ST LO BASE (1)
- 1ST LO EMITTER (2)
- 1ST LO OUT (3)
- Vcc 1 (4)
- 2ND LO EMITTER (5)
- 2ND LO BASE (6)
- 2ND MIXER OUT (7)
- Vee (8)
- LIMITER IN (9)
- DECOUPLE 1 (10)
- DECOUPLE 2 (11)
- RSSI OUT, DCPL 3 (12)

Right side:
- (24) VARICAP C
- (23) VARICAP A
- (22) 1ST MIXER IN 1
- (21) 1ST MIXER IN 2
- (20) 1ST MIXER OUT
- (19) Vcc 2
- (18) 2ND MIXER IN
- (17) AUDIO OUT
- (16) OP AMP OUT
- (15) OP AMP IN (-)
- (14) OP AMP IN (+)
- (13) QUADRATURE COIL

with very low power consumption, and are designed for use in battery-operated devices like cordless phones, handhelds, paging receivers, etc.

Despite these similarities, the MC13135 is configured somewhat differently than the MC3362 and offers significant improvements. For example:

• The MC13135 logarithmic-response RSSI circuit has a 75-dB dynamic range. This feature is especially useful for instrumentation applications like low-cost spectrum analyzers, DF receivers, and other precision signal-strength measuring devices.
• The mixer third-order intercept approaches − 10 dBm at 50 MHz, providing significantly improved resistance to intermod.
• The first local oscillator (LO) is a user-friendly colpitts circuit providing faster start-up and improved reliability with crystal control. Fast start-up is especially important for data transceiver applications.
• The audio output is greater; as high as 300 mV RMS at 25 ohms. This provides plenty of drive for an earpiece or line-level compandering device.
• The first IF works up to 21 MHz, permitting improved image rejection at higher VHF frequencies.

There are other minor improvements as well. For example, power consumption is slightly lower (3.5 mA), and improved gain distribution makes pc layout somewhat less critical.

Circuit description

An internal view of the MC13135 is shown in Figure 1. Nearly everything you

need to make a working receiver is contained within the device. As a designer, you'll supply Vcc, an RF input circuit, frequency control for the first and second LOs, band-pass filters for the IFs, a quadrature coil, and a gated audio amplifier to drive the speaker. Figures 2 and 3 show a functional schematic for a 50-MHz narrowband FM receiver and audio system. Here's a stage-by-stage rundown of how it works and what you'll need to make it play.

First (high) mixer

Referring to Figure 2, the first mixer is an active Gilbert Cell balanced configuration that converts incoming VHF signals to the first-IF frequency. With an input frequency of 50 MHz, the mixer provides about 10 dB of conversion gain to 10.7 MHz. Conversion gain is intentionally held low to improve IMD performance.

Overall receiver sensitivity at 50 MHz is specified at 0.7 μV for 12-dB SINAD. At 150 MHz, predicted sensitivity is somewhat lower (approximately 1 to 1.5 μV). Preamplification is probably unnecessary in the 50-MHz region because mixer sensitivity approaches the spectrum noise floor. However, weak signal applications in the upper VHF region may require a preamp. The maximum frequency of operation is specified at 200 MHz, but reception at higher frequencies is possible with LO injection from an external source.

The preliminary data sheet from Motorola rates the MC13135 mixer input impedance at 460 − j350 for 50-MHz input. At 150 MHz, the predicted input impedance
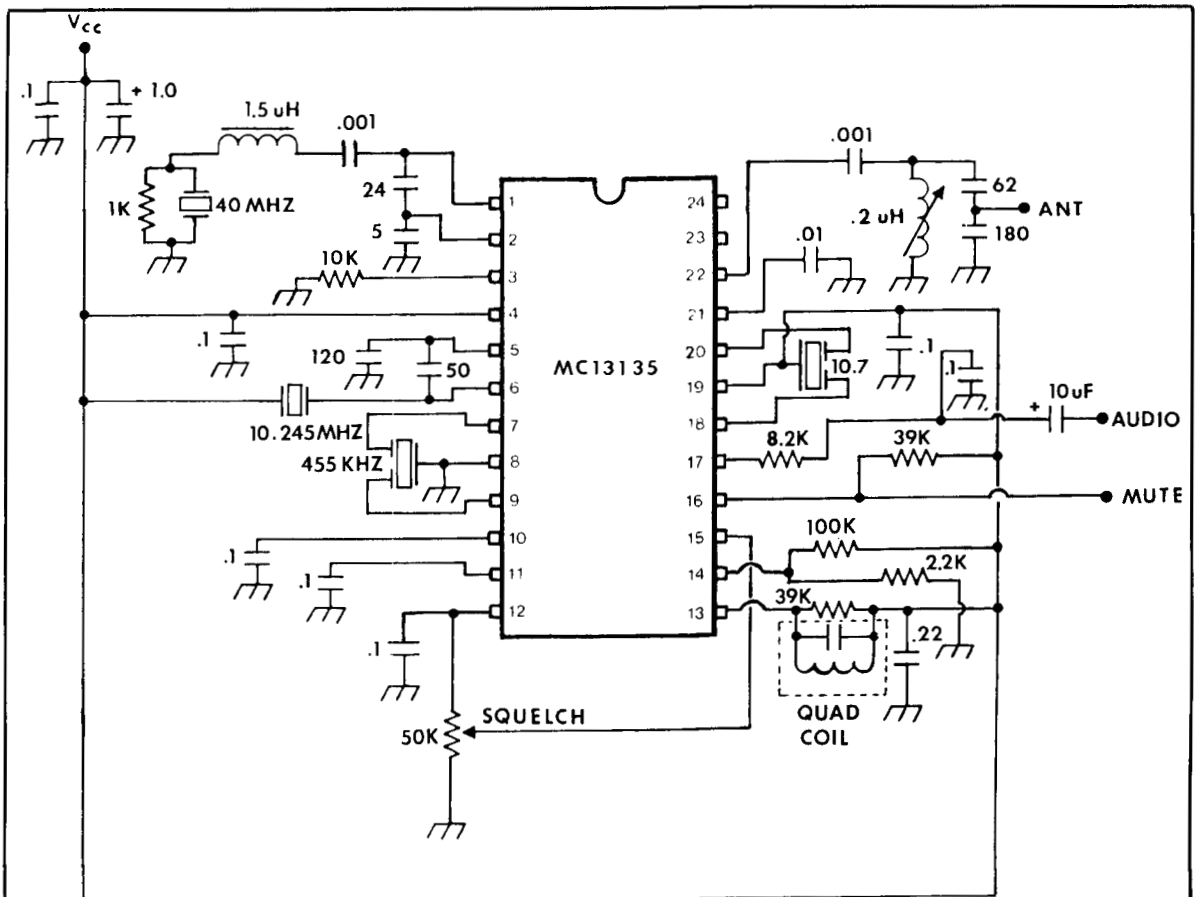
PARTS LIST
MC13135 50-MHz narrowband FM test receiver parts.

1 - MC13135 IC	1 - 120 pF
1 - 1.5 μH, Coilcraft SLOT TEN-4-02 or equivalent.	1 - 180 pF
	2 - 0.001 μF
1 - 0.2 μH, Coilcraft M1175-A	1 - 0.01 μF
1 - Quad Coil, Toko 7MC-8128Z	7 - 0.1 μF
1 - Xtal, 40 MHz range, 3rd overtone, series res. (ICM 471260)	1 - 0.22 μF
	1 - 1.0 μF
1 - 10.245 MHz Xtal, 25 ppm, 32-pF para., HC-18 case (ICM 434265)	1 - 10 μF
	1 - 1 k
1 - 10.7 MHz filter, muRata SFE10.7SM3-A	1 - 8.2 k
1 - 455 kHz filter, muRATA CFU455F2 (12 kHz), CFU455B2 (30 kHz)	1 - 10 k
	2 - 39 k
1 - 5 pF	1 - 50 k potentiometer
1 - 24 pF	1 - 100 k
1 - 50 pF	1 - 2.2 k
1 - 62 pF	

Figure 2. Schematic of the MC13135 50-MHz FM test receiver.

is somewhat lower than at 50 MHz—around 300 ohms. If you are designing for 2 meters, or for the 150-MHz region, use a 6:1 transformation to match the input into a 50-ohm system.

First (high) local oscillator

The first LO is an internally biased col-

pitts oscillator which provides several possibilities for frequency control. Access to the oscillator transistor base and emitter are provided at pins 1 and 2. A circuit for crystal control of the first LO is shown in **Figure 2**. Crystal control of the chip's onboard LO will generally be limited by the upper frequency range of readily available crystals. Third overtone types are typically
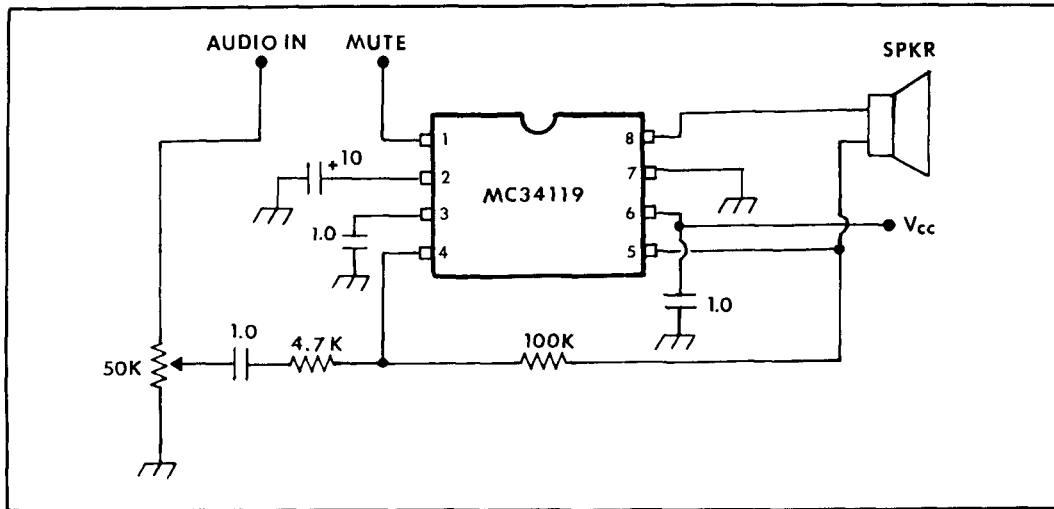
Figure 3. Schematic of the MC34119 gated audio amplifier.

available through 75 MHz; fifth overtone types are available through 150 MHz.

In addition to crystal control, the first LO also operates in an LC configuration that provides reception up to 200 MHz. In the LC mode, the oscillator tank can be voltage-tuned by the chip's onboard varactor diode. **Figure 4** shows a circuit configuration and a VCO linearity curve for LC operation in the 50-MHz region. The varactor is temperature compensated to minimize frequency drift due to ambient temperature changes. To ensure drift-free performance, the inductor you select must also be temperature stable. Make air-wound coils mechanically rigid and mount them securely. If you plan to use a premanufactured tunable inductor (shielded or unshielded), select one that approximates the desired inductance at its minimum value and remove the slug.

Ultimately, the MC13135 VCO may be driven by a synthesizer chip, DAC (digital to analog converter), or some other precision voltage source. However, for sake of experimentation, a good quality pot with a reduction drive—or a surplus 10-turn instrumentation pot—will work fine. Use fixed resistors on the high and low side of your pot to keep tuning within the linear range of the varactor diode (bias range for linear

tuning is typically 1 to 3.5 volts DC).

As a third frequency control option, you may externally drive the LO port of the first mixer through the oscillator transistor. Connect your external LO signal source to pin 1 through a 50-pF coupling capacitor, and set LO injection level for approximately 100 mV p-p. Avoid overdriving the LO input; this will degrade mixer IMD performance.

First (high) IF

Amplification at the first-IF frequency takes place within the mixer section's active-DBM network. Normally, 10.7 MHz is selected as the high-IF frequency because of the large number of filter options available to designers. However, frequencies as high as 21 MHz may be used in special applications requiring greater image rejection.

The first IF is especially designed to accommodate a wide range of ceramic or crystal bandpass filter types. Low-Z output to pin 20 and high-Z input on pin 18 ensure that filters of almost any input and output impedance will work as long as they are properly terminated. For general purpose FM applications, I recommend using a 180-kHz bandwidth ceramic device like the muRata SFE10.7MS3-A. Connect this between pins 18, 19, and 20 as shown in **Figure 2**. To be "technically correct," you may wish to install a 360-ohm terminating resistor at the filter output to minimize bandpass ripple. However, omitting this resistor provides a bit more IF gain without noticeably degrading performance.

Second (low) mixer

The second mixer is also an active Gilbert Cell balanced configuration providing ap-
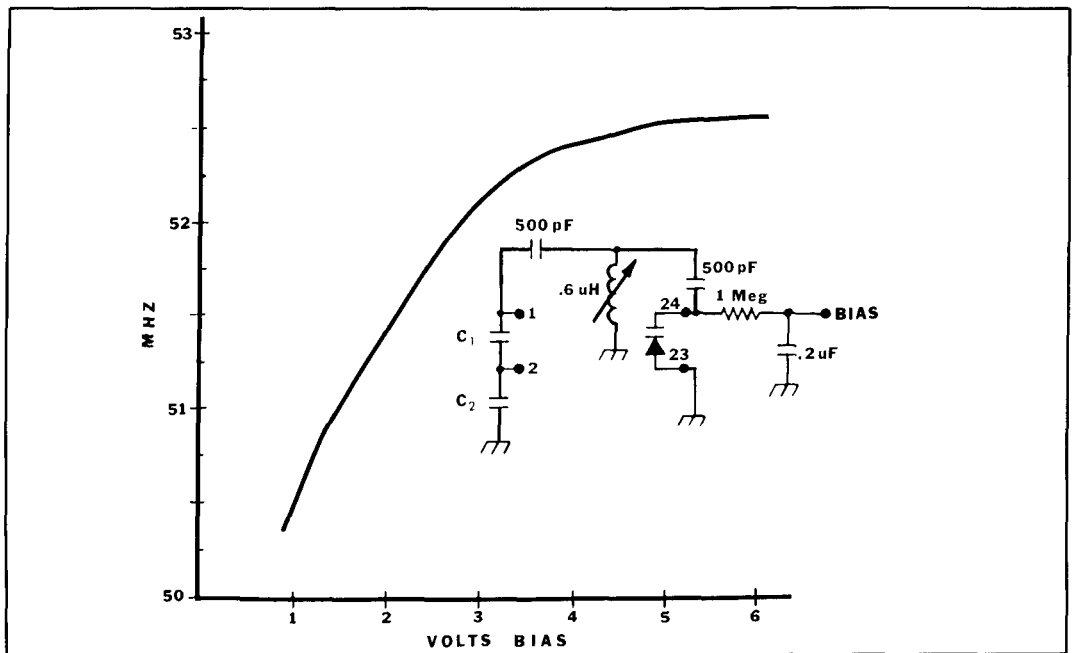
Figure 4. Schematic 50-MHz VCO tuning configuration using MC13135 onboard varactor diode, plus resultant tuning curve.

proximately 12 dB of conversion gain between 10.7 MHz and 455 kHz. The second IF may be operated as high as 3 MHz, but 455 kHz operation is typical for most designs because of the wide selection of low cost ceramic filters available at this frequency.

Second (low) local oscillator

The second LO is an internally biased colpitts oscillator. This oscillator is typically

crystal controlled to run at 10.245 MHz— the frequency required for conversion from 10.7 MHz to 455 kHz. Feedback capacitors and the crystal are connected to the second LO transistor via pins 5 and 6.

Second (low) IF

The second IF provides a large proportion of the receiver's overall gain plus signal limiting to remove AM components from
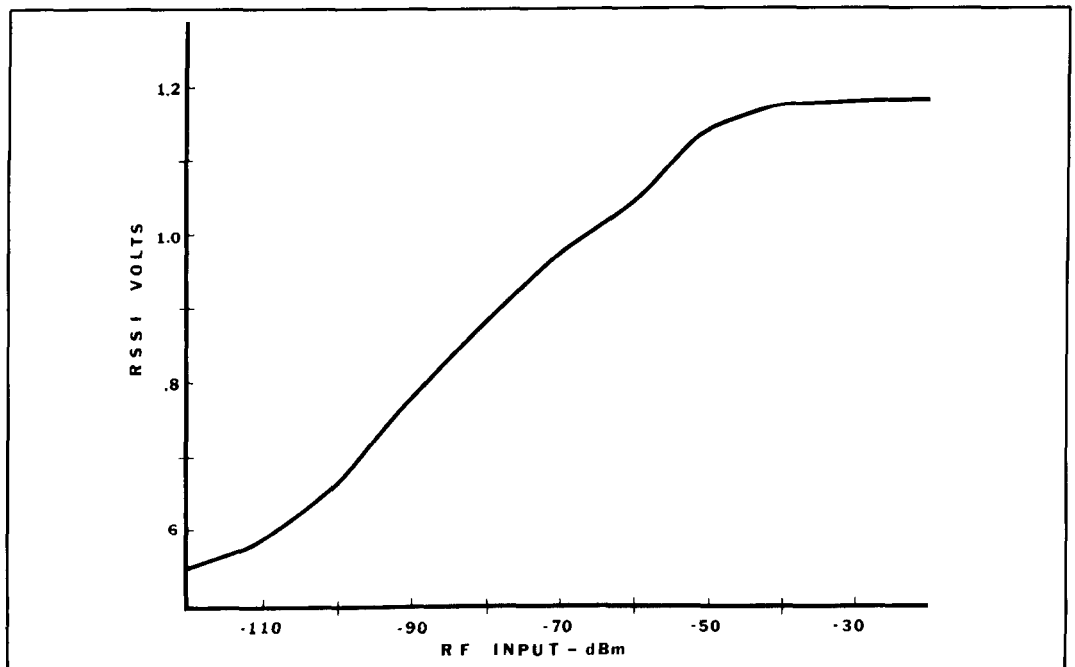
Figure 5. RSSI output in volts versus RF input in dBm at 50 MHz.

Lt   X1

MC2833

1

Variable
Reactance   RF
Osc.   16

56 p

2   15   51 p

4700 p

100 k

3   Buffer   14   47 p   0.22 µ   C1   Cc1

4700 p

Mic Amp   13   390 k

Deviation
Adjust   100 k   4   1.0 k

120 k   Q2   12   C2

Audio
Input   Dynamic
Mike   2.7 k 1.0 µ   5   L1   1000 p

11   C3

4.7 k   6

1.0 k   Re1   7   VREF   10   470 p

470 p   Q1   RF Output
-5.0 to -10 dBm
(see Note 4)

Electret
(alternate)
Microphone
and biasing   8   9   C4
C5
L2

Rb1   Cc2

1.0 µF
tantalum

VCC   9.0 Vdc

PARTS LIST
MC2833 FM transmitter.

1 - MC2833	1 - 51-pF silver mica
1 - 0.22 µH, Coilcraft M1175-A	1 - 56-pF silver mica
2 - 1 k	2 - 470 pF
1 - 2.7 k	1 - 0.001 µF
1 - 4.7 k	2 - 0.005 µF
1 - 100 k	2 - 1.0 µF
1 - 120 k	1 - Xtal, 25 ppm, 32-pF parallel,
1 - 390 k	HC-18 case (ICM 434265)
1 - 100-k trimpot	

See Table 1 for special parts that determine 6 or 2 meter operation.

Figure 6. Internal structure and schematic of the MC2833-based narrowband FM transmitter.

the FM signal. An external 455 kHz band-pass filter is inserted between the ICs amplifier and limiter stages at pins 7, 8, and 9.

For narrowband FM voice operation, I suggest using a filter with fairly tight bandpass characteristics such as the muRata CFU445F2. This filter provides a 12-kHz bandpass at -6 dB. For high-speed data or spectrum analysis, a wider filter might prove more desirable. The muRata CFU455B2 provides a bandpass of 30-kHz at -6 dB.

Quadrature detector

Output from the receiver's limiter stages is demodulated by quadrature detection. The quad coil is connected between Vcc and pin 13, as shown in **Figure 2**. Almost any 455-kHz quad coil might work here, but I suggest using a Toko 7MC-8128Z 7-mm inductor to ensure good audio recovery and drift-free detector performance.

As a "no-tune" alternative to the quadra-

ture-based IC, Motorola will offer a MC-13136 receiver chip employing a high-level ceramic discriminator later this year. Preliminary data on that new IC should be available by the time you read this column.

Squelch

Unlike some FM chips, the MC13135 doesn't have a conventional noise-operated squelch circuit. Instead, RSSI drive is used to trigger an audio gate to satisfy the squelch function.

To keep the squelch simple, I suggest pairing the MC13135 with the Motorola MC34119 low-power audio amplifier IC.[1] The MC34119, shown in **Figure 3**, is a low-voltage 200-mW amplifier IC originally designed for speakerphone amplification. One particularly attractive feature of the MC34119 is a built-in audio gate (also called a "disable function"). When this gate sees a low logic level (<0.8 volts) on pin 1, the chip amplifies. When the gate goes high (>2.0 volts), the amplifier is switched into a power-down state. This feature makes a great squelch!

The MC13135 generates disable voltage for the MC34119 gate via an undedicated op amp available at pins 14, 15, and 16. To accomplish this, the op amp is configured as an inverting comparator (level detector). One comparator input is connected to the output of the RSSI, while the other is biased by R1 and R2 to a suitable reference level. Any time the comparator circuit detects a received signal through a rise in RSSI output, its output goes low—switching the gate of the MC34119 into the enable mode. When the incoming signal disappears, the RSSI output drops and the comparator goes

high—disabling the amplifier. The comparator or "squelch" threshold level is set via a 50-k linear pot.

Audio output

Recovered audio (pin 17) is passed through an RC low-pass filter and coupled to a power amplifier through a DC blocking capacitor. A practical audio output circuit employing the MC34119 is shown in **Figure 4**. A 50-k audio-taper pot provides volume control. Loop gain of the MC34119 is controlled by the 100-k feedback resistor connected between pins 5 and 4. Note that the MC34119 has differential output, and no coupling capacitor is used between the amplifier and speaker. To prevent instability, speaker leads should be formed into a twisted pair and kept relatively short—12 inches or less. If you install a headphone jack, isolate it from chassis ground.

RSSI

The RSSI voltage appearing at pin 12 is proportional to the log of received signal amplitude. This means you can accurately depict the strength of any incoming signal between −115 dBm and −40 dBm on a buffered voltmeter or oscilloscope in dB. **Figure 5** shows a typical response curve for RSSI output versus signal input. In serious instrumentation applications, the MC13135 undesignated op amp may be configured as an RSSI linear line driver rather than as a comparator. The RSSI output may also be used for AM signal detection.

MC13135 applications

One reason I chose the MC13135 for this

ITEM	6 METERS	2 METER
Xtal	16 MHz	12 MHz
Lt	3.3 to 4.7 μH Coilcraft M1289-A	5.6 μH Coilcraft M1312-A
L1	0.22 μH Coilcraft M1175-A	0.15 μH Coilcraft M1284-A
L2	0.22 μH Coilcraft M1175-A	0.10 μH Coilcraft M1283-A
Re1	330 ohms	150 ohms
Rb1	390 k	220 k
Cc1	33 pF	47 pF
Cc2	33-pF silver mica	10-pF silver mica
C1	33-pF silver mica	68-pF silver mica
C2	470 pF	1000 pF
C3	33 pF	18 pF
C4	47-pF silver mica	12-pF silver mica
C5	220-pF silver mica	33-pF silver mica

Table 1: Components versus output frequency
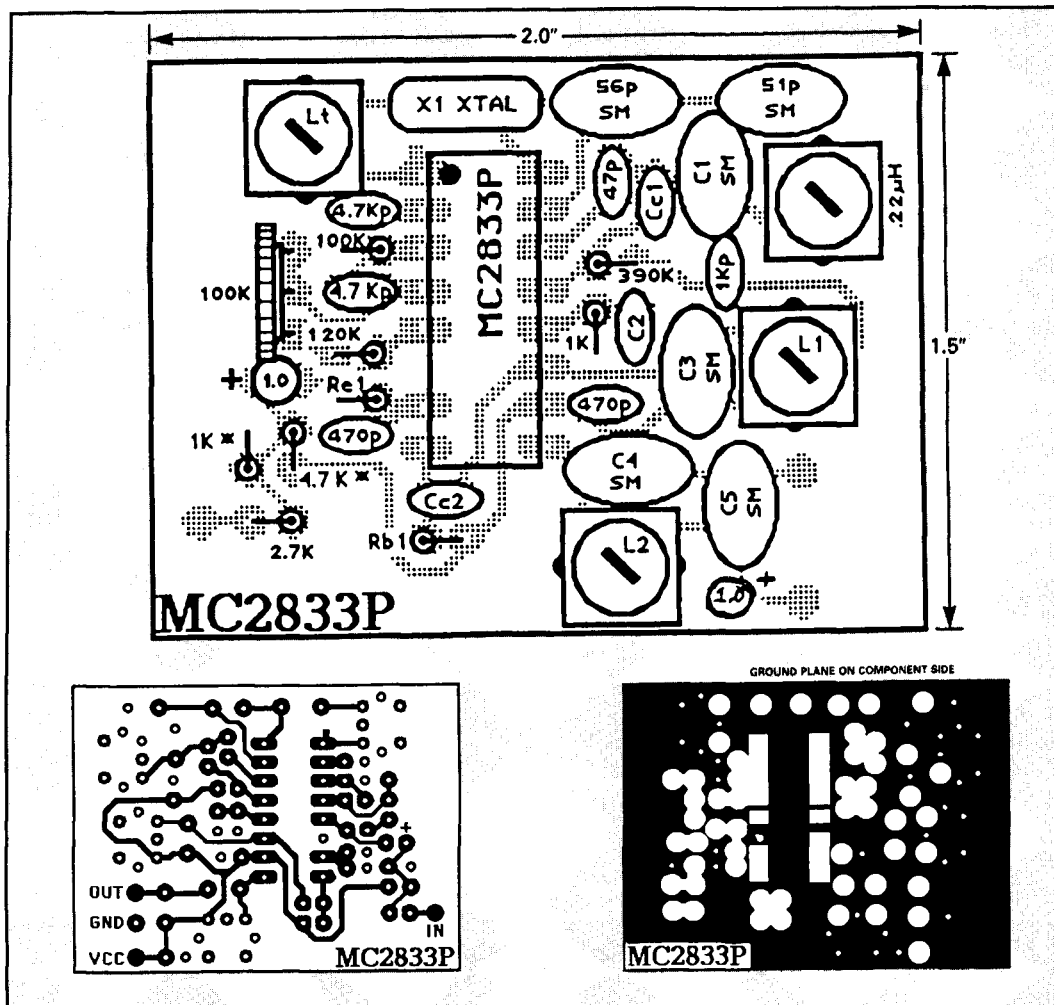
Figure 7. Parts placement layout and pc board art for the MC2833-based narrowband FM transmitter.

first installment of "Quarterly Devices" was because it's new. So new, in fact, that updated performance data from Motorola was being incorporated into this article just hours before deadline! But, by the time you read this, the MC13135 will be ready for market—complete with finalized application notes and engineering samples. You may obtain these directly from Motorola.

The other reason I selected the MC13135 is because this most recent generation of FM receiver ICs can be applied to a number of especially interesting construction projects. For example, with RSSI output that's essentially linear over a 75-dB range, these chips are perfect for a high-sensitivity antenna range field-strength meter, high-resolution spectrum analyzer, calibrated ADF receiver, etc. Furthermore, the MC13135 (and others) seem ideal for a simple 6 or 2 meter FM transceiver. When is the last time someone on your local 2-meter repeater said, "The rig here is homebrew?" Well, with inexpensive chips like the MC13135

and MC2833 transmitter available, why not? Which brings me to the Motorola MC2833.

MC2833 transmitter chip

The MC2833 FM transmitter isn't a new product; it's been available since 1987. It is widely used in many familiar consumer devices, and has been applied to innovative amateur radio applications like the Kantronics Data Transceiver. I recently built a 2-meter test exciter using the MC2833 and found it both easy and forgiving to work with. Best of all, it sounds great on the air—and would make a fine companion transmitter for the MC13135 in a homebrew voice or data transceiver! Without going into a lot of detail, here's some information to help you get one up and running.

MC2833 circuit description

The MC2833 is an FM-transmitter-on-a-chip designed for use in cordless telephones

and battery-operated communication transceivers. It operates over a wide range of voltages (2.8 to 9.0 volts DC) and draws only about 3 mA. On 2 meters, it uses inexpensive 12 or 18-MHz crystals and will deliver about +5 dBm of RF into a 50-ohm load with a Vcc of 8 volts.

The schematic in **Figure 6** shows a block diagram of the chip's interior—plus Motorola's recommendations for attaching external components. The chip contains a crystal oscillator and buffer, speech amplifier which adapts to a dynamic or electret microphone, reactance modulator, and two auxiliary transistors for frequency multiplication. **Table 1** provides the specific component values Motorola recommends for operation on 6 and 2 meters.

I built my version of the MC2833 transmitter on a prototype board which is available from Motorola. (You can get one by calling them at 602-897-3615 and requesting the MC2833 prototype kit.) If you wish to make your own board, pc pattern art and parts placement is given in **Figure 7**. I substituted a mix of monolithic and disk ceramic capacitors for the silver micas specified on the data sheet with no apparent degradation in transmitter performance. If

you decide to use an 18-MHz crystal instead of the specified 12-MHz type on 2 meters, reduce the values of the 51 and 56-pF capacitors to 33 and 39-pF, respectively. The chip's speech amplifier may be driven with a low-Z dynamic or electret microphone. If the electret element has a separate power lead, connect it to Vcc. All that remains is adding enough RF power amplification to hit the repeater of your choice.

Conclusion

I hope this first installment of "Quarterly Devices" conjures up visions of exciting new test gear and FM transceiver projects. If so, perhaps you'll share your inspirations with us here at the quarterly. Special thanks to Motorola's Rich Potyka, W7DTL, for his technical assistance and willingness to let us "peek under the tent" at the MC13135. Next issue, I'll tackle low-power audio amplifier ICs. The LM386 is nice, but there are a number of new devices on the market (like the MC34119) that might provide a better fit for your next project.

REFERENCES
1. Rodney Kreuter, WA3ENK, "2-Meter Monitor," *Ham Radio*, July 1988, page 64.

PRODUCT INFORMATION

High-Resistance Meters for Bench-Top and System Applications

Hewlett-Packard has introduced two high-resistance meters: the one-channel HP 4339A, designed for bench-top and automated-system applications; and the four-channel HP 4349A, designed primarily for automated-system applications. The HP 4339A has a 1,000-ohm to 10E16-ohm measurement range; the HP 4349A has a 1,000-ohm to 10E15-ohm measurement range.

The HP 4339A provides outputs from 0.1 to 1,000 volts DC, uses an ammeter to mea-

sure the current flow, and displays the results. Five digits of resolution can be displayed for resistance, current, volume-resistivity, and surface resistivity measurements.

The HP 4349A offers precise resistance and current measurements. Option 001 deletes two channels from the instrument. External power supplies must be used for each channel, allowing users to select the power source that best meets their needs.

For more details on either high-resistance meter, contact Hewlett-Packard Company Inquiries, 19310 Pruneridge Avenue, Cupertino, California 95014

Marv Gonsior, *W6FR*
418 El Adobe Place
Fullerton, California 92635

TITAN 425 RELIABILITY ENHANCEMENTS

Modifications that really pay off

The TEN-TEC Titan model 425 linear amplifier is a high quality all-brand, grounded grid 1500-watt peak-envelope power (PEP) output design, which easily meets all its performance specifications. More than 2000 of these units are in use worldwide. It incorporates a comprehensive range of conventional protective circuits found in good design, including high-voltage interlocks and shorting bars, overdrive visual warning and bias fold-back, automatic-level control (ALC) generation, and step-start.

Here are some ancillary modifications and suggestions to enhance the Titan's performance while increasing its survivability and reliability. Some of these basic concepts have universal application in linear amplifier design.

Cooling protection

Cooling protection for the amplifier's two tubes—the excellent, but expensive, Eimac 3CX800As—is one worthwhile addition you should consider providing. Should a rare fan failure* or serious tune-up problem occur without your knowledge, you could be faced with a catastrophic dollar loss and be knocked off the air at an inopportune time. Therefore, it's prudent to protect this valuable investment in every way possible.

I investigated a number of alternative cooling protection methods. These included thermal cutoffs, snap switches and fuses, air-pressure sensing, and a diode and a thermistor bridge. Finally, I decided on the circuit shown in **Figure 1**. It's a multifunctional

combination of a reasonably fail-safe and noninvasive visual remote-temperature readout and audio alarm. The circuit uses a low-cost LM324 quad op amp which performs three functions: U1A, bridge sensor-thermometer; U1B, audio oscillator; and U1C and U1D, push-pull audio amplifier. The total power requirement, at 12 volts, is approximately 20 mA when the oscillator and amplifier are on, excluding the regulator current. In the quiescent state, power consumption is only 2 mA.

Because of voltage drift in power supplies and/or the effect of zener-diode regulators on the calibration, I used a three-terminal regulator. The plus voltage may be either 12 or 15 volts, as chosen by the regulator. Compensation is provided in the calibration by R3. This sensor may be used with any power amplifier or other heat source, and requires no modification in its application. I built it on a 3 × 4-inch single-sided copper clad perfboard, with 0.10-inch holes on centers, using a pad cutter. Many of the connections can be made directly at the pin-outs of the wire-wrap type IC socket used. Note that a number of the connections are common to each other so they can be made easily and directly at the socket.

Sensor circuit

The basic parameters and functions of the sensor are as follows: Resistor R1 is a garden-variety, negative coefficient thermistor. When its value is 2k, which is equal to R2, the point of maximum sensitivity—75 degrees C—is reached. Resistor R3 establishes the meter calibration point for the

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*TEN-TEC reports that it has not experienced cooling-fan failures in the Titan model 425 linear amplifier. Ed.
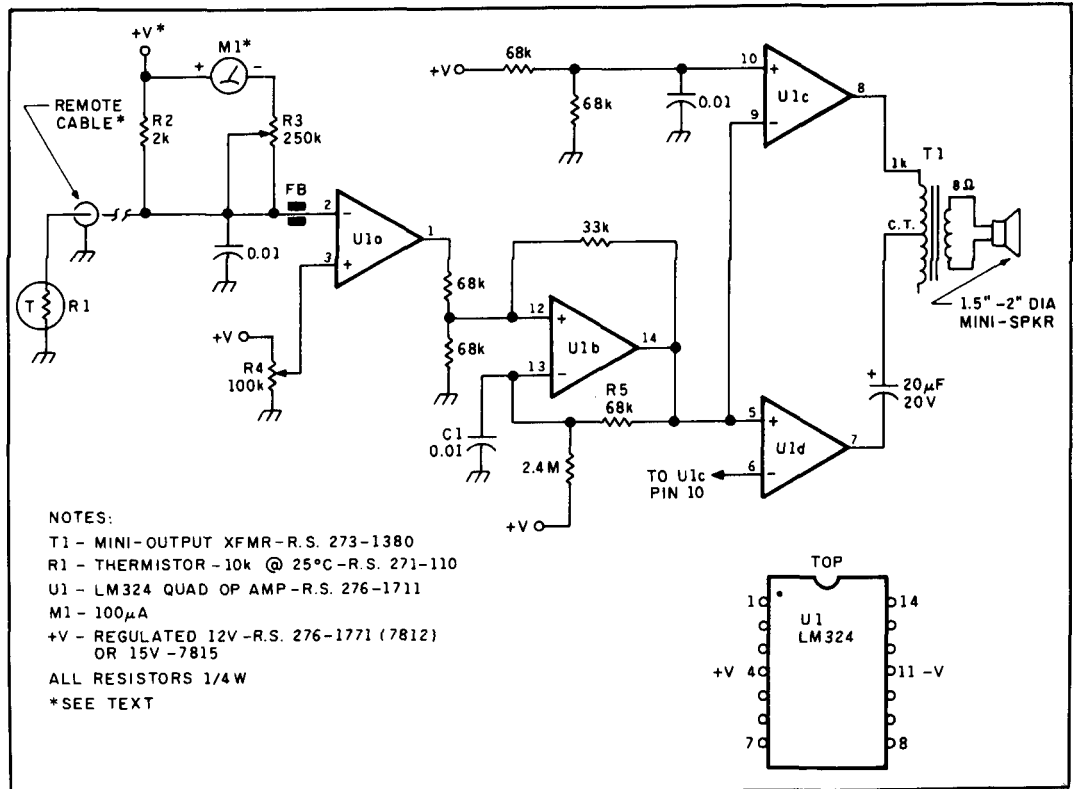
Figure 1. Temperature sensor schematic.

microameter, turned "thermometer," which is calibrated for temperature; R4 sets the trigger threshold for the tone oscillator. Capacitor C1, in conjunction with resistor R5, establishes the tone-oscillator frequency of approximately 800 Hz, which is in the most sensitive frequency range of the adult human ear.

The values shown in **Figure 1** are for my particular meter—a 3-inch Simpson model 27 with a 100-μA full-scale movement. This scale is most convenient and provides a calibrated range from 0 to 100 degrees C. Any other relatively low-range microameter will do, but you'll need to adjust R3 for meter calibration. The temperature tracking between my meter and a good thermometer was within 2 degrees C at all points between 25 and 80 degrees. The cable for remoting the sensor circuitry to thermistor R1, may be any shielded type. I used 174A/U coax with some ferrite beads for decoupling at the connection to U1A. External power supply leads should be similarly decoupled as a precaution.

Thermistor installation and meter calibration

Position thermistor R1, when connected to the remote cable, at one of the exhaust openings with appropriate insulation. Use one of the large screws holding the two

screens protecting the exhaust ports, or a small piece of tape, to secure the cable with a clamp to the top panel. Once this is done, calibrate the meter scale with 25 μA representing 25 degrees C, as I did, or a calibration of your choosing. Note that there will be a very slight increase in temperature indicated at first turn on. This is normal, as the thermistor is actually dissipating a few milliwatts through the circuitry. The calibration may be done with a heat gun or hair dryer for the high end, in conjunction with a good thermometer to adjust R4 for the trigger point of the audible warning tone. Once you've made these "set-and-forget" adjustments, you can feel a little more comfortable with respect to thermal protection for the Titan or any other amplifier.

Air-pressure sensor

An air-pressure sensor provides an alternative, and more positive, method of fan-loss protection. A low-torque, rotary micro-switch combined with a small "sail" is activated by the air pressure in the plenum.* This type of protection is frequently used in commercial equipment. I recommend an Omron type D2MC-01E (S1, **Figure 2**) with clockwise rotation (D2MC-01EL counter-

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*Enclosure inside which air is at a pressure greater than that outside the enclosure. Ed.

clockwise rotation). It makes an ideal switch, as its torque requirement is only 5.1 gm-cm. This is critical because its rotating action, under air pressure, is essential for reliable operation. It may be interlocked with any circuit as the contacts are in a single-pole, double throw configuration. The cost of the switch is minimal. It's available from distributors.* It's very small and can be mounted in place with silicone cement (RTV) or double-backed tape, so no drilling is required. Any additional control leads which go through the plenum, should be thoroughly decoupled for RF using three terminal capacitors. A suggested layout is shown in **Figure 2** along with a drawing of the "sail." I've used this system in my home-brew linear amplifier for many years with complete success.

Inrush protection

The basic design of the amplifier's high-voltage power supply properly takes into account the problems associated with relatively high inrush currents experienced with cold, low-resistance heaters and the charging of large capacitors, because the two are involved on a common core in the power transformer.

The stress caused by the surge-current limitations of the diode rectifiers is also taken into account, and a step-start circuit is used (see **Figure 3**). This circuit consists of relay K7, whose time constant for closure is established by parallel resistors R3, R4, and R5 in conjunction with capacitor C5. When activated, the relay shorts out resistor R1 in

*Omron distributors: (800)626-6676.

the power transformer primary circuit, providing basic inrush current protection. However, I felt that the time constant of approximately 157 ms, or a sixth of a second, was relatively short considering the cold cathodes of the 3CX800As and, in particular, the current to charge 50 μF of filter capacitance. I incorporated a circuit change—the addition of a 1000-μF capacitor in parallel with C5 **(Figure 3)**—for greater inrush current protection. This should enhance the life of the related components. After the change, the time constant increased to one third of a second, or about 240 cycles. Bill Orr, W6SAI, has written an excellent, comprehensive article on the subject of inrush protection parameters.[1] All amplifier users should follow the universal principles he outlines.

Cathode inrush and voltage protection

The Eimac specification for the cathode of a 3CX800A is 13.5 volts ±0.6 volt at 1.5 amperes. Eimac has established that excess heater voltage will *increase* emission, at the considerable expense of shortening tube life. Marginally less than nominal voltage will enhance it without the loss of emission.

Because of this relationship, it's prudent to *accurately* measure the heater voltage at or near the tube socket, under load. Since each of the 3CX800s requires only 1.50 amps of current, you may measure the voltage at the power connector on the amplifier rear panel with reasonable accuracy. The amplifier must be under load and at a nominal line voltage, using an accurate meter, preferably a true root-mean-square type.

Figure 2. Cooling protection.
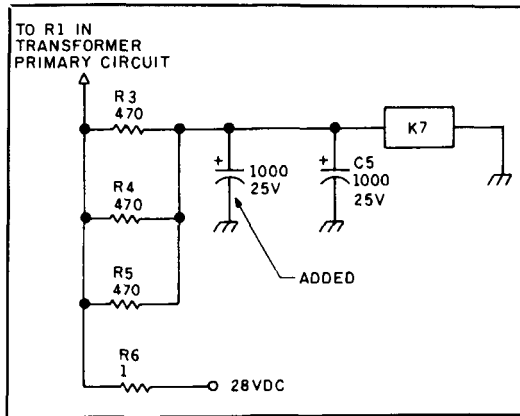
Figure 3. High-voltage step-start modification.

It's best to take several test measurements, at different times, with these two parameters in mind. **Figure 4** shows the test point on the power connector on the rear panel of the amplifier. The Titan's nominal heater voltage, from what I have been able to determine, tends to be very near normal under load—*if* the line voltage is close to 240 volts under load.

One further consideration regarding over-voltage on the cathode is that any heat dissipation above the nominal is transmitted to the grid structure, increasing its existing, relatively low level, 4-watt dissipation. So, play it safe. If you determine that the voltage is on the high side, or above Eimac's specification, take appropriate steps to correct the situation.

Voltage-protection circuit

Small series resistance provides one easy way to accomplish a voltage reduction using a short length of nichrome wire to yield a few tenths of an ohm. For convenience, it may be wound using a large-value 2-watt resistor as a form. Nichrome is difficult to solder so use a clean, hot iron and good flux on *very clean* wire surfaces.

In addition to establishing the nominal cathode voltage through a dropping resistor, a resistor provides some minor reduction of
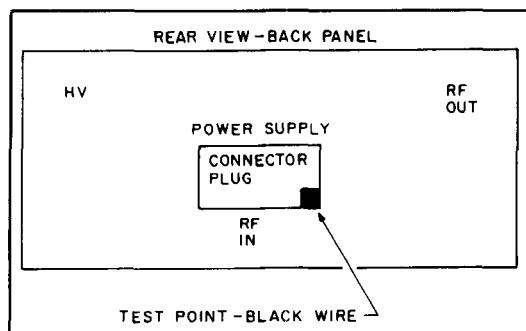
Figure 4. Cathode voltage test.

the thermal shock to the tubes. This resistor may be placed in the power supply or in the amplifier. If the cathode voltage, under load, falls below Eimac's specification, it would obviously dictate the use of an auto-transformer. Or, after checking your primary power wiring, you might consult with your power company to determine whether the wiring conforms to their specifications.

Alternative cathode inrush protection

An alternative method of inrush protection is provided by a Keystone Carbon Company's "inrush current limiter." This excellent, inexpensive negative-temperature coefficient thermistor can serve a dual purpose of lowering the cathode voltage while providing a substantial, *very slow* step-start capability. These devices are typically used in power-supply primaries or secondaries, and are extremely effective. You may wish to consider Keystone's type CL-40, which provides about 0.27 ohms at 3 amps. You can also try their CL-50. It offers about 0.32 ohms at a 3.0 amps of current.

These devices provide a thermal time constant of 100 and 120 seconds, respectively—well within the minimum 3-minute warm-up period. Keystone's inrush current limiters are available from a number of distributors.* They are about the size of a nickel and dissipate nominally, in this application, about 3 watts. Operating temperature is approximately 140 degrees C. Because of this, you should mount the current limiter in an appropriate location to minimize heat radiation to other components. The Keystone Company states that the life testing of the CL-40 devices, based upon 3000 cycles of 5 minutes on and 5 minutes off, resulted in a resistance change, at 25 degrees C, of within 5 percent of the starting resistance. Obviously, if given the choice, I'd use the inrush current limiter, because it does it all!

Caution

A word of caution is in order here to remind anyone attempting any internal examination of this equipment that **high voltage kills!** Remember to disconnect the high-voltage power cable from the amplifier *before* beginning any work and activating the power switch. Once you activate the main power switch (to measure the cathode voltage internally under load, for example), high voltage is present at all times! The high-voltage shorting bars and microswitch inter-
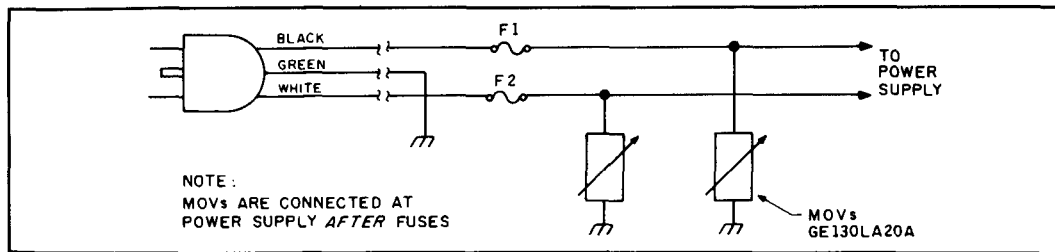
Figure 5. Power-line protection.

locks are there for your protection and should not be defeated. It's also important to note that the high-voltage bleeder current is quite small, and takes a relatively long time to fully discharge the 50-μF of filter capacitance.

Cathode warm-up delay

Eimac recommends a minimum 3-minute warm up to allow the 3CX800A cathode to come up to temperature. Note that this specification is *strictly a minimum*. A conservative design would require more than 3 minutes to be optimal, if you've got the patience to wait for the cathode to fully come up to even temperature before operating. This is desirable for longest tube life, so it pays to plan ahead. Otherwise, hot spots will develop along the cathode surface, degrading it, and eventually reducing the life of the tube. Therefore, a warm-up period closer to 5-minutes would be optimum.

In my unit, the 3-minute timing circuit, consisting of R10 and C8 on the relay control board, was short 15 seconds for the closure of relay K5, which is well within the component tolerances. However, being on the conservative side, I elected to allow for greater warm-up time so I paralleled C8 with a 5.6-μF, 25-volt capacitor, which added 30 seconds to the delay. Regardless of the timing circuit, I always allow at least four minutes, if possible.

A prudent corollary to the warm-up consideration is the often-neglected cooling-down period. Out of conservatism, I allow several minutes for cool down, depending upon the outlet air temperature, because an internal thermal lag occurs after operation. With the exhaust temperature sensor in operation, this is relatively easy to achieve.

RF drive control

The Titan has an overdrive protection fold-back biasing circuit and red light-emitting diode (LED) warning light. A high level ALC circuit is also available for drive power control. The 3CX800As only require a *maximum* of 65 watts peak-envelope power (PEP) RF drive to produce 1500 watts

of clean PEP output. Generally, about 50 watts are all that are required, depending upon the frequency and impedance matching.

I strongly recommend the use of the amplifier's ALC loop to preclude overdrive, which only produces waveform flat-topping. Above the 65-watt level, the red overdrive LED flashes brightly, all the meters dance up beautifully, and the output waveform is limited—creating distortion and the resulting excessive bandwidth. I have verified this on my spectrum analyzer. Thus, a high-level RF clipper is created without benefit of subsequent filtering.

I also use a small black paper mask, which I cemented in place with RTV, vertically along the left side (top front view) of the red overdrive LED warning indicator. This eliminates the side lighting to the LED from the panel lights, which are next to the diode, resulting in improved contrast and visibility.

Grid-current interlock

It's been suggested that an overdrive grid current interlock circuit be incorporated. This circuit could be used in conjunction with the existing overdrive sensing circuit to trigger a selenium-controlled rectifier to operate a relay, which would lock out the amplifier by placing it into its standby mode until reset after the proper correction. I think this is a superior way to go but I didn't care to get into the circuit boards, so I rely on the ALC loop, which does an excellent job for me. I have full power from the exciter in the stand-by mode with its normal self-generated ALC, and the same with the amplifier in operation—except that the exciter is then shut down at approximately 50-watts of PEP of drive when the high-level ALC control is properly set . . . no muss, no fuss!

Currently, with most exciters capable of 100 watts PEP output or more, it's a comfort knowing that you're essentially automatically protected, both from the standpoint of the relatively fragile 4-watt grid dissipation of the tubes, and from excessive distortion products. Obviously, the ALC loop is an ideal protective circuit which should be used universally.

Power-line protection

Lightning strikes or other anomalies, both direct and indirect, can cause great spikes and/or wide fluctuations in power-line voltage. This, in turn, can cause costly direct damage and/or latent failures due to the earlier stress on components. To help preclude this, it's wise to place a metallic oxide varistor(s) (MOV) across the power-line input to the power supply. I used a pair of GE type 130LA20A MOVs, one from each leg of the 220 to neutral, as shown in **Figure 5**. These will clip hard at 130 volts AC on either side of the 240-volt line. If your line voltage tends to be generally above that value, which is bad for all equipment, you must select the next higher voltage limit. If this is the case, all the more reason to protect the heaters from over-voltage; or, better yet, call your local power company to have them correct it! I should point out that MOVs lose some of their capability as a result of the number and amplitude of the surges they dissipate. So, if your area is subject to frequent power surges, you should replace these devices periodically. The use of MOVs is an extremely modest investment to protect a substantial one.

High-voltage power supply mods

I felt that the peak reverse voltage (PRV) of the high-voltage diode bridge should be increased for greater reliability. Consequently, I placed an additional Motorola MR510 (any other 1000-volt, 3-amp diode will do) in each leg of the bridge by soldering it between two of the existing diodes along with a 680-k resistor for equalization. However, I would use 2-watt resistors instead of the existing ones. If you wish to go one step further during the same operation, I recommended that you shunt a 1,000-volt 0.01-μF disc capacitor across each diode for the conventional capacitive-voltage equalization and spike protection. TEN-TEC has now changed from 3 to 6-amp diodes, and no failures have been reported.

Summary

These modifications may be performed with minimum effort and expense, and without the need for any cutting or drilling. The results are well worth your time, and will provide additional peace of mind and protection for a rather large investment. I feel that TEN-TEC did a good job in offering very reasonable protection for their amplifier. However, there's nothing like buying a little bit of extra insurance, at minimal cost and effort, which you hope will never pay off. Should you ever need it, you'll be glad you did! ∎

REFERENCES
1. Bill Orr, W6SAI, "Radio Fundamentals: Long Life for Your Amplifier Tubes," CQ, January 1991, page 68.

PRODUCT INFORMATION

Microwave Pulse Power Transistor Line Expanded

Motorola has added to its line of microwave pulse power transistors with the introduction of four new devices, expanding the available output power to 500 watts. The MRF10070, MRF10150, MRF10350, and MRF10500 microwave power transistors are intended for operation in the 1025 to 1150-MHz frequency band in applications such as TCAS, TACAN, and mode-S transmitters.

Each transistor operates from a power supply of 50 volts and has a typical amplifier gain ranging from 9 to 10 dB. All four parts are supplied in industry-standard hermetic metal-ceramic packages and are internally matched for broadband operation.

For more information on these new tran-

sistors, contact Motorola Literature Distribution, P.O. Box 20912, Phoenix, Arizona 85036. Ask for data sheets MRF10070/D, MRF10150/D, MRF10500/D, and MRF10-350/D.
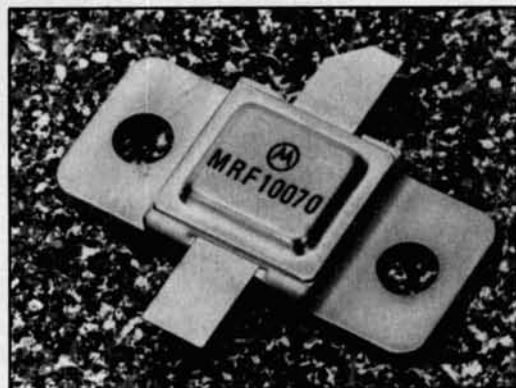
By Henry Elwell, *N4UH*
2701 Rary Road
Cleveland, North Carolina 27013

THE 160-METER SEMI VERTICAL: A CASE HISTORY

A look at a little-known DX antenna

H ave you ever come across an antenna that was so simple, yet so successful, that you wanted to tell everybody about it? Since 1936, when I was first licensed as W2JKH, I've tried many different antennas. I began with a simple Zepp antenna, and after many years, moved on to a complicated controlled vertical radiation rhombic,[1,2] among other radiators, now in use. My most recent antenna, however, is so simple, so easy to erect, and so effective, I must tell you about it! But I'm getting ahead of the story of why and how this antenna came about.

I wanted to obtain DXCC status on 160 meters. To accomplish this goal, I shunt fed an 80-foot tower supporting one leg of the rhombic and arranged 36 quarter-wave radials around the base of the tower. After two winters of work, I had fulfilled my ambition. Unfortunately, the subsequent addition of a 2-meter antenna on top of the tower destroyed its 160-meter radiation pattern. Although I could have done certain things to correct this problem, my inquisitive amateur spirit urged me to try something new.

The new antenna had to meet the following criteria.

• Vertical polarization of the radiated signal.
• Radials not required.
• Physically easy to feed.
• Supported by an 80-foot tree.
• Operation equal to or exceeding that of the 80-foot tower.

For your own purposes, the fourth requirement is flexible, although I recommend a minimum tree height of at least 35 feet.

MININEC computer analyses

Using my MININEC computer program for study of thin-wire antennas, I began building different configurations in the computer, to see what would suit my situation. For comparison, I started out with a half-wave dipole, 80 feet high, using the program's built-in ground environment (Virginia clay soil), which is similar to that at my location in North Carolina. I used this environment for all configurations. The horizontal antenna produced the expected pattern with maximum radiation straight up. While it didn't have a good figure-eight pattern in the horizontal plane, I knew it would be great for domestic contacts (see **Figure 1**). For all analyses, I chose a frequency of 1.83 MHz, and used 20 segments for each wire length. When using MINI-NEC on the more complex antennas, be sure that the number of segments you choose is sufficient to do the job. Although it can take over three hours to run the greater segmented configurations, fewer segments per wire length can lead to errors. My computer is a Swan AT12 running at 12 MHz.

A quarter-wave vertical

Next, I tried a quarter-wave vertical 127.5 feet tall. The plot consisted of a circular horizontal pattern, a vertical pattern with maximum radiation at 22 degrees, and an

160-METER DIPOLE — 80 FEET HIGH

127.5' 127.5'

80'

Z = 39.2 - j30.6
90° 5.99 dBd

ELEVATION ANGLE 5°
FORWARD GAIN -12.78 dBd

1.830 MHz

160-METER DIPOLE — 80 FEET HIGH

AZIMUTH ANGLE 180°
MAXIMUM GAIN 6.00 dBd AT 90°

1.830 MHz
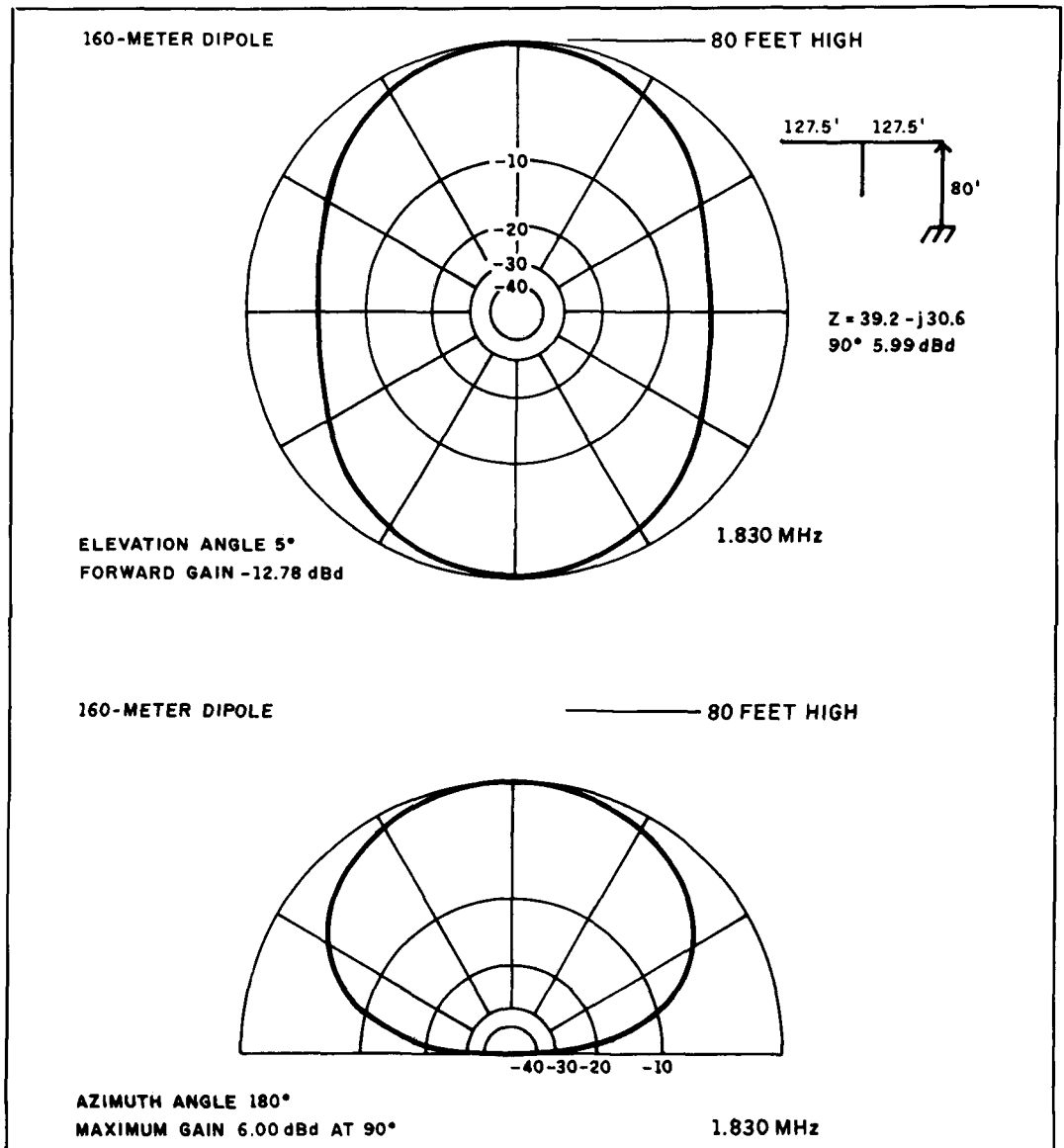
Figure 1. Radiation pattern of a 160-meter dipole 80 feet high, using MININEC computer program.

excellent null at the 90-degree radiation point (**Figure 2**). I performed the same study using an 80-foot vertical with similar radiation patterns. The input impedance was down to about 10 ohms, whereas the quarter-wave antenna was about 28 ohms. These verticals would require an extensive ground system for efficient operation.

Quarter-wave inverted L

I've heard several 160-meter stations using quarter-wave inverted L antennas with radial systems which achieve very good results. Going through the task of laying radials again held little interest for me, but I was curious about what could be expected from them in the vertical radiation plane. I simulated a normal quarter-wave L, then an

inverted L with surprising results. **Figures 3** and **4** show the radiation patterns. Maximum gain at 26 degrees for the normal L was greater than that of the inverted L, which had maximum gain at 24 degrees. However, the vertical radiation pattern was different, showing a front-to-back ratio for the normal L—hence its gain margin. The input resistance of the normal L was about 4 ohms versus 16 ohms for the inverted L.

Half-wave inverted L

While using that particular configuration, I decided to see how the half-wave inverted L pattern looked. This antenna was commercially popular many years ago. Having the current at the topmost part of the straight vertical portion results in greater ra-

160-METER QUARTER-WAVE
VERTICAL

——————— GROUND MOUNTED

-10
-20
-30
-40

Z = 26.7 -j5828
22° -0.73 dBd

ELEVATION ANGLE 5°
FORWARD GAIN -5.34 dBd

1.830 MHz

160-METER QUARTER-WAVE
VERTICAL

——————— GROUND MOUNTED

-40 -30 -20   -10

AZIMUTH ANGLE 0°
MAXIMUM GAIN -0.73 dBd AT 22°

1.830 MHz
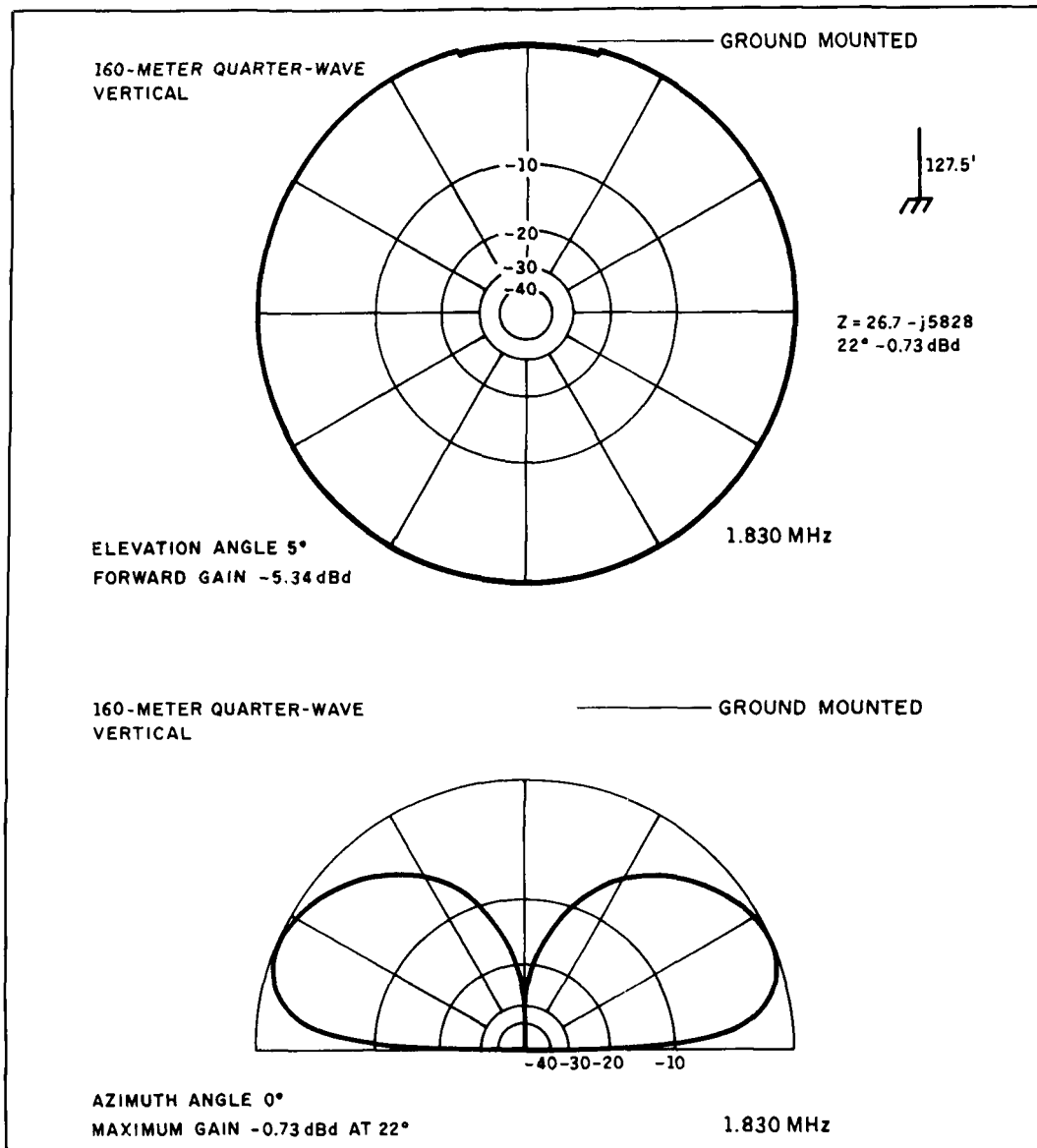
Figure 2.Quarter-wave vertical, 160-meter simulated radiation pattern. Maximum vertical radiation is at 22 degrees.

diation efficiency. Still, considerable energy leaves the antenna at relatively high radiation angles due to the current flow in the horizontal section; the flat top is only a quarter wavelength above ground. **Figure 5** shows the considerable radiation at the 90-degree point.

This article isn't written from the technical standpoint, so don't expect a theoretical discussion. It will give you a good idea of what you can do with some wire and a couple of trees, even on a small property. As I've mentioned, I wasn't interested in putting down more radials under a new antenna. Thus, my enthusiasm turned to a half-wave antenna version which I felt would be less dependent on ground-return currents. The ground currents cause power losses, which reduce the radiation efficiency. I

chose a length of wire 255-feet long to be resonant in the 1.83 MHz range.

If you look at the distribution of current on a half-wave antenna **(Figure 6)**, you'll see that the current maximum is in the center of its length. With that in mind, I reasoned that I should put the center of the half wave in the vertical plane. By using an 80-foot tree, and allowing the bottom end of the vertical portion to be about five feet off the ground, I had a 75-foot section for vertical use. This meant I had 180 feet of wire left to play around with in the horizontal plane. My first simulation was a Z shape: 90 feet extended five feet above the ground, 75 feet vertically, and 90 feet horizontally at the 80-foot level—all in the same plane. (Unless stated otherwise, the excitation source to the antenna is always at its
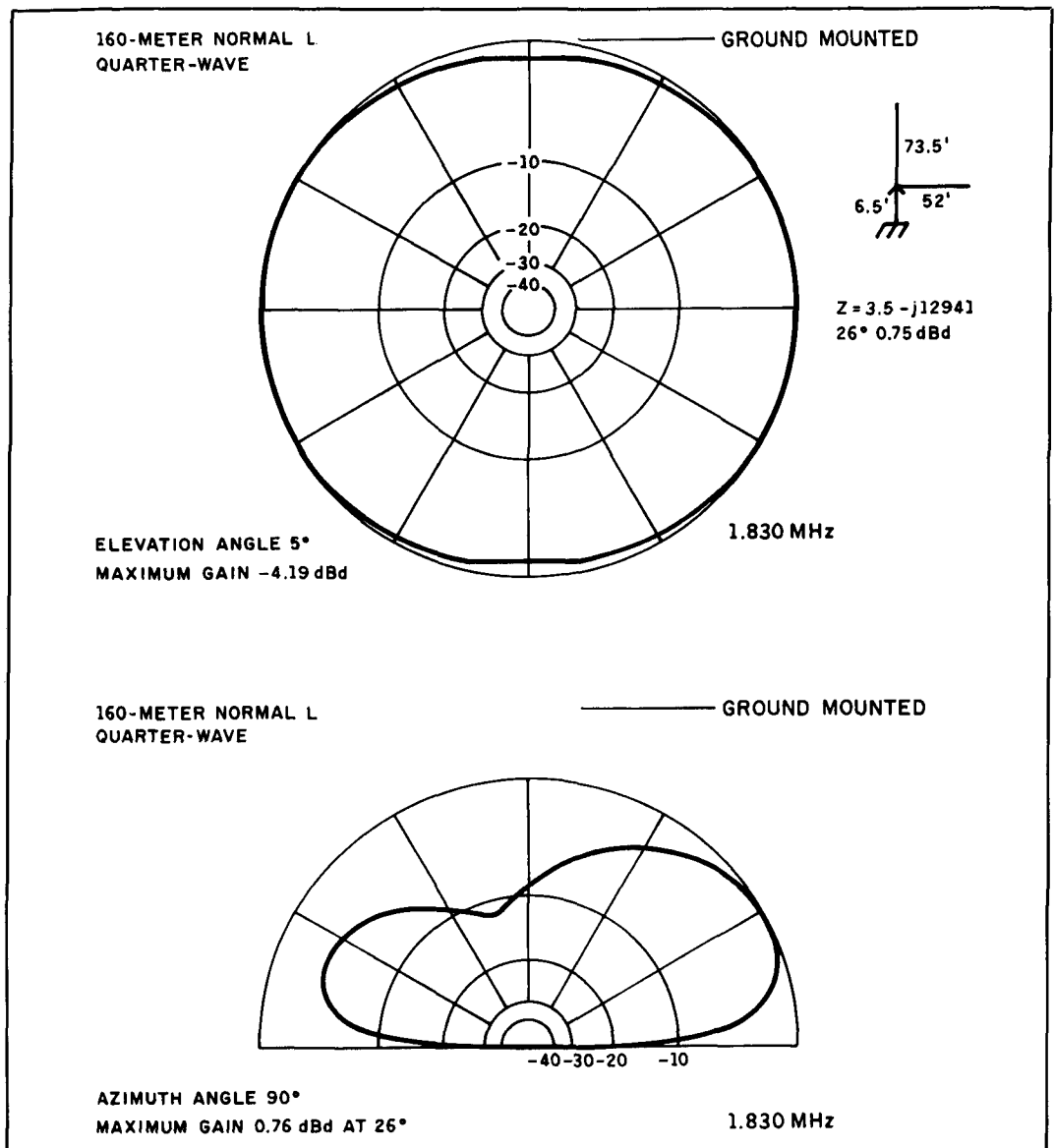
Figure 3. Radiation patterns for a simulated normal L quarter-wave 160-meter antenna.

end.) The Z antenna showed a maximum radiation at 32 degrees with a front-to-back ratio of 3 dB (see **Figure 7**). It had a lopsided vertical pattern, and I was looking for a symmetrical pattern in the vertical plane.

Lazy U

Then I thought, "Why not try to cancel the horizontal radiation and see what can be done to improve the vertical pattern?." The quickest way to do that, I reasoned, was to reverse the direction of the lower horizontal portion to place it under the upper section. The antenna would look like the letter U laying on its side—hence the name the Lazy U antenna. K2FE suggested that the current distributions would be as shown in **Figure 8A**. Currents A and C would cancel, leaving

the vertical radiator of **Figure 8B**. The configuration resulted in a computer-derived symmetrical pattern, with a vertical radiation about 27 degrees above the horizon, but with a 90-degree null about 6 dB down (see **Figure 9**).

Erecting the Lazy U

The results looked exciting! I wanted to put up the antenna and see what it would do. I used a sling-shot to run a light line over a tree from a fishing reel/rod combination. I used that line to pull over a heavier cord, followed by a rope. With the strong rope, I was able to haul the antenna wire into position. Because I had to pull the wire through trees, I chose an insulated, stranded length of no. 12 wire on the recommen-

INVERTED L QUARTER-WAVE
160-METER EXTRA LENGTH
ON TOP

79.5 FEET HIGH

52'

73'

6.5'

-10
-20
-30
-40

Z = 15.7 - j9623
24° - 0.72 dBd

ELEVATION ANGLE 5°
FORWARD GAIN -5.52 dBd

1.830 MHz

INVERTED L QUARTER-WAVE
160-METER EXTRA LENGTH ON TOP

79.5 FEET HIGH

-40 -30 -20   -10

AZIMUTH ANGLE 0°
MAXIMUM GAIN -0.73 dBd AT 24°

1.830 MHz
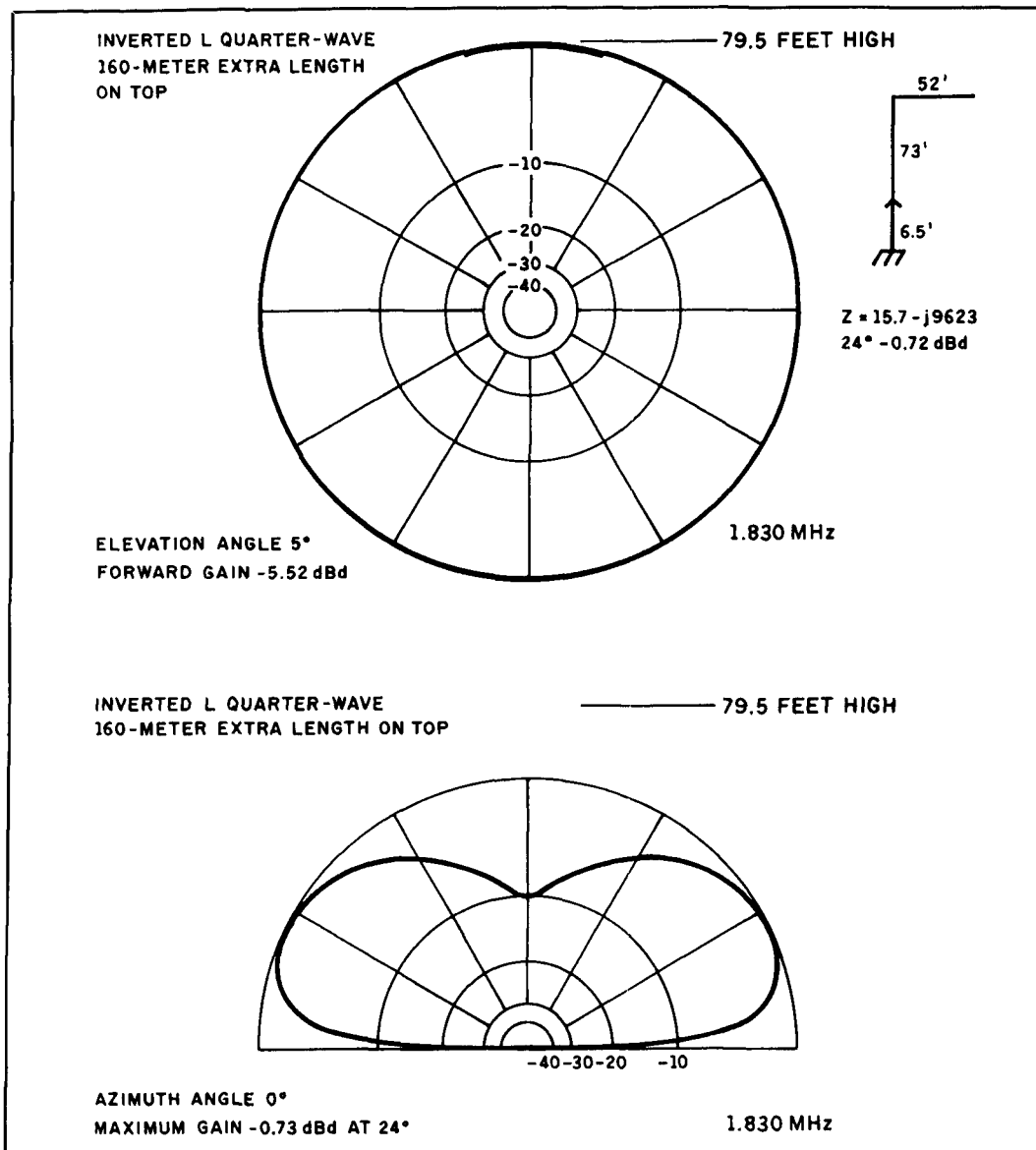
Figure 4. Radiation patterns for a simulated inverted L quarter-wave, 160-meter antenna.

dation of NA2M. You can purchase the wire (called TRHN) at any electrical supply house. It comes with any color insulation; I chose green. The wire has a very tough covering and is available in either stranded or solid form. I paid $20.80 for 260 feet.

I used an insulator at the top of the 75-foot vertical section (see **Figure 10**). This obstructs the movement of the wire as the insulator hits the top branch. When it hits, you know the vertical section is in the correct position. At W2LL's suggestion, I used a Millen solid-state grid dipper to find antenna resonance by winding 3 to 4 turns of the antenna around the dipper coil. I then lengthened and/or shortened the lower horizontal section to produce resonance at 1.83 MHz.

All my models were excited at the ground

end of the antenna for convenience in running the program. Because it would be difficult to feed the antenna at the center in real life, W2KXD advised trying a Pi network between the 50-ohm coax cable and the high-impedance end of the antenna (see **Figure 11**). I adjusted the network by placing a VSWR bridge at its input and tuning the network for a null.

Lazy-U antenna results

My first couple of nights on the air, I used 100 watts to work stations up the eastern coast to Canada, and obtained as good or better reports than I was giving. I then began using my 700-watt linear, reaching Utah and the West Coast plus an English station on CW. I knocked off Germany,

Spain, and a second English station on another night with good signal strength. Using 100 watts, I worked a couple more Europeans during the 1990 ARRL DX CW contest. Those contacts assured me the antenna was doing as well as the original vertical. I was finally convinced that I had a winner during the CQWW 160-meter SSB contest, when I worked everything I called while running 100 watts output. All this was possible without the labor of installing ground radials.

In the fall of 1990, grazing cattle took over the pasture supporting the lower 90 feet of the Lazy U. I reversed that section 90 degrees onto "safe" property, and the antenna became a Zee shape. It's pattern is shown in **Figure 7**. Operation was not noticeably affected. In the 1990 CQWW CW

contest, I worked stations in an area encircling North Carolina (V73-HC8-PY-D4-G-VE7-KH6)—a total of 30 countries. The secret to my success lay in mounting the vertical section of the antenna in the center of the total antenna length.

Improving the Lazy U 90-degree null

I wanted to improve the null, and canceling the horizontal radiation component seemed the way to go. Thus I modeled a shape where the lower 90-foot section was folded back upon itself to cancel its radiation (see **Figure 12**). There was negligible difference in the radiation patterns.

Since this tactic accomplished little, I
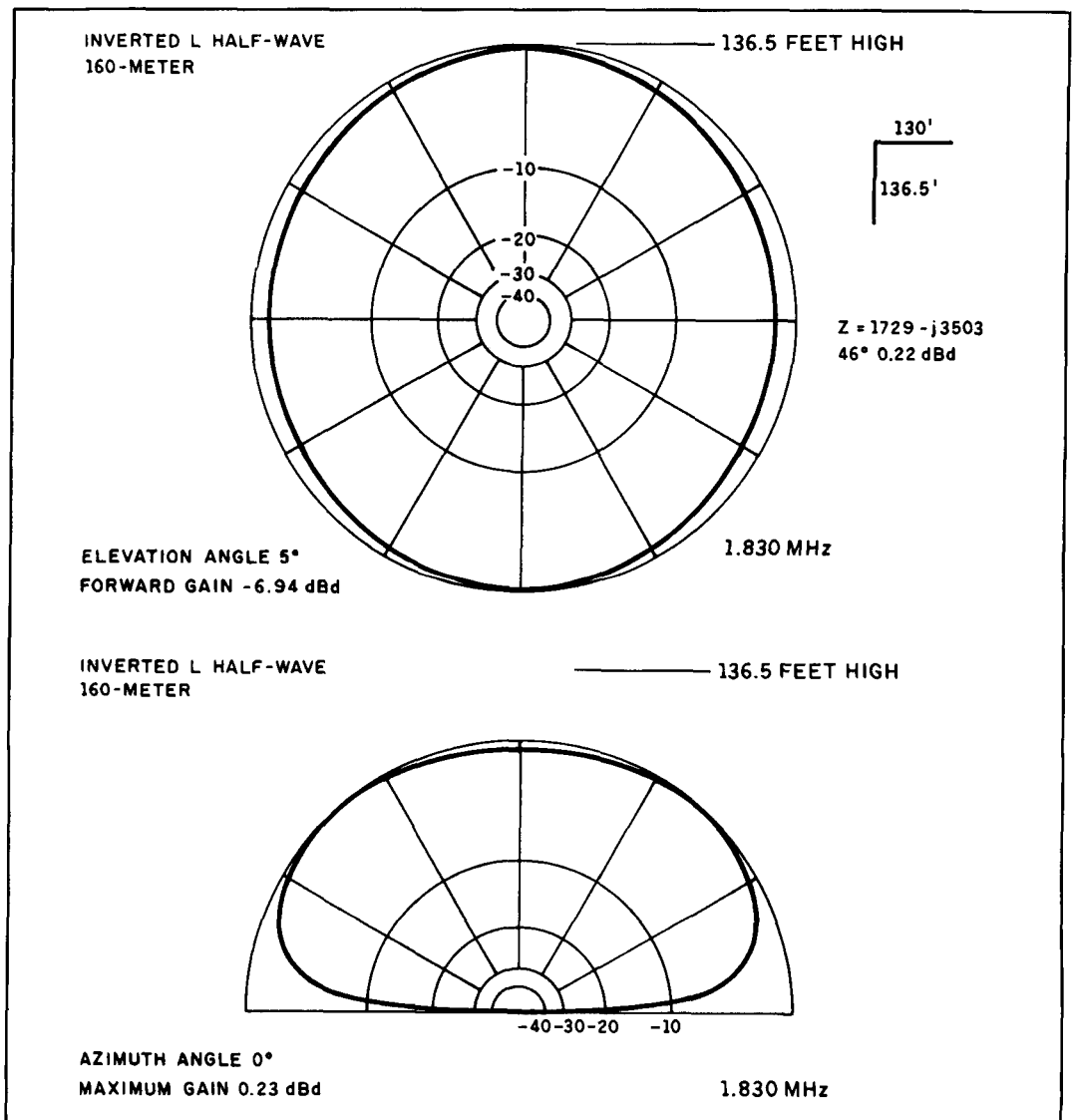
Figure 5. Simulated radiation pattern for a 160-meter inverted L antenna, one-half wavelength long.
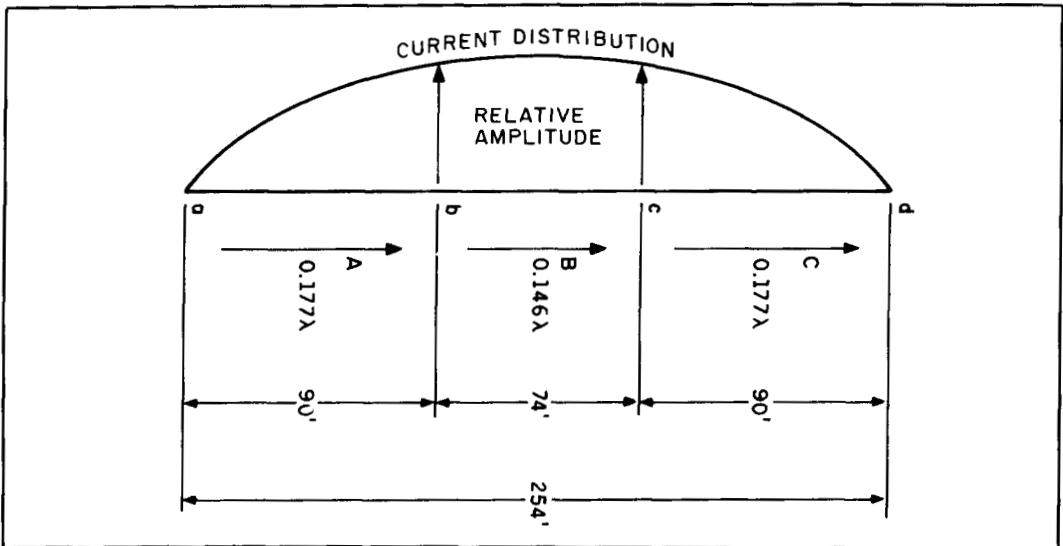
Figure 6. Current distribution and assumed current flow in a half-wave dipole. Maximum current is at antenna center. Dimensions shown are for 160 meters.
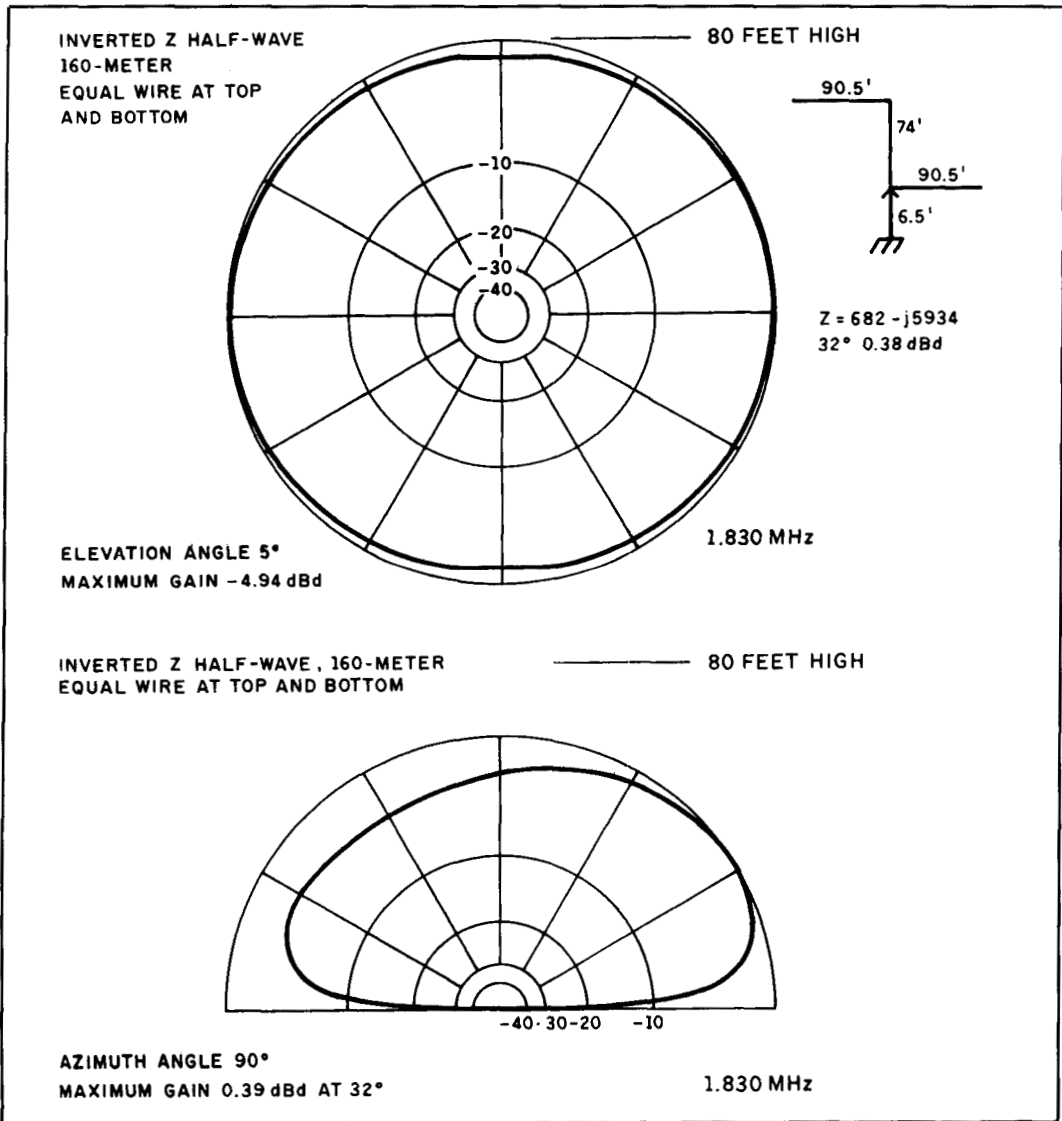
Figure 7. Simulated radiation pattern for 160-meter Z-shape antenna with equal length wire at top and bottom.
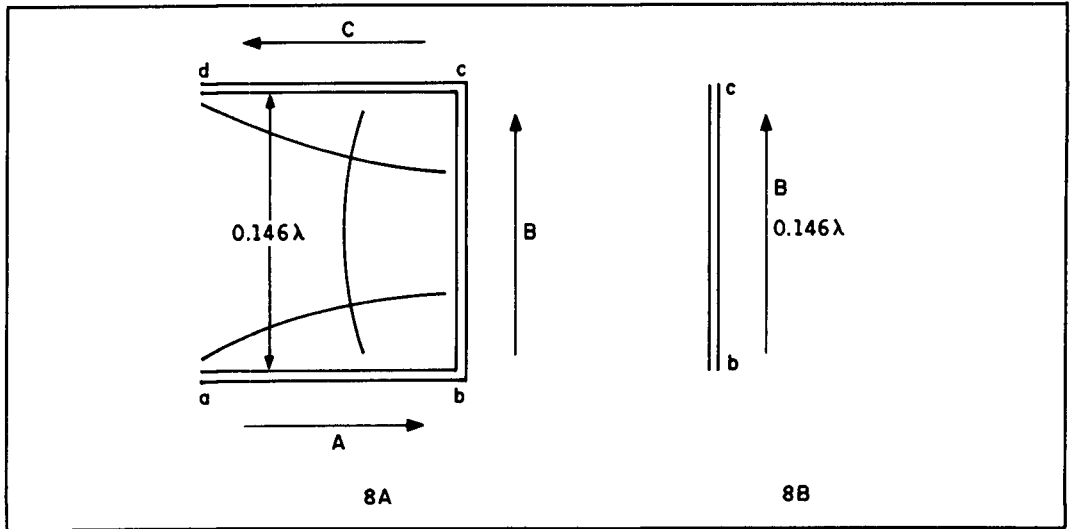
Figure 8. Current distribution in Lazy U showing cancellation of current flow in sections A and C of *Figure 8A*, resulting in the vertical radiator of *Figure 8B*.
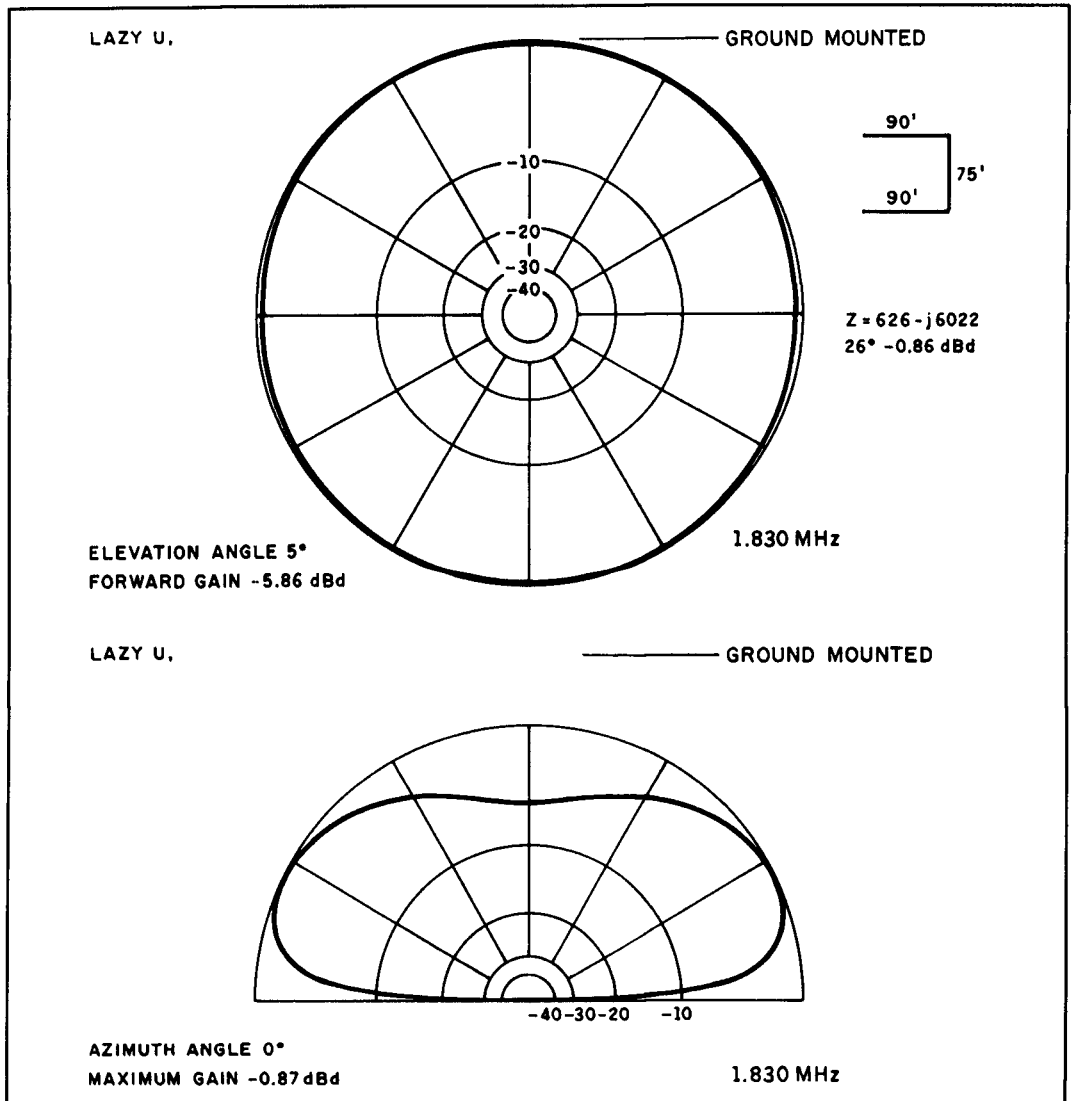
Figure 9. Computer-derived radiation pattern of the 160-meter Lazy U configuration. Wire lengths are 90, 75, 90 feet on sides.
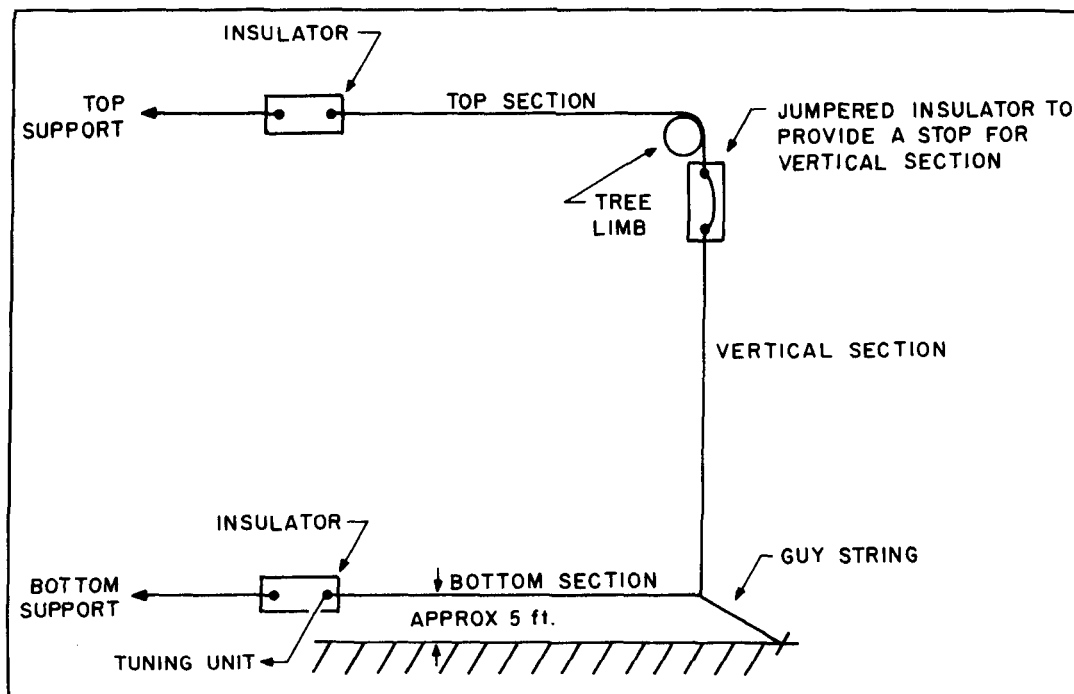
Figure 10. Lazy-U configuration. Total length of antenna is 254 feet for 1.83 MHz. Use highest tree available to maximize length of vertical section.

decided to straighten out the lower section and put the folded section at the 80-foot level. That configuration resulted in a vertical radiation pattern which looked like Mickey Mouse's ears, with a null of almost 13 dB at the 90-degree pattern point. It had a radiation angle of about 27.5 degrees (Figure 13). I liked the results, so I pulled down the physical antenna and restructured it. Be careful when assembling the folded-back section; it could easily loop around itself.

I suggest you use the construction shown in **Figure 14.** Insulator number 1 is used as the stop which lets you know that the vertical section is at its limiting position. The small weight in the center adds stability to the section when it's pulled up and keeps it from twisting about during windy weather.

While I can't offer an explanation as to the cause, the antenna resonance point became 1.94 MHz after I folded back the upper section. I needed to add approximately 15 feet of wire to the lower horizontal section to resonate the antenna at 1.83 MHz. The Pi network produced an input null on the VSWR meter with little difference in its tuning. On-the-air operation seems to be the same as the original Lazy-U configuration.

It's very difficult to operationally determine the value of the additional 10 dB of null improvement at the 90-degree radiation angle. The improvement in null suggests more power is going into the lower-angle lobes. Thus there's some question in my
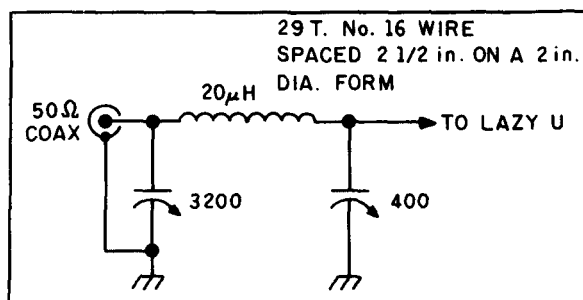
Figure 11. Pi network to match 50-ohm coax to bottom section end of Lazy U. This is a high-voltage point for the output capacitor; use spacing consistent with output power.

mind as to the benefit of complicating the construction of the basic Lazy U. Actually, the folding back of the upper section doesn't produce complete radiation cancellation because the currents in the two wires aren't equal. In fact, the current at the end of the open section is zero.

Testing the Lazy U on 80 meters

I explored other configurations with different feed points, but I won't discuss them here. However, I was interested in determining whether there is an improvement in the vertical pattern when the vertical section is increased in length. Because of the limitation of the 80-foot tree height, I found it
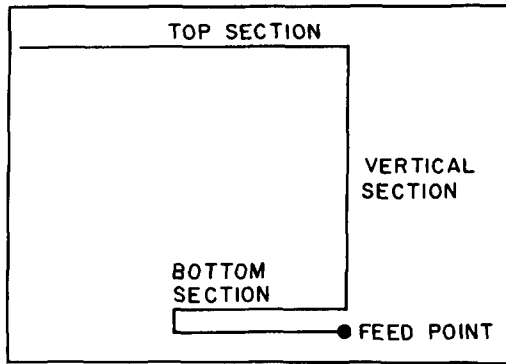
Figure 12. Modified Lazy U to minimize horizontal radiation.

necessary to move to a higher frequency to make such a test. Thus, I made a study of the Lazy U on 80 meters using a total length of 113 feet. With the 75-foot vertical limitation, that meant the upper and lower sec-

tions would be 19 feet long. As **Figure 15** illustrates, an almost 26-dB null occurs at the 90-degree radiation point. As the vertical length is increased, the 90-degree null becomes deeper. Just for fun, I made a run using an antenna length greater than a half wave to see how the pattern was affected. The 80-foot vertical restriction still applied. Because of the current reversal beyond the half-wave point, an unbalanced horizontal radiation occurs and the 90-degree null suffers.

Operational check on 80 meters

Once the initial Lazy U shape is up, it's fairly easy to change things. With that in mind, I took down the 160-meter antenna, and put up the basic Lazy U for a frequency
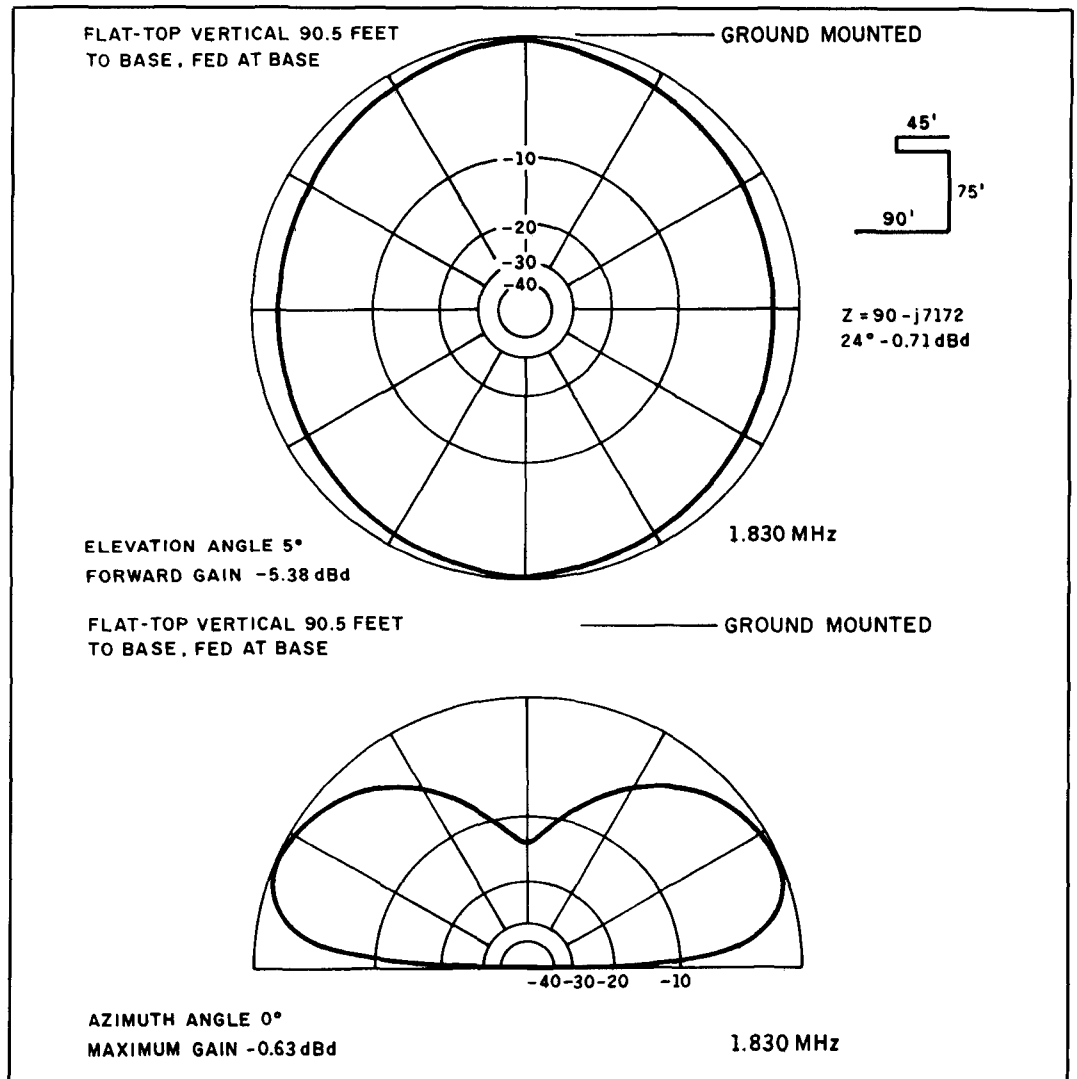
Figure 13. Simulated radiation pattern for flat-top vertical for 160 meters. Length is 90.5 feet to base. Antenna is base fed.
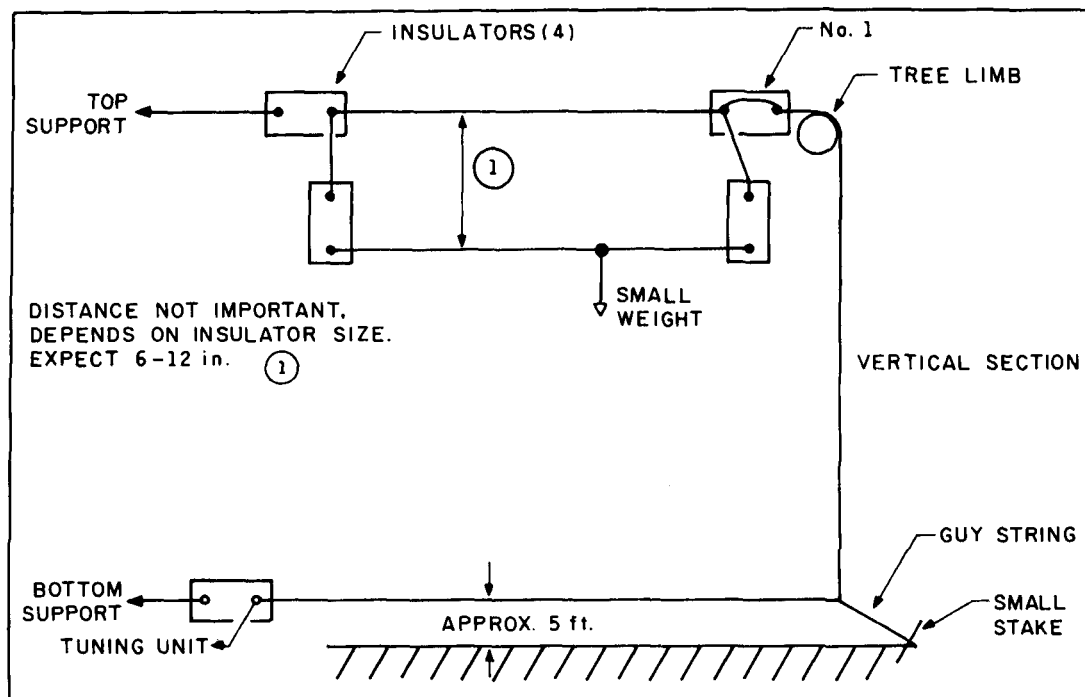
Figure 14. Construction of Lazy U with top section folded for horizontal radiation suppression.

of 3800 kHz. That frequency required about 123 feet of wire. This meant that 75 feet would be vertical, with top and bottom sections of 24 feet each. Grid dipping the antenna showed resonance was too low, and a value of 119 feet was the final length for 3.8 MHz. Only the lower section had to be cut. To get the proper top section length, I had to pull the antenna down 2 feet, and cut off 4 feet from the bottom end. Four feet is the difference between 123 and 119 feet, producing an equal shortening of 2 feet for upper and lower sections.

The Pi network required a new inductance value, which I procured by shorting out turns. I used the VSWR bridge at the Pi network to perform the matching from the 50-ohm coax cable to the antenna end, and obtained a reading of 1:1.

I performed tests comparing the Lazy U with a W2LL quarter-wave sloper. The quarter-wave sloper is a 60-foot tower topped with a TH7 tribander. The tower has eight uninsulated guy wires for support. The coax feed is through a 1:1 balun, whose output is connected at the 60-foot level. The effort required to model that configuration for the MININEC program was more than I wished to do. Experience has shown that the W2LL antenna has considerable high-angle radiation.

The first operation at 100-watts output occurred just before sundown. A station in east Tennessee, 200 miles away, said the W2LL antenna was S9 + 10 dB, while the Lazy U was S6 to S9. Simultaneously, a station in full sunlight in Houston, Texas, 966 miles away, said the Lazy U was stronger, but storm static was too great to give an S-meter reading.

Three hours later I contacted the east-Tennessee station again, and was told the W2LL antenna was 15 dB stronger than the Lazy U. However, stations in Mississippi, Louisiana, and Kansas (all 500 to 825 miles away) said the Lazy U was stronger. When on receive during the same evening, I found a station in Mexico was copiable through heavy static on the Lazy U, not copiable on the W2LL antenna, and readily copiable on the rhombic pointed in his direction. This was at a distance of approximately 1500 miles. The results show the Lazy U is putting out a low radiation-angle signal, expected of a vertical with reasonable efficiency. The gain of the rhombic is estimated to be about 5 to 6 dB over a dipole at this frequency, although its vertical radiation angle is about 45 degrees compared to the 26 degrees for the Lazy U.

The results, although of qualitative value, gave me confidence that the operational results agreed with what the computer analyses predicted. From my many years of ham operating, I can say that I'm elated with the performance of the Lazy U antenna. So if you have only a 30-plus foot tree on your property for operation on 160 meters, your Lazy U antenna will give you much greater low-angle radiation (see **Figure 16**)
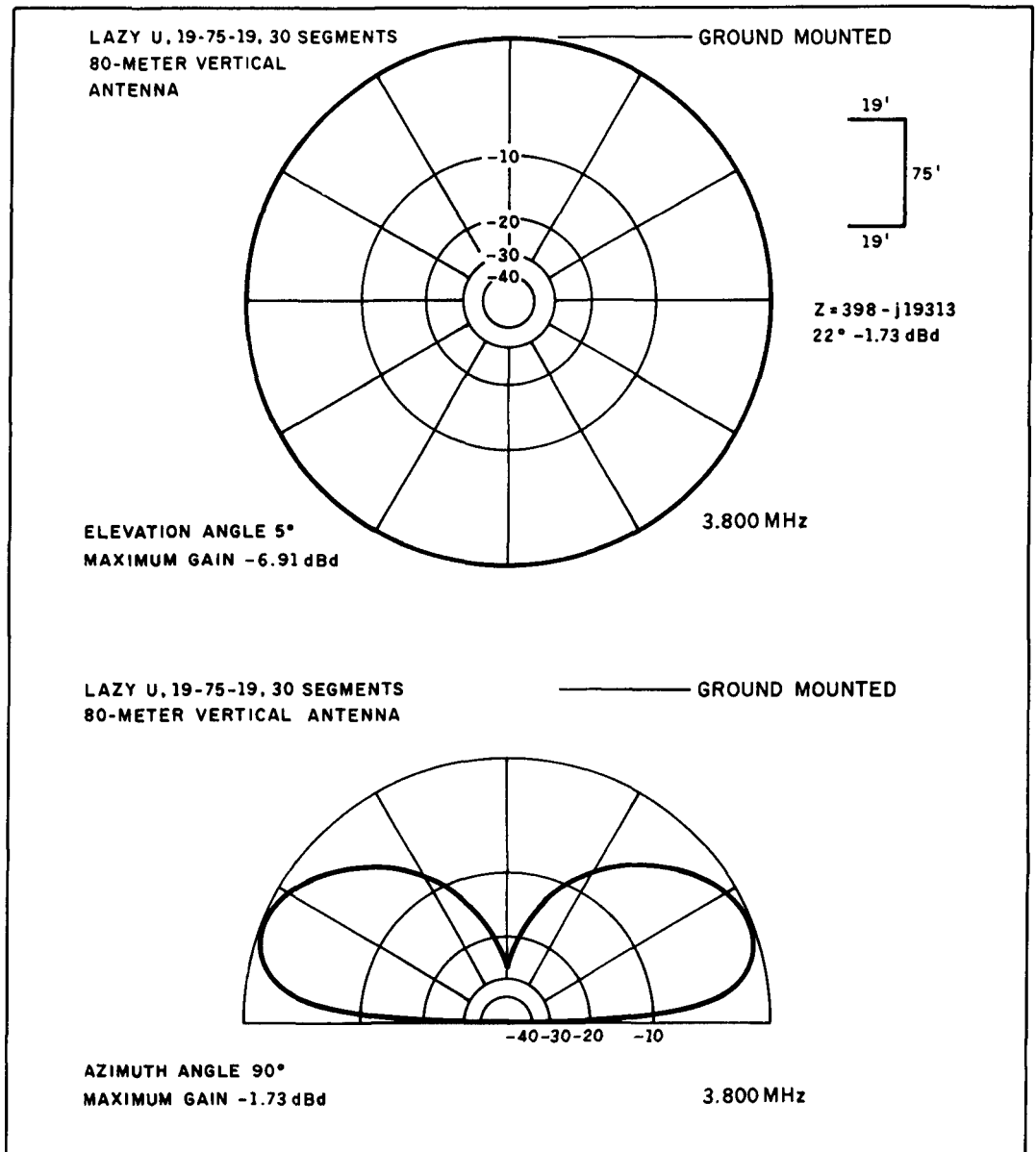
Figure 15. Radiation patterns for the Lazy U antenna on 80 meters.

than a half-wave horizontal at 80 feet (see **Figure 1**).

Past studies

Have you ever thought you've discovered something new, only to find out that your idea is old? Having finished my work and manuscript, I mentioned my computer studies to Bob Eldridge, VE7BS, a long-established 160-meter DXer, with whom I correspond. I asked him what he thought of the idea of the antenna that I called the Lazy U; the name is similar to that of the old antenna called the Lazy H. Much to my surprise and dismay, Bob sent me a short article he had written for the *West Coast 160 Bulletin*,

briefly describing an antenna developed from the all-band G8ON antenna. The original article on the G8ON antenna appeared in the September 1957 issue of the *RSGB Bulletin*. Bob wrote a manuscript for *73* magazine, which they titled "The Lazy U Dipole," and published in September 1986. Bob's article was mainly about the Lazy Z, a half-wave antenna in the shape of a Z (or snake) with no sections in the horizontal plane. No great attention was given to the Lazy U; Bob simply suggested it be fed from the middle with a coax line, or at the top with a Zepp feeder. So although I cannot lay claim to the creation of an original antenna, I feel this article does give greater insight to a little-known DX antenna.

LAZY U 112-30-112 (CALLED 160-METER SHORT ANTENNA)

GROUND MOUNTED

112'

30'

112'   5'

Z = 80.6 -j4957
32° -0.71 dBd

1.830 MHz

ELEVATION ANGLE 5°
MAXIMUM GAIN -6.25 dBd

LAZY U 112-30-112 (CALLED 160-METER SHORT ANTENNA)

GROUND MOUNTED

1.830 MHz

AZIMUTH ANGLE 30°
MAXIMUM GAIN -0.71 dBd
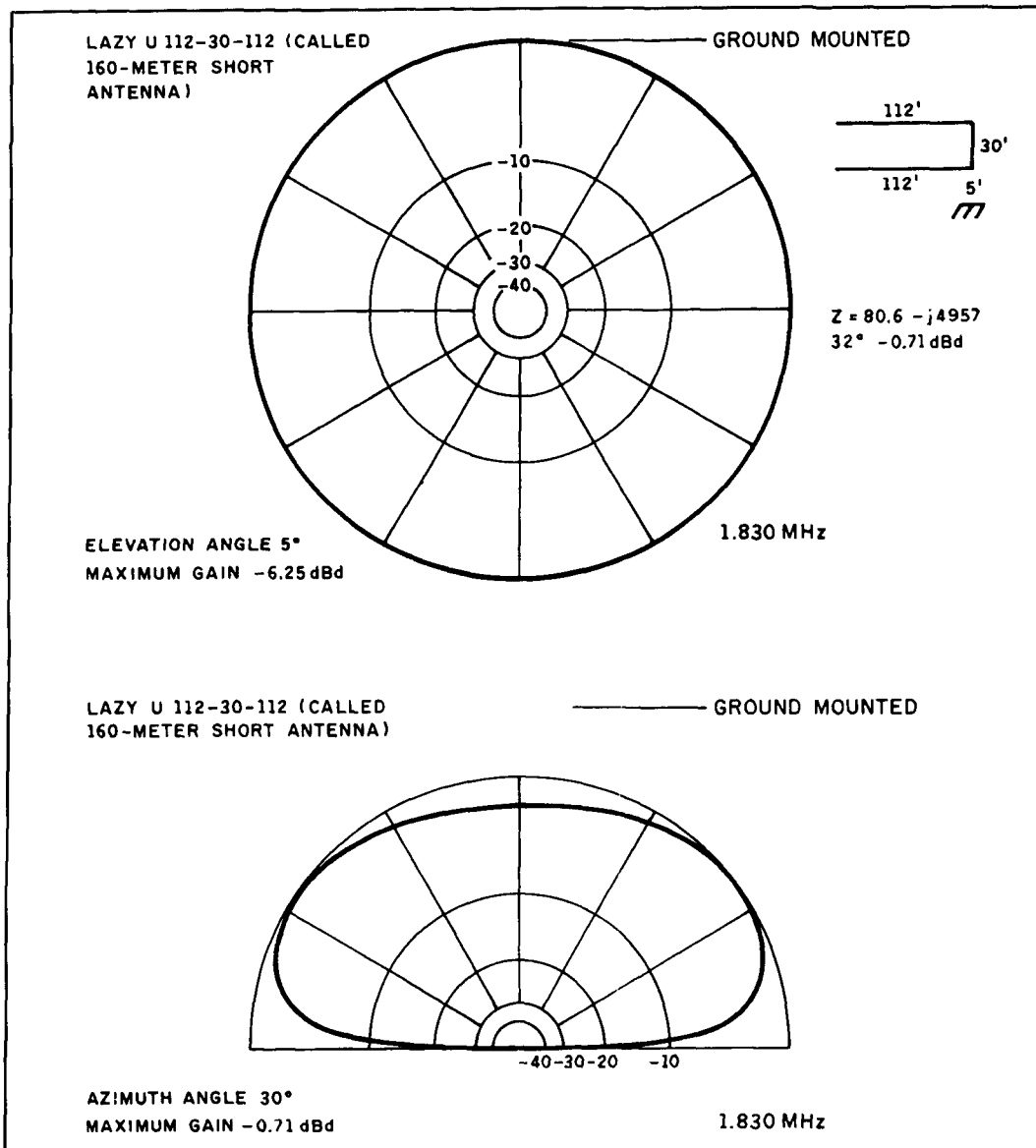
Figure 16. Radiation patterns for a 160-meter Lazy U with 30 feet in the vertical plane (also called a 160-meter short antenna).

Summary

I've attempted to show that a tree-mounted wire can provide effective long-distance operation on 160 meters without the use of radials, given the following conditions:
• The wire is one-half wavelength on the operating frequency.
• The wire has three separate sections: two in the horizontal plane, and one in the vertical plane.
• The horizontal sections must be parallel to each other, one above the other, for maximum cancellation of the horizontal component.
• The center of the wire must be in the vertical plane for as great a length as possible. Use the tallest tree available.

• The lower horizontal section should be 3 to 5 feet above the ground and protected from inadvertent contact.
• The antenna should be fed at the bottom end. A Pi network is suggested.

The Lazy U antenna has proven a very effective antenna for long-distance contacts, even without a radial ground system. It was designed and analyzed using the MININEC computer program. The results of the analysis were put into operational use and are consistent with the computer results.

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2. Henry Elwell, N4UH, "Controlled Vertical Radiation Rhombics, Part 2: Antenna Erection Performance," Ham Radio, April 1985.

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McKinney, Texas 75070

PROPAGATION LOSSES BETWEEN RANDOMLY LOCATED ANTENNAS AT UHF

Calculating path loss for terrestrial RF links between portable equipment under variable conditions

W hen performing a link analysis to predict the operating range between terrestrial stations, the losses caused by terrestrial effects modify the path-loss factor significantly. Free-space calculations are no longer applicable. This was discovered decades ago when portable equipment first came into use and leading communications equipment manufacturers sought ways to quantify the problem. Volumes of experimental data have been reduced to empirical curves that provide "average" and "typical" path-loss factors based on variables such as ground conductivity, antenna height, and classification of surrounding environment.

Unfortunately, the tendency to use free-space propagation and adjust for "terrestrial losses" still exists among the academia.

This results in predicted ranges that often are orders-of-magnitude greater than actual operating ranges encountered when a communications system is finally fielded.

I will summarize some published experimental data and provide reference material often neglected by engineers when making terrestrial range predictions for the first time. (It is rarely neglected a second time.)

Predicting the strength of the received signal

Here's a list of the steps that are followed when predicting the strength of the received signal.
• Effective Radiated Power. The actual power (ERP) that leaves the antenna in the direction of the receiver.

• Free-space path. Propagation due to spreading from an isotropic loss source.

$$P.L. = \frac{1}{4 \times \pi \times RR} \qquad (1)$$

• Field intensity at receiver. This is given in watts per square meter and is numerically = (ERP)/(4 × π × R × R)
• Receive antenna aperture. This is given in square meters and is numerically = (Gr × 9 E16)/(4 × π × F × F)
• Putting it together, we have a received power of:

$$Pr = \frac{ERP \times Gr \times 9\,E16}{(4 \times \pi \times F \times R)^2} \qquad (2)$$

This equation is a form of the radar equation (see **Figure 1**), and the basis for most link analyses for an RF link. But what about propagation loss?

Equation 2 will work well in free space unless receiving 60 GHz in the presence of oxygen, or 24 GHz in a rainstorm. However, when the transmitter and receiver are close to the ground, there are additional losses due to the following:
• Multipath. Reflections from any object including vegetation, earth, metal objects, rocks, trees, birds—absolutely everything that doesn't have a 377-ohm impedance.
• Absorption. A large factor at certain frequencies, but not very important below 1 GHz. After all, a transistor radio will pick up stations through the walls of your house.
• Shadowing. Reception of the signal even if you can't see the transmitter. The losses are unpredictable.
• Ground conductivity. The bending and loss of signal strength when the antenna is close to the ground. These losses are significant and very variable.
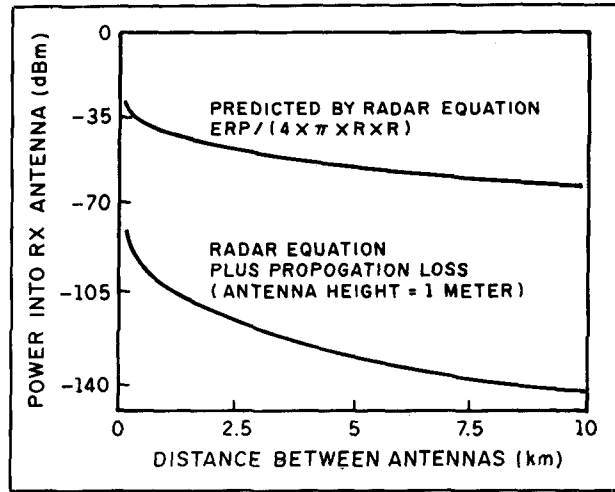
Figure 1. Comparison of received signal strengths for 1-watt ERP and 1 square-meter RX antenna. (Source: Reudink[4]).

• Cross polarization. For randomly located antenna sites, the polarization of the signal at the receiver is the vector sum of the reflections. The resulting polarization will be unpredictable.
• Antenna pattern. The inability to predict where the antenna will point. The patterns of both transmit and receive antennas change with location.
• Noise Temperature. If − 174 dBm is used for the noise-floor calculation, the receiving antenna may have directional properties due to reflections and be electrically pointed at a hot part of the sky, like the sun. Although this may not change the received signal strength, it may greatly modify the final signal-to-noise, especially at UHF and below. This includes atmospheric and manmade noise.

The list goes on. It's obviously impossible

$$\frac{E}{Eo} \quad = \quad 1 \quad + \quad Re^{jD} \quad + \quad (1 - R)\,Ae^{jD} \quad + \quad \ldots\ldots\ldots\ldots$$

| | Direct Wave | Reflected Wave | "Surface Wave" | Induction Field and Secondary Effects of Ground |

where   R = Reflection coefficient of the ground
A = "Surface Wave" attenuation factor

$$D = \frac{4 \times \pi \times H1 \times H2 \times F}{3E8 \times d}$$

F - Frequency
H1 - Height of Tx ant location
H2 - Height of Rx ant location
d - Distance

Table 1. Bullington's attempt to describe the phenomenon of losses which occur when the transmitter and receiver are close to the ground.

$$\text{Propagation loss due to terrestrial conditions in addition to the radar equation loss.} \quad = \quad 10 \, \text{Log} \left(\frac{Ht \times Hr}{(\text{Distance})^2} \right)$$

where Ht and Hr are the elevations of the transmitting and receiving antennas above ground.

Note that this factor results in an overall (Distance^4) characteristic for power loss instead of the square law predicted by the radar equation.

Table 2. Manipulation of Bullington's description, with some approximation for grazing paths, polarization, and antenna heights yields Equation 4.

to model completely the strength of a received signal at nonsited (random) locations. However Bullington[1] attempted to describe the phenomenon in the manner shown in **Equation 3**, found in **Table 1**.

Manipulation of Bullington's description with some approximation for grazing paths, polarization, and antenna heights yields **Equation 4**, in **Table 2**.

The typical effect of **Equation 4** is as follows: For antennas 1/3 meter above the ground and 100 meters apart, the average additional loss will be 30 dB greater than predicted by the radar equation.

At 1000 meters, this additional loss becomes 70 dB greater than that predicted by the radar equation.

Figure 1 compares the expected field strength at a receiving antenna for free space and for Bullington's propagation loss for transmitter and receiver antennas located 1 meter above the ground. However, the propagation loss information is extremely dependent upon the terrain and type of structures, as well as the antenna heights. **Figure 2** and **Figure 3**[2] give the propagation loss for specific environments with different antenna heights. In these cases, a 1-GHz nonsited terrestrial link was characterized from between 1 and 500 km.

These curves are typical of UHF because the differences due to frequency are somewhat insignificant compared to the differences caused by the terrain. **Figure 4** shows the median received signal power versus frequency for a constant transmitted power to many suburban locations.[3] Note that the difference in mean path loss from 400 MHz to 1000 MHz is only about 6 dB, while the propagation losses exceed the free space loss from 25 to 80 dB.

Note also that the urban and suburban curves are much worse due to the presence of multiple scatterers. However, even smooth plains (an almost ideal terrestrial situation with no trees and obstructions) provide propagation losses of 20 to 30 dB over 1 to 10 km if the antennas are close to the ground (2 meters).

If one of the antennas is considerably higher (15 meters), the loss drops to 10 to 20 dB. Still, this is a significant reduction of the power margin. Bullington's methods predict that as the antennas get closer to the ground, the losses skyrocket, although the attenuation versus distance remains fourth-order. A vertical antenna sitting on the flat earth at ground level doesn't provide a good situation for short range UHF links with low power levels. However, the radiation lobes that create sky-wave and E/F-layer
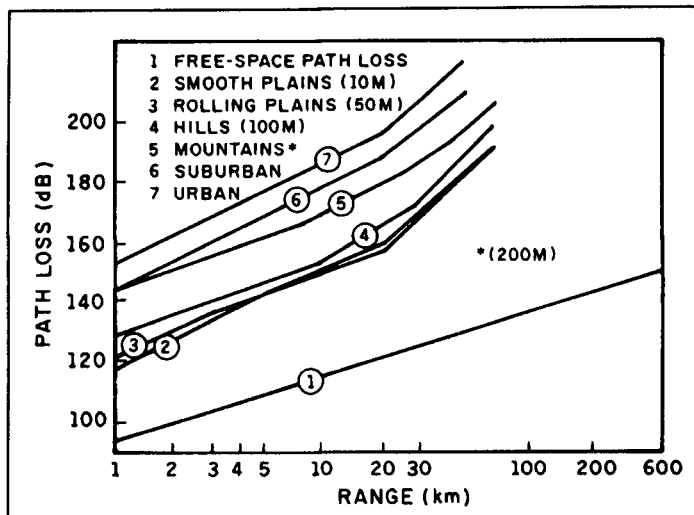
Figure 2. Path loss (dB) versus range (km) antenna height equals 2 meters for RX and TX.

long-distance propagation may work well. It all depends on the signal path.

Inside buildings, the problem presents about the same level of complexity due to the reflections of wiring, air-conditioning, and various materials and shapes. **Reference 4** tries to deal with this and provides a few equations as a guide. Anyone attempting a link design should get a portable signal source and calibrated receiver, and work out a data base for this problem. It's easy to do and the calibration accuracy of the equipment is insignificant when compared to the expected losses you'll be measuring.

The figures presented do not represent worst case. They represent "average," "typical," and "use-me-if-you-want-to-be-realistic" situations. If you don't consider the terrestrial propagation losses, there will be a large discrepancy in the operating range.

Effect of terrestrial multipath on bit-error rate

There's another source of havoc created by the propagation and multiple reflections in trying to demodulate the data transmissions. A receiver will normally process a received waveform one symbol at a time, using a matched-filter technique tailored to the transmitted waveform. The presence of multipath signals—that is, reflections from any object illuminated by transmitted power—greatly affects the accuracy with which decoding can be performed due to the overlapping returns. Intersymbol interference is created when a bit being decoded is overlapped by a previously transmitted bit that was delayed due to reflections.

How prevalent are these reflections? Here's an example:

Your eye is an antenna sensitive to electromagnetic energy (light). The sun is the transmitter. Look out the window. Everything you see is electromagnetic energy reflected off grass, trees, and other objects. If your eye lacked spatial orientation (directional gain) and color discrimination (frequency), it would behave as an unfocused photocell. How, then, could you then tell a tree from a bush? Our entire vision, and the "vision" of a receiver, is a composite of reflections—not just one or two prevalent sources.

On the other hand, if you look at a clear sky, it's easy to detect a dot that may be a bird or airplane. That's the difference between terrestrial nonsited and ground-to-air links.

When an overlap of transmitted signals
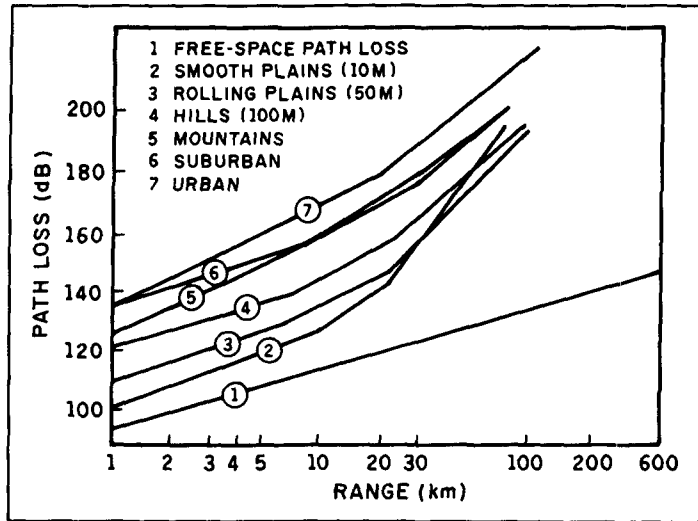
Figure 3. Path loss (dB) versus range (km) antenna height 15 meters TX, 2 meters RX.

occurs at the receive antenna, there's an increase in the bit-error rate which cannot be improved by increasing the transmitted power or lowering the receiver noise floor. There are, however, techniques that can improve performance by using adaptive equalization to suppress the multipath components. This requires substantial signal processing capability and may not be cost-effective in some systems.

There are two primary methods of transmission that can minimize this decoding loss. One method involves using an estimator-correlator type receiver typified by the RAKE structure.[5]

An alternative method is to use M-ary signaling to increase the bit rate while keeping the symbol rate slow to avoid intersymbol interference. However, the M-ary receiver is substantially more complex than conventional binary schemes.
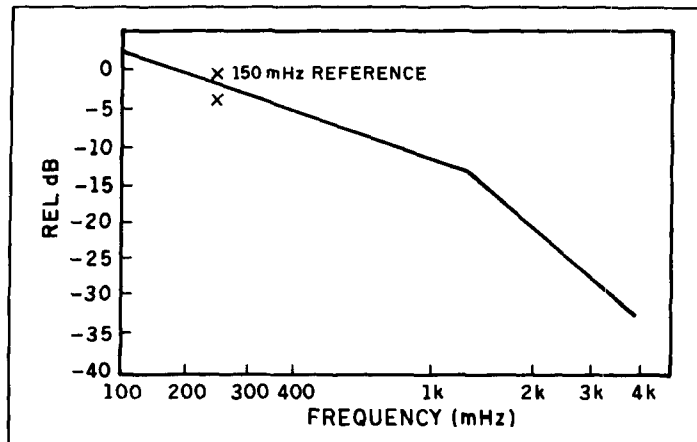
Figure 4. Median values of received signal power at suburban locations (dipole for TX, whip for RX, constant power at all frequencies).

Reference 2, "Advances in Packet Radio Technology," is an excellent article which deals with this modulation/demodulation.

Summary

If you're doing a link prediction for mobile, portable, nonsited systems, use the reference material listed at the end of this article. The material is the result of years of effort by people who have field tested their equipment. It is recognized by those in the business as an accurate statement of what will happen in the real world.

Try it yourself, as I have done, and you'll be a believer, too. You can't afford to ignore it. An additional 50 or 60 dB loss in a mobile or portable link, which hasn't been allowed for, can wreak havoc with a system's operation.

Remember, this data should only be applied to systems operating under the conditions where the measurements were taken. Space, air-to-air, and most air-to-ground or fixed-sited terrestrial links do not have these losses. ■

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5. G.L. Turin, "Communication Through Noisy, Random Multipath Channels," *IRE Convention Record*, Volume 4, Part IV, 1956, pages 154-166.

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2. A. Longley and P. Rice, "Prediction of Tropospheric Radio Transmission Over Irregular Terrain: A Computer Method," *ESSA Technical Report*, ERL 79-ITS 67, 1968.

PRODUCT INFORMATION

PSPLOT Program Available

PSPLOT, a printer utility program for Brian Beezley's MN,YO, and NP programs, takes K6STI's .PLT format and converts it into PostScript commands to send to a PostScript-compatible laser printer. Developed by Paul Terwilliger, NX1H, PSPLOT produces plots that adhere closely to those generated by the PLOT program supplied with MN, YO, and NP.

The printer utility program gives you the ability to print up to 18 plots on one page, overlay up to four plots on one coordinate chart, read command inputs from a DOS text file, use linear-dB or log dB scales, and use polar or rectangular coordinates.

PSPLOT is available for $25. To obtain a copy, or for additional information, contact Paul Terwilliger, NX1H, 11 North Shore Road, Chester, New Hampshire, 03036.

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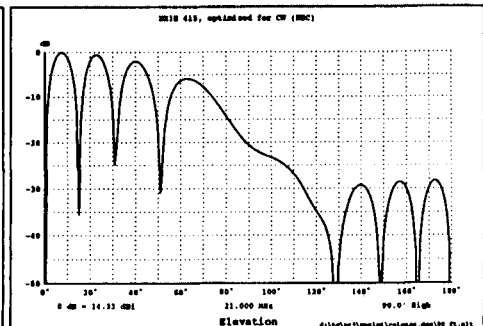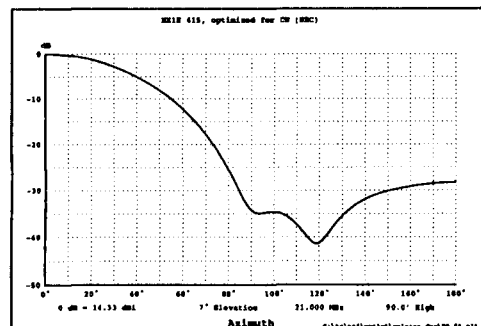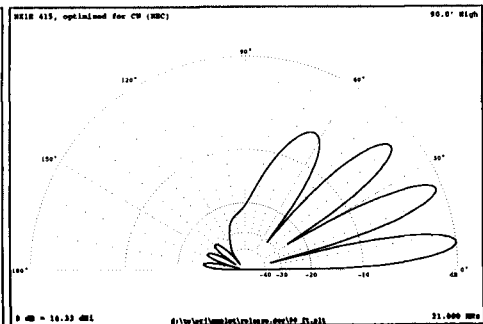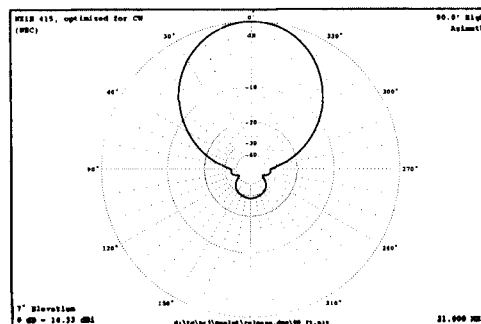
THE FINAL TRANSMISSION

Software—An American Industry At Risk?

This month Communications Quarterly introduces "The Final Transmission"—a forum for our readers and authors. The opinions expressed in this column are those of the writer and not necessarily those of the staff of Communications Quarterly. Ed.

As a designer and user of both hardware and software systems, I can't help but observe the explosion of software titles and wonder at the rationale behind the pricing structure. Somehow it doesn't seem reasonable that a software product price should be equal to, or greater than, that of the hardware on which it is run. What effect will this pricing discrepancy have on our ability to compete in a global software market? Could it be that if we in the United States aren't creative and efficient in our design of high-performance software, and very aggressive in the pricing of these products, we may soon find ourselves facing the same predicament in the global software market that we now face in the consumer electronic industry? This piece will first look at the current U.S. software/hardware pricing structure and then review its ramifications in the global marketplace.

Pricing versus development costs

In many respects, the basic business operations in the development of both hardware and software products are quite similar in many respects. It would be difficult to distinguish the actual final product as hardware or software from the management functions alone. The initial product-definition brainstorming, market targeting, top-level program management, design and production planning, tooling, project tracking, production, sales and distribution, maintenance, and so on, are all tasks common between hardware and software product development programs. There are certainly differences on a micromanagement scale. Tooling for a hardware project may include a state-of-the-art NC milling machine, while a software project may require a state-of-the-art software development system. But to the companies' comptrollers, those two "tools" are identical—a line-item property cost on the balance sheet.

The differences between hardware and software product design are even obscured at the basic design level. A hardware engineer very often breadboards various concepts to test the design of a circuit, a mechanical construction, or other physical concept. A software engineer may also breadboard a new algorithm or new screen presentation to test his design. From a management perspective, such tasks represent the general process of breadboarding. The medium is incidental.

There are those who argue that software is just more difficult to design than hardware. This argument is loosely supported by the inordinately long development times required for the design of some software. However, unexpectedly long and costly development in any product, hardware or software, is less a function of the peculiarities of the task and more a function of ineffective management during the development phase and inefficient use of the human resources dedicated to the task. There's no clear and obvious fundamental difference in the design progression of hardware and software systems. Consequently, inefficient development strategies can't be used as a justification for high software development costs. The apparent design complexity of software over hardware may very likely be more a consequence of software design management than of differences in complexity.

Hardware and software product distinctions

There are some relevant distinctions between hardware and software product development in the initial production stages. For example, it's generally easy to discover a hardware problem, say, a hole in the wrong place. Design flaws may be much more difficult to isolate in a software product. But in both cases, a fix must be engineered after the problem is isolated. In a hardware product, all of the defective product must be discarded or reworked. In a software product, new distribution media must be produced. Hardware problems are comparatively simple to isolate, but difficult to repair. Software problems are comparatively difficult to isolate and simple to repair. Of course, a well-engineered product should encounter few catastrophic problems in production. Therefore, it's reasonable to argue that even production-related troubleshooting is similar between hardware and software product developments.

Another very significant distinction between hardware and software products is apparent in the production phase. A hardware manufacturer must deliver a final piece of hardware, usually made from an array of physical materials (like sheet metal, electronic components, wire, screws, bolts, paint, etc.). Software product production typically requires little in the way of raw materials, other than disks for the actual code and paper for the associated manuals and packaging. A hardware manufacturer generally requires a force of skilled and semiskilled labor to assemble the various physical items into a finished product. The manufacture of a software product is rather accurately described as a simple printing process.

Hardware and software products also differ in the area of after-sale support. When someone purchases a hardware product, he is pretty much on his own in learning to use it. Some form of service support is provided during the warranty period, but there's rarely an 800 number to call should one have any questions about product operation. Even if a person wishes to repair his hardware item, no detailed support other than a service manual is typically available. (Of course, an individual is often not even legally "allowed" to repair his software products.).

The software industry has created a vast support service structure to explain the intricacies of software programs to their users. Why? The need for such a system could be a result of either less than diligent

users or poor product design and documentation. Most likely it's the result of some of both.

Hardware and software warranty support take different forms. If a new automobile breaks down during the warranty period, you can return it to the dealer for warranty service. The dealer must maintain a vast service infrastructure, including trained labor and costly specialty equipment, to support the service task. However, hardware (like an automobile) demands continuous service over its lifetime, so the warranty support infrastructure may be more effectively amortized.

If a software product fails to perform properly during the warranty period, one generally telephones the software *manufacturer*, not the dealer. Very often, the software dealer is simply a distributor with little or no knowledge of the products it vends. That places the warranty-support burden on the manufacturer. Software warranty-support service still requires professionally trained personnel but, in general, needs little in the way of costly equipment. Again, a well-engineered product should require little warranty support.

Product liability is also significantly different between hardware and software products. It's rather unlikely that a software user of a spread-sheet package or a word processor will be mortally injured by the software. The user *may* suffer a coronary when the software mysteriously locks up the system after hours of laborious effort, but that's his problem. Software manufacturers simply indemnify themselves against any injury or damage suffered as a result of using their products. Even if a software package is directly responsible for some damage, such as a defect in some end product because of bugs in a drafting or CAE program used in its design, the software manufacturer holds itself harmless.

Hardware manufacturers don't seem to be able to insulate themselves from damage caused by their products. Hardware manufacturers are almost always held liable for the safety and proper design and operation of the items they sell. Product liability is a very significant risk in hardware manufacture. However, even under this burden, manufacturers of personal computers manage to market their products at prices that are lower than that of some of the software products that may be used with them.

To summarize, it would appear that basic management systems used in the development of hardware and software products are comparable. Similarly, the design of hardware and software products should be

comparable in well-managed environments. After-the-sale warranty maintenance of hardware and software products are perhaps difficult to compare, but could be considered similar in warranty liability—albeit with displaced cost centers—for well-engineered products. So, in terms of basic development and maintenance, hardware and software products can be considered roughly similar. However, manufacturing cost and product liability are significantly more costly for hardware than for software products.

Once the product is developed and the non-recurring engineering (NRE) is complete, the total development cost is amortized over total production volume of the product—independent of whether the product is hardware or software. Products of comparable development cost and sales volume should be similarly amortized independent of the medium of the product.

What are we paying for?

The purchaser of a $2000 oscilloscope receives a real, physical piece of material—the oscilloscope. The purchaser of a $2000 software product receives little in the way of physical product (other than disks and manuals). If both of these hypothetical products have similar development costs and volume distributions, the amortization schedule for each should be comparable. Why then does the software product price equal, or exceed, the hardware product price?

One answer proposed to this question is that software programs provide improved productivity and, in turn, improved profitability for the user. Or, in other words, the software product provides a service which results in value added in business conduct. Apparently, it's considered reasonable for a software manufacturer to demand a significant share of that value added by using a pricing structure generally independent of product cost.

This an interesting concept. Rather than base product price on product cost, why not base it on the value the purchaser accrues from the product? If this rationale is considered a sound premise for defining product price, what should an actual software manufacturer be prepared to pay, in terms of a percentage of its gross profit, for such indispensable services as telephone communication, transportation services (ie., shipping), or even electric power service? Indeed, without any one of those services, contemporary software manufacturers couldn't function. Perhaps the price of those services should be adjusted to allow the service providers to collect their fair

share of the value added by the service rendered. Because services like communications and transportation are absolutely essential, the price of such services should logically be high enough to pretty much capture all profit, or at least a significant part. Of course, with the manner in which electric power has permeated all our lives, no one should be able to afford electric power.

To carry the argument further, what if this concept were applied to health care services? How should hospitals and surgeons determine the cost of excising a potentially fatal diseased appendix or performing emergency cardiac surgery? Should the patient be charged on the basis of value added to his life by the operation? It seems that basing the price of a product on the *potential* value added is totally subjective and therefore generally unrealistic—unless, of course, you can get away with it.

It has been suggested that some software manufacturers actually charge different prices for substantially the same software product, depending on the environment in which it is used.[1,2] If this is true, should automobile dealers review the station of potential purchasers and the final application of the product when computing the price of an automobile? Should a salesperson purchasing a vehicle for sales calls be required to pay a much higher price than an ordinary commuter pays for the identical vehicle?

The concept of price based on value added could be extended to the actual computer on which software is run. Clearly, the software can't function at all without the machine. Therefore, no matter what the price of the software, the value added by the machine is at least equal to the software price. In fact, because numerous different software products may be run on a single machine, the value added by the machine is indeed much greater than that of any one software product installed. Perhaps hardware vendors should *license* machines to end users rather than sell them, just as is common with software. Each time someone loads a software product, he sends a value-added fee to the hardware manufacturer. To make it simple, the amount of that fee could simply be the purchase price of the software product installed.

This may sound ludicrous, but isn't this equivalent to paying additional fees for each network workstation served by a single software product? When a manufacturer sells a machine, that manufacturer makes it possible for a single individual to operate a very large number of software products of almost limitless variety. When a software

vendor sells (licenses) a software product, the software does only a single task (if one can call word processing or computer-aided engineering "single tasks"). The hardware manufacturer doesn't penalize the end user if multiple software products are run on its product—even if they are run concurrently. Purchasers also pay no penalty if multiple users operate the same hardware, either as a stand-alone unit or concurrently as part of a network. Why then should it be reasonable for a software manufacturer to penalize the end user (require multiple licenses) for multiple users of a single software product?

If this is considered reasonable, shouldn't hardware vendors demand that each user of a specific hardware unit purchase a valid hardware operator's license for that particular system. Consumers would not stand for such a practice in terms of their hardware, but they support it with virtually all of their software. There is clearly a value-added inequity between software and the hardware environment in which it is used.

Another issue often raised to defend software pricing structures is that of intellectual property. It is argued that software manufacturers maintain that program users benefit from the "intellectual property" of the software creators. Consequently, those software creators have the right to be compensated for each use of their intellectual property.

This, too, is an interesting concept. Just who is the software creator? It's not a company. An individual creates the intellectual essence of a software program. A company may provide a favorable infrastructure and a catalytic environment for innovative thought, but the actual intellectual property is the creative inspiration of an individual—not the company. Such is true of both hardware and software designers. Note that patents, by Constitutional direction, are *in all cases* issued to the inventor—not the employer. So if the creators of intellectual property should be compensated for each use of their property, the actual authors of a software program should receive a royalty from each product license vended that contains some of their intellectual work. Is this normal practice in the compensation of software developers? Likely not.

There's no less intellectual property content in a hardware product like a computer system, an automobile, or a building, than there is in a software product. One could reasonably argue that there's a much more significant amount of intellectual property resident in many hardware items. This is especially true when one considers the creative expression that goes into the styling of those

products—the artistic essence that invites customer appeal. Nevertheless, in hardware as in software, the actual creators of these products are rarely compensated equitably for the use of their intellectual "essence." Therefore, the argument that software pricing structures serve to reward intellectual property creators just isn't credible.

Dealing with "bugs"

An annoying feature of many software products is that they seem to include "bugs" and nonfunctional features. Generally, we wouldn't tolerate such quirks in a hardware product. We'd take the item back to the place where we purchased it and demand a replacement or a fix for the problem. But the purchasers of software products are expected to accept bugs as part of the inherent characteristics of software.

Suppose the CEO of, say, a major software company purchases a nice new "state-of-the-art" sports car. Upon examining the car in detail, he discovers that the passenger door won't open. Then, upon contacting the dealer to report this problem, he's told: "The 'door problem' is a documented 'bug' in this particular release. It will be corrected in future releases. There are two work-arounds: 1) use the driver's door for passenger ingress and egress, or 2) have the passenger climb over the door when the top is down. This is simply a minor 'bug' in the product. The product is otherwise totally functional." Oh yes, the CEO also discovers that the advertised AM/FM Compact Digital Quadraphonic 500-watt per channel audio system wasn't installed. When he inquires about this omission, he is advised that this particular feature has not yet been implemented. Unfortunately, there's no suggested work-around and no targeted release date.

It's likely CEO would probably be less than satisfied with his purchase. How strange, then, that his company expects those who buy their software to accept bugs as an inherent part of the product. Should the purchaser of a piece of hardware be more receptive to accepting some resident bugs and nonfunctional features in the hardware, if they are to be accepted in software?

It's quite possible that most software users could cope with a hardware defect more easily than they could a bug or nonfunctional feature in a software product. Suppose someone has a CAE structural design package that mysteriously errs on rare occasions, but is correct most of the time. If he uses that product to design a 100-story office structure or a suspension bridge, the

results could be disastrous (of course, the software manufacturer is held harmless). About the only recourse the user has is to confirm the design by some independent means, hopefully with a product of another manufacturer. This, of course, renders the original product valueless. One must also accept the possibility that there may be bugs in the verification product.

Total Quality Management (TQM) is one business strategy that has been used successfully by Japanese companies. Numerous U.S. hardware manufacturers are adopting TQM principles in an effort to compete in the global market. Perhaps TQM applied to software development would result in more reliable, bug-free software products, if not more competitive pricing. **Reference 3** details the consequences of poor software design and the need for quality-assurance software-design tools to assure quality in software development. Interestingly enough, those who choose to read this reference will note that the software quality-assurance design tools and related software reviewed were developed by Hewlett Packard, but in Bristol, England—not the United States.

From 800 to 900-number support services

In an effort to reduce cost, or perhaps to profit from some perceived value added, some software manufacturers are eliminating their 800-number support and instituting 900-number support services. If this practice is being implemented to charge the actual users of the support service directly and provide a significant price reduction to the basic purchasers of the product, the practice can be excused—even applauded. But it's unlikely that we'll see any significant price reductions. It seems irresponsible for a software manufacturer to market a poorly documented, buggy, complicated software product and then proclaim that customer support constitutes value added, for which they are entitled to additional revenue. This is particularly true when the support is often "vapor support." Of course, it's equally irresponsible for the software user to badger the manufacturer's customer support service with simple operational questions that are easily answered by reading the documentation. Indeed after-the-sale support is a major sticking point for both the manufacturer and the end user.

A possible solution

To eliminate this problem let's consider the following suggestion. What if software products were offered under two different pricing structures? The first would operate under current practices: The price covers the cost of the product and some form of reliable, effective end-user support. The second pricing structure would be governed by a type of caveat-emptor policy: You knew what you wanted, you bought it, and now you're on your own. The only support one gets is a warranty that the media is readable. However, the manufacturer would have a binding legal obligation to repair bugs and implement advertised, but unimplemented, features on any release. Any product vended must be, in all ways, the product advertised. Beyond that, the manufacturer doesn't want to hear from you again—except when you purchase more products.

Suppose that the price of the second option—zero user support—were perhaps one-tenth of the full service price. A word-processing package with a full-service price of $495 would retail for $49.50, with zero service. Which would must people be likely to purchase? Based on the arguments presented here, it's difficult to defend a much higher price for *any* software package—regardless of its full-service price. In general, the majority of users do not make significant use of the manufacturer's support services, and on those occasions where they do encounter some difficulty, the manufacturer is often of little help. Many users would very likely prefer a no-service option, at significantly reduced price.

The $49.50 price tag is in the range of the price of typical limited production technical text books, and about twice that of typical third-party computer information texts. The design and development of such technical works may entail somewhat less labor effort than the development of a software product, but such works also enjoy much more limited distribution. Amortization of software is likely to be much more effective. Also, technical works certainly include similar intellectual property content as software works. Therefore, a no-support market price for software products on the order of that for other typical written technical matter seems reasonable.

If this suggestion were seriously proposed to software manufacturers, they would surely balk. User support is touted by some manufacturers as a very major cost in their operation and, correspondingly, a justification for the comparatively high prices charged for software. But if one suggests its elimination with a corresponding price reduction, customer support might suddenly become of negligible importance. If custom-

er support service is a major cost, why not eliminate it and give the end user a corresponding price break? If it isn't a significant expense, why not improve and expand the 800-number service?

In both domestic and global markets there are certainly a vast number of products where price is reasonably based on value added. There are also products where price seems to have no relation to anything concrete; take art, for example. In general, competition is the feedback mechanism that regulates price in our domestic and global commerce. When it comes to software, it seems that domestic software manufacturers use the competition to determine what the market will bear, and simply price their products at that point. There seems to be less activity in adjusting price to gain a market share, as is common in other highly competitive markets. That approach is an acceptable and ethical business practice. However, it can only be successful as long as there are no other outside agents with designs on our markets and who have the wherewithal to aggressively price their products.

Based on the comparative costs of development, production, marketing, maintenance, and liability of hardware and software products, it doesn't seem reasonable that a software product should have a price equal to or, in many cases, more than twice that of the hardware system environment in which it is to be used. But the purpose of this piece is not to make a plea for low-price software, but rather to suggest that our vast and very lucrative software industry may be in serious jeopardy; in part, because of its pricing practices.

Is our software industry in jeopardy?

The United States was at one time a world leader in many industries—automotive and consumer electronics are two. However as global markets matured, our domestic industries failed to keep pace with the winds of global competition. In effect, our domestic policy was: "The consumer will buy what the manufacturer builds, at the manufacturer's price." Concern for such product features as customer wants, product quality, price, and price/performance ratio were of little domestic concern. Certainly they were of *some* concern to manufacturers, but much less so than a favorable RONI (return on net investment) on a quarter-by-quarter basis. But others in the world market were not indifferent to the potential market value of those seemingly insignificant product features, and were willing to take the risk of long-term investment. Long-term strategies were formulated by our global competitors exploiting attributes dismissed by our domestic manufacturers. Such strategies, in concert with others, proved quite successful for our foreign competitors.

Based on the historical record in the hardware industries, one can logically predict that an efficient, aggressive, creative, cost-effective approach to software design will eventually prevail. Software will become a commodity with a highly competitive pricing structure; that is, cost-based pricing.

Unfortunately, as with hardware, such creative business practice probably won't be a fruit of this country. If we are not very careful, diligent, and innovative in our business practice, we will face the same outcome in our software industry that we now face in virtually every aspect of our hardware industries.

It's virtually impossible to purchase a truly American-made, high-performance, high-technology consumer-electronic product. In the future, will high-quality, high-performance, cost-competitive software be a product of India rather than America? Or of Japan? Or China? Or even the Soviet Union? The dominant players in the software game can't yet be predicted accurately, but based on the historical record, it's logical to predict that America will soon cease to be a significant software supplier to the global market. Even now, some American software manufacturers are having significant software developments engineered outside the United States.[3] Such products are only superficially "American made." They are more accurately foreign made and American purchased.

Japan is truly a modern-day Phoenix. It literally rose from the ashes of total destruction following World War II to become a (perhaps the) world influence in the global market. Oddly enough, the United States was instrumental in the rebirth of Japan. We are, in part, an instrument of our own demise. Japan literally rebuilt its national infrastructures from virtual total physical destruction and near total emotional and cultural devastation to capture numerous facets of the global hardware market. The country accomplished this feat in less than 30 years, with the comparatively archaic technologies of the 1950s and 1960s.

Even now, there seems to be another Phoenix fanning the faint, infant embers of rebirth—the Soviet Union. Unlike post-war Japan, the Soviet Union, although presently

in a state of disarray, doesn't have to contend with a ruined physical infrastructure. Antiquated, surely, but not destroyed. The political and civil difficulties will obviously be very difficult to overcome, but there's little doubt that this people will persevere. It's also reasonable to assume the Soviet Union possesses equally as many innovative and creative assets in the form of engineers, scientists, and managers as post-war Japan, and even contemporary America. With access to today's technologies, rebirth may take a matter of years rather than decades. And unlike Japan, the Soviet Union has a vast wealth of untapped natural resources with which to purchase the tools it needs from global vendors—though not necessarily American vendors. What they sorely need are business strategy tools. Once again, we are supplying these tools. The secrets of McDonald's and PepsiCo and Domino's and Microsoft and Hewlett Packard and AT&T and...will help this Phoenix rise anew from its economic ashes. It would be irresponsible to simply assume that the Soviet Union cannot recover its economic means. To idly ignore that threat or, worse yet, fail to capitalize on that opportunity would be foolhardy. Japan may soon own us. Will the Soviet Union soon own Japan? And what if the organizational stirrings of the European Community (E.C.) are successful, or even marginally successful.

For a people struggling with an antiquated physical infrastructure, a software industry is a natural for consideration as a target development industry in the modern technological world. Software development and production requires no costly factories or facilities. Indeed, a sterile office/factory environment is about the least productive environment for creative innovators. Rural campus compounds tend to be much more catalytic to innovation. And rural is one thing of which the Soviet Union enjoys plenty. Compared with factories dedicated to competing in a global hardware market, the tooling cost required for mounting a globally competitive software industry is almost negligible. A globally competitive software development posture lies primarily in the creativity and innovation of engineers and managers, and less in the physical tools of the trade. Unfortunately, the Soviet Union is very likely only one of several raptors surveying global software market opportunities.

How can the U.S. maintain a competitive edge?

When Japan suddenly became a serious challenger of our domestic hardware products, one approach to countering that threat, rather than innovation and head-on competition, was an attempt to simply limit the import of Japanese goods. This approach has had little success domestically, and cannot be applied to a global market.

If we are to compete successfully on a global scale, we must muster all of our innovative strategies. But even now, software manufacturers are diverting very significant resources away from critical research and development and pouring those resources—the lifeblood of the industry—into copyright litigation.[4] One can certainly approve of a company working to protect its property. However, when the life cycle of the property being protected is likely shorter than the court actions taken to defend it, one must question the wisdom of litigious proceedings. Historically, the patent and an aggressive defense against patent theft hasn't protected the American hardware industry from foreign capture. Indeed, about half of United States patents are issued to foreign agents, and many of those are to Japanese concerns. The litigious actions aggressively pursued by American software developers may simply serve as the very instrument of the destruction of our software industry. Such actions serve to dilute resources critical to innovation and at the same time may very well be forging the legal weapons that our adversaries require to capture our software industry.

The efficient, cost-effective development of high-performance, reliable software with highly aggressive, competitive, cost-based pricing may not be as much a simple benefit to the consumer as an essential strategy for the survival of one of the last innovative industries in which we still enjoy a global influence.

All is not yet lost. But can the management of our software industry show the same creativity, innovation, and wisdom in their business practices needed to maintain our global lead in the software market that the software design engineers have shown in their creations that have given us this transient supremacy? Are we a nation at risk with respect to our software industry?

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the gamma match, and ran three no. 14 wires up to the top of the tower, spaced about 6 inches apart. The tower is about 60-feet high, with a 205BA at 60 feet and an interlaced 4 element for 15 and 4 element for 10 at 68 feet. The three wires are tied together at the bottom, and fed via L networks for 160 and 80. On 160, one network gives better than 1.5 over the whole band, while on 80, the 2:1 SWR bandwidth is about 100 kHz. The gamma matches were very narrow on both bands—on 80, about 25 kHz. I came to the conclusion that a larger feed rod (ie, more capacity to ground) helped drop the Q. This system certainly works very well although the next experiment is to be elevated radials. At the moment, adding radials has no measurable effect. The system is quarter-wave resonant at 2.55 MHz.

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