COMMUNICATIONS QUARTERLY

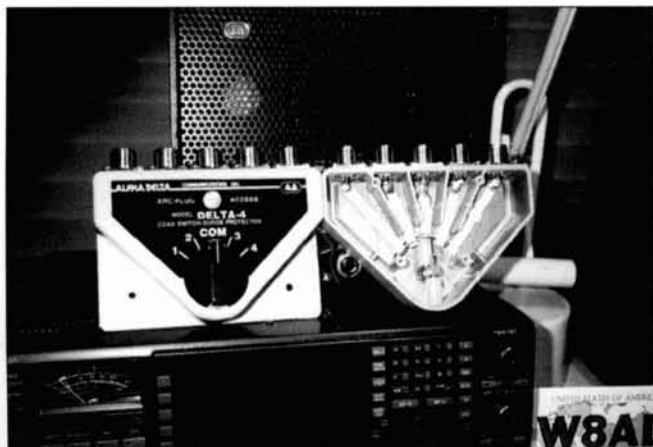Summer 1998 $9.95

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COMMUNICATIONS QUARTERLY

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CONTENTS

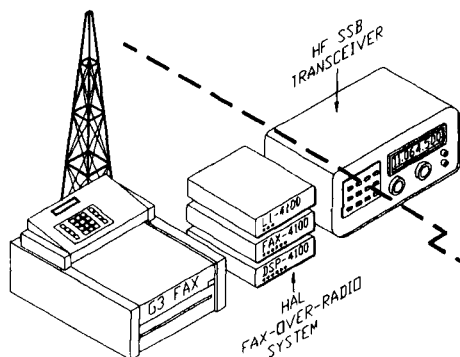Volume 8, Number 3   Summer 1998

page 19

page 96

On the Cover: Go first class: add a BASIC Stamp to your next ham project. See Mike Hall, WB8ICN's article "A BASIC Stamp Project for Amateur Radio," beginning on page 9. Cover art by Bryan Bergeron, NU1N.

EDITORIAL

Are You a Writer?

Each time a new copy of *Communications Quarterly, QST, Sprat,* or *RadCom* comes over the transom, I pounce on it—eager to scope out what's inside. Will this be a classic "anthology" issue packed with intriguing projects and seminal experiments? Or will it be a "ho-hummer" that survives on a back shelf to a ripe old age without so much as a single dog-eared page? While the answer may have *something* to do with the magazine's editor, it also has *a lot* to do with its readers. After all, as writing instructors are constantly reminding us, all good writers were—and are—*avid readers* first!

Most *Communications Quarterly* readers like to devour ideas in order to synthesize, envision, and generate more ideas. We may do this in the privacy of our own scratch pads, over the Internet, on the air, or in technical forums at hamfests. From all this mental activity, important technical problems often get solved and interesting projects come to light. Clinical psychologists call this "creative" or "generative" behavior. Writing teachers call it a good start.

This brings me to the main point of this editorial. You don't have to be a great writer to create a fascinating article. All you really need is something interesting to share, and a desire to share it.

As a case-in-point: When I graduated from college, I was convinced that I was a lousy writer and that I would never publish anything. Writing was pure agony, so why would I ever want to do it? But the fact remained that I liked to read and I loved to build.

Then, something changed. One day while in QSO with another ham on my new home-built rig, my contact suggested, "Why don't you write up that design and submit it to one of the ham magazines?"

"I can't write. They'd never accept it," I said.

He replied, "Look, I'm a technical writer by profession, and I'm telling you it's easy! Just analyze two or three articles you really like, and make your article look like their articles!"

So, I read a few pieces closely, noting how most were divided into various sections such as introduction, theory, construction notes, testing and alignment, conclusion, etc. But I still needed one more push. Fortunately, in those days, two of the four major U.S. ham magazines were published here in New Hampshire, where I live. I subscribed to *Ham Radio*, so I called Editor Jim Fiske and asked what I should do to get started.

Jim said, "Drop by the office. I'll show you."

Drop by the office?? That's heady stuff for a kid just out of school! To make a long story short, Jim spent about three hours showing me around the facility—explaining each step of the publication process and pointing out all the things *he* liked to see in articles to make them more valuable, useful, and appealing to readers. To me, it was red-carpet treatment for a total unknown! To Jim, it was simply part of the editor's job—nurturing potential new authors and encouraging them to share their work.

The point is that very few technical authors start as competent, fluid writers. That's part of the editor's job—to pound out the dents, guide, and encourage us while we hone our skills.

Are you a writer? Your editor may be the only one who knows for sure!

de Rick Littlefield, K1BQT

Enter the Editor

One of the things I enjoy most about editing *Communications Quarterly* is working with our authors. Over the 10-plus years I've been editing ham magazines, I've dealt with all kinds of writers; but I really enjoy working with the novice writers who have entrusted me with their first ever articles.

A favorite "first-timer" wrote the lead article for this issue, "A BASIC Stamp Project for Amateur Radio." Mike Hall, WB8ICN, and I first met on line when he was trying to locate an article in a back issue of the *Quarterly*. He later invited me to check out the QRP project he had posted on his Web site. I could see he had something of value to share with our readers, and let him know that I thought his project had merit. Then I put a folder with his name on it in the file drawer and waited to see what would materialize.

About a month later, when Bryan Bergeron, NU1N, and I were trying to come up with a cover idea, I mentioned the possibility of an article using the BASIC Stamp. In no time, he had created the dynamite cover you see on this issue. I lost no time in e-mailing a copy of the picture to Mike, and he rose to the bait.

After a few months of diligent testing and retesting his project, then writing and rewriting, Mike sent me his story. Everything about it showed how excited he was about his project and how much he wanted to share what he had learned. Because this was his first attempt there were a few things to prune and a few things to smooth out; but, when all was said and done, we both were happy with the result.

In all my years as an editor, I've enjoyed the relationships I've forged with my authors. In the months that we work together refining articles, I learn a lot about them. There are many authors I've worked with for years—people who've become old friends. But the authors who mean the most to me are the ones who say: "I love what you did with my article. It sounds just like me." Your article is your article; I'm just the technician, making a little tweak here and there to help you sound your best.

So if, like Rick years ago, you've been toying around with the idea of writing an article for *CommQuart,* just give me a call. I'll be happy to give you a "tour" around the office and all the encouragement you need!

Terry Littlefield, KA1STC
Editor

TECHNICAL CONVERSATIONS

In Support of resonant elevated radials

Dear Editor:

In my article on elevated radials (Winter 1998 issue of *CommQuart*), I stated that "resonant radials can be used with practical elevated ground-plane type antennas, and to simulate 'connection' to ground for numerical modeling programs like NEC-2 which does not allow a wire to touch lossy ground."

Why do we want to simulate ground "connection"? To model a GP-type antenna (using NEC-2) in its operating environment. One example is to model the effect of a grounded tower (which may support a Yagi antenna) used to support a wire antenna. In this case, the radials associated with the GP-type antenna should not be too close to the ground. To model the grounded tower, we could use 4- or 6-resonant radials, at a very low height (a few wire diameters) over real ground, simulating a stake ground—for a well-grounded tower use a larger number of resonant radials.

Dick Weber, K5IU, disagrees with my conclusions (see "Technical Conversations," Spring 1998 issue). In fact, he concludes that "it is hard to find anything positive to say about 90-degree radials." On this point, we strongly disagree.

Let me model (using the industry standard MOM code NEC-4D) a quarter-wave vertical wire (AWG #10) fed against four radials, height 2 meters over average ground (5 mS/m, 13). For a frequency of 3.75 MHz, the radiator height is 20 meters. As the length of the radials is changed progressively—0.25, 0.238 (resonant length), 0.2, 0.15, 0.1 and 0.05 wavelengths—the space wave gain decreases (reference is the case where the radials are resonant), + 0.02, 0.0, - 0.13, - 0.32, - 0.54, and - 0.83 dB, respectively.

The above assumes we have matched the antenna, coil Q-factor 300, since when the radials are shortened the monopole is no longer resonant. The base reactance changes progressively, + j 8.1, - j 1, -j 30.9, -j 79.2, -j 159, and -j 374 ohms, respectively.

The ground wave field strength changes in a like manner to the space wave gain, + 0.03, 0.0, - 0.13, -0.33, - 0.56, and - 0.86 dB, respectively. For the radio amateur, these are small gain changes. The important factor, particularly for the case of phased arrays, is that a GP-type antenna (half-loop or monopole) that is resonant when fed against an infinite number of buried radials (simulating a PEC ground) is approximately (very closely) resonant when fed against resonant elevated radials.

John S. (Jack) Belrose, VE2CV
Aylmer, QC
<john.belrose@crc.ca>

Comments on elevated radials

Dear Editor:

In the spirit of Letters to the Editor approaching article-length proportions, I'd like to comment on part two of the "Elevated Radial Wire Systems for Vertically Polarized Ground-plane Type Antennas" by Jack Belrose (VE2CV), appearing in Spring 1998 *Communications Quarterly*, and a letter to the editor from Dick Weber (K5IU).

Elevated radials remain controversial for good reason. The only measurement that tells us field strength is, not surprisingly, a field strength measurement! The only measurement accurately indicating ground system efficiency change at a given location with a given antenna is, not surprisingly, an "A-B" comparison of one system to a control system with only a radial system change. Neither Belrose or Weber offer conclusive evidence of performance change, since neither has offered an A-B test with control over errors.

The person Belrose credited for installing a 0.17-wavelength-tall radiator with a successful elevated radial system was also the engineer responsible for delivering WVNJ a properly functioning radio station. The owners of WVNJ were repeatedly assured the WVNJ radial system was operating with near-perfect ground system efficiency. They were repeatedly told unexpected signal shortfalls at WVNJ (several dB) were due to ground losses unrelated to the elevated radial system.

WVNJ's signal, despite results of several NEC-based modeling sessions and written engineering assurances to the contrary, improved several dB when a conventional ground system was installed.

To understand why such conflicting data might occur, readers must first understand how the theoretical field strength curves (like those Belrose published) are determined.

1. With fixed power, often with a non-directional pattern from a test antenna, field strength readings are taken along several paths extending radially from the antenna site.

2. The slope of field strength readings taken in step (1) are "matched" to the closest slope of

an FCC ground conductivity graph. This graph is used to estimate average or typical ground conductivity along many radials at the operating frequency. (Radial, in this context, is not a ground system radial. It describes a radial direction from the transmitting antenna.)

3. Using an approximation published in the FCC rules, or FCC supplied graphs, estimated ground conductivity from step (2) is used to estimate field strength versus distance normalized to the power levels in the working system. This provides an estimate of field strength levels that should be achieved if the antenna is providing acceptable efficiency.

Every step in this long complex process contains tolerances and errors. Errors include, but are not limited, to the following:

1. How the field strength meter was used (adjusted, positioned, and read). Even a slight movement of the meter in any direction will often affect readings.

2. Absolute calibration of the FS meter in dB/mV. Calibration error is typically 1 percent of full-scale voltage anywhere on the scale for properly calibrated FS meters in a controlled environment. For an example of how seemingly tight tolerance affects results, using a 25-mV/m scale to measure 5 mV/m (a normal condition) results in almost 1 dB of measured FS tolerance!

3. Conditions at or near the measurement point (including multipath, re-radiation from surrounding conductors, and soil conditions). While doing broadcast proofs and field strength readings, I've observed hundreds of measurement points where moving only a few feet can change FS readings by several dB. This is a common problem and is illustrated by the scattering of readings in Belrose's graphs (and Belrose picked the "best" examples with the least scatter).

4. Errors and compromises as the engineer goes through the various steps looking for a conductivity curve with "approximately the same slope" as measured FS, and then manipulates that data through various steps to determine efficiency.

We are fortunate if a reading falls within a few dB of "predicted values!"

For example, **Figure 5** of Belrose's article has so much scattering that it's difficult to tell what the "real" FS is at any distance. Looking carefully at that graph, we find FS readings

Mike Hall, *WB8ICN*
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Dayton, Ohio 45418-2916

A BASIC STAMP PROJECT FOR AMATEUR RADIO

Digital control of the Ten Tec SSB/CW board

Many recent articles have featured a new approach for homebrewers who wish to digitally control their stations. This approach centers around Microchip's 16C56 and 57 series microcontroller chip (uIC).[1] These uICs are used in many applications, including a rather unique mini-processor system on a chip called a "Stamp"—more precisely put, the BASIC Stamp I (BS1) and BASIC Stamp II (BS2). Both chips are produced by Parallax, Inc.[2]

I first discovered these little jewels about nine months ago during a random search of the Internet for a simple, yet effective, interface controller. I had heard about the Stamp a few years back at the Dayton Hamvention™, but was told it was far too slow for any practical ham radio use. I have since learned a very important lesson: Always do your own research and form your own judgments! The Stamp may be too slow for applications such as a frequency counter in the RF range; but, considering the broad spectrum of controller uses I've come across for the Stamp in just a few months, I'd say we've only scratched the surface of the applications for Stamp opportunities in the ham shack.

Introduction

I've divided this article into four sections. The first describes what led me to this project
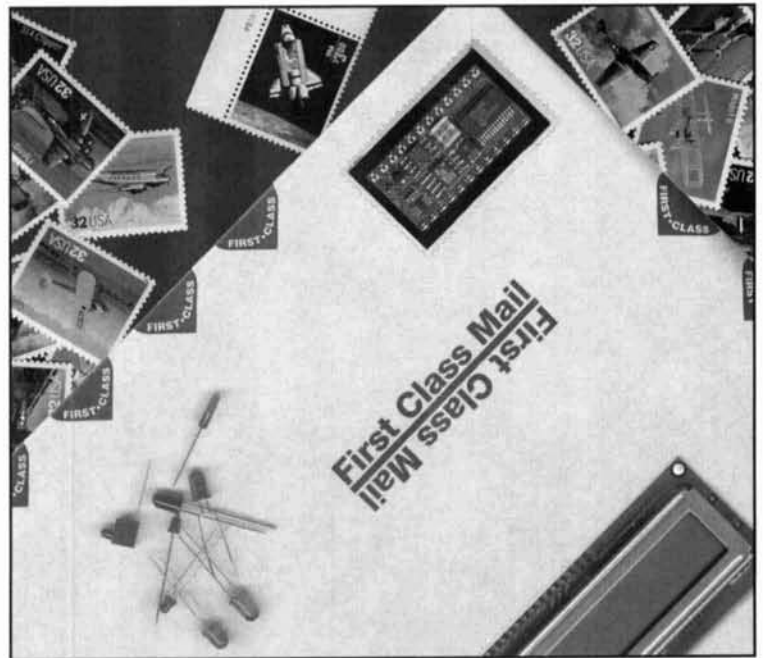
The BASIC Stamp is a first-class addition to your ham project.

and to the BASIC Stamp. In the second section, I detail both BASIC Stamps I and II, and the editors used to write, debug, and program code for the Stamps. Third, I discuss my project—a replacement of analog controls (RF gain, coarse and fine-tuning, and bandpass) on my Ten Tec SSB/CW receiver board with digital pots controlled by the Stamp—explaining both the hardware configuration and the code used to make it all work. Finally, the fourth section provides a general wrap up, some thoughts on enhancement of the interface project, and some

1. Microchip Technology, Inc., 2355 West Chandler Blvd., Chandler Arizona, USA 85224-6199. Internet URL: <http://www.microchip.com>.
2. The BASIC Stamp is a registered trademark of Parallax, Inc., and is available from Parallax, Inc., 3505 Atherton Road, Suite 102, Rocklin, California USA 95765. Internet URL: <http://www.parallaxinc.com>.

Figure 1. Schematic diagram of the BASIC Stamp 2 IC, rev. A.

rather exciting ideas for uses of the Stamp and its actual processor: the PIC 16Cxx series chips.

The beginning

I bought the Ten Tec CW-SSB Receiver kit Model 1056[3] at the 1996 Hamvention. It had been years since I'd put something together for fun, and I thought this kit might help me get back into homebrewing. I built the receiver and had many hours of fun using it to copy CW and SSB stations. The kit lets you wire the local oscillator to the band of your choice; I chose 40 meters. After a while, I pushed the receiver

3. Ten Tec, 1185 Dolly Parton Parkway, Sevierville, Tennessee USA 37861. Internet URL: <http://www.tentec.com>.

aside to make room on the bench for other interests. Then, late last summer, a project at work involved developing a simple motion control with a limited budget outlay. After some research, I found many robotics applications that used the BASIC Stamp. At the same time, I ran across several *Nuts & Volts*[4] magazine articles that showed a wide range of other uses for this nifty device.

Being a big kid at heart, I sort of got lost in the spirit of this little novelty and began an earnest search for any and all info on the Stamp. After surfing the Parallax Web site, I discovered an e-mail list that was an ongoing virtual conversation of Stamp enthusiasts from all over of the world. The Parallax description of this e-mail list was something on the order of: "The BASIC Stamp's e-mail list keeps you in touch with BASIC Stamp enthusiasts around the world." They weren't kidding! I learned lesson number two here: Unless you have a good size e-mail box, get the digest form of the Stamp list! I normally receive one to three digest versions of posts to this list daily, which amounts to about 12 to 30 messages in each digest.

For the most part, the Stamp I chip is the most popular version. However, when looking for a way to digitally control my Ten Tec receiver board so I could manipulate it with my PC for a digital mode, such as packet or SSTV, I ordered the more versatile of the two: the Stamp II. But after playing with the Stamp II for a few days, I discovered that most of the application notes for the Stamps, at least those available at that time, were for the Stamp I.

The Stamp II was relatively new and just starting to catch on. Because I wanted to ensure the most applicable platform for as many users as possible, I decided to rewrite the Stamp II code and hardware design written by "Stampster" Guru Beau Schwabe (he also chose the Stamp II for this type of application) for the Stamp I. Schwabe, an Electronic Research Specialist with Sabolich Prosthetic Research and Development in Oklahoma City, thought up the concept of using the Dallas Semiconductor digital pot (DS1267) chip with the Stamp to replace analog pots with a digital counterpart.[5] The reengineering of the hardware portion was rather easy, as I do this sort of thing for a living, but the code conversion was a challenge. Now that I've gone both ways with the code, it seems rather simple and enjoyable—sort of like a crossword puzzle. Lesson number three: Get your hands dirty and have some fun. Ask for help and advice.

I completed my first Stamp I project in February 1998 and posted it on my Web page. Only one problem remained after the initial work was done. Unfortunately, the only real fix was a band-aid, which I detest using unless

there's no other solution. The Stamp I only has eight I/O lines—not enough for my needs. I did find a way to use just one I/O for the four function switches needed for the receiver board. I created the hardware to determine which switch I had pushed using an RC network and a decision-making logic table in software. The problem was that the RC network caused a clicking noise in the receiver, which I couldn't easily mask. I found a way to use a lot of capacitors to pull all the low-frequency AF (RC time constant clicking noise) to ground, but I didn't really care for this approach. The band-aid saved the patient but was a messy fix to the problem.

Because I couldn't sell myself on this solution—which meant that no one else would ever buy it—I rewrote the code and redid the hardware for the receiver board using the Stamp II. Because the Stamp II has twice the number of I/O lines, I had plenty of lines for switches and didn't need an RC network to determine which switch had been pushed. Versatility provided the solution. I had enough I/O lines so each of the four switches had separate access to the CPU of the Stamp processor.

What is a BASIC Stamp?

The BASIC Stamp is nothing more than a small computer, actually a microprocessor that runs Parallax BASIC (PBASIC) programs. The PBASIC language is proprietary software developed by Parallax for use in the Stamp I (BS1) and Stamp II (BS2). It's burned into the 16C56 or 57's EPROM, depending on which Stamp you're using. Each Stamp has fully programmable I/O pins which can be used to directly interface to TTL-level devices, such as buttons, LEDs, speakers, potentiometers, and shift registers. By adding a few extra components, these I/O pins can be connected to non-TTL devices, such as solenoids, relays, RS-232 networks, and other high current/voltage devices. Though available in various physical packages, all versions of the BASIC Stamp have the same logical design, consisting of a 5-volt regulator, resonator (an oscillator that serves as the system clock), serial EEPROM, and PBASIC interpreter.

When programming a Stamp, the editor (free from Parallax; I downloaded mine before I got my first Stamp) "tokenizes" (that is, it adds symbols, called tokens) the PBASIC program and stores it in a non-volatile serial EEPROM, which can be read from and written to by the interpreter chip. This interpreter chip fetches

4. *Nuts & Volts*, 430 Princeland Court, Corona, California USA. Internet URL: <http://www.nutsvolts.com>.

5. Dallas Semiconductor Corp., 4401 South Beltwood Parkway, Dallas, Texas USA 75244. Internet URL: <http://www.dalsemi.com>.
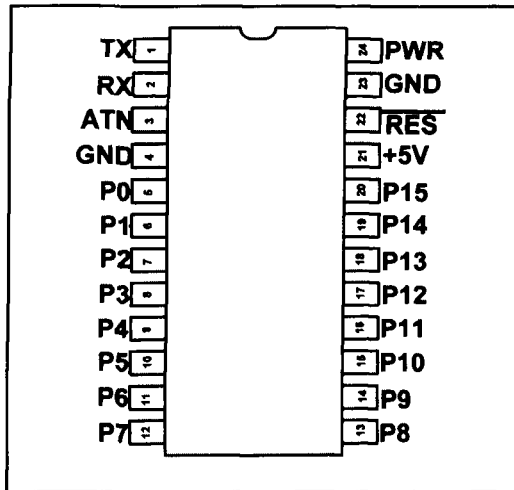
Figure 2. The 24 DIP module seen here with a pin-out table and pin descriptions.

the instructions one at a time and performs the appropriate operation on the I/O pins or internal structures within the interpreter. Because the PBASIC program is stored in an EEPROM, it may be programmed and reprogrammed almost endlessly without the need to first erase the memory. To program BASIC Stamps, you simply connect the Stamp to an IBM PC or compatible (someone has finally developed an editor for LINUX users) and run the editor software to edit and download your program into the Stamp's EEPROM.

The BS1 has eight I/O pins, room for 256 bytes of data and program instructions in EEPROM (about 80 to 100 instructions), and executes an average of 2,000 instructions/sec. There are seven word registers, all of which are byte addressable; two are bit addressable. The system clock is internal and runs at 4 MHz. The I/O lines are TTL at speeds of 300 to 2400 baud. The PC-to-Stamp interface is via the PC parallel port.

The BS2 has 16 I/O pins plus two synchronous serial pins, can hold about 500 instructions, and executes an average of 4,000 instructions/sec. There are 16 word registers, all of which are byte, nibble, and bit addressable.

The system clock runs at 20 MHz. The I/O lines are TTL running at 300 to 50 Kbaud; 0 to 19200 if you're using flow control and use the PC serial port for a programming interface. Current requirements are modest for both the BS1 and BS2—2 and 7 mA, respectively. Cost seems to be the only real hang up for folks like me. The BS1 runs about $34 and the BS2 is priced at $49, no matter where you shop. But considering the ease of programming, the simplicity and number of times you can reprogram, and the support network, I'd venture to say it's worth the price.

My idea was to develop applications using

the Stamp and then burn the code into a PIC once I converted it to assembly language. The new Parallax editor (version 1.2) is supposed to allow you to burn BS1 code into a 16C58, but I haven't tried. The instruction sets differ somewhat between the two Stamps, although many of the instructions are used for both. I won't go into a lot of detail here, but I suggest you check out the free, downloadable manuals for each Stamp.

Code, hardware, and interfacing to a receiver

First, let's take a quick look at the BS2 chip itself. I used the BS2-IC, Rev. A for my project. **Figure 1** shows the workings of the BS2. All of this actually fits on a small 24 DIP module, shown in **Figure 2** along with a table giving the pin names and a short description of the use for each pin Look at **Figure 1** again. The BS2 CPU is a custom Microchip PIC 16C57 (labeled U1). U1 is permanently programmed with the PBASIC2 instruction set. When you write a program for the BS2, you use a text editor or the Parallax BS2 editor and then download it to the BS2 through the serial port on your PC. When you download your program to the BS2, you're actually telling U1 to store your program in PBASIC symbols, called tokens, in the 2-KB EEPROM (U2, which is a 24LC16B). When you run the program, U1 retrieves these tokens, interprets the PBASIC instruction you wrote, then carries out that instruction. U1 executes internal instructions at about 5 MIPS. However, it takes several internal instructions to make up one PBASIC2 instruction; therefore, the real-time processing speed of the BS2 is around 3,000 to 4,000 instructions.

U1 has 20 I/O lines, 16 of which are available to the user for BS2 I/O use. Two of the remaining lines are for serial communications, and the final two are for interfacing with U2. The 16 BS2 I/O lines are 5-volt logic lines, usable from TTL to CMOS. They have approximately the same characteristics as the 74HCTxxx-series logic devices. Those lines which you direct to be an "input" I/O have little effect on any devices connected to them. Think of them as tristate, high-impedance input connections. Bits P0 through P7 make up one port and P8 through P15 make up the second port. These pins should be considered as two 8-bit pin ports and should never sink more than 50 mA or source more than 40 mA. Any unused I/O pin should be made into an output pin and connected to +5 volts.

It takes several milliseconds to write data into the EEPROM (U2), and the limitation on the number of write cycles before it begins to fail is around 10 million. But you can do an
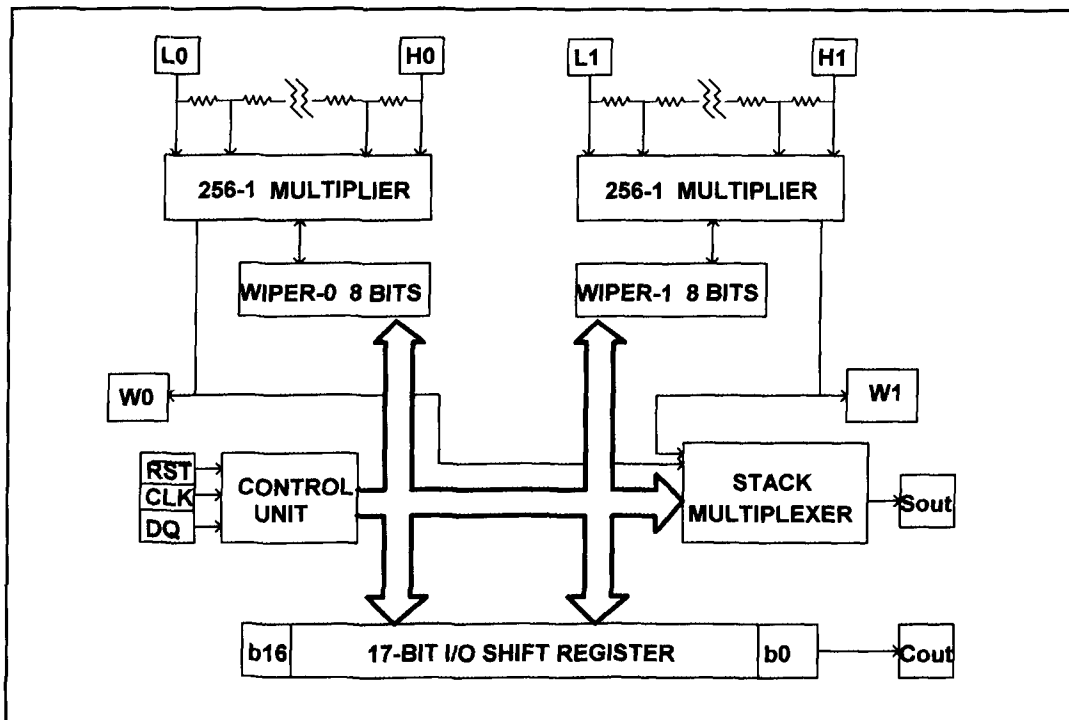
Figure 3. Block diagram for the Dallas Semiconductor DS1267.

awful lot of writes and downloads before you hit that number!

U3 is used as the reset circuit and is rather important during power up. For those who want an even more detailed description of the BS2, I recommend that you download the BS2 manual (*BASIC Stamp Programming Manual 1.8*) and read pages 207 through 212. For now, the important thing to remember is that pins 1 through 4 are used to communicate with your PC for writing, debugging, and downloading your BS2 programs. Pins 5 through 20 are the I/O pins you use to communicate with the real world—in this case, the Ten Tec receiver board—and pins 21 through 24 are power, ground, and system reset connections.

You can power the Stamp with a 9-volt battery connected to pin 24, which will provide you with a 5-volt regulated power source at pin 21, or just connect a regulated 5-volt power supply to pin 21. I used a 9-volt "transistor" battery connected to pin 24 because I didn't want to build a 5-volt power supply. If you use a 9-volt battery, a 5-volt source is automatically available through the Stamp itself.

The DS1267 (1267) is produced by Dallas Semiconductor and you can download the data sheet in PDF format from their Web site.[6] The 1267 consists of two digitally controlled solid-state potentiometers. Look at **Figure 3**, the block diagram for the 1267. You'll note that each pot consists of 256 resistive sections, each equaling about 39 ohms. When you transfer a

data word from the BS2, you're actually sending two bytes (16 bits) of information that tells each pot to move up or down in resistive value. The word is sent, along with a dummy bit (the storage resister in the 1267 is 17 bits long), to the 17-bit I/O shift register in the 1267. The RST I/O line from the BS2 enables the correct 1267 to receive data, and the data word is sent one bit at a time on the DQ I/O line, using the CLK (CLK I/O line) signal for timing synchronization on the DQ I/O line. Once in the shift register, the individual bytes are moved to their respective wiper controls through the stack multiplexer (controlled by the control logic). After the RST line goes low, the wiper position is changed according to the data byte sent from the BS2. **Figure 4** is the project schematic and will help you see how this all works.

As previously mentioned, the project itself was to provide a means to digitally control my Ten Tec 7.0- to 7.3-MHz Direct Conversion SSB/CW, Model 1056, receiver board. I removed the RF gain (RF), the coarse and fine tuning (CT and FT), and bandpass (BP) 10-k potentiometers and replaced them with the 1267s that would create the same effect as the 10-k pots. I used two DS1267s to replace the four pots on the receiver board. I connected the two chips to the receiver board via a cable harness. The function switches RF, CT, FT, and BP are momentary pushbutton switches sal-

6. Internet URL for DS1267 Dual Digital Potentiometer: <http://www.dalsemi. com/DocControl/PDFs/1267.pdf>.

vaged from an old VCR. The up/down switch is simply a toggle switch used to tell the pots whether to increase or decrease their values when I push one of the four function switches.

The idea is simple enough. I wanted the ability to control my receiver either from a PC or a keypad. But to get to that point, I needed to create this project with switches to prove the concept and later, as it turned out, to learn more about using the Stamp in many different ham applications.

I'll list and discuss each segment of code. I've laid out each segment in the text instead of having you jump back and forth from the text to another figure on another page. You may find it helpful to have **Figure 5**, the program flowchart, handy when going through the code. It provides a shorthand way to see what a program does as you step through it. (If you would like to receive a copy of the complete program, send your request, along with an SASE, to Terry Littlefield, KA1STC, at the Editorial Office address listed in this magazine.)

```
'Program: RCVR.BS2
'This program is an upgrade of WB8ICN's BS1
   Receiver/DS1267 Project
'P11 thru 14 are inputs for the BP, FT, CT, and
   RF pot controls
'p15 is used to determine if pots values increase
   or decrease
'P0 thru 3 are outputs to the DS1267
'Special thanks goes to Beau Schwabe for his
   contributions
'of the initial idea of using the DS1267 for this
   type of interface.
```

This first segment contains the comment lines I always put in my code to give the user an idea of what the program is all about. It also helps me recall why I wrote the program when I haven't looked at it for many months. Lesson number four here: Always include a lot of comments in the programs you write. When you compile your code, or in the case of the Stamp, tokenize it, the comments are left out so no extra memory is used or wasted. The comments will help you during debugging and later help you recall why you wrote what you did.

```
DQ con 0 'Pin 0 is for data connection
CLK con 1 'Pin 1 is for Clock
RST1 con 2 'Pin 2 is for reset of pot1&2
   (RF,CT)
RST2 con 3 'Pin 3 is for reset of pot2&3
   (FT/BP)
INPUT 11
INPUT 12
INPUT 13
INPUT 14
INPUT 15
```

This next section tells the Stamp what connections you're using. The "con 0", "con 1",

etc., tell the Stamp that bit 0, bit 1, bit 2 are used for the data (DQ), clock (CLK), and the pot you're selecting (RST1 is for pots 1 and 2, etc.). The "INPUTxx", where xx = bit number, tells the Stamp that bits 11 through 15 will be input pins only. These five input lines are from the five function switches. In a few minutes, you'll see what the function switches do. A word of explanation and caution is warranted here: Don't confuse "bit" numbers with actual chip pin numbers. The Stamp has no "idea" what a *pin* number is, nor does it care. The software tells the Stamp what to do by referencing *bit* numbers. Remember that the BS2 Stamp uses 16 I/O lines, so you're dealing with a 16-bit word for programming considerations. The nice thing about the BS2 is that you can address each bit by itself, or address a nibble which is 4 bits, a byte which is 8 bits, or the whole word which is 16 bits. I didn't use 7 bits (bits 4 through 10) on the BS2 chip, or pin numbers 9 through 15.

```
DSPOTA var word 'Word variable holding
   pot1&2 values
DSPOTB var word 'Word variable holding
   pot3&4 values
DSPOT1 var DSPOTA.lowbyte 'Variable for
   RF pot
DSPOT2 var DSPOTA.highbyte 'Variable for
   CT pot
DSPOT3 var DSPOTB.lowbyte 'Variable for
   FT pot
DSPOT4 var DSPOTB.highbyte 'Variable for
   BP pot
DIRA=%0000111111111111 'Output bits
   0,1,2,3 to DS1267's,
      'bits 4 thru 10 tied high since they are
         unused,
      'bits 11 thru 15 are input bits from the
         function switches
```

This part of the code tells the Stamp that I'll the use labels (nicknames) DSPOTx for storing certain values that I will then use to control the DS1267s. These labels make it rather easy to create storage locations for the values you need to use to keep track of in a program. The Stamp decides where to store the values under the labels you give it. The last line, "DIRA", tells the Stamp that bits 0 through 10 are output bits and bits 11 through 15 are input bits. Seven bits (bits 4 through 10) aren't used in the program, but I still declare them for safety's sake—a good habit for any programmer.

```
DSPOT1=200 'Presetting RF pot
DSPOT2=200 'Presetting CT pot
DSPOT3=200 'Presetting FT pot
DSPOT4=200 'Presetting BP pot
```

These four lines were included just to get the pot values into the ballpark. Otherwise, at power up, you'll be starting at zero for all four
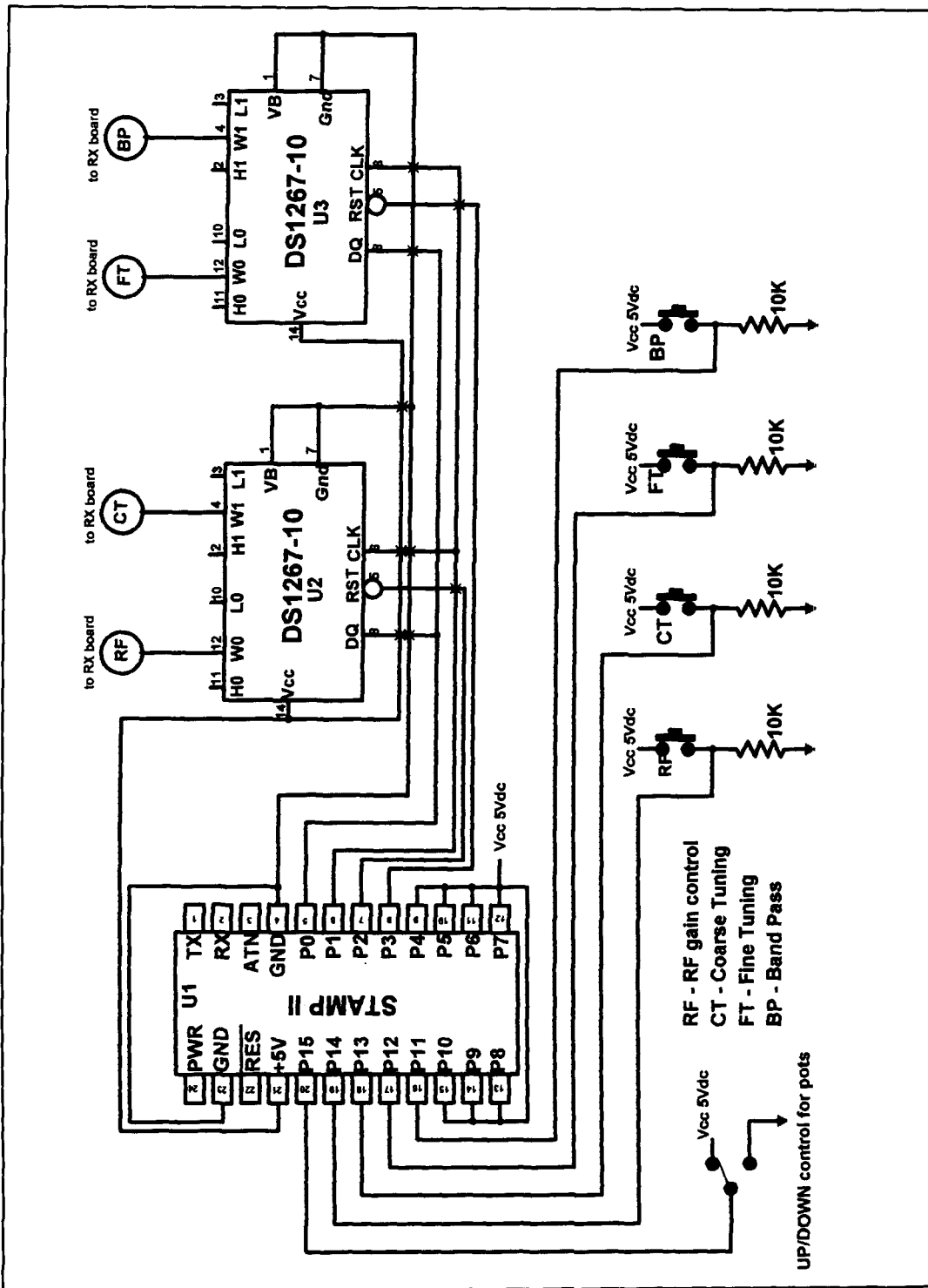
Figure 4. Schematic of the author's project.

pots and will have to get the RF and BP into usable ranges before you can start tuning. You may set these values to your liking once you get a feel for a good initial setting.

```
begin:
if in14 = 1 then rfsw
if in13 = 1 then ctsw
```

```
if in12 = 1 then ftsw
if in11 = 1 then bpsw
goto begin' if no switch was pushed, then start
    scan over again.
```

Believe it or not, this is the main program. If any of the "if...then" statements are true, say if you pressed one of the four function switches
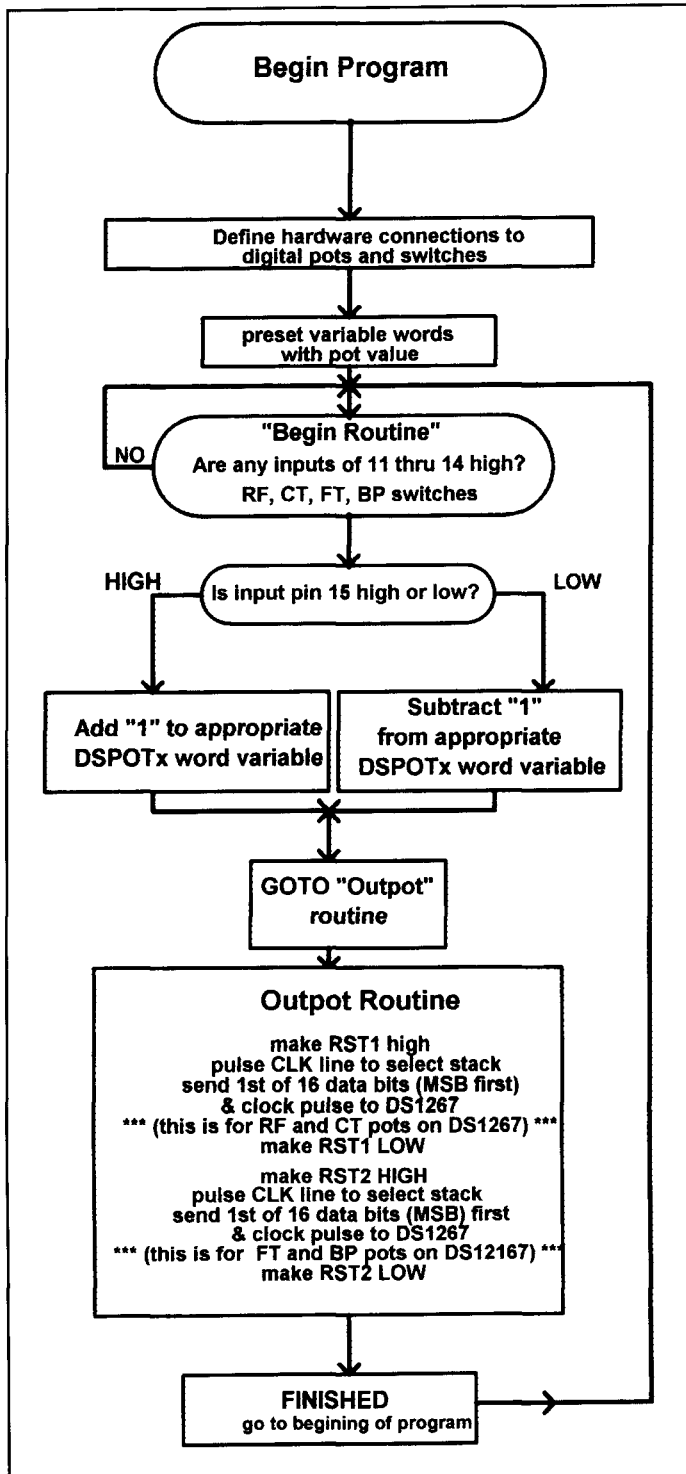
Figure 5. Program flowchart.

(this puts a high on that particular input I/O line), you then branch to a subroutine for that true condition. If no function switch was pushed, the last line, the "goto begin" statement, takes you back to the beginning of this loop and starts examining each switch again. Bits "in11" through "in14" are the input bits for the function switches.

```
rfsw:
    if in15 = 1 then rfup
    goto rfdwn
ctsw:
    if in15 = 1 then ctup
    goto ctdwn

ftsw:
    if in15 = 1 then ftup
    goto ftdwn

bpsw:
    if in15 = 1 then bpup
    goto bpdwn
```

You'll encounter this portion of the code if one of the function switches has been pushed. Depending on which switch is pressed, you enter one of the four subroutines above. Each does the same thing: it checks to see if the Up/Down switch is set to 5 volts (indicating the need for an increase in pot value) or to ground (indicating the need for a decrease in pot value). For instance, if you pressed the CT button to change the coarse tuning of the receiver, the if/then statement would have been true; i.e., "in13 = 1". This action would have caused you to branch to the "ctsw" subroutine to see if you wanted to increase or decrease the CT pot value. This, in turn, would either increase or decrease the tuning in coarse increments on the receiver board. For our example, we'll say "in15 = 1" (meaning we want to increase the frequency). This will cause you to branch to the next set of subroutines (shown below), in this case the "ctup" subroutine. If "in=0", then the "goto ctdwn" routine would be selected.

```
rfup:
DSPOT1 = DSPOT1 + 1
gosub outpot
goto begin

rfdwn:
DSPOT1 = DSPOT1 - 1
gosub outpot
goto begin

ctup:
DSPOT2 = DSPOT2 + 1
gosub outpot
goto begin

ctdwn:
DSPOT2 = DSPOT2 - 1
gosub outpot
goto begin

ftup:
DSPOT3 = DSPOT3 + 1
gosub outpot
goto begin
```

```
ftdwn:
DSPOT3 = DSPOT3 - 1
gosub outpot
goto begin
bpup:
DSPOT4 = DSPOT4 + 1
gosub outpot
goto begin

bpdwn:
DSPOT4 = DSPOT4 - 1
gosub outpot
goto begin
```

Note: The "+1" and "-1" values are what I chose for my project. By using your own settings, you can increase or decrease the resistive values in chunks to your liking.

Three separate actions occur in each of these subroutines, all similar in nature depending on how you arrived at that particular subroutine. In our example, we want to increase the pot value for the CT pot. The "ctup" subroutine first adds 1 to the variable word "DSPOT2". Had we wanted to decrease in frequency (in15=0), the subroutine "ctdwn" would have subtracted 1 from the variable word "DSPOT2". After the 1 is added, you "gosub" to the "outpot" subroutine, which moves the variable word values out of the Stamp and into the DS1267's digital pots.

I chose to move all four pot values to the 1267 each time a switch was pressed, so I can use just one routine to move word values to the DS1267 instead of writing four separate move routines. Remember, you only have 2K of data/program storage capability in the BS2, so being stingy is okay—at least when it comes to using the Stamp memory. Once the values have been moved out to the DS1267 (using the routine shown below), you'll "return" to the "ctup" routine (or whatever routine you were using in this section). You'll then execute the last line of the subroutine, which is to "goto" the main loop of the program and start over.

Actually, you can save eight lines of code by using "goto outpot" and deleting the "goto begin" statements for each of the eight "gosub" commands and then changing the "return" statement (in the "outpot" subroutine) to a "goto begin" statement. I've kept my code as it is for debugging and enhancement ideas I'm currently playing around with. It's your call whether you want to save the eight lines of code.

```
outpot:
    high RST1 'Make RST1 high to start data
        transfer to pot1&2
    pulsout CLK,1 ' Pulse for stack select of
        DS1267
    Shiftout DQ,CLK,msbfirst,[DSPOTA\16]
        'Shift out values for pot1&2
```

```
    low RST1 ' Make RST1 low to end data
        transfer
    high RST2 'Make RST2 high to start data
        transfer to pot3&4
    pulsout CLK,1 ' Pulse for stack select of
        DS1267
    Shiftout DQ,CLK,msbfirst,[DSPOTB\16]
        'Shift out values for pot3&4
    low RST2 ' Make RST2 low to end data
        transfer return
```

This is the last part, and one of the most important. This segment of code is used to get the pot values out to the digital pots so you can change gain, frequency, or the bandpass. The first four lines move the RF and CT pot values out to the first DS1267 (U2) and the second four lines move the FT and BP pot values out to the second DS1267 (U3). Because I'm moving all four pot values each time I press a function switch, and each DS1267 has two pots on it, I must perform the actual move routine twice in this subroutine.

Let me address just the first four lines, as the second four lines work the same way. The first line selects the correct 1267 for data transfer. The RST command is used to enable the three-wire (DQ, CLK, RST) serial port operation of the device. The RST signal is an active high input (to the DS1267) and is required to begin any communication to the DS1267. The second line is really just a dummy line and serves no purpose in this application other than to send the dummy bit. Remember that the storage register in the 1267 is a 17-bit register and the data word is only 16 bits. The third line moves the data over, one bit at a time. The fourth line turns off the communication link between the Stamp and the DS1267. The "return" statement (last line) takes you back to the "ctup" routine, or whichever routine you came from. As mentioned earlier, you could change this to a "goto begin" statement and, with the corresponding changes in the xxUP or xxDWN routines, save eight lines of code.

Well that's the deal. It's not really difficult, and it's a neat way to go from analog to digital control for the Ten Tec receiver board. This concept can also be applied to any similar project; for instance, the DS1267 can be used to vary the 10-k pots in 256 step increments. This means that each time a function switch is pressed, you change the pot value by roughly 39 ohms. You can lower this value by paralleling a fixed resistor with the wiper pin on the DS1267. This will limit your total movement (resistive range) ability, but, for my project, this didn't present a problem.

Operating instructions

It's quite simple to operate the BS2-1267 interface. Unless you intend to change some

aspect or values in the program, once you've programmed the BS2 it never needs to be reprogrammed—even if you disconnect the 9-volt battery.

To operate this homebrew project, connect a 12-volt DC source to the Ten Tec board (you can operate the receiver with a 9-volt battery; see page 18 of the manual) and connect a 9-volt battery to the BS2 circuit. Once power is applied to the BS2, you're in the operational mode. The "Up/Down" switch controls the direction in which the receiver is tuned, the direction of the bandpass control, and the increase or decrease in RF. Five volts will cause all four function controls (RF, CT, FT, and BP) to increase in value. Ground connection (0 volts) will cause all four function controls to decrease in value. I've found that I seldom touch the RF control unless the signal is really strong. Once I have the receiver in the CW frequency range using the CT control (I prefer CW to SSB for QRP operation), I normally only use the FT control. The bandpass, once tuned for CW signals, will probably never need to be adjusted.

I've used a few tricks to enhance this project. These include paralleling either 5- or 10-k resistors with the CT and FT control lines going to the receiver from the 1267 chip. This provides much finer tuning control and is great for zero beating weaker stations. You can also change R5 on the receiver board and do the same thing. By all means, put the receiver board in a metal enclosure and ground it to the power supply's ground terminal. This helps reduce RFI and some broadcast interference if you live close to any commercial radio towers. I've been kicking around the idea of building a 40-meter bandpass filter using some toroidal cores. It won't cost much, but, because the receiver is a direct-conversion design, this filtering will surely help.

Conclusions

After a while, it can become boring to just turn on the rig and talk. Even with a DA1 or I7 callsign (those were some fun days—working the world on 5 watts and a vertical antenna), ragchewing has it limits and one can only talk about the weather, antennas and rigs, and the last few contacts just so many times. But there are lots of options when it comes to hamming. The BASIC Stamp is just one. I see the Stamp as the prototype platform for designing and implementing new ideas. To that platform, you can add a cheap (less that $5) PIC chip to use

as the final working microcontroller. To do so, you must learn a little PBASIC and assembly language programming, interface to the analog world, and use the digital devices that are getting cheaper everyday.

There are numerous homebrew projects you can create with a Stamp kit. For example, how about a Stamp or PIC packet station that fits in your shirt pocket; a home automation system that controls and monitors your heat, A/C, security sensors, and water usage; or a digital weather station you can access with your HT. You could add a PIC-controlled DSP box to your station which you can adjust, in real-time, to meet the needs of the current operating conditions, allowing you to hear those weak and rare stations. You could build a CW keyer, a phone dialer, and so on. Well, you get the idea. I hope to bring some of these ideas to my workbench, my Web page,[7] and maybe to future articles.

Is this the ultimate in a QRP receiver design? Definitely not. But it's a project I had loads of fun with and it provides an avenue for future experimentation. I learned a lot about receiver basics, how to use a BS2, and how to build up the courage to write an article and submit it for publication. I plan on using the receiver board for some other projects and eventually a QRP station. I'm not really sure what I like best: operating my projects after I've finished them, or just working on the projects because it's so much fun. I've had very few projects work as intended when first powered up. If they did, I probably wouldn't enjoy the homebrewing part of our hobby so much.

Acknowledgment

There's one gentleman who has played a major role in this project. Though not a ham, he is definitely a "Stampster" guru who deserves a mention of praise and thanks. Beau Schwabe is an Electronic Research Specialist with Sabolich Prosthetic Research & Development in Oklahoma City, Oklahoma. As I mentioned at the beginning of this article, he thought up the concept of using the Dallas Semiconductor digital pot (DS1267) chip with the Stamp to replace analog pots with a digital counterpart. Beau is a font of knowledge for Stamp applications and is always there to offer advice and encouragement. My personal thanks to him for the concept, but more so for his enthusiasm. He keeps me going on my project ideas. Thanks Beau! ∎

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MODIFYING A 160-METER ELEVATED RADIAL VERTICAL

Ensure near equal radial currents with optimal performance

In the latter part of 1997, Duane Walker, KE7BT, put up a 160-meter vertical with three quarter-wavelength elevated radials. Even though the antenna had only been up and in service for a few months, Walker decided to modify the system after reading an article by Dick Weber, K5IU, in the spring 1997 issue of *Communications Quarterly*.[1]

According to Weber's article, verticals using elevated quarter-wavelength radials will have unequal radial currents. When the radial currents aren't equal, the radiation pattern may be skewed and may have significant radiated energy in a higher angle lobe.[2] This is due to horizontally polarized energy radiated from the radial system as a result of the unequal radial currents.

This article shows the steps taken to determine the modifications Duane made to his system. These modifications were made to ensure that the radials would have near equal currents and that the overall system would provide optimal performance. We hope this information will help others who may be planning to make modifications with a similar goal in mind or

who are going to install a new elevated radial vertical system.

Modification background

Weber's article shows radial currents that were measured using the 160-meter elevated radial vertical at W7XU and the two 160-meter elevated radial verticals at WXØB. In addition,
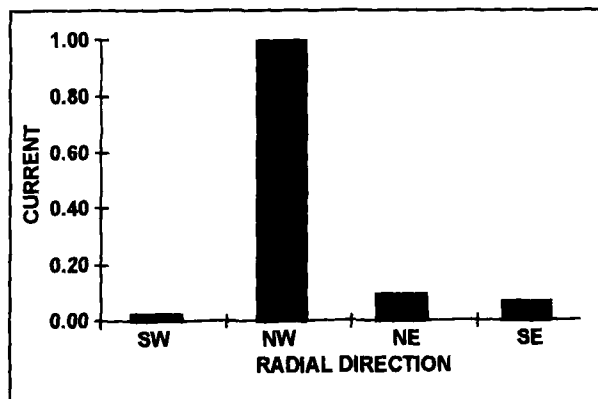
Figure 1. W7XU's radial currents (normalized).

A) All Current in One Radial   **B) Equal Radial Currents**
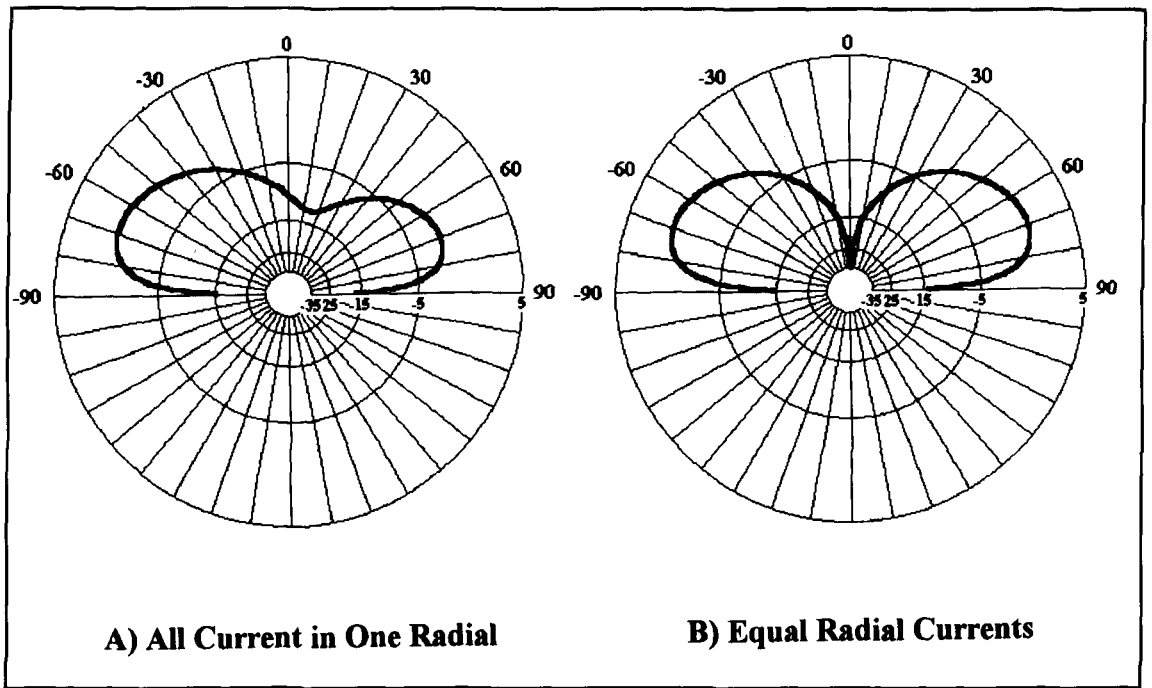
Figure 2. 160-meter patterns: (A) All current in one radial; (B) Equal radial currents.

he provides similar measurements for his 80-meter elevated radial vertical. **Figure 1** gives the measured currents in the four 15-foot-high, 127-foot-long elevated radials used by W7XU along with his quarter wavelength vertical located out in a field with no other towers nearby.[3] It's striking to note how almost all the current is in one radial. With this current partitioning, the radiation pattern shown in **Figure 2A** will result. W7XU had originally set out to have an omnidirectional pattern as in **Figure 2B**, but wound up with something different.

WXØB's radial currents were measured at two frequencies and are shown in **Figure 3**.[4]

Each antenna has two elevated radials. Clearly, these antennas weren't working as originally intended either. The radial currents measured at W7XU and WXØB are very interesting and show that 160-meter verticals using quarter wavelength elevated radials aren't necessarily giving the performance they are capable of providing.

Weber went on to show that an elevated radial system using quarter-wave length radials is highly likely to have unequal radial currents. He attributes this propensity to the sensitivity of current partitioning due to small differences in a radial's impedance when the radial has a
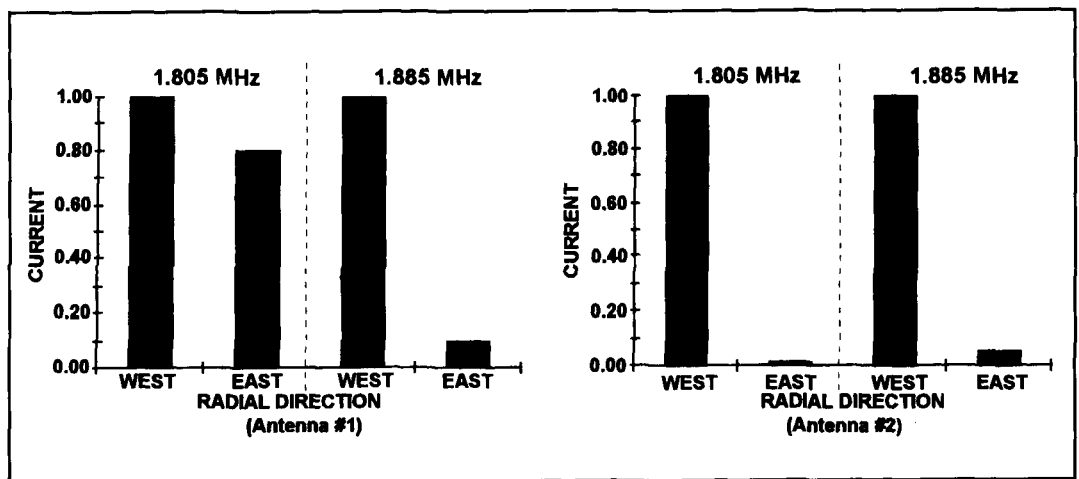
Figure 3. WXØB's radial currents (normalized).

Photo A. 160-meter vertical with three elevated radials.
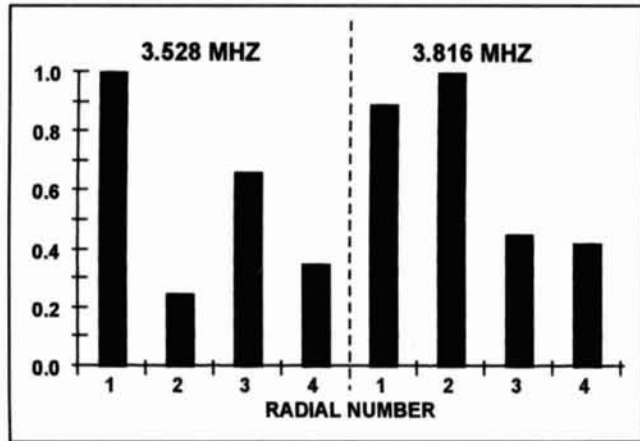
Figure 4. Weber's measured currents in quarter-wavelength radials on 80 meters (normalized).

length near 90 electrical degrees.[5] He also shows that radials longer and shorter than 90 degrees are much less sensitive and will more readily produce near equal radial currents. To demonstrate that this was the case, Weber tested a range of radial lengths with his 80-meter elevated radial vertical, which is in close proximity to a tower. **Figure 4** shows the currents he measured with quarter wavelength radials. **Figure 5** shows radial current measurements taken with approximately 45 and 120 degree long radials.[6] This data is compelling, but Weber had no such data for 160-meter elevated radial antennas.

Before making any modifications to Duane's antenna system, we wanted to see if it had unequal radial currents. Radial current measurements were taken at 1.800 and 1.900 MHz with the existing 130-foot radials using a Palomar PCM-1 snap-on current meter. We found the currents were unequal, as shown in **Figure 6**. In view of this, we mapped out a plan

a) 45 Degree Radials

b) 120 Degree Radials
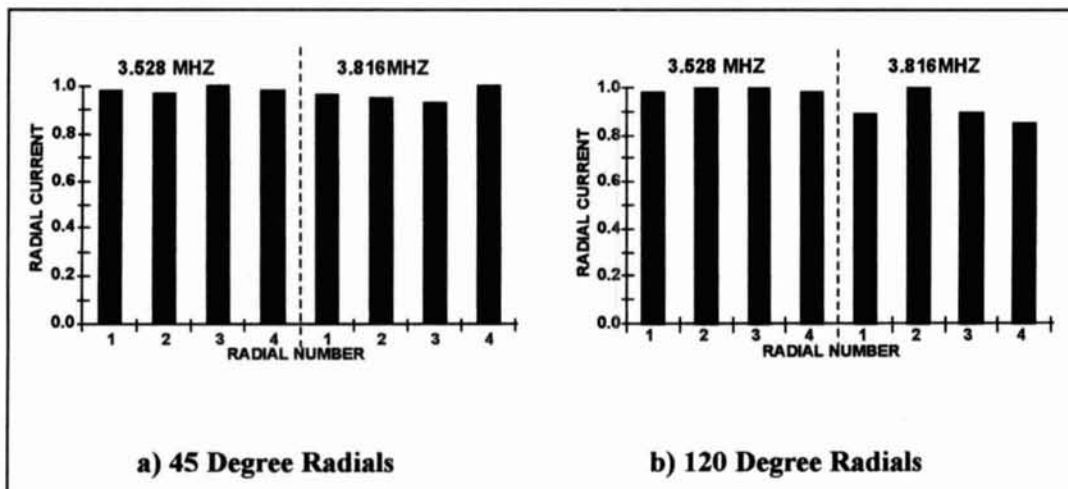
Figure 5. Weber's plots for 45 and 120-degree radials on 80 meters (normalized): (A) 45-degree radials; (B) 120-degree radials.
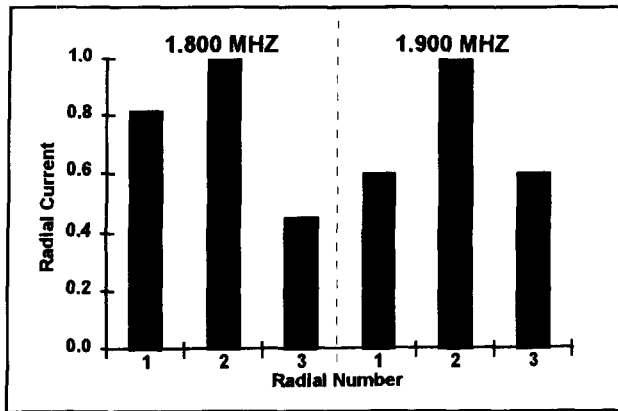
Figure 6. Radial currents in 130-foot-long elevated radials (normalized).

that would lead us to modifications that could be made to his antenna system, and which would yield near equal radial currents with optimal performance.

Antenna description

Duane's 160-meter vertical is shown in **Photo A**. The bottom of the 13-inch-tall Racal-Decca base insulator is 17 feet above ground. One hundred and eighteen feet of Rohn 25G tower sections are used as the vertical element. An aluminum mast at the top adds an additional 10 feet for a total vertical radiator length of 128 feet. Guying is three-way at five levels; the lowest set of guys are pressed into service as elevated radials as shown in **Photo B**.

Due to property restrictions, the elevated guy posts couldn't be set out far enough for the
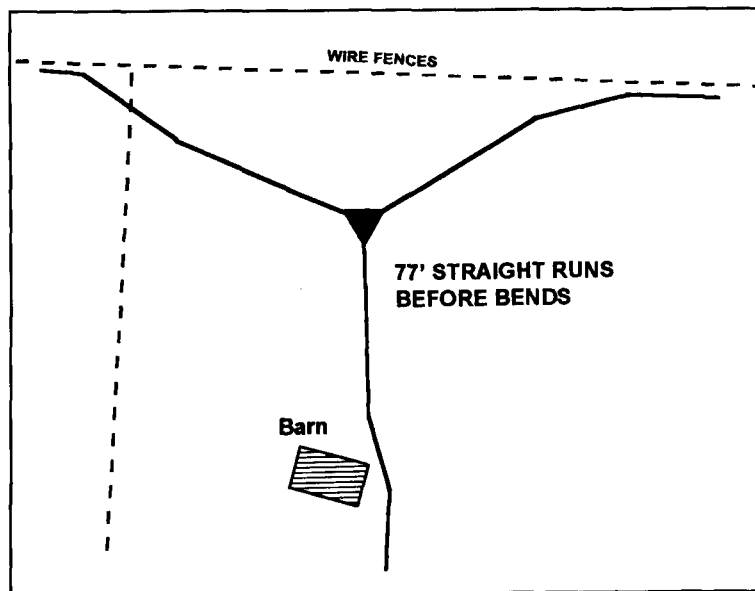
lower set of guys to be a quarter wavelength long. To make each guy 130 feet long, insulated wire pigtails supported by 6-foot-tall fiberglass poles were added to the 77-foot guy wires. The three resulting radials are 17 feet high at the tower end and slope down to 6 feet where the pigtails are connected. Two of the radials are 3 feet above a wire fence on the property line and the remaining radial routes around a barn. Other than two wire fences illustrated in **Figure 7**, the antenna is relatively in the clear with a 70-foot tower about 125 feet away.

Our approach

With the vertical radiator and guys already in place, we were somewhat limited as to which parts of the antenna system could be altered. We decided to focus on modifications to the radial system—still using three radials, while leaving the vertical radiator as is. One other limitation we imposed was that lower guys could not be shorter than 77 feet. Duane didn't want to modify his guys, although a range of longer radial lengths could be created using different length pigtails. Once system limitations were established, we set out to determine the radial length that would have near equal radial currents that also resulted in an optimally performing antenna system.

We defined a three-step plan. The first step was to measure the radial currents with a range of radial lengths, shorter and longer than a quarter-wave. This would tell us if longer and shorter radials had more similar radial currents, and if there were specific lengths that had near equal currents.

Our second step was to run NEC simulations using the same range of radial lengths to generate a graph of gain versus length. From this graph, we would select the radial length resulting in the highest gain antenna system, which testing showed produced near equal radial currents. NEC modeling during this step would use straight radials for simplicity.

Our third step was to observe the effect of using radials of the selected length which were not installed in a straight line, but routed as needed. This would be determined with a NEC model using the actual radial paths. We would compare the predicted performance for the routed and straight radials. If we were convinced the differences were minimal, the radial system would be modified to the selected length.

Executing the plan

Radial current measurements were taken with 77, 98, 112, 140, 154, and 168-foot radials. Radial lengths longer than 77 feet used pigtails

Figure 7. Radials with extensions and nearby fences.

similar to the ones originally used to make the radials 130 feet long. Results of these tests are shown in **Figures 8** through **13**.

The data in **Figures 8** through **13** indicate that radial lengths longer or shorter than a quarter wavelength will have more similar radial currents. From a practical standpoint, we now knew the radial lengths that could be used for this particular 160-meter antenna. The 77-foot radials apparently weren't short enough to yield near equal radial currents. At the same time, radial lengths longer than 140 feet readily produced near equal radial currents. After reviewing this data, we made a decision to use a radial length in the 154- to 168-foot range. To determine if there was a gain advantage for a particular radial length within this range, we moved on to step two of our plan.

The antenna was modeled using NEC2 (NEC-WIN) with radial lengths ranging from 77 to 180 feet. We used this model to predict the gain of the major, lowest elevation angle lobe for each radial length. This model simulated average ground conditions using a ground conductivity of 0.005 Siemens/meter and a relative permitivity of 13. For radial lengths longer than 90 degrees, the radial capacitor used to null the radial set's inductive reactance was assumed to be lossless. With radial lengths less than 90 degrees, the radial inductor used to null the radial set's capacitive reactance was assumed to have a Q of 200. Results of these model runs are shown in **Figure 14**.

From **Figure 14**, we learned that there was 0.08 dB difference between using the radial lengths that produced the maximum theoretical gain and the gain using 154-foot radials. Because equal radial currents were measured using 154-foot radials, we felt confident we could get the predicted performance. On the other hand, testing showed that 130-foot radials had unequal radial currents. This implies the predicted gain of 0.78 dBi can't be obtained. Considering these findings, 154-foot radials were tentatively selected and we moved on to step three.

Was there an effect on the pattern because the radials were not in a straight line as illustrated in **Figure 5**? Did the two fences have an effect? These questions were answered with NEC2 models using average ground conditions and a frequency of 1.85 MHz. **Figure 15** shows two patterns using 154-foot straight radials with the maximum gain of the major lobes labeled. The left pattern is broadside to one of the three radials and the right pattern is for the case where the same radial is viewed from its end. With three radials there is an azimuthal asymmetry of 0.22 dB. The patterns shown in **Figure 16** take into account the bent radials;

Photo B. Radial feedpoint connection.
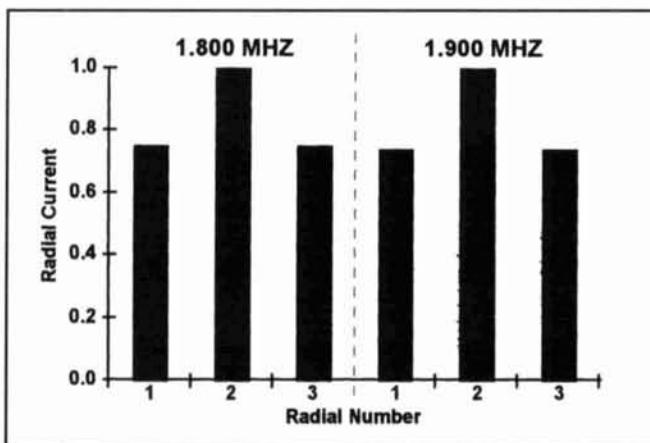
Figure 8. Radial currents in 77-foot-long elevated radials (normalized).
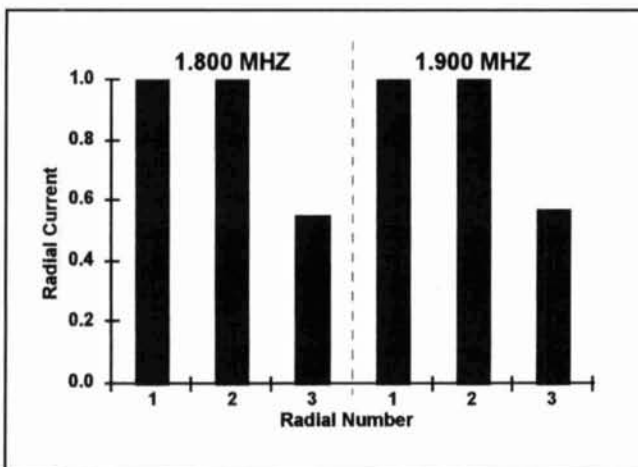
Figure 9. Radial currents in 98-foot-long elevated radials (normalized).
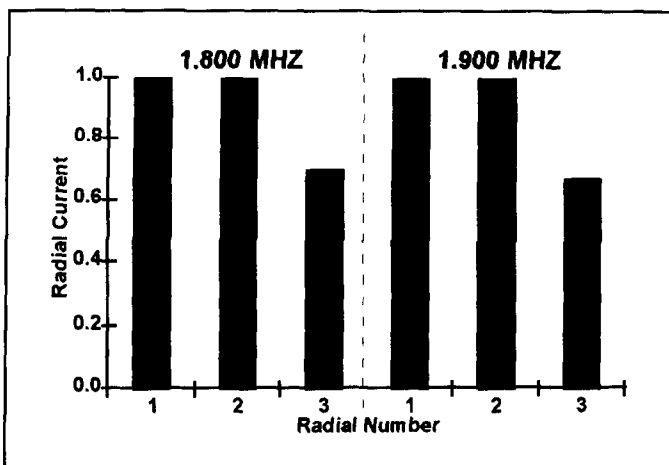
Figure 10. Radial currents in 112-foot-long elevated radials (normalized).
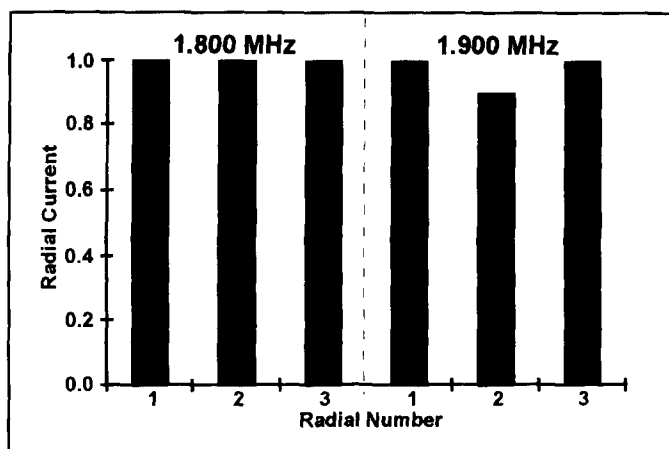
Figure 11. Radial currents in 140-foot-long elevated radials (normalized).
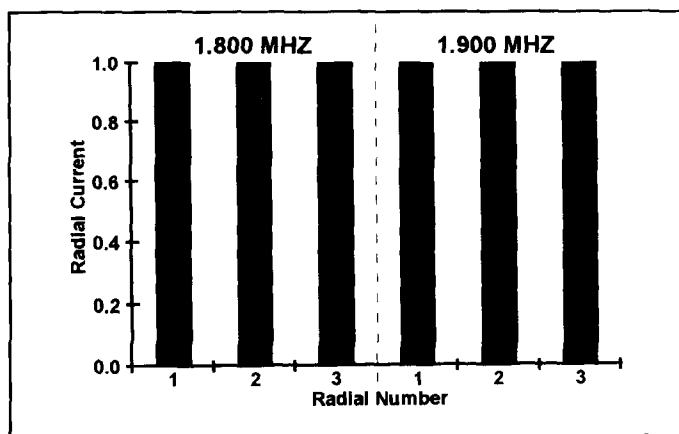
Figure 12. Radial currents in 154-foot-long elevated radials (normalized).

Figure 17 takes into account the bent radials and the fences.

We compared these plots and concluded there was no problem with the radials being bent and near the fences. Pattern asymmetry and gains were essentially the same. We feel

confident making this conclusion because the radial currents with 154-foot radials were measured and found to be equal as illustrated in **Figure 12**. Our conclusion is reinforced by the NEC-predicted radial currents for each case. As expected, the predicted radial currents were equal for straight radials. For both the bent radial configuration and bent radial with fence configuration, the radial currents were within 0.8 percent of each other. With the measured radial currents essentially equal, the actual pattern will be within the range of those bounded by **Figures 15, 16**, and **17**.

After completing our three-step program, we felt sure that changing from 130 to 154-foot radials would yield the best performance this type of system could produce. As a result, the radials were lengthened to 154 feet using pigtail extensions as illustrated in **Figure 5**.

Related issues and comments

You may notice that we referred to radial lengths in terms of feet. We did not use electrical degrees for a reason. As a radial is moved closer to the ground, it must be made shorter to have the same electrical length. The conversion of a radial whose length is in feet to degrees can't be done directly. It is a function of ground conditions and the height above ground. The impact of this is that when a radial is cut to a quarter wavelength in free space, it will actually be electrically longer than 90 degrees when over ground. This is illustrated by the graphs in **Figures 18** and **19**. These graphs show the physical length of a #12 copper wire 90 degrees long at 1.85 and 3.75 MHz, respectively, for various heights above average ground. The graphs were generated using NEC2 over average ground. If we had reported the 130-foot-length radials as 90 degrees, we would have been off. In recognition of this, we did not use degrees as the unit of measure for radial lengths.

In the process of working through our three-step plan, we've learned several things. First, don't use radial lengths near a quarter wavelength. With lengths near 90 degrees, the currents will be unequal. The degree of nonuniformity will depend on the exact installation, ground conditions, and how you originally picked your radial lengths. You should especially avoid the lengths shown in **Figure 18** for 160 meters and the lengths shown in **Figure 19** for 80 meters. An excellent example on 160 meters is shown in **Figure 1** where the radials were 127 feet long. The antenna may appear to work well with unequal radial currents, but it won't perform at its full potential. Second, measure the currents in your elevated radials. If the currents aren't reasonably near equal, con-

sider shortening or lengthening the radials to get near equal currents. Doing so will suppress the radiation of horizontally polarized energy from the radials and provide a more optimally performing vertically polarized antenna.

If you find that your elevated radials don't have near equal radial currents, and the antenna is near a tower, you may have a skewed pattern mainly due to currents flowing in the tower. You may have unintentionally built a parasitic array. In this case, you may want to find a way to choke the tower's currents. Regardless, you should change to lengthened or shortened radials to eliminate horizontally polarized energy from being radiated by the radial system, as illustrated by modeling Weber's 80-meter system.

When identical quarter-wavelength radials were used in a NEC2 model over homogeneous, average ground with no tower present, equal radial currents were predicted and the pattern wasn't skewed. When Weber's tower was put in the model 35 feet from the vertical radiator, the pattern was skewed and the radial currents shown in **Figure 20** were predicted. When the model was changed to use 45-degree radials, the model predicted the currents shown in **Figure 21**. The pattern was still skewed because the tower was acting as a parasitic element, but the horizontally polarized energy being produced by the radials was suppressed. The actual measured currents are shown in **Figures 4** and **5A**. Even in the perfect world of the computer, unequal radial currents were predicted and then made near equal. With length and ground conductivity differences seen in the real world, and possible conductive anomalies in the near vicinity of the antenna system,
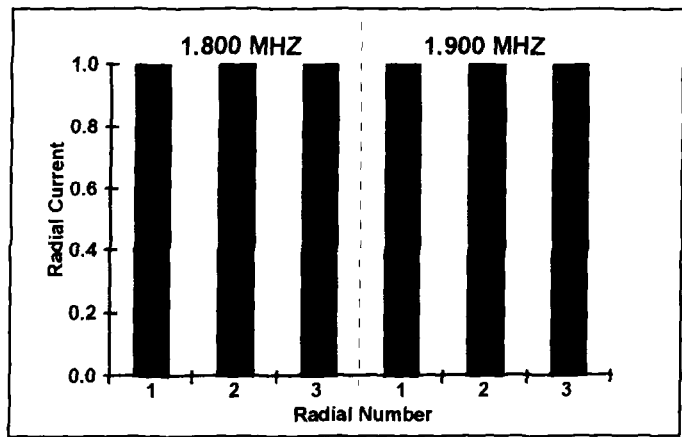
Figure 13. Radial currents in 168-foot-long elevated radials (normalized).
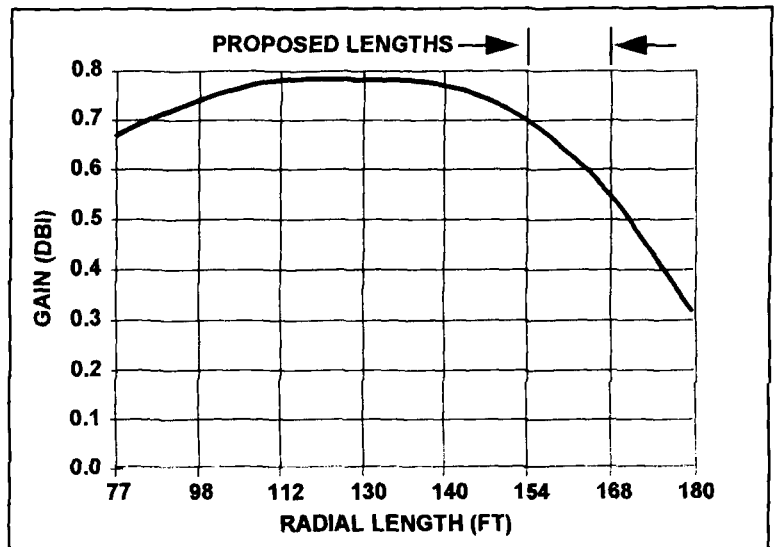
Figure 14. Gain with different length radials.

Broadside to Single Radial (Radial to Right)

Into Radial End
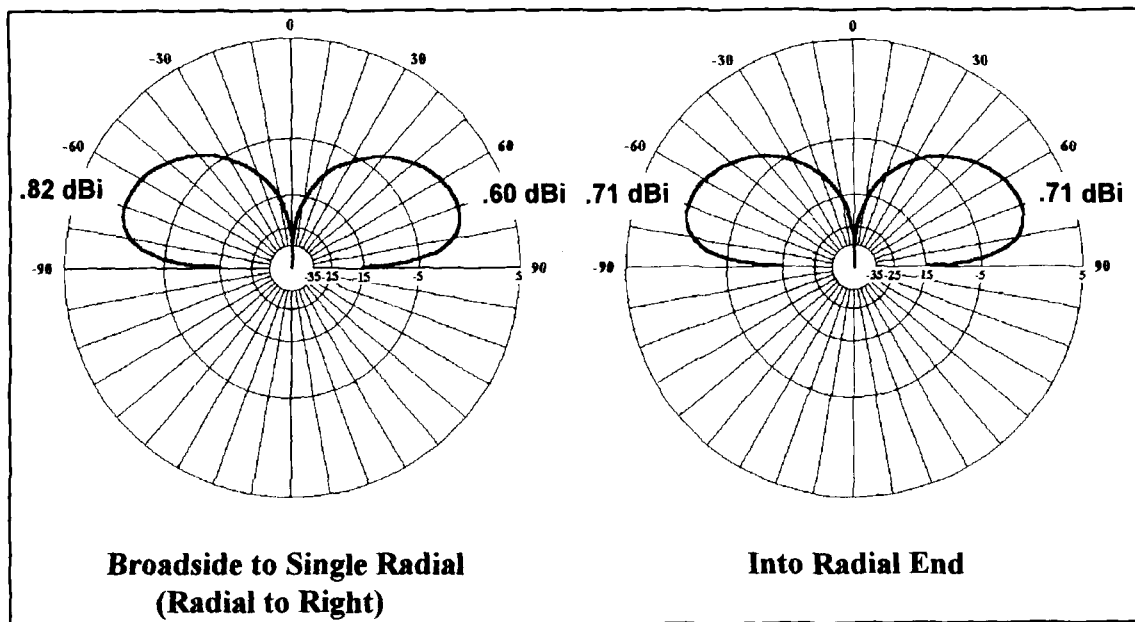
Figure 15. Predicted patterns for 154-foot straight radials.

Broadside to Single Radial
(Radial to Right)

Into Radial End
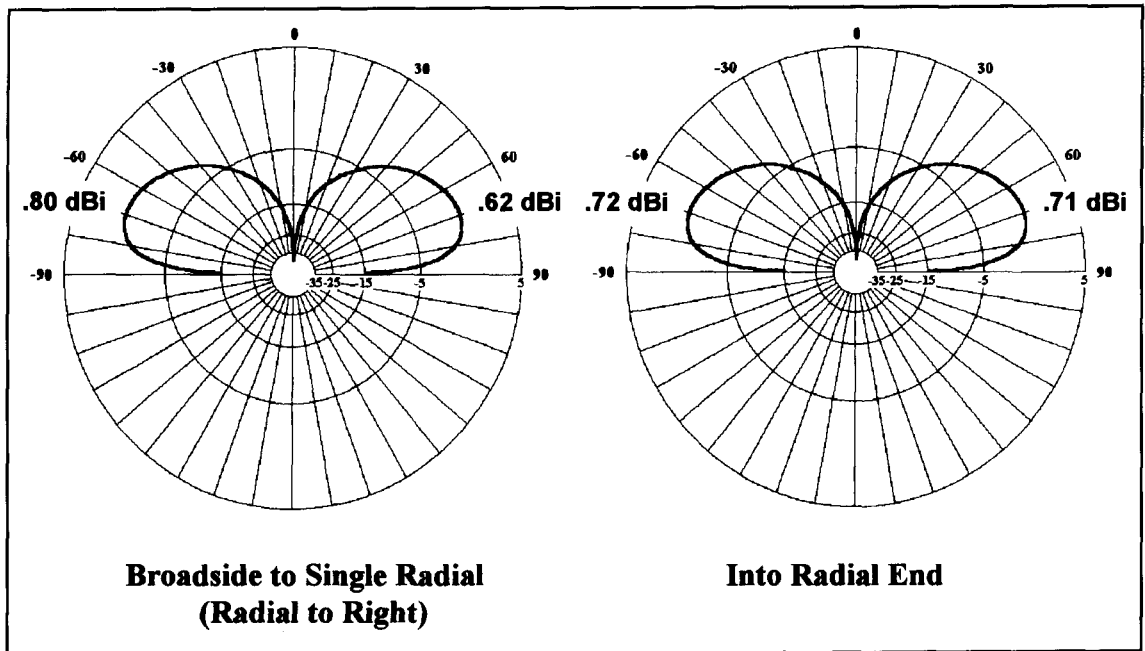
Figure 16. Predicted patterns for 154-foot radials with bends.

Broadside to Single Radial
(Radial to Right)

Into Radial End
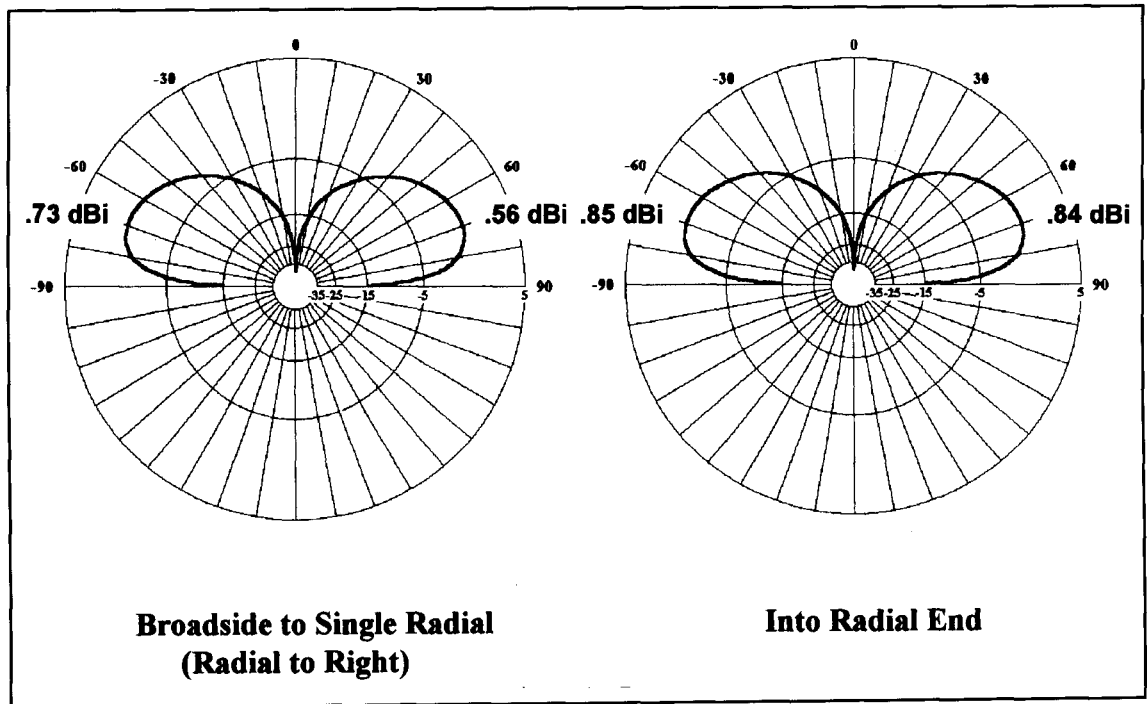
Figure 17. Predicted patterns for 154-foot radials with bends and fences.

you're going to have unequal radial currents when using a radial length near 90 degrees. As examples, you have at one extreme the measured radial currents at W7XU where the antenna is well in the clear. Toward the middle are the currents seen at KE7BT for a system relatively in the clear. Finally, you have K5IU's 80-meter system, which is 35 feet from a tower, at the other extreme.

Was there a noticeable difference in the performance of Duane's antenna after modification? This is a hard question to answer because Duane wasn't able to perform a direct comparison of the original and modified configura-
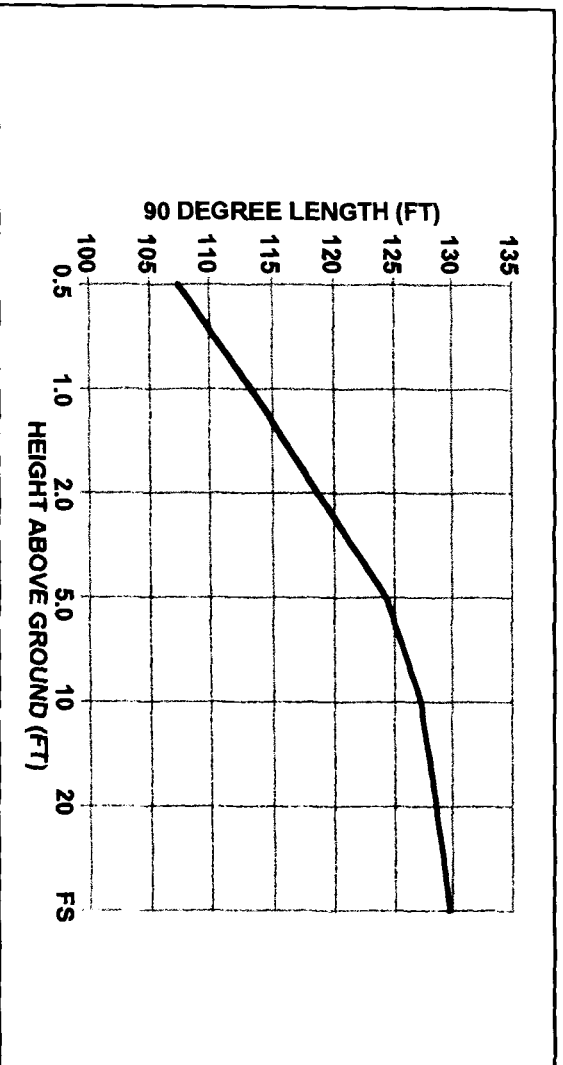
Figure 18. Length of 90-degree radials at 1.85 MHz above average ground.
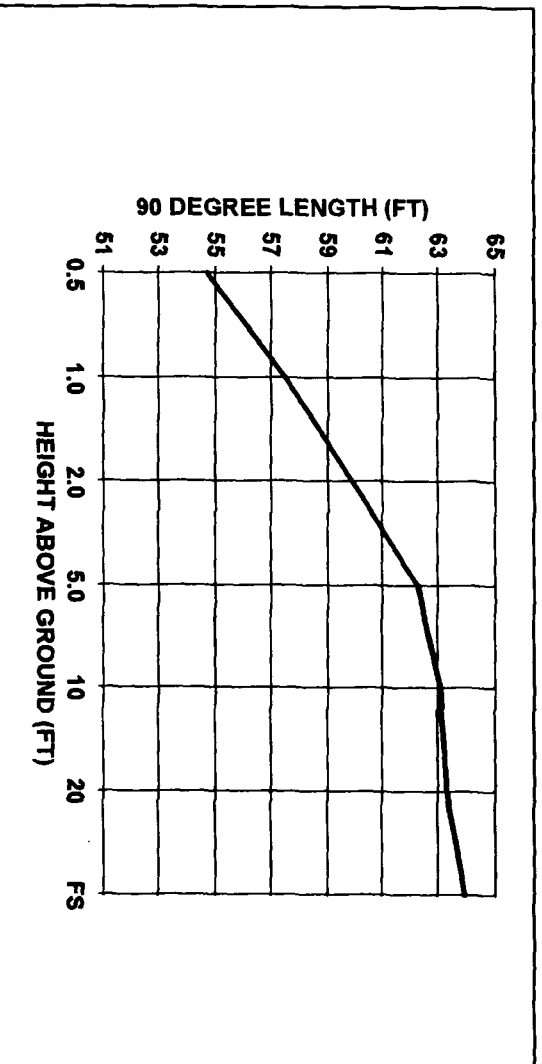
Figure 19. Length of 90-degree radials at 3.75 MHz above average ground.
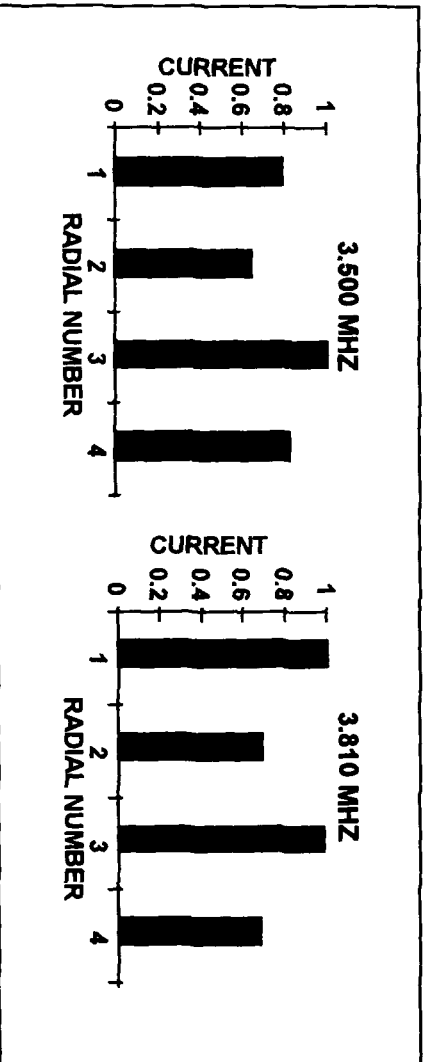
Figure 20. Predicted quarter-wavelength radial currents, near tower (normalized).
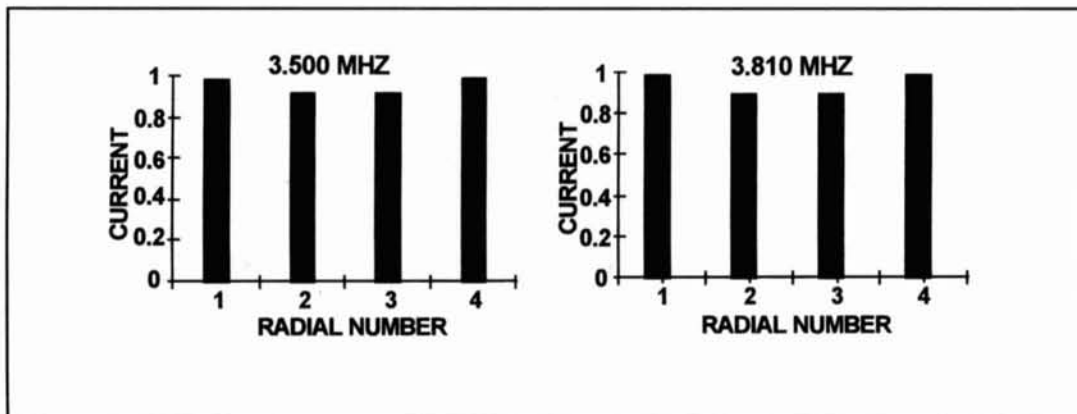
Figure 21. Predicted 45-degree radial currents near tower (normalized).

tions. On-the-air tests seem to indicate there is no noticeable difference. This result was anticipated. Because differences would most likely be in the 1 to 4-dB range, they would be hard to detect.

Was the effort spent on this project worth it? The answer is yes. After investing all the expense and trouble of putting up an antenna of this size, it's rewarding to know that its full potential is being realized. If the original radial currents had been like those measured at W7XU or WXØB, there might very well have been a noticeable difference. If you're using an elevated radial vertical, we recommend you measure your radial currents. If your radials have reasonably near equal currents, your antenna may be providing near optimal performance. Without measuring your radial currents, you won't know. ∎

REFERENCES
1. Dick Weber, K5IU, "Optimal Elevated Radial Vertical Antennas," *Communications Quarterly*, Spring 1997, pages 9–27.
2. Ibid., page 18.
3. Ibid., page 17.
4. Ibid., page 16.
5. Ibid., page 15.
6. Ibid., pages 22–24.

PRODUCT INFORMATION

Hitachi's New Three-CCD Camera Provides High-Quality Digital Video Processing

Hitachi's HV-D15 three-CCD color camera features a single-chip digital VLSI (very large scale integration) which provides both video processing and encoding and can be remotely controlled from a PC. With its high-resolution, low-light imaging and remote control capability, the HV-D15 can handle many applications, such as broadcasting, teleconferencing, remote observation, medical, and microscopy.

Hitachi's 13-bit digital VLSI DSP (digital signal processing) chip provides higher quality video processing than analog. It incorporates multiple functions like detail, color correction, noise reduction, color balance, shading, and auto level control. All signal processing from the processor section through the encoder section is accomplished within the single-chip VLSI. The camera's high signal-to-noise ratio (63dB NTSC, 61dB PAL) and wide range are complimented with 10-bit A/D converters as well as its 13-bit internal digital signal processing. The camera provides both high resolution and high sensitivity. With three half-inch, 410,000-pixel NTSC (470,000-pixels PAL)

CCDs and double-sampled Y channel digital processing, the HV-D15 provides 800 TV lines of resolution. High sensitivity is assured with micro-lens CCD technology. Adding +20-dB high gain and Ultra Gain provides operation down to an illumination level of just 1.01x.

The HV-D15 can also adapt to wide variations in light levels. The micro-processor in the HV-D15 controls auto gain, auto electronic shutter, and auto lens iris to control the video level, even in applications with extreme light level fluctuation. Additionally, an auto tracking white circuit maintains proper color balance changing light levels.

Multiple HV-D15s can be remotely controlled from a PC simultaneously via an RS-232C interface. The camera data can also be transferred to a PC for storage and later recall.

For further flexibility, three application files within the HV-D15 can store user-selected set up information. Switching between application files results in each menu item being reset according to the information previously stored in the file.

For more information on the HV-D15, call (516) 921-7200.

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E-mail: <DrBingo@compuserve.com>

PHASED ARRAY ADJUSTMENT

For amateur radio antennas

This article describes a lumped-parameter tee-network model of the coupling between radiating elements in a phased array, which provides a physical and intuitive understanding of phased array operation and adjustment interactions. Using this model, I will analyze the performance of a practical phased array antenna that appeared in a *QST* magazine construction article.[1]

The technique of connecting mismatched feeders to the driven elements in a phased array is an expedient method for obtaining relative current phases in elements that approximate the desired currents. I will discuss a simple way to provide instant pattern adjustability with a variable capacitor later in the article. This practical technique helps compensate for the coarse method of using mismatched transmission lines to provide phasing.

Basic theory

A phased array is an antenna having more than one driven element. The desired radiation pattern is obtained by adjusting the current magnitude and phase in each element. Ideally, you want to be able to adjust the current magnitude and phase in each element *independently* and obtain a perfect impedance match to the transmission lines carrying power to each element. It is nearly impossible to do this by fanning out different length transmission lines from the common feedpoint to the driven elements, but a compromise can be reached.

An antenna design is defined by a set of desired performance features. This set may include gain, beamwidth, departure angle, polarization, front-to-back ratio, front-to-side ratio, impedance and pattern bandwidths, component stress (current, voltage, compression, tension, shear), overall size and weight, and feedpoint impedance. Because many of the de-

sired features interact, and some may be improved only at the expense of others, priorities must be assigned to these performance criteria.

To efficiently optimize your design, it helps to understand how antenna performance features interact on both the quantitative and qualitative levels. For example, it is generally true that there's a tradeoff between gain and bandwidth. A slight improvement in gain may cost you a major degradation in bandwidth. Therefore, you must find a balance, and this depends on the relative importance of each of the antenna's performance criteria. Obviously, if you're operating CW at a single frequency, you don't care much about bandwidth. If you're only operating in the SSB portion of the 20-meter band, you only care about antenna performance in the upper half of the band, and so on.

Optimizing algorithms have been around for a long time, and can readily be applied to antenna design problems. A formula is written to create an error sum based on each of the
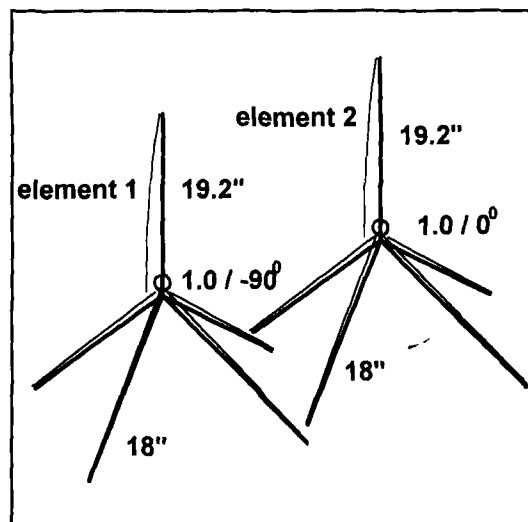
Figure 1. Two quarter-wave elements.
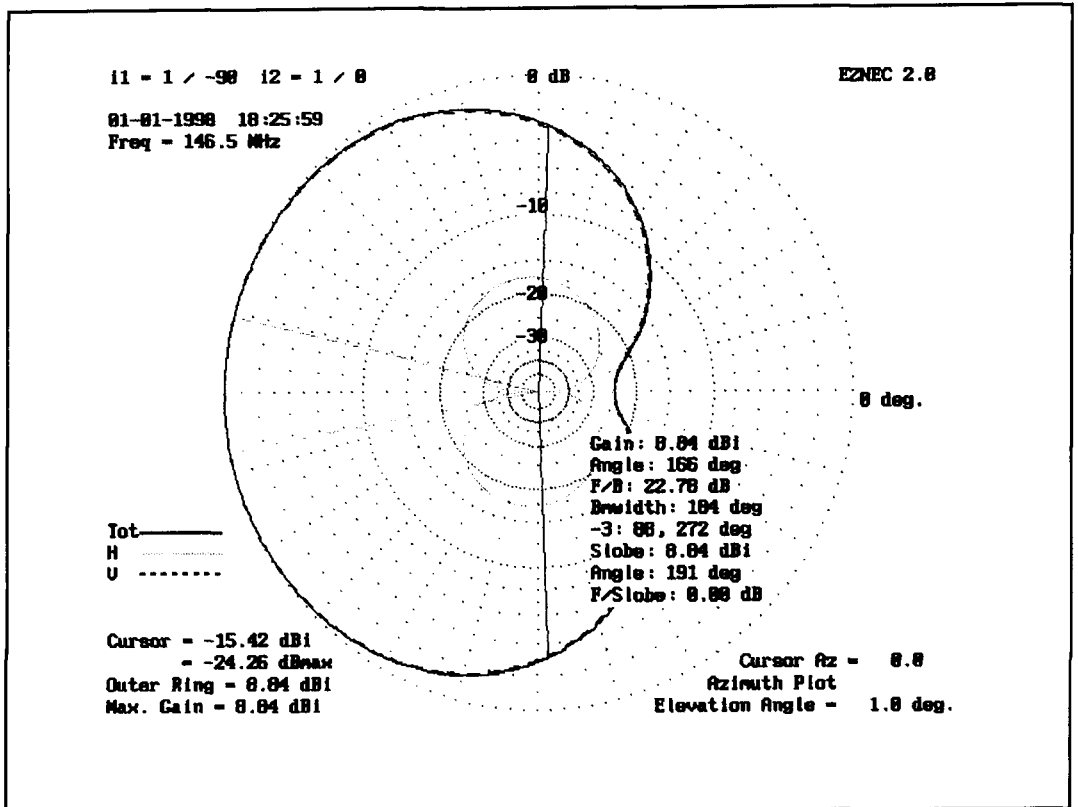
Figure 2A. 90-degree phasing.
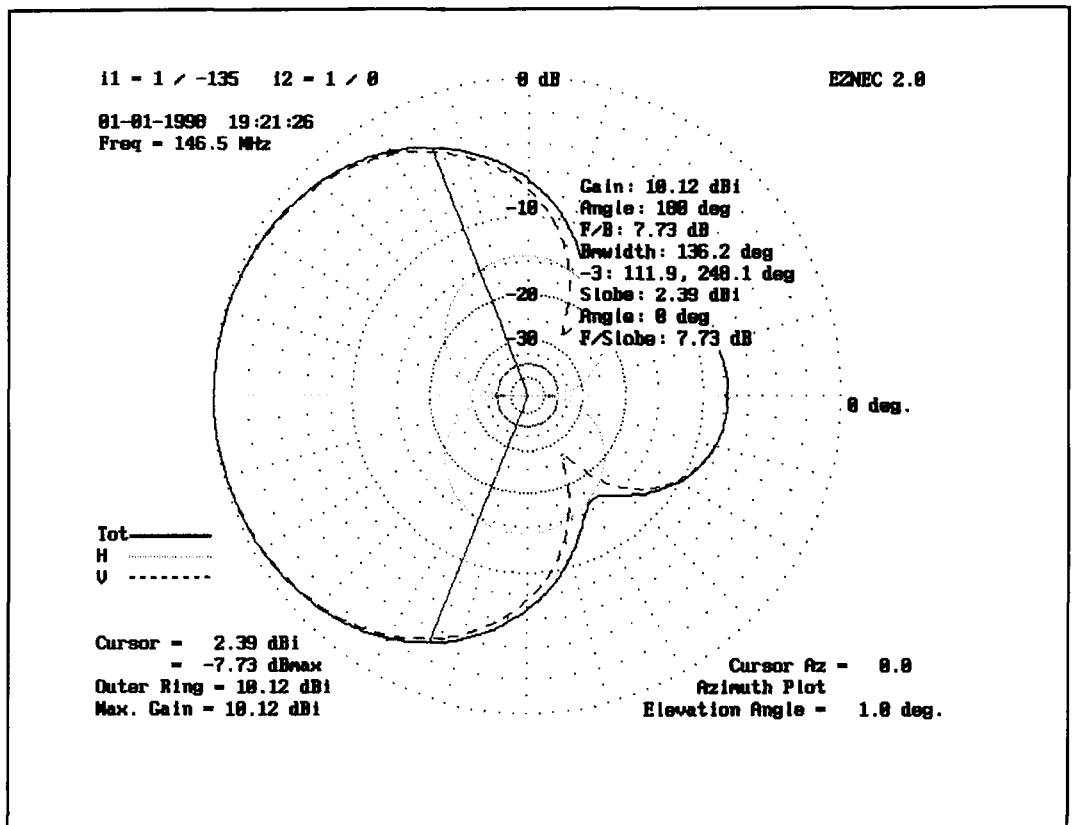
Figure 2B. 135-degree phasing.

weighted antenna performance criteria. Various parameters in the antenna are adjusted until a minimum error is found. This iterative process is both an art and a science, and it has a jargon all its own. There are brute-force methods that may take a long time and never reach the best solution, and there are "intelligent" algorithms that use advanced mathematical concepts and come together quickly.

Because the outcome of such optimization depends on the number of parameters that can be varied, the more the better (usually). Typical high-gain amateur radio antennas consist of one driven element and many parasitic elements. The input parameters to a typical parasitic antenna design would include element length, diameter and spacing, height above ground, and perhaps some transmission lines and other discrete components. If you also introduce multiple RF sources—adding both the source current magnitudes and phases to the adjustable parameter list—then you increase the possibility of finding a better compromise between all the desired performance features. In other words, for a given field intensity, you may get twice the bandwidth from an array of two driven elements compared to an array of one driven and one parasitic element. Or perhaps, you might obtain a certain gain from a driven array that requires only half the volume of a conventional parasitic array with the same gain.

Let's explore the physical interactions and rules-of-thumb that apply when designing an antenna with more than one driven element. Admittedly, the complexity of the problem increases exponentially as one adds more knobs to the "box," but a human operator will eventually develop a feel for the problem by simply turning the cranks and observing what happens. Since the 1960s, the "box" has often consisted of a method-of-moments antenna analysis program.

Some general rules apply, and if you read this article carefully, you will find some useful shortcuts to the antenna design process. Rule number one: Write everything down in a notebook so you don't end up repeating yourself. Rule number two: Enter enough detail so you can verify your results later. Rule number three: Define a simple set of goals and try to stick with them.

A starting point

The behavior of a single isolated antenna element, such as a dipole in free space, is a simplified or reduced model of an actual antenna. In reality, the behavior of a dipole is influenced by the presence of the Earth, weather, support structures, transmission line, insula-
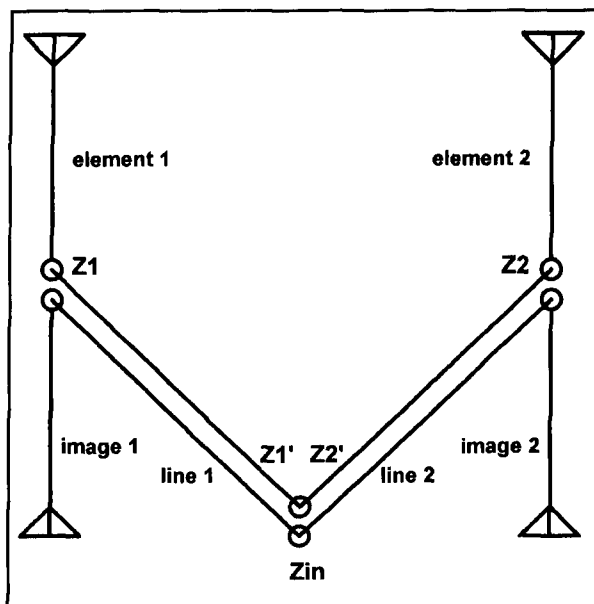
Figure 3. Transmission line "feeders."

tors, etc. Often these influences are minor, but when an additional radiating element is deliberately added to an antenna, major changes occur in the first element. When both elements are physically aligned and are within a wavelength or so of each other, they are said to be strongly coupled.

Everything can loosely be tied together by observing the current in the antenna elements and remembering Ohm's Law. The electric field intensity produced by an antenna is directly proportional to the currents in that antenna. For a given power input to an antenna, high-gain multi-element wire antennas have relatively high currents compared to single-element antennas of comparable dimensions. This higher gain is manifested as a decrease in the input resistance to the antenna.

For example, assume you have a 144-MHz half-wave dipole fed at the center. You measure the input impedance and find it to be
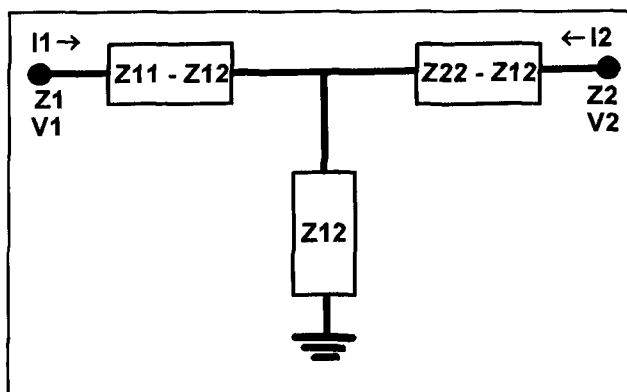
Figure 4. Phased array impedance; tee-network model.
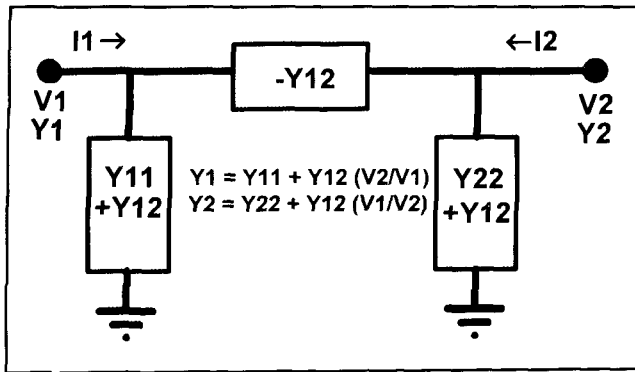
Figure 5. Phased array impedance; pi-network model.

about 70 + j0 ohms. Then you add a second
antenna element consisting of a half-wave-
length of wire in parallel with your driven ele-
ment. As you move this wire closer to your
dipole, you'll see the input resistance to the
dipole drop. For a spacing of 30 centimeters,
the input impedance becomes about 25 + j25
ohms, and the field intensity increases by about
3.6 dB in one direction compared to the dipole
alone and decreases in the opposite direction.
The parasitic wire acts as a reflector at this
spacing. At much closer spacing, it can act as a
director and produce even more gain, albeit in
the opposite direction.

For a spacing of 10 centimeters, the input
impedance is about 4 - j20 ohms, and the gain
is about 4.0 dB. However, because the currents
are much higher and the input resistance much
lower for this close spacing, you can lose quite
a bit of your input power in the form of I_2R
losses, depending on the resistance of the wires.
I assumed copper for the gain figures cited; alu-
minum would yield a gain of perhaps 3.9 dB;
and steel would just about eat up all of the gain
for the 10 centimeters spacing.

Never assume that maximum current means
maximum gain. If you were to narrow the spac-
ing between the two antenna elements so the
second element shorted out the driven element,
the input resistance would approach zero and
the current would get very high, but the field
would approach a theoretical minimum.
Remember, real wires and real insulators have
losses, and the higher the currents in these
wires, and the higher the voltages across these
insulators, the more RF power will be lost as
heat. Worse yet, you may encounter voltage
breakdown, especially under wet conditions
during modulation peaks.

Also, the field produced by a wire depends
on the current distribution along that wire, and
it's not just the maximum current magnitude
that defines the electric field intensity. In fact,
if you have equal magnitude but opposite phase
currents on a wire, your total field at some
locations can approach zero, no matter how
large your currents are. The field at any point in
space is the sum of the fields from the currents
in all the antenna elements, delayed by the indi-
vidual times each field takes to reach that point.
Because the current is not uniform over the
length of each wire element, it's useful to apply
a little integral calculus to determine the area
under the current distribution curve to get an
accurate handle on the electric field produced
by an antenna.

This is one reason why computer programs
are so useful for antenna analysis. They can
integrate by brute force. Antennas may have
very complex shapes and may operate in very
cluttered environments, so it is nice to have a
computer program that can incorporate many of
the physical details—reducing the number of
assumptions and simplifications that must be
made to create a practical model.

In the world of antenna design, gain is traded

Table 1. Effects of Adjusting Element Currents

Input Impedance at bases		Complex current at bases	
Z1(ohms)	Z2 (ohms)	i1 (amps)	i2 (amps)
75.5+j35.7	23.1-j15.8	1.00/-90°	1.00/0°
49.6+j46.7	12.5+j10.3	1.27/-135°	1.27/0°
55.1+j36.2	0+j0	1.35/-123°	0.98/0°
53.0+j26.8	-27.5-j5.7	1.46/-123°	0.69/0°

Watts		Gain	
P1	P2	dBi	dBf
77	23	8.8	22.8
80	20	10.1	7.7
100	0	9.6	8.7
113	-13	8.9	5.6

for just about every other performance feature. Higher gain typically is paid for with an increase in one or more of the following: antenna size, weight, windage, current, voltage, bandwidth, feed complexity, difficulty of impedance matching, importance of manufacturing tolerances, etc. At one extreme, if you want your antenna to present a resonant 50 ohms across several octaves, you might be better off building a dummy load out of noninductive resistors! At the other extreme, you may end up with an antenna Q so high that, when you try to modulate, the reflected power at the sidebands knocks you off the air.

Returning to the analysis of the *QST* construction article,[1] consider the case of two quarter-wave elements cut for the 2-meter band and spaced a quarter wavelength, each having four ground radials bent downward at a 45 degree angle (**Figure 1**). If the phase of the current in one element is 90 degrees out of phase with the other, the fields radiated from these elements are in phase in one direction and out of phase in the opposite direction. That is, the signal reinforces in the beam direction and cancels out the back. If the current magnitudes are equal, the null out the back goes to zero.

In **Figure 1**, if the left element's current phase is -90 degrees relative to the right element's phase, then by the time the left element's electromagnetic wave arrives at the right element (a quarter wavelength distant), its overall delay is 180 degrees. The signal from the left element cancels that from the right element, and the pattern develops a null to the right. The opposite occurs to the left bearing (**Figure 2A**). For other relative phases the pattern may have higher gain, but a secondary lobe may appear out the back, compromising the front-to-back ratio. Phasing of 135 degrees maximizes forward gain for the given quarter-wavelength spacing (**Figure 2B**).

Physical particulars

The two vertical antenna elements are 19.2 inches long; all calculations were done at 146.5 MHz. The elements are spaced 19.2 inches, and the image-plane radials are 18 inches long.[1] The ground is 30 feet below the element feedpoints. The ground conductivity is 5 mS/m and the relative dielectric constant is 15. All wire diameters are 1/4 inch. The transmission lines are lossless RG-8, which has a surge impedance of 52 ohms and a velocity factor of 66 percent. At 146.5 MHz a wavelength in free space is about 80 inches, but it is only 53 inches in RG-8. By convention, we assign 360 degrees of phase shift to a wavelength, so a quarter wavelength is 90 degrees. A piece of RG-8 physically a quarter wavelength long (in
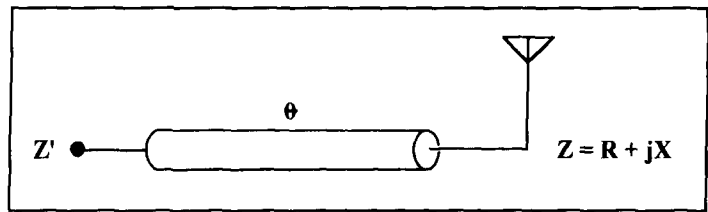
Figure 6. Theta is 90 degrees for a quarter-wavelength line. Phase shift phi associated with this line is -90 degrees for the matched condition.

this case, 20 inches) actually has a phase shift of 90/0.66 = 136 degrees, but only if the VSWR is 1.0 (more about this later). The outer conductors of the RG-8 were included in the moment-method antenna model, and formed a Vee shape beneath the ground radials. This affected the pattern only slightly, perhaps 0.2 dB. The gain figures cited here are for an elevation angle of one degree above the horizon. See the **Appendix** for a typical vertical pattern.

Phase shift sign conventions

There's some confusion about the sign of the phase shift across a transmission line. Is it 20 inches of RG-8 minus 136 degrees or plus 136 degrees? In the commercial broadcast industry, most engineers assign a minus sign to the phase shift across a transmission line. This is in keeping with the concept of "delay." That is, the arrival of the RF energy at the end of the line is delayed by 136 degrees in this case. And, if that same energy were traveling through 20 inches of space, it would only be delayed by 90 degrees. If you were to model this 20-inch length of transmission line as a tee-network, it would have a capacitor in its shunt leg, and identical input and output inductors in its series arms. If you chose a pi-network model, it would have identical capacitors in the shunt input and output arms, and a coil in the series arm.

I think some of the confusion about the sign convention of phase shift stems from "Circuits 101," where a current entering an inductor is said to lag the voltage applied to that inductor by 90 degrees, and the current entering a capacitor is said to lead the voltage across that capacitor by 90 degrees. The current in the inductor is defined as V/jX, so its phase shift is -90 degrees relative to the applied voltage, or -jV/X. I suppose it all goes back to the way waveforms are plotted against a time axis. If time increases from left to right on the graph, the inductor current sinusoid starts 90 degrees later, which is to the right of the voltage sinusoid. And, in a Cartesian coordinate system, one typically assigns positive values to the right of the vertical axis, and negative values to the left—which tends to contradict the value of -90 degrees just defined on the time axis. Again,
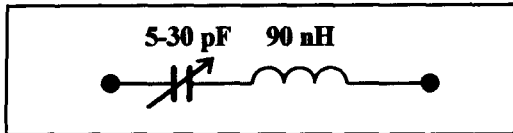
Figure 7. In the 2-meter band, a 5–30 pF varicap in series with a 90-nH air-core coil will work.

this is just convention, and if you're consistent with whatever convention you choose, your results will be okay. Above all, remember that you're interested in comparing one current with another current, not with a voltage.

The outer ring of a Smith chart contains a legend stating which direction is "towards the generator." When you move in this direction (clockwise) on the Smith chart, you're moving away from the load, which, in the case of the feeder lines, is away from the base of the antenna elements. To check your math, make a rough plot of your impedance results on the Smith chart to see if everything makes sense. Note that the phase shift around the complete circumference of a Smith chart is 180 degrees, not 360.

So, how does one obtain equal current magnitudes and a relative phase of 90 to 135 degrees in the two antenna elements? Perhaps the first approach that comes to mind is to feed the two elements with different lengths of transmission line, one a quarter wavelength longer than the other (**Figure 3**). However, on closer inspection, this technique is flawed because it assumes the transmission lines are terminated in their characteristic impedance. You see, the current phase shift through a transmission line depends on its load impedance. And the feedpoint "operating" impedances of the two elements in the array are not equal to each other, nor are they equal to their individual or "self"-impedances. This is due to the electromagnetic coupling between the two elements. They are close enough physically to influence each other significantly. So every time you change the length of one transmission line, the mismatch on *both* lines changes, the relative element currents change, and so on. In fact, just about everything interacts in a multi-element antenna. You can't make a change in one element without influencing the others.

You can get a handle on the amount of coupling and its exact impedance effects by using the tee-network model in **Figure 4**. Think of it as a transformer model, where the self-inductance of the primary and the secondary are the same (the two antenna elements are the same), and the mutual inductance between transformer windings is an analog of the coupling between the two elements in the array. However, in the world of antennas, the resistance is more significant, so we refer to self- and mutual *imped-*

ance rather than inductance. Of course, the mutual impedance between antenna elements can also be capacitive.

Where:

Z11 = self impedance of element 1
Z22 = self impedance of element 2
Z12 = mutual impedance between both elements
Z1 = operating input impedance to element 1
Z2 = operating input impedance to element 2

In this case Z11 = Z22 because the array is bilaterally symmetrical.

If you prefer, you can model the coupling between elements as a pi-network equivalent (**Figure 5**) of the tee-network just defined. This might provide some additional insight into the workings of a phased array. In this case, Z1 = 1/Y1, Z11 = 1/Y11, etc. The port input voltages and currents are the same in **Figures 4** and **5**. The pi-network has the advantage of having one less "node" than the tee-network, making for a more efficient computer analysis model. I tend to think in series R and X, rather than shunt G and B.

Keep in mind that these tee- and pi-networks aren't the same as the standard impedance matching networks used in antenna tuners, as those have only one driven port. Also, the impedances seen looking into these antenna models aren't resonant, whereas matching networks typically are. So, you can't make too many safe inferences about antenna behavior based on normal antenna match behavior.

You can measure the *self*-impedance of an antenna element by connecting an impedance meter to the feedpoint of one of the elements when the other element's feedpoint has nothing connected to it. You must do this in the presence of the other element, because the self-impedance of a single element alone in free space won't be the same. You can then determine the *mutual* impedance by measuring the feedpoint impedance when the second element's feedpoint is short circuited (in this special case, Z1 is called Zsc), and plugging that value into **Equation 1**:

$$Z12 = \sqrt{Z22\,(Z11 - Zsc)} \qquad (1)$$

Alternatively, you can terminate the base of element 2 in an open circuit and measure the complex voltage across this termination, V2. You can also measure the complex input current at the base of element 1, I1. Then you can use Equation 2 to determine the mutual impedance. This isn't too practical because most of us don't have access to a vector voltmeter. However, it's something you can easily model with a moment-method antenna analysis program.

$$Z12 = V2\,/\,I1 \qquad (2)$$

The closer the antenna elements are to each other, the higher the value of mutual impedance, and the more Zsc will differ from Z11. If the elements were very far apart, Z12 would be zero and Zsc would be equal to Z11. Using **Equations 3** and **4**, you can determine the actual base operating impedances. In this case, they are very different from the self-impedances of the two elements, and the total RF power input to the antenna only rarely splits evenly between the two elements in a phased array. Remember that the operating resistance and power of an individual element can be zero, and can also be negative. However, this is never true at the overall input to the array.

$$Z1 = Z11 + Z12 \ (i2/i1) \qquad (3)$$

$$Z2 = Z22 + Z12 \ (i1/i2) \qquad (4)$$

Where:

i1 = complex base current in element 1
i2 = complex base current in element 2

The general form of the operating impedance equation for n antenna elements is:

$$Zi = \Sigma \ (Ij/Ii) \ Zij, \text{ for } j = 1 \text{ to } n$$

Math example

The self-impedance of these elements is 49.2 + j10.0 ohms, per NEC2, when all wire diameters are 1/4 inch and 30 feet above 5 mS/m earth. The self-impedance is determined by placing a current source at the feedpoint of the first element and an open-circuit load at the feedpoint of the second element. (I used a 10-k resistor.) The pattern is essentially omni-directional and the gain is 1.4 dBi when the second element is left "floating" this way.

However, it's important to realize that self-impedance isn't determined by detuning the other elements in the array to obtain an omni-directional pattern. With electrically taller elements, say a half wavelength, you may have to short-circuit the parasitic element's feedpoint to

obtain an omni pattern, and this would yield very inaccurate self-impedance values for the driven element. So just ignore the pattern shape you get when "floating" the other elements in the array. All you care about is the impedance of the driven element obtained with this test, which we define as self-impedance (Z1 becomes Z11 when there is no current in the output arm of our **Figure 4** tee-network model).

Using **Equation 2**, mutual impedance Z12 is equal to the voltage across 10-k resistor V2 divided by the input current, I1. If you use a source current of 1 amp at zero degrees, and NEC to determine the current through the 10-k resistor to find V2, be careful you don't get the sign wrong, or your value for Z12 will be off by 180 degrees. This can happen because, in our coupling model of **Figure 4**, I2 is going into the port; but in the NEC open-circuit model, I2 is coming out of the port. A vector voltmeter would verify this, but as I said, most of us don't own one. It's always a good idea to compare your measured or calculated mutual impedance with textbook values to be sure you haven't made an error somewhere.

Most of us can get our hands on an inexpensive impedance measuring device, like those advertised in the amateur radio magazines. When the open circuit at the base of element 2 is replaced with a short circuit, Z1 = Zsc = 55 + j36.2 ohms. The mutual impedance is then calculated from equation 1, thus Z12 = 36.7/-45.6° = 25.7 + j26.2 ohms. By the way, when the second element is shorted out, it acts as a reflector—which produces a gain of 4.7 dBi. This isn't bad compared to the 1.4 dBi value obtained earlier for the omni-directional pattern.

If you drive both elements, you can obtain a maximum gain of 5.2 dBi when the currents are equal in magnitude and phased 135 degrees. If you phase them 90 degrees, you get a better front-to-back ratio, but less forward gain (3.9 dBi). Of course, you can obtain pretty good gain by feeding one element and tuning the other as a parasitic reflector. But if you want the sharpest possible null out the back, you can-

Table 2. Effects of Feeding with Mismatched Lines

Case	Input impedance at bases		Current at bases	
	Z1 (ohms)	Z2 (ohms)	i1 (amps)	i2 (amps)
1	73.0+j41.4	30.4-j18.3	0.91/-91°	1.11/0°
2	32.1+j41.7	15.6+j8.8	1.40/-150°	1.54/0°

Case	Power (watts)		Gain	
	P1	P2	dBi	dBf
1	61	39	9.2	22.3
2	63	37	10.9	5.5

Table 3. Mismatched Feeder Lines

line 1	Input 1 Z1'(ohms)	VSWR	line 2	Input 2 Z2'(ohms)	VSWR	Combined Zin (ohms)	VSWR
23 in	39+j31	2.08	17 in	104+j4	2.00	31.6+j15.2	1.85
23	20+j18	2.96	3	108+j6	2.09	18.5+j12.7	3.00

not obtain it unless you independently control the phase and magnitude of the currents. And when the QRM originates from the back lobe of your antenna's pattern, you might wish for a better front-to-back ratio, regardless of forward gain. A second, perhaps more important, reason for having a simple method to adjust the relative current magnitudes and phases is to compensate for stray capacitance and inductance within the feeder system and the antenna, undesired radiation from the feeder lines, construction irregularities, and so on.

Table 1 describes the differences for various element currents when 100 watts of power is delivered to the array. Note that dBi is the maximum gain of the pattern relative to an isotropic radiator, and dBf is the front-to-back ratio of the pattern. The parasitic case (where the second element isn't driven and has no insulator between it and its counterpoise radials) looks attractive. An element is parasitic when it has current but no power; i.e., when its operating resistance is zero, or it is short-circuited.

Negative resistance

You may wonder if it's possible to create a situation where one of the operating resistances is actually negative. In this case the power in one element would be negative, and the power in the other element would actually be greater than the total input power to the antenna. This

may sound strange, but it does happen. The sum of all the element powers must equal the total input power to the antenna. The negative element passes power back down its transmission line and the positive element passes some of its power to the negative element.

To see this in the simple array, all you need to do is change the relative current. You know the parasitic case has zero resistance at the base of the reflector, so start from there by reducing the magnitude of the element 2 current while maintaining the same phase relative to element 1. The pattern obtained with this particular negative resistance case isn't that great, and the bandwidth is likely to be narrower compared to a comparable pattern obtained from an array having all positive operating resistances. The negative resistance implies an increased circulating current. But there are some instances where a very sharp null is desired in the pattern, and a negative resistance element is the only way to obtain it. The correct method of returning the negative power to the common feedpoint in phase is widely misunderstood, but isn't something we need to discuss in detail at this time.[2]

Feedlines

How can you feed this array with the desired complex currents (i1 = 1.0/-90° and i2 = 1.0/0°)? Start with a pair of transmission lines

Table 4. Pattern Adjustment
with Series Coil and Capacitor at Element 1

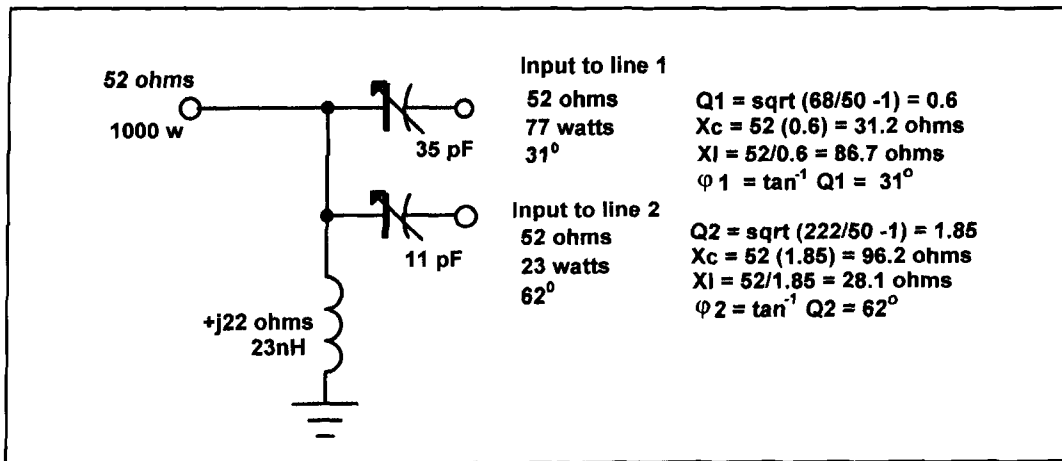| #1 Tuning Reactance	Zin (ohms)	Current at bases i1 (amps)	i2 (amps)	dBi	dBf
-30 ohms	35+j11	0.6/-56°	1.10/0°	7.9	4.5
-20	34+j12	0.65/-65	1.08/0	8.2	6.7
-10	33+j13	0.80/-76	1.08/0	8.6	10.9
0	32+j15	0.91/-91	1.11/0	9.2	22.3
10	31+j18	1.02/-107	1.21/0	9.9	19.6
20	31+j21	1.08/-123	1.36/0	10.4	12.0
30	33+j25	1.08/-137	1.53/0	10.6	8.1
40	36+j27	1.02/-146	1.65/0	10.4	5.7

Figure 8. Power divider.

as in **Figure 3**, and consider two cases. In both cases, 23 inches of RG-8 is connected to element 1. In the first case, line 2 (connected to element 2) is 17 inches long, but in the second case, it's 3 inches long. In both cases, the RF current takes longer to get from the common feedpoint of the two transmission lines to element 1 than it takes to get to element 2. You would expect the relative phase in the first case to be less than it is in the second case, as the difference in physical line lengths is 6 inches in the first case and 20 inches in the second. A wavelength at 146.5 MHz in RG-8 is 53 inches, so the phase difference in the first case is (6/53) 360 = 41° if the VSWR is 1.0; in the second case it is 135 degrees.

In practice, the actual phase shifts turn out to be much different because the lines aren't matched. In fact, the relative phase in the first case is 91 degrees, and in the second case it's 150 degrees. So you missed the boat by 50 degrees in the first case and by 15 degrees in the second (**Table 2**). But because you actually wanted a relative phase of 90 degrees, rather than 41 degrees, the first case looks pretty good as a practical feeder solution. When I say relative phase, I mean that i2 leads i1 by 90 degrees, or i1 lags i2 by 90 degrees.

Remember that the sign of the electrical length of the transmission line is normally construed as positive, but the current phase shift is normally considered negative to connote the time lag associated with travel through the line. So in **Figure 6**, if you have a quarter-wavelength line, you would say that theta (Θ) is 90 degrees. However, the phase shift phi (φ) associated with this line is -90 degrees, for the matched condition. The formulas listed in this article assume that the value for Θ is positive. So think electrical length when you see Θ, and think time-lag phase shift when you see φ.

The transmission lines are terminated in the

impedances above, so they are mismatched. The lines transform their load impedances (**Equation 5**) to the Z1' and Z2' values in **Table 3**. The combined input impedance, Zin, at their common junction is simply Z1' in parallel with Z2'. The current phase shift through the mismatched transmission lines can be determined with **Equation 6**. If you change the line lengths in an attempt to produce higher gain, comparing case 2 to case 1 above, your impedance match is even worse—as the element currents increase and the operating resistances decrease farther below the 52-ohm characteristic impedance of the transmission line. Increased currents for a given power is the price you pay for higher gain.

You don't have to use a moment-method antenna analysis program to determine the currents in the individual array elements. Instead you can simply plug the tee-network model of the coupled antenna elements per **Figure 4**, and the two feeder lines, into an AC network analysis program. If the network analysis program doesn't have a transmission-line model, you can make your own pi- or tee-network equivalent.

$$Z' = \frac{Z - jZo \tan\Theta}{1 - j [Z/Zo] \tan\Theta} \qquad (5)$$

$$\varphi = \tan{-1} \frac{R}{X + Zo/\tan\Theta} \qquad (6)$$

Note that the current phase shift through the transmission lines given by **Equation 5** is only part of the story. There's an additional phase shift that occurs at the common end of the transmission lines because the input current doesn't split evenly between the two lines. That is, the input impedance to each line is different (Z1' and Z2'). Finding this additional phase

Table 5. Pattern Adjustment with Series Coil and Capacitor at Element 2

#2 Tuning Reactance	Zin (ohms)	Current at bases			
		i1 (amps)	i2 (amps)	dBi	dBf
-40 ohms	24+j22	1.02/-85°	0.84/0°	8.9	13.2
-30	26+j21	1.08/-86	0.90/0	9.0	15.6
-20	28+j20	1.04/-87	0.96/0	9.0	19.1
-10	30+j18	0.98/-89	1.03/0	9.1	23.4
0	32+j15	0.91/-91	1.11/0	9.2	22.3
10	33+j11	0.83/-87	1.20/0	9.2	16.9
20	32+j6	0.73/-98	1.29/0	9.2	12.6
30	30+j1	0.62/-111	1.38/0	9.1	9.3

shift requires a simple exercise of Ohm's Law, where the relative phase of the two input currents can be determined by **Equation 7**.

$$i2'/i1' = Z1'/Z2' \tag{7}$$

What you've done by decreasing the length of line 2 is to compromise the impedance match to obtain maximum gain; but in neither case do you have a sharp null out the back, because the element current magnitudes are never equal. Is there some other combination of transmission line lengths and types that would offer a better impedance match and also produce the desired pattern? Maybe, but finding it could be a painfully iterative process.

Practical adjustment techniques

Before getting into a full-blown phasing and coupling network design, look at what a single variable capacitor installed at the input to element 1 buys you in the way of adjustability. You know from **Table 1** that the inductive input reactance to this element varies from 27 to 47 ohms for the range of radiation patterns of interest. If you tune this out with a capacitive reactance, you reduce the VSWR on line 1 and afford some adjustability of the pattern (**Table 4**). The pattern degrades with a capacitor, but gives you what you want with an inductor. Apparently, the design is relying heavily on mismatch to produce additional phase shift in the transmission line.

The following information is for the case where you have 23 inches of RG-8 connected to element 1 and 17 inches of RG-8 connected to element 2. Note the touchiness of the pattern; this is an indication to be very careful with your transmission line wiring, because even a little inductance has a big effect at the base of the

driven elements. It also indicates that you need a practical adjustment handle if you want the pattern to arrive when and where you expect it. **Table 4** illustrates the range of forward gain and front-to-back values that can be had with a simple turn of a variable capacitor in series with a fixed inductor.

It may be that if an inductor in series with element 1 provides the desired adjustment range, then a capacitor in series with element 2 may do the same. A small variable capacitor seems better suited to the application than a variable inductor, so to cover the reactance range of -40 to +40 ohms, you could use a variable capacitor in series with a fixed coil. In the 2-meter band, a 5- to 30-pF variable capacitor in series with a 90 nH air-core coil would work (about three turns on a half-inch diameter, see **Figure 7**).

Table 5 shows what happens when you tune the base of element 2. It should be no surprise that, once again, the results aren't as expected. Specifically, you just don't get the big forward gain you obtain by tuning at the base of element 1. So, it looks like you need to use an inductive reactance in series with element 1 to get the maximum gain from your array. It could be that the adjustment at the base of element 2 is less effective because there is less power in this element.

Perhaps a variable capacitor at the input to one of the transmission lines will do the trick. Maybe a variable capacitor at the input and output of both transmission lines will provide enough adjustment latitude. I leave this as an exercise for the reader, because I don't believe in taking all the fun out of a project and leaving no room for experimentation. Remember that at these relatively low power levels, the use of mismatched feeder lines isn't a bad practice, so take advantage of it.

There's nothing egregiously wrong with hav-

ing a high VSWR on the feeder lines. In most cases, the extra losses and higher voltages and currents in short feeder lines caused by the mismatch aren't too significant at ham radio power levels. But remember, the higher the gain, the higher the currents in the feeder lines. If you're in doubt, be sure to calculate all component voltages and currents, and be aware that the wet voltage ratings at the ends of the lines are a lot lower than the ideal dry sea-level ratings (yes, you must de-rate for altitude).

Make some effort to match the main transmission line at the common junction of the feeder lines. The main line is a lot longer than the feeders, and its overall power loss could be significantly higher in the mismatched condition. A couple of stubs will do the trick, as long as you're careful they don't become additional parasitic radiating elements in your array.

As an alternative to the mismatched feeder approach, you can match the operating base impedance of each element to the transmission line surge impedance and provide a controlled phase-shift power divider at the common input to the array. However, you need to keep track of the phase shift across the element matching networks, too. And, it's not easy to make an impedance matching network that works well in the 2-meter band, and has independent phase and impedance transformation adjustability.

You can make a tee-network phasing and coupling unit out of three stubs. A quarter-wavelength ladder-line stub with a sliding short will work well as an inductor, and an adjustable length open-circuited stub would work as a capacitor—but then you have to deal with the problem of parasitic reradiation from the stubs. That is, the stubs start to act as undesired antenna elements. A lumped parameter tee-network may be the best approach, because these components will be considerably smaller than the stubs.

An adjustable coaxial capacitor is easy to fashion from a couple inches of half-inch copper pipe stuffed with a plastic dowel bored and tapped to accept a 3/8-inch diameter screw. This provides a surge impedance of about 10 ohms, which yields about 14 pF per inch. A 5/16-inch screw yields about 9 pF per inch, and a 1/4-20 screw about 6 pF per inch. A DC capacitance meter will get you into the ballpark, but an impedance meter at VHF—if used properly—is safer.

The time-honored method of playing an air-core solenoidal coil like an accordion provides a practical form of adjustable inductor. You could also try an adjustable ferrite slug, if you can find one that's not too lossy at 146 MHz. In any case, you need to make a good estimate of the initial component values. This means using the formulas and measurement techniques pre-

sented here or running a moment-method analysis to determine what your driving-point impedances are likely to be.

Phasing and coupling network example

Assume that you really want equal current magnitudes in each element and a relative phase of 90 degrees. You already know the driving point impedances and powers to expect. Now you want to match those impedances to 52 ohms, and you also want 52 ohms at the common feedpoint to the array. Start at the power divider, using what's called an Ohm's Law configuration.

You know that element 1 radiates 77 watts and element 2 radiates 23 watts. For more power to be transferred to line 1, you need to present a lower resistance to the common input point looking towards line 1 as compared to line 2 (V2/R). You also know that the paralleled resistances looking toward the two lines is 52 ohms. So for 100 watts input to the array the math looks like this, where V is the input voltage across the power divider:

$V^2 = PR = 100(52) = 5200$

$V^2/R1 = 77$ watts, so $R1 = 68$ ohms

$V^2/R2 = 23$ watts, so $R2 = 222$ ohms

Checking your work,

$R1R2/(R1 + R2) = 52$ ohms.

So, if you connect an L network between the common point and each transmission line, you might have a network that looks like the one in **Figure 8**. Note that the shunt elements of the two high-pass L networks combine to form one coil. I chose "leading" phase shift networks, as opposed to "lagging" phase shift, low-pass networks simply because it's cleaner and easier to adjust a coaxial screw capacitor than it is to adjust a coil. So I wound up with two capacitors and one coil, instead of two coils and one capacitor.

In my opinion, a coaxial screw inductor would be physically too long for this application because it would begin to act as another radiating element in the array. However, if you're willing to add more-or-less a quarter wavelength of line to your capacitor, you can make it look like an inductor. Note that the two capacitors in the power divider aren't connected to ground, so you must be careful how you mount them. If you make them from 1/2-inch copper pipe, you could align the pipes in parallel and solder them together; this would form the common side of the power divider.

Now that you've taken care of the front end of your phasing and coupling design, you must determine the phase shift needed in the impedance matching units at the other ends of the feeders. Note that the phase shift across the two legs of the power divider are not equal; that is, leg 1 lags 31 degrees behind leg 2 (62 - 31 = 31). To simplify the design, use equal length feeder lines. This has the benefit of creating a symmetrical perturbation of the antenna structure and will tend to minimize pattern distortion and other side effects. You want the relative current in element 1 to lag that of element 2 by 90 degrees. You already have 31 degrees of lag provided by the power divider, so you need an additional -59 degrees. You can obtain this by selecting a phase shift across matching network 1 of 50 degrees, and across matching network 2 of 109 degrees per **Figure 9**. The component values shown are for 146.5 MHz.

During adjustment, it may also be useful to have a dual-trace, wide-band oscilloscope handy to monitor the currents and phases in each element using a small sample loop located near, but not connected to, each element. However, you would have to be very careful

with the placement and detuning of your sample lines so as not to disturb the array. This gets complicated at VHF, and, unless you're extremely patient, this may be more trouble than it's worth. However, it is quite practical at lower frequencies.

It would be nice to have two helpers with handheld 2-meter transceivers—one aligned with the main bang of the antenna and the other on the other side. You could communicate with them using your transceiver plugged into the antenna under test. As you adjusted your phasing and coupling components, your helpers would relay their S-meter readings back to you. You could also use a single helper, and rotate the antenna 180 degrees of azimuth to measure the forward and reverse fields. Be sure to maintain a constant input power to the antenna when you do this; an auto-tuner would be ideal for these circumstances.

If there's no one around to help, you could keydown the handheld and leave it radiating in a nearby open area while you adjust your antenna for maximum received signal in the forward direction, and minimum signal in the reverse direction. Keep a clipboard and pencil
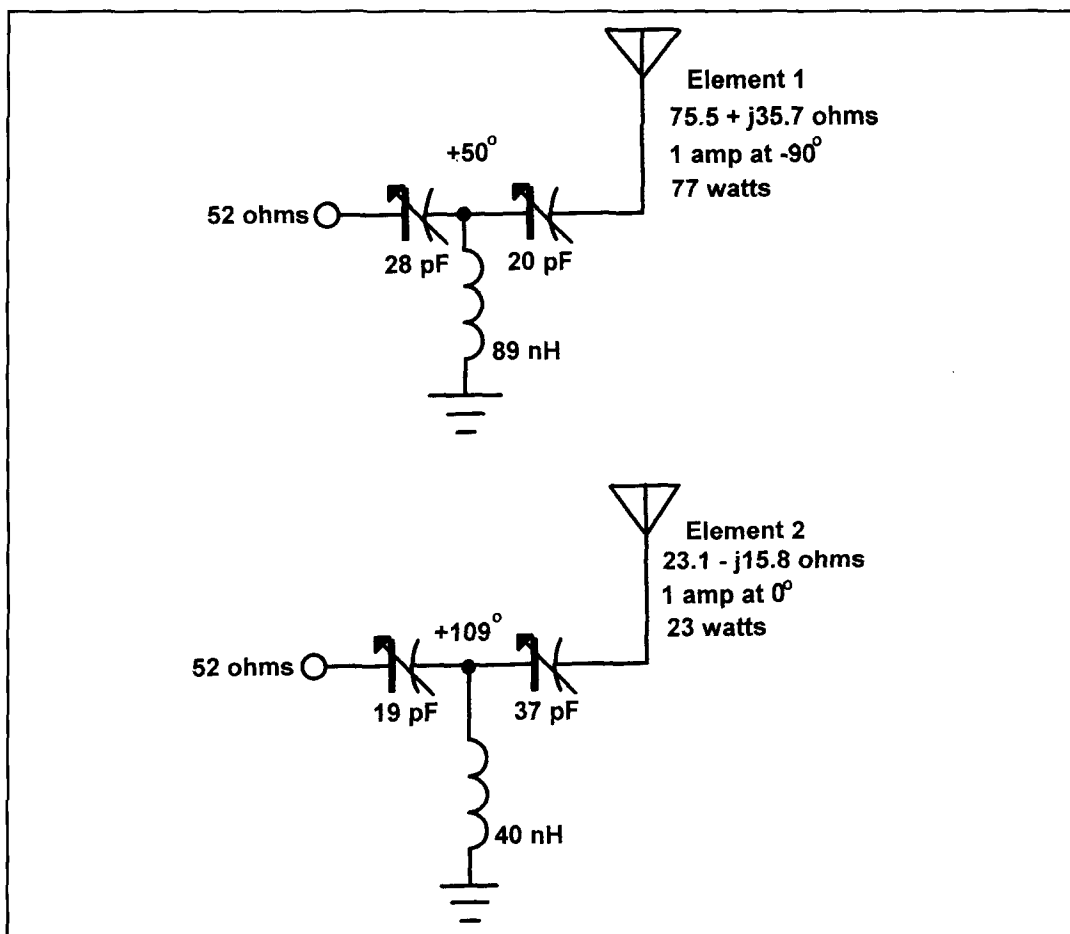
Figure 9. Matching networks.

at your side to keep track of component settings and the resulting field intensities.

Summary

The technique of connecting mismatched feeders to the driven elements in a phased array is an expedient method for obtaining relative current phases that approximate the desired values. A variable capacitor in series with one or more of the lines will often provide a good degree of adjustability; however, it won't always work. In this case, you may need to insert phasing and coupling networks between the elements and the feeder lines, and construct a power divider at the input to the feeders. For this approach to be successful, it's important to have a knowledge of the operating feed impedances of the driven elements, and the way they interact. In actual practice, it can be a real "can of worms."

I suggest a simplified approach that will allow easy adjustability of the pattern. In this approach, a single series coil and/or capacitor is inserted at the feedpoint of one or more of the driven elements. In all cases where lumped parameters are added to the design, impedance and pattern bandwidth may suffer, but this needs to be addressed on a case-by-case basis. If your moment-method program doesn't allow you to enter all the network components in your phasing and coupling equipment, a net-work analysis program can do the job if you model the antenna self- and mutual-impedances using the multi-port network models discussed here.

You don't actually need an antenna analysis program to measure and model the interaction between radiating elements in an antenna. All you need is an impedance measuring device, some math skills, and some network analysis software. However, a moment-method program is certainly helpful, and should be part of your arsenal of design tools. Sometimes the stray reactances and pattern "scattering" within and around an antenna can mask what's really going on, and if you can approach an antenna problem simultaneously from different angles, you'll find a design solution in less time.

A note about the mutual coupling model. I used a tee-network in this article, but a special pi-network requiring some fancy matrix math is the practical form when more than two antenna elements are used.[3] Topologically a pi-network has one less voltage node than a tee-network, and this has advantages over a tee-network. However, a detailed discussion of the generalized antenna impedance model is beyond the scope of this article; I have, however, attached an appendix providing the basics. ∎

REFERENCES
1. Harold Thomas, "A 2-Meter Phased-Array Antenna," *QST*, January 1998.
2. Grant Bingeman, "Negative Towers," *Broadcast Management/Engineering*, November 1980.
3. Dane Jubera, "Toward Improved Control of Medium Wave Directional Arrays," *IEEE Transactions on Broadcasting*, December 1981.

Appendix

A rigorous multi-port network model of a phased array containing "n" radiating elements may be developed using the following procedure. First, form a matrix of the mutual impedances between all the elements. This matrix will contain n^2 complex values, and look like this:

$$
\begin{matrix}
Z11 & Z12 & ... & Z1n \\
Z21 & Z22 & ... & Z2n \\
\cdot & \cdot & & \cdot \\
\cdot & \cdot & & \cdot \\
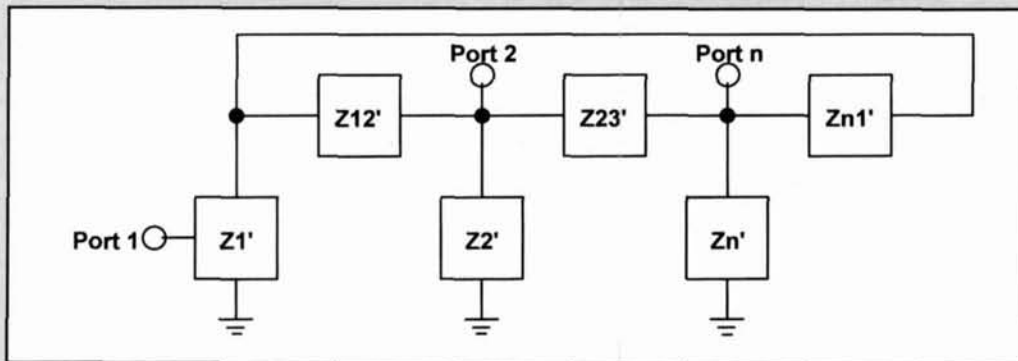Zn1 & Zn2 & ... & Znn
\end{matrix}
$$

Figure 10. Multi-port phased array network model.

Then you must invert the matrix, keeping in mind that matrix inversion is a tedious math process. You probably won't want to program it yourself, but simply call it up as a function from a math library. In other words, an individual matrix element Yij doesn't equal 1/Zij.

$$
\begin{array}{cccc}
Y11 & Y12 & \dots & Y1n \\
Y21 & Y22 & \dots & Y2n \\
\cdot & \cdot & & \cdot \\
\cdot & \cdot & & \cdot \\
Yn1 & Yn2 & \dots & Ynn
\end{array}
$$

Next, you must calculate the lumped-parameter values for the network shown in **Figure 10**, using these equations:

$$Zi' = 1 / [Yi1 + Yi2 + \dots + Yin]$$

$$Zij' = -1 / Yij$$

Figure 11 shows the vertical or elevation pattern of the array described in **Figure 1** when element currents are equal in magnitude and phased 135 degrees. The forward gains and front-to-back ratios vary depending on the elevation angle of interest. The values throughout this article are for an elevation angle of one degree above the horizon plane. Because communication at these frequencies is almost always line-of-sight, this is a valid angle. Note that the far-field goes to zero when the elevation angle is zero, but the near field does not. For 100 watts input to the array, you can expect a vertically polarized field intensity at one kilometer, two meters above ground of about 1.5 mV/m for the pattern shown in **Figure 11** on a bearing of 270 degrees true over 5 mS/m earth, assuming no reflections. This value would be slightly higher if the ground conductivity were better. Over perfect ground the field would be about 12 mV/m.
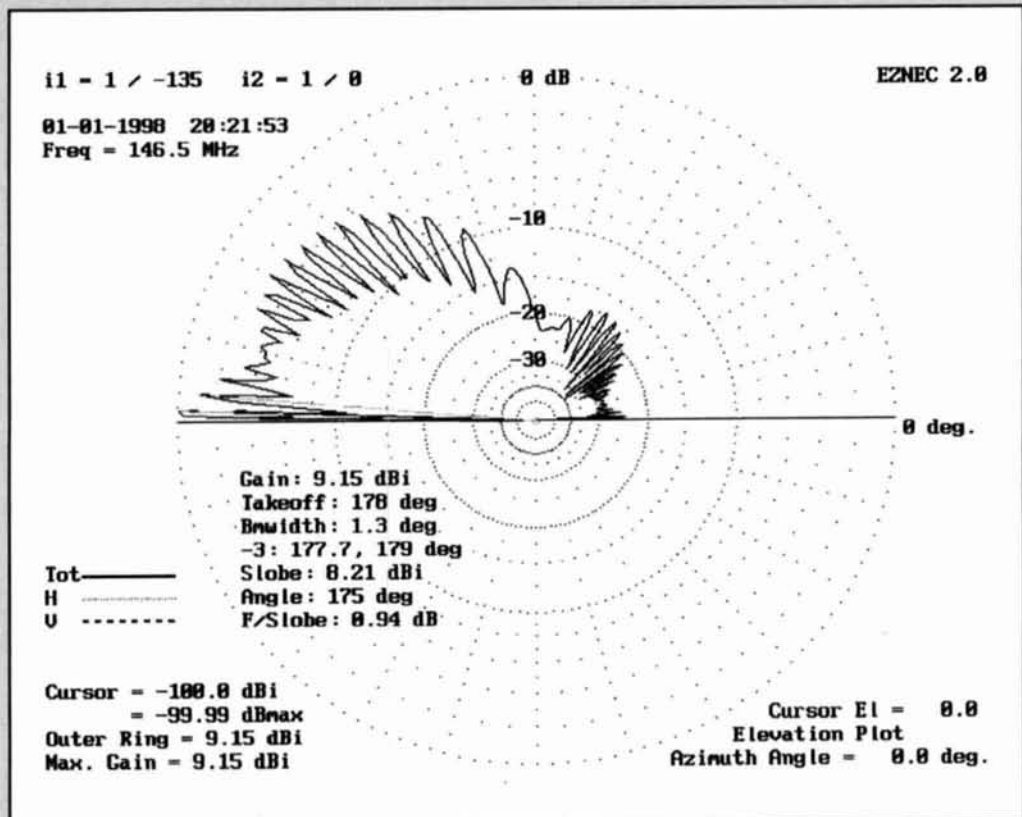

Figure 11. Elevation pattern of array described in Figure 1 when elements are equal in magnitude and phased 135 degrees.

Quarterly Computing—Using Electronic Workbench to Analyze Audio Circuits

By M.A. Chapman, KI6BP

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Summary: A description of the features of *Electronics Workbench* software (EWB) as they apply to audio circuit design. The column emphasizes the use of EWB to design audio filters.

PRODUCT INFORMATION

MFJ-4035MV Power Supply

MFJ Enterprises has announced the MFJ-4035MV 35/30 amp adjustable regulated DC power supply. This heavy-duty power supply features 35 amps surge and 30 amps continuous, and is adjustable and regulated. Voltage is front-panel adjustable from one to 14 Volts DC with detent set 13.8 volts DC. The power supply has lighted front panel meters and an ON/OFF switch.

Features include built-in circuit protection that automatically shuts down the power supply when it's drawing too much current and a quiet internal fan to keep components cool.

The MFJ-4035MV is priced at $149.95, and it carries MFJ's *No Matter What*™ one-year limited warranty. For the name of your nearest dealer or to order, call (800) 647-1800; Fax: (601) 323-6551; E-mail: <mfj@mfjenterprises.com>; or visit the Web site at <http://www.mfjenterprises.com>.

Belden® 50-Ohm Transmission Plenum Coax for Wireless, Microwave, and RF Products

Belden Wire & Cable Company has recently introduced Brilliance 7733A, a 50-ohm, low-loss transmission cable that's designed for delivery of wireless, microwave, and radio frequency (RF) signals in plenum applications. The coax cable can also be used for radio tower connections.

To impart low smoke and flame-spread properties, Belden jackets the cable in fluorocopolymer (FPC), a high-quality, plenum-grade compound that offers a practical and economical alternative to traditional plenum jackets. As a result, Belden 7733A passes the UL 910 (Steiner Tunnel) flame set and is rated NEC CMP.

Belden 7733A 50-Ohm Transmission Coax is ruggedly constructed with a high-velocity, gas-injected foam FEP insulation over its 10-AWG solid copper conductor. Foam FEP offers better water migration than semi-solid constructions, so when the cabling is used in outdoor applications, such as antenna masts and towers, it is less susceptible to damage from rain and moisture should the connector seals be compromised. The uniform, small-cell foam FEP insulation also provides a high velocity of propagation (84%), a high degree of crush resistance, and resists conductor migration.

Belden 7733A cable is shielded with Belden's exclusive Duofoil® shield, featuring aluminum/polyester/aluminum laminate tape, plus a 90-percent tinned copper braid shield for 100% coverage.

For more information, contact Belden Wire & Company, P.O. Box 1980, Richmond, Indiana 47375; Phone: (800) BELDEN-4; Home Page: <www.belden.com>.

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Voltronics Corporation announces its ultra-miniature surface mount chip-size trimmer capacitors with a new range of 0.4 to 1.0 pF. The "JS" Series measures 2.8 mm x 2.2 mm x 1.0 mm and is ideal for telecommunications applications. High stability makes it perfect for frequency oscillator applications.

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William J. Byron, *W7DHD*
240 Canyon Drive
Sedona, Arizona 86336

THE ARC METHOD OF PRODUCING CONTINUOUS WAVES

> *In the spring of 1994, Bill brought us "Arcs and Sparks, Part 1: How Did They Work?" As he settled down to write Part 2 recently, Bill realized he had covered all the pertinent information in this article he had written for the* AWA Review. *The powers that be at the* AWA Review *have graciously allowed us to reprint it here so our readers can learn "the rest of the story."—Ed.*

Two important basic methods were developed in the early 1900s for producing radio waves. The first, using spark technology descending directly from the experiments of Heinrich Hertz, was more immediately successful. Spark transmitters were the first in the world to span large distances, covering most of the industrially developed nations, and first to cross the Atlantic Ocean. The other technique was the arc method, the main subject of this article.

This is a story about arc-transmitter development, and of the men who did it. It sprang from discussions with other amateurs—discussions which revealed that *none* of us knew much about arc technology, except the "old wives' tales" that have been passed down over the years. Many of those are erroneous, and call for correction; the ones that are correct deserve elaboration.

It didn't take long for experimenters and practitioners alike to attempt the logical extension of the wire telephone: radio telephony. Many of the early experiments—and some formal patents, all of which eventually proved to be worthless—involved the use of spark-gener-ators adapted for telephony. It became obvious that there were few modifications of spark technology of use in practical radiotelephony. That required continuous waves. Spark systems were of the damped-wave variety, no matter how high the charging frequency. Fessenden's early concept of the high-frequency alternator was as a carrier generator. Nevertheless, he modified plans to use the first alternator built for him as a high pulse-rate synchronous rotary gap spark generator when it turned out to be capable of reaching only 10 kHz.[1] Other experimenters attempted to modulate high-pulse-rate spark transmitters. Subsequent alternators were used as carrier generators and some were modulated—but they came several years later.

Arc technology developed concurrently with that of the spark. Early experiments date from about 1900. *Experiments* they were; poorly grounded in theory, and subject to so many variables that it was difficult to predict much about their performance. Although the power outputs of arc transmitters were initially very low, they did have the property necessary for voice transmission: *continuous waves*. Many low-powered arc sets were successfully modulated, with their signals spanning many miles. Their development will be covered only cursorily here; the full history of radio-telephony would be the subject of a very large book. The development of arc technology led to the most powerful transmitters in the world; they featured unmodulated CW, however, not voice transmission.

The "Poulsen Arc," much quoted, and devised in the earliest part of the century by Valdemar Poulsen, did not spring forth mature and ready for operation. He did use a magnetic
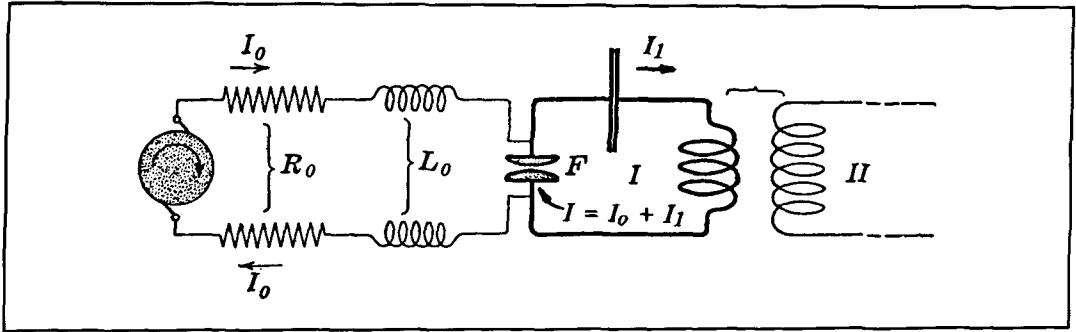
Figure 1. Basic arc circuit.

field, but that had been originated by William Duddell, in England. The method was called "Duddell-Poulsen" in some early references. His major early contribution was the use of a "spirit lamp" in the vicinity of the arc, which provided one of the essentials in further arc development—that of using a highly mobile ion (hydrogen) that enabled scavenging of the arc within each radio-frequency cycle. Each experimenter built on the results of prior work, on which much technical development relies even now.

There can be no complete story that does not relate the accomplishments of the Federal Telegraph Company and its chief engineer during the most exciting period of arc-transmitter development. He was Leonard F. Fuller. He spanned the era from days of spark transmitters up to the beginning of the nuclear age, to which he made a coincidental but important contribution. His activities are necessary parts of this story.

The arc

The "arc" method of producing radio-frequency energy has its roots in experiments with spark technology and attempted improvements in the state of the art. Some of the earlier spark transmitters used direct-current sources, the

most notable being those of Marconi. He used 6000 two-volt storage cells in series for some early high-powered transmitters, though they eventually were to use various sources of alternating current. It is no accident that the circuitry for the arc systems parallels that of the spark (see **Figure 1**), and it is possible that some of the very early experiments resulted from problems associated with steady arcing at the spark-gaps. **Figure 2** is of the form that the arcs employed in operation.

Early experimenters in the field were Elihu Thomson, Reginald Fessenden, and Nikola Tesla. The first two became famous for other activities. Thomson was the first; others were interested, too, so the experiments proliferated. It had been discovered that an arc would oscillate if a series-resonant circuit were placed around it. It was functional only at audio frequencies, however—the "singing" arc. Fessenden held the original patents on the mechanical device which evolved as the Alexanderson alternator, and he was very active in the development of all transmission methods from spark up through alternators. Tesla never received adequate credit during his lifetime for the contributions he made—which were numerous, and *very* important. Every time a polyphase motor is run, we are seeing Tesla in both the motor and the two or three-phase power that runs it; even in the alternating cur-
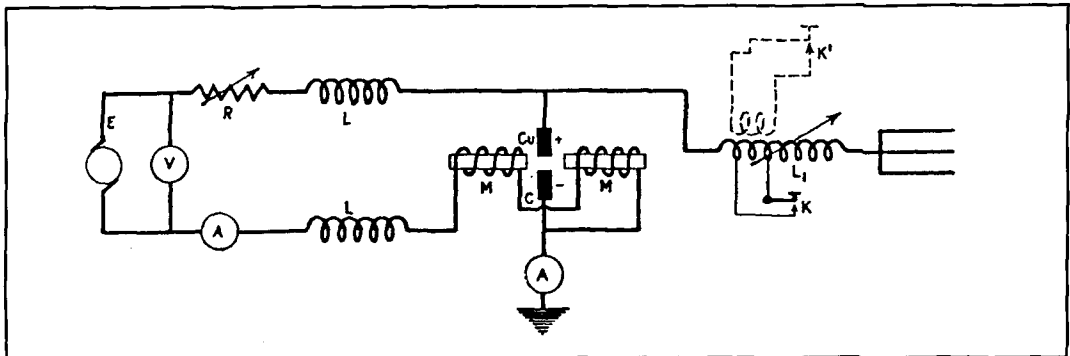

Figure 2. Typical arc transmitter.

rent itself. Tesla was one of the first to use carbon electrodes, and was among the earliest to produce oscillating arcs. Charles Steinmetz himself was active in the center of things during those times.

Shortly after the turn of the century, the state of the art permitted real-time studies of arc (and spark) characteristics with the Braun tube, which was the cold-cathode precursor of today's oscilloscope. **Figure 3** shows the method used. Note that both electrostatic and electromagnetic deflection are employed. The technique dates circa 1908. **Figure 4** is a photograph of the face of a Braun tube displaying a damped (spark) waveform. Similar studies were made on the arc, **Figure 5** being one such result. Note that linear sweep is neither desirable nor necessary for the latter display.

Use of a primitive oscilloscope enabled those engaged in fundamental research to examine very well some of the characteristics of the arc discharge, just as we might do today.

Physics of the arc

While the spark systems will function equally well on AC or DC, the arc systems *require* direct current. This results from the physical and electrical characteristics of the arc.

Although the spark systems required that the spark be extinguished between charging pulses (but not within radio-frequency cycles), the arc method requires the arc to be extinguished *within RF cycles*. One would suspect that, since DC is required, the arc must therefore be constant. It isn't: it is extinguished and reignited during each and every RF cycle. This necessitated the use of "blow-out" magnets and the provision of alcohol (or other organic) drip-feed to the arc chamber, as described later.

An essential characteristic of the arc must be emphasized. The characteristic curve (the
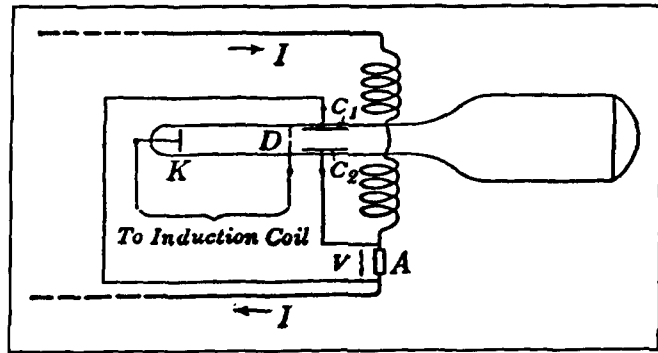
Figure 3. Braun-tube oscillator.

hyperbolic curve in **Figure 6**) shows that for an increase in voltage there is a corresponding decrease in arc current—and vice versa. That is the "negative resistance" that promotes oscillation. Arc transmitters simply would not be possible without it. Some early tetrode vacuum-tube oscillators functioned because of the negative slope in the screen current curve (in many tube types), and tunnel-diode oscillators work on the same principle.

The voltages applied to arc transmitters were much lower than in big spark transmitters: 2000 volts or less for the arcs, 50,000 or more for the sparks. Some small arc converters used voltages as low as 110 from DC house mains—common in the Teens—and most operated from supply voltages in the vicinity of 550. The first Poulsen-constructed arc transmitter purchased and brought to this country by Elwell operated from 110 volts DC. It was rated at 100 watts input.

The L/C ratios were inverted: high capacitance in the spark systems versus high inductance in the arcs. Since the primary circuit (Zenneck and others called this the "condenser circuit") must be resonant at or near the transmitting frequency (see below), an increase in inductance necessitated a corresponding
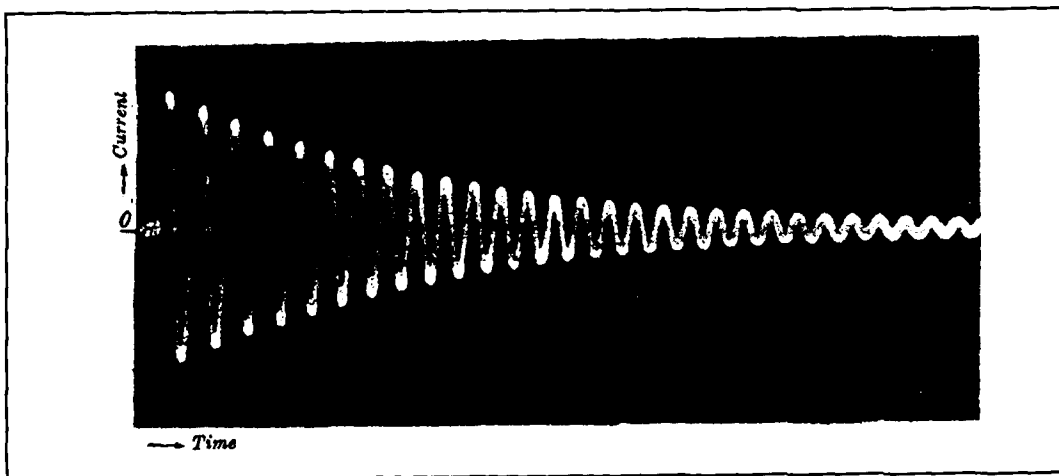

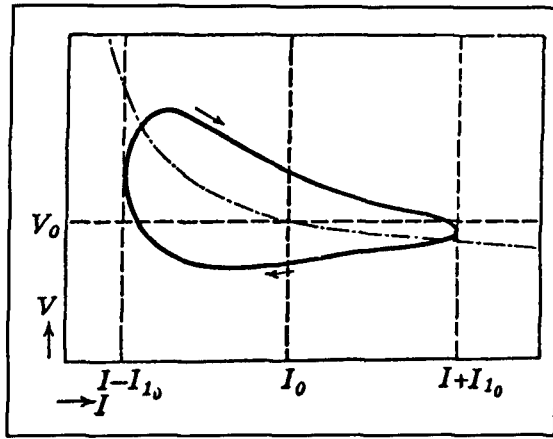Figure 4. Damped-wave oscillation made with Braun tube.

Figure 5. V-I characteristic of an arc.

decrease in capacitance, and vice versa. Thus, although most of the contemporary descriptions of the arcs spoke in terms of capacitor voltages, they were really current-driven devices. Many of the smaller transmitters (and the largest ever built, as will be seen) had no "condenser circuit"; the antenna and tuning inductances were connected directly to the anode of the arc, while the cathode was grounded. The "R, L, and C" were those of the antenna circuit. The largest arc transmitters of the Federal Telegraph Company used no magnetic coupling in order to circumvent certain Marconi patents. The intrinsic capacitances of the largest fixed transmitting antennas (unavailable to shipboard stations, incidentally) seldom exceeded 0.05 μFs. The bulk of the large energy storage initially must have been in the inductance. There is frequent reference to that stored in the capacitor, but no mention of the energy in the inductor. In none of the references are the inductor-related expressions

$$E = 1/2LI^2 \text{ and} \qquad (1)$$

$$V = L(di/dt) \qquad (2)$$

mentioned, although Fuller appreciated the inductive effect, and explained it in different terms. It should be reiterated that this discussion is about the cyclic reignition of the arc; it is not a question of energy, per se, but of the production of enough voltage for reignition without manually restriking the arc.

Calculation of the power input for *spark* systems can be done by taking the peak charging voltage (V) and the capacitance in the primary circuit (C) in farads, and computing the energy (E) in joules, or watt-seconds. If the energy in joules per charge is multiplied by the charging pulse-rate (δ, per second) the primary power input can be calculated:

$$\text{Spark power input (watts)} = δ \times C(1/2)V^2 \qquad (3)$$

The capacitors in the arc transmitters could not possibly deliver the necessary voltage for reignition with a charging potential of less than 2000 volts, assuming that the voltage comes from the initial charge. Zenneck stated in 1915, however, that "As a matter of fact, it [the capacitor voltage] may under certain conditions rise to a much higher value [than the source voltage]."[2] Today we know why, but it may have been difficult then to explain the phenomenon. It is sufficient that they knew it happened, and documented it.

Arc transmitters eventually were operated at the same power levels as spark transmitters; even higher, because there were several converters of 350- to 1000-kW size built and operated during and after World War I.

The static behavior of an arc can be characterized by

$$V = a + (b/I) \qquad (4)$$

where V is the voltage across the arc, I is the current, and a and b are constants. At the extreme (I ≈ 0) the equation does not hold, of course, and the voltage is simply V_z, the cyclic reignition potential of the arc. The equation is of the form of an hyperbola, as shown in **Figure 6**. However, an examination of **Figure 5** shows that the current and voltage do not follow the shape of the hyperbola, but rather envelop it. It is a similar case when one compares a class-A vacuum tube amplifier which operates into a pure resistance, wherein the current-voltage characteristics are linear, with the same amplifier operating into a reactive load. In the latter case, we find an ellipse which surrounds the projected linear load line. **Figure 5** came from measurements made with the Braun tube and circuitry in **Figure 3**, with the arc placed where the symbol "A" appears.[3]

The currents (and symbols) shown in **Figure 2** are graphed in **Figures 7** and **8**. It can be seen that the current I_0 follows very nearly a sinusoidal path in **Figure 7**. The flat portion is the period in which the arc is extinguished. It is labeled T_2 in **Figure 8**. The arc conducts during period T_1. If one examines the dashed lines in **Figure 8**, it could be concluded that the voltage across the arc is constant during the period T_1 but follows the sine curve during the period T_2. This is the time wherein Zenneck postulates that the capacitor is charging (the ordinate in the figure is voltage). At the time a voltage is reached across the gap corresponding to the point V_z, the arc is *reignited*, and the voltage drops to the constant value V_b. All this assumes that the voltage across the gap is constant during the arc.

Experimental evidence revealed, however, a "boat-shaped" voltage characteristic along this

supposedly flat line. The actual characteristic is shown in **Figure 9**, along with the shape of the current.

The voltages across both the inductance and the capacitance are actually iX_L and iX_C. These are the voltages that previously were said to occur "under certain conditions." In fairness to the authors, the energy in both the reactive elements trades back-and-forth in the usual resonant manner, and the energy in the capacitor *at the peak* is indeed

$$E_p = 1/2CV_p^2 \qquad (5)$$

Where V_p is the peak *instantaneous* voltage, not the supply voltage, a value of 30 or 40 kV in the largest transmitters. (In the Federal Telegraph systems the capacitance was *only* that of the antenna—so the high voltage was on the antenna itself, and the value of that capacitance was usually 0.015 μF maximum.) This is where the suppositions in the references may have led readers astray. Zenneck *did* say in his book, however:

That this is possible may be understood that a change in I_0... may produce higher potentials because of the self-induction of L_0, by adding to the voltage across the condenser terminals...*Whether this is always the sole explanation* is a question which need not be further investigated here.[4] [emphasis added].

The duration of T_2 in **Figure 8** depends on several factors, but mainly on how fast the gap
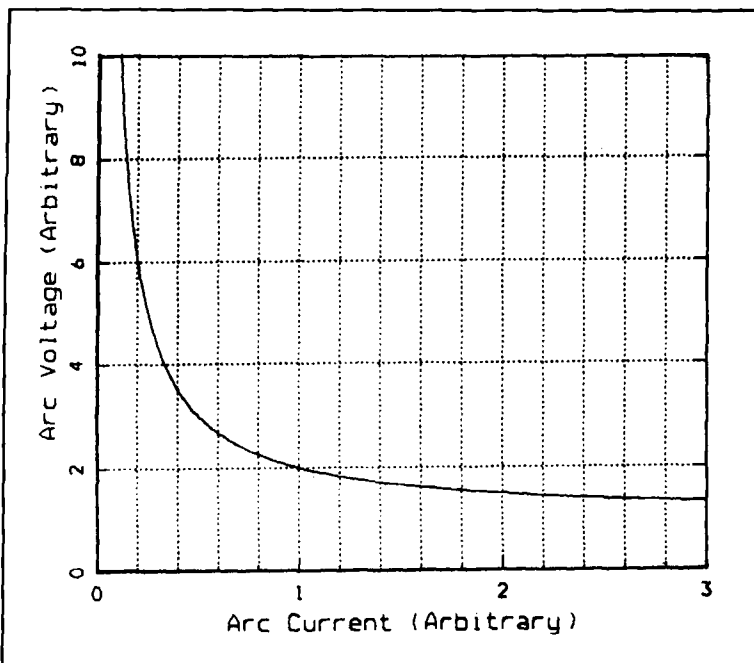
Figure 6. Arc electrical characteristics.

can be cleared of residual ions. Suffice it to say that the frequency of the arc transmitter is not controlled entirely by R, L, and C (R includes arc, circuit, and radiation resistances). The period of the cycle is actually T_1 plus T_2, and T_2 may vary appreciably. From an operating standpoint, it is a matter of adjusting the trans-

Table 1. Common liquid organics used in arc converters.

Organic	Molecule Shape
Ethyl Alcohol*	H H H-C-C-OH H H
Methyl Alcohol	H H-C-OH H
Gasoline (Typical)	H H H H H H H H-C-C-C-C-C-C-C-H H H H H H H H
Kerosene (typical)*	H H H H H H H H H H H H-C-C-C-C-C-C-C-C-C-C-H H H H H H H H H H H H
Acetylene (Gas)	H-C≡C-H

*Most commonly used by Federal Telegraph Co.
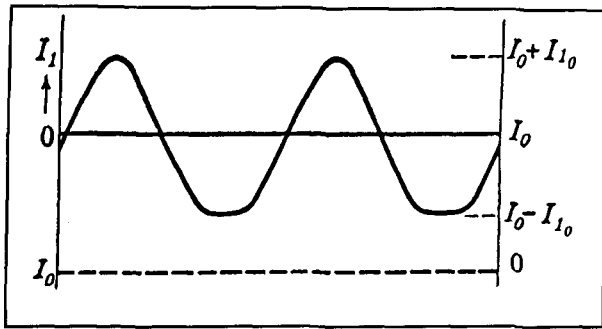
Figure 7. Capacitor current.

mitter to the desired frequency and power output. It also shows the discriminating examiner that the spectrum of the arc transmitter was not really all that pure—though much better than that of the spark, and adequate for the time. Arc systems were used well into the vacuum-tube age.

Operational concerns

It is shown in the foregoing figures that arc is extinguished for a period during each RF cycle. In order to obtain a constant frequency, unaffected by varying conditions in the arc, it is imperative that its environment be controlled so that the arc is encouraged to reignite at exactly the same time during each RF cycle (this would fix T_2). There was enough "jitter" in arc converters, however, to increase the signal width, and to produce "mush" or "hair on the signal" in the terms of those who have heard them. We would characterize that today as phase modulation.

To attain the desired regularity, residual ions must be swept from the gap. Hydrogen ions are
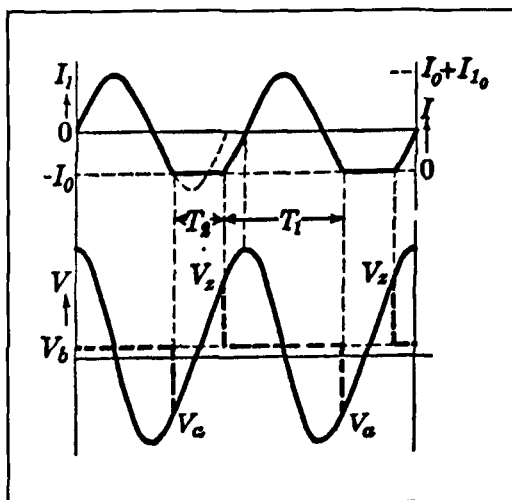
Figure 8. Arc current-voltage.

most mobile, having the lowest atomic and molecular weights (both ions and molecules must appear in the gap). This can be done by making sure that the environment is mostly hydrogen by dripping alcohol or another organic compound into the chamber, with the hydrogen being stripped from the organic molecule. With a high-mobility ion, the "blow-out" magnetic field, perpendicular to the arc current, sweeps the residual ions from the gap. Pure hydrogen and "illuminating gas" were used in some transmitters, the hydrogen being generated by the electrolytic decomposition of water. The alcohol-injection method was much less complicated—and certainly less dangerous.[5] There was always a "puff" or an outright explosion on the initial startup. It was vented by a flapper valve and a pipe to the outside. The larger arcs used an additional centrifugal exhaust fan.

Other organic compounds were used in the arc chambers. Besides the common alcohols (ethyl and methyl), there are references to the use of gasoline, kerosene, and even acetylene. None of these was particularly desirable, though the Federal Telegraph Company and U.S. Navy did use kerosene in their larger converters; the choice depended on the frequency and the power level (see **Table 1**). There is a characteristic of alcohols (both methyl and ethyl) that should be mentioned. Ethyl alcohol decomposes in the arc to a mixture of CO_2 + $6H_2$ + $3C$.[6] CO_2 has a molecular weight of 44, compared to 2 for H_2 and 12 for carbon. It requires very high flux density to clear it out of the gap. Its drawbacks and advantages will be shown later—in the section on Leonard Fuller and Federal Telegraph.

Kerosene, as it developed, became necessary to operate at both higher power levels and shorter wavelengths, despite depositing prodigious amounts of carbon all over the interior of the arc chamber. Acetylene has only two atoms of hydrogen for two atoms of carbon, whereas the object of the use of organics was to place as much hydrogen as possible in the gap relative to carbon. Furthermore, acetylene is second only to hydrogen in its explosive qualities. Acetylene was abandoned early on.

Maintenance was the principal reason for installation of duplicate converters in larger stations. One unit could be operating while the other one was undergoing maintenance. Replacement of the carbon cathode in the large Federal arcs required only a minute or two; cleaning the chamber only a few minutes. This was not feasible in large spark stations—it was impossible to stay in the same room with a large rotary spark. Reduction of down-time is significant in commercial operations, and vital in military communications.

The standard compositions of the arc electrodes were copper for the anode (usually the upper electrode in vertical arcs), and commercial-grade carbon for the cathode. The copper electrode was concave. The graphite cathode tip was convex, so that it would fit concentrically inside the anode tip. The copper electrode was cooled. Fans and fins would suffice in the smaller transmitters, but water cooling was required in the large ones. Even the jacket surrounding the arc chamber in large converters was water-cooled.

The maximum efficiency of an arc system was barely 50 percent.[7] This, of course, meant that in the typical 100-kW transmitter the heat to be carried away from the arc (mostly from the copper electrode) corresponded to 50 kW. The one-megawatt transmitter mentioned earlier must have required either an outside spray pond or a small cooling tower. No matter what ultimate heat-sink was used, there had to be an interim heat-exchanger because the coolant in contact with the anode was distilled water; it was necessary for both high electrical resistance and freedom from deposits.

Carbon electrodes were sacrificial; they had to be adjusted frequently. Methods of obtaining uniform erosion of the electrodes usually involved rotating the carbon cathode by means of an external motor, and all successful transmitters used one such method or another. Some transmitters employed a small centrifugal pump
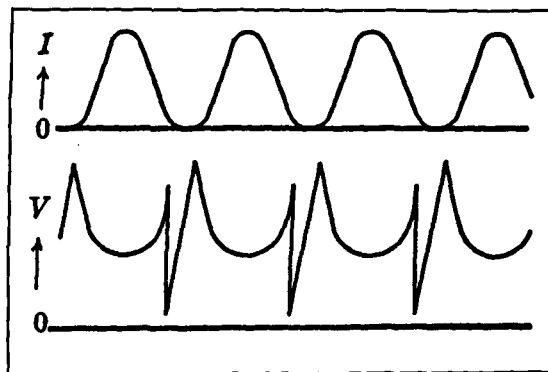
Figure 9. Experimental arc voltage and current.

attached to this same motor, which circulated anode-cooling water. Some smaller shipboard units were cooled by thermal siphoning through a tank mounted on the wall.

Various other electrode shapes were used, too. One used a copper electrode in the form of a thick-walled tube and a flat cylindrical graphite cathode. An axial magnetic field was used to force the arc to rotate slowly around the copper electrode. This form was used in at least one audio-modulated converter. In this method the necessity for mechanical rotation supposedly was avoided. It couldn't have been too successful, because there is only one transmitter cited as using this method out of several documents. Some variants may have been developed
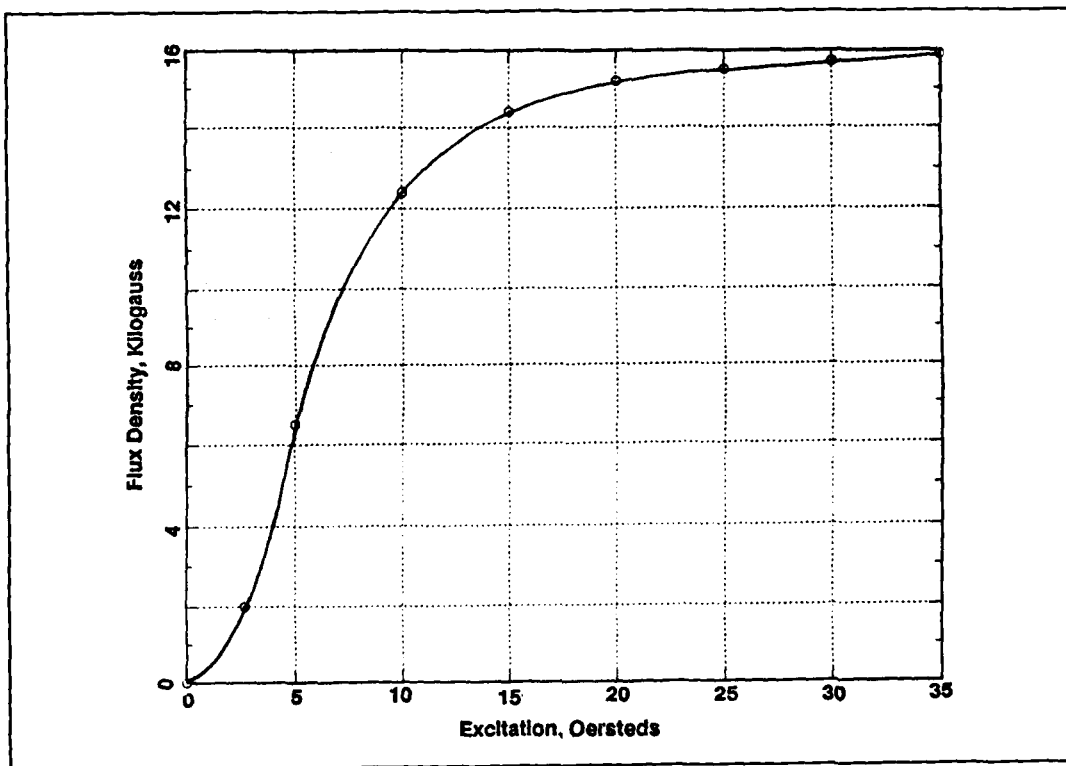
Figure 10. Magnetization curve, cast steel.
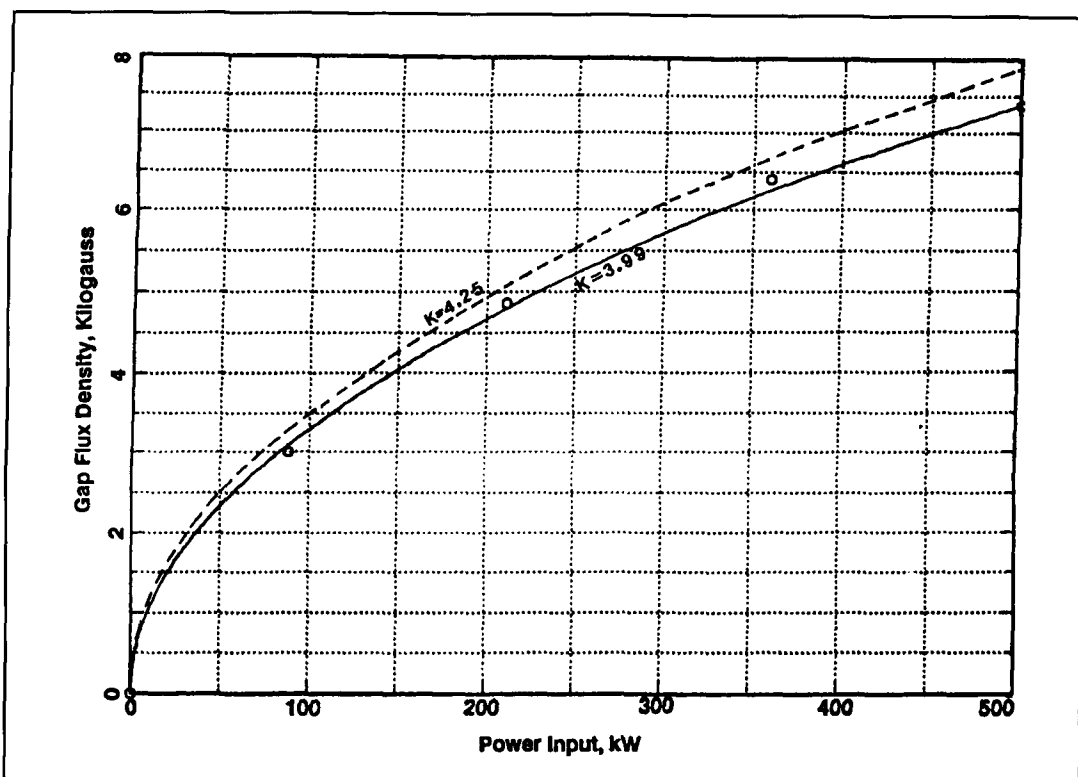
Figure 11. Arc characteristics—kerosene, wavelength 12,100 meters.

to circumvent paying royalties on existing patents. The Goldshmidt alternator (German) and at least one French design were developed to bypass the Alexanderson-Fessenden patents, for example. The era is peppered with patent-infringement suits involving nearly everyone, including Marconi. Fessenden was seemingly in constant litigation; he gave the U.S. Navy a great deal of trouble in this regard. Much of it, unfortunately, was justified.[8]

Fuller and the Federal Telegraph Company

Two men were predominant in the history of the Federal Telegraph Company. The first was C. F. Elwell, who founded the company in 1908, and who was the driving force behind its early development. Elwell negotiated for and purchased the U.S. rights to the Poulsen arc. The other was Leonard Fuller, who replaced Elwell as chief engineer. He applied different talents to the then problems of Federal, and carried the art of arc-transmitter design to its zenith.

Elwell had graduated from Stanford University, and was involved in the development of high-power electric furnaces. This work was being done on the campus proper. During this period he was approached by the Henshaw brothers, local bankers. They had

purchased a patent covering radiotelephony using a DC induction coil (with interrupter) which had a second primary winding to couple a carbon-button microphone magnetically with the spark circuit.[9]

Elwell was approached by the bankers twice, and refused the first proposal. He accepted, however, a second offer by the Henshaws while still occupied by the electric-furnace work. A trip to the East Coast to acquire furnace materials offered the opportunity to pick up components for a spark transmitter, which he did. His first work with the spark device was done on the Stanford campus. It met with disapproval from the Trustees, so he moved to a small bungalow in Palo Alto. By this time he was becoming very interested in radio.

His work with the patent (originally filed by a man named McCarty) was barely successful. He did manage to transmit somewhat intelligible signals over a distance of about one mile, but only when the spark gap was adjusted so narrow that it sustained an "arc"—normally an undesirable characteristic. The fact that the devices approached operability only when the gap was arcing did point to another technique which was not well publicized in the United States. A few technical articles, and several in the popular magazines, had been written about the Poulsen arc. Elwell picked up on them; they seemed to point to the next step.

Elwell cabled Poulsen and asked if the U.S. rights had been sold. They had not. An appointment was made, and Elwell, at his own expense, made the trip to Denmark. The original price for the rights was set at $250,000. When it developed that the arc transmitter could be keyed at a 200-WPM rate, Poulsen increased the price to a half-million dollars. That was too much for Elwell to raise. After failing to get financing (the Henshaws lost interest when the McCarty patent proved worthless) he finally negotiated (during a second visit to Denmark) for the rights by promising stock issues and $1,000 in cash. He also ordered two transmitters of 5 and 12 kW. The first transmitter he purchased at that time and brought back is the first arc introduced here.

Those are the essentials of the formation of what eventually became the Federal Telegraph Company. Elwell remained at the helm through 1912, improving on the Poulsen arc, principally in the provisions for cooling. Poulsen's transmitters had to be "rested" after continuous operation in order to cool down. Elwell eventually raised the power to 30 kW—as had Poulsen in Denmark. That was the end of the power escalation until further research and development.

Elwell had modulated the 100-watt Poulsen arc and others of Federal design. He had demonstrated it to local businessmen and many of the staff at Stanford University. He eventually established a network of stations extending from Northern California to Dallas, Texas, with Federal-designed arcs. In the beginning, the network was Sacramento-San Francisco-Palo Alto, and was a radio-telephone/CW circuit. This accomplishment enabled the financial backing necessary for future development. However, the growth of large arc networks occurred in the CW mode.

The U.S. Navy desired to establish a far-flung communications network. The Federal network by then extended to Honolulu, and was operational day and night—after the power level had reached 30 kW. This was not lost on the Navy. Though there were influential Navy officers who were hanging their hats on the large sparks, Elwell finally secured permission to put a 30-kW Federal arc alongside a 100-kW Fessenden spark transmitter at Arlington, Virginia. It was granted grudgingly. Elwell was told that "no nails or screws were to be allowed in the walls, floor or the ceiling." The Federal arc and all its equipment were built on a skid (a common technique today) and installed. Fessenden had required three years to complete his installation. Elwell's unit required only days.

The comparison was overwhelmingly in favor of the Federal arc. It produced less than half of the antenna current of the Fessenden spark, but its performance astounded the Navy personnel. The first day of tests enabled direct communication with San Francisco, and the
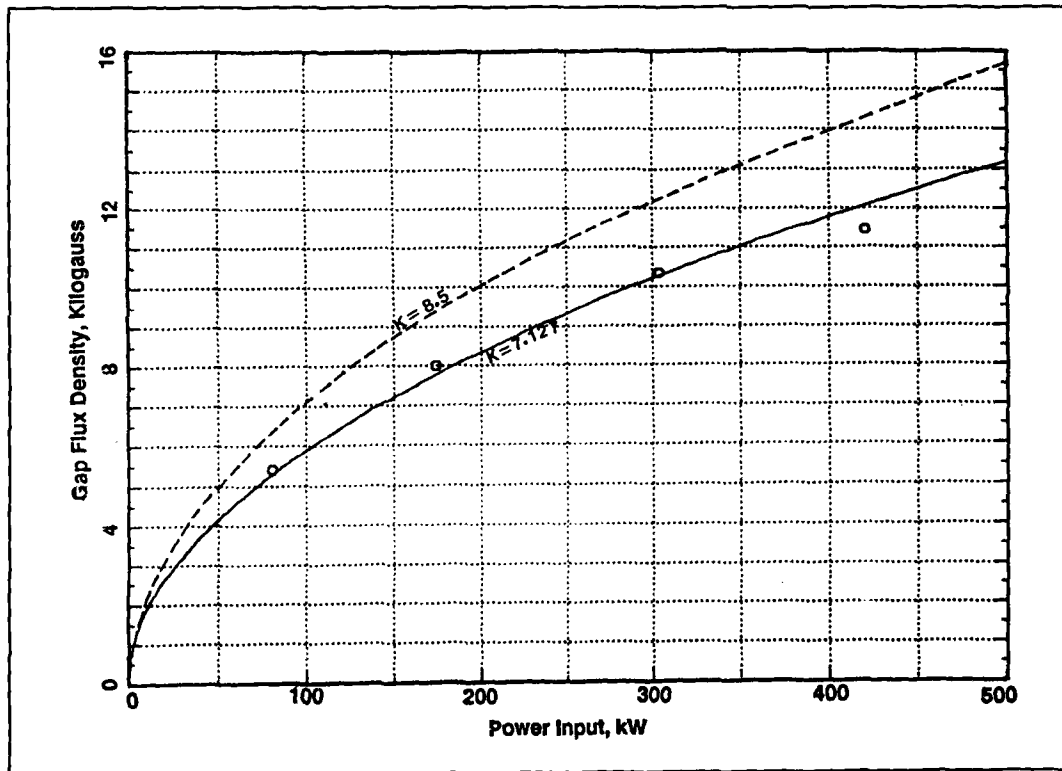
Figure 12. Arc characteristics—alcohol, wavelength 12,100 meters.
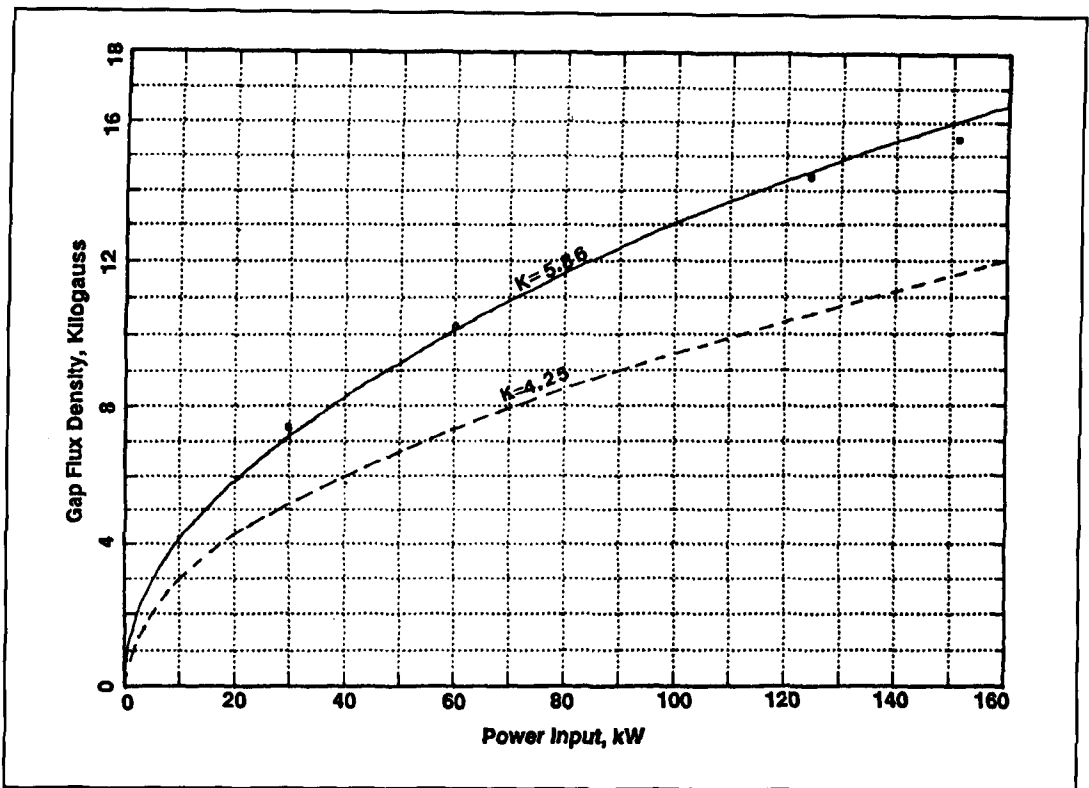
Figure 13. Arc characteristics—kerosene, wavelength 4500 meters.

next night, with Honolulu; the latter location *still in daylight. Of course, we know that* 100 amperes of antenna current distributed over 10 kHz of spectrum cannot compare with 50 amperes concentrated in only one or two kHz. The arcs were the nearest thing to "single-signal" communication to date. The Navy immediately ordered a 100-kW arc.

It was here that Elwell's engineering success was halted. His idea of power increase was to "use a magnifying glass" on the 30-kW arc design drawings. It was not successful. No increase in power output was obtained, though the input was raised to 60 kW. No one knew why.

The thorough engineering and research effort required for the answer was provided by Leonard F. Fuller, who had been hired by Elwell two years before. It was he, and virtually he alone, who determined what, why, and how.

Fuller initiated a thorough investigation of the relationship among the arc current, magnetic flux in the gap, wavelength (frequency), and organic used for the gap atmosphere. He was the first to do so. He found that, for instance, increasing the arc current with a constant magnetic field only doubled the number of residual ions; the field could not clear the gap quickly enough. So the flux had to be increased. He was on his way to the first comprehensive understanding of just how the arcs worked.

It was about this time that Elwell and the

board of directors of Federal had an unresolvable difference in just how the company should proceed. The result was Elwell's resignation. He had been among other things, the chief engineer. That mantle fell on Fuller. He was able also to interface with the Board in a more effective way than Elwell.

Arc-transmitter development reached its highest point under the direction of Fuller. From the time he took over the engineering aspects of Federal, he had increased the size of the transmitters from 30 kW to better than one MW. The path to this level is a classic study in competent engineering design.

The Federal arcs of sizes above 100 kW all had closed magnetic circuits, except for the gap itself. Fuller had determined the necessary fluxes for all planned transmitters, and had done enough experiments to enable him to extrapolate the power levels to as much as two and five megawatts This work was done by the use of a scale model that could be carried easily by one hand[10]—another indication of Fuller's engineering talent. The last two designs never came to fruition. It was too far into the age of vacuum tubes. The "E-frame" magnetic yokes—not laminated, but of cast steel—were as heavy as 60 tons in the 1-MW unit. Including the copper windings, the weight was about 85 tons.

The relationships among the main variables were determined empirically. First, however, it

is necessary to define what Fuller was working with. **Figure 10** displays the magnetization curve for cast steel.[11] The source from which the figure came had several such curves in one figure; values were picked off the original and replotted with the aid of a curve-fitting routine.[12] That explains the circular points on the curve. Notice that cast steel is very near saturation at 16 kilogauss. This will promote better understanding of the other figures.

Figures 11, 12, and **13** are for the two predominant arc organic atmospheres, kerosene and ethyl alcohol, and for two wavelengths (12,100 and 4500 meters, or 24.8 and 66.7 kHz, respectively). These curves, fitted from Fuller's published data from 1917–18, are but three out of literally dozens randomly selected from Fuller's 1919 IRE paper,[13] most of which figured in his Ph.D. thesis.

The figures containing the curves were generated by a curve-fitting routine on an MS-DOS PC.[13] It is possible to fit a smooth curve through every point for which Fuller tabulated data. **Figure 14**, a replot of the points in **Figure 13** fitted to a cubic polynomial, demonstrates that it can be done. These "fits" are perhaps useful for predicting the necessary magnetic flux for a given power input for the atmosphere selected, for a specific transmitter. They do little, however, to corroborate the theoretical aspects postulated by Fuller. The coefficients for the cubic (third-order) polynomial don't mean anything. There is nothing in the theory

as it existed then to justify any term that involved the cube of the input power. On the other hand, the expressions that Fuller generated and which are described below do show (1) that the shape of the curve is correct [that is, the points can be described by the quadratic of **Equation 12**] and (2) they demonstrate the uncertainties in quasi-scientific measurements made on a large, operational transmitter, a 500-kW Navy unit at Pearl Harbor. Fuller confirmed that it is difficult to control the arc atmosphere to "laboratory standards" in a large operational unit. He tried to minimize the uncertainty by using much more kerosene and alcohol than usual for operation; this, to approach the lab conditions he would have liked to have. Even so, the flux predictions were good to within only 18 percent.[14]

There are probably other reasons. Although ethyl alcohol is a "pure" hydrocarbon, it always exists as the azeotrope unless the atmosphere is carefully controlled (dry); that is necessary for the preservation of "absolute" (i.e., 200 proof) alcohol. In practice, it therefore has up to 5 percent of water in it, and that winds up in the arc chamber. Kerosene is a "cut" from a refinery fractionating tower. While it centers about C_{10}, there are lighter and heavier hydrocarbons present: a few percent of C_9, as well as C_{11}, and probably others as well.

These cases were selected to demonstrate the limitations of arc transmitters of *any* power input, at least for the two atmospheres listed
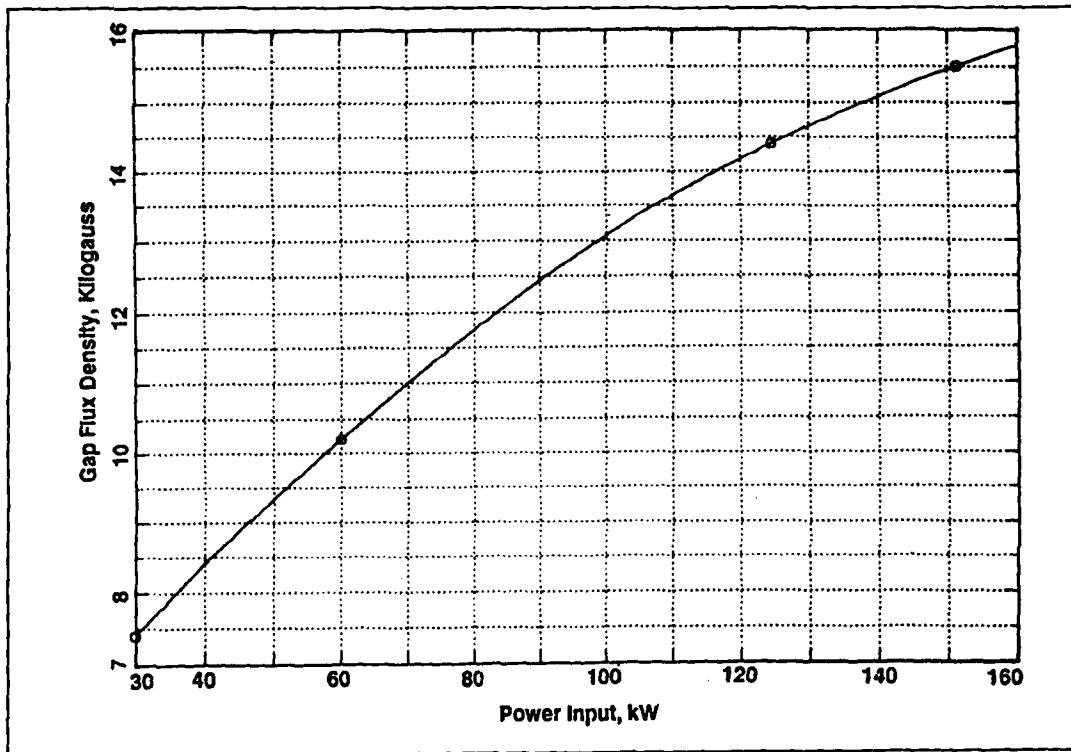
Figure 14. Arc characteristics—kerosene, wavelength 4500 meters.

and for the specific design of the converter in use at Pearl Harbor. It will be seen that for the wavelengths of 12,100 and 4500 meters, using kerosene in the arc chamber, the required magnetic fluxes in the gap are 4 and 15.4 kilogauss, respectively, for 150 kW power input. The data reveal that to increase the power above 150 kW at 4500 meter wavelength would require (on that transmitter) more flux than the saturation flux of the yoke itself. (Use the solid lines on the curves for extrapolation.) The dashed lines represent the constants (K) from **Equation 12**.

Now let us examine the curve for alcohol at 12,100 meters (**Figure 12**). At 450 kW, the required gap-flux is about 11.6 kilogauss. For a kerosene atmosphere the required magnetic flux density at a wavelength of 12,000 meters is only 7 kilogauss. That explains why Federal and the Navy used kerosene in all of their high-power transmitters. It is possible that Fuller's follow-on designs for multi-megawatt transmitters would have used hydrogen.

Fuller did a remarkable job for Federal as chief engineer. His approach combined considerable engineering talent with a good understanding of the theoretical aspects of the devices with which he was working. At the start, there were so many variables that it was difficult to predict arc-converter performance; he resolved much of the mystery. Though Fuller knew more than most people then constructing arc transmitters, there were still other aspects that eluded his total understanding of their behavior. Understanding of major principles nevertheless he had; there is no other way to explain his success.

It had been known for some time that, for a fraction of every cycle, the arc is extinguished. The time available for complete scavenging of the gap was therefore inversely proportional to the frequency. On the other hand, the shorter the available time, the stronger was the necessary magnetic flux. It can be said, therefore, that the requisite magnetic flux β_g, is directly proportional to the frequency. The "frequency" of a radio wave in those days was stated as its inverse; wavelength was the preferred parameter. Therefore,

$$\beta_g \propto 1/l, \tag{6}$$

where l is the wavelength in meters.
In Fuller's own words,

The molecular velocity of the atmosphere in which the arc burns controls the value of β_g. To extinguish the arc properly and deionize the gap in the time available, there is no gain in raising the ion velocity [above that] necessary to break up the ionic stream in the time allowed. Hence, if the molecular velocity of the [atmosphere] surrounding the arc is high, it is unnecessary for the magnetic field to increase the velocity as much as would be the case

were their velocity lower. Therefore the necessary field strength is inversely proportional to the molecular velocity of the medium surrounding the arc.[15]

$$\beta_g \propto 1/r, \tag{7}$$

where r is the molecular velocity.
Fuller further explains that the temperature of the flame is so high compared to that of the surrounding gas that no appreciable error is introduced by assuming that the absolute temperature of the arc flame is proportional to the input power $(E_{da}I_d)$. It follows that

$$r \propto \sqrt{E_{da}I_d}. \tag{8}$$

Since the best (just adequate) strength of β_g is proportional to the electrostatic field tending to maintain the arc, it also follows that

$$\beta_g \propto E_{da}. \tag{9}$$

The number of ions to be removed from the gap is proportional to the arc current I_d. Therefore,

$$\beta_g \propto I_d. \tag{10}$$

From equations 9 and 10,

$$\beta_g \propto E_{da}I_d \tag{11}$$

Now an expression for the required magnetic flux can be stated:[15]

$$\beta_g = (K/l) \sqrt{E_{da}I_d} \tag{12}$$

Where:

β_g = Gap magnetic flux density in kilogauss,
K = A derived constant averaging 4.25 for kerosene and 8.5 for ethyl alcohol,
l = Wavelength in kilometers,
E_{da} = Arc supply voltage in volts, and
I_d = Arc current in amperes

The increase in soot deposit resulting from the use of kerosene was accepted. Alcohol, with its lesser soot problem, could be used at longer wavelengths. The results are one of two possible main reasons for predominant communications remaining at VLF. The other has to do with propagation and reception under water—the latter a Navy requirement.

Equation 12 implies that not only must the antenna loading-coil taps be changed for a frequency shift; the magnet-coil taps had to be changed, too. Fuller called it "tuning the magnetic circuit." Those transmitters apparently were not easily "band-switched."

Fuller was at a technical disadvantage in two areas, and had to work around both of them. First, Federal transmitters were not coupled magnetically to the antenna because a Marconi patent (the "four-sevens," British patent No. 7777) existed. The antenna remained directly connected through loading coils to the convert-

er anode. It prevented Federal from gaining the harmonic reduction available in a double-tuned coupling system.

Second, the heterodyne local-oscillator system had been patented by Fessenden. Elwell originally had assigned him to investigate those circuits. It is interesting to note that Fuller had worked for the National Electric Signaling Company (NESCO), the residual of Fessenden's original company in New York. He actually was working on the heterodyne system. He was hired by Elwell from NESCO, at Fuller's instigation—he had written Elwell, asking if there might be a place for him with Federal. At the same time Fuller joined Federal, Lee de Forest was associated with the company; indeed, he devised the oscillating Audion circuit at Palo Alto. One of de Forest's associates, incidentally, developed the motor-operated "tikker." Fuller's receiving apparatus used both systems.

Thus the major manufacturer of arc transmitters (converters) was the Federal Telegraph Company. It supplied many units to the Navy. The smallest commercially available converter sold by Federal was of 2-kW size. Many units of this size were in shore stations. The U.S. Government also purchased several hundred converters of 5 to 30-kW size for ships, both Naval and commercial; the latter were being pressed into wartime service.

The smaller converters were started by operating a spring-return handle or a push-button attached to the cathode. It allowed the electrodes to contact momentarily to start the arc. In the larger units a separate smaller arc was ignited near the main electrodes, ionizing the gap, and thus starting the main arc. Even this auxiliary arc-starter was remotely operated. Manually shorting a 2000-volt, 500-amp DC power source is intimidating in its own right, and that does not say anything about the inevitable explosion in the chamber on startup.[5] **Figure 15** shows the circuitry of a typical arc transmitter. A chopper was used to modulate (actually to key) some smaller transmitters. The chopper was necessary because passive crystal and other diode-type detectors were still much used.

There are many records and photos of 2- to 30-kW arc transmitters, but the most powerful transmitters in the prealternator, pretube era were of the arc type. Tube transmitters did not approach their power levels for many years— and it was two more decades before those power levels were surpassed.

A one-megawatt arc converter was furnished (as the Annapolis transmitters in **Photo A** were) by Federal for the Bordeaux station. This station was conceived in 1916, intended to provide military communications between the U.S. and the AEF in France. In late 1918 hos-
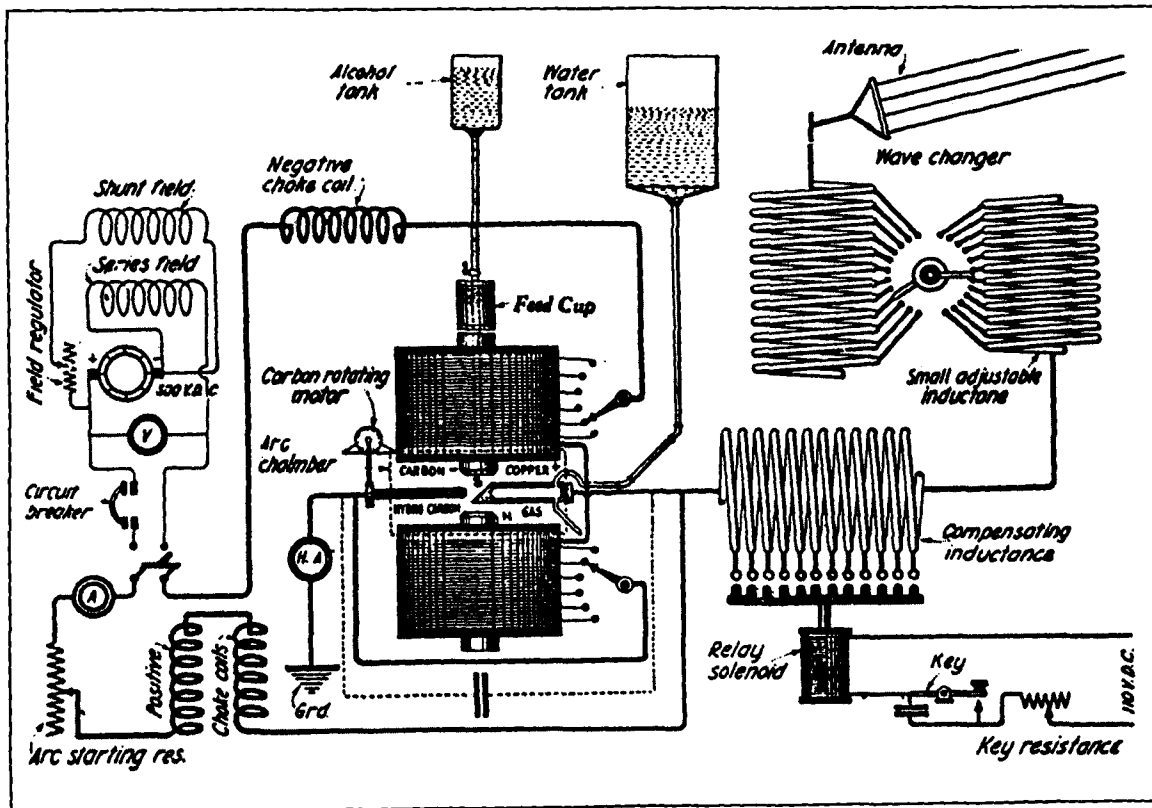
Figure 15. Typical arc transmitter.
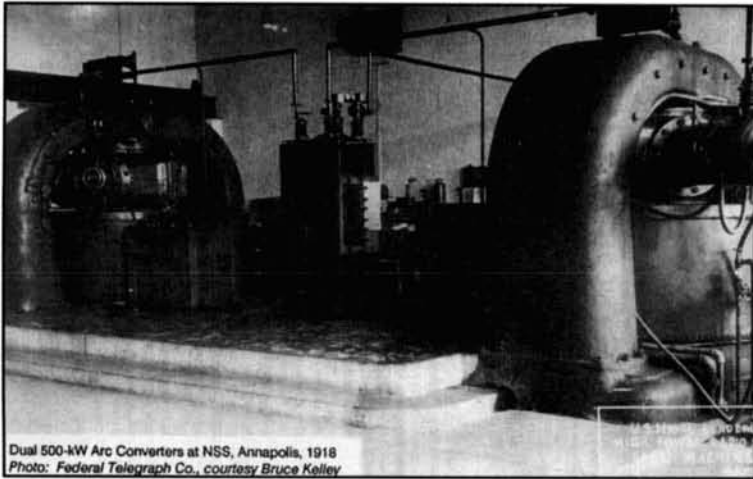
Photo A. Dual 500-kW converters at NSS, Annapolis, 1918. Photo: Federal Telegraph Co., courtesy Bruce Kelley.

tilities ended, and it appeared that the project probably should be shelved. But the French were very interested in the station, named "Lafayette Radio," and it eventually was built at Croix d' Hins near Bordeaux. It was completed in 1920.[16]

The Lafayette station is unique in history. There are many facets to its intent, design, construction, and operation, in addition to its being the most powerful transmitter in the world at the time. It would be, even today, among the "power-houses" of VLF transmitters. Occupying the same part of the spectrum as do modern NATO stations, it operated at a similar power level.

At the forefront of U.S. VLF activities was the Navy. The original conception of the Lafayette station involved the U.S. military. As the only knowledgeable organization here, the Navy became responsible for the transmitter, the antenna, and its support structures. The design concept—indeed the plant design itself—was attributable eventually to the Navy. Beyond 1919 (the year of the resumption of construction) the French were heavily involved, and one does not have to be reminded that Chauvin was French!

The agreement between the U.S. and French governments stipulated that the French were to assume all costs of construction; the U.S. Navy was to be the prime contractor. Transmitter and antenna design was delegated to Federal. The Pittsburgh-Des Moines Company, steel fabricators, was subcontracted to erect the tower supports for the antenna. Uniformed Navy personnel and some civilian engineers completed the organization.

Eight antenna-support towers were arranged in a quadrangle 1312 feet 4 inches by 3937 feet. Each was 820 feet tall; the tallest structures in the world after the Eiffel Tower. These supported the enormous top-hat for the antenna.

The station was completed on November 15, 1920, and was formally turned over to the French on December 18 of the same year, after having transmitted receivable test signals to a distance of 12,000 miles. Its total cost was $3,500,000 in 1920 dollars.

There is more to the story, however. The designer of the antenna system was none other than L.F. Fuller. Even so, the French carped continuously at the antenna and ground-system design. They flatly refused to accept it, and assumed total responsibility for that part. The Federal design was installed eventually, with very minor changes, but not before the French had copied Fuller's drawings line-for-line. The title "Federal Telegraph Company" was removed from the drawings, replaced by "designed by Captain Brassier."

The French were miffed at the installation of a powerful Poulsen arc machine. They would have preferred a rotary machine of their own design. Their complaint was that the harmonics from the powerful arc transmitter would be heard all over France. They were probably right!

Keying of arc transmitters

Arc transmitters were keyed continuous-wave devices. Quite a few were modulated by voice and music; they were modifications of CW transmitters. The arc has been shown to be extinguished during part of every RF cycle. The arc current supply cannot be cut off, however; the extinction is a part of the cycle-by-cycle physics of the device. Then how does one key the transmitter? A schematic diagram of the principle is shown in **Figure 16**. Other methods were tried; even keying the antenna via the ground lead. One method diverted the transmitter output via an SPDT relay to a circuit containing exactly the same R, L, and C as the antenna. The "dummy antenna" had to dissipate the full output of the transmitter, but it certainly eliminated the back-wave!

All methods presented serious obstacles; the antenna currents of the large transmitters were of the order of 1000 amperes. Even here, the large currents and high voltages switched (at keying speed, incidentally) would severely tax the relays. Federal merely broke up the total inductance shunted in the "key-up" position into several smaller intervals. If there were four relays, then each one would switch a fraction of the total antenna current, at open-contact potentials corresponding to the amount of inductance shunted by the contacts. Presumably this method also introduced a measure of reliability. The failure of any one relay did not shut down the station.

The antenna (and with it, the entire resonance system) was merely detuned during the

"break" (key-up) portion of the transmission, producing a shift in frequency and "back-wave." The age of FSK dates from that time; technically even earlier, say before 1908. There were some so-called "secrecy" systems proposed for spark transmitters using the same scheme. Successful application of FSK had to await better receivers.

Audio modulation

Arc transmitters emitted nearly pure CW. There were RF harmonics (of both the main frequency and the back-wave) and some phase modulation brought about by the inevitable arc instability, but those are of little consequence from the transmitter side when it is being modulated.

All of the modulation methods up to about 1918 or so were of the passive character; that is, they all absorbed energy. Even the magnetic amplifiers (saturable reactors) used with Alexanderson alternators were passive in the sense that they added no energy to the radiated signal. All of them, in one way or another, effectively placed a dynamic shunt on the signal at an audio-frequency rate.

The carbon microphone had been available for decades. It was natural that it should be tried on the new radio-frequency generators. It was used first in spark transmitters with very limited success, but the arc provided the first vehicle for successful application. At this time several new types of carbon-button microphones were developed [17]. There were other methods, too, some involving liquid-flow schemes, and even one which funneled carbon granules hour-glass style between two plates.

There was a fundamental problem that was very difficult to overcome. The resistance of the typical telephone button was (and is) about 50 ohms, and it could dissipate only two watts or so continuously. The circuits in which the microphones were operated were of much lower impedance. So lower resistances were necessary for a sort of impedance match. Some "RF" microphones were built to exhibit resistances as low as ten ohms. The main problem was that the button itself couldn't dissipate much power, no matter what its resistance. It was known that the equivalent resistance had to match the "driven" impedance; that corresponds to a loss equal to half the power in the unmodulated circuit.

Many circuits were used, most of which appear in **Figure 17**. The microphones in the figure are shown as single units. For the reasons explained above, many times they were actually sets of several such buttons connected in parallel or series-parallel; partly to match impedances, but mostly to dissipate power.

Photos of microphones of this period show different arrangements of multiple parallel buttons common then. One of them had 25 microphones in parallel! Any combination required that the audio sound power be diverted equally, and in phase, to each of the buttons. That meant that the connecting tubes had to be exactly the same length.[18]

Most of the transmitters were built with the microphone as part of the transmitter panel, for at least two reasons. The microphones were dissipating half of the transmitter output, and the connections to the buttons were in RF circuits. The higher the power of the transmitter, the higher the heat dissipation became. It would be frustrating not to be able to hold the microphone.

There are records of microphones with cooling fins or water-cooling. Some had centrifugal blowers. At least one scheme involved the rotation of spare microphones in a carousel or "lazy Susan." In this one, there was a two-minute allowable limit before the next set of microphones had to be selected. Such methods were necessary to carry away the heat developed in the buttons.

Most of the successful voice transmitters probably did not exceed 100 watts input, though Fessenden was operating higher power levels with the aid of a "microphone relay." An example of such a device would be a regular button-battery-receiver system where a secondary, much larger, button was driven by what would have been the diaphragm of an ordinary headphone. He also experimented with a forerunner of the modern condenser
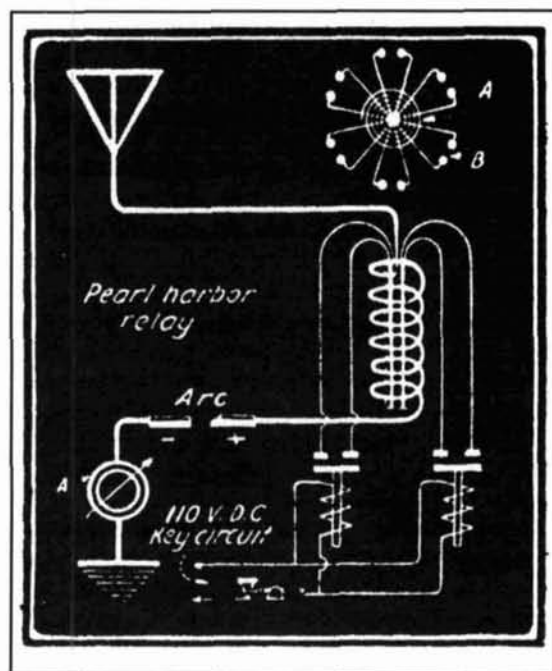
Figure 16. Typical keying system.
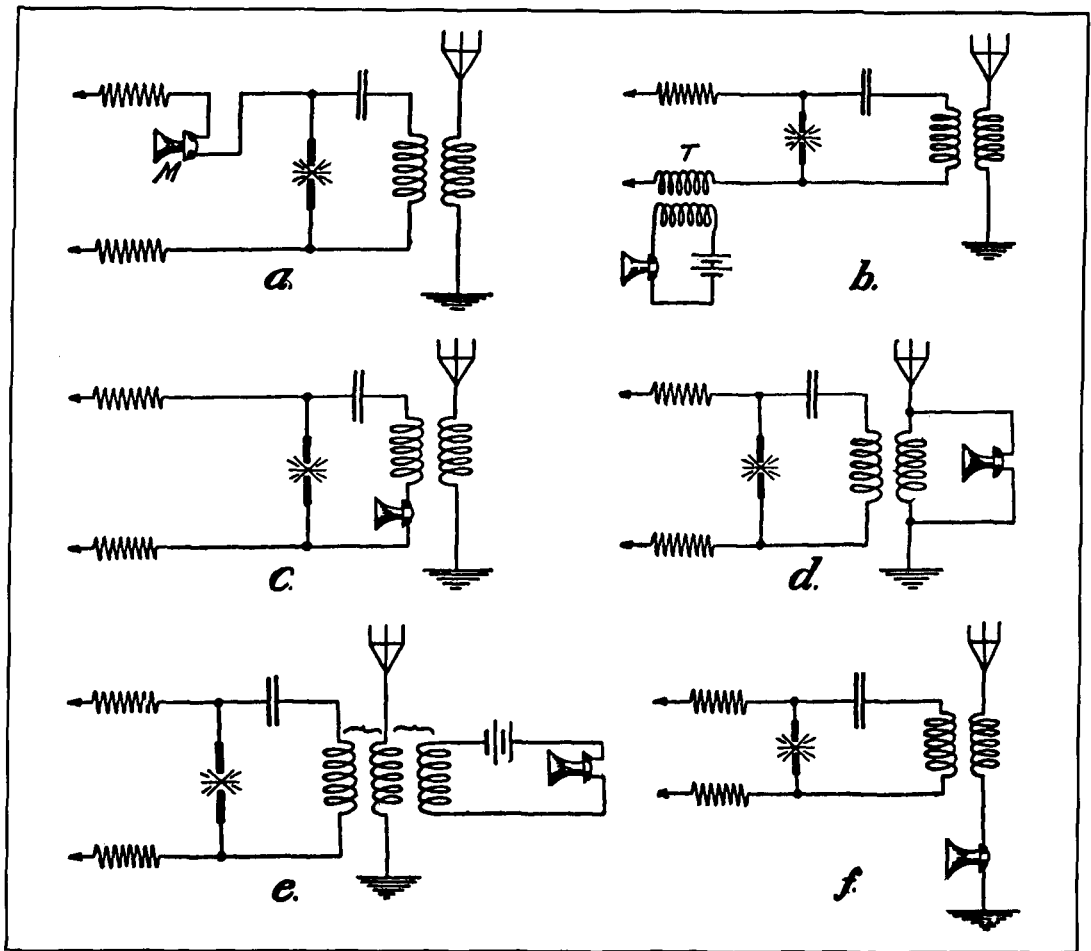
Figure 17. Circuits for modulating an arc transmitter (from Goldsmith).

microphone. Even with lower power, the microphones had to dissipate 50 watts in the usual 100-watt systems. Such transmitters did a good job, considering the frequencies used (about 15 to 30 kHz), the inefficient antennas, and their low power levels. Records exist of readable signals of distances up to 400 miles. There was one occasion in which one of Fessenden's transmissions was heard in Europe. The transmitting station didn't keep a log, so he couldn't prove that the transmission was made. What is unusual about it is that the power to the antenna was only six watts. It was a "freak." All of us, particularly amateurs, know how that can happen. However, barely readable signals were of no use to companies establishing commercial stations; the signal had to be of "commercial quality."

Arc *CW* transmitters at the time were capable of tens, and later, hundreds of kilowatts. The low power levels of the modulated arcs were strictly due to power-dissipation limits of the microphones. A version of the saturable-reactor used on the Alexanderson alternators could have been adapted for an arc transmitter.

Alternators could be modulated (even though it was "downward modulation") by a swing of a least *50 kilowatts*. There was no advantage in further development of modulated-arc systems. High-power transmitting tubes were in the near future, and they would effectively terminate further development of any other RF-generation method.

After the Bordeaux station, the days of new arc transmitter designs of *any* power level or mode of operation came to an end. A follow-on design by Federal for a 2000-kW arc was abandoned in 1920, as was a 5000-kW converter design concept. There is no reason to doubt the probable success if they had been built. Fuller knew how to do it.

Reception

Modulated signals could be received by crystal detectors from the very beginning; they were envelope detectors. Arc and other CW signals, however, required receivers of comparatively sharp selectivity in addition to some

kind of heterodyne or oscillating detector. There were some exceptions; devices called "choppers" and/or "tikkers" were introduced at one time or another in transmitters of lower power and in receivers. A chopper was merely a rotating device having a barred commutator which broke up a transmitted signal at an audio-frequency rate. It didn't actually interrupt the antenna current; it was placed across, or magnetically coupled to, the same coils as the key (and sometimes in series with it) which produced the same FSK effect as does the key alone. It was a modulating system.

The "tikker" usually was a rough wheel and spring brush, to accomplish a similar effect for receivers. With a tikker, one could receive CW—after a fashion—with a crystal detector. Those receivers lacked the selectivity to separate two RF signals close in frequency, and so could not suppress the back-wave sufficiently. The receiver tikker did work reasonably well on vacuum-tube transmitter signals which operated without the back-wave. The later Federal arc stations used both the tikker and the oscillating Audion which de Forest had discovered while he was working at Federal in Palo Alto. So receiver incompatibility is not a modern phenomenon; it is nearly as old as radio itself.

Auxiliary uses of the arc

Arc converters of small size were utilized, also. It was stated above that effective reception of arc CW signals required a heterodyne detector. A separate small converter was used for the injection signal; Fessenden was doing this as early as 1908. He also patented the method; it was the major factor in Federal's concentration on other methods to detect CW. These small units also served as signal generators for circuit alignment or for any such tasks as were required in operation of then-modern stations. They served the same purpose as do our own signal generators, although they were not nearly as portable! Spark-station equipment was adjusted by using a simple buzzer in the same manner.

VLF antennas

Any transmitter, regardless of type, obviously requires an antenna. There hasn't been much discussion of VLF antennas (an esoteric art) in the present literature. As a matter of fact, antennas for those frequencies still present problems. It wasn't until the mid-30s that the advantages of radial ground systems were demonstrated. Perhaps the closest to modern practice was the ground system for POZ at Nauen, Germany. The antenna there was a symmetric single cen-

Photo B. A large steel casting for a megawatt-sized arc transmitter donated to Berkeley and Lawrence by Fuller.

ter-mast umbrella about 1000 feet in diameter, and the ground was a circular mat buried to a depth about a half-meter. There is no description that tells whether it was a radial system. It is believed that it was constructed of square mesh, however, about 3,000 feet in diameter.

Consider the problem in more familiar terms. Reduced to a size commensurate with amateur frequencies, say 3.8 MHz, the VLF antennas corresponded to verticals two feet high, and a ground system that might be, at most, three feet in diameter. The radiation resistances were usually between 0.03 and 0.05 ohms; never higher. Feedpoint resistances were typically a few tenths of an ohm, still resulting in hundreds of amperes current at that point—and radiation efficiencies generally less than ten percent.

Dozens of ground rods, doctored up with "blue vitriol" (copper sulfate), and usually installed in a pattern of stars, were common for ground systems. Then there were "stars of stars" that compromised the total current-collecting system. Some actually used buried conductors (notably on "Ls") where parallel 1000-foot-long wires laid, fish-bone style, at right angles to the axis of the L, along its length. The concentration of return current at the stars occasionally started brush fires; the buried wires apparently had no such problems. Alexanderson developed the multiple-tuned antenna to reduce ground loss. It, too, was related to return-current concentration and antenna radiation efficiency. It is interesting that reproductions of the "rod and salt" grounds

were found in many publications and were promoted for amateur use up to the '50s or '60s. They really were inappropriate for anything but lightning and fault-current grounds at MF and above, amateur operation on long wavelengths (where rods were at all effective) having been prohibited for decades.

The largest antennas presented a capacitive reactance corresponding to that of about 0.05 microfarads. *The objective (and it was proper) was to make that capacitance as large as possible.* Therein lies the reason for the great lengths of the "inverted Ls," and the large diameters of the umbrellas. There was no great directivity in the Ls, despite some ideas that still exist. They were essentially top-loaded verticals, and thus almost completely nondirectional. The fact that the antennas did occasionally exhibit some directivity was certainly the result of differences in the nearby (off-site) soil conductivity—and it was minimal.

Given the dimensions of those antennas, their natural resonant frequencies were at least two times higher than the operating frequencies; let's assume twice as high. In the Ls used for the Alexanderson transmitting stations, the base-loading coils were constructed to have a reactance equal and opposite to that of the top-capacitance, *thus achieving resonance. At VLF,* then and now, Litz wire (litzendraht, or "stranded wire")—or rather ropes composed of Litz wire—enabled the construction of base-loading coils with Qs of between 5000 and 10,000. The radiation resistance and the inductance of the vertical portion of the antenna were comparatively insignificant. Therefore one probably can say safely that *all* of those antennas were equally top and base-loaded, with the radiating portion—*the vertical part only*—placed close to the electrical center of the system, or at 45 degrees. The vertical part was perhaps only three degrees long (out of 90 degrees, since they were all Marconis). It meant also that the voltages over the entire exposed system were nearly the same value, and that was usually upwards of 100,000 volts; higher in some. Consider further that even if a ground system of *zero resistance* existed, and with a base-loading coil with a Q of 10,000, at a frequency of 20 kHz the maximum antenna efficiency would be only 65 percent. That's because of the very low radiation resistance of VLF antennas.

Antenna wires were numerous, long, and correspondingly heavy. Even without considering ice-loads, the insulators were called upon to support heavy weights. In installations such as the Ls, the forces were directed largely downward; the wires were hung from horizontal structures. The insulators used in the umbrellas and other systems where there was a combination of wire weights, as well as the forces resulting from catenary suspensions, were loaded more heavily. Many times the insulators had to be installed in pairs or triples, just to accommodate the load. The leakage resistances of these insulators were therefore in parallel. Some antennas used hundreds of insulators, each dissipating perhaps several watts. None of the design problems in those days was solved easily. They *were solved,* however. The art of insulator construction steadily improved, and corona rings date from that period. Fuller himself was involved in insulator research and design for a period after he left Federal Telegraph.

Ice storms still make problems for stations located at higher latitudes. Today's antenna systems are designed so that they can be lowered to the ground. Resistance heating of the antenna wires is used in some places; that art originated in the Teens. Multiple-wire feeds were used, and the far ends usually were connected together. This allowed a low-voltage, high-current transformer to be connected to them, producing ohmic heating that melted the ice. Some umbrellas were built deliberately to exhibit three-fold radial symmetry; three-phase power then could be applied efficiently to them. In either case the transmitter must be shut down during an ice-storm. Again, the ancients *produced a solution that we still use, eighty* years later.

The end of an era— and the beginning of another

The age of arc-transmitter development spanned about 1900 to 1920. That is a short time in the scheme of things. The early part of this century, say from 1900 to 1930, was a period of tremendous technical and industrial advancement. The arcs were only a part of it.

So was the development of high-vacuum, high-powered transmitting tubes. It was this development that heralded the end of the arcs as well as the sparks, and that of the mechanical alternators, though the arcs operated through the Thirties, and a few of the Alexandersons were working throughout World War II. There were even some residual emergency spark transmitters (on 500 kHz) on the hundreds of cargo ships that operated—and were lost, too—throughout WW II. To illustrate the acceleration, consider that the time from the early radio experiments to the megawatt-sized arc transmitters (1900–20) was exactly as long as that from the first experiments with large high-vacuum tubes to the appearance of the cavity magnetron (1920–40).

In the later years of Leonard Fuller's career, he was head of the Electrical Engineering Department of the University of California at

Berkeley. Ernest O. Lawrence was on the physics faculty at the same time. The science of nuclear physics had advanced dramatically, starting shortly before the publication of Einstein's landmark paper on the Special Theory of Relativity in 1905, and continued with the published works of Becquerel and the Curies.

Lawrence was performing fundamental work in high-energy physics. It was his idea to use a combination of high-frequency RF power combined with a very high magnetic flux to accelerate ions into the mega-electron-volt range-the "Cyclotron" concept. During one lunch period at the University, he expressed to Fuller his desire to obtain a large electromagnet.

Of the four steel castings for the megawatt-sized arc transmitters, only two converters had been completed. The Chinese government had ordered a two-converter station—the common configuration—but it and Federal were unable to complete the transaction. Fuller was aware[19] that those sets of magnet castings had remained on Federal property for 18 years.

He subsequently arranged for the two castings to be donated, one to Stanford University, the other to Berkeley and Lawrence (**Photo B**). Stanford never used its magnet, and it is believed that the steel was sold for scrap. The other was well used. It became the magnet for the very first cyclotron. Fuller was well versed in magnet design. It has not been presented here in detail, but the design of the pole-pieces on all of the Federal arcs was very complex. Fuller participated in the design of the pole-pieces for the cyclotron, which were 40 inches in diameter. There could not have been a more fortunate situation.

The Federal Telegraph Company, along with Leonard Fuller, raised the art of arc transmitter development to its highest point, ending with the large station installed in France at the end of World War I. The art lay in limbo in succeeding years. Even so, the unique circumstances that existed when Fuller and Lawrence were together at Berkeley enabled the former to contribute significantly to the field of nuclear physics. Thus the art developed in the early field of radio—starting with unsteady steps early in the century—contributed significantly to the beginning of the Nuclear Age.

Acknowledgments

Besides a general interest in history, I have a special attachment to radio history. But to do an article such as this required not only effort from me, but a great deal of help from others.

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A 40-Meter Novice Band HBR

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Summary: The article describes a project involving a receiver for the Novice/Technician portion of the 40-meter band. The receiver is a single-conversion design employing low-noise mixers and affordable computer-grade crystals in the IF stage.

Please contact the author for additional information.

PRODUCT INFORMATION

HP Provides Multichannel CDMA Capability for Its Digital RF Signal Generators

Hewlett-Packard Company's line of digital RF (radio frequency) signal generators now includes multichannel code-division multiple access (CDMA) capability and a built-in dual-arbitrary waveform generator.

The new CDMA capability gives engineers working on CDMA base stations and mobile equipment the ability to test the devices under real-world conditions. With the optional new features, HP's ESG-D series provides internally generated, statistically correct multichannel CDMA signals. Up to 256 Walsh-coded channels can be generated to qualify components under realistic, worst-case conditions, such as a fully loaded base station with interference from adjacent base stations. This new capability is designed to reduce test uncertainty and conflicts with customers regarding different testing methodologies.

The enhancements come with an intuitive and powerful user interface for easy, one-button access to common CDMA signals that meet specific test requirements. Additional features of the HP ESG-D series signal generator include flexible CDMA-signal-simulation capabilities (for investigating and designing new communications systems or for testing proprietary or non-standard systems) and a built-in dual-arbitrary waveform generator (for generating complex).

The multichannel CDMA personality and internal dual-arbitrary waveform generator are now available for HP's ESG-D line of signal generators with frequency ranges of 250 kHz to 1 GHz, 2 GHz, 3 GHz or 4 GHz. Base prices range from $14,600 to $19,100. Upgrades are available for current customers of HP's ESG line of signal generators.

Specific information about HP's ESG-D line of signal generator products can be obtained on the World Wide Web at <http://www.hp.com/go/esg>.

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OBSERVATIONS OF 3/4-METER RADIO PROPAGATION ACROSS TEXAS

A quantitative approach

In the late 1970s, a group of hams began experiments in linking UHF amateur radio stations. We enthusiastically began construction and siting of some stations. Our naive approach probably kept us from discouragement given the task ahead. Daunting problems of beyond-the-line-of-sight (BLOS) path losses, cast-off equipment pressed into unintended use, audio distortion, and system control were essentially ignored. Like true hams everywhere, we just hooked it up and prayed for success. We were rewarded with innumerable site visits for tweaks and repairs, unreliable links, and an increasing appreciation of the need for a more quantitative approach.

When we started, we had no reference points. We knew that casual BLOS QSOs were possible, but how reliably? Duplex or simplex? Over what kinds of terrain? In all kinds of weather? Day and night? In all seasons? How good does the equipment have to be? How expensive? Can we make do with surplus land mobile radios? Are antennas and duplexers available off the shelf, or will we have to pay extra for "engineering" to get them to work in the low end of 420-450 MHz amateur bands?

The "usual" ham literature offered little that was helpful. Point-to-point linking wasn't a subject for ham magazine articles, or for chapters in amateur radio handbooks. At the time, we were unaware of earlier work in this field,

and had no idea how to deal with the piles of dirt that obstructed our proposed paths.

This article describes a quantitative approach born of our experiences. Others wishing to achieve linking over the radio horizon should be able to use our approach to evaluate their own existing or proposed systems. We'll compare the theoretical and actual capability of several UHF (425 MHz) duplex radio links. These links span the state of Texas over an airline distance of 908 kilometers (564 miles). The communications links are coupled in a chain beginning in southeastern New Mexico and terminating at Corpus Christi on the Gulf of Mexico.

The terrain varies from mountainous, with ground elevations as high as 2,040 meters (6,693 feet), down to 7.6 meters (25 feet). Ground conductivity includes that of salt-water marshes to volcanic or sedimentary rock. In between are grasslands, brushy semi-arid pasture, deciduous trees, and a few conifers, as well as shifting desert sands whose contours appear as dotted lines on geodetic survey maps. There's a repeater at each way point. Hence, this system provides an unusual opportunity for radio amateurs to observe and study centimeter wavelength propagation and to document diurnal and long-term seasonal variations.

We can readily compare predicted with actual performance. Since the links are in year-

Photo A. Remounting antennas. Guests from Houston, Ft. Stockton, El Paso, and Monahans, Texas, assist members of Intertie, Inc., in relocating antennas as the Mt. McElroy site in the Davis Mountains of west Texas, 6,693 feet, AMSL. This station overlooks Pecos and Ft. Stockton, as well as the junction of I-10 and I-20, en route from Houston and Dallas to Los Angeles. One of the telephone poles had fractured and feedline on one of the link antenna coax runs was soaking wet and corroded. We replaced all hard line runs and connectors. We secured each to the poles with #14 insulated solid copper wire, taped all connectors with Scotch 23™, then thoroughly varnished each taped connection with Scotchkote™.

round use, only free time and financial considerations constrain the frequency and duration of measurements. Short- and long-term variations can be observed within a broad range of geography and climate.

The system currently consists of 13 radio stations. The oldest link was established in 1975. The trans-Texas link was completed in the fall of 1994. In **Photo A**, you can see us hard at work relocating some of our antennas.

Background

In 1947, Bullington[1] thoroughly explored the state-of-the-art of propagation above 30 MHz. His paper contains a series of nomograms for path loss calculation. With these, we can compute the additional losses over free-space along a specific terrestrial path. Similar, but more extensive, graphic aids were provided by

General Electric Company.[2] They offer additional strategies to estimate the success of a proposed radio link.

In 1950, Day and Trolese[3] reported a unique set of measurements taken in the Arizona desert. They observed both line-of-sight (LOS) and BLOS paths over 43 and 76 kilometers (27 and 47 miles). Their antennas and equipment were mounted on elevators within 61-meter (200-foot) towers. With this equipment, they proved the existence of anomalous behavior of the index of refraction with consequent trapping of radio waves within ducts. Patterson[4] reviewed their data at a conference in Austin, Texas, which brought together many of the current contributors to BLOS literature. The syllabus references allow the novice to review UHF propagation literature from 1947 to the present. Also presented were papers from path measurements made in Texas and Alaska. Both these studies were conducted within the amateur 420- to 450-MHz band.

Of considerable value were the *Microwave Handbook*[5] from the Radio Society of Great Britain and recent monographs by Rohan[6] and by Doble.[7]

Propagation modeling

Propagation modeling is usually conducted with one or more models:

1. Standard propagation models. These simple models take advantage of the linear decrease in refractive indices at low altitudes. Models have been published which describe propagation in the near Fresnel zones, short over-the-horizon ranges and with tropospheric scatter.[1,8]

2. Non-standard models may include ray tracing, waveguide models, and parabolic equation models.[9,10,11]

3. Computer programs have been written that use more than one method. Two of these are described in the BLOS Conference (op.cit.); commercial programs are available.[12]

Our approach will be limited to the standard models. Bullington[1] states that the decibel path loss between two dipoles with increasing distance will be the sum of losses due to each propagation mode. The losses along a given path are multiplicative. Use of the decibel, a logarithmic (exponential) unit, allows simple addition and subtraction to replace multiplication and division. Path power requirements are thus separated into manageable sub-units. Each of these steps will be explained later:

1. Calculate the free-space attenuation.
2. Calculate distance to radio horizon from each end of the proposed path.

3. Calculate losses in between.
4. Calculate additional losses due to shadowing by obstructions.
5. Add safety factors based on probabilistic considerations.
6. Consider the possibility of anomalous propagation methods.

Free-space propagation. In a vacuum, radio waves propagate at the speed of light. Their power decreases as the square of distance. Therefore,

• Free-space signals are optically attenuated 6 dB for every doubling (octave) of distance. (This is a different way to use "octave"—a word usually associated with musical pitch or frequency.

• Free-space propagation is very predictable.

Diffraction. Because of diffraction (a phenomenon that occurs in a vacuum, too), the attenuation doubles again.

• The attenuation over plane or smooth earth is 12 dB/octave distance.

Scattering. Terrestrial propagation occurs within our atmosphere. In this situation, further propagation occurs as a result of scattering in the troposphere, the nearest atmospheric layer. The ionosphere is largely transparent to UHF radio waves; however, rain—and even oxygen—can account for additional attenuation at very short wavelengths.

• If scattering occurs, the attenuation increases 20 to 40 dB/octave.

Obstructions. Most obstructions will be apparent when viewed on a *path profile*, to be described below. Some obstructions will be obvious. However, we must remember that the wave front radiated by the antenna is football shaped (an ellipsoid with the two antennas at the foci of the ellipse). Reflection from the ground occurs not at a point, but over an elliptical area, essentially the first Fresnel zone. Simple geometric clearance is necessary, but not sufficient, for an ideal LOS path. This first zone is "bounded" by points for which the transmission path from transmitter to receiver is greater by one-half wavelength than the direct path.[13] We require that the rays clear the intervening terrain by 0.6 of the first Fresnel zone. The clearance is calculated by way of **Equation 1**:

$$h = 1316 \sqrt{\frac{d_1 d_2}{f(d_1 + d_2)}} \, , d_1 \le d_2 \tag{1}$$

where:

d_1 = distance to obstruction in miles from one antenna

d_2 = distance to obstruction in miles from the other antenna

f = frequency in MHz

h = 0.6 Fresnel zone clearance in feet
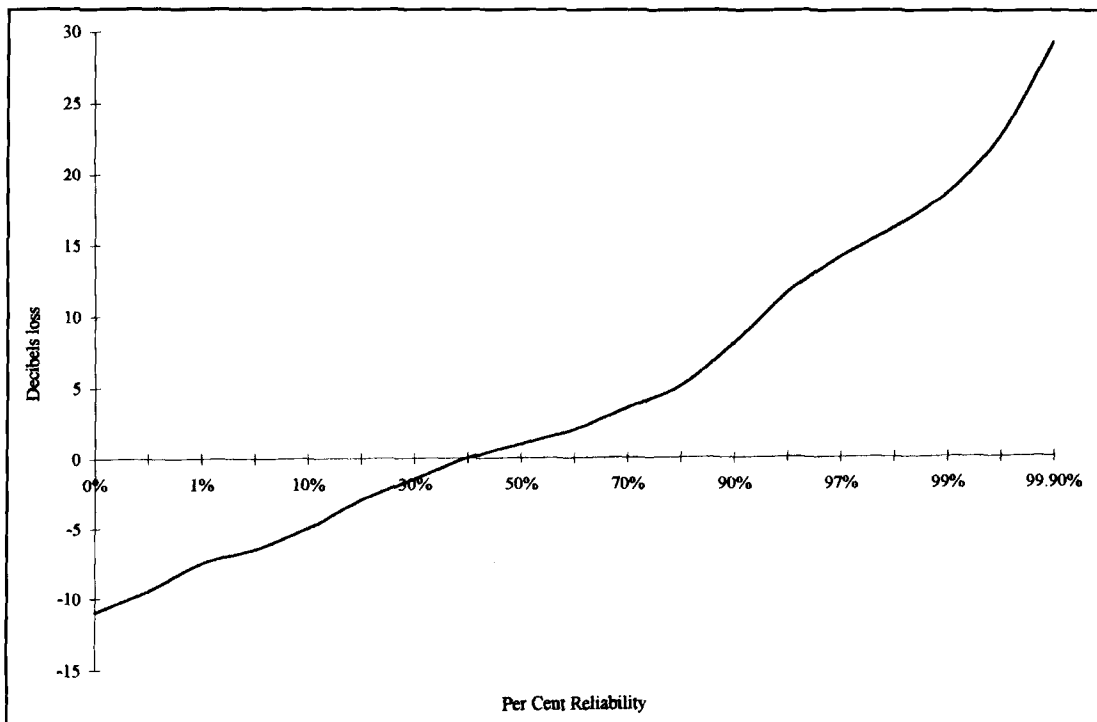
Figure 1. Percent path reliability versus decibels loss. This graph illustrates the Rayleigh distribution of path failures. For 90-percent reliability, the path loss capability (PLC) must be increased by about 8 dB. (Redrawn from Shepherd.[2]) The fade margin climbs very rapidly with small increments in percent reliability.

• If the necessary clearance is not present, additional losses will be incurred due to diffraction around the obstacle. Shadow losses may account for an additional 10 to 40 dB.

The sum of all losses, expressed in decibels, represents the power required to traverse the path. From the other side, the path loss *capability* (PLC) must be **measured** and compared with the **calculated** path loss. *Path loss capability* is the power available to traverse a given radio path. The PLC must at least equal the path loss, or no communications will be possible. Any excess PLC is insurance against fading from any source. We will look at this concept in detail below.

Reliability of Path. The formulas and nomograms in the literature typically provide for **50 percent** path reliability. To increase the probability of success over each day, and day to day, requires more power (best achieved with bigger antennas and better receivers). It has been shown that the "electromagnetic field scattered by a large number of small scatterers" has a Rayleigh distribution.[2,13] **Figure 1** illustrates this concept in relation to path loss cost and percent reliability. Note the sharply increased path loss required when we ask for near-totally reliable paths, and that for 0 dB margin, we only achieve a 50 percent reliability. We will settle for 90 percent, which carries a path loss penalty of 8 dB.

Probability of dropouts. Van Donkelaar[14] offers a formula for estimated annual path dropout, based on terrain, climate, distance, and fade margin. Note that distance carries a cubic relationship.

$$U = 0.07884 \; \alpha \; \beta f d^3 \; * \; 10^{\frac{-FM}{10}} \qquad (2)$$

where:

U = path downtime, seconds/year
α = terrain factor   β = climate factor
f = frequency in MHz
d = path length in miles
FM = fade margin in dB

• In summary, we assess the losses along a given path by determining the radio horizon for each station, the terrain between these two horizons, the 0.6 Fresnel clearances, additional losses due to shadowing, and the reliability loss factor. (In the special case of non-BLOS propagation, the path should still be tested for 0.6 first Fresnel zone clearance.)

Anomalous propagation modes. There is some good news! Sometimes, the path loss seems much lower than expected. A common reason for this anomaly is the phenomenon of ducting, in which the propagated wave becomes trapped in a layer of air of uniform

refractive index near the Earth. This area can be created by a number of mechanisms related to solar activity and the atmosphere.

This leads us to a discussion of **radio refractive index** (RRI). This is a measure of the bending of radio waves by the Earth's atmosphere. The refractivity of the atmosphere has been extensively measured as a function of altitude and time over most of the world. At sea level, the average index, *n*, is 1.00034. Because this is a cumbersome number to work with, a more convenient *N* has been defined:

$$N \equiv (n\text{-}1) \times 10^6 \qquad (3)$$

N has been found to be a function of atmospheric pressure, temperature, and water vapor pressure:[7]

$$N = (77.6 \; P/T) \; (3.73 \times 10^5 \times e/T^2) \qquad (4)$$

where:

P = atmospheric pressure in millibars
T = temperature, degrees Kelvin
e = vapor pressure in millibars

In a clear sky, the lapse rate is **linear** over the first kilometer of height, with a slope of *-40N units/km*. This refractivity, starting with *340 units* at sea level, results in the path of a VHF or UHF signal being bent toward the Earth's surface with a radius approximately four times that of the Earth. This results in propagation of the signal beyond the geometric horizon. For point-to-point work, this effect reduces the number of stations required. The variability of N can significantly affect propagation.

Because the radio refractive index is (after the first kilometer of height) nonlinear, and plots of this value are difficult to analyze under field conditions, a new variable, *M*, is defined as a function of height *h*, in kilometers, and is to be added to the N value:

$$M(h) = N(h) + 157h \qquad (5)$$

When *dM/dh=0, dN/dh=-157* N/km. In this situation, a plot of M against h is vertical and ducting will take place. When a perfect duct is formed, the loss within the duct drops as low as 3 dB per octave.[5] (Again, when speaking of distance, an *octave* is 2x the distance.)

The *k* profile is a second alternative to the use of N units. Bullington[1] defined *k* as the ratio of effective Earth's radius to true Earth's radius. Localized differences in atmospheric refraction may add to or subtract from this radius. With the addition of a refracting atmosphere, the effect is to extend radio wave propagation about 15 percent over that in a vacuum.

Table 1. Effects of Various Degrees of Refraction on Radio Propagation[5,7,14]

Atmosphere	dN/dh	k	Radio Horizon	Anecdotal Report
Super-refractive	<-40 u/km	>1.33	Extended	Enhanced distance
Standard	-40 u/km	1.33 (4/3)	≈15%>optical	Can't see; hears
dM/dh=0	-157 u/km	Infinite	Unlimited	Ducting
Sub-refractive	>40 u/km	<1	Decreased	See, can't hear

A current definition of $k(n)$ is:

$$k = \cfrac{1}{1 + \left(\dfrac{a}{n_s} \times \dfrac{dn}{dh}\right)} \tag{6}$$

Substitute $n_s = 1.0003$ and $a = 6371$ km; multiply numerator and denominator by n_s/A, or 157; and substitute N for n and we obtain a suitable formula for k:

$$k = \cfrac{157}{157 + \dfrac{dN}{dh}} \tag{7}$$

The expression dN/dh is called the **gradient of refractivity.** As noted above in the discussion of lapse rate, dN/dh is about -40N units/km, and when substituted in the formula above, yields a k value of 1.34 (about 4/3). This is the basis for the so-called "4/3 Earth profile" mapping of point-to-point links. This value for k is very nominal and may vary a great deal with geography and time of day and year. Small changes in k can enhance or spoil an otherwise usable path, particularly if this path grazes the Earth's surface. k is greatly affected by N, and N is a function of atmospheric pressure, temperature, and water vapor pressure. (cf. Equation 4 above). When N decreases more than 40µ/km, enhanced propagation results, because the Earth appears flatter to radio waves. When N goes positive, there is diminished propagation, a condition of subre-

fractivity occurs, and on-path obstructions appear taller. The variability of k with terrain is not at all well emphasized. In the sub-refractive case, it might be possible for aircraft to escape radar detection (radio "hole"), while they are optically visible.[9] **Table 1** summarizes these possibilities; **Table 2** shows k values as a function of geography.

Examination of **Table 2** suggests that our west Texas sites will have an average k of 1.17 to 1.20; the low mountains just east about 1.25; central Texas sites will probably be close to the Standard Atmosphere of 1.34; the Gulf Coast sites will be 1.50 in the winter and 1.60 in the summer. Hence, we will be penalized in west Texas, just where we need the best refraction; perhaps the low sand hills leading to coastal stations will not be as obtrusive.

Radio refractive index and anomalous propagation. There are four conditions (radiation nights, subsidence, advection, and frontal systems), which may modify the clear sky lapse-rate, or rate of change of radio refractive index.[4]

• **Radiation nights** occur with conditions of sunny days/clear skies at night. The rapid loss of Earth heat into the atmosphere causes the Earth to cool faster than the air above. The layer of air nearest the Earth will exhibit a linear decrease of 1°K/100 m, provoking a non-uniform lapse rate. This phenomenon does *not* occur over water, because water has a better thermal conductivity than Earth. These are the conditions observed by Ray and Trolese[3] in 1950 in the Arizona desert

Table 2. Typical k-values Arranged by Geographical Location*

k	Terrain
1.17	Dry mountainous areas above 2290 meters (7500 ft) elevation.
1.20	Dry mountainous areas between 1525 and 2290 m. (5000–7000 ft.).
1.25	Dry mountainous areas up to 1525 m (5000 ft.).
1.30	Inland plains during the winter.
1.34	Standard atmosphere k.
1.50	Inland plains and northern coastal areas during summer, southern coastal plains during winter.
1.60	Southern coastal areas during summer.
1.75	Extreme southern coastal areas during summer.

* Adapted fron Haarsager[15]

Table 3. Site Geographic Summary

Azimuth↓	Location	AMSL	HAAT	Distance	Azimuth↑	Site Abbr.
154°		SE NM	2118m		0	0°
123	McElroy	2040	440m	145km	334	MCP
99	Glass	1801	317	106	303	IMR
128	Pandale	729	125	177	279	PAN
99	Brackett	520	58	115	308	BKT
85	Muffin	351	59	64	293	UVA
45	SAT W	287	31	111	265	AFV
15	SAT N	294	32	23	225	SAT
135	Sierra	113	42	98	320	KDY
167	Odem	122	143	98	347	ODM
135	Corpus Chr	53	51	27	315	CRP
ΣDistance			964km			

(our west Texas sites should be subject to almost daily thermal inversions).

• **Advection effects.** These effects result from the sweeping of warm air from land masses over water. These tend to occur with the movement of air around an area of terrestrial high pressure. If a second land mass is nearby (50 to 250 kilometers), the mass of air can be swept over it and there is a second type of temperature inversion. These conditions were reported at 420.5 MHz at Prudhoe Bay, Alaska, by Kennedy and Ruper.[16] Our sites near the Gulf Coast could see similar advection effects.

• **Subsidence** occurs when descending air from a high-pressure system gets heated by compression. Although these phenomena occur at greater than 1 kilometer above the Earth's surface, the layer of dry air over cooler, moister air can descend much lower, provoking major propagation anomalies.

• **Frontal systems** can provoke localized inversions. A wedge of cold air may be driven beneath warm air as a cold front moves into an area. Because these fronts move fast, their effects are brief.

Atmospheric conditions for ducting have been touched on briefly at several points above. Ducting can have two effects: enhanced propagation or excessive losses (denial of path, in which transmit and receive stations are not within the same duct). Ducting also requires that the radiated wave "fit" inside. Ducts are like wave guides; the duct must be wide enough to allow propagation within it. Priedigkeit[9] reported the following empirical formula, also used in metric form by Kennedy and Rupar:[16]

$$f\,min = \frac{7.13 \times 10^6}{T^{1.5}\sqrt{-\frac{dN}{dH} - 48}} \qquad (8)$$

where:

f_{min} = minimum frequency in MHz

T = duct thickness in feet

dN/dH = gradient of refractivity in N units/1,000 feet

Substituting 425 MHz for f, refractivity of 160 units/1,000 feet, and solving for thickness,

Table 4. Path Loss Estimates

Location	Dist km	20 Log Dist	Loss dB	Excess Loss	Total Loss
SE N. Mexico	0				
Mt. McElroy	145	43.2	128	0.4	128.4
Glass Mtn.	106	40.5	125	0.6	125.6
Pandale	177	45.0	130	11.6	141.6
Bracketville	114	41.1	126	27.4	153.4
Uvalde	64	36.1	121	30.7	151.7
SAT West	111	40.9	126	32.2	158.2
San Antonio	23	27.2	112	-1.9	110.1
Kenedy	98	39.8	125	27.7	152.7
Odem	98	39.8	125	44.0	169.0
Corpus Christi	27	28.6	114	-3.7	110.

Table 5. Equipment Specificatons

Instrument	Manufacturer/Type	Accuracy	Precision
RF Step attenuator	Telomic Ind. TAD50-A	About 2%	±1 dB
Audio Voltmeter/Attenuator	Motorola S1500A	2%	± 1 dB
Signal Generator/RF Attenuator	IFR AM/FM 1200	±2.5 dB	±1 dBm
Directional wattmeter	Bird 43	5% F.S.	100 watt scale ± 1 watt
			25 watt scale ±0.5 w

we obtain a value of 136 feet to sustain communication. Priedigkeit's example used a 300-foot duct, which gives an f_{min} of 130 MHz.

Duct formation in Texas. Because of weather conditions prevalent in desert environments, ducting along our west Texas links should be common. During the day, the sun heats the desert floor to a high temperature. If there is no cloud cover at night (the rule in the desert) rapid cooling occurs after sunset, and then there is a layer of cooler air where an upper boundary forms in an area of refractive capacity, promoting propagation. Because narrow ducts occur more frequently than wide ones, and because duct propagation is wavelength dependent, ducting phenomena at higher frequencies are more common.

Anomalous propagation occurs with subsidence and advection. The former occurs with anticyclonic weather systems; the latter over water or wet areas. Advection could be a factor on our southern-most linkages near the Gulf of Mexico.

Tropospheric scatter. This mode requires a common volume of atmosphere within the bore sight of the two antennas. Although tropo scatter is incidentally possible in our links, we have not designed them specifically to take advantage of this propagation mode.

Method

As noted at the beginning of this article, our equipment is sited in a series of locations extending from the mountains of west Texas to the Gulf Coast. The backbone for this path contains 10 stations. Another series of links carries the signal north into the Texas hill country. **Table 3** summarizes the main linkage east from New Mexico to the Gulf. Prior to permanent equipment installation, every site was path-tested at least once to its proposed neighbors.

Our equipment is mainly reconditioned 30-plus-year-old land mobile FM transmitters and receivers whose performance is enhanced by receiver preamplifiers and homebrew transmitter power amplifiers. Band pass filters sometimes saw first service in the old 460-MHz Citizen's Band. In most cases, duplex capability is provided by small mobile band-reject duplexers, often augmented with additional

bandpass filters in transmit and/or receive legs as required. We use circulators and isolators wherever required to offset intermodulation interference. Our antennas are commercial-grade units. These and all duplexers are purchased new, but most of our feedline has seen 25-plus years service, usually in a different location, prior to acquisition. The entire system is remotely controllable from any way point.

Equipment

The receivers at Mt. McElroy, Glass Mountain, and Pandale are Motrac™ units. We have M series at all except Pandale, where there are L series radios. The Brackettville site has a GE MASTR Exec-II™. Preamplifiers, when used, are Advanced Receiver Research P432VD types*, or units designed and home-brewed by WA5VJB that have a measured NF of 0.9 dB and a gain of 20 dB.[17] Transmitters are either Motorola Motrac MST™ exciters (10 watts), or a hybrid of GE ExecII™ exciters driving Motorola MHW710™ "bricks" barefoot (10 watts). For QRO to 25 watts, we add an MRF 646 final.[18]

For antennas, we use Sinclair Technologies SRL 307™ series with one, two, or four bays (10, 12.5, 15 dBd), or the SRL 360 Radaflector™ (15 dBd). Feedline is 7/8-inch Andrew Heliax™, 1/2-inch for short runs. Duplexers are either Celwave™ PD-633-6A, Sinclair MR-356, or Wacom WP-678.

Isolators are made by EMR Corp™ (75514 IM panel). All in-cabinet RF connections are made with RG-142 or RG-214 double-shielded coaxial cable. All between-series adapters are shunned. Lightning arrestors are Celwave™ PD-395 or Poly-Phaser™ IS-50NX-C2. GE "Pre-Prog," Motorola™ T1500, or Decibel Products™ DB 4001 series bandpass cavities are used as required at each site. Bandpass filters are always used in front of the preamps, in series with the duplexer receiver port.

Measurement protocol

We made measurements every 90 days for two years on four sites of our 10-site backbone.

*Published specs: ARR™ P432VD:N.F., 1.8 dB; Gain 15 dB; P432VDQ (GaAsFET):N.F. <0.5; Gain 16 dB.
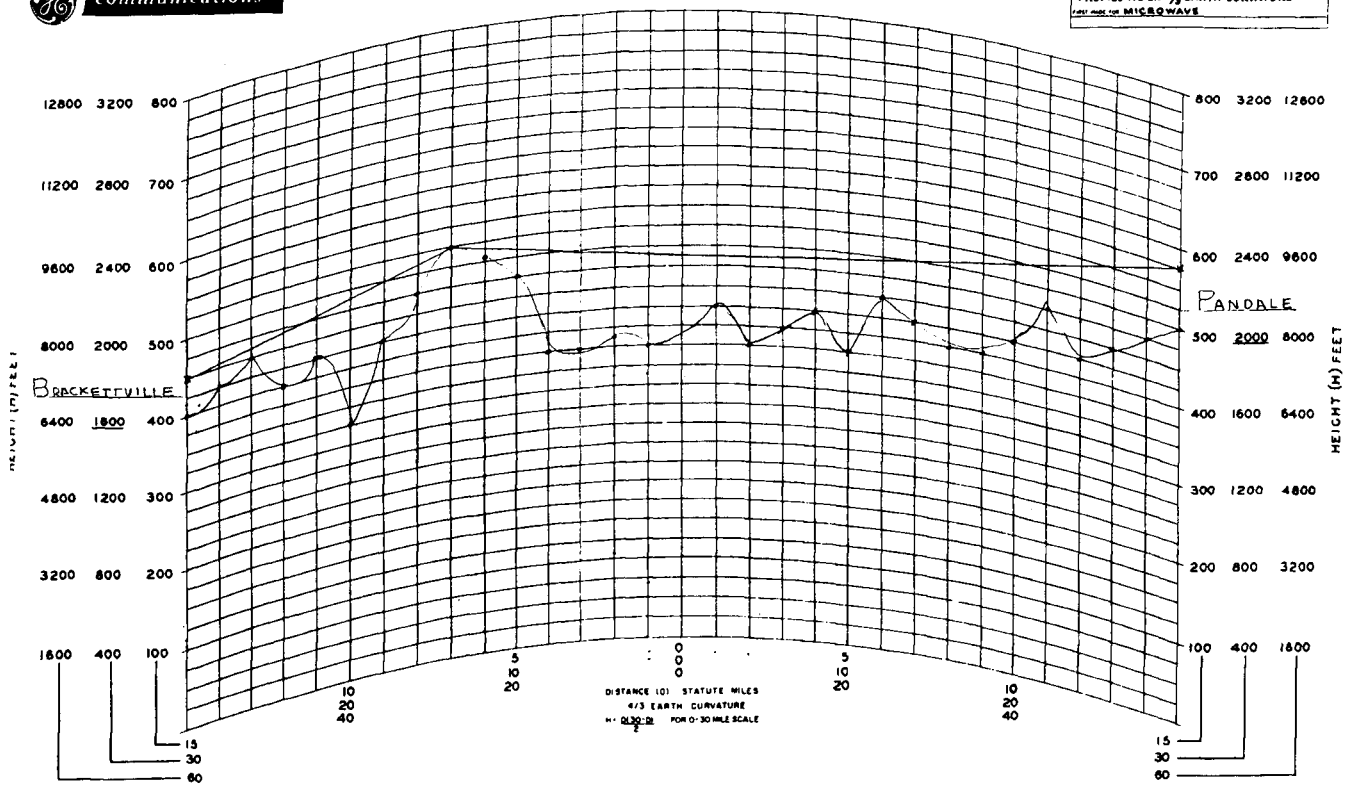
LOCATION	LEFT SIDE	RIGHT SIDE
LOCATION	Brackettville	Pandale
LATITUDE	29° 30' 28"	30° 08' 52"
LONGITUDE	100° 23' 51	101 19 33
REFLECTOR BEARING	308°	128°
GROUND ELEVATION *	1612'	2090'
DIPOLE HEIGHT	200'	300'

* REFERENCE - SEA LEVEL

CUSTOMER	Intertie, Inc.
PROP/REQ. #	San Antonio, Tx.
DATE	5-01-92
PROFILE MADE BY	LSH
SOURCE OF DATA	USCGS DELRIO & SONORA

SCALE USED	0-15 MI.	0-30 MI.	0-60 MI.

Figure 2. 4/3 Earth profile of proposed path from Brackettville to Pandale, Texas. The straight line connects the two antennas at the top of their respective towers. Note the obvious obstruction near the Brackettville tower. Not so obvious is the significant 0.6 Fresnel zone obstruction and the Pandale end of the link (see Equation 1). We apply this test to all but the most obviously clear line-of-sight (LOS) and all beyond-line-of-sight (BLOS) paths.

These measurements were made in clear, cloudy, and rainy weather, in calm and high winds, and in temperatures ranging from 4.4 to 35 degrees Celsius. Most observations were made at midday. Single measurement sessions were implemented at various times at the remaining sites (in preparation for a more deliberate schedule).

Our group performed the following activities:

1. We calculated the theoretical transmission loss for each path in each direction.

2. We measured Effective Receiver Sensitivity (ERS), Effective Isotropic Radiated Power (EIRP), and, from these data, computed the Path Loss Capability (PLC).

3. We measured the fade margin of each path and compared this with the PLC.

4. We assessed the possibility of anomalous propagation modes.

Later phases will address the paths from central Texas to the Gulf of Mexico. We will then compare the observed path transmission variability with our west Texas observations.

Estimation of path power loss from path plot-based calculations

In order to assess losses due to a rough Earth surface, we plotted the three-dimensional ray path from site to site in profile (two-dimensional). Ordinarily, this is done on special graph paper with a curved abscissa, representing the Earth's bulge. The radius of curvature is calculated at 4/3 of the Earth's radius. We purchased 1:250 000 U.S. Geodetic survey maps for each link segment. After carefully locating each projected site on the appropriate map, we drew a

straight line connecting each one. (Site locations were determined using the 7.5 minute USGS maps and the Decibel Products "Measuring Stick".*)

We will present the Brackettville to Pandale path here. This path was selected for further study because it shows serious obstruction and preliminary path testing was only borderline satisfactory. The altitudes of each site and of all land features along the way (determined from the contour lines) above mean sea level (AMSL) were transferred to the 4/3 graph paper, after the x and y axes were properly scaled. When these points are connected, we see a side view of the Earth's crust along a path. This approach avoids the calculation of the Earth's bulge as a separate problem. However, if special paper is not available, Haarsager[15] has a suggested formula.

The 0.6 first Fresnel zone clearances, as well as shadowing losses for in-path obstructions were added to path loss calculated for smooth Earth to each calculated radio horizon and for the zone between the horizons. Nomograms initially published by Bullington[1] (reproduced by Rohan[6]) were used to make these determinations quickly. Distance to radio horizon is graphically presented in Figure 7 of Bullington's 1947 article; his Figure 6 allows calculation of decibel loss relative to free space transmission at points BLOS over a smooth Earth. If 4/3 Earth profile plots are not used, his Figure 5 allows calculation of diffraction loss caused by Earth's curvature. Shadow loss relative to smooth Earth is obtained from his Figure 10.

Figure 2 shows the path plotted on 4/3 Earth profile graph paper. The path is 115 km (71.4 mile) long through fairly rugged country. The towers at each end are 61 and 91 meters tall (200 and 300 feet). Free-space attenuation over this path would be 125 dB. Radio horizon from the shorter tower is 32 km (20 miles); from the taller tower, 40 km (25 miles); the distance between is 44 km (27 miles). The dB loss for these three distances is 18, 3, and 22, for a total of 43 dB. Examination of the 0.6 first Fresnel zone clearance requires 91 meters (300 feet), which, from the path plot, is clearly not present. Thus, shadow losses must be added in, and are estimated at 10 dB. *This gives a total predicted path loss of -178 dB.*

We made similar determinations for the remaining links using a computer program[12] borrowed by an associate. These programs compute free-space losses, to which terrestrial losses are added. They operate from a digitized data base of altitude versus geographic coordinates

*Available from Decibel Products, Inc., 3184 Quebec, Dallas, Texas. Similar latitude/longitude rulers can be found a map shop.

and represent the data as a 4/3 Earth profile ray map with the pertinent numerical data thereon.

Results: path loss estimates

We compared the -178 dB figure with the PLC predicted from equipment in place at the two sites. From Brackettville to Pandale, PLC is 187 dB; in the opposite direction, it is 177 dB (detailed descriptions will be given later). Average measured fade margins are 14.2 dB for Pandale to Brackettville; 8.75 dB in the other direction, indicating conditions are better than predicted. A commercial program, using a k-value of 1.33, predicted 27.4 dB loss over 126 dB free space for a total of 153 dB.

Results: path power loss predictions

The formula for free-space loss is:

$$\text{Loss in dB} = 32.45 + 20 \log_{10}d + 20 \log f \quad (9)$$

d = distance in kilometers
f = frequency MHz

For these calculations, we will use a nominal frequency of 425 MHz (20 log 425 = 52.57 dB). As noted earlier, the term "excess path loss" was obtained from a computer-calculated path analysis at $k = 1.33$ (Standard Atmosphere). The 8-dB additional loss for 90 percent path reliability has not been added in here. **Table 4** shows these estimates.

From **Table 4**, we see that the shortest path is from Odem to Corpus Christi, and that it is essentially LOS, which accounts for the *negative* excess path loss. The worst path is from Kenedy to Odem. This path has a 0.6 first Fresnel zone obstruction just south of Kenedy (Pettus ridge) and continues obstructed for more than 1/3 of the distance to Odem. The link from San Antonio to Kenedy suffers because of two large humps of dirt. Brackettville looks at 0.6 Fresnel obstructions in all directions. Our longest path is from Glass Mountains to Pandale, a distance of 177 kilometers (110 miles).

The path grazes the Earth's surface most of the distance and has a 0.6 first Fresnel zone obstruction at the Pandale terminal. There are clear shots from SE NM to Mt. McElroy, Mt. McElroy to Glass, SAT to SAT W, and from Odem to Corpus Christi. Only SAT W to SAT and to Corpus Christi are LOS; the others are all BLOS. We now present the collected data for the path losses thus far calculated, site specific ERS, and EIRP, and PLC calculations, and compare the path predictions with our

measured fade margins taken every 90 days for two years.

Field measurements

Measurements were limited to those possible with relatively simple equipment, available to advanced amateurs. The precision and accuracy is for the most part, dependent on the quality of the three attenuators. Only the signal generator/RF attenuator carried a reference laboratory certification throughout the test period. All instruments were relatively old. The measurement equipment used at each site is listed in **Table 5**.

The same coaxial cable jumpers were used at each measurement session at each site.

Field measurement protocol. At each session, we tried to make the same measurements in the same order, such that cable plumbing setups were invariant.

• Order of measurement:

1. Transmitter output power, 50-ohm load. (Directional wattmeter) (*Tx Po*)

2. Transmitter output power and reflected power at the antenna port of the duplexer. (*Dplxr Po*)

3. Receiver measurements. (RF signal generator with RF attenuator, Audio voltmeter with switchable 20-dB audio voltage divider, part of Motorola S1500A test set):

a. Receiver 20 dB quieting sensitivity measured at duplexer antenna port. (*Basic Rcvr Sens*)

b. Receiver 20-dB sensitivity measured at 50-dB attenuator slug with 50-ohm load connected to directional wattmeter line section. (*Sens 50Ω*)

c. Receiver 20-dB sensitivity measured at 50-dB attenuator with antenna connected to directional wattmeter line section (*Sens ant*)*

4. System flat-fade margin[7] (RF step attenuator). We measured link fade margin by placing an RF step attenuator in series with the receiver spigot of the duplexer. We progressively attenuated the signal to drop out, as indi-

*In the event of inability to measure 20-dB audio quieting, we substitute the threshold of an op-amp squelch circuit (Motorola Micor™), for which we have an LED indicator at each radio receiver.

Table 8. Path Loss Capability

	West Texas to/from New Mexico				Field Measurements 1995-1997			
	←McElroy→		←Glass Mtn.→		←Pandale→		←Bracketville→	
EIRP	17.86	19.68	18.84	25.48	26.81	24.06	25.48	19.06
ERS	-129.74	-143.54	-131.97	-159.08	-159.90	-161.40	-152.76	-134.14
PLC		163	152	186	185	187	178	

cated by a squelch threshold LED (speaker audio was monitored, as well). Fade margin could be assessed in *simplex*, in which we obtained a signal from the distant site (signal generated at the distant site) and then extinguished by the attenuator, or *duplex*, in which the signal was generated locally and looped back through the distant site. Simplex testing was used only when we were unable to place the distant site in loop around. (We were always able to obtain a simplex test signal.) The duplex test is clearly superior, because our self-generated noise from the repeater and any associated link transmitters at both ends of the link is factored into the measurement.

• Calculations from field measurements:[5]

a) ERS. From the measurements above, we calculate:[5,19]

ERS = Basic Receiver Sensitivity + (Sens Ant -

Sens 50 Ω) + Feedline Loss - Ant gain   **(10)**

This calculation takes the intrinsic capability of a receiver and indicates the degree to which the receiver sensitivity is "spoiled" by real conditions at a specific site. The ERS is the **measured** capability of that receiver while it lies immersed in all the noise generated by other transmitters and receivers at that site, as well as any off-site locally strong emitters. The measurements in parentheses allow us to evaluate

the receiver as it looks into the noise intrinsic to a 50-ohm resistor, and then at the real world outside as presented to the receiver at its antenna lead-in. The difference subtracts from the intrinsic receiver sensitivity. Further degradation results from the required plumbing on the way to the antenna. Antenna gain is where the power to recover resides.

All measurements are entered as dBw, that is dB over 1 watt. Since the RF generator/attenuator reads out in dBm, we subtract 30 to obtain dBw. Antenna gain, usually manufacturer-specified as decibels gain over a dipole, or dBd, is entered as dBi (decibels over an isotropic radiator, which emits RF equally well in all directions). To make this correction, add 2.15 to the dBd. By making all measurements at the duplexer, we include all losses and gains up to that point from incidental filtering and external preamplifiers, if used. From that point onward, we have only feedline (including duplexer) and antenna, items which in our case are the same for each link transmitter-receiver combination. A typical site using a 15-dBd Yagi antenna might show the following:

ERS = -149 + (-76.5 - (-77.5)) + 5.25 - 17.15
    = -159.9 dBw   **(11)**

This example shows that there is 1 dB ambient noise degradation of basic receiver sensitivity of -149 dB over 1 watt; from 5.25 dB total

Table 9. Average Fade Margins: Calculated and Measured

	West Texas to/from New Mexico				Field Measurements 1995-1997			
	←McElroy→		←Glass Mtn→		←Pandale→		←Bracketville→	
Distnce	145	106	106	177	177	114	114	64
Terrain	Smooth	Smooth	Rough	Skims	Skims	Vrough	Vrough	Exroug
PLC	-1.0	163	152	186	185	187	178	
CPLoss	-128	-126	-126	-142	-142	-153	-153	-152
CFade	12.15	37	26	44	43	34	25	
MFade	26	36	10	15	21	14	9	13

Distnce = Distance between sites in km
Vrough = Very rough; Exroug = extremely rough
CPLoss = Path Loss in dB calculated from 4/3 earth surface plots;
CFade = PLC minus CPLoss in dB
MFade = Average fade loss in dB, measured about every 90 days over 2 years.

loss from the duplexer and feedline, and 17.15 dB antenna gain over isotropic.

    b) EIRP.

$$EIRP = Tx\ P_o - Feedline\ loss + Ant\ gain$$
$$TX\ P_o = Transmitter\ output\ power \quad (12)$$

Again, antenna gain is in dB over isotropic, or dBi; transmitter power is expressed in dBw. This is a simple calculation in which we convert transmitter antenna terminal power to decibels over 1 watt, subtract duplexer and feedline losses, and add the gain of the antenna over isotropic. A typical site runs a 25-watt (13.97 dBw) link transmitter. Duplexer and feedline losses are about 5.7 dB. At this site, there is a 12.5-dBd (14.65-dBi) antenna. Note that the antenna gains go **negative** on receivers and **positive** on transmitters; feedline-duplexer-connector losses do the opposite.

$$EIRP = 13.97 - 5.72 + 14.65 =$$
$$22.9\ dBw\ (52.9\ dBm) \quad (13)$$

The EIRP is the total transmitting capability of the site expressed as decibels over 1 watt. The power is increased with reference to an isotropic source by gain of the antenna in the chosen direction. Power is lost during the transition from the transmitter antenna port to the antenna through filters, duplexers and isolators, connectors and feedline.

    c. PLC is the ratio of the effective power of the link transmitter to the minimum power to required to operate the associated link receiver.[5] In dB:

$$PLC = EIRP - ERS \quad (14)$$

This very important calculation expresses the capability of our link as a number. This is the power we have available to overcome the losses over a specified path. With this relatively simple set of measurements we can size up a pair of link sites. With the exception of the contributions of site-specific noise (**Equation 10,** items in parentheses), reasonable predictions can be made before construction begins.

Continuing with the example, if the distant receiver has an ERS of -159.9 dBw, the *PLC* is:

$$PLC = 23 - (-160) = 183\ dB \quad (15)$$

Equation 15 says that, probabilistic considerations aside, we can overcome 183 dB of path loss with the transmitting and receiving systems specified. This is a key concept! If we use Bullington's[1] method for estimating path power loss, we can make a reasonable prediction of link performance.

Results: effective receiver sensitivities

We made calculations of EIRP and ERS for each visit to each site. From these data, we calculated the PLC; ERS was calculated using **Equation 9** from the three field measurements noted above. The results from the west Texas

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Table 10. Link Dropout Predictions in Hrs/Year as Function of Terrain, Climate and Distance[14]

| | West Texas to/from New Mexico | | | | Field Measurements 1995-1997 | | | |
	←McElroy→		←Glass Mtn→		←Pandale→		←Bracketville→	
d	90	66	66	110	110	71	71	40
α	4	0.25	1.0	1.0	1.0	1.0	1.0	1.0
ß	0.125	0.125	0.125	0.125	0.125	0.125	0.125	0.25
CFM		37	26	44	43	34	24	
FM	32±8	33±13	16±10	17±12	31±22	14±5	9±5	18±14
UmaxFM	<1	<<1	<1	2	<<1	<1	17	<<1
Umedium	2	<1	3	31	1	17	42	<1
UminFM	14	<1	25	465	155	42	167	7

d = distance in miles
α = Terrain factor: Smooth or water: 0.25; Rough: 1.0; Mountainous: 4.0
ß = Climate factor: Very dry or high altitude: 0.125; Normal inland temp: 0.25; Hot humid: 0.5
f = 425 MHz
CFM = Calculated fade margin
FM = Median observed fad margin, plus or minus range of measurements
U = Link down time in hours/year

system site visits from October 29, 1995, to October 11, 1997 were averaged and appear in **Table 6**. Local noise, as assessed by measurements of receiver sensitivity on the antenna, compared with sensitivity with a 50-ohm load, showed degradation of <3 dB at all sites.

From **Table 6**, it is apparent similar radios do not have similar effective sensitivities. On-site noise could be a factor and could be checked by reversing the radios paired at the site. However, this experiment would be disruptive. The variation of sensitivity is shown, too. One radio failed (and was replaced) at McElroy to New Mexico. There was one very strange result at Glass, which accounts for a large negative variation there. The most steady results were at McElroy to/from Glass, Pandale to Glass, and Brackettville to/from Pandale.

Results: effective isotropic radiated power

After each measurement was made, EIRP was calculated from **Equation 12**. The results are shown in **Table 7**.

The table shows that the apparent duplexer losses seem to vary quite a bit from the factory specification of 1.4 dB for mobile duplexers. This is likely due to reactive power and our failure to measure the power output with a 50-ohm load on the receiver port of the duplexer each time. We did measure reflected power at the duplexer antenna terminal. The differences were small—never more than a decibel—and we present the actual measured values on the transmit side, only the duplexer losses in the ERS calculation are manufacturers' specifications. (All duplexers were bought new and are factory tuned.) The lower power levels were selected for the shorter links. We wish we could have the 17-dB antennas on the PAN-BKT link, but financial considerations rule.

Results: path loss capability

With ERS and EIRP in hand, we were prepared to calculate the PLC (**Equation 14**). These results are shown in **Table 8**; all figures are in decibels over 1 watt (dB). **Table 9** shows the PLC versus calculated and measured path losses, based on path profile calculations.

In all cases, our measured fade margins (*Mfade*) exceed our 8-dB standard, based on the Rayleigh path reliability criterion. The predictions are within 10-dB of measured values in a distinct minority of paths—McElroy to Glass. Our least reliable path is from Brackettville to Pandale, and furthermore, it seems 15 dB short of the predicted value. In this respect, it appears that the handmade 4/3 plot path prediction out-

performed the computerized version. (Recall that for 50-percent path reliability the Bullington estimate is -178 dB; the computer estimate is -142 dB; and the PLC is 187 dB northwest, 177 dB southeast.) When radio waves skim the terrain, as they do between Glass and Pandale, the path is likely to be very sensitive to minor changes in *k* index.

Probability of path dropout

Path success probabilities: path dropout predictions. From **Figure 1**, we selected a fairly optimistic 90-percent path success as our goal. This translates to a minimum fade margin of 8 dB. Examination of **Table 9** shows that this is not too ambitious. We used **Equation 2** to give us another way to look at our results—this time, in terms of downtime expected on each link. **Table 10** summarizes the calculations for our west Texas measurements to date.

These data can provide the basis for further observations, in which actual downtime is observed and compared with the values calculated via the prediction formula. Examination of raw data reveals uniform fade margins only on the smooth path from Mt. McElroy to southeast New Mexico. Otherwise, the variations are remarkably wide, suggesting major changes in radio refractive index (RRI) from visit to visit. Extreme values were occasionally obtained, suggesting anomalous propagation at that time. Nearly all measurements were made near midday. These extremes are seen in **Table 11**, a summary of fade margins on the Pandale to Glass Mountains path.

These results suggest that the radio path attenuation changed. Frankly, we do not really understand the considerable variability in measured fade margin. This is a very simple measurement, requires no calculation, and is the least susceptible to observer error. Were the changes in ERS responsible? Probably not, because the path could be *worse* at a time when the ERS had *improved* (lower dBm). The SWR of the feedline and antenna did not appreciably change throughout the two years of observations and was 1.22:1. We wonder if anomalous propagation was present on October 1995, January 1996, and January 1997. (The ERS may have been spoiled by an undetected spurious emitter on site, or may have been incorrectly recorded.)

We lack sufficient data to attempt to correlate these apparent path enhancements with other paths at the same time. The distances involved prohibit simultaneous measurements, or even measurements on the same day. We will continue to observe fade margins with site maintenance visits to see if there is some cyclic change. Furthermore, we have already

reassessed the antenna connections at Glass, looking for **water in connectors**, a common cause of loss of ERS. None was found; Pandale will be next!

Day and Trovese[3] compared LOS and BLOS paths over relatively smooth Earth in the Arizona desert at frequencies of 170, 520 MHz, and higher to 24 GHz. Anomalous propagation was observed regularly at sunset and early evening on both paths.

Summary and conclusions

We present here a quantitative exercise in the theory of UHF radio propagation and path power predictions. We reviewed pertinent literature published since 1947 and presented the key information needed to deduce path losses and the effects of atmospheric refraction, optical diffraction, and shadowing. We took note of the probabilistic nature of these predictions and applied the information. We demonstrated the necessary steps to predict the performance of amateur point-to-point centimeter wavelength relay systems.

Theoretical path loss determinations were compared with field measurements of several currently operating 425-MHz radio links over a two-year period. We used a hand-drawn Earth profile map with Bullington's algorithms to predict path losses over one of four operating point-to-point paths. A computer-generated profile significantly underestimated real path losses; the **older** approach seemed closer to reality in the semiarid desert of west Texas. Use of 1.20 for the k index might improve the reliability of the computer-generated predictions. To finish our investigation, we examined the basics of anomalous propagation.

An inconstant radio refractive index could account for fading or dropouts. The paths examined so far appear to work without the help of atmospheric ducts, on scatter propagation alone. Dropout probability insurance asks for an additional 8 dB of path loss to be overcome, if we are to achieve 90-percent reliabili-

ty. In spite of significant losses from obstructions, every path has significant fade margin.

We plan to extend our measurements to our mid-zone paths, whose k should be nearer 1.34, and whose paths are not quite as rough but have more vegetation. We will then observe performance southeast to the Gulf of Mexico, wherein we may enjoy somewhat better propagation, as well as different causes for duct formation. Path power measurements provide the field experience necessary to evaluate a possible borderline linkage.

Radio amateurs contemplating linked systems need not guess at success or failure of a proposed system. If there is a single LOS link involved, success is likely—unless there are 0.6 Fresnel zone hills in the way.

Our group is constrained to erect systems on borrowed space on someone else's site. The distance to the next site may exceed LOS, cover rough terrain, and be severely limited as to available antenna height and space. In this case, we would definitely counsel some time spent poring over maps and evaluating available apparatus before committing friends to an unreliable or nonfunctioning system. Don't forget to add some dB for fade margin—remember that the path loss versus path loss capability equation provides 50-percent reliability only. Given our regulatory (and financial) constraints, good antennas and receivers will always trump higher-powered transmitters.

Early on, we found that good audio quality and system controllability are totally dependent on reliable RF links. Hence, getting the RF part of the picture right is paramount and must precede all other considerations in system planning. This paper should give help and encouragement to hams contemplating anything from a stable RF connection to a licensed family member, a voting RF-linked repeater receiver system, or a multi-hop station such as ours. Aside from our operations, there are currently over 200 channels coordinated to linking in Texas alone.* This is not an isolated pursuit!

*Data available from Texas VHF FM Society, Inc.

Maintenance of our linked systems takes much weekend energy. The measurements we have described allow us to watch our links, and sometimes prepare a problem before it knocks us down. Lightning, rain, and snow constantly challenge radiators, feed systems, and radios. We use lightning arrestors on all RF feeds, **as well as** power line protection coupled with the most compulsive grounding we can manage. We specify good quality flexible double-shielded coaxial cables or hard line as a "must-have." Proper assembly of connectors and taping of joints[20] can save hundreds of troubleshooting hours. Not only will a bad joint lose signal, it may become a radiator of spurious RF.

This work could be easily duplicated elsewhere in other existing or proposed systems. And, to those so inclined, the further step of comparison with theoretical determinations and comparison with the work of others provides challenge, opportunity, and real excitement.

Acknowledgments

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Intertie, Inc. is an affiliate of the Cactus Intertie, headquartered in California. Cactus members regularly provide substantial technical assistance to our group. ■

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PRODUCT INFORMATION

Alinco Introduces DJ-C5T VHF + UHF "Credit Card" HT

Alinco has introduced the DJ-C5T, a "credit card" Handy-Talkie (HT) Transceiver designed to operate on the popular 2-meter (144~148 MHz) and 70-centimeter (420~450 MHz) bands. The new HT, only slightly larger and thicker than a credit card, features an internal speaker, runs on an internal lithium-ion battery, has 50 memories, CTCSS encode and decode, and includes AM aircraft band receive.

Featuring a removable flexible rubber-coated antenna, the transceiver can be inconspicuous until needed. The lithium-ion battery recharges in two hours or less. Lithium-ion batteries are also free of the "memory effect" often associated with NiCd power sources.

The DJ-C5T sports its own small speaker.

The transceiver can also utilize a variety of speaker/mic combinations and its accessories are interchangeable with the DJ-C1 and DJ-C4 models.

Among the features included are CTCSS encode and decode, programmable European tone bursts, 50 memory channels plus a "call" channel. adjustable offset (to 15.995 MHz), AM Airband receive, extended VHF receive (118~173.995 MHz), full coverage of the UHF amateur band (420~449.995 MHz), adjustable tuning steps, auto power off feature, battery save setting, MARS/CAP capability, and packet radio capability. The radio comes with a clear plastic case and a "snap-in" charging unit.

For more information, contact Alinco, 438 Amapola Ave., Suite 130, Torrance, California 90501.

Reviewed by Peter J. Bertini, *K1ZJH*
Senior Technical Editor

QUARTERLY REVIEW

Radios by Hallicrafters

Many hams enjoy dabbling in vintage technology, whether it's collecting and restoring early radios or reassembling their original novice stations. These hams remember nostalgically how Bill Halligan's Hallicrafters products have played a prominent role in amateur radio stations since the company's inception in 1932.

The Hallicrafters moniker was a composite of two words: "Halligan" and "Handcrafted." Bill's earliest sets were hand-produced, one at time, using the highest quality parts. Unfortunately, Hallicrafters immediately became mired in patent infringement battles with RCA. This, combined with the depression, almost spelled an early demise for what was to become one of amateur radio's major manufacturers and supporters.

The Hallicrafters legend

Chuck Dachis' book, *Radios by Hallicrafters*, traces the Hallicrafters legend from its birth until its demise in the 1980s. The book gives a good account of the history of Hallicrafters throughout the World War II years, a 1948 African expedition (equipped with the latest Hallicrafters products), up to and until the rights to the Hallicrafters name were placed in trust by a Texas bankruptcy court. The company's last amateur radio product, the FPM-300 transceiver, was introduced in 1972.

The book reveals many behind-the-scenes activities over the history of Hallicrafters. Early photos of the various Hallicrafters plants are included, as are many rare historical photos of the folks involved behind the scenes. You'll also find a price guide, but it's dated. Prices now fluctuate wildly according to geographical area and what has unfortunately grown into a speculative "investment" market.

Product lines spotlighted

Radios by Hallicrafters is over 200 pages long. Most of the volume is devoted to covering the Hallicrafters product lines, which were not just limited to amateur radio-related products. Hallicrafters home radios, televisions, military sets, two-way radios, and other offerings share the limelight, but Hallicrafters' extensive line of receivers is given full and proper coverage. Page after page is devoted to black-and-white and color plates of Hallicrafters receivers, and the photographs are of excellent quality. Each photo is accompanied by one or two descriptive paragraphs that provide the receiver's coverage, tube complements, and other pertinent data.
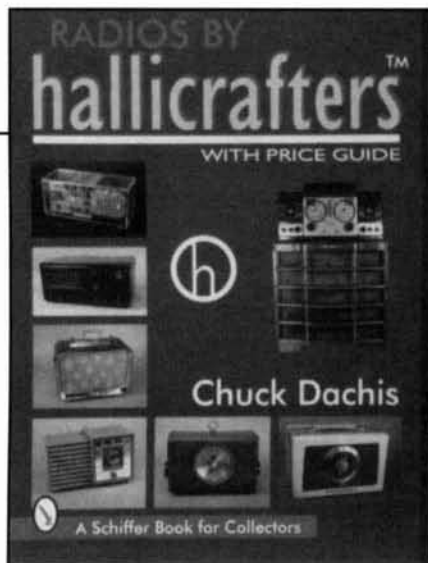
Many of the photos are from the author's extensive collection of Hallicrafters products and memorabilia. Many old-timers will probably feel a strong nostalgic twinge or two when they see the venerable SX-28 *Super Sky Rider* receiver—for some of us the epitome of a communications receiver. Others may fondly recall saving paper route money to buy their first S-38 receiver, one of Hallicrafters humbler offerings for those on a tight budget.

Radios by Hallicrafters was an ambitious undertaking and is a long overdue and fitting tribute to one of the prominent names in amateur radio history. Without any reservation, I give the book a two "thumbs up" rating. Now, if you'll pardon me, I have a sudden urge to go tinker with my SX-28 and SX-42 receivers.

Radios by Hallicrafters, by Chuck Dachis, WD5FOG, is softbound and published by Shiffer Publishing, Ltd. It sells for $29.95. ISBN number: 0-88740-929-6. ∎

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DESIGNING FREQUENCY SYNTHESIZERS

Optimum phase noise performance

Today frequency synthesizers are an integral part of the communications scene. They offer many advantages in terms of stability, flexibility, and convenience. They are ideal for operation with processors, which enables a host of facilities like dual VFOs, memories, scanning, and the like to be incorporated. Yet one of their major drawbacks is the levels of phase noise which can be generated—particularly by those synthesizers that are poorly designed or designed to a budget.

Fortunately, it is possible to predict the levels of phase noise likely to be produced by the synthesizer. With this knowledge, it is possible to optimize the design to provide the best performance, and to place additional effort into improving those circuit areas most likely to affect that performance.

While it is possible to conduct this analysis through complex math alone, most engineers prefer a more visual approach which is simple to use and very effective.

What is phase noise?

Noise is present on all signals, both in terms of amplitude and phase noise. The amplitude component is normally small and can be ignored, but this is not so for phase noise. It gives rise to noise sidebands which spread out symmetrically on either side of the carrier, as
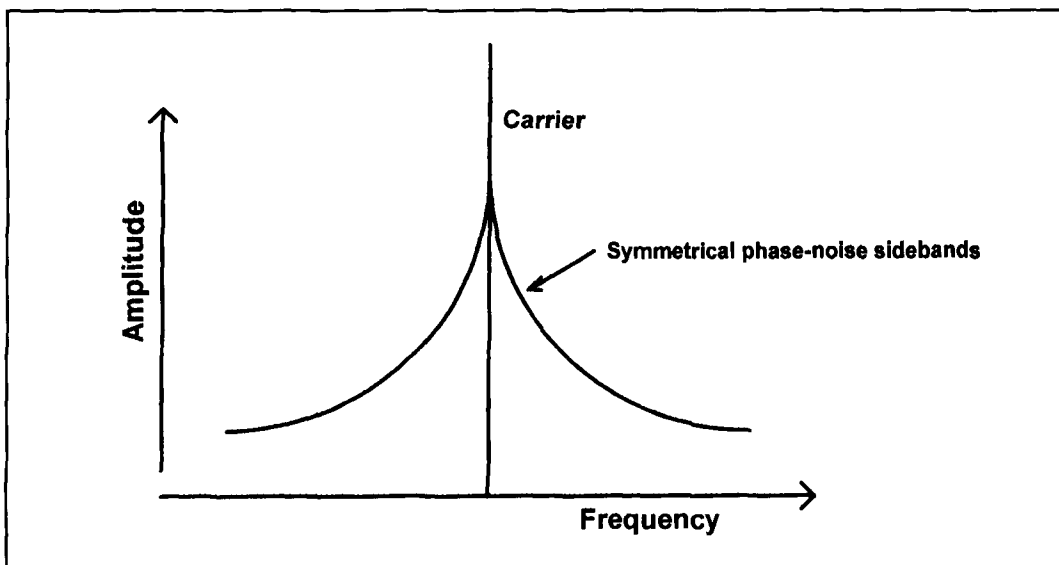
Figure 1. Noise spreading out on either side of a carrier as a result of phase noise.
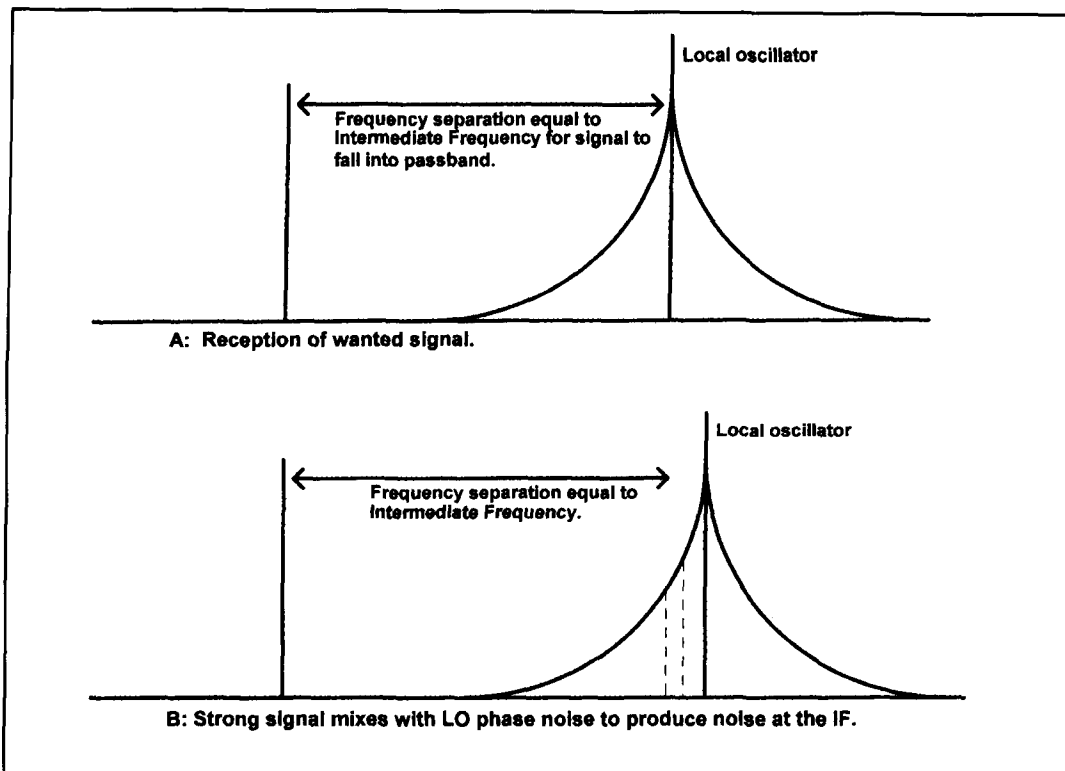
Figure 2. Reciprocal mixing. (A) Reception of wanted signal. (B) Strong signal mixes with local oscillator phase noise to produce noise at the IF.

illustrated in **Figure 1**. This diagram shows the performance of a typical oscillator. However a synthesizer, being rather more complicated, has its noise profile modified by the action of the loop, as will be seen later.

Different oscillators have different amounts of phase noise. One of the major factors is the Q of the tuned circuit. Accordingly, crystal oscillators perform very well, and variable frequency oscillators using traditional variable capacitors and inductors do not perform as well. Voltage-controlled oscillators using varactors are often comparatively poor because of the inherent low Q of the varactors.

Phase noise arises from the short-term instabilities on a signal. These instabilities are sometimes referred to as phase jitter. This results in the signal being phase modulated, which in turn gives rise to the noise sidebands spreading out on either side of the wanted signal.

It is often necessary to quantify the level of phase noise. Unlike a perfect carrier that occupies a single frequency, noise spreads out over the frequency spectrum. This means that noise power is proportional to the bandwidth. Consequently, a certain bandwidth must be specified to measure noise. In addition, its measurement position must also be specified if the noise varies with frequency. In the case of phase noise, it is generally specified in a 1-Hz bandwidth and at a certain offset from the carri-

er. For example, a level of noise may be quoted as being 50 dB down on the carrier in a 1-Hz bandwidth at 1 kHz offset; i.e., -50dBc/Hz at 1 kHz offset.

Alternatively, the phase noise can be considered in terms of integrated phase jitter and measured in RMS radians. Although this approach is not as widely used, it is sometimes more convenient when phasor calculations are to be performed. It is often used when bit error rates must be estimated when phase modulation is used.

The effect of phase noise

Phase noise affects communications systems in a number of ways. In a transmitter, it will result in the signal spreading out over a wide bandwidth corresponding to the phase noise characteristic of the synthesizer. For an SSB transmitter this can be heard as "splatter" on either side of the signal, which can cause interference to other users of the spectrum.

For a receiver, phase noise gives rise to a phenomenon known as reciprocal mixing. Under normal conditions, the incoming wanted signal mixes with the local oscillator to change the frequency down to the intermediate frequency. Here, the difference between the local oscillator frequency and the incoming signal frequency is equal to the intermediate frequen-

cy, as shown in **Figure 2**. If the local oscillator is tuned off frequency, then the incoming signal may mix with the phase noise on the local oscillator to produce a signal within the IF bandwidth as shown. Obviously, the lower the level of phase noise, the less this effect will be. However, in the presence of nearby strong signals, this reciprocal mixing effect can considerably raise the noise floor of the receiver and impair its sensitivity. It is for this reason that a receiver local oscillator must have a very good phase noise performance.

Phase noise is also important in a number of other areas. It is obviously important for systems using phase modulation. Here, large amounts of phase noise can degrade the bit error rate and reduce system reliability and performance. Radar systems that use the Doppler principle to pick out only the moving targets, find that phase noise appears as a signal on a different frequency to the main carrier. As a result, it introduces clutter onto the screen.

These, and many other applications, require signals with low levels of phase noise and, as a result, phase noise is a fundamental parameter in the design of any frequency synthesizer. Accordingly, it must be considered at the earliest stages of the design along with other basic requirements like frequency coverage, step size, and so forth.

The basic synthesizer loop

Apart from direct digital synthesizers, which are becoming more popular, most synthesizers are based upon the phase-locked loop (PLL). As shown in **Figure 3**, the PLL consists of three main blocks: a voltage controlled oscillator (VCO), phase detector, and a loop filter. A reference signal is also required, but this is not strictly part of the loop.

In operation, signals from the reference and VCO enter the phase detector. Here, an error voltage proportional to the phase difference is produced. This signal is passed through the loop filter and then applied to the control termi-

nal of the VCO. The error voltage is such that the phase error between the reference and the VCO is reduced. During the loop locking process, a point is eventually reached where a steady error voltage is present. This means a constant phase difference exists between the two signals. As the phase difference is unchanging, it means that the two signals are on exactly the same frequency.

To make the loop run on a different frequency to the reference, a divider can be placed between the VCO and the phase detector. Because the phase detector will try to ensure that the phase difference between the two signals entering it is made constant, the VCO will run at a multiple of the reference frequency as governed by the divider (see **Figure 4**).

As a basic PLL synthesizer of this type will only step at increments of the reference frequency, a divider is often placed after the reference oscillator to reduce its frequency to a suitable one. In this way, one may use a crystal oscillator running at a frequency of a Megahertz or so to ensure optimum stability.

Apart from placing a divider into the loop, it is also possible to place a mixer into the loop (see **Figure 5**). This effectively adds a frequency offset onto the generated frequency and can be used to generate multi-loop synthesizers, resulting in improved performance over a single loop.

The loop and phase noise

Phase noise is generated at different points around the synthesizer loop. Depending on where it is generated, it affects the output in different ways. For example, noise generated by the VCO has a different effect from that generated by the phase detector. This illustrates that it is necessary to look at the noise performance of each circuit block in the loop when designing the synthesizer so that the best noise performance is obtained.

Apart from ensuring that the noise from each part of the circuit is reduced to an absolute min-
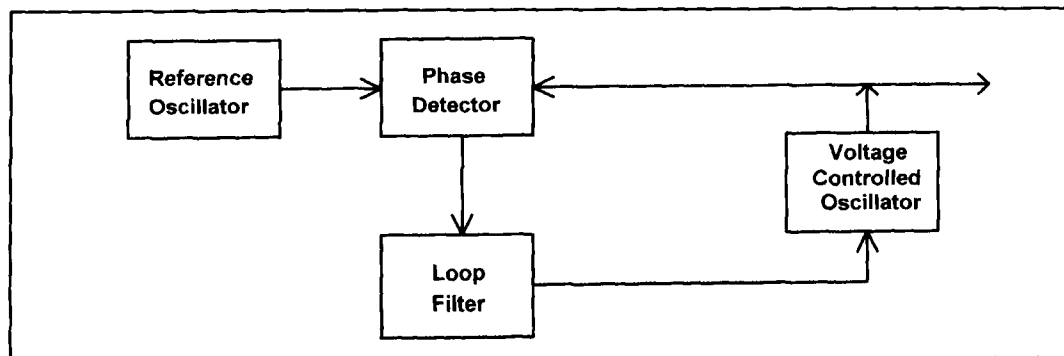
Figure 3. A phase-locked loop.
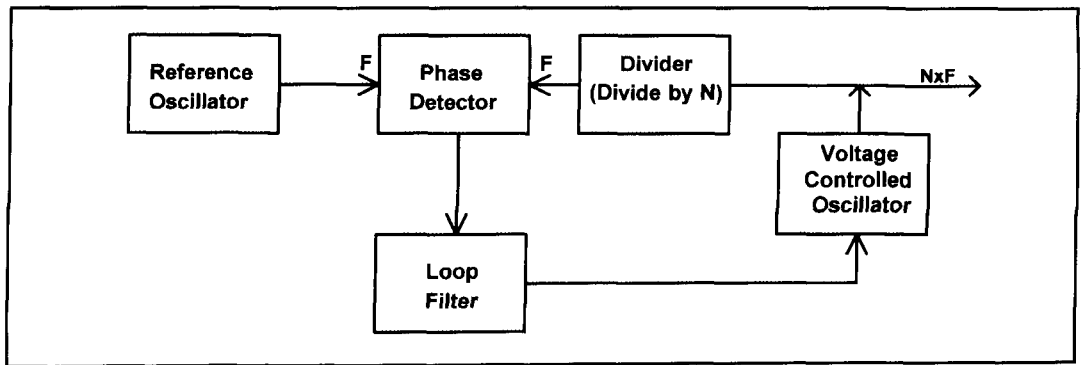
Figure 4. A synthesizer using a divider to multiply the reference frequency.

imum, it is the loop filter which has the greatest effect on the final performance of the circuit. The loop filter determines the break frequencies where noise from different parts of the circuit start to affect the output.

To see how this happens, take the example of noise from the VCO. Noise from the oscillator is divided by the divider chain and appears at the phase detector. Here it appears as small perturbations in the phase of the signal and emerges at the output of the phase detector. When it comes to the loop filter, only those frequencies below its cut-off point appear at the control terminal of the VCO to correct or eliminate the noise. From this it can be seen that VCO noise which is within the loop bandwidth is attenuated, but that which is outside the loop bandwidth is left unchanged.

The situation is slightly different for noise generated by the reference. This enters the phase detector and again passes through it to the loop filter where the components below the cut-off frequency are allowed through and appear on the control terminal of the VCO. Here, they add noise to the output signal. So noise from the reference is added to the output

signal within the loop bandwidth, but it is attenuated outside the loop.

Similar arguments can be applied to all the other circuit blocks within the loop. In practice, the only other block which normally has any major effect is the phase detector, and its noise affects the loop in exactly the same way as noise from the reference. Also, if multi-loop synthesizers are used, the same arguments again apply.

Effects of multiplication

As noise is generated at different points around the loop, it is necessary to discover what effect this has on the output. As a result, it is important to relate all the effects back to the VCO. Apart from the different elements in the loop affecting the noise at the output in different ways, the multiplication in the loop also has an effect.

The effect of multiplication is very important. The level of phase noise from some areas is increased in line with the multiplication factor (i.e., the ratio of the final output frequency
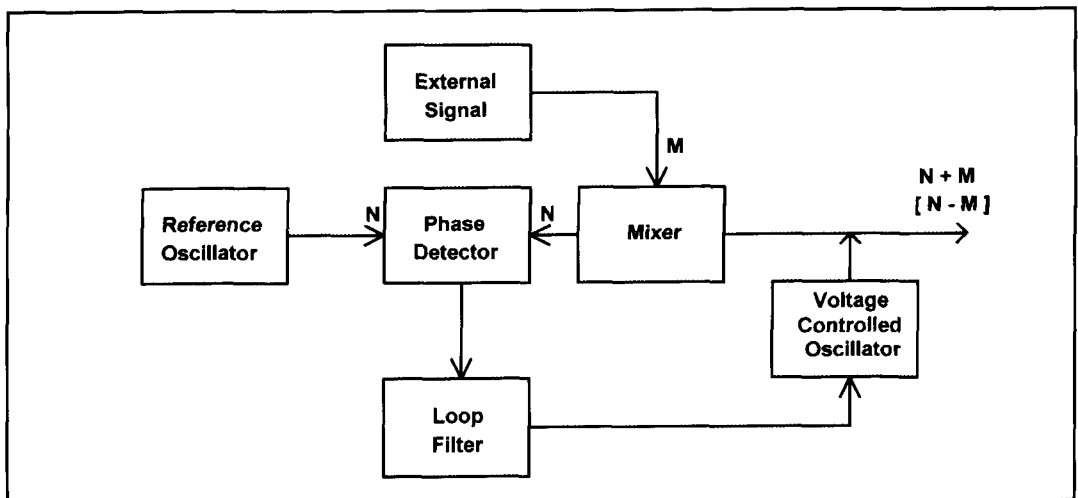
Figure 5. A synthesizer loop using a mixer to provide a frequency offset.
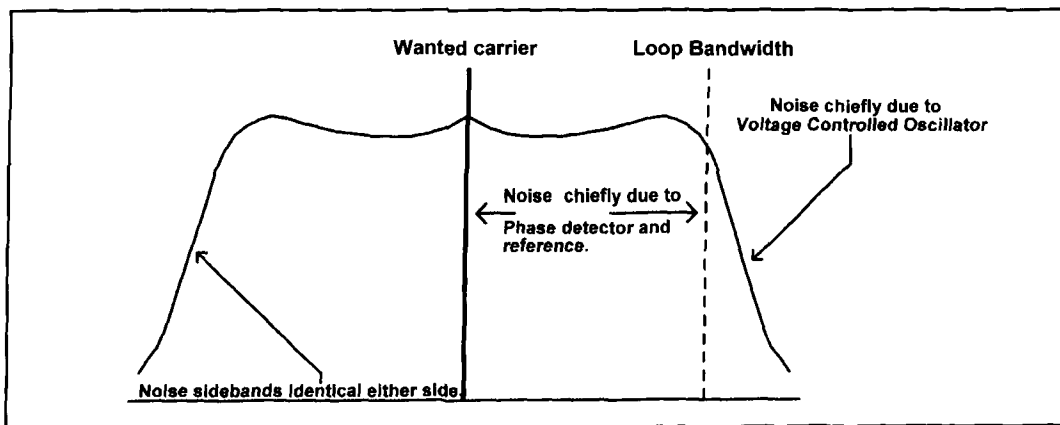
Figure 6. Outline of the noise characteristic of a typical one loop synthesizer.

to the phase comparison frequency). In fact, it is increased by a factor of $20 \log_{10} N$, where N is the multiplication factor. The VCO is unaffected by this, but any noise from the reference and phase detector undergoes this amount of degradation. Even very good reference signals can be a major source of noise if the multiplication factor is high. For example, a loop which has a divider set to 200 will multiply the noise of the reference and phase detector by 46 dB!

Using this information, it is possible to build up a picture of the performance of the synthesizer. Generally, this will look like the outline in **Figure 6**. From this it can be seen that the noise inside the loop bandwidth is due mainly to components like the phase detector and reference, while the VCO generates the noise outside the loop. A slight hump is generally evident at the point where the loop filter cuts off and the loop gain falls to unity.

By predicting the performance of the loop, it is possible to optimize the performance or to look at areas which can be addressed to improve the performance of the whole synthesizer before the loop is even built. T analyze the loop further, it is necessary to look at each circuit block in turn.

Voltage-controlled oscillator

The noise performance of the oscillator is of particular importance because the noise performance of the synthesizer outside the loop is totally governed by its performance. Also, its performance may influence decisions about other areas of the circuit.

The typical noise outline for a VCO is shown in **Figure 7**. At large frequency offsets from the carrier the noise is flat and determined largely by factors such as the noise figure of the active device. The performance of this area of oscillator operation can be optimized by ensuring the circuit is running under optimum noise perfor-

mance conditions. Another approach is to increase the power level of the circuit so the signal-to-noise ratio improves.

Close in, the noise starts to rise—initially at a rate of 20 dB per decade. The point at which this rise begins is determined mainly by the Q of the oscillator circuit. A high Q circuit will ensure a good noise performance. Unfortunately, VCOs have an inherently low Q because of the Q of the tuning varactors normally used. Performance can be improved by increasing the Q, but this often results in the reduction of oscillator coverage.

Still further in towards the carrier, the noise level starts to rise even faster—at a rate of 30 dB per decade. This results from flicker or $1/f$ noise and can be improved by increasing the level of low frequency feedback in the oscillator circuit. In a standard bipolar circuit, a small unbypassed resistor in the emitter circuit can provide significant improvements.

To assess the performance of the whole loop, it is necessary to assess the performance of the oscillator once it has been designed and optimized. While there are a number of ways to achieve this, the most successful is generally to place the oscillator into a test loop having a narrow bandwidth and measure its performance with a spectrum analyzer by holding the oscillator steady. The results are only valid outside the loop bandwidth, but, as a test loop is likely to have a much narrower bandwidth than the loop being designed, the noise levels in the area of interest will be unaltered.

Reference

The noise performance of the reference follows the same outlines as those for the VCO, but the performance is naturally far better. The reason for this is that the Q of the crystal is many orders of magnitude higher than that of the tuned circuit in the VCO.
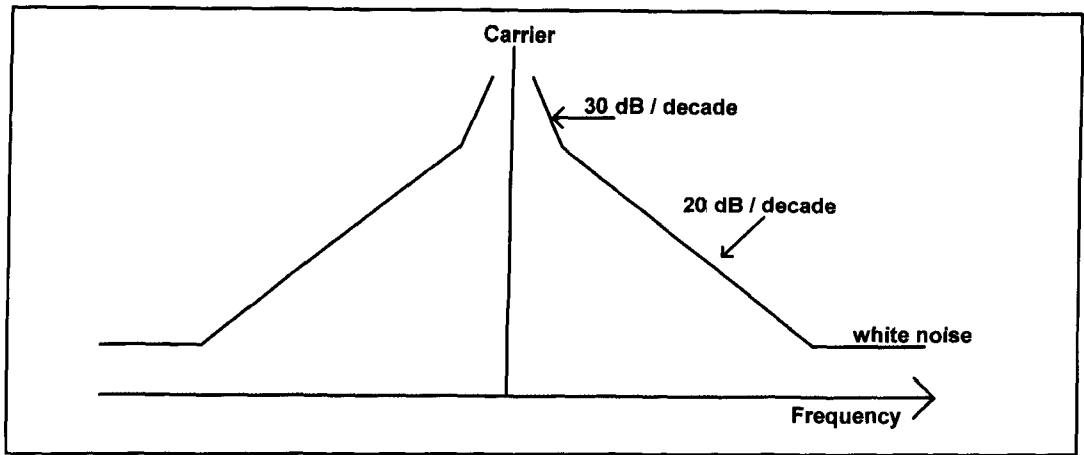
Figure 7. Noise outline for an oscillator.

Typically, it is possible to achieve figures of -110 dBc/Hz at 10 Hz from the carrier and 140 dBc/Hz at 1 kHz from a crystal oven. Figures of this order are quite satisfactory for most applications. If lower levels of reference noise are required, they can be obtained—but at a cost. In instances where large multiplication factors are necessary, a special low noise reference may be the only option. However, considering the cost, they should be avoided wherever possible. Plots of typical levels of phase noise are often available with crystal ovens giving an accurate guide to the level of phase noise generated by the reference.

Frequency divider

Divider noise appears within the loop bandwidth. Fortunately, the levels of noise generated within the divider are normally quite low. If an analysis is required, it will be found that noise is generated at different points within the divider. Each of these points will be subject to a different multiplication factor depending upon where in the divider the noise is generated and the division ratio employed from that point.

Most divider chains use several dividers, and, if an approximate analysis is to be performed, it may be more convenient to only consider the last device or devices in the chain as these will contribute most to the noise. However, the noise is generally difficult to measure and will be seen with that generated by the phase detector.

Phase detector

Like the reference signal, the phase detector performance is crucial in determining the noise performance within the loop bandwidth. There are a number of different types of phase detector (see **Figure 8**). The two main categories are called analog and digital.

Mixers are used in analog phase detectors. If the output signal-to-noise ratio is to be as good as possible, it is necessary to ensure the input signal levels are as high as possible within the operating limits of the mixer. Typically, the signal input may be limited to around -10 dBm and the local oscillator input to +10 dBm. In some instances, higher level mixers may be used with local oscillator levels of +17 dBm or higher. The mixer should also have a low NTR (noise temperature ratio). As the output is DC-coupled, it is necessary to have a low output load resistance to prevent a backward bias from developing. This could offset the operation of the mixer and reduce its noise performance.

It is possible to calculate the theoretical noise performance of the mixer under optimum conditions. An analog mixer is likely to provide a noise level of around -153 dBc/Hz.

There is a variety of digital phase detectors that can be used. In theory, these give a better noise performance than their analog counterparts. At best, a simple OR gate-type will give figures about 10 dB better than an analog detector will give, and an edge triggered-type (e.g., a dual D type or similar) will give a performance of around 5 dB better than the analog detector.

These figures are the theoretical optimum and should be treated as a guide, although they are sufficient for initial noise estimates. In practice, other factors may mean that the figures are different. For instance, power supply noise, circuit layout, etc. can degrade the performance from the ideal. If very accurate measurements are required, then results from the previous use of the circuit, or from a special test loop, can provide the required information.

Loop filter

There are a variety of parameters within the area of the loop filter that affect the noise performance of the loop. The break points of the

A: Analog Mixer (Phase Detector)

B: Exclusive OR as a phase detector.

C: Edge-triggered dual D type comparator.
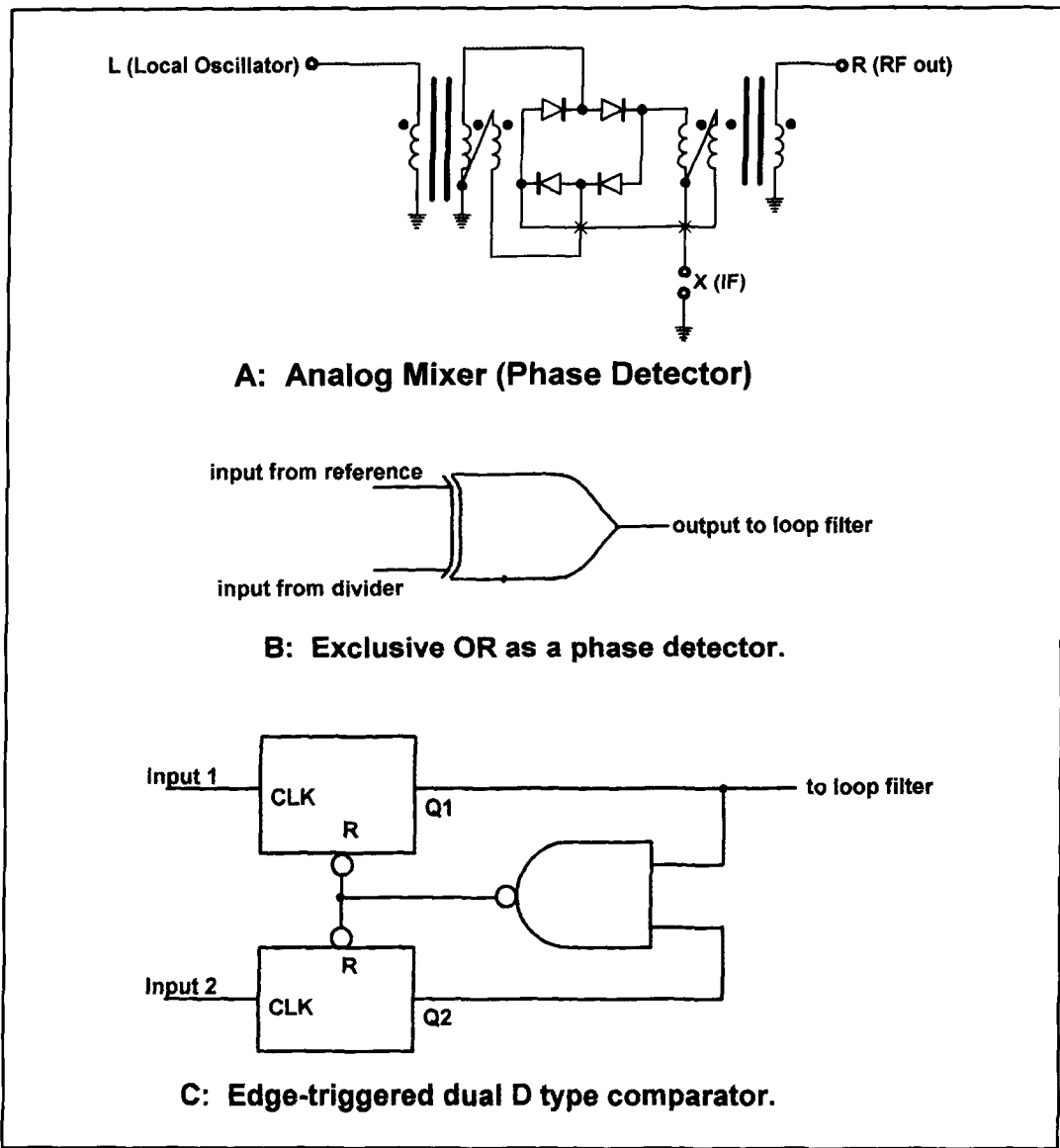
Figure 8. Phase detectors.

filter and the unity gain point of the loop determined by the filter govern the noise profile.

In terms of contributions to the noise in the loop, the major source is likely to occur if an operational amplifier is used. If this is the case, a low noise variety must be used—otherwise the filter will make a large contribution to the loop phase-noise profile. This noise is often viewed as combined with that from the phase detector, appearing to degrade its performance from the ideal.

Plotting performance

From the investigation of the noise components from each element in the loop, it is possible to construct a picture of how the whole loop will perform. While this can accomplished mathematically, a simple graphical approach quickly reveals an estimate of the performance and shows which are the major elements contributing to the noise. In this way, some redesign can be undertaken before actual construction, so the best option can be chosen on the drawing board. Naturally, it is likely that some optimization will be required after building, but this method enables the design to be made as close as possible beforehand.

First, it is necessary to obtain the loop response. This is dependent upon many factors, including the gain around the loop and the loop filter response. For stability, the loop gain must fall at a rate of 20 dB per decade (6 dB per octave) at the unity gain point as illustrated in **Figure 9**. Provided this criterion is met, a wide
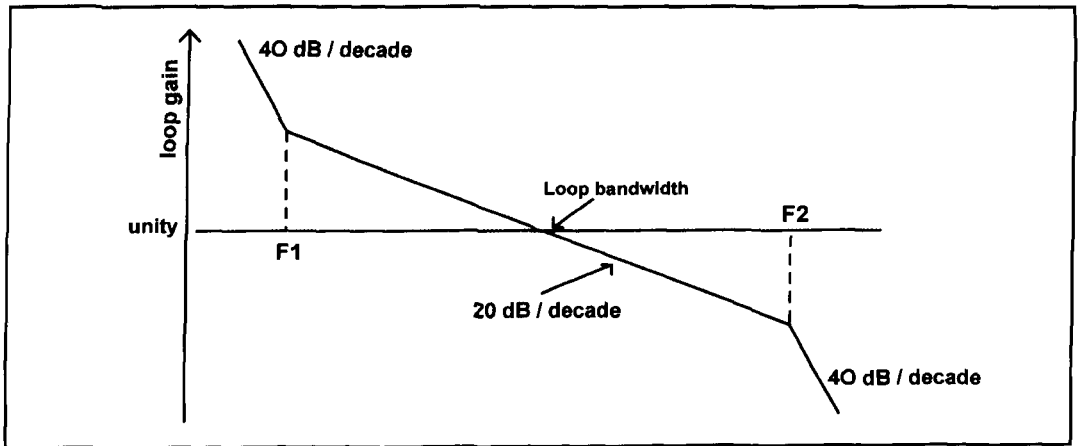
Figure 9. Loop response showing loop filter break points.

variety of filters can be used. Often it is useful to have the loop response rise at a greater rate than this inside the loop bandwidth, allowing the VCO noise can be attenuated further. Outside the loop bandwidth, a greater fall-off rate may suppress the unwanted reference sidebands further. From a knowledge of the loop filter chosen, the break points can be calculated; and with a knowledge of the loop gain, the total loop response can be plotted.

With the response known, the components from the individual blocks in the loop can be added, as they will be affected by the loop and seen at the output.

First let's look at the VCO. Outside the loop bandwidth its noise characteristic is unmodified. However, once inside this point, the action of the loop attenuates the noise, first at a rate of 20 dB per decade, and then at a rate of 40 dB per decade. The overall affect of this is to modify the response of the characteristic

(see **Figure 10**). Outside the loop bandwidth, the noise profile is left unmodified. Far out, the noise is flat, but further in the VCO noise rises at the rate of 20 dB per decade. Inside the loop bandwidth, the VCO noise will be attenuated—first at the rate of 20 dB per decade, which, in this case, gives a flat noise profile. Then, as the loop gain increases at the filter break point to 40 dB per decade, there is a drop in the VCO noise profile of -20 dB per decade. However further in, the profile of the stand-alone VCO, noise rises to -30dB per decade. The action of the loop leads to an overall drop of -10 dB per decade.

The effects of the other significant contributions can be calculated. The reference response can easily be deduced from the manufacturers' figures. Once obtained, these must have the effect of the loop multiplication factor added. Once this has been calculated, the effect of the loop can be added. Inside the loop, there is no
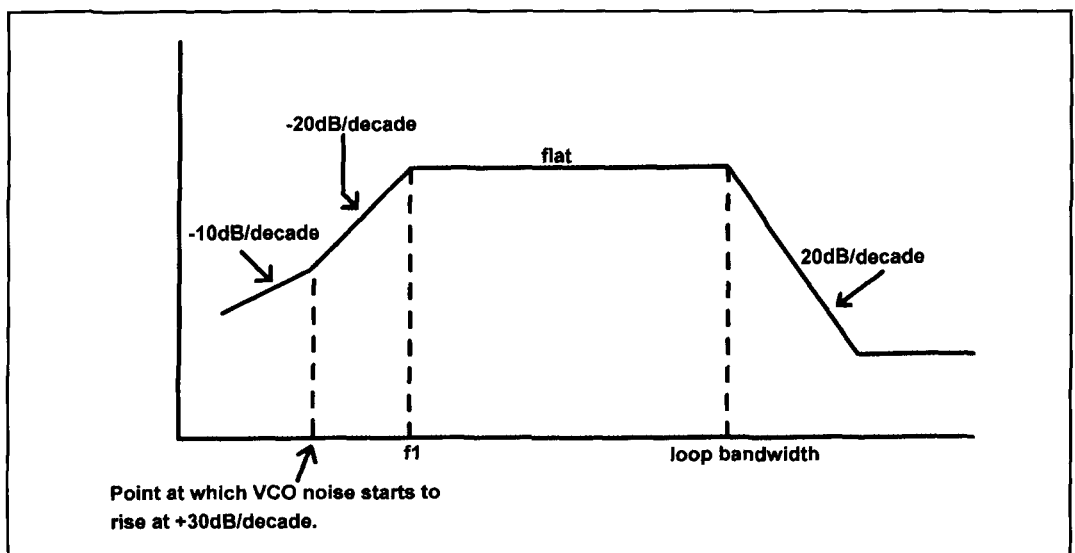
Figure 10. Overall effect of the VCO noise contribution when modified by the action of the synthesizer loop.
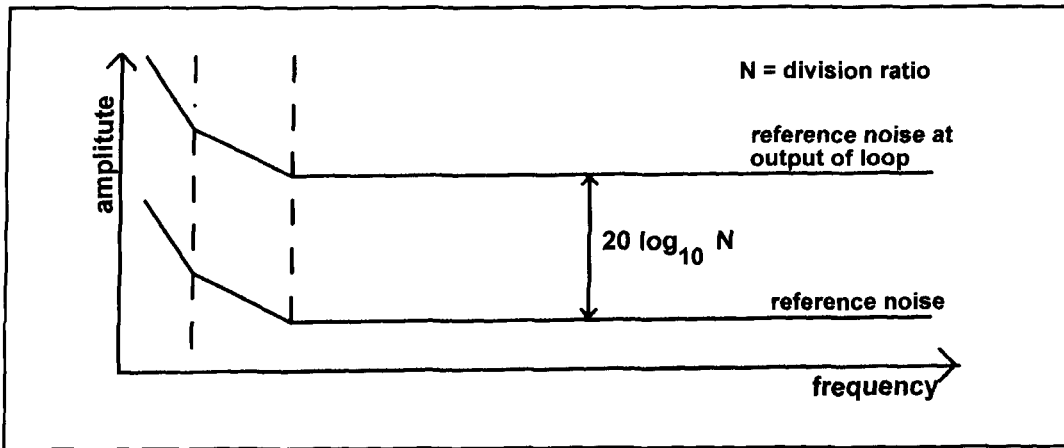
Figure 11. Noise from the reference source as seen at the output of the loop.
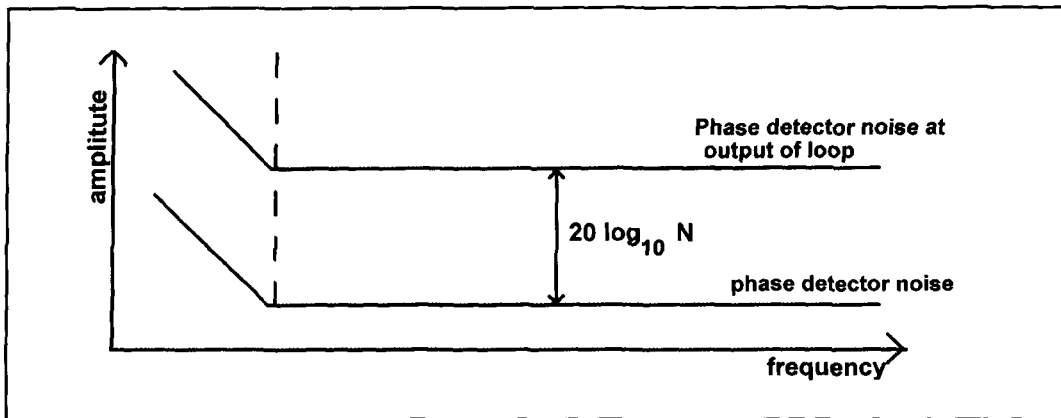
Figure 12. Noise from the phase detector as seen at the output of the loop.

effect on the noise characteristic; however, outside this frequency it will attenuate the reference noise, first at a rate of 20 dB per decade and then after the filter break point at 40 dB per decade as shown in **Figure 11**.

The other major contributor to the loop noise is the phase detector. The effect of this noise is treated in the same way as the reference, having the effect of the loop multiplication added and then being attenuated outside the loop bandwidth as shown in **Figure 12**.

Once all the individual curves have been generated, they can be combined onto a single plot for a full picture of synthesizer performance. When doing this, remember that it is necessary to produce the RMS sum of the components because the noise sources are not correlated.

Once this has been done, it is possible to optimize the performance by changing factors like the loop bandwidth, multiplication factor, and possibly the loop topology to obtain the best performance and ensure that the required specifications are met. In most cases, the loop bandwidth is chosen so a relatively smooth transition is made between the noise contribu-

tions inside and outside the loop. This normally corresponds to lowest overall noise situation.

Summary

Although this approach may appear to be slightly "low tech" in today's highly computerized engineering environment, it has the advantage that a visual plot of the predicted performance can be put together very easily. In this way, the problem areas can be quickly identified, and the noise performance of the whole synthesizer optimized before one commits to a final design. ■

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Edited by Peter J. Bertini, K1ZJH
Senior Technical Editor

This month we have two projects from Quarterly staffers. First, Peter Bertini, K1ZJH, our senior technical editor and editor of this column, presents his answer to expensive hot-air soldering stations for handling surface mount parts. Second, Rick Littlefield, K1BQT, our resident QRP aficionado, presents his latest little transceiver: the "Teeny Twoer."

—*de KA1STC*

A Hot-Air SMD Soldering Station for the Home Workshop

Here's a homebrew alternative to pricey commercial units.

Peter J. Bertini, K1ZJH
Senior Technical Editor

As a technical editor, I occasionally receive product samples from chip manufacturers. Some of these, like the Analog Devices AD607 "receiver on a chip" would be fun to experiment with. But, alas, the latest generation of integrated circuits sport surface-mount designs, and are no longer offered in the familiar DIP-style packages.

Conventional soldering methods don't work with these tiny multiple-lead surface-mount devices. What's needed is a "hot-air soldering workstation" that plays a stream of superheated air over the work, heating all the device leads and the pc board to a temperature where *reflow* soldering takes place. A small handheld hot air system easily runs $500, bench units start at $1,000 (you supply compressed air), and prices continue into the five-digit range. Most hams aren't willing, or able, to spend anywhere near

this amount for a soldering tool. So, I decided to apply some ham ingenuity to the problem, and designed a low-cost unit using commonly available items.

This project is the result of several weeks of experimentation. I sacrificed quite an assortment of soldering irons and desoldering tools before realizing success. While the result isn't the ultimate solution, my invention readily handles small SMD resistor, capacitor, and IC packages. The information presented here is more of a guideline to building a similar device, and not so much a detailed blueprint for producing a carbon copy of my unit. There's room for improvement, and I hope others will offer suggestions for future improvements.

My unit was constructed from discarded Weller soldering stations. Of course, I don't recommend hacking up functional units for this project! I had accumulated several year's worth of worn out and discarded Weller solder station component-power packs, soldering iron handles, tips, etc. at very attractive prices while cruising the local hamfests. There are two common designs. The original Weller iron contains a magnetic temperature sensor housed in the soldering iron heating chamber. Newer versions use electronic temperature control. A temperature sensor is housed in the heating chamber near the tip, and the electronics are either in the base or soldering iron handle. Either iron style is suitable for this project.

Modifying the EC1201A Soldering Station

Figure 1 shows the construction of a Weller EC1201A soldering station. Salvage the barrel assembly (1), the heater assembly (3), the heater mount and socket (4), and the iron handle and cord assembly (5). Discard the elec-

1: Barrel nut assembly
2: Iron soldering tip
3: Heater element
4: Heater assembly socket
5: Handle
6: Power Cord
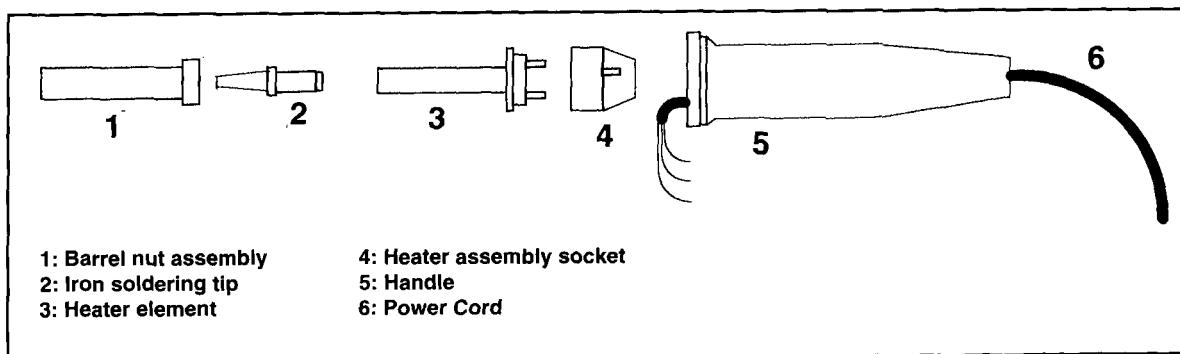
Figure 1. The major components of a Weller controlled temperature soldering iron. The temperature-sensing element and switch are discarded (not shown), and the tip (2) is replaced with a modified Weller type "ETA" tip.
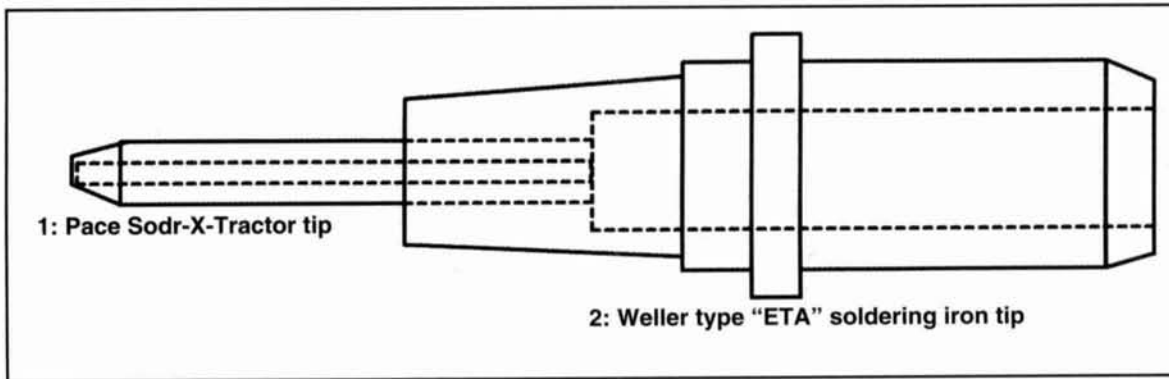
Figure 2. The Weller ETA style tip is designed for irons that have electronic sensing. The tip is partially hollow and is easily modified for airflow by filing the tip end flat and drilling an air hole through the tip. To permit working in close tolerances, the tip is extended using a Pace "Solder-X-Tractor" tip as a heat pipe to direct airflow. The tip has an inside 0.095-inch orifice; the outside diameter is 0.125 inch.

tronic sensor and any other electronics found in the newer models, or the mechanical magnetic temperature sensor and control contacts from the earlier models. Wire the heater directly to the supply leads on the power cord, placing the full supply voltage from the Weller powerpack continuously and directly on the 24-volt resistance heater.

I'll admit operation probably exceeds the design specifications of the Weller heater. It glows faintly in a darkened room, indicating the high temperatures reached by the heater element. I estimated that the airflow needed to reach about 400 to 500 degrees C was adequate. I suspect the operating life of the heater assembly is greatly shortened by this abuse, but its relatively low-cost component and my infrequent SMD soldering needs have yet to result in a failure. For safety reasons, I *never* leave the iron unattended or idling without compressor airflow.

Use silicon adhesive to seal the cord entry point into the handle, and close the gap between the heater socket assembly and handle to make everything airtight. Note that an open air chamber now exists from the handle through the heater assembly to the soldering iron tip. Next, modify the Weller tip to allow for airflow. The earlier magnetic-controlled tips aren't well suited for this modification, so I used a Weller "ETA" tip. This tip has a hollow core to accommodate the electronic sensor used in later irons.

File the tip flat, and drill it to permit a Pace SX-300 desoldering tip (0.095 inch inside, 0.125 inch outside, part number 1121-0091)* to be force-fitted as a tip extension (see **Figure 2**). If the Pace tips aren't readily available, simply drill the Weller ETA tip with a #46 wire-gauge drill to form a suitable hot-air soldering tip.

*Pace, 9329 Fraser Street, Silver Spring, Maryland 20910. Solder-X-Tractor tip, part number: 1121-0091.

The Pace tip extension is nice as it provides a very narrow air pipe, making it easy to maneuver in tight areas. At this point, pressurized air applied to the Weller handle will flow through the handle, solder heating element, and the modified tip.

Recently, I added an additional modification: a small strip of metal, with a width that matches the Weller heater element inside diameter. The strip is twisted about 2-1/2 turns, and, when inserted into the heater element, it diverts the airflow to produce higher air temperatures.

A Pressurized Air Supply

Next, I had to find a suitable means of supplying pressurized air. This was the most costly portion of the project. Visiting a pet store, I noted that several large aquariums were being fed air from a single compressor—a rather quiet device in a small package. This compressor was moving a rather large quantity of air, while under the backpressure of many feet of tubing. I settled on a model in the $40 price
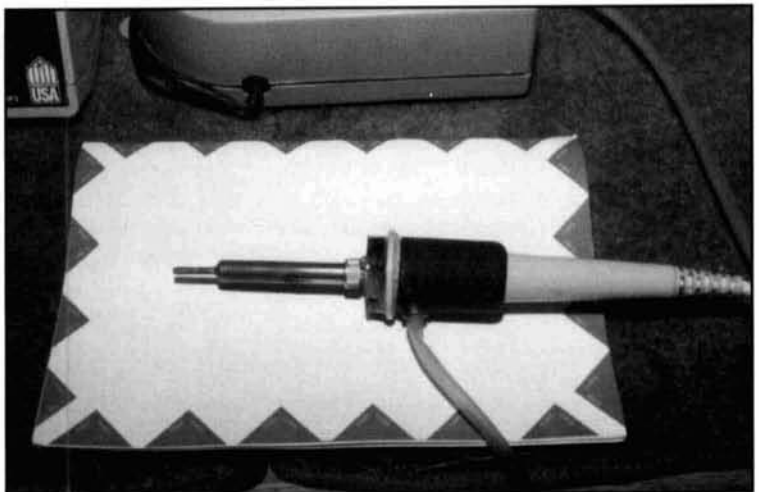
Photo A. Modified Weller soldering iron. Note attachment of air hose.

Photo B. The Weller soldering-station power supply, modified iron, and aquarium air pump used in the author's home SMD soldering setup.

range and also grabbed an assortment of aquarium air hose, hose splices, and brass control valves while the charge card was out. At home, exploratory surgery showed how the fellow worked.

The magnetic field from an electromagnet vibrates an iron armature in sync with the 60-Hz AC line voltage. The armature travel is sufficient to drive two thick, and opposing, rubber diaphragms. The diaphragms, with some simple internal reed valves, form simple air pumps, each alternately driving air to the supply nipple on each AC voltage crest. The pump also features a variable air-pressure control. The pump air pressure can be controlled externally by providing a brass aquarium valve as a bleed-off for excess air. However, according to the manufacturer's directions, placing the valve inline to restrict the airflow will damage the diaphragms. These little pumps are surprisingly powerful!

The Weller handle must be modified to accept air from the compressor air hose (**Photo A**). This is easily done by drilling a small hole into the side of the Weller handle and gluing a plastic air-hose splice over the opening. Choose an area where the handle taper is widest, so the power cord doesn't interfere with the airflow. I also drilled out the splice to a larger diameter because the original opening overly restricted the airflow and created excessive backpressure on the pump.

As shown in **Figure 2**, the air is carried into the soldering line by a length of aquarium air hose. I chose a more expensive grade that's more pliable and lightweight than the standard grades. To keep things neat, tie-wrap the air

hose to the iron's power cable over a major portion of its run.

Check It Out

Of course, the proof of any project is in how well it works. Like most SMD workstations, some experimentation is needed to find the techniques that will allow you to get the most out of the hot-air station. In general, a good warm-up period is needed to allow the heater and iron to reach operating temperature. The airflow rate isn't overly critical, but too little or too much air will hinder soldering efficiency.

When using a hot-air system, it can take a while before the SMD device and pc board are heated sufficiently for reflow soldering to occur. I spent several hours removing and resoldering SMDs from a discarded Motorola handheld radio circuit board. These Motorola boards are usually a bit difficult to work with; however, even the smallest parts were easy to remove or reflow solder to the board. Larger ball-grid SMD IC packages were way beyond the capabilities of this, or any, simple soldering device. But, with practice, I was soon soldering and removing the AD607 chip from the development board.

I sent the unit (**Photo B**) to Rick Littlefield, K1BQT, for evaluation. He confirmed my observations. Considering the modest cost, I'm very pleased with this hot-air system. Its scope is limited to small devices, but similar limitations apply to many commercial stations I've tried in the past.

The "Teeny Twoer" AM Transceiver

More AM nostalgia in a small package.

Rick Littlefie!d, K1BQT

In keeping with the current wave of AM nostalgia, here's a 2-meter companion for the Nor'easter-Six described in the Winter 1998 issue of *Communications Quarterly*. Smaller and simpler, the "Teeny Twoer" is a complete 1-watt QRP AM rig with a tunable-superhet receiver and a crystal-controlled AM transmitter. All parts are low-cost experimenter items and the entire transceiver is constructed on a single 3.8 x 4.0-inch two-sided board.

Circuitry

As shown in **Figure 1**, the receiver is a simplified version of a popular single-conversion airband circuit originally described in *Radio*
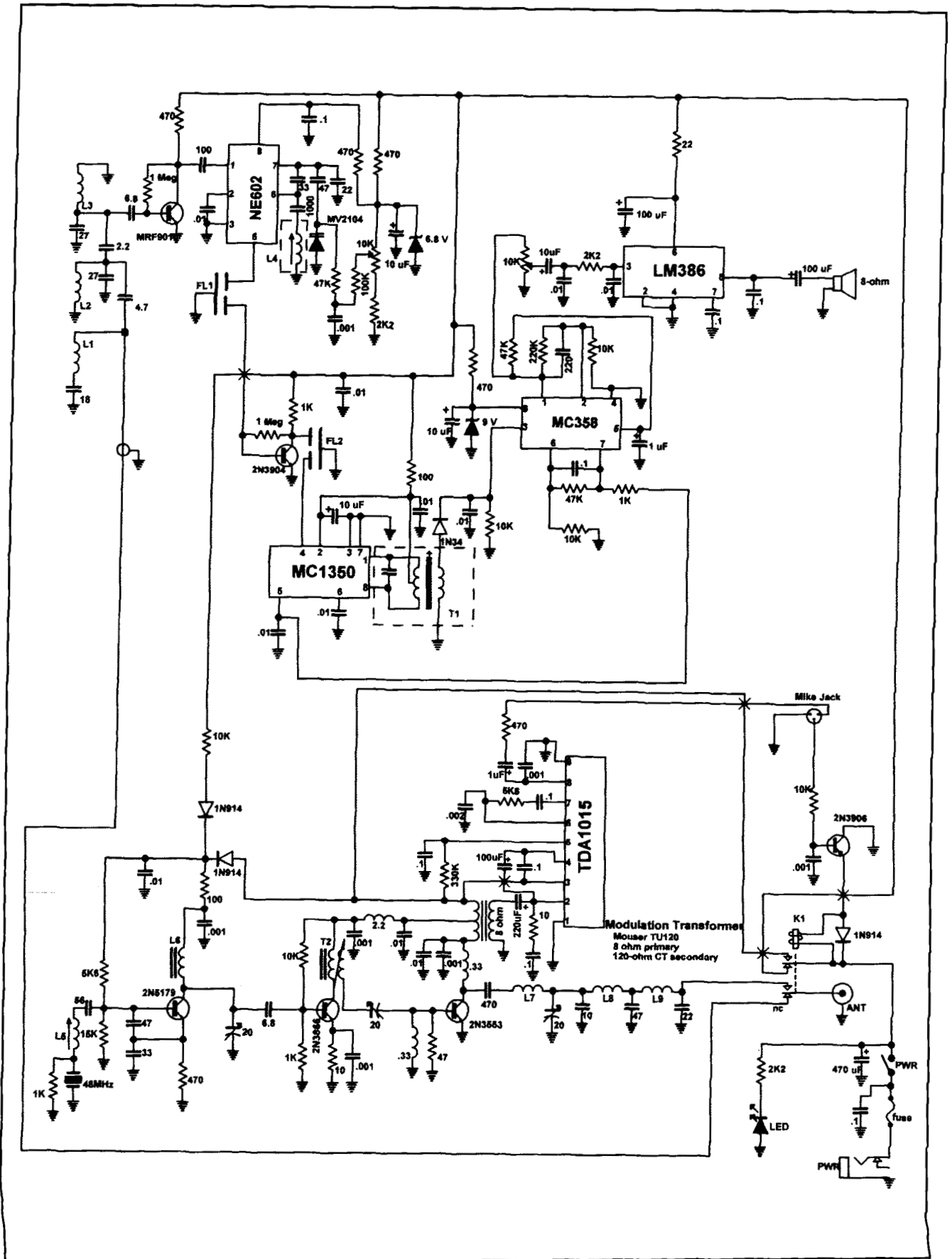
Figure 1. The Teeny Twoer schematic diagram.
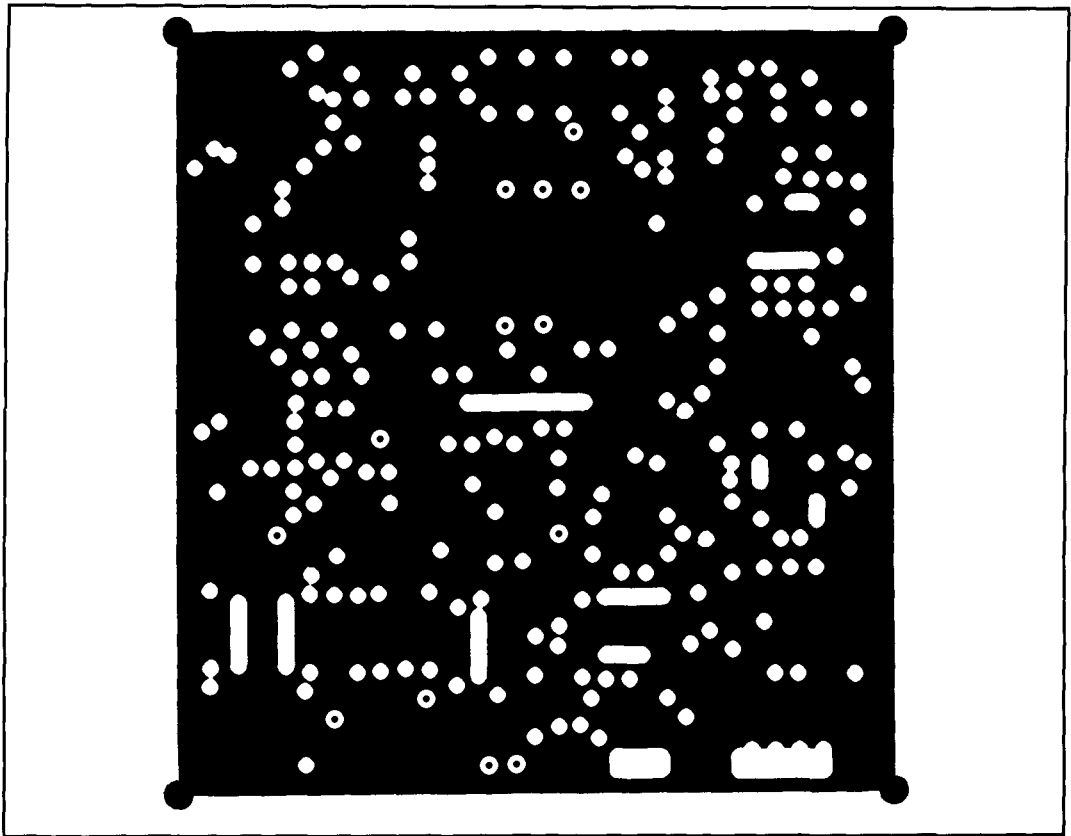
Figure 2. Teeny Twoer x1 component-side mask.

Craft Magazine in 1993. This particular rendition doesn't include squelch, a modification that shrinks and simplifies the layout. Also, unlike the Nor'easter, the Teeny Twoer uses broadband IF filtering to circumvent the inevitable drift associated with using a varactor-tuned 133-MHz local oscillator. An MRF901 preamp provides a quiet receiver front-end that delivers better than 0.3-dB AM sensitivity.

The transmitter has only three RF stages and is high-level modulated by a monolithic audio IC. Oscillator/tripler Q4 uses a 48-MHz third-overtone crystal to generate a 144-MHz signal, with frequency netting provided by L5. A spot-switch activates the oscillator at reduced output to aid receiver tuning. Driver Q5 operates in class AB for maximum gain, and is partially modulated from the center-tap of T3 to ensure full envelope depth on voice peaks. PA Q6 operates in class C and delivers up to 2.5 watts PEP output at 100 percent modulation. The transmitter is matched down to 50 ohms through an L-network and filtered for harmonic content with a half-wave low-pass filter.

Monolithic audio chip U5 provides a high-gain microphone preamp and AF power amplifier with few external components. T3, a low-cost 400-mW output transformer with a 120-ohm primary, provides a respectable match to the transmitter's resistive load.* U5's 100-k input impedance favors Hi-Z microphones, but Lo-Z dynamics may be used with passable results. With its 5-watt output capability, U5 has plenty of headroom and is conservatively rated for this application. Any quality Hi-Z dynamic or crystal microphone should deliver near broadcast-quality audio.

Construction

Masks for the Teeny Tower's printed circuit board are provided in **Figures 2** and **3**. Tinned pre-drilled boards are also available from FAR Circuits.** Once you have a board, construction involves little more than gathering up the parts and installing them as shown on parts-placement diagram **Figure 4**. Part substitutions are generally okay as long as the items fit the available space. For example, miniature ceramic discs like those provided in Radio Shack's 272-806 capacitor assortment may be substituted for many of the smaller-value monolithics. The circuit board doesn't provide plate-

$$*R_L = \frac{V^2}{2P_O}$$

Parts List

Capacitors
- 1 2.2-pF ceramic disc
- 1 4.7-pF ceramic disc
- 2 6.8-pF ceramic disc
- 1 10-pF monolithic
- 1 15-pF monolithic*
- 2 22-pF monolithic
- 2 27-pF monolithic
- 3 33-pF monolithic
- 3 47-pF monolithic
- 1 56-pF monolithic
- 1 100-pF monolithic
- 1 220-pF monolithic
- 1 470-pF monolithic
- 1 1000-pF monolithic (or 0.001)
- 7 0.001-µF ceramic disc
- 11 0.01 ceramic disc
- 4 0.1 ceramic disc
- 3 0.1 monolithic
- 2 1-µF electrolytic
- 4 10-µF electrolytic
- 3 100-µF electrolytic
- 1 220-µF electrolytic
- 1 470-µF electrolytic
- 3 3-22-pF trimcaps, Mouser 24AA022 or equivalent, 0.2-inch pin spacing

Resistors (1/4-watt, 5 percent)
- 2 10-ohm
- 1 22-ohm
- 1 47-ohm
- 2 100-ohm
- 6 470-ohm
- 4 1 k-ohm
- 3 2.2-k ohm
- 2 5.6-k ohm
- 6 10-k ohm
- 1 15-k ohm
- 3 47-k ohm
- 2 100-k ohm
- 1 220-k ohm
- 1 330-k ohm
- 1 1-Meg ohm
- 2 10-k ohm linear pot, 17-mm, Mouser 31CW401

Diodes
- 3 1N914 or 1N4148
- 1 1N34
- 1 6.8-volt 400-mW zener (1N5235B or equivalent)
- 1 9.1-volt 400-mW zener (1N5239B or equivalent)
- 1 1N4001
- 1 MV2104 varactor
- 1 green LED

Semiconductors
- 1 MRF901
- 1 2N3904
- 1 2N3906
- 1 2N5179
- 1 2N2866 or 2N5109
- 1 2N3553
- 1 NE602
- 1 MC1350P
- 1 LM358
- 1 LM386
- 1 TDA1015

Chokes
- 2 0.33-µH molded choke
- 1 2.2-µH molded choke

Transformers
- 1 10.7-MHz IF transformer, Toko 25:1, green core (T1)
- 1 3 turns #24 bifilar on T25-12 (T2)
- 1 AF transformer, 8 ohms to 120 ohms CT, Mouser TU120 (T3)

Coils
- 1 10 turns #24 air-wound on a 4-40 thread (L1*)
- 2 4 turns #24 air-wound on a 4-40 thread (L2,L3)
- 1 2-1/2 turns slug-tuned shielded, Coilcraft (L4)
- 1 14-1/2 turns slug-tuned unshielded, Coilcraft (L5)
- 1 6 turns #24 on T25-12 (L6)
- 1 4 turns #22 air-wound on a 10-32 thread (L7)
- 2 6 turns #22 air-wound on a 4-40 thread (L8, L9)

Other
- 2 SFE10.7 ceramic filter, 110-kHz, 50-kHz, or 25-kHz harmonic mode (FL1, FL2)
- 1 3rd-overtone crystal, operating frequency/3, ICM 471270 (Y1).
- 1 miniature DIP relay, DPDT with 12-volt DC coil (K1)
- 1 DPDT push-on push-off power switch, Mouser 612-7A211 or equivalent
- 1 SPST momentary, RadioShack 275-1571
- 1 5-pin pc-mounted DIN jack (or builder's choice, panel-mounting mic connector)
- 1 coaxial-type 2.1 x 5.5-mm power jack (or fused black/red wire pair)
- 1 SO239 coax jack (or BNC)
- 1 8-ohm speaker, 2-1/4 inch or larger
- 1 pc board, FAR Circuits
- 1 TO-18 high thermal resistance slip-on heatsink
- 1 mini-type crystal socket

Many of the specified items are available at reduced prices over the Internet from Dan's Small Parts at: <http://www.fix.net/dans.html>.
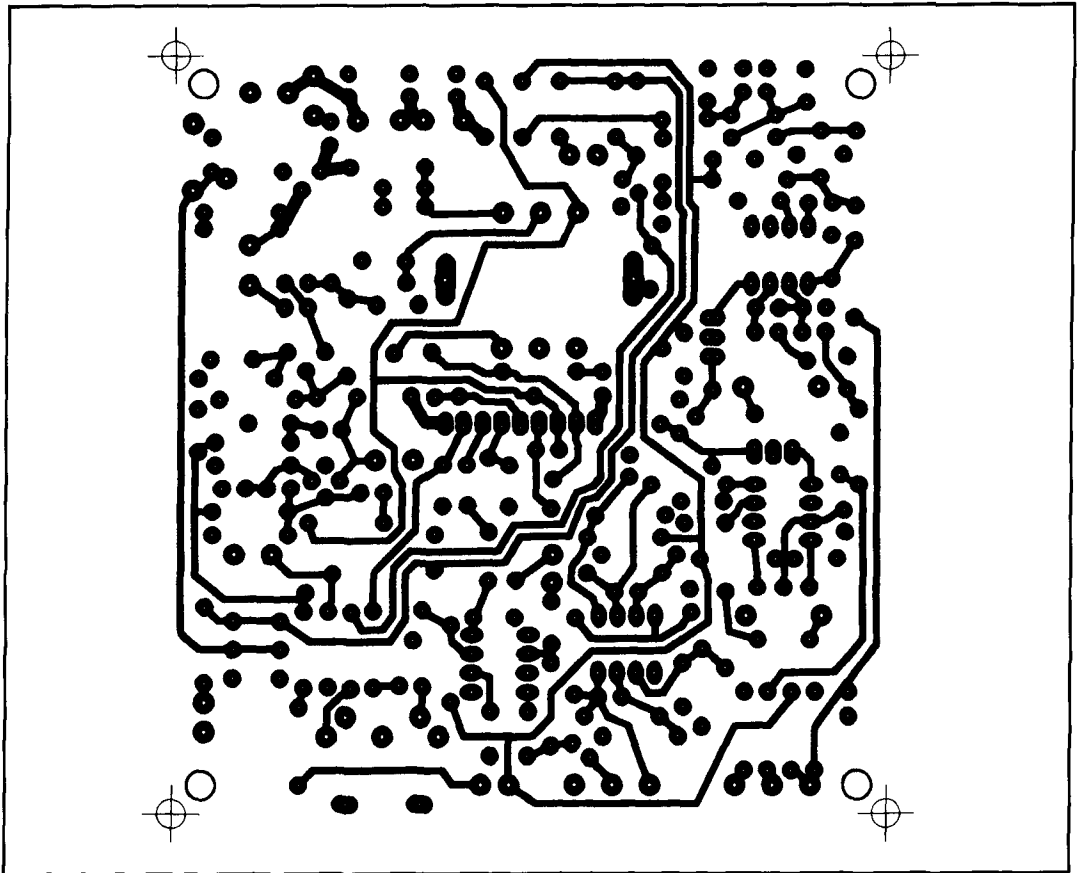
Figure 3. Teeny Twoer x1 solder-side mask positive.

throughs, so components with grounded ends must be soldered on both sides of the board. Remove excess flash from ceramic-disc capacitor leads before installation to ensure short lead-lengths and a good joint on top.

When mounting L4 and T1, it helps to rough up the metal just above the mounting tabs to ensure good solder wetting without overheating the coil. Note that the specified 10-k volume and tune controls are manufactured with the mounting tabs on the back side of the case. If you use a different type with tabs on front, positioning at the front panel mounting holes may be incorrect. A scale front-panel layout is provided in **Figure 5**. This drawing assumes 0.25-inch spacers are used to support the PC board in the case.

Filters FL1 and FL2 should be broad enough to tolerate mild LO drift and the normal frequency differences between AM stations. At the same time, they should be sharp enough to reject SSB, beacon, and packet activity up and down frequency from the AM operating frequency. Type SFE10.7 ceramic filters come in a variety of bandwidths. The readily available 110-kHz WBFM versions are adequate for casual operation, but narrower 50- and 25-kHz provide a better balance. I used the MuRata

SFE10.7MV 25-kHz harmonic-mode filters in my prototype with excellent results; but these remain somewhat difficult to find in the U.S. except as engineering samples.[1] If tandem 110-kHz filters prove too wide for use in your operating area, consider replacing FL2 with an inexpensive two-pole 25-kHz channel-spacing monolithic crystal filter. This is available from Digi-Key for $3.41 and may be installed as a direct drop-in replacement at FL2 without special matching.

When mounting the driver and PA transistors, be sure to space the cases about 0.1 inch above the grounded surface of the pc board. It's always a good idea to keep transistor leads short at VHF, but parasitic capacitance between the metal case and pc board will make tuning difficult if the gap is too small. Also, the 2N3553 and modulator IC require heatsinks. Any high thermal-resistance TO-18 slip-on should provide adequate heat dispersion for the PA, and a rectangular piece of aluminum about the size of a large postage stamp should take care of U5. When installing the MRF901, note that the long lead is its collector (also marked "M" on the top of the case).

All air-wound coils are formed by winding enameled wire on the threaded shank of stan-
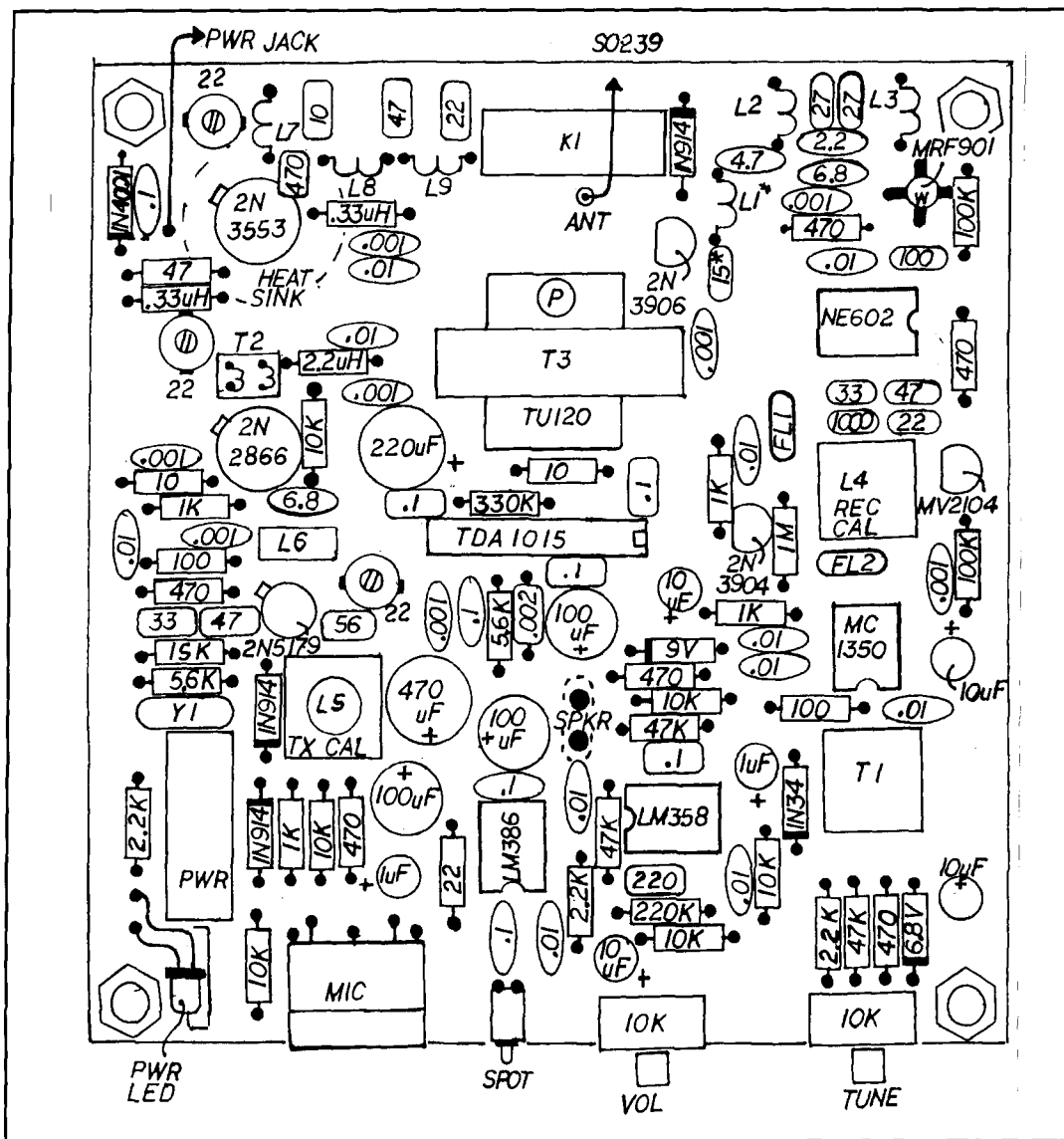
Figure 4. Parts placement diagram.

dard-size machine screws. T2 is a bifilar-wound toroid coil, as shown in **Figure 6**. To wind, hold two short lengths of #24 wire together and pass them through the center of the toroid three times (turns on toroid forms are counted on the inside). Make sure the paired wires remain parallel during winding and are pulled snug against the form with each turn. Use an ohmmeter to pair the lead ends, as shown on the construction detail. Windings may be separated and spread out over the form after installation. When installing T3, position the side marked "P" toward the rear of the pc board. This is the 120-ohm side of the transformer.

Plastic cases look good and are easy to prepare, but they provide no immunity from computer noise on the operating desk. For that reason, I installed the Teeny-Twoer pc board in a small 4 x 4.25 x 1-1/2-inch homemade alu-
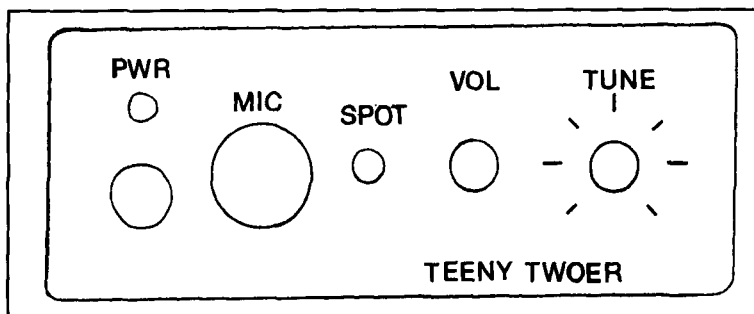
Figure 5. Front panel layout.

minum box (see **Photo A**). A built-in 2-1/4-inch speaker provides good, if not live-and-in-person, audio reproduction (there's plenty of space to install a speaker/phone jack on the back panel if you crave "hi-fi" reproduction.)
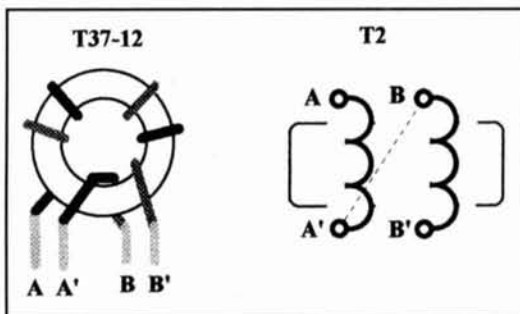
Figure 6. Before installation, use an ohmmeter to pair windings, as shown. When the inductor is installed, the A and B windings will be phased by a trace underneath the board.
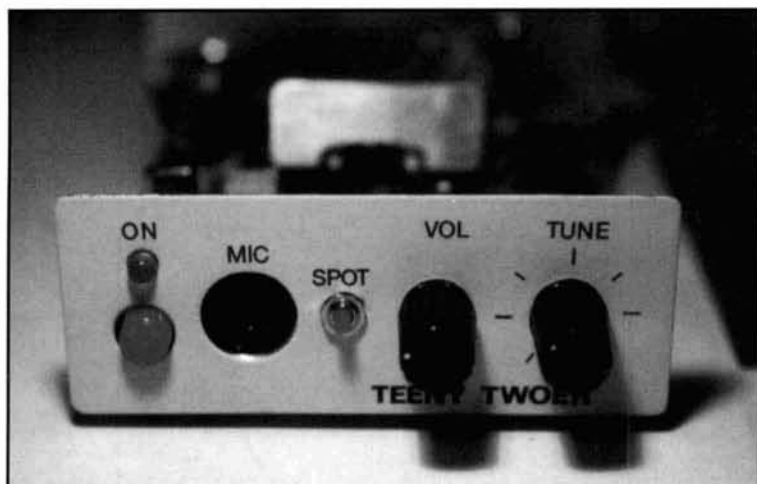
Photo A. Installation in an aluminum box will shield the Teeny Twoer from outside noise.

When laying out the box, take care not to mount the rear-panel power jack where it will obstruct access to the PA tuning trimcap.

Tune-up and Operation

Receiver tune-up requires a calibrated weak-signal source, preferably a VHF signal generator with AM modulation capability. To start,
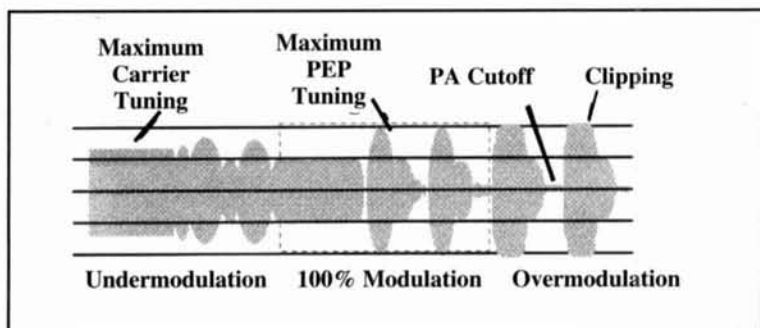
Figure 7. Adjust for maximum carrier and "talk-power" suffers because the transmitter is not tuned for peak power levels. When correctly tuned, modulation peaks reach *twice the amplitude and four times the power level* of the unmodulated carrier. Overmodulation should be avoided—this condition compresses or clips the envelope on peaks and cuts off the PA on nodes, causing distortion and splatter.

adjust T1 for a noise peak. Then, set the generator up for 144.4 MHz at -80 dBm. To align the local oscillator, set the receiver's tuning potentiometer at mid-range (12:00), and adjust L4 until you pick up the generator signal. To peak the front end, reduce generator output for noisy copy and adjust RF coils L2 and L3 by stretching or compressing turns for maximum signal. Finally, repeak T1. The receiver tuning range should be just under 1 MHz, allowing you to check the entire bottom portion of the band for activity with a quick sweep of the dial.

If you choose to install the optional airband image filter, set the generator to 122.9 MHz and slowly increase the output level until you locate the image signal. Compress or stretch L1 to find a sharp pronounced null. Image rejection should exceed –70 dB when the filter is tuned, but there may be a 3 to 4-dB in-band penalty with it in place.

Begin transmitter alignment by pressing the spot switch and adjusting the oscillator trimcap for strongest carrier in the receiver. Next, connect a power meter and dummy load, and adjust all three transmitter trimmers for maximum carrier output. If everything is working correctly, output should measure 1 watt or more with a 13.8-volt supply.

To complete the tune-up, sample the transmitted signal with a high-frequency scope and readjust the PA output trimcap for maximum envelope amplitude with modulation applied. This tuning step is important because it optimizes the transmitter for peak modulated output rather than for best carrier strength (see **Figure 7**). If you don't have access to a 100-MHz scope, tune the PA for minimum downward modulation on a power-meter while speaking into the microphone. When properly adjusted, the transmitter should develop approximately 2.5 watts PEP on voice peaks with an unmodulated carrier level of 0.7 watts (see **Photos B** and **C**).

Before operating on the air, there are two things left to do. First, net the transmitter on frequency by sampling its output frequency with a counter and adjusting VXO coil L5. Second, check modulation percentage with normal speech using your microphone of choice. This may be done by viewing the waveform on a scope or by tuning the sidebands for splatter. The Teeny Twoer doesn't provide a microphone gain control and, with up to 5 watts of modulation available, the combination of a strong voice and a high-output Hi-Z mic can easily overdrive it! If adjusting your mic technique is insufficient to control modulation level, pad down the input circuit at pin 8 of U5, or alter the value of the 5.6-k coupling resistor on pin 6 for reduced gain. I use an old EV-674 dynamic microphone wired for high-Z output,

and find that speaking naturally about 6 inches from the element yields 100 percent modulation with studio-quality results.

Picking a Good AM Operating Frequency

In the old days, it was common practice to pick a spot somewhere near the traditional 144.4-MHz AM calling frequency and blast away. Today, however, Packet and APRS have moved in, pushing AMers in some parts of the country to other frequencies. If you have a drawer full of 144.4-MHz crystals, this change may not come as welcome news. However, unlike 50.4 MHz, 144.4 has never been designated as "official." Consequently, no replacement was considered during recent APRS discussions involving the ARRL Board. This means you're on your own as you try to find existing local activity, or to choose an AM channel that causes minimal interference to existing services. When operating near the weak-signal portion of the band, please avoid transmitting in the 144.275 to 144.300 propagation beacon zone.

Using the Teeny Twoer with an All-mode Brick

At first glance, you might conclude the Teeny Twoer's output level is too low to drive an all-mode brick. Not so! A close inspection of the P-in versus P-out specifications for several 10-watt input amplifiers revealed that most can be driven into gain compression with 5 watts or less. Indeed, some imported HT boosters may saturate with as little as 2 watts. This means it's best to ignore the 10-watt input specification, a claim that usually refers to maximum permissible drive for non-linear FM service. For AM linear service, the ideal carrier level is much lower than that—normally only 25 percent of the input power required to produce 1-dB gain compression. For many two-stage amps, that's 1 watt or less—exactly what the Teeny Twoer delivers! Any more drive will yield only wasted carrier energy, lower modulation percentages, and excessive heat.

Conclusion

As any veteran Benton Harbor Lunchbox owner will tell you, 1 watt on 2 meters can cover amazing distances—especially with the aid of a good antenna and a boost from Mother
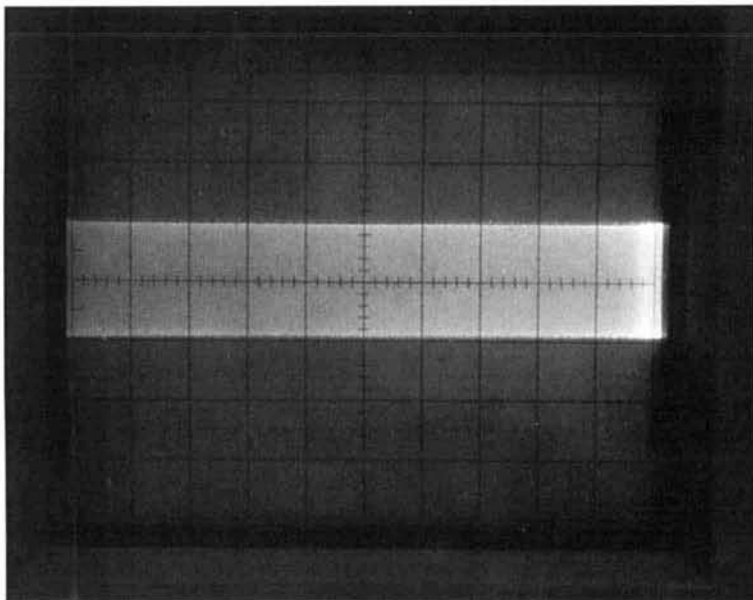
Photo B. Unmodulated carrier.
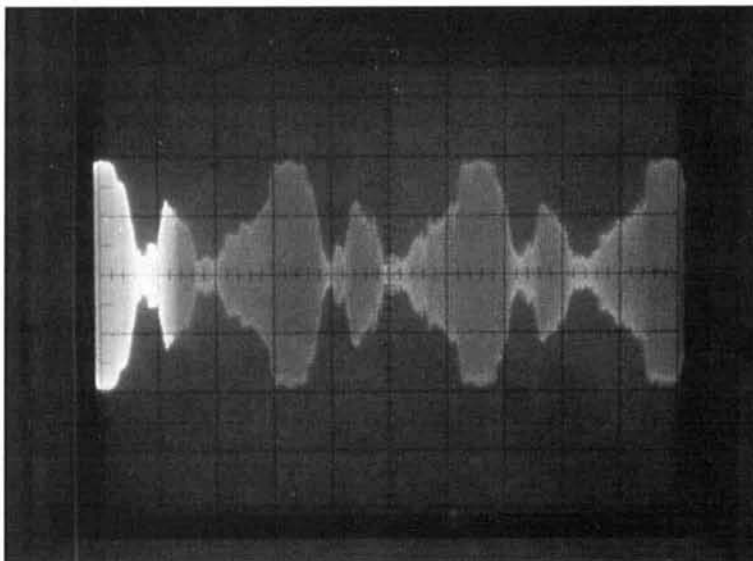
Photo C. Carrier modulated by an audio signal.

Nature. Consequently, when you hear distant action on the band, there's no need to be shy about jumping in. If any of the big shots on boat-anchor alley make fun of your little radio, you can always grab a brick and give 'em an S-meter bending reminder that big things sometimes come in small packages! ∎

REFERENCE
1. Rick Littlefield, K1BQT, "Quarterly Devices: New Narrowband 10.7-MHz Filters from Murata," *Communications Quarterly*, Summer 1994, page 89.

both three dB below theoretical and nearly three dB above theoretical.

In his **Figure 6**, at 10 kilometers, FS is over 3 dB down from theoretical, indicating less than 50 percent antenna efficiency. At the start of the curve, measured FS was between 1/2 and 1 dB more than theoretical (indicating more than 100 percent efficiency, an error in FS or conductivity estimates, an error in FS readings, or a combination of any or all).

If we move the measured FS at the start of the curve down to the theoretical 100-percent efficiency point for the ground conductivity selected, one FS reading at 9-1/2 kilometers indicates about 5 dB shortfall while the other predicts a 3-1/2 dB shortfall!

In **Figure 7A**, one 9 kilometer reading is just over 4 dB below theoretical. At **Figure 7B**, the shortfall is about 4 dB at 7 kilometers. This scatter is typical in FS readings of this type, and a valid reason why indirect FS comparisons *should never be used* to absolutely "prove" performance differences. Data like this have been the only "real proof" offered of model and real world agreement, even though there are far too many variables influencing results.

Error-prone measurements should never be used to verify models, especially when errors found in the data approach or exceed the amount in dispute. This is especially true when better methods with less error are readily available.

If measurements were made at fixed points while changing only the radial system, and if the operator carefully made measurements using the same equipment the same way on the same scale, the only error would be how closely the scale tracked over a very small range. Errors would be greatly reduced or eliminated, since many variables would be removed from the measurements (including use of the graphs and FS slope with distance). This is why I endorse direct A-B testing of one system, with only a radial system change.

Weber assumes even more, believing current measured at one point tells everything about a complex resonant system's performance. He did not supply a single simple field strength measurement supporting his claims, even though the only practical way to measure efficiency change of a large antenna is through the field strength change.

One doesn't need to look far into Weber's theory to find flaws. The third paragraph plainly indicates quarter-wavelength long elevated radials present an "impedance of $0.0 + j0$ ohms." That statement is repeated throughout the text. A radial with a 0-ohm impedance, however one might measure a single terminal system, could have no power loss through radiation (either as heat or a field of some type). If radial impedance is indeed $0.0 + j0$ as Weber

suggests, the radial isn't radiating radiation fields. It also isn't producing induced losses and can't be involved in any form of energy conversion to heat. Current balance isn't an issue at all in such a radial, because the radial isn't radiating anything!

Of particular importance is the statement on page 99, "Using 45 degree radials essentially eliminates unequal radial currents and produces no extraneous energy." Every real-world radial radiates in every system, unless the radial is viewed from one particular point in space where the field under consideration is canceled by some other field. We shouldn't confuse far-field conditions with near-field conditions where earth, trees, telephone wires, and other conductors or lossy dielectrics are closer to one radial wire than another radial generating an opposing and equal field at that point. The radials being discussed are in the near-field zone of earth, and no matter what we do with current balance, the radials still couple to earth (and other objects near the radials) by radiation and induction fields.

Weber's data does not "prove" efficiency or performance improvements, and might very well prove just the opposite. Neither ground conditions or surrounding objects are perfectly distributed and symmetrical in characteristics around the antenna. A consequence of this is that radials will almost certainly radiate differing amounts of energy even when carrying exactly equal currents. Re-radiation from the ground and other objects around the antenna also changes the far-field pattern. Forcing equal currents can actually result in lower efficiency and more undesired radiation in a non-homogenous environment!

Like everyone else, I don't like to waste money and time installing needless radials. Several weeks ago (in preparation for installing a new 160 array on my farm), I measured FS at 12 carefully fixed far-field points free of multipath. (I have an almost ideal location, as there are no buildings, power lines, or other vertical obstructions for miles in nearly every direction.) With four 1/8-wavelength long radials resonated by high-Q loading inductors (Q>450), base impedance was about 32 ohms. With no change (including no change to true applied power), removal of loading and extension of radials to a 1/4 wavelength improved FS by almost two dB! (At the same time, base impedance increased to just over 40 ohms, dispelling the popular hammyth base impedance changes substitute for far-field FS measurements.)

Grounding the common point of the elevated system to three 6-foot ground rods spaced around the tower produced almost 1 dB of additional loss!

The same radiating system, with no change

other than radial system changes, showed nearly 1-dB average improvement when 16 on-ground radials were substituted for the four elevated radials. Base impedance was notably higher with the 16 radials, yet far-field FS was stronger and the system could be grounded for lightning protection with no loss in FS.

Part of my test involved A-B over-the-air comparisons between the test vertical and a dipole 110 feet in the air (located about 500 feet east of the test antenna), and another 1/4-wavelength vertical over a large ground mat several hundred feet from the test antenna. The test antenna was installed over virgin pasture, with no conductors within two wavelengths of the antenna. In every case, ground wave field strength measurements tracked over-the-air results. When 60 full-length radials were added, the test vertical almost always tied the other two antennas in performance. When four elevated radials (isolated from ground at the common point) were used, virtually every contact reported the test vertical was noticeably weaker. When four inductor loaded radials (with carefully balanced currents) were used, performance was even less satisfactory.

Based on two groups of measurements I made at two very different locations (one over mostly rock in a suburban location, one over

the very thick topsoil of an open pasture), WVNJ's results, and verbal reports of other test results, I can't state beyond any doubt that small elevated radial systems will never work in every situation. Belrose's data, selectively taken from one particular station and processed through a series of estimates, prove even less. Weber's endorsement of an even smaller counterpoise system, based on almost no data at all except unimportant current measurements and personal opinion, means even less.

It would be wonderful if proponents of "space and labor saving systems" supplied field strength readings from controlled A-B tests, tests where ONLY radials are changed. Until they do, and until others repeat or confirm their results, I'll keep using a system that measures better in A-B tests, and has always worked without question or controversy for years and years.

Tom Rauch, W8JI
Barnsville, Georgia

A response to W5OLY on the Conjugate Match

Dear Editor:
The statement by Bruene on page 28 of the Spring 1998 issue, "They then reference Sabin,

who is also wrong on this subject," is an incorrect statement that should be retracted. The *QEX* article, written by me, that Belrose et al referenced, was incorrectly used by them. My *QEX* article is very definite on the subject of the Norton equivalent circuit. It specifies an AC current source and a dynamic AC plate resistance. When the AC load is varied until it is equal to the AC plate resistance then, for a certain fixed value of AC current, the AC power into the AC load is a maximum. This point was also discussed in my Letter to the Editor in the Winter 1998 edition of *Communications Quarterly.*

Also, regarding Bruene's statement on page 30, regarding "Direct measurement of output impedance": this is another misrepresentation. The method that I proposed, and used in my Letter to the Editor (performed at 3.8 MHz to reduce detuning), was described in my *QEX* article, and that article carefully explains that the change of load resistance must be quite small so that the tuned circuit will not be significantly detuned, and that if detuning is a problem then some minor retuning may be needed. Also, the peak-to-peak plate/collector/drain voltage must not invade highly nonlinear regions. The load resistance change that I have used is less than 9% low, not the 50% low or 100% high that Bruene mentions, which would be very wrong. The 9% low will cause a small detuning effect, but not enough to invalidate an "estimate" of the "magnitude" of the output impedance. This test has to be done carefully, but it is at least as good as any other method, and a lot simpler, in my opinion.

William E. Sabin, WØITH
(via Internet)

Some suggestions

Dear Editor:

Here's a response to your suggestion that you be contacted online. I've been a subscriber from the beginning and enjoyed most issues and learned a lot from them. But for a magazine devoted to communications, the S/N ratio could be greatly improved in many articles with a bit of editorial attention to the communications aspect.

Consider Joe Reisert's article. There's lots of useful figures to help explain his test, but some-

body (art editor or electronic composition manager with no concept of information transfer?) has lost sight of the fact that they are important to the text. We don't get mention of the figures until page 58 and then have to turn back to page 49 to find **Figure 1**, with the other figures equally disembodied from their parent text. That isn't the way to make reading or learning easy! John Evan's article on Jodrell Bank goes in the opposite direction and **Figures 8** and **9** mentioned in the text on page 71 appear to have been just lost. It almost went correctly for all of Mac Chapman's article (the "Quarterly Computing" one)!

I appreciate that the process of translation from computer screen to paper can lead to such confusions. But if production staff perceptual processes get a bit mired in such transfers, what's wrong with taking what one might label the trf approach and laying out a magazine of the relatively small size of *Communications Quarterly* with scissors and paste to get it right? The ease of imparting information and not the appearance is what is most important.

The second paragraph of your editorial affirms your purpose in looking for advances in communications technology. That's why I subscribe, but please do it in a way that makes the magazine instructive and helpful rather than as though the aim is the magazine equivalent of a coffee table book.

Hector Maclean, VK3CEC
(via Internet)

I agree with W7DHD

Dear Editor:

I am in complete agreement with Bill Byron, W7DHD, in his letter on elevated radials in the Spring issue of *Communications Quarterly.* We do need to see measured results.

Einstein once said that theory merely tells you what to measure. His theories were not completely accepted until they were proven experimentally.

Of course, some important theories can never be proven by measurement, such as the Big Bang and evolution. All we can say is that with what measurements and data that we do have, they seem most plausible.

Harry R. Hyder, W7IV
Tempe, Arizona

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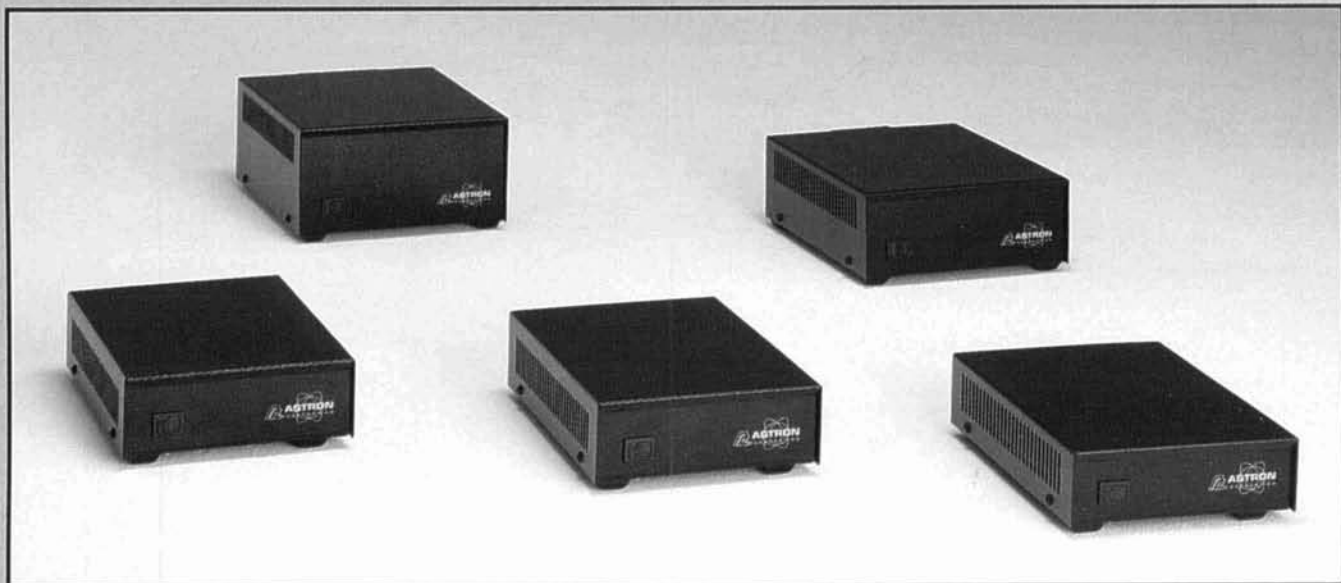
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