QEX

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July/August 2009

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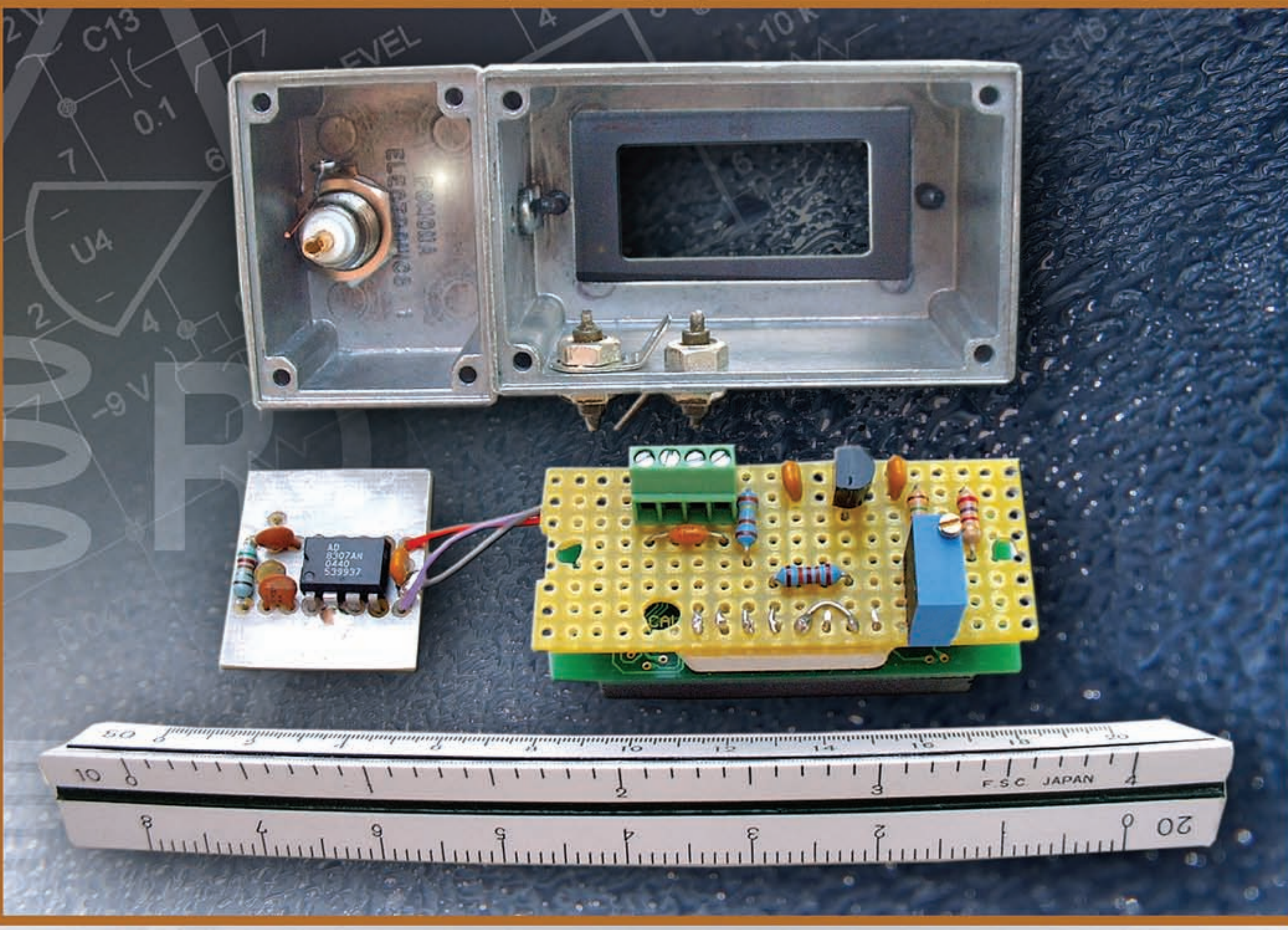

AF8L describes an RF power meter that is part of his Cybernetic Sinusoidal Synthesizer. An AD8307 logarithmic amplifier board is isolated from the rest of the circuitry by using separate cast aluminum project cases.

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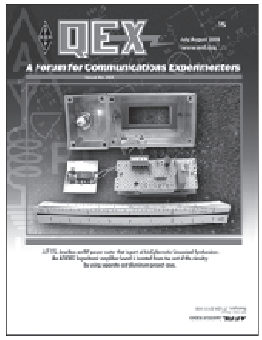

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2) document advanced technical work in the Amateur Radio field, and

3) support efforts to advance the state of the Amateur Radio art.

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Larry Wolfgang, WR1B

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Empirical Outlook

Dayton

As I write this, the 2009 Dayton Hamvention and ARRL National Convention are recent history. The ARRL Expo area had a number of new features this year. I can't take any credit for the success of any of part of the Expo, but I was sure proud to be able to participate in some of the excitement!

ARRL Lab Test Engineer, Bob Allison, WB1GCM, organized the first kit building section of ARRL Expo. He was assisted by ARRL Education and Technology Program Coordinator, Mark Spencer, WA8SME and Nathan McCray, K9CPO, all weekend. Several others assisted with a number of of the building "classes," including yours truly. Bob had two kits available for the builders — either a 24 hour digital clock or an electronic keyer. These kits were designed by Mark Spencer, and packaged by Mark's wife Doris, KF6QKL.

Bob reports that over 40 kits were built during the weekend. While in raw numbers that doesn't sound like many, the fun and excitement on the faces of those builders was well worth the effort! Builders ranged from pre-teen to 70+ year olds, with many first timers and a few who just hadn't built anything in a while and wanted to try their hand again under a watchful eye. At one point there was a young pre-teen boy who was building a clock kit with some help from his mother. After he carefully soldered his first resistor and then clipped the excess leads he looked up with a huge grin and exclaimed, "I learned how to solder!" His smile was even wider when he first applied power, and the clock display came on. There were several teens, including at least one General class ham, who had some building experience, and wanted to work on their kits with little or no additional help.

I was able to offer a few tips, such as always wear your safety glasses when soldering or clipping component leads, point the circuit board slightly away from your face and place a finger on top of that lead before you cut it, to help control it and prevent it from flying away (or towards an eye).

I also talked to quite a few Hamvention attendees who were very pleased to see ARRL sponsoring an activity like the kit building, and helping teach some youngsters (and some not so young) the joy that they have experienced over the years. Of course the real lesson here is that ARRL is not just a group of Headquarters employees and a few volunteers doing this at a major convention. ARRL is every member, and those members — both individually or as part of a club — can take advantage of many opportunities to share the thrill of building equipment. I hope we helped inspire many more hams to teach others about the fun of building a project.

In addition to Amateur Radio, the Scouting program is one of my favorite pastimes. ARRL has a long history of supporting Scouting and Amateur Radio programs, including the Radio Merit Badge, the annual Jamboree on the Air (always the third full weekend of October) and an Amateur Radio presence at National Jamborees, typically held every four years. As part of this year's Dayton ARRL Expo area, a group of dedicated Scout Leaders set up and staffed a space for Scouts and Leaders to meet and share ideas for introducing Scouts to the fun and excitement that we know as ham radio. While I did not help to staff this area, I was able to meet with many Scout Leaders, and it was a lot of fun to share ideas and learn from many of them.

I also spent quite a bit of time around the TAPR booth. This was an opportunity to talk about some of the projects that TAPR members are working on, and to help explain some of the displays operating there. While a 15 foot contact on 6 meters isn't all that exciting, making that contact with a pair of transceivers as part of the high performance software defined radio (HPSDR) project was exciting. I have seen the HPSDR setup a few times now, at a couple of ARRL/TAPR Digital Communications Conferences and last year at Dayton. (For more information about the 2009 DCC in Chicago, Sep 25 - 27 see Upcoming Conferences in this issue and **www.tapr.org**.)

The HPSDR project has definitely evolved! This year, the radios included the "Atlas" backplane with "LPU" power supply, "Penelope" ½ W transmitter, "Mercury" receiver, "Ozy" interface/controller cards plugged in along with "Alex" external filter cards. The systems were mounted in "Pandora" metal enclosures that were being sent for screen printing of labels at the end of Hamvention. Most of the boards are either available from TAPR or expected to be available shortly, including the cabinets (**www.tapr.org**). There was also a "Pennywhistle" 20 W amplifier module prototype on display.

One common question about the HPSDR display was, "Why should I be interested in HPSDR rather than a (commercial product)?" The simple answer is that if you want a finished, commercial package there are a number of software defined radios available, but if you want something that requires some building, and invites experimentation and more learning about the hardware and software, then you might be interested in HPSDR. I am pretty sure that many *QEX* readers are interested in learning about and experimenting with technology, in one form or another. I hope you are having as much fun with our hobby as I am!

Gary Steinbaugh, PE, AF8L

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A Cybernetic Sinusoidal Synthesizer: Part 3 An RF Power Meter

This part in the series describes some advanced control theory and the RF power meter module

The crystal oven of the ovenized crystal controlled oscillator (OXCO) presented in Part 2 employs a proportional controller that maintains the crystal temperature at its upper turning point for a reasonable range of ambient temperatures. Since this oven is an easily controlled process, a more sophisticated controller would be unproductive. All proportional controllers, however, have a weakness that is recognized by many control engineers, but fully understood by few. More demanding processes, like the upcoming RF level controller and the phase locked loop (PLL) frequency synthesizer, require attention to this problem.

Proportional Action's Dark Secret

To help with visualizing the proportional controller's weakness, I have invented a liquid level control example comprised of a tank, a float-operated input valve, and an identical output valve that is manually operated (see Figure 26). Now, in order to maintain a constant liquid level, both valve handles must be in the same position (liquid in equals liquid out) as shown in Figure 26A. Increasing the load by moving the output valve handle will result in the condition shown in Figure 26B; since liquid out is greater than liquid in, the level will fall, as will the float, opening the input valve (Figure 26C). The level will continue to drop until the position of the valve handles are the same. But now the liquid level is no longer at the original setpoint! This phenomenon, called proportional

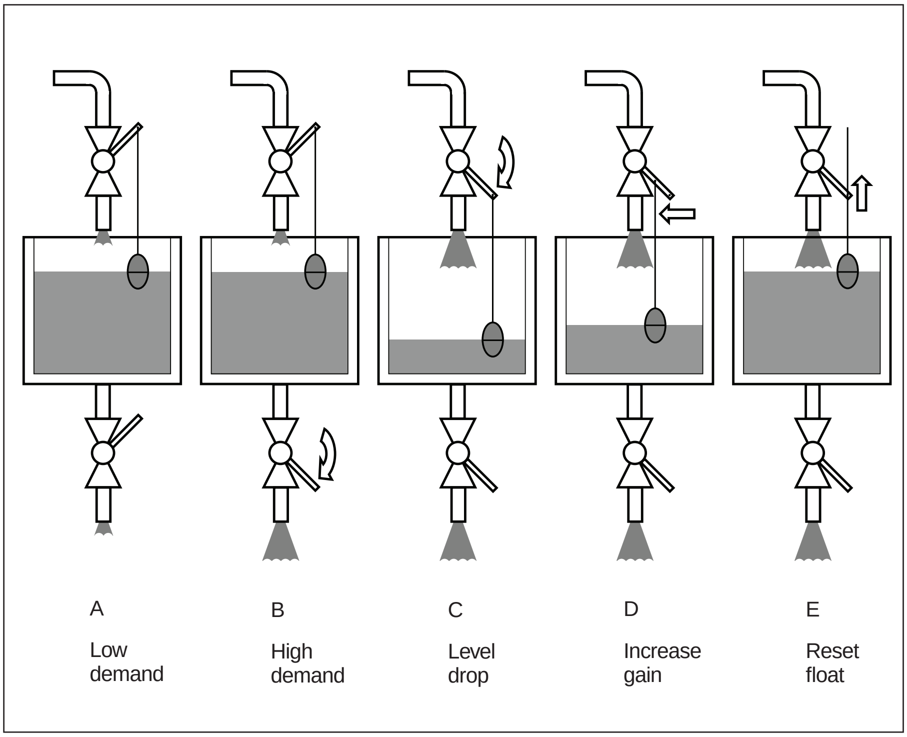

Figure 26 — This mechanical water-flow control system demonstrates why proportional control systems will always exhibit proportional droop. In Part A, both the input and output valves are set to the same low flow, and the water level in the tank will be maintained at the desired setpoint. At Part B, the output valve has been opened to satisfy a high demand, but the input valve has not yet been opened to maintain the tank water level. Part C shows that as the water level in the tank drops, the float will drop and open the input flow valve. The water level in the tank will again remain at a steady level, but the new level is far below the desired setpoint. At Part D, the sensitivity of the control is increased by moving the float rod closer to the valve body on the handle. Part E demonstrates the effect of adding *proportional-integral* control action by periodically resetting the length of the float rod.

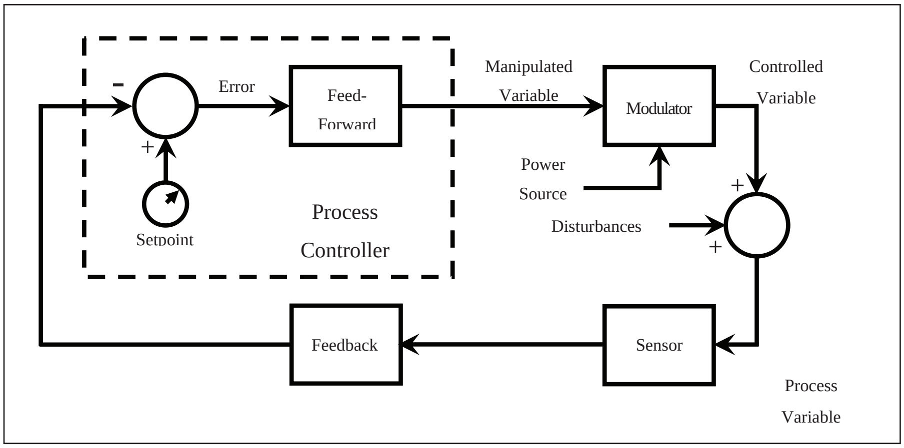

Figure 27 — This slightly revised version of Figure 1 from Part 1 in the Mar/Apr 09 issue shows a generalized feedback loop. Samples taken at the loop output are fed back and subtracted from the loop input to produce the error signal in the process controller section.

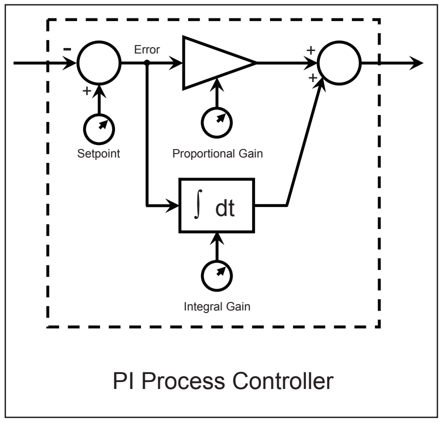

Figure 30 — This diagram represents the addition of integral action to proportional action in the process controller diagram. This is known as a proportional integral (PI) controller.

droop, *must* occur with any proportional controller, regardless of its implementation (liquid, pneumatic, mechanical, electrical, or whatever).

To return the level to the intended setpoint, one might try increasing the sensitivity of the input valve by moving the float rod closer to the valve handle's pivot point (Figure 26D). This partially raises the level, but any slight surface ripples will jiggle the input valve's handle, causing spurts of liquid that will jiggle the handle even more, possibly resulting in instability. High proportional gain settings will *guarantee* oscillations, since the system will operate as an on-off (control engineers say "bang-bang") controller.

A better approach is periodically to reset the length of the float rod (Figure 26E); this produces a *proportional-integral* control action. To explain this, I have rearranged the generalized feedback loop diagram (Figure 1 from Part 1) by isolating the process controller from the other loop components, as shown in Figure 27. Figure 28 focuses on the controller as a stand-alone component. The proportional gain setting essentially determines how *unhappy* a loop is with a given error (the difference between the setpoint and the actual process variable). In the OXCO described in Part 2, the proportional gain is preset to 18 by the ratio of R16 and R17 (see Figure 5 in Part 2); the inquisitive reader might replace these resistors with a 250 kΩ potentiometer to experiment with the effects of varying the proportional gain.

Integration is a simple mathematical operation that can be modeled by a hopper partially full of pebbles (Figure 29). If we throw one pebble per second into the hopper, the level will rise slowly; if we throw two pebbles per second into the hopper, the level will rise twice as fast; if we remove pebbles from the hopper, the level will fall, and so on.

An integration over time is symbolized by $\int dt$; the fancy "s" means "sum," and dt refers to infinitesimally small amounts of time. Incidentally, the Latin word for pebble is "calculus." (The Laplace transform of $\int dt$ is 1/s, where $s = \sigma + j\omega$.)

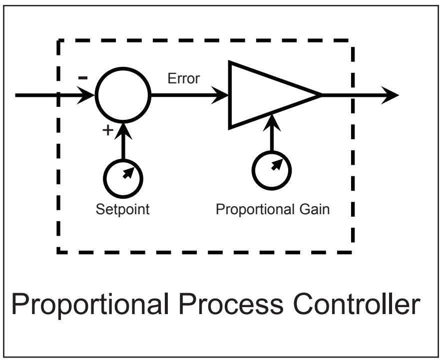

Figure 28 — Here we have added proportional control to the process controller block.

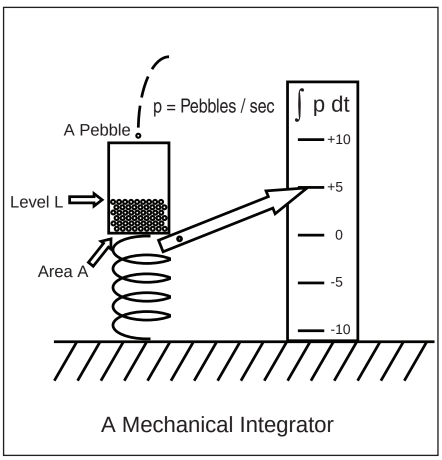

Figure 29 — This drawing illustrates a mechanical integrator. As pebbles are added or removed, the output reading will increase or decrease.

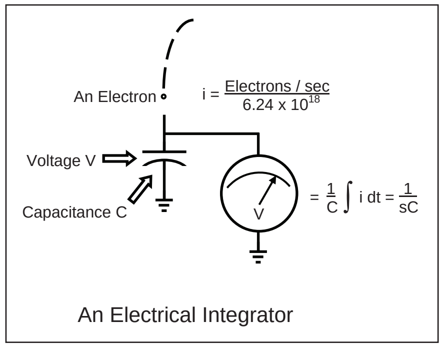

Figure 31 — An electrical integrator consists of a capacitor and some electrons. As electrons are added to the capacitor, the voltmeter indication will increase.

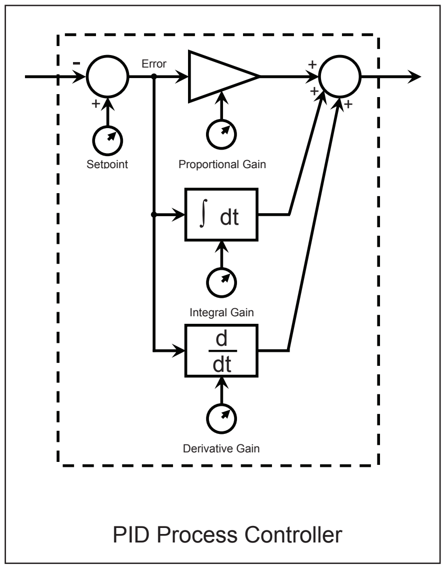

Figure 32 — Adding differential (rate) action to the process controller, here we have the proportional, integral, derivative (PID) controller. The derivative action will react to fast changing errors in the control process. The Cybernetic Sinusoidal Synthesizer does not require a PID controller. Remember to use the simplest control scheme that will produce satisfactory results.

Think of the error as pebbles, added faster for larger errors. As long as there is any error, the pebble level (the integral) will increase, and the larger the error, the faster the integral grows. If we drive the input valve in Figure 26 from both the float and the weight of the pebbles, the valve will continue to open as long as the droop error is present, opening slower as the error decreases. The valve will eventually open enough to force the error from proportional droop to zero.

Adding an integrator produces the PI controller shown in Figure 30. There is now another adjustment — the integral gain setting — that essentially determines how *impatient* a loop is with a given error. If the integral gain is set too high, the hopper level will grow so quickly that the valve will still be opening when the setpoint is reached, resulting in an overshoot while the excess pebbles are removed. Oscillations may again result, similarly to the shower example in Part 1, in which *you* were a PI controller with too much integral gain, twisting the hot water knob instead of making small, patient corrections. Of course, too little integral gain allows proportional droop to reoccur. In control engineering terminology, adding integral action can improve steady-state conditions at the expense of transient response.

An electrical version of integration can be implemented with a capacitor and some electrons (see Figure 31). Note that the voltage V is analogous to the hopper's level L, and the capacitance C is analogous to the hopper's base area A; a larger capacitor will fill up more slowly. Also, recall that a current I of one ampere equals the flow of one coulomb (6.24×10^{18} fundamental charges) per second, and all the units will come out correctly. The familiar RC low-pass filter can be thought of as a kind of integrator (its Laplace transform is $(1/s\,C) / (R + 1/s\,C)$, or $1/(s\tau + 1)$ where τ is the RC time constant). Such filters will be used in the RF level controller and the frequency synthesizer.

Although not used in these RF modules, most industrial controllers offer *proportional-integral-derivative* action (PID), as illustrated in Figure 32. A differentiator is basically a speedometer; the object of adding derivative action is to catch fast-changing errors before the integral has a chance to grow too large. This adds yet another adjustment, the derivative gain setting, which essentially determines how *nervous* a loop is with a given error. Certain types of loops can

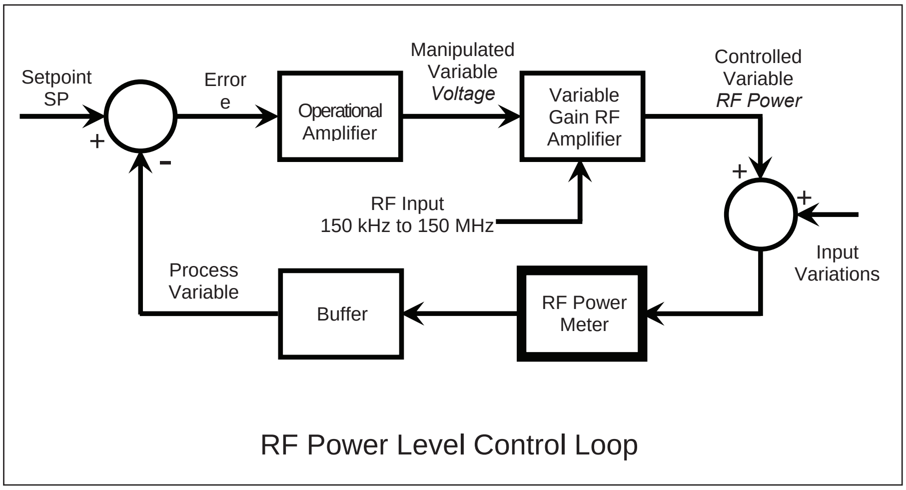

Figure 33 — This block diagram illustrates the control system for the RF power level control loop. We will concentrate on the RF power meter section in the remainder of Part 3 of the article series.

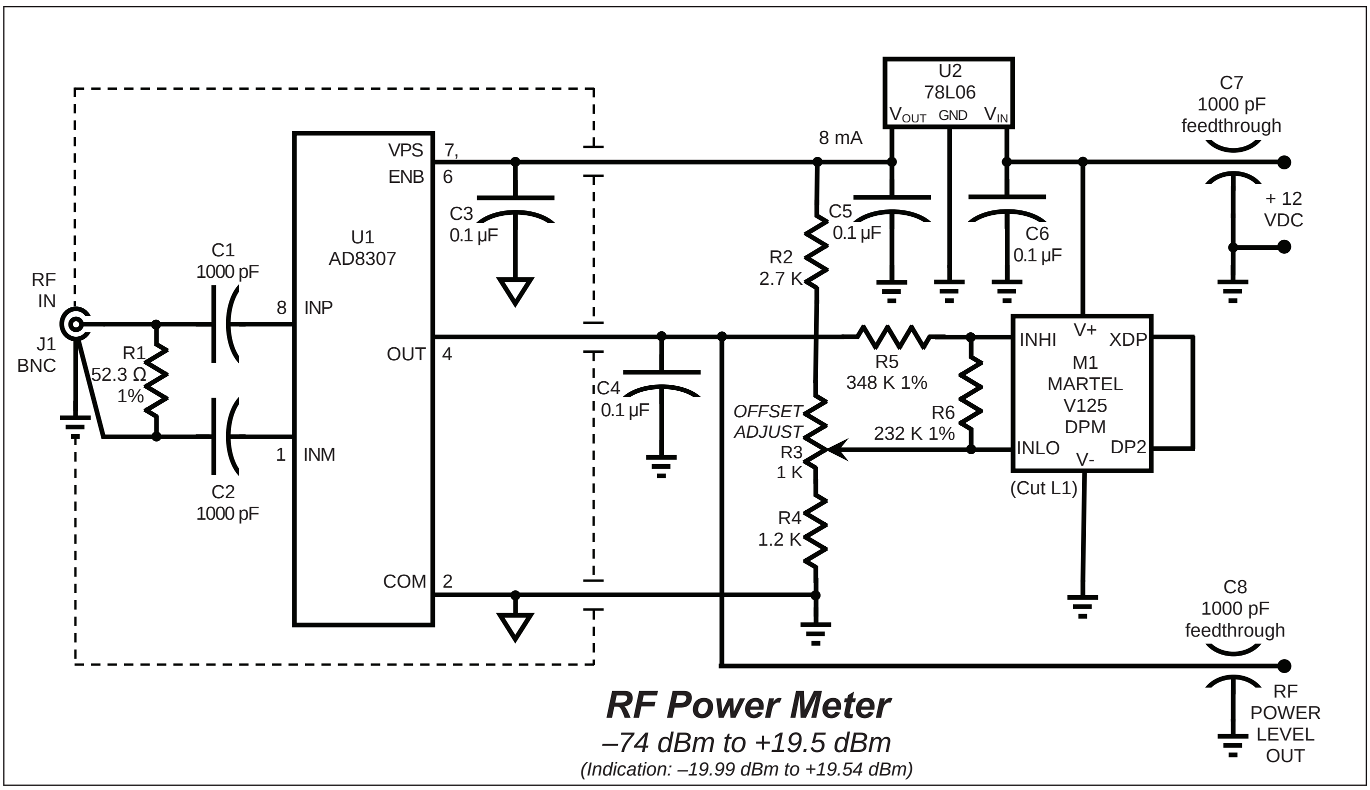

Figure 34 — The schematic diagram for the RF power meter is given here.

benefit greatly by the addition of derivative (rate) action. Because differentiators are sensitive to noise, however, too much derivative action can result in twitchy loops.

Setting all these gains is not straightforward. If you are interested in loop tuning, I would refer you to the landmark paper by Ziegler and Nichols.[13] Reading this will underscore the wisdom of using the simplest control scheme that will produce satisfactory results for a given process.

An RF Power Meter

Now that we have produced a stable, low distortion sinusoid, the next step is to make its amplitude adjustable, and to immunize it against the effects of load changes. This is accomplished by an automatic RF power level controller, comprised of an RF power meter and a variable-gain RF amplifier. Its control loop is diagrammed in Figure 33. As I mentioned in Part 2 of this series, a control loop is no better than its sensor, so I will first highlight the power meter.

Figure 34 shows the RF power meter schematic, and Figure 35 is a list of components. This module uses the familiar Analog Devices AD8307 logarithmic amplifier, but with a couple of twists. You may be surprised that I power this component with 6 V instead of the usual 5 V. I wondered why the second paragraph of the data sheet gives the upper input power limit as +17 dBm, while the very next paragraph says +16 dBm. It also says that the peak-to-peak input voltage may not exceed the supply voltage, which is specified from 2.7 V to 5.5 V, with an absolute maximum rating of 7.5 V. I ran some experiments on a few AD8307s, measuring output voltage versus input power at several supply voltages; the results are graphed in Figure 36. Note that the upper input power limit appears to be a function of V_S. I submitted these results to the applications engineers at Analog Devices, who were unwilling (perhaps understandably) to comment on my discovery. Nevertheless, given the V_S limit of 7.5 V, I felt confident about operating the component from 6 V and extending the maximum input power to almost 20 dBm.[14]

Another trick is the scaling of the AD8307 output for direct digital indication without using active circuitry. Rather than multiplying the output voltage with an operational amplifier, I used a digital panel meter with a full-scale input of ±200 mV, moved the decimal point one digit to the right, and *divided* the output voltage by 2.5, resulting in a direct readout in dBm. Considering the relatively high output impedance of the AD8307, at 12.5 kΩ, I decided to use a resistive voltage divider with an input impedance on the order

[13]Notes appear on page 7.

RF Power Meter Component List

R1	52.3 Ω	1/4W, 1% Metal Film
R2	2.7 K	1/4W, 5% Carbon Film
R3	1 K	Trimpot
R4	1.5 K	1/4W, 5% Carbon Film
R5	348 K	1/4W, 1% Metal Film
R6	232 K	1/4W, 1% Metal Film
C1	1000 pF	Ceramic
C2	1000 pF	Ceramic
C3	0.1 µF	Ceramic
C4	0.1 µF	Ceramic
C5	0.1 µF	Ceramic
C6	0.1 µF	Ceramic
C7	1000 pF	Feedthrough
C8	1000 pF	Feedthrough
U1	AD8307	Logarithmic Amplifier (Analog Devices)
U2	78L06	6 V @ 100 mA Regulator
M1	V125	± 200 mV Digital Panel Meter (Martel or equivalent)

Pomona Model 3754 Miniature-size Cast Aluminum Enclosure
Pomona Model 2400 A-size Cast Aluminum Enclosure
Phoenix MPT 0,5/ 4-2,25 4-position Terminal Block

Misc.: BNC Connector, SIP Connector, Hardware

Figure 35 — The RF power meter parts list.

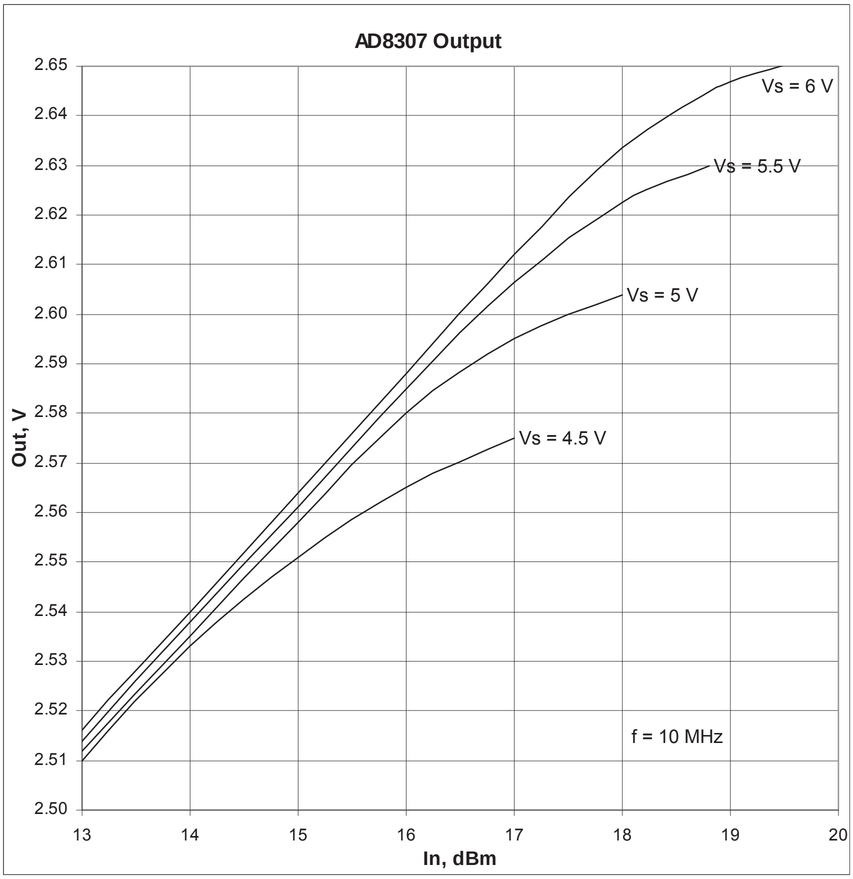

Figure 36 — This graph shows the results of my experiments with a sampling of AD8307 logarithmic amplifier ICs. Higher supply voltages resulted in an extended output range.

of 1 MΩ. Trying to avoid potentiometers whenever possible, I looked at the EIA E96 1% resistor standard to find combinations in a 1.5:1 ratio (for voltage division by 2.5), and chose 348 kΩ and 232 kΩ. But then I asked myself the question: what other exact division ratios are possible using 1% resistor values? After making nearly a hundred spreadsheets, I produced Table 1, exhibited in the Appendix. Of course the resistor pairs may be scaled by decades, such as 34.8 kΩ and 23.2 kΩ, or 348 Ω and 232 Ω, depending upon the input impedance desired. I hope that this hard-won table will be of benefit to you some day.

Figure 37 shows the RF power meter components before assembly, and Figure 38 depicts the completed module. The digital panel meter has an adjustment to trim the span (slope); the trimpot is accessible via a hole in the perfboard, but I found this adjustment to be unnecessary. An offset (intercept) adjustment was required, however, and this is accomplished with R3; you may wish to use the calibrated –10 dBm output from your new OXCO to make this adjustment.

The analog output of the RF power meter will be used to provide the process variable input to the variable-gain RF amplifier described in Part 4 of the series.

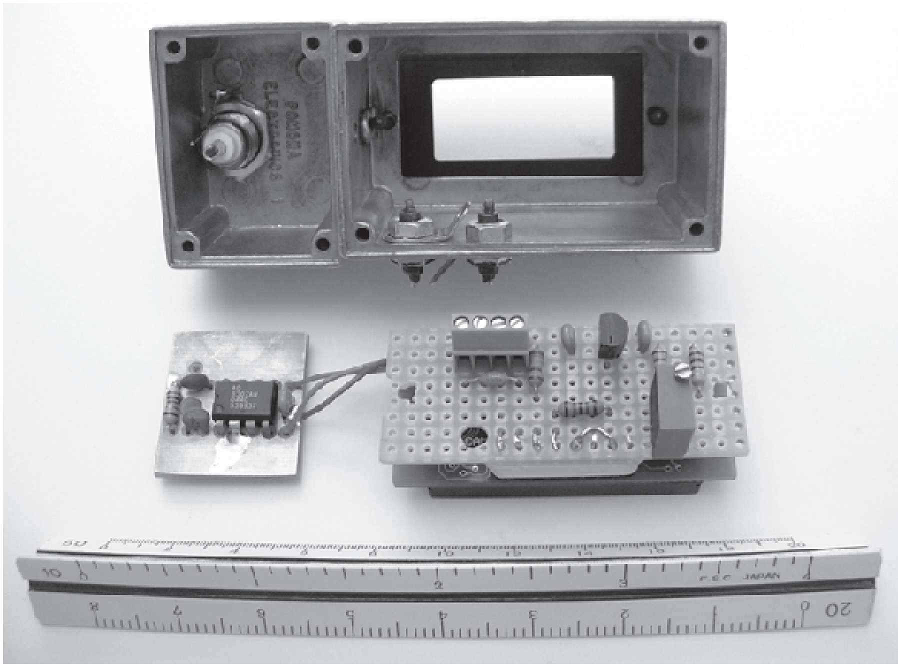

Figure 37 — This photo shows the AD8307 amplifier board and the rest of the RF power meter circuitry ready to be installed in the Pomona aluminum project cases.

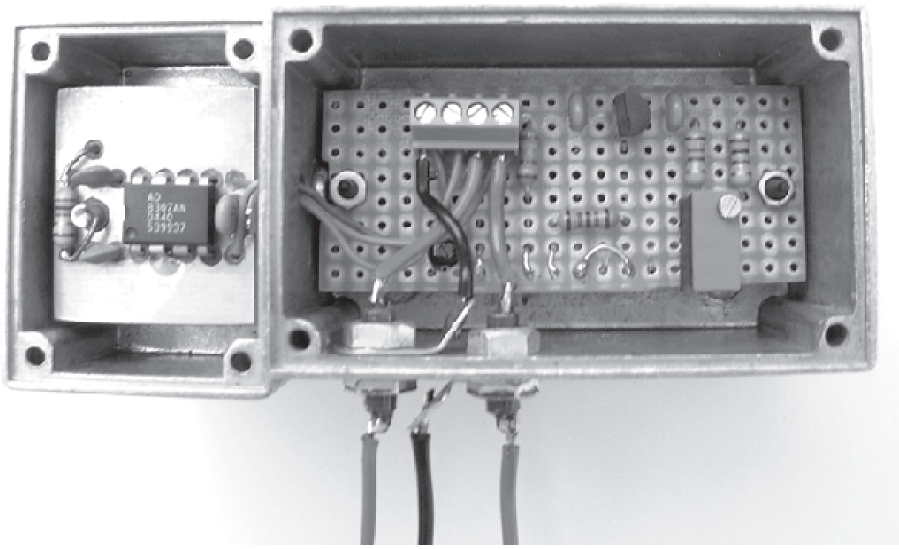

Figure 38 — The RF power meter circuitry is installed in the project cases. Notice that the leads from the amplifier board go through matching holes in the project case sides and then connect to the rest of the circuit using a four-connector terminal strip.

Appendix

The following resistive voltage divider example refers to Figure 39.

Desired: $V_I = 5$ V, $V_O = 1$ V, ≈ 10 kΩ input impedance.

5 V / 1 V = > ÷ 5, so $R_A / R_B = 4:1$.

A logical choice for R_A would be 10 kΩ, but this requires 2.5 kΩ for R_B, and the nearest 1% resistor value is 2.49 kΩ.

Consulting Table 1 (on the next page):

4:1 (÷5)

102 25.5

so $R_A = 10.2$ kΩ and $R_B = 2.55$ kΩ, yielding a voltage division of exactly 5 (within the resistor tolerances, of course).

Notes

[13]John G. Ziegler and Nathaniel B. Nichols, "Optimum Settings for Automatic Controllers," *Transactions of the American Society of Mechanical Engineers*, Vol 64, 1942, New York.

[14]Using a shortcut from Part 2, Note 11, dBm = 10 log (2.5 (6 V)²) = 10 log (90) = 19.54 dBm.

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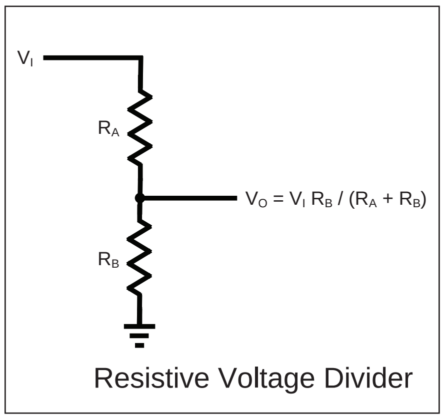

Figure 39 — This is a resistive voltage divider, as used in the RF power meter to scale the digital panel meter display so it will read power directly in dBm.

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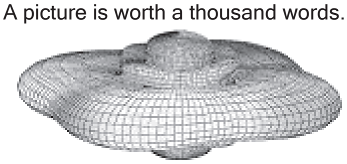

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Table 1
Exact Ratios Using Standard 1% EIA E96 Resistor Values

1.5:1 (for ÷2.5)

15	10
16.5	11
21	14
23.7	15.8
24.3	16.2
26.1	17.4
26.7	17.8
29.4	19.6
34.8	23.2

2:1 (÷3)

20	10
21	10.5
22.6	11.3
27.4	13.7
28	14
29.4	14.7
31.6	15.8
32.4	16.2
34.8	17.4
37.4	18.7
39.2	19.6
44.2	22.1
46.4	23.2
118	59
133	66.5
143	71.5
150	75
165	82.5
169	84.5

3:1 (÷4)

102	34

4:1 (÷5)

102	25.5

5:1 (÷6)

59	11.8
66.5	13.3
71.5	14.3
75	15
82.5	16.5
84.5	16.9
100	20
105	21
113	22.6
137	27.4
140	28
147	29.4
158	31.6
162	32.4
174	34.8
187	37.4
196	39.2
221	44.2
232	46.4

For 6:1 See the “Not Exact But Useful” section at the end of the Table.

7:1 (÷8)

93.1	13.3
105	15
140	20
147	21
196	28

For 8:1 and 9:1 See the “Not Exact But Useful” section at the end of the Table.

10:1 (÷11)

100	10

And so on

11:1 (÷12)

110	10
121	11
143	13
154	14
165	15
374	34
649	59
825	75

13:1 (÷14)

130	10
143	11
169	13
182	14
442	34

14:1 (÷15)

140	10
147	10.5
154	11
182	13
196	14
210	15
280	20
287	20.5
294	21
301	21.5
357	25.5
392	28
511	36.5
665	47.5
931	66.5
1050	75

15:1 (÷16)

150	10

And so on (See 1.5:1)

17:1 (÷18)

187	11
221	13
255	14
340	20
357	21

20:1 (÷21)

200	10

And so on (See 2:1)

21:1 (÷22)

210	10
294	14

22:1 (÷23)

1650	75

25:1 (÷26)

255	10.2

28:1 (÷29)

280	10
294	10.5
392	14
1330	47.5
2100	75

34:4 (÷35)

340	10
357	10.5
374	11
442	13
2550	75

35:1 (÷36)

357	10.2

51:1 (÷52)

1020	20

55:1 (÷56)

649	11.8
715	13
825	15
1100	20
1540	28
1870	34

59:1 (÷60)

590	10
649	11
1180	20

62:1 (÷63)

806	13
1240	20

65:1 (÷66)

715	11
806	12.4
845	13
1300	20
1820	28
2210	34

68:1 (÷69)

1020	15

75:1 (÷76)

750	10
825	11
1050	14
1470	19.6
1500	20
1740	23.2
2100	28
2370	31.6
2430	32.4
2490	33.2
2550	34
2610	34.8
2940	39.2
3090	41.2
3240	43.2
3480	46.4
4020	53.6
4320	57.6
4530	60.4
5490	73.2
5760	76.8
7320	97.6

77:1 (÷78)

1540	20

79:1 (÷80)

1580	20

81:1 (÷82)

1620	20

87:1 (÷88)

1740	20

89:1 (÷90)

1780	20

91:1 (÷92)

1820	20

95:1 (÷96)

1330	14

Not Exact but Useful:

6:011 (÷7)

107	17.8

8:021 (÷9)

150	18.7

9.027:1 (÷10)

102	11.3

19:027 (÷20)

215	11.3

QEX

John A. Magliacane, KD2BD and **Bill Walker, W5GFE**

1320 Willow Drive, Sea Girt, NJ 08750; kd2bd@amsat.org; PO Box 696, Ada, OK 74820; bw@cs6.ecok.edu

Splat!: An RF Signal Propagation, Loss and Terrain Analysis Tool

Analyze the coverage area of your local repeater or the propagation path between two stations with this multi-platform analysis tool.

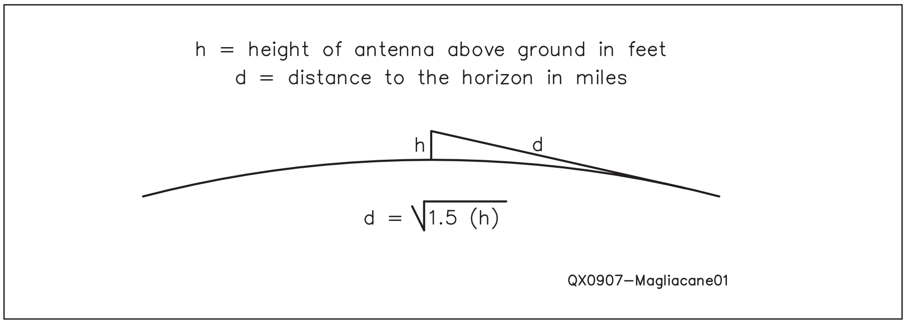

Figure 1 — Simple line of sight model over a smooth Earth.

SPLAT! is a powerful terrestrial signal propagation and terrain analysis tool created by KD2BD for the spectrum between 20 MHz and 20 GHz. It has many useful applications in the field of wireless electronics.

SPLAT! can accurately determine minimum antenna height requirements between transmitters and receivers to establish line-of-sight paths and Fresnel Zone clearances. *SPLAT!* can estimate path loss, field strength and received signal levels based on the Longley-Rice Irregular Terrain Model. *SPLAT!* can produce signal contour maps suitable for visualizing repeater system coverage areas, assist in designing point-to-point control links, wireless Wide Area Networks (WANs) and help perform frequency coordination and interference studies. Its capabilities and performance rival those of commercial software packages costing tens of thousands of dollars.

SPLAT! is free software designed primarily for operation on *Unix* and *Linux*-based workstations. Alternate versions for other operating systems are also available.

Recently, Dr. Bill Walker, W5GFE, has created a Web based interface allowing many of the *SPLAT!* functions to be executed through any Internet connected platform that supports a graphical Web browser.

SPLAT! History

SPLAT! was originally created in 1997 to estimate the line-of-sight operational coverage of the N2SMT/R 70 cm ATV repeater at Brookdale Community College in Lincroft, New Jersey. The program was rewritten several times before finally being released to the public via the Internet under the GNU General Public License in April 2002.

About a year after its release, Dr J. Douglas McDonald, ex-WA5HDC, integrated point-to-point Longley-Rice Irregular Terrain Model code into *SPLAT!*, which enabled accurate estimates of RF path loss to be made. Doug is co-author of several patents relating to adaptive equalization and echo cancellation techniques employed in digital television (DTV) reception, and used *SPLAT!* in some of his DTV reception studies.

In early 2005, Doug Lung, AH6DL, Vice President for Engineering at Telemundo/NBC, wrote an article for *TVTechnology* magazine discussing the use of *SPLAT!* in illustrating the coverage area of several DTV transmitting facilities. Follow-up correspondence with Doug provided valuable suggestions and technical guidance for adding support for high gain directional antenna systems employing both electrical and mechanical beam tilt used in the television broadcast industry. Doug was also able to provide "sanity checks" during code development by verifying results obtained through *SPLAT!* with those of actual ground-based measurements and of commercial software packages of similar function.

The use of *SPLAT!* reached beyond the planet in 2006 when Vaughn Cable, K6ZTA, at NASA's Jet Propulsion Laboratory in Pasadena, California obtained a copy of *SPLAT!* with the intention of modifying it for use in developing path loss and fade margin policies for communication links of interest to NASA in preparation for future manned lunar and Martian surface operations and exploration.

How It Works

Communication paths on VHF and higher frequencies take place through a variety of mechanisms such as line-of-sight propagation, diffraction over the horizon and the Earth's terrain, atmospheric refraction and various forms of scatter. The easiest of these to consider is a line-of-sight path.

Figure 1 illustrates how the distance of a line-of-sight propagation path between an antenna and the horizon can be determined. The associated equation actually takes atmospheric refraction into account, which extends the radio horizon to a distance slightly beyond that of the optical horizon.

When the height of the second antenna in

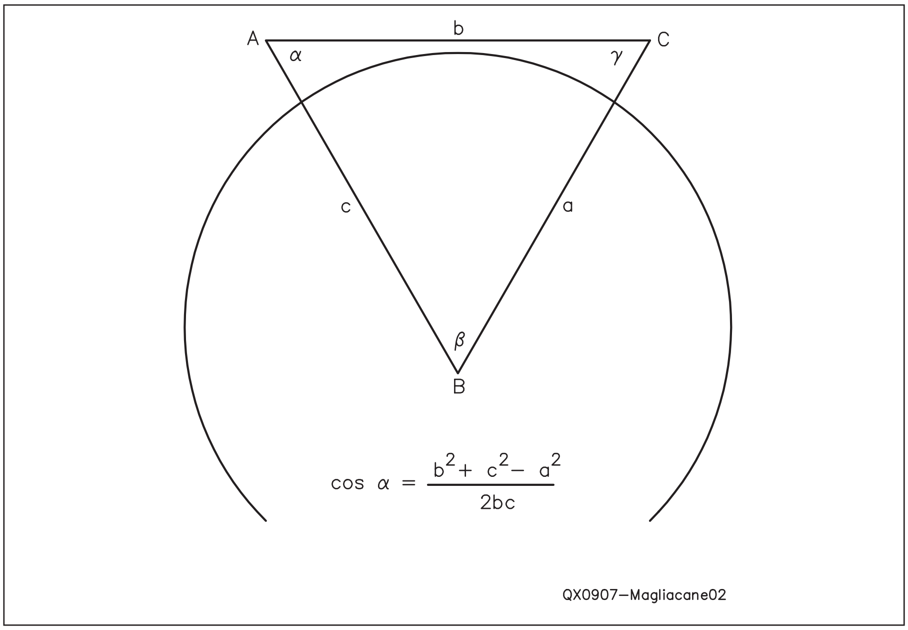

Figure 2 — Simple line of sight model with both antennas above the Earth's surface.

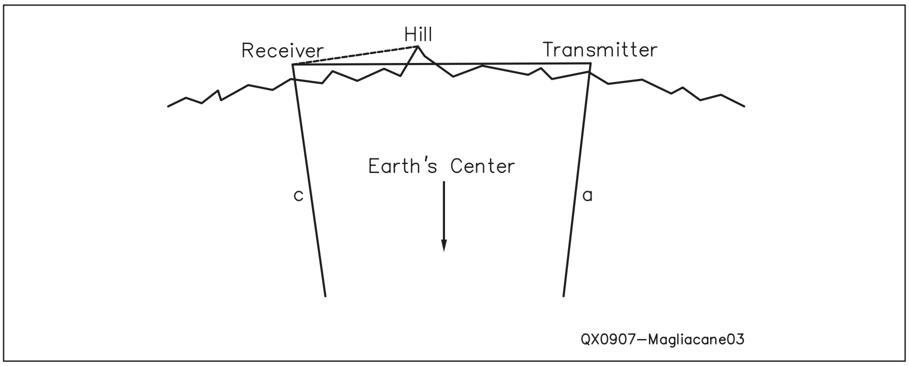

Figure 3 — Simple line of sight model over a rough Earth.

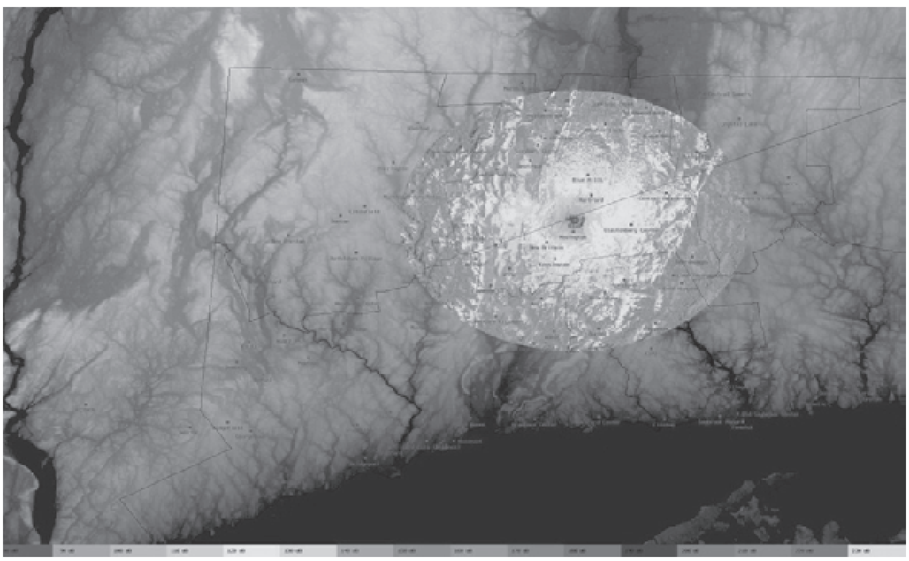

Figure 4 — A Longley-Rice Terrain Shielding Plot, 146 MHz, centered on W1AW (cropped).

the communications link extends to a known distance above the Earth's surface, the maximum length of the propagation path can be determined through simple trigonometry. Figure 2 illustrates how the Law of Cosines can be used to solve for all angles in the triangle formed between Antenna A, Antenna B and the Earth's center.

A line-of-sight path exists if the angle from Antenna A to Antenna B is greater than that of the angle between Antenna A (or B) and that of the Earth below. Increasing the Earth's radius by 33 percent over its actual amount effectively compensates for the effects of atmospheric refraction on the propagation of the signal between the two points.

Real life, however, is much more complicated. Unless we limit ourselves to communication links between isolated ships at sea, there's terrain with which we must contend. Having an accurate model of the Earth's terrain permits a determination as to whether a terrain feature will cast a shadow between the two points in question. As before, line-of-sight determination becomes a trigonometry exercise, and every terrain feature along the path between transmitter and receiver must be taken into account to qualify the path. See Figure 3.

Is Line-of-Sight Enough?

Whether a line-of-sight path is required for reliable communications depends on many variables. High power commercial VHF and UHF broadcasters will claim that a line-of-sight path to their transmitting antenna is not a necessity for adequate reception of their signals. On the other hand, a WiFi operator using low-powered unlicensed transceivers will claim that a line-of-sight path is in and of itself, insufficient information for ensuring reliable operation of their equipment. Both claims are correct. A terrestrial line-of-sight condition, in fact, provides us very little detail regarding the actual RF attenuation experienced over that path, or its long-term reliability.

The Longley-Rice Propagation Model

In the 1960s, Anita Longley and Phil Rice developed a general purpose propagation model based on electromagnetic theory and statistical analysis capable of predicting the median path loss of terrestrial radio signals propagating across irregular terrain. Their model has become the FCC's defacto standard in predicting path loss for the spectrum between 20 MHz and 20 GHz, and is heavily used in the commercial broadcast industry for predicting service contours, and performing interference studies.

Longley-Rice propagation modeling is used in Amateur Radio applications as well.

The April 2008 issue of *QST* carried an article by Dan Henderson, N1ND, describing the efforts by ARRL and concerned repeater owners to resolve a problem that has arisen concerning shared usage of the 420 MHz band by amateurs and large (and classified) USAF PAVEPAWS radars operating within close proximity to each other.

The USAF must have, and is lawfully entitled to, an interference-free environment in which to operate PAVEPAWS radars. At the same time, amateur repeater owners are interested in sharing the 420 MHz band with these radar systems when it is technically possible to do so.

Dan Henderson's article mentioned that one of the tools employed by ARRL in an attempt to resolve the conflict was "Longley-Rice terrain shielding calculations."

Longley-Rice Terrain Shielding Calculations

The FCC has long recognized Longley-Rice terrain shielding calculations as acceptable evidence when considering waivers for geographic restrictions in granting commercial broadcasting licenses.[1]

Figure 4 shows the results of a 146 MHz Longley-Rice terrain shielding calculation centered on W1AW, assuming a target antenna height of 5 feet (typical for a mobile antenna).

Until recently, such calculations have, of necessity, been the province of large mainframe computers, with substantial engineering talent available for interpreting the results. "*SPLAT!*" makes the analysis of terrain, and the determination of Longley-Rice model path-loss available to the average Amateur Radio operator.

SPLAT! has two drawbacks for the average user. The first is the fairly large number of "command line" parameters that are necessary to control the program, and the second is the rather large data sets, which describe geographic terrain data to the program so that accurate topographic maps may be created. (We comment, though, that the first "drawback" is precisely the reason why it was easy to create a Web interface to the program.)

The terrain data itself, a product of the STS-99 Space Shuttle Radar Topography Mission, is available in the public domain from several sources, and covers a large portion of the Earth. In addition, data that provides geographic locations for cities and political boundaries in the United States is available through the United States Bureau of the Census. The volume of data is intimidating, and depending on their needs, not everyone will be interested in downloading it, much less storing it.

[1]Notes appear on page 14.

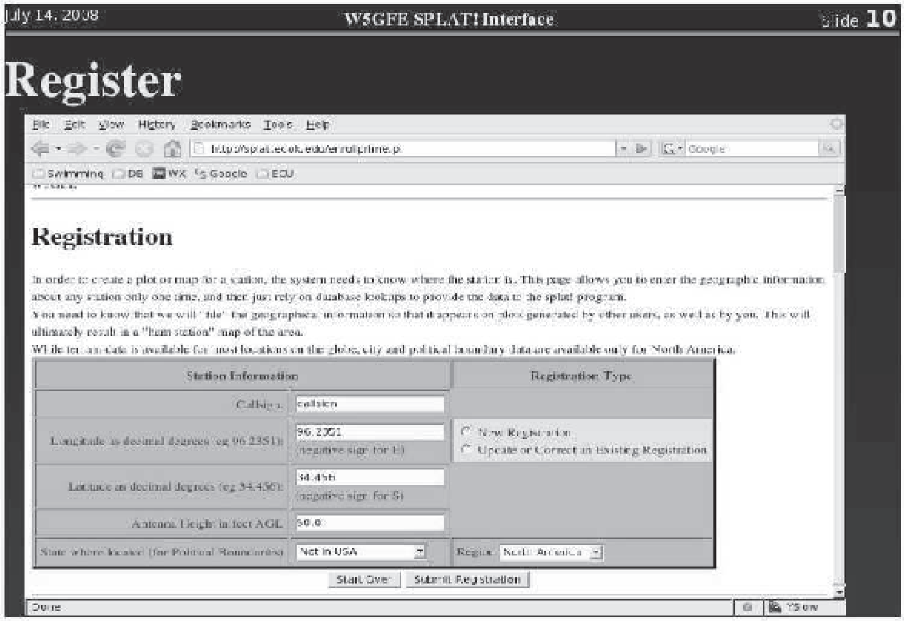

Figure 5 — This screen shot is the Registration Page from the *SPLAT!* Web site.

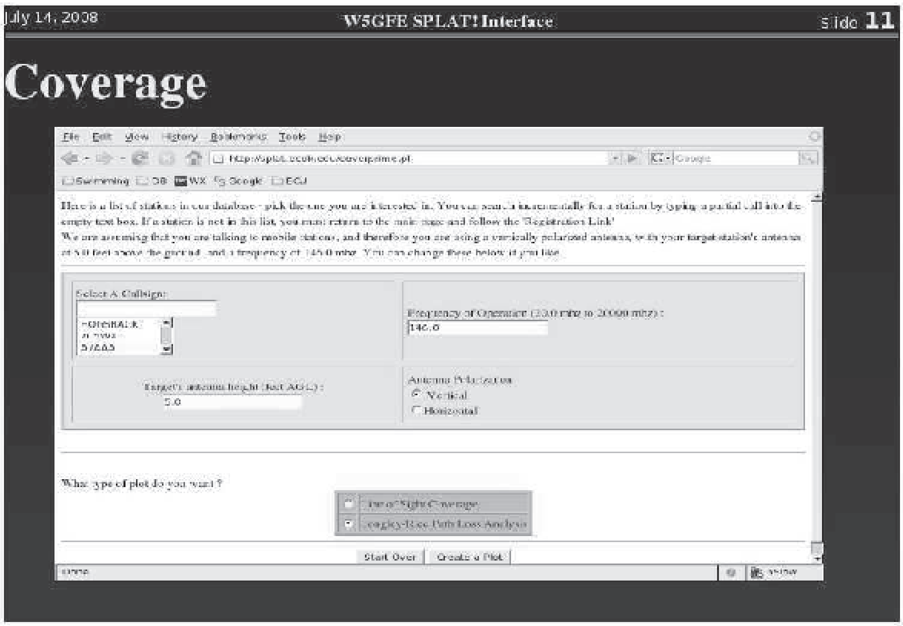

Figure 6 — Here is a screen shot of the Coverage Page from the Web site.

The *SPLAT!* Web Interface

Having access to a good-sized LAMP (*Linux*, *Apache*, *MySQL*, *Perl*) system, W5GFE created a general-use *SPLAT!* server capable of making the benefits of "*SPLAT!*" available to Amateur Radio operators worldwide.

W5GFE's system allows the big LAMP to serve as a repository for the *SPLAT!* data files.[2] It also provides the computational "heavy lifting" that *SPLAT!* demands, and presents the results to anyone with a Web browser via the Internet. The site is graciously hosted by the Computer Science Department at East Central University, in Ada, Oklahoma (see **http://cs.ecok.edu/**).

By following a series of Web menus, anyone can generate Longley-Rice plots, topographic maps, line-of-sight plots, profile plots, and multiple transmitter plots.

The resulting graphics are presented on a Web page that can be viewed at leisure, or saved for further manipulation. Note that some of the topographic maps are large, and may need a sophisticated image viewer with a "zoom" feature for detailed inspection. Most Web browsers will present the images at a size that is convenient for direct viewing.

How to Use the *SPLAT!* Web Interface

The Web interface does not require any special browser, nor does it require any particular computer or operating system. Users will want a fairly fast Internet connection to be able to download topographic maps in reasonable time, but profile plots and text reports should be accessible even at dial-up speeds.

Registration

The first step is to visit the site at **http://splat.ecok.edu/** with your browser.

Users should first visit the "Registration" section of the site. See Figure 5. You will need to know the latitude and longitude of your station, as well as the height of your antenna. This data is stored in a database on the LAMP server, enabling other stations to see the location of your station on maps they generate.

The geographic data provided by each registrant is entered in a database. Each successive *SPLAT!* computation consults that database when displaying geographic locations. As a result, once a certain station is "registered," the location of that station will be plotted on succeeding maps. If your friend N1XXX registers, and you later create a plot for your own N1YYY station, the map you generate will show the locations of both N1YYY and N1XXX.

Eventually, a "ham map" of a given area is generated. Be advised that other users of the site will be able to discover the general location of your station, and your data is stored permanently on the site. The "state" information is used to plot the various cities and political boundaries of your location directly on topographic maps. This information may be somewhat incomplete.

If you wish to explore the effects of using antennas at different heights, just register several different stations at the same location (for example, W5GFE-1, W5GFE-2, W5GFE-3, and so on). If you accidentally register incorrect information, you can correct it on this page by using the "update" button.

After you have registered your site, you are ready to create plots.

Coverage Plots

Follow the "Coverage Plots" link from the main page (see Figure 6). You will see a page with a pull-down list of registered stations, and some fill-in-the-blank options. The blank "Select a Call Sign" box can be used to incrementally search for a station that is already registered. Just begin typing the call sign in the box. This is a feature of all of the pages that generate plots.

The resultant plot (Figure 7) is useful for graphic depictions of coverage areas.

Profile Plots

Profile plots are not very large, and can probably be handled even by a slow Internet connection. The plots are "normalized," which in this case means that the stations on each end of the circuit appear at the same level on the graph, and the various paths are traced against these "normalized" locations.

As you can see in Figure 8, the opportunity to select two stations is provided, as is the ability to create signal profiles of the terrain between those two stations. Selecting the topographic map option will create a map containing the two stations, and a line of site path between them (if it exists).

Two plots created by the "profiles" page appear in Figures 9 and 10.

Multi Station Plots

It is interesting to plot combined coverage areas of linked repeater sites by using the "Multiple Station Plots" link. The pull-

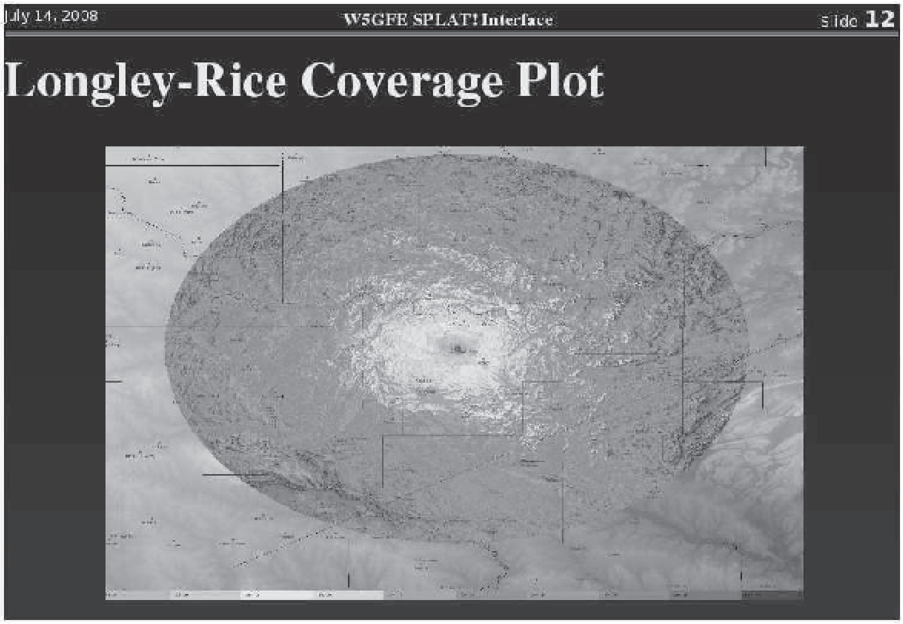

Figure 7 — This Coverage Graph shows the results after entering the station information at the Coverage Page.

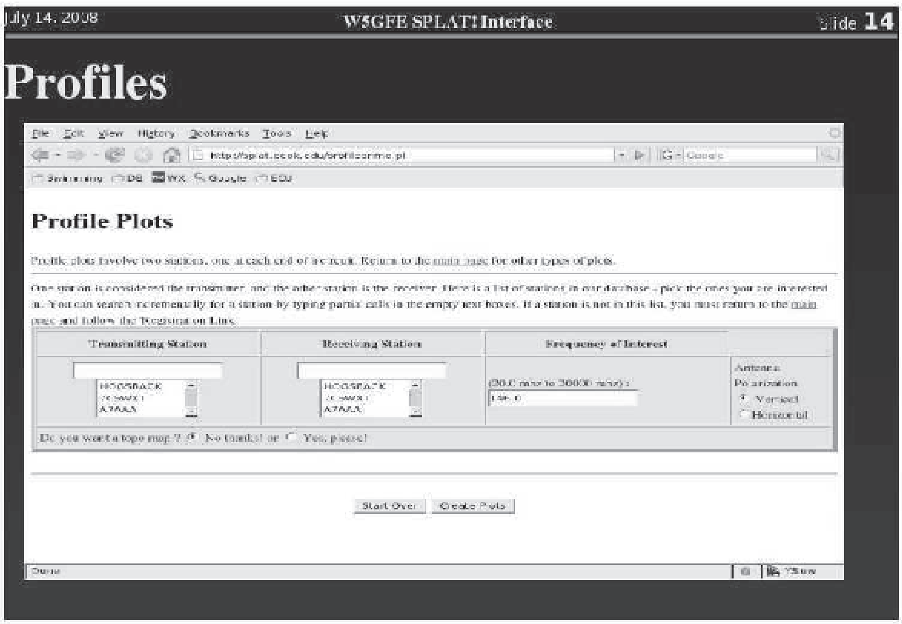

Figure 8 — Use the Profiles Page to create a profile of the path between two stations.

down menus of Figure 11 allow selection of up to four transmitting sites simultaneously. Colors are used to delineate the coverage of each transmitter and to show the combined coverage area. See Figure 12. Interestingly, two different antennas on the same tower may be treated as though they are "multiple transmitters."

Installing Your Own

Even though the command line version of *SPLAT!* is easy enough to use, the Web interface is sufficiently convenient that users may want to install it on their own equipment, especially if they are plotting a large number of stations, and want successive maps to include previously plotted stations.

The computational demands of the Web interface are not trivial, though it is certainly possible for a club to host the interface on a local Web site. The complete installation, including world wide maps, comprises about 17 GB of disk space. A "North America only" installation would run about 1.5 GB, and a "local area only" installation might take less than 250 MB.

The ability to execute CGI programs under a Web server will be required, as will the administrative authority over a database under *MYSQL*. Users will need to be able to offer "write permissions" for the Web server on at least one directory, and will need to be able to compile the *SPLAT!* source code, or download the binaries from one of the many on line repositories.

Some of the larger plots may require more time than your Web server can provide to a single connection. There is a setting in the Apache configuration that can adjust this "timeout" parameter. Documentation is available in the "HOWTO" file.

This list of requirements sounds more daunting than it actually is. Most *Linux* distributions come with all of the required applications pre-installed. The *SPLAT!* source code and the Web interface, with a "HOWTO" document is available for downloading through the W5GFE Web site at East Central University, Ada, Oklahoma: **http://splat.ecok.edu/**.

Acknowledgments and Disclaimer

Stan Horzepa, WA1LOU, kindly publicized the W5GFE *SPLAT!* Web site in his excellent on-line "Surfin'" blog on March 7, 2008. The site scored a "virtual WAS" over the subsequent weekend. You can still find that blog by going to the ARRL Web site (**www.arrl.org**) and typing SPLAT in the "Search Site" window.

Walter K1CMF pointed out an error that prevented the original Web site version (though not *SPLAT!* itself) from functioning correctly on frequencies other than 146.0 MHz. Walter also confirmed that the corrections, when made, allowed agreement with other relevant materials in his possession. Walter is directly involved with the effort to avoid interference to the PAVEPAWS system on Cape Cod, Massachusetts.

The definitive Web site for *SPLAT!* belongs to John, KD2BD, and can be found at **www.qsl.net/kd2bd/splat.html**.

John's program is a wonderful creation, and a major contribution to ham radio. Bill's Web site could not exist without John's *SPLAT!*

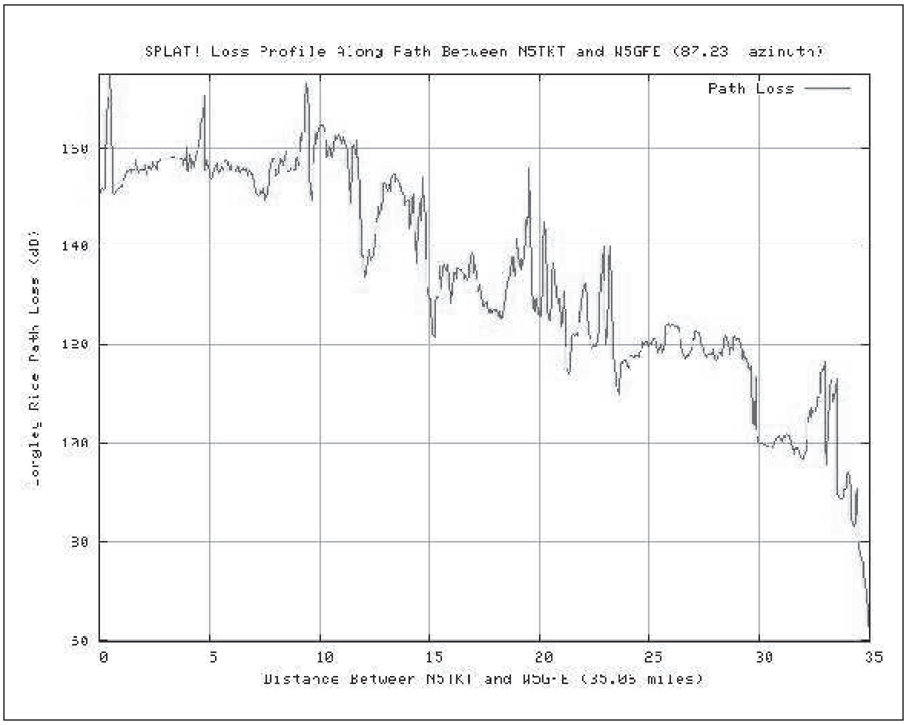

Figure 9 — This graph shows the Path loss (in dB) between two stations.

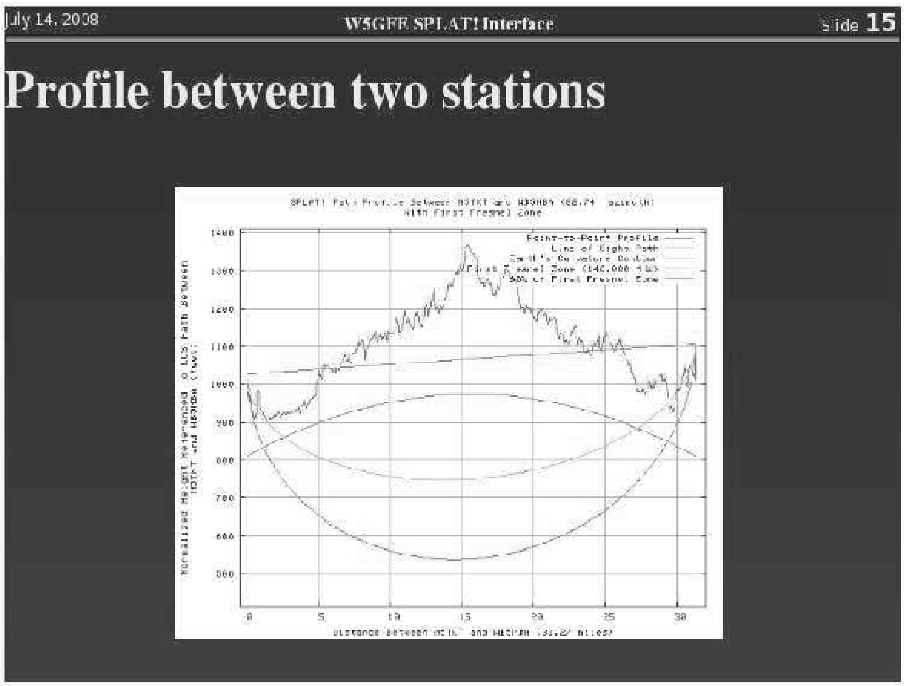

Figure 10 — Here is a screen shot of the program results for "Normalized Terrain and Path Between Two Stations."

Notes

[1]FCC Office of Engineering and Technology Bulletin No. 69, "Longley-Rice Methodology for Evaluating TV Coverage and Interference," and others, **www.fcc.gov/Bureaus/Mass Media/Public Notices/LPTV Notices/pnmm0039.doc**.

[2]Quad dual-core Opteron 64 bit processors, with 32 GB of main memory, and 1.5 TB of disk storage, running *RedHat Linux*.

John Magliacane, KD2BD, has held an Advanced class license for over 25 years, and a Commercial FCC Radio License since 1994. John holds Associate Degrees in Electronics, Computer Science and Mathematics as well as a Bachelor's Degree in Electronics Engineering Technology.

Since 1987, John has been employed as a Learning Assistant for electronics courses offered at Brookdale Community College, Lincroft, NJ and has served as advisor for the College's Amateur Radio Club since 1991. John has also worked as a freelance technical writer for about 15 years. His weekly "Space News" newsletters enjoyed great popularity across the packet radio networks during the 1990s.

John prefers the technical aspects of Amateur Radio, having engineered BPSK and FSK modems for Pacsat satellite communications, and Linux*-based satellite tracking and related communications software. In recent years, John has been active in frequency measuring tests using receiving apparatus, frequency measuring methods and instrumentation of his own design.*

Bill Walker, W5GFE, was first licensed as a "Conditional Class" in 1961, at the age of 14, and has held the same call sign since then. He has upgraded to Amateur Extra class. He is an ARRL Life Member.

Dr Walker received a BS in mathematics from West Texas A&M in 1968, and MS (1970) and PhD (1974) degrees in mathematics from Texas Tech. He currently is appointed Adolph Linscheid Distinguished Teaching Professor and is Chair of Computer Science at East Central University in Ada, Oklahoma. He was founding chair of the Oklahoma Computing Consortium.

Dr Walker has published in RTTY Journal, 73 Magazine, QST, QEX, Byte *and professional journals. He has authored three textbooks (all in the computing field) and presented many papers at professional meetings.*

He is a 32nd degree Mason, and although not a Native American, belongs to the Comanche Little Ponies, the Comanche War Dance Organization, and the Intertribal Singers Drum. He also serves on the editorial board for Whispering Winds magazine.

Bill has been married 29 years (to the same woman!) and lives in rural Oklahoma with his wife Anita, their son Dalton and spotted dog Leia.

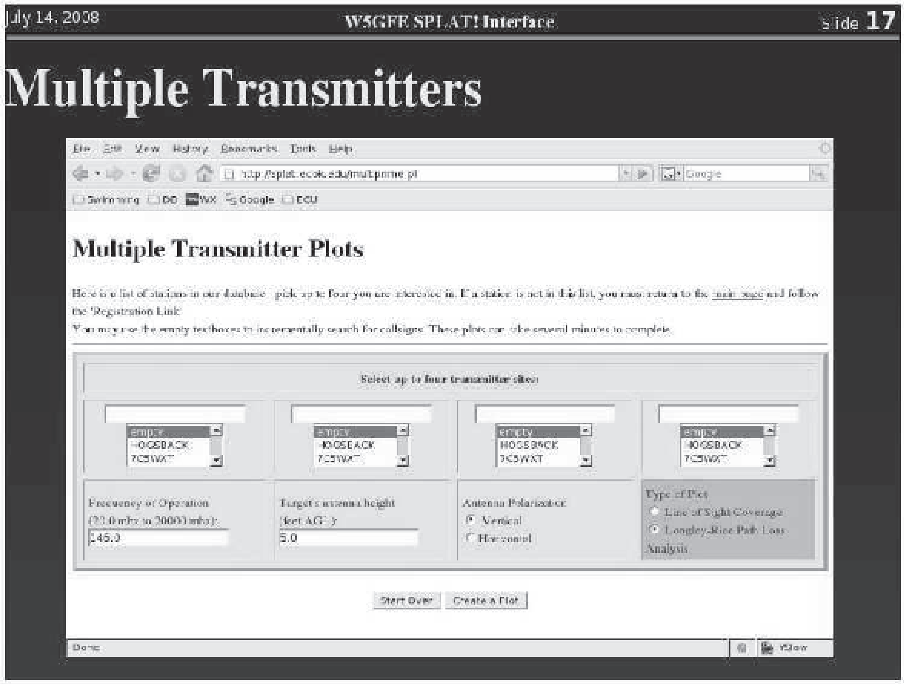

Figure 11 — You can analyze the coverage for multiple transmitters.

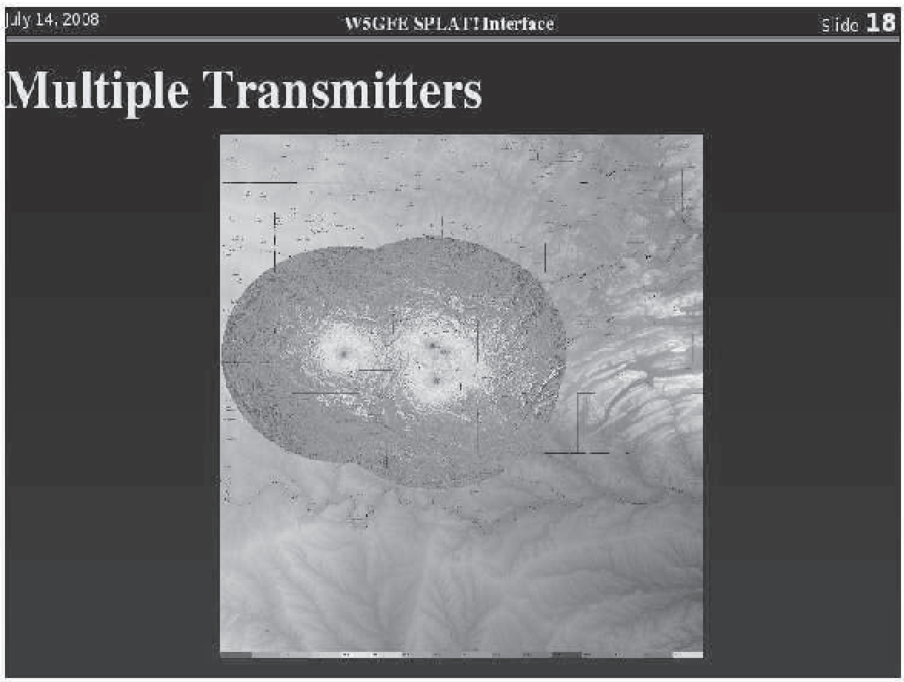

Figure 12 — Here is an example of the output plot created when you enter data for multiple transmitters.

QEX

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Experimental Determination of Ground System Performance for HF Verticals Part 5 160 Meter Vertical Ground System

How much will the signal strength and feed point impedance change as radials are added?

This experiment was actually the first of the series of experiments on ground systems that have been the subject of this series of articles. The experiment involved measuring the change in signal strength as radials are added to the ground system of a vertical antenna, beginning with four radials and going up to 64 radials. The intent was to determine the additional gain in signal for each doubling of radial number, and to determine the point of vanishing returns. In addition, the changes in feed point impedance due to changing radial number were of interest.

While the results of this initial experiment were quite interesting, a more important result was an appreciation of the difficulties of making these measurements accurately. This experience led to a modification in the test procedure and a shift to 40 m verticals, which have been described earlier.

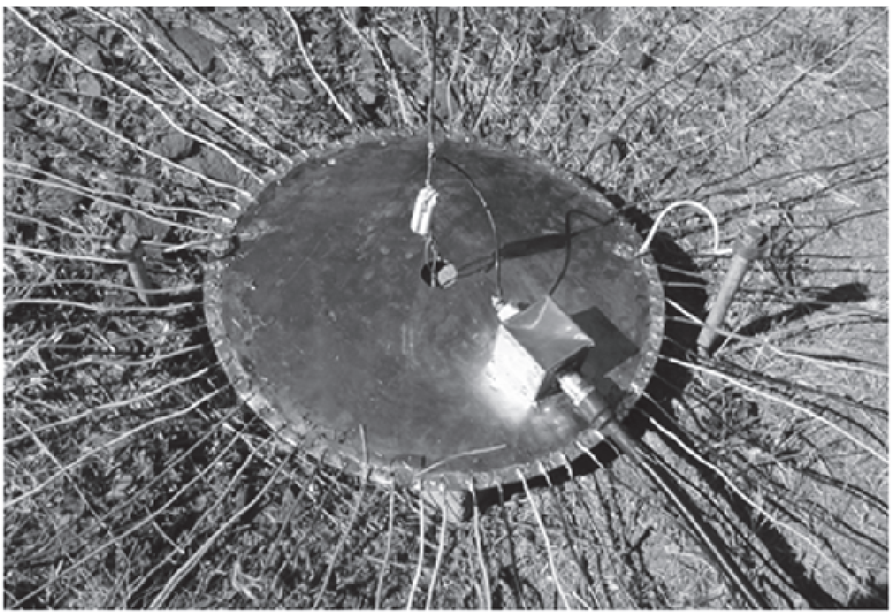

Figure 1 — This photo shows the antenna base with radials attached.

Test Antenna Description

The test frequency for this experiment was 1.800 to 2.000 MHz. The vertical was 125 feet of no. 12 AWG insulated copper wire suspended from a Dacron line hung between two 150 foot poles.

At the base of the antenna there was an 18 inch diameter copper disk, as shown in Figure 1. The inner ends of the radials and the shield of the coax feed line were attached to the disk. There were also two galvanized ⅝ inch × 4 foot ground stakes connected to the disk. The radials were 130 foot lengths of no. 12 insulated (THHW) wire lying on the ground surface. Radials were put down in the sequence of 4, 8, 16, 32 and 64.

The terrain around the antenna was not flat, but rather on a narrow ridge about 40 to 50 feet wide. The result is that many of the radials were in part bent down at about a 45° angle as they ran down the steep slope on either side. Along the ridge, however, the radials are more or less level.

The test antenna was erected 700 feet to the east of my house with a 50 foot deep gully in between. The ridge is in a Douglas fir forest with 100 plus foot trees within 50 feet of the test antenna at some points. The radial system ran along the ridge and also down the sides of the ridge into the forest.

To excite the test antenna, between the house and the antenna there was a 700 foot length of 1⅝ inch coax, with an additional 75 feet of ½ inch coax. Both were Andrews heliax.

Measurement Equipment

The signal source was a Yaesu FT1000MP transceiver with two Bird Model 43 wattmeters on the output (forward and reflected power). The wattmeters were used to set the forward power to a constant 50 W and also to measure reflected power to calculate SWR. The SWR measurement is needed to correct for the power reflected from the antenna and not radiated. This correction was applied to the received signal amplitude.

The receiving antenna was a 10 foot vertical wire driven against a 4 foot ground stake, next to my house. The receiver was an HP3585A spectrum analyzer. The amplitude resolution was about ± 0.1 dB.

Base impedance measurements were made at the antenna using an N2PK vector network analyzer (VNA). The impedance measurements were accurate to better than 1%.

The test procedure was very straightforward. For each number of radials, the FT1000MP output was adjusted to 50 W and received signal strength on the spectrum analyzer recorded along with the SWR for that measurement and the input impedance at the base of the antenna.

Test Results

Three complete runs were made to verify repeatability of the measurements. Each run included a complete stepping through the number of radials in the sequence, 4, 8, 16, 32 and 64. Typical received (and corrected for SWR) signal strengths versus radial number are given in Table 1. This data is graphed in Figure 2.

The data in Figure 2 has one obvious oddity. You would expect that the incremental difference as the radial numbers are doubled would be monotonically decreasing as the radial number rises. The step between 16 and 32 radials does not do this and it appears that the value for 16 radials is too small. This anomaly was noted during the experiment, however, and checked carefully as the radial count was redone three times. The anomaly was there in all three cases. I have no explanation for this other than the irregularity of

Table 1
Typical Test Data for Received Signal Strength with P_o = 50 W.

Number of Radials	*Corrected Signal Strength*	*Relative Signal Strength*
4	–30.1 dBm	0.0 dBm
8	–29.3 dBm	0.8 dBm
16	–28.9 dBm	1.2 dBm
32	–28.0 dBm	2.1 dBm
64	–27.7 dBm	2.4 dBm

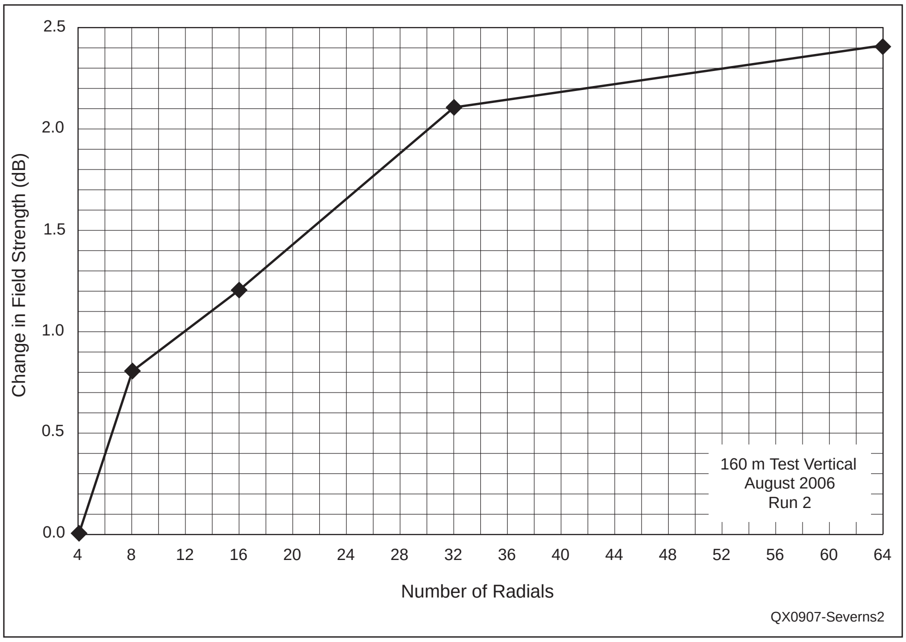

Figure 2 — Here is a graph of the typical signal strength change with radial number.

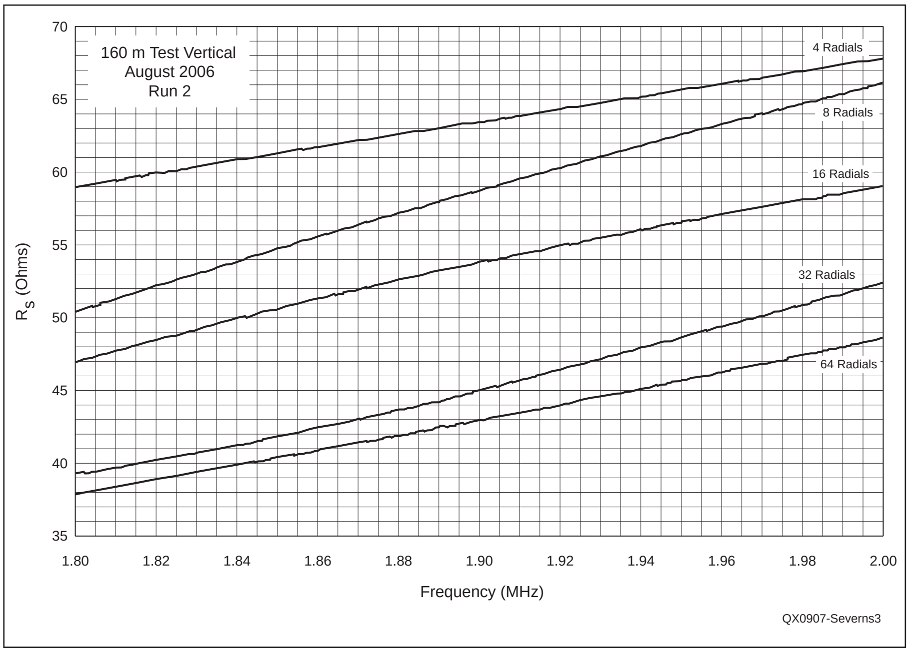

Figure 3— This graph gives the resistive part of the base impedance over the 160 m band for different radial numbers.

the site, which forced the radial layout to be far from flat or level. Later experiments with more regular radial systems on other antennas all showed the expected monotonic decrease in improvement with increasing radial number.

In any case, it's pretty clear that 32 radials do a good job and by 64 radials you are well into the region of vanishing returns. I certainly could not justify doubling the radial count to 128!

The results of feed-point impedance measurements are given in Figures 3, 4 and 5.

As discussed in Part 2 of this series, we would expect the resonant frequency to vary with the number of radials, due to the shift in radial resonance because of soil loading. The 40 m experimental work was done over an essentially flat pasture and the resonant frequency change was regular and monotonic. The gross irregularity of the ground surface in this earlier experiment, however, resulted in the erratic frequency changes shown in Figure 5. This problem was a primary reason for moving the experimental site from the narrow ridge to a pasture. Unfortunately, the 150 foot support poles were not available in the pasture so it was necessary to change the experimental frequency to 40 m to make the vertical height manageable.

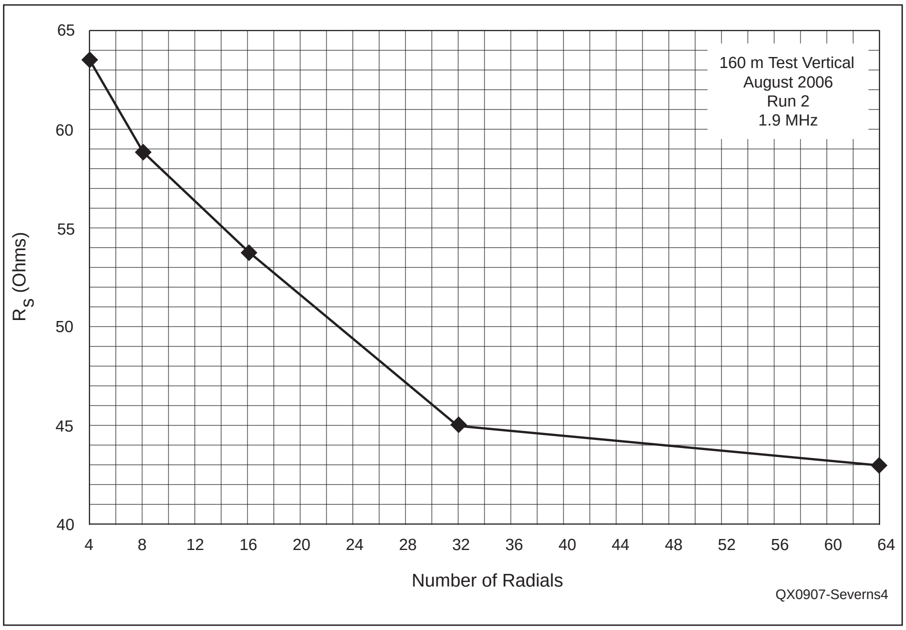

Figure 4 —This graph shows the base resistive component versus radial number at 1.9 MHz.

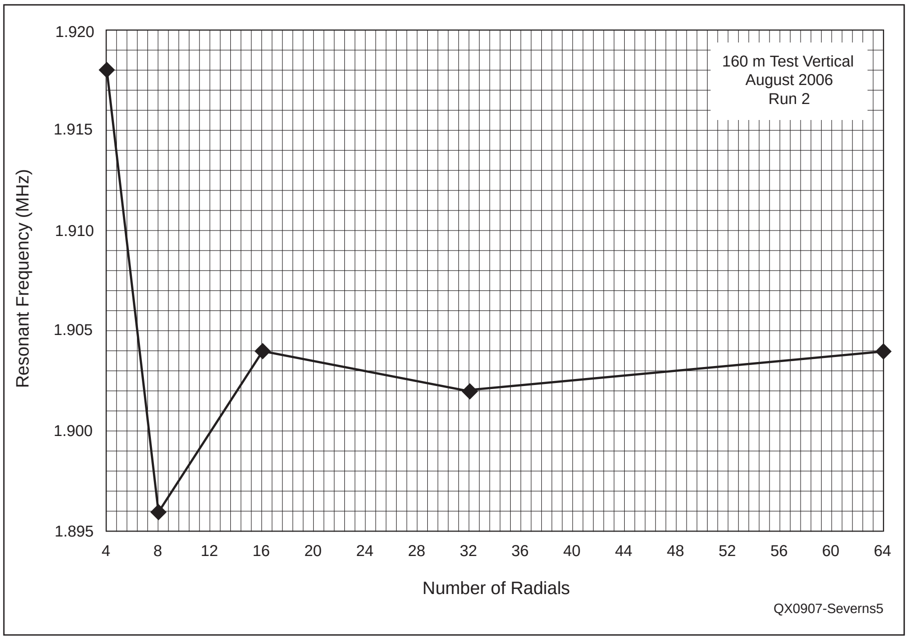

Figure 5 — This graph shows the antenna resonant frequency for different numbers of radials.

Summary

This initial experiment helped me to understand the problems inherent in making accurate comparisons between different ground systems. I had to change the site, the test frequency, the test instrumentation and the test methodology to get to the point where I could have confidence in the test results and draw conclusions from them.

This experiment was by no means a failure, however. We can see that the change in signal strength is very much in line with what we saw in the 40 m work. It also supports the conclusion that we should use at least 16 radials, but when we use more than 32 radials we are definitely reaching the point of vanishing returns. For most amateur installations the Standard Broadcast ground system of one hundred twenty 0.4-wavelength radials could not be justified by any useful increase in signal strength.

QEX

Rudy Severns, N6LF, was first licensed as WN7WAG in 1954 and has held an Extra class license since 1959. He is a consultant in the design of power electronics, magnetic components and power-conversion equipment. Rudy holds a BSE degree from the University of California at Los Angeles. He is the author of two books and over 80 technical papers. Rudy is an ARRL Member, and also an IEEE Fellow.

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The "Chicken Wire Wonder" — A Unique Broadband Vertical Antenna for the HF Bands

I'll bet you've never seen a Tapered Area Small Helix (TASH) antenna!

Many so called broadband "wonder" antennas advertised to the Amateur Radio community dissipate most of the applied power in resistive loading. Nevertheless, these "wonder antennas" do have a few benefits.[1] They produce very little TVI or other interference, since they radiate almost no power! In addition to these broadband "wonder" antennas, there are several truly efficient broadband antennas suitable for the Amateur Radio bands. These antennas include the bicone, discone, helix and others. Each of these broadband antennas has certain benefits and drawbacks.

Also, a few years ago a new efficient broadband antenna was featured in two trade journal articles.[2,3] The antenna was called a "Tapered Area Small Helix" or TASH for short. Apparently, the antenna was given that "catchy name" to help the article readers remember the antenna. TASH rhymed with trash. Although TASH sounded somewhat derogatory, it should have made the antenna memorable. Either the TASH name was really not that memorable or, at the time, there wasn't much interest in wideband antennas. In any case, the articles received little response from the antenna design or amateur community.

The TASH antenna performs similarly to a quarter wave vertical at most frequencies. A quarter wave vertical is one of the most popular HF antennas, because it provides low angle radiation superior to a beam or dipole that is not far above ground.[4] As described in the journal articles, the original TASH antenna consisted of a right triangle of conductive material rolled to form the TASH element, as shown in Figure 1. This first generation TASH antenna provided vertical performance and low SWR over only a single octave frequency range. More compact versions were later designed with better SWR bandwidth. Most of the improvements were obtained by decreasing the TASH element height-to-diameter ratio while reducing and reshaping the element area. Low SWR bandwidths of more than 10:1 have recently been demonstrated.

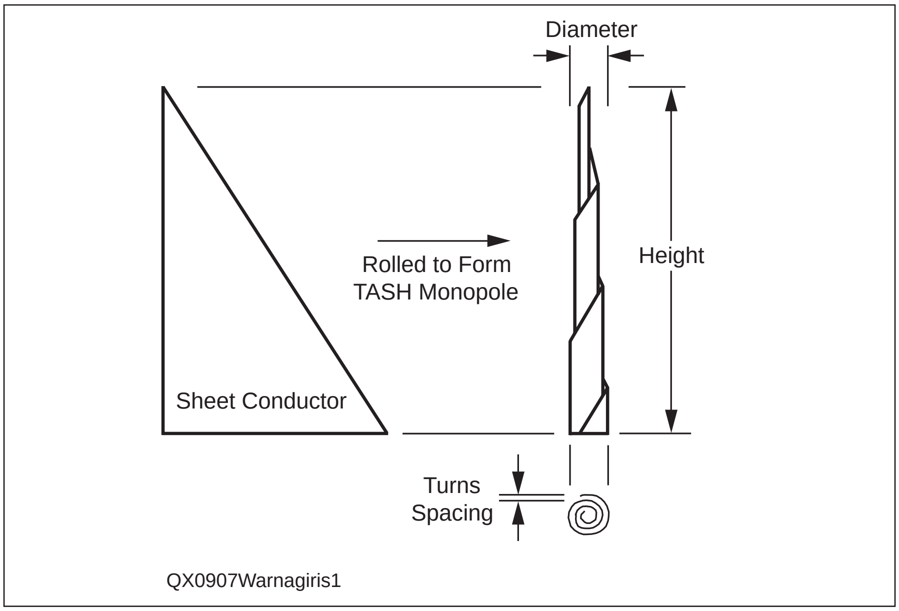

Figure 1 — Construction of the original TASH antenna element.

Figures 2A through 2D show several TASH variants, leading up to the most recent version. Figure 2A is one of the original 10:1 height-to-diameter-ratio designs with good SWR over a single octave. Figure 2B is a low profile variant with fair multi-octave SWR. Figure 2C is a wide spaced variant with good multi-octave SWR. Lastly, Figure 2D is a smaller element variant, also with good multi-octave SWR. Although these earlier versions provided wideband low SWR, they were not as compact as the most recent version, which successfully provides multi-octave SWR with a 3:1 height-to-diameter ratio, with only a one turn element.

Although a TASH antenna resembles a helix antenna it doesn't share its electrical properties. Both helix and TASH antennas have wide SWR bandwidths, but a helix antenna is a circularly polarized narrow

[1]Notes appear on page 24.

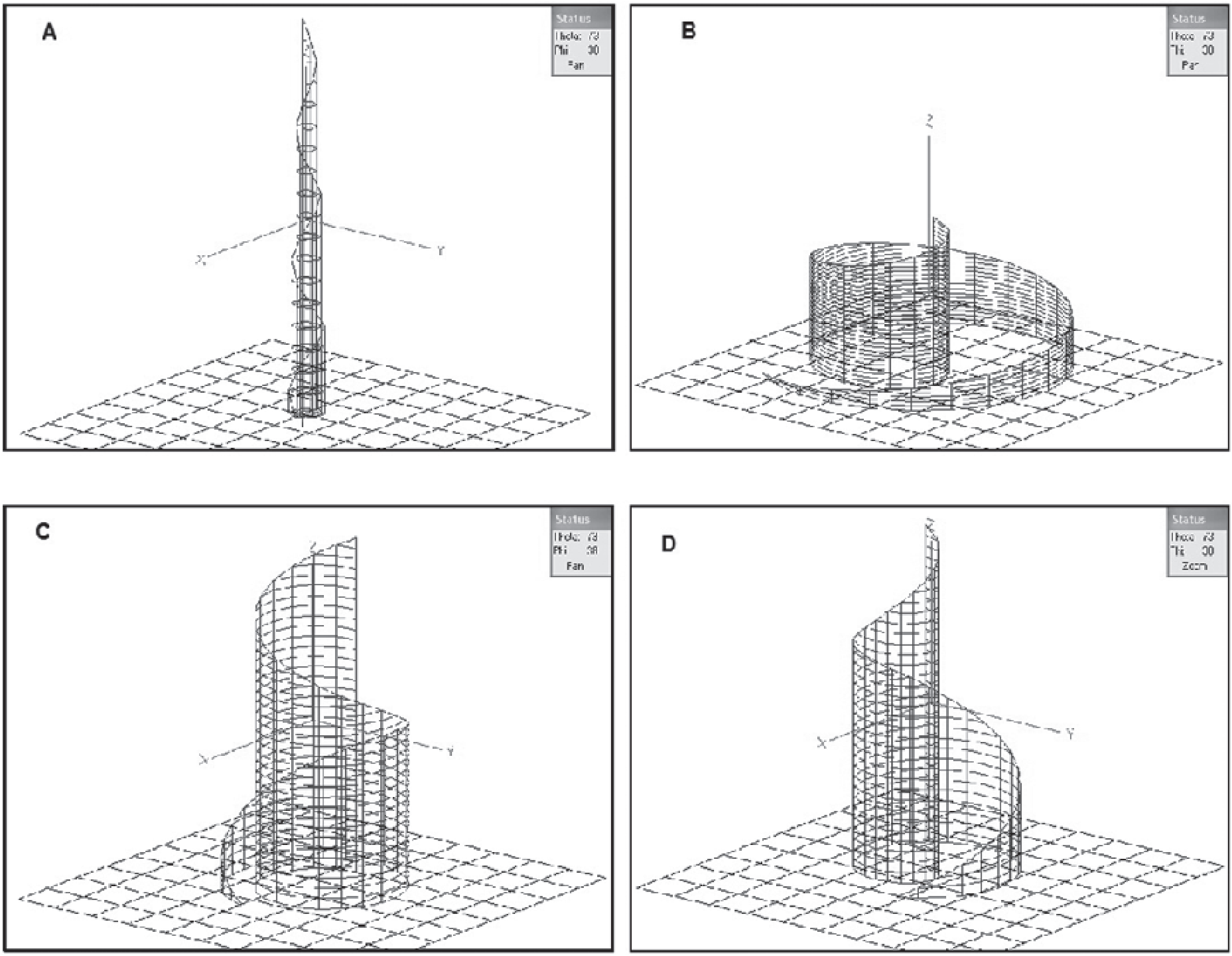

Figure 2 — Construction of several early TASH antenna variants.

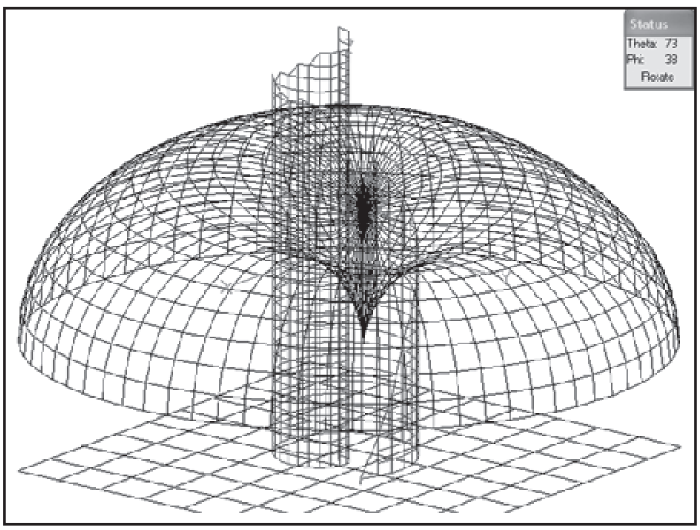

Figure 3 — Typical TASH antenna radiation pattern.

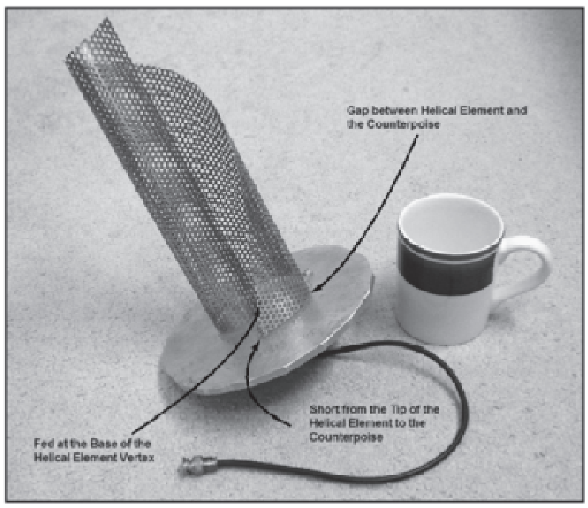

Figure 4 — Improved 225 MHz to 2.5 GHz TASH antenna (cup for scale).

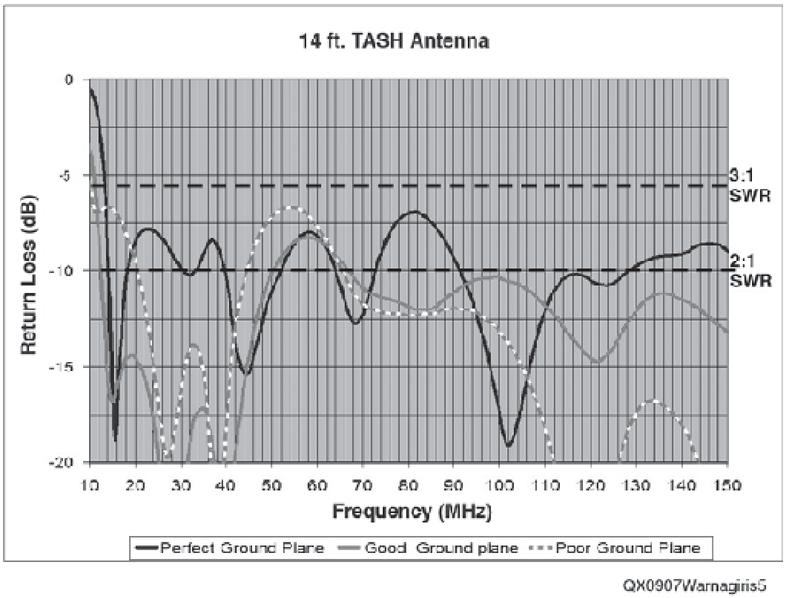

Figure 5 — Return loss and SWR simulations for a TASH antenna over various grounds.

Table 1
TASH Measurements

Band	*Test Frequency*	*Power Out*	*Power Reflected*	*SWR*
20 Meters	14.083 MHz	100 Watts	< 1 Watt	<1.2:1
17 Meters	18.080 MHz	100 Watts	9 Watts	1.85 :1
15 Meters	21.080 MHz	21 Watts	2.2 Watts	1.96:1
12 Meters	24.920 MHz	16 Watts	1.4 Watts	1.84:1
10 Meters	28.100 MHz	50 Watts	4.8 Watts	1.90:1

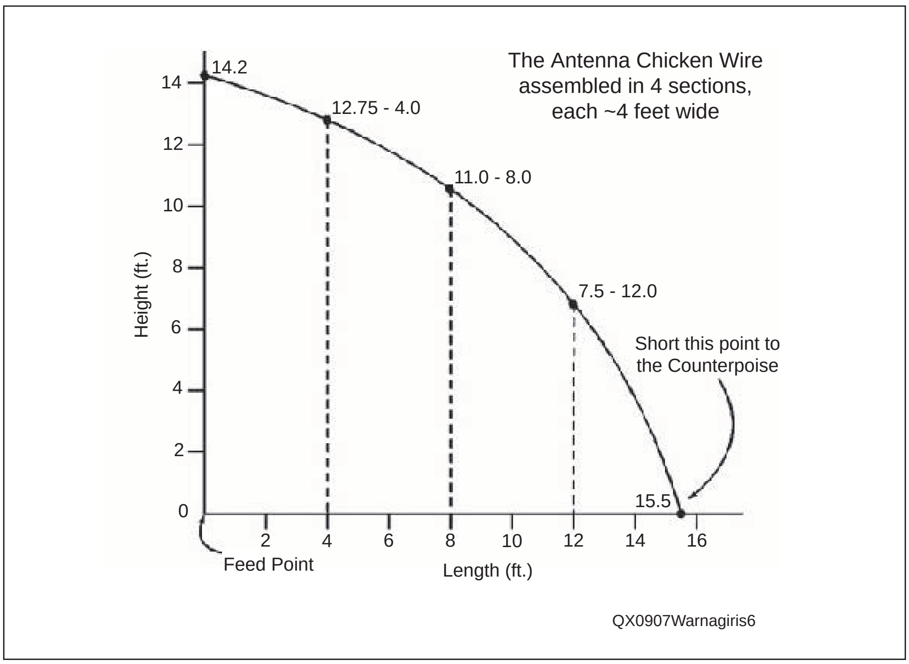

Figure 6 — Dimensions of the chicken wire TASH antenna element.

beam radiator, while the TASH antenna is a mixed polarization omni-directional radiator. Figure 3 shows a typical TASH radiation pattern.

Features and Problems

When compared to conventional discone or cone (half a bicone) antennas, the TASH antenna offered some physical as well as electrical benefits. The original TASH configuration had a few serious problems, however. First, it was difficult to maintain the optimum spacing between the turns of a multi-turn TASH element. The triangular "tapered area" was large and the spacing between turns fairly critical (see Figure 2B). Also, the height-to-diameter ratio made the antenna somewhat bulky. So, further software simulations were made in an attempt to reduce the number of turns and the element area.

Even after much effort the antenna simulation program did not give the desired result. Eventually some drastic changes to the configuration led in the right direction. The changes were placement of a ground plane short at the tip of the TASH spiral and tapering the element in a logarithmic fashion. A short at the element tip produced a TASH antenna with a much greater height-to-diameter ratio than was previously seen. The elements of all earlier TASH antennas were right triangles. The logarithmic taper proved to be the key to diameter reduction. It's now possible to build TASH monopoles with an SWR less than 3:1 over a 10:1 bandwidth, with a diameter only a third of the antenna height.

The first TASH antenna constructed, based on the revised simulations results, was a 13 inch high TASH antenna with a cut off frequency of about 225 MHz. It was intended for military applications at 225 MHz and above. Figure 4 shows the early prototype with a coffee cup for scale. This antenna gave acceptable SWR at frequencies of about 300 MHz with just the counterpoise shown in Figure 4. Doubling the area of the counterpoise moved the low frequency cutoff down to about 225 MHz without appreciably affecting the SWR at higher frequencies. This was in good agreement with computer simulations.

The TASH antenna was simulated using several antenna simulation packages, but the package used most frequently was one called *GNEC*.[5] *GNEC* allows antenna surfaces to be constructed as a grid similar to actual wire antennas assembled from expanded metal or even poultry netting. (*GNEC* is only available to user's who are licensed from Lawrence Livermore National Labs to use NEC4. — *Ed.*) Figure 5 shows the simulated

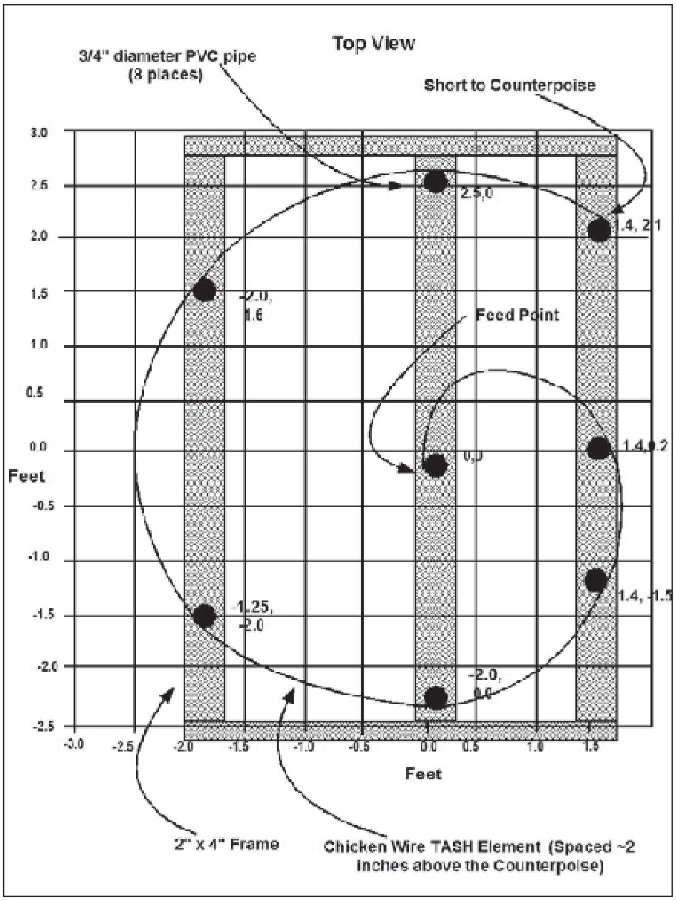

Figure 7 — Dimensions of the TASH antenna base and element curvature.

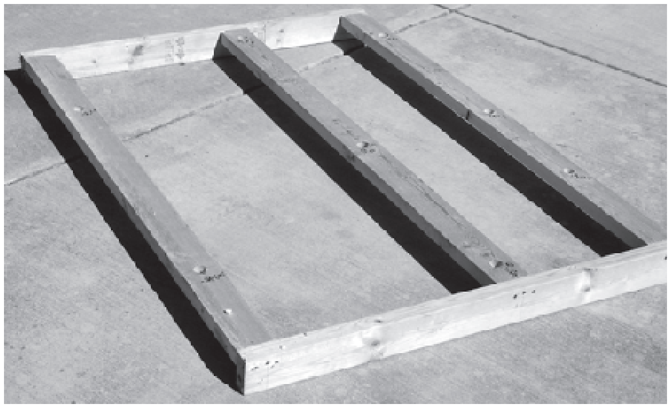

Figure 8 — TASH antenna base prior to assembly of the PVC framework.

return loss and SWR for a 14 foot tall TASH antenna over various grounds. Note that the ground condition does affect the SWR, but not significantly.

Construction of an HF TASH Antenna

Successful simulations of a 14 MHz TASH antenna over various ground planes situations eventually led to design, assembly, and test of a large TASH antenna suitable for operation on the 20 through 10 meter HF bands. Most HF antennas are assembled using copper wire or aluminum tubing. Both materials were out of the question for fabrication of the large element necessary for a 14 MHz TASH antenna. This led to consideration of other suitable materials, such as galvanized poultry netting, better known as "chicken wire."

Chicken wire is definitely a rather unique antenna material. Although chicken wire is not the best conductor (about ⅓ that of copper), its point to point conductance is still rather good. This is due to the large number of conductive paths through the many wire hexagons making up the chicken wire. Using an ohmmeter with the leads several feet apart, the measurement of the dc resistance of a large section of chicken wire gave a reading of less than a few ohms. Fortunately, since the TASH element has such a large surface area, the surface conduction loss is minimal even when fabricated using a poor conductor.

At HF, the chicken wire element need not be cut to the exact dimensions shown in Figure 6. In fact, "chicken wire" and words like precise and exact should probably not be used in the same sentence. Keeping the wire within an inch or so of the antenna design is not easy and really doesn't matter greatly. Both simulations and measurement of a test TASH antenna have shown that input impedance and radiation pattern are rather insensitive to distortions of the TASH antenna structure. The key dimensions of ground plane spacing and overall size/shape of the TASH element seem to be the only aspects showing some criticality. Fortunately, these dimensions are not all that difficult to maintain on the large chicken wire TASH antenna element designed for a frequency as low as 14 MHz.

So, the HF TASH antenna eventually built was simply a scaled up version of the successful 3:1 height-to-diameter ratio 225 MHz TASH design. As mentioned, chicken wire was selected for the TASH element since simulation showed that the surface conductivity of the TASH element was not critical. PVC tubing was assembled as the framework for the element. Chicken wire was also used for the counterpoise. Figure 6 shows the dimensions of the chicken wire element and Figure 7 shows the dimensions of the base and the spiral spacing of the chicken wire element.

The antenna base consisted of several 2×4 boards cut and assembled as shown in Figure 8. The TASH antenna framework was assembled from short sections of PVC pipe interconnected by PVC T-sections. The approximate dimensions of the pipe are shown in Figure 9, in two dimensions. The assembly of the framework is shown in Figure 10. Part of the chicken wire counterpoise was placed on the wood frame prior to assembly of the pipes. Eight vertical ¾ inch diameter PVC tubing was placed in holes drilled at the locations shown in Figure 7. Each vertical piece of tubing was held in place by a wood screw into the 2×4 perpendicular to the tubing (see Figure 11).

Once the framework was assembled, sections of chicken wire were stitched together using galvanized wire to form a single element as shown in Figure 12. Before the element was wrapped on the tubing framework,

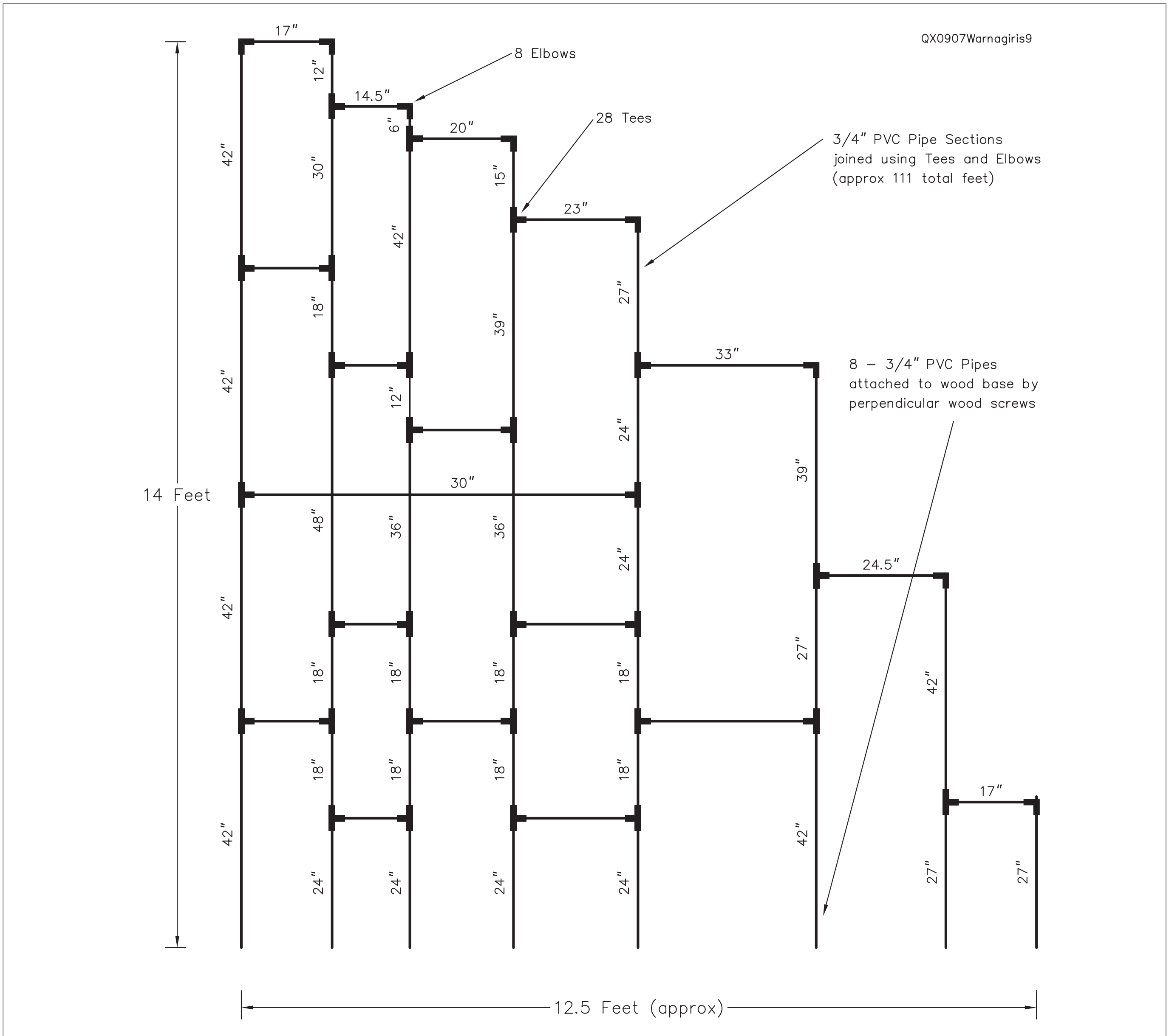

Figure 9 — Dimensions of the TASH antenna PVC framework laid flat for illustration.

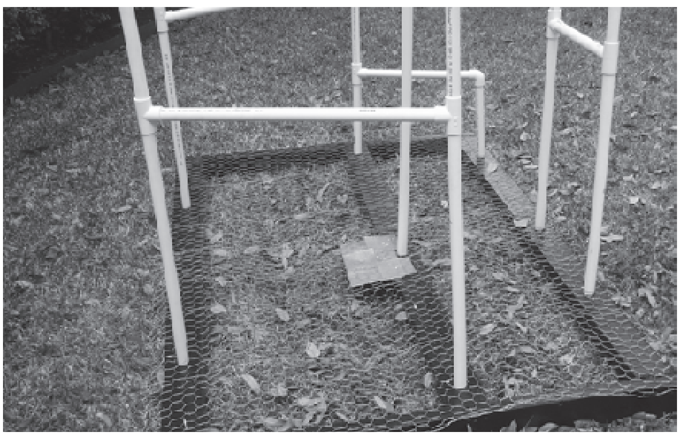

Figure 10 — PVC framework prior to installation of the TASH antenna element.

Figure 11 — PVC pipe secured to the TASH antenna base by a wood screw.

Figure 12 — Chicken wire sections stitched together to form the TASH element.

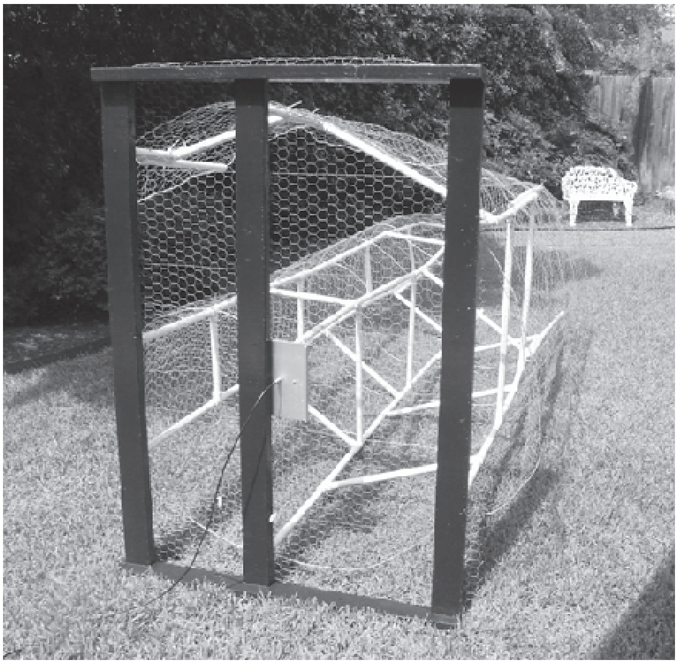

Figure 13 — The complete TASH antenna on its side showing the feed point.

a layer of chicken wire was stapled to the 2×4 frame to form a counterpoise for the TASH antenna element. Then the TASH element was wrapped on the framework spaced about 2 inches above the counterpoise chicken wire. The coaxial feed and element shorts locations are shown in Figure 6. Figure 13 shows the feed point with the antenna on its side. The complete HF TASH antenna is shown is Figure 14.

HF TASH Test Results

The return loss of the HF TASH antenna was measured using an HP 8754 Network Analyzer. Figure 15 shows the measurements with return loss and SWR scales overlaid. Note the similarity in the

Figure 14 — Complete HF through VHF TASH antenna ready for operation.

simulation data for return loss and SWR of the measured data with the simulated data shown in Figure 5.

On air tests were performed using an old ICOM IC-720A transceiver with no antenna tuner. The tests were performed on the bands shown in Table 1.

Full 100 W output was not obtained above 20 MHz due to the transceiver SWR fold-back circuits. Minor antenna adjustments or a simple tuner could easily reduce the SWR to allow full power from the IC-720A. Since the SWR was less than 2:1 on the five bands, newer transceivers could probably handle the SWR without a tuner.

Conclusions

The amateur radio community has several assigned bands at various frequencies in the 2 to 30 MHz range. Unfortunately, they are spaced several megahertz apart, so that few antennas can cover all bands of interest. A TASH antenna can be an excellent choice for situations requiring multi-band operation. The wide bandwidth of the TASH antenna would even allow operation on the VHF bands from the same HF antenna.

One of the best features of the TASH antenna is its dc to ground impedance at the antenna input. It's a short circuit to the counterpoise, which makes it inherently short circuited at frequencies below the lower frequency cutoff. This can reduce the occurrence of intermodulation products from broadcast, appliance hash, and other sources of low frequency noise. The short can also protect the rig from lighting damage and charge build up (a problem with cone and discone antennas), but a slight modification to the antenna might be necessary if lightning protection is a real concern.

The distance between the antenna connection center conductor and the outer return is approximately 20 feet. Measured dc input impedance at the connector of the large TASH antenna shown in Figure 14 was less than 4.0 Ω even after a few weeks in the weather. This is not a very good short to ground, but it in no way reduces the effectiveness of the antenna. Lower impedance for lightning protection is easily provided by adding heavier wire to the edge of the TASH element and shorting it directly to ground. This does not affect the RF performance of the antenna.

Cone and discone antennas, on the other hand, are inherently low pass, or all pass antennas. At dc, the input impedance approaches an open circuit. These antennas often build a charge when left unterminated and can easily damage a receiver without a dc input return. Also, lightning, broadcast signals and power noise can be easily coupled to a receiver. True, filtering at the receiver input can reduce the potential for interference problems, but it's always best to attenuate extraneous signals prior to reaching any receiver front end component including the coaxial input feed line.

Although comparable to a discone or cone antenna electrically, it has a smaller footprint for the same lower frequency cutoff. Another advantage of the TASH antenna over the discone or cone is its physical configuration. Besides requiring a smaller footprint and vertical height for a given lower frequency cutoff, it has most of its mass located near the base. This helps stabilize the TASH antenna during high winds.

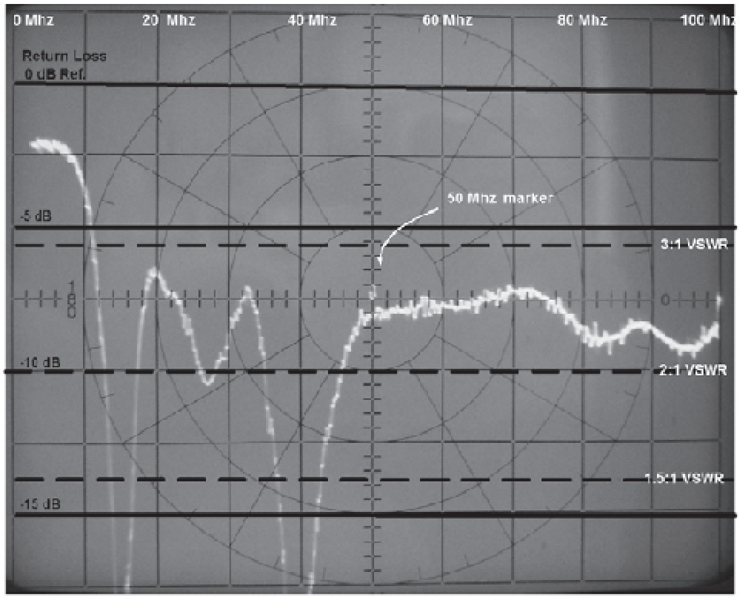

Figure 15 — Measured SWR and return loss of the complete chicken wire TASH antenna.

A TASH antenna is a new option in wide band antennas. It has features not available with other wideband antennas and can be assembled using inexpensive materials, such as chicken wire. Chicken wire can really work wonders on the HF bands.

Notes

[1]One source of information about various types of wide bandwidth antennas is this Web site: **www.g3tpw.co.uk/Page5OtherMulti-bandAntennas.htm**

[2]Thomas J. Warnagiris, K3GSY, "A Monopole with a Twist," *Microwave Journal*, Vol 44, No. 9, Sep 2001, pp 120-137.

[3]Thomas J. Warnagiris, K3GSY, "A Monopole with a Twist Revisited," *Microwave Journal*, Vol 48, No. 7, July 2005, pp 54-74.

[4]There is a brief discussion of antenna polarization at the Web site: **ham-shack.com/polarization.html**

[5]There is more information about GNEC at the Nittany Scientific Web site: **www.nittany-scientific.com/**

Tom Warnagiris, K3GSY, was first licensed in 1958 as KN3GSY. He received his BSEE from the Pennsylvania State University and has performed graduate work at the University of South Florida. He is currently a retired engineer from the Signal Exploitation and Geolocation Division of Southwest Research Institute (SwRI). He previously worked as an engineer at ARINC Research, the ECI division of NCR, C-COR Electronics, and HRB Singer. He holds several communications related patents and is a Senior member of the Institute of Electrical and Electronic Engineers (IEEE).

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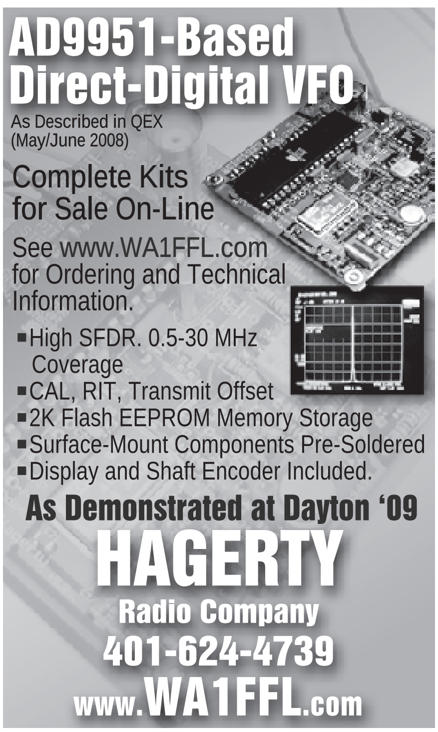

Next Issue in *QEX*

William Moneysmith, W4NFR, describes "A Homecrafted 440 MHz Duplexer." There have been published designs for 2 meter duplexers, but W4NFR could not find a published design for a 440 MHz unit. Using junk-box parts and hardware store supplies, he fashioned this set of duplexers for a new 440 MHz repeater installation. This article will lead you through the construction process as well as the steps to tune the four cavities. With two cavities on the transmit side and two on the receive side, W4NFR measured throughput losses of about 3 dB and an isolation of about 52 dB for his duplelxer.

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Alternatives to *Octave*

In this series of articles, we've been examining the use of *GNU Octave* to solve a variety of Amateur Radio related problems.[1] While I'm a fan of *Octave*, I must admit that there are some good free alternatives for solving numerical problems, such as *Scilab*, *Python*, and *Maxima*.[2, 3, 4, 5]

We should also take note of *Mathcad*,[6] an alternative to *Octave* that, while not free, is a real bargain when purchased from ARRL in connection with a book by Bill Sabin, W0IYH.[7] Bill recently introduced the use of *Mathcad* 14 for signal analysis in *QEX*.[8] Although I tend to toot *Octave*'s horn, I have owned several revisions of *Mathcad* in the past and I have found it to be an excellent design and analysis tool.

Why, then, do I favor *Octave*? Being retired, I no longer have an employer to bear part or all of the cost of software, and *Octave*'s price (free) is a big incentive. In addition, I made a move to *Linux* some years ago and *Mathcad* is not available for *Linux*. Another factor in my choice was my former employer's use of *Matlab*.[9] *Octave* is more compatible with *Matlab* than with *Mathcad*. In addition, *Octave* allows the use of the same operators for dealing with vectors or scalars and *Octave*'s code more closely resembles C than do some of the other choices.[10] These are significant advantages from my viewpoint.

These, of course, are personal reasons and others may make different choices based

[1]Notes appear on page 27.

Table 1
Python Code for Attenuator Design

```
# /usr/bin/python -qf

import math

print "\n\n *** BALANCED H ATTENUATOR DESIGN ***"

# enter data required to design attenuator from keyboard:

loss = float(raw_input("\n    ENTER ATTENUATOR INSERTION LOSS IN dB: "))

zin = float(raw_input("    ENTER ATTENUATOR INPUT IMPEDANCE IN OHMS: "))

zout = float(raw_input("    ENTER ATTENUATOR OUTPUT IMPEDANCE IN OHMS:
"))

# calculate attenuator power ratio from insertion loss:

n = 10. ** (loss / 10.)

# calculate resistance values:

r3 = 2. * math.sqrt(n * zin * zout) / (n - 1.)

r1 = zin * (n + 1.) / (n - 1.) - r3

r2 = zout * (n + 1.) / (n - 1.) - r3

# print out results:

print "\n *** BALANCED H ATTENUATOR RESISTANCES ***\n"

print "     R1/2 = %f" % (r1/2)

print "     R2/2 = %f" % (r2/2)

print "     R3 = %f\n\n" % (r3)
```

on personal circumstances, on personal preferences, and on the types of problems they wish to solve. Although we haven't dealt with symbolic algebra in this series, you might choose *Mathcad*, *Maxima*, or *Python* with *Sympy* if you want to manipulate equations symbolically.[11, 12] If you want a system modeling tool comparable to *Matlab*'s *Simulink*, you might choose *Scilab* with *Scicos* (see Note 3).

Since some of us might already have *Python* installed on our systems, let's compare *Python* with *Octave*. What do we have to change to use *Python* for some of the computations we've been performing with *Octave*? Well, not much for some of them. Table 1 contains a *Python* version of the *Octave* code we used to design a balanced H attenuator in Table 1 of *Octave Calculations for Amateurs*.[13]

Differences between the *Octave* and *Python* code include:

1) No semicolons at the ends of the lines of *Python* code because semicolons are not required to suppress output in *Python*;

2) A line "import math" in *Python* so that the math module will be imported to support our call to math.sqrt;

3) The use of the operator "**" for exponentiation, as *Python* doesn't support "^": *Octave* supports both forms of exponentiation;

4) Hyphenated function calls: a call to a function in a *Python* module must include the name of the imported module that contains the function. Where you might call "sin" for the sine function in *Octave*, you need to call "math.sin" or "cmath.sin" in *Python*, depending on whether the argument is real or complex. *Python* features an object oriented language and if you're familiar with *C++* you should recognize the format of *Python* function calls.

What if we want to use *Python* for something more than its built-in modules support, such as Fourier analysis? We can add capability to *Python* by installing any one of several external libraries. One that supports Fourier analysis is NumPy, which adds a number of advanced mathematical functions to *Python*'s repertoire.[14]

Once NumPy has been installed and you've verified that you can import NumPy into *Python*, you can allow *Python* access to NumPy's functions by including the line "import numpy" in your code. To try this out we'll modify the code in Table 1 of *Octave for Signal Analysis* so that it will run as a *Python* script.[15] The modified code is listed in Table 2.

Note that rather than plotting from *Python* as we did from *Octave*, we're writing data to two files, one with time domain data and one with the corresponding frequency domain

Table 2
Python Code for Signal Analysis (Fourier Series)

```
#!/usr/bin/python

import sys

import math

import cmath

import numpy as np

print "\n\n  ENTER MODULATION FACTOR: ",

m = raw_input()

A_cxr = 1  # carrier peak amplitude

v_cxr = range(1024)

v_mod = range(1024)

v_cxr_out = range(1024)

s_cxr_out = range(60)

# Calculate time domain modulated waveform and write to file

fileout = open("v_cxr.txt", "w")

for n in range(1,1024):

    v_cxr[n] = A_cxr * math.cos(2 * math.pi * n / 32)

    v_mod[n] = (1 + float(m) * math.sin(2 * math.pi * n / 512))

    if v_mod[n] < 0:

       v_mod[n] = 0

    v_cxr[n] *= v_mod[n]

for n in range(1,1024):

    v_cxr_out[n] = str(str(n) + ', ' + str(v_cxr[n]) + '\n')

    #print v_cxr_out[n]

    fileout.write(v_cxr_out[n])

fileout.close()

# Calculate magnitude of frequency domain modulated waveform

#    and write to file

fileout = open("s_cxr.txt", "w")

s_cxr = np.fft.fft(v_cxr) / 512

for n in range(1,60):

    s_cxr_temp = abs(s_cxr[n])

    s_cxr_out[n] = str(str(0.5 * n) + ', ' + str(s_cxr_temp) + '\n')

    #print s_cxr_out[n]

    fileout.write(s_cxr_out[n])
```

Table 3
***Gnuplot* Commands for Plotting Time and Frequency Domain Vectors**

```
# The time domain data in v_cxr.txt may be plotted by
# executing the following commands from within Gnuplot:
set title "TIME DOMAIN SINUSOID"
set xlabel "TIME: 1.95 us PER SAMPLE"
set ylabel "AMPLITUDE"
set grid
plot "./v_cxr.txt"
# The frequency domain data in s_cxr.txt may be similarly
plotted
# using the following commands from within Gnuplot;
set title "AM SIGNAL FREQUENCY SPECTRUM"
set xlabel "FREQUENCY IN kHz"
set ylabel "AMPLITUDE"
set grid
plot "./s_cxr.txt"
```

data. Each file stores two columns, one column with data for the abscissa (X axis: time or frequency) and the other column with data for the ordinate (Y axis: amplitude).

We'll plot the data by first starting *Gnuplot*.[16] We'll then execute the lines shown in Table 3 from the *Gnuplot* command line and we'll end up with graphs of time domain and frequency domain functions that are essentially the same as those in Figures 3 through 6 of *Octave* for Signal Analysis.[15, 17]

There are some advantages to using a datafile as an interface between the math utility and the plotting program even if we're using *Octave*. The developers of *Octave* have been working on an internal plotting functionality that is intended to replace *Gnuplot*. Some of the *Gnuplot* functions have been deprecated in *Octave*, and with the move to *Octave* 3.x, some of them are no longer available. We can rewrite our code to use *Octave*'s internal plotting if we would like to do so, but it's still under development, while *Gnuplot* is a mature, stable plotting tool. Note that if we use datafiles to transfer calculation results to a plotting utility, we can use any combination of calculating and plotting tools, giving ourselves some versatility should we desire to use more than one of either type of software tool.

In addition, data stored in a datafile can be analyzed or graphed by importing it into another program that may be more appropriate for the problem or problems under consideration.

Hopefully, that gives us some insight into what's out there in the way of math software. While most of the commercial packages are too expensive for the average amateur mathematician (or radio operator), some of the free ones are excellent performers. You might try downloading more than one and comparing them on problems that interest you. Personal comparisons can give you some insight into the differences among the various utilities and may give you reasons to favor a particular implementation.

Notes

1. See **www.octave.org**
2. If I've left out your favorite math utility, I apologize. I've included the ones referenced in this article because I've had opportunities to use them and to compare them with *Octave* and with commercial utilities such as *Matlab* and *Mathcad*. There are other excellent math utilities.
3. See **www.scilab.org**
4. See **www.python.org**
5. See **maxima.sourceforge.net**
6. See **www.mathcad.com**
7. William E. Sabin, *Discrete-Signal Analysis and Design,* 2008, Wiley, (**www.wiley.com**) Interscience Division, available from your local ARRL Bookstore, Order no. 0140. Telephone toll free in the US 888-277-5289 or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**
8. William E. Sabin, W0IYH, "A Modern Discrete-Method for Signal Analysis and Design," *QEX*, Nov/Dec 2008, pp 36-38.
9. See **www.mathworks.com**
10. The "normal" operators in *Octave*, such as "+," will not operate element by element on vectors or matrices, but will perform matrix operations, which may not always be what we want. The "dot" operators, such as ".+," will, though, operate element by element on scalars, vectors and matrices. You'd be safe in using the dot operators for everything except where you specifically want matrix operations.
11. See **en.wikipedia.org/wiki/Comparison_of_computer_algebra_systems**
12. See **wiki.sympy.org**
13. Maynard Wright, W6PAP, "*Octave* — Calculations for Amateurs," *QEX*, May/June 2005.
14. See **www.numpy.org**
15. Maynard Wright, W6PAP, "*Octave* for Signal Analysis," *QEX*, Jul/Aug 2005.
16. *Gnuplot* is available at: **www.gnuplot.info**. You should already have *Gnuplot* installed if you've been using *Octave*.
17. The paths associated with the filenames in the plot commands in Table 3 are based on the use of the *Linux* file structure and on having the files located in the logged-in directory. You may need to change this to fit your operating system and file structure.

Maynard Wright, W6PAP, was first licensed in 1957 as WN6PAP. He holds an FCC General Radiotelephone Operator's License with Ship Radar Endorsement, is a Registered Professional Electrical Engineer in California, an ARRL Member, and a Life Senior Member of IEEE. Maynard has been involved in the telecommunications industry for over 45 years. He has served as technical editor of several telecommunications standards and holds several patents. He is a Past Chairman of the Sacramento Section of IEEE. Maynard is Secretary/Treasurer and past President of the North Hills Radio Club in Sacramento, California.

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Maximizing Radiation Resistance in Vertical Antennas

There is more than one way to increase antenna efficiency. Here is one that is not often considered.

Statement of Problem

The typical Amateur Radio HF vertical antenna is installed with a less than optimum ground radial system. AM broadcast vertical arrays use 120 buried radials. The purpose of these radials is to minimize ground resistance, and thus maximize antenna efficiency. The time and cost of such an installation can be prohibitive, not to mention the complete disruption (lawns, gardens, sidewalks and so on) of the circular land parcel involved. For the 160 meter band, the area of this optimum ground radial field is about 1 acre! In the majority of installations, amateurs attempt to optimize their antennas and ground systems for a given set of limitations. The whole point of all the work, time and money invested in vertical ground systems is to reduce the loss resistance, *R*. Since *R* rarely approaches very low values (<1 or 2 Ω) in amateur installations, raising the radiation resistance, R_r, can have an equally beneficial effect. Indeed, a single piece of wire (capacitive hat) can have an equally profound effect on antenna efficiency as burying scores of radial wires.

Some recent empirical studies have provided ever more precise insights into some of the trade-offs of compromised ground systems.[1] Little if any work has been published in amateur literature on the other half of the efficiency problem — radiation resistance. The first purpose of this paper is to derive a general equation for the calculation of R_r in vertical antennas. Then, through theoretical analysis, I attempt to answer the question: *What is the optimum design of a given physical vertical height to maximize radiation efficiency?*

There are really two issues with ground losses in vertical antenna systems. The first is lowering *R* by installing a ground radial system. The lower the ground losses, the more efficient the vertical will be. The second is the far field effect of the ground on low angle gain and overall antenna directivity. The higher the ground resistance in the far field, the more attenuated the low angle energy will be, and thus maximum gain will be at some elevated angle above the horizon. Therefore, the directivity of a vertical antenna of a given height will be constant given the terrain characteristics in the far field. The *gain* of the antenna will be simply directivity × efficiency (this is a standard equation). We have some degree of control over the first ground problem (efficiency), but nearly no control over ground conductivity (not to mention slope, structures, vegetation and so on) multiple wavelengths from the antenna (directivity). We can have a 100% efficient antenna and still be victim of what happens in the far field, like poor ground conductivity. The best we can do is maximize the efficiency of our vertical within our given limitations. This paper focuses on the often neglected R_r. Finally, we can certainly tailor directivity by using multiple-element arrays. This paper will concentrate on the single vertical element array, but hopefully its implications will also aid in increasing efficiencies in multi-element vertical arrays.

An antenna consists of all objects (including the ground) that lay within the near field. Therefore, losses encountered in the near field will affect antenna gain, efficiency and directivity. Objects may be intentional parasitic elements, unintentional parasitic elements, energy absorbers, ground losses (series *R*) and/or shunt losses (parallel *R*).

Losses encountered in the far field affect propagation losses and reflections. Coupled with the gain and directivity of the antenna proper, the far field also shapes gain and directivity; essentially, the two effects add. Of course there is no clear boundary between the near and far fields, but the far field can loosely be defined as the region where the relationship between the magnetic and electric fields "settle" according to Maxwell's closed form equations (several wavelengths away from the "antenna").

Secondary to series ground losses within the near field of a vertical antenna are losses due to objects that appear as parallel shunting resistors. These objects may simply be attenuators or may act as resistors connected to the ground, such as trees, buildings, and so on.

Critical Parameters

Radiation resistance is often confused with feed point impedance. There are some instances where the two values may be quite similar (for example, a base-fed ¼ λ vertical). Our first task is to define radiation resistance, R_r, with precision. Krauss gives a complete but brief discussion of R_r for vertical antennas.[2] We need some derivation to define R_r. From Krauss we get:

$$A_e = \frac{h_e^2 Z_0}{4R_r} \quad \text{[Eq 1]}$$

[1]Notes appear on page 33.

where:
A_e is the effective antenna aperture
h_e is the effective antenna height
R_r is the radiation resistance
Z_0 is the impedance of free space (377 Ω)

Solving for R_r we get:

$$R_r = \frac{h_e^2 Z_0}{4A_e} \quad \text{[Eq 2]}$$

Thus, we can simplify the equation to:

$$R_r = 94.25 \frac{h_e^2}{A_e} \Omega \quad \text{[Eq 3]}$$

Notice the absence of any reliance upon feed point impedance on R_r in this general equation.

For those who may be interested, the derivation of this equation is found in Chapter 2 of Krauss's text. I will explain the various terms in this equation, however, so the reader may develop an intuitive understanding of the terms, including the all-important R_r.

It often comes as a surprise that "free space" indeed has a characteristic impedance, much like a transmission line. (See the Impedance of Free Space sidebar.) This value is derived from μ_0 and ε_0, the permeability and permittivity of free space.

Calculation of R_r

Now that we have a simple equation that can be used to directly calculate R_r, we only need to know the antenna effective height (h_e) (not to be confused with electrical length or physical height) and the effective antenna aperture, A_e. As implied by the terms, h_e is defined in one-dimension space, usually defined in meters or feet. Aperture is an equivalent area (two-dimension space), usually defined in square feet or square meters. So the linear space terms cancel and the solution is in ohms.

A further point of confusion is differentiating *antenna R_r* with R_r at some *point on the antenna*. Antenna R_r is a function of the *average current along the antenna element*, while R_r at any given point (usually a high current point) is simply calculated using the current value at that point. Antenna R_r is the more important value, as it is used for *antenna* loss calculations.

In a previous *QEX* paper I presented a brief explanation of antenna aperture and its relationship to gain, so I will only review the absolute basics here.[3]

Antenna Aperture (A_e)

The aperture of an isotropic antenna is:

$$A_{iso} = \frac{\lambda^2}{4\pi} \quad \text{[Eq 4]}$$

This value is of primary importance for anyone who wishes to pursue antenna theory in depth. The power gain of any antenna compared to an isotropic antenna can be expressed as an equivalent area (aperture) compared to the aperture of the isotropic radiator (Equation 4).

The aperture principle is very simple: the more "area" your antenna occupies (oriented toward the desired direction) the more power you will receive for a given field strength, or you will concentrate in a given direction when transmitting. An excellent analogy is the solar cell. The greater the area (lots of solar cells) and the better the orientation to the sun (perpendicular), the more solar power you

Impedance of Free Space

During the mid 19th Century the values of μ_0 and ε_0 were known, since they can be measured using steady state (dc) values. James Clerk Maxwell in the 1870s noticed that $1/\sqrt{\mu_0 \varepsilon_0}$ was very close to the then-known speed of light. In fact, this equation later proved to be the speed of light that we know today (about 300,000,000 m/s). Maxwell, therefore, was the first to propose that light was a form of electromagnetic radiation and was also the first to predict the existence of radio waves. Even more significantly, his set of closed-form equations defined exactly the relationships of the electric and magnetic fields. His equations remain today the basis of all study in electromagnetism, written a least a decade before radio waves were discovered. His work is perhaps the greatest scientific achievement of the 19th Century.

Series and Parallel Losses

When both series losses (such as ground losses) and parallel shunt losses (such as from nearby trees and buildings) are present we can calculate overall efficiency by:

$$efficiency = \frac{R_p}{R_p + R_r}\left(\frac{\rho}{R_s + \rho}\right) \quad \text{[Eq 11]}$$

where:
R_r is radiation resistance
R_p is parallel R losses
R_s is series R losses
and

$$\rho = \frac{R_r R_p}{R_r + R_p}$$

When only parallel losses are present we get:

$$efficiency = \frac{R_p}{R_p + R_r} \quad \text{[Eq 12]}$$

To minimize R losses in antennas we want to maximize R_p (and minimize coupling to this parallel resistance) and minimize R_s. Equation 12 is simplified in that R_p represents the actual equivalent resistance at the operating frequency of the building or tree taking into account the decrease of resistance due to coupling (usually capacitive) into the object from the antenna. A more accurate and more complex model would include the actual reactive coupling from the antenna and the resistance of the object at the operating frequency. It is important, however, to distinguish between the parallel and series losses, especially in vertical antennas and this note is provided to present a simplified math model for what is actually happening in a typical amateur vertical antenna, where trees, buildings and ground losses are all present.

Notice that when a parallel resistive loss is encountered, overall loss can be minimized by lowering R_r! Lowering R_r, however, also often results in a lowering of the feed point impedance and of course increases series R losses. Therefore this statement must be interpreted with great care. When both are present, some compromise R_r will be optimum. For most amateur installations, however, the dominant loss will be series resistance; therefore maximizing R_r will be the usual goal.

can capture. Double the area, and you double the power received. Of course, the aperture of an antenna is not the actual physical area of the antenna element(s). For example, the thin wires of a ½ λ dipole provide a rectangular aperture "area" about ½ × ¼ λ broadside to the dipole. Placing a reflector near the dipole, thus creating a 2 element Yagi increases the aperture. Placing the dipole near a dense tree decreases the aperture.

If the gain is given in dBi, we can derive the power gain over an isotropic radiator by using the inverse-log or anti-log equation:

$$G_{iso} = 10^{\frac{dBi}{10}} \quad \text{[Eq 5]}$$

To calculate dB from power we use a log function. To go backwards, we use the anti-log function of Equation 5.

An antenna's aperture *(A_e)* is then simply:

$$A_e = G_{iso}\frac{\lambda^2}{4\pi} \quad \text{[Eq 6]}$$

In other words, if we have an antenna with a gain of 3 dBi (3 dB over an isotropic radiator), G_{iso} (the power gain) is about 2. So the aperture of a 3 dBi antenna is twice the aperture of an isotropic radiator, or 2 ($\lambda^2 / 4\pi$).

It is possible to directly calculate the aperture of any antenna using rather complicated vector calculus. This process is laboriously performed in Chapter 5 of Krauss's text. Fortunately, we can use a much easier source. We can "cheat" and use *EZNEC*!

EZNEC will give us gain in dBi (or other dB references). Gain is calculated by *EZNEC* using perfect ground and zero loss antenna elements. The dBi gains at the 0° elevation are then used to derive the values of A_e by using Equations 5 and 6. These values are, in practice, not attainable because of losses, but we are calculating R_r. After we calculate R_r, we can then also calculate antenna efficiency, which factors in these losses by Equation 7:

$$efficiency = \frac{R_r}{R_r + R} \quad \text{[Eq 7]}$$

Notice that Equation 7 is for *series R* losses only. Ground losses in verticals are usually assumed to be series losses, since there is an equivalent resistance along the ground radial lines. *Parallel R* losses may be represented by nearby objects like trees and buildings located within the near field. These "circuits" may be modeled by an equivalent capacitive coupling in series with a grounded resistor or a simple attenuator. The mathematical model for the effect upon vertical efficiency when both series and parallel losses are present is provided in the Series and Parallel Losses sidebar. In the main body of this paper, however, *R* is assumed to be series resistance losses.

Effective Antenna Height (h_e)

As we can see from Equation 2, R_r is proportional to the square of h_e, and this becomes the key term. In effect, if we maximize h_e, we maximize R_r. We must carefully define h_e, however. It must be separated from the actual physical height of the vertical, h_p.

The verbal definition of h_e is the *average* RF current along the vertical physical length, h_p, multiplied by the actual physical height, h_p. *It is important to note that the effective height of an antenna element can never exceed the physical height, but the two values can approach near equal values.* This fact is in contrast to claims that vertical folded dipoles can actually increase R_r for a given vertical height. Again, it is important to separate feed point impedance from antenna R_r.

There are two current functions that are important for our purposes: a sinusoid distribution and a linear distribution along the length of the vertical antenna. Sinusoid distributions are found along antenna lengths that are usually found in fixed amateur antennas (where the lengths are about ⅙ λ or longer). The linear distribution is a short cut we can use for very short antenna lengths (low band mobile whips) since sin *z* is nearly equal to *z* for very small angles—"angle" here is represented by (h_p / λ) × 360°. So for very short antennas, the angle is very small.

For our discussion we will focus on the sinusoid distribution, but the calculation of the linear distribution warrants simultaneous discussion, since the linear case is an easier calculation and an understanding of how short antennas function is very useful

An *EZNEC* Model

A relatively simple modeling experiment using *EZNEC* will demonstrate several of the key points made in this paper. Model a 70 foot tall perfectly conducting vertical antenna over perfect ground with 37 segments. Feed the antenna at Segment 1 (the base). At 3.5 MHz, *EZNEC* yields 5.17 dBi gain with a 3 dB beamwidth of 38.9°, and a feed point impedance of about 38 Ω resistive.

Now place a 1 Ω resistor in each of the 36 segments above the feed point. The gain has dropped to 3.47 dBi and the feed point impedance is now 56 Ω resistive, but the 3 dB beamwidth is still exactly 38.9°, indicating no change in directivity. Identical directivity implies that *there is no reduction in h_e with series loss.* If we assume that the 56 Ω represents the equivalent $R + R_r$ term, and from the first lossless experiment above, R_r is 38 Ω. Using Equation 7, we get about 68% efficiency, which agrees exactly with the gain ratios of the two examples. Here again, *EZNEC* is used to perform difficult calculus, this time relating a linearly distributed series resistance to the real portion of the feed point impedance (36 Ω series dc resistance versus a Δ 18 Ω in feed point impedance). We are left with simple algebra and anti-log functions.

We can change the resistive values to relatively high values, such as hundreds of ohms and see very little change in directivity (38.9° beamwidth). But the gain fluctuates in a deterministic manner with the real part of the feed point impedance according to Equation 7. Of course, if we raise a series resistance to a sufficiently high value, the directivity will begin to reflect a "true" physical height, due to an effective opening in the antenna element, but for losses commonly experienced in amateur ground systems, this rule of thumb holds: *in a vertical antenna, directivity and h_e are independent of series losses.* This is in contrast to the often mistaken assumption that adding more ground radials will lower the radiation angle. It may indeed improve efficiency and gain, but it will not affect directivity.

The result of this method is fully reciprocal for either a perfect ground and lossy vertical (as shown above) or an equivalent lossy ground and perfect vertical (the usual situation) by the principle of superposition. For a practical > ⅛ λ vertical antenna, we can assume near-perfect conduction in the vertical, and the bulk of losses will be in the ground.

Now if we use the same two verticals (lossy and perfect) and model them at 7 MHz, we can see the advantage of a ½ λ versus ¼ λ over an identical lossy ground. The gain numbers are 6.1 and 6.9 dBi respectively. The ½ λ vertical is 84% efficient versus the ¼ λ at 68% efficient for the same linear loss radiator (0.514 Ω/foot). The 3 dB beamwidth, however, remains the same for the two ½ λ models, again, showing directivity and h_e did not change with increased loss.

Aside from providing insights into relative efficiency of verticals over similar grounds, proving a simplified method of measuring ground resistance, this exercise also demonstrates the independence of R_r and R in practical vertical antennas, and quantifies the difference between ½ λ and ¼ λ verticals operating over identical ground losses.

for many amateurs, particularly on the low-frequency bands. The derivation of R_r from Krauss's equation is also shown to converge on more frequently published abbreviated equations for very short vertical antennas. This will be shown in detail later.

Effective height for a sinusoid current distribution is thus:

$$h_e = \frac{1}{I_0}\int_{\theta}^{\phi} \sin(z)\, dz = \frac{I_{av}}{I_0} h_p \qquad \text{[Eq 8]}$$

This equation provides a scalar (dimensionless) value that reflects the average current as a percentage of the maximum current along the radiator for the sinusoidal case. We use the given phase points (θ and ϕ) found at the top and bottom of the vertical. For example, a ¼ λ vertical is 90° high, therefore we use 0° and 90° for θ and ϕ.

We must define another convention before we actually solve this integral for the ¼ λ vertical. Rather than using degrees for angle measurement, we will use radians, which are much more convenient than degrees in these types of calculations; 360° = 2π radians. So for a ¼ λ vertical, the equivalent 90° height is π / 2 radians.

We find the integration results in a current of 1 A. I_0 is then π /2 × 1 (radian height of the vertical × average current). So the average current is 1 / (π /2) or 0.637, or I_{av} / I_0. So, the average current over the length of an unloaded ¼ λ vertical is 63.7% of the maximum current on the vertical (maximum in this case being at the base). So if the vertical is 0.25 λ high, then the effective height *(h_e)* of the vertical portion of the antenna (ignoring ground for the monument) is 0.64 × 0.25 λ or 0.159 λ.

Data for Various Unloaded Vertical Heights

In Table 1, h_p is the physical height in wavelengths of the vertical element. We can calculate *he* by including the ground image (multiplying h_p calculated in Equation 8 by 2). The gain in dBi is read directly from the 0° elevation from *EZNEC*. G_{iso} is then derived from Equation 5, A_e from Equation 6, and finally R_r from Equation 3.

Table 1 shows calculated data for various vertical physical heights with no top loading. For the three shortest lengths, the current distribution is assumed to be linear, therefore the average current is simply 50% of maximum (we don't need to solve the integral in Equation 8). For ¼ λ and higher, the sinusoid distribution is assumed and Equation 8 must be used.

Notice that the apertures (gain) for very short verticals (less than about ⅛ λ) remain nearly constant as we change h_p. For very short verticals (from 0 to about 1⁄50 λ), the dBi value derived from *EZNEC* modeling was exactly 4.77 dBi (a power gain of exactly 3) which agrees *exactly* with Krauss's calculus. Also, since the gains — and thus apertures — do not change with antenna height for very short heights, we can rearrange terms by substituting A_e from Equation 6, and multiply the isotropic aperture by 3:

$$R_r = 94.25\frac{h_e^2}{A_e} = 94.25\frac{h_e^2}{\frac{3\lambda^2}{4\pi}} = 395\left(\frac{h_e}{\lambda}\right)^2 \qquad \text{[Eq 9]}$$

This is the equation commonly published for short vertical antennas. (Again h_p is 2 × the vertical height to include the ground image.) With no top loading, the average current is about 50%. If appropriate top loading is installed, then the current average approaches 100% and h_e is now nearly equal to h_p. Sometimes, h_e is taken as just the vertical's height (ignoring the ground contribution to h_e). In this case, Equation 10 should be used. Again, be careful when defining h_e!

$$R_r = 1580\left(\frac{h_e}{\lambda}\right)^2 \qquad \text{[Eq 10]}$$

Bottom Fed ½ λ Antenna Feed Point Impedance

It should be noted that the feed point impedance of ½ λ base fed verticals is quite unpredictable in practice. It is better to use a ⅝ λ vertical or shorter than ½ λ to avoid excessively high feed point impedances. Indeed, running full power into a Bobtail curtain results in very high voltages at the feed point. Very heavy duty lumped matching elements and switches must be used to prevent often spectacular component failures!

Beside excessive ground losses in very short verticals, *R* losses in matching lumped elements (particularly inductors) and in the vertical radiator itself must also be carefully considered. The lower R_r becomes as we shorten the radiator, the more significant ground and other losses become.

From Table 1 it is apparent that maximum R_r occurs with a ½ λ vertical. This will be the most *efficient* vertical antenna possible over any non-perfect ground. Notice that R_r for the ½ λ vertical is about 3 × R_r for the ¼ λ vertical. The efficiency difference between the two can be calculated using Equation 7, assuming you have determined a value for *R*.

A further possible advantage of high verticals (greater than ¼ λ) may be that ground losses also decrease. For example, if the ground current and voltage are "images" of the vertical current and voltage, the high ground current peak will be removed from the vertical base by an electrical ¼ λ for a ½ λ vertical. The displacement current is thus being "spread" over a greater area of the ground's equivalent square resistance. In principle, this should reduce losses, but good empirical data is needed to prove or disprove this hypothesis. This is yet another complexity to the already very difficult task of quantifying ground losses in verticals. No matter what the ground losses are, and if or how they change with different current distribution, raising R_r will *always* improve antenna efficiency over an imperfect ground.

A crude estimate for *R* over a given ground system can be calculated by measuring the base feed impedance of a ¼ λ vertical operating over the ground system. The real portion of the impedance should show a value somewhat more than 37 Ω. The difference is a crude estimate of *R*. For example if we measure 60 Ω at the feed point, *R* will be 23 Ω and the antenna efficiency will be about 62%, as calculated by Equation 7. (See the sidebar, An *EZNEC* Model for more information.)

Recent empirical data agrees with this inference. When experimenting with different ground radial lengths, the feed point impedance changed (and therefore ground resistance) in close correlation with a change in antenna gain (within 1 dB of predicted values).

Empirical support for the independence of Rr and R is provided by Part 3 of the N6LF series (see Note 1), in the Mar/Apr 2009 issue of *QEX*, page 29 and 30. The data in Figure 1, Figure 2 and Table 1 of that article

Table 1
***R_r* For Various Ground-Mounted Vertical Physical Heights With No Loading**

h_p (λ)	h_e (λ)	dBi	G_{iso}	A_e	Rr (Ω)
1/32	0.031	4.78	3.01	0.240	0.4
1/16	0.063	4.78	3.01	0.240	1.6
1/8	0.125	4.86	3.06	0.244	6.0
1/4	0.32	5.17	3.29	0.262	36.8
3/8	0.54	5.79	3.79	0.302	92.1
1/2	0.64	6.88	4.88	0.388	99.5
5/8	0.587	8.12	6.49	0.516	62.9

can be used to demonstrate that the measured vertical antenna gain is within 0.2 dB of the value calculated by Equation 7 in this article. To use Equation 7 with complex impedances, ignore the imaginary part and use only the resistive values for the measured impedances. The ratios of these efficiency terms represents the ratios of the radiated power terms. Converting to dB shows the near-perfect agreement. This theoretical agreement with measured results indicates Rudy's scrupulous experimental techniques. His data from Part 1 is a bit more off theoretical values (0.5 dB), but as he points out, use of only 4 radials can be a bit "flakey."

Relative measurements can be taken using relatively simple field strength meters. A receiver can also be used by operating in its linear region, the AGC turned off, and the detected audio voltage ratios of a CW signal recorded (using a scope or a good audio voltmeter). The *power* difference in dB is then 20 log of the voltage ratios. More accurate measurements are possible using a vector voltmeter. See Note 1.

Efficiency does not translate into the highest gain possible, especially at low elevation angles. For example, the ⅝ λ vertical will have the highest gain at lower elevation angles for any given far-field ground loss, but notice that R_r has dropped, thus lowering efficiency compared to the ½ λ vertical for any given non-zero ground loss. In most cases, *R* losses won't be so high as to make the ½ λ vertical outperform the ⅝ λ antenna at lower angles. (See the Bottom Fed ½ λ Vertical Feed Point Impedance sidebar.)

It should also be pointed out that this average current calculation ignores relative phases of the currents along the vertical axis. For the ⅝ λ vertical, the relative phase has to be corrected to yield the proper result. Therefore, this method needs modification (segmentation and summing) for antenna heights over ½ λ. Longer vertical lengths are usually avoided in amateur applications since the radiated elevation angle quickly rises as the length is made longer than ⅝ λ. (Obtaining a low radiation angle is often the primary incentive to using a vertical antenna.) For example, if we calculate the average current for a 1 λ vertical antenna, we notice that the average current is zero (implying zero effective height). Indeed, there is no radiation at the 0° elevation angle (also implying zero aperture), but there *is* radiation at higher angles. So this simplified analysis only holds for vertical heights less than or equal to about ½ λ, with the ⅝ λ case requiring a phase adjustment for current averaging. (See the sidebar, Integration of Currents in Vertical Antennas.)

Optimizing Top Loading

It is well documented that top loading raises R_r and thus improves efficiency of short vertical antennas. In effect, top loading changes the average-to-maximum current ratio from about 50% to nearly 100% when proper loading is employed. This has the effect of doubling the all-important h_e term, and thus increasing R_r by a factor of four, as we found with Equations 9 and 10. This is of extreme importance for short antennas.

What about higher vertical antennas, however? Can we improve vertical efficiency by top-loading even resonant verticals? If so, is there an optimum loading structure for a given physical height and operating wavelength? Is there any simple guideline for developing a loading structure? In addition to quantification of R_r, the second main purpose of this paper is to provide a "yes" answer to these questions.

For a given physical height limitation for a vertical antenna (a very real limitation for most amateurs operating at low frequencies) the goal should be to maximize the average current over the length of the vertical element *(exactly the goal for very short antennas)*. For physical lengths less than ½ λ, this can be simply accomplished by placing top loading in such a manner that the maximum current appears halfway up the vertical element. For the ½ λ vertical, by definition, the highest current point is at the center, and needs no further loading for optimization. One feature of *EZNEC* is the illustration of current along an antenna element. This is a very convenient feature while modeling a vertical antenna and experimenting with different top loading configurations. You know you'll be close to optimum when you can "eyeball" the current maximum near the vertical's center.

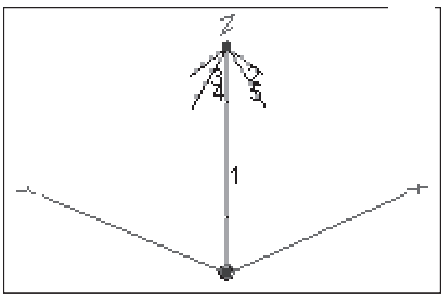

Figure 1 — An EZNEC representation of a ¼ λ vertical antenna with top loading wires.

Example: Optimization of a ¼ λ Vertical Antenna

We calculated from the sinusoid definite integral that the average current/maximum current on a ¼ λ segment is 0.64. Now suppose we use a ⅛ λ top loading wire (45° or π/4) on the same ¼ λ vertical, to configure the familiar "inverted L." The current maximum now appears halfway up the vertical rather than at the base, since the angles defining the definite integral are now 45° and 135° instead of 0° and 90° (the antenna height remains 90°). This results in a new average current of 90%, and h_e increases from 0.32 λ to 0.46 λ.

Let us compare the performance of an unloaded ¼ λ vertical and a ¼ λ vertical with an electrical equivalent of ⅛ λ top loading. See Table 2.

Although we only see a modest lossless gain increase from the addition of the top loading, R_r doubles. This can have a significant overall gain improvement over lossy grounds. Similar dramatic improvements for R_r can be realized for any vertical height less than ½ λ, but this is particularly true for all heights less than about ⅜ λ. For the ⅜ λ case, a 1⁄16 λ top loading (22.5°) will be optimum.

Even more important is the implication for phased vertical arrays. Using top loaded ¼ λ elements rather than simple ¼ λ elements can reduce losses dramatically over lossy ground systems, since R_r and the feed point impedances can be considerably lower in phased arrays. Similar optimization is possible for vertical lengths higher or shorter than ¼ λ. There is no advantage to top loading for verticals ½ λ high and higher. Armed with this idea, we can shorten modeling experiments if we know to place the current maximum at the center of the vertical's height.

The actual optimum placement of the current maximum is often a little above the midpoint of the vertical, because of loading effects, but the difference is usually insignificant.

Example: Practical Design of Top-Loaded ¼ λ Vertical Antennas

Many monopole vertical antennas use guy systems for support. Let us assume four guy lines connected to the top of a ¼ λ high vertical and spaced 90° apart around the vertical. From *EZNEC*, the optimum length of these four wires is 2⁄7 × ¼ = 0.071 λ. For example, an 80 meter CW vertical exactly 70 feet high would be optimized with four

Table 2
Comparison of a Simple ¼ λ Vertical Antenna With a Top Loaded ¼ λ Vertical Antenna

Antenna	*hp*	*he*	*dBi*	*Giso*	*Ae*	*Rr*
1/4λ simple	1/4 λ	0.32	5.17	3.29	0.262	36.8
1/4 λ with 1/8 λ top loading	1/4 λ	0.46	5.29	3.38	0.269	74.1

Integration of Currents in Vertical Antennas

Up to a ½ λ height, the integration of all current elements over the vertical length sum at the zero degree elevation angle. For heights above ½ λ opposite phase currents along the vertical will subtract. At higher elevation angles, these vectors will begin to reinforce, and thus create higher angle responses. We can approximate h_e for the ⅝ λ vertical by weighting the average current for the ½ λ case (multiplying by 4) and adding the additional current average for the ⅛ λ case *over the high voltage segment*. This gives us an average current value of (0.64 (4) + 0.375) / 5 = 0.587. The effective aperture, h_e, is then 0.587 × 1 λ (h_p) or 0.587 λ. Notice that h_e is actually shorter for the ⅝ λ versus the ½ λ height. This is because more of the length of the ⅝ λ antenna has high voltage and low current. Also notice that h_p is only 1 λ for the ⅝ λ vertical. The length contributed by the ground in longer verticals only extends out to the voltage maximum along the ground surface, in this case ⅜ λ.

20 foot loading wires. In practice, there is a very small advantage (about 0.01 dB) to running the top loading wires horizontally compared to pointing them downward at a 45° angle as part of a guy system.

The *EZNEC* drawing of this antenna is shown in Figure 1. Wire 1 is physically ¼ λ high, and the four loading wires are 0.071 λ long. These wires are electrically connected to the top of the vertical, and insulated from the rest of the guy wires. This antenna shows the characteristics of the optimized top loaded vertical in Table 2.

Multi-band optimized verticals can be modeled using band traps at optimum points either along the vertical and/or horizontal (loading) segments. For any given vertical height, inverted "L," "T," or more complex capacitive hats can be optimized using *EZNEC*. The modeler will notice that the low-angle gain becomes maximum when the current maximum(s) approach the element centers. This is the purpose of the "loading wires" in the multi-band "Curtain-Zepp."[4] In this miniaturized curtain antenna, the capacitive loading on the end elements are close to the ground rather than up in the air.

Another important point is that traps need not always be placed to form resonant lengths at different bands. Traps may also be used to set electrical lengths (and thus effective heights) for maximum gain, such as in creating multi-band extended double Zepp antennas (see Note 4), or optimum current distribution along a vertical length and/or in the top loading structure. For example, I have built multi-band verticals for the higher bands by placing traps to create ⅝ λ heights at the various bands. This uses the physical height to much greater effectiveness (higher gain *and* efficiency). The disadvantage, of course, is the need for a tuner.

Optimization should be done using good antenna modeling software, but optimization can proceed much quicker if you know *a priori* that you want the current maximum at or near the middle of the vertical element. I have found that top loading wires should be kept shorter than about ⅛ λ to minimize horizontal radiation from the "hat." We can increase the capacitive loading by increasing the number of wires and/or adding a series inductor at the very top of the vertical structure. Here again is an example where a trap can be used to limit the length of a top loading wire at a higher frequency and the additional length at lower frequencies automatically optimizes the current distribution. Again, using traps in the conventional multi-band vertical has the sole advantage of normalizing feed point impedances for the various bands of interest, but "wastes" the additional height (and therefore gain and efficiency) above the traps at the higher frequencies.

This discussion only treats the "real" part of impedances and losses. It is assumed that any reactive components are tuned out in the matching circuit. Also, the intent of this paper is to quantify radiation resistance and thus the feed point impedances; antenna matching and related issues are only mentioned when necessary. Of course it is necessary to provide efficient well designed matching units and becomes increasingly critical as the vertical is made shorter. These topics, however, are exhaustively treated in many publications.

Conclusions

1) A simple general equation is presented for calculation of R_r.

2) The derivation of the commonly published R_r equation for very short verticals is derived from the general case.

3) Everything in the near field must be considered part of the antenna.

4) The physical characteristics of the near field — including the "intentional" antenna element(s) — will determine the directivity, efficiency and gain of the antenna. This includes Radiation and loss resistances.

5) The physical characteristics of the far field will affect the gain and directivity by effecting propagation losses. The effects of the near and far fields add.

6) Antenna Radiation Resistance (R_r) is independent of feed point impedance.

7) The efficiency of any fixed height linear vertical antenna can be optimized by orienting the maximum current point at the halfway point along the vertical element. This holds for not only very short antenna heights, but also for any vertical antenna height less than ½ λ. The placement of this current maximum is accomplished by careful design of the top loading element(s).

8) Radiation Resistance is a function of the physical height of a vertical antenna and the current distribution along that linear height, nothing else.

9) Series and parallel resistive losses are present in practical vertical antennas, however the series loss is usually more severe.

10) Radiation resistance (R_r) and effective height (h_e) are independent of loss values commonly found in amateur vertical antenna installations.

11) Lowering ground resistance and raising radiation resistance will result in higher overall efficiency, but often the latter is much easier to accomplish for a similar improvement.

12) Use of *EZNEC* and other antenna modeling tools can provide a short-cut to very complex analytical problems, including calculation of radiation resistance.

Notes

[1]Rudy Severns, N6LF, "Experimental Determination of Ground System Performance for HF Verticals — Part 1," *QEX*, Jan/Feb 2009, pp 21 – 25.

[2]John Krauss, *Antennas,* 2nd edition, McGraw Hill, 1988.

[3]Robert J. Zavrel Jr, W7SX, "How Antenna Aperture Relates to Gain and Directivity," *QEX,* May/Jun 2004, pp 35 – 38.

[4]Robert J. Zavrel Jr, W7SX, "*The Curtain-Zepp*" to be published, *QST*.

Bob Zavrel, W7SX, was first licensed in 1966 as WN9RAT and then WA9RAT. He has been W7SX since 1977. His primary interest in Amateur Radio is low band DXing, designing and building antennas, tuners and amplifiers. He is particularly interested in "maximum performance for the least money." His current home is on top of a ridge in the Oregon Coast Range Mountains. He uses 140 foot Douglas Fir trees as masts to support a wire antenna farm on the 3 acre property. An ARRL Life Member, Bob holds 5BDXCC, 5BWAZ (200), has 334 mixed, 324 CW, 99 on 160 m, 210 on 80 m and 299 entities confirmed on 40 meters.

Bob has a BS in Physics from the University of Oregon and has worked in RF engineering for 30 years. He has 5 patents, and has published over 50 papers in professional and amateur publications, including the first block diagram of an SDR receiver in 1987. He was involved with the first generation of RF integrated circuits for cellular phones, and worked extensively with DDS, WLAN, and passive mixer development. Bob currently works as an independent RF engineering consultant.

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Tech Notes

On Elevated Radials

I was pleased to read the series of articles by Rudy Severns, N6LF, on elevated radials for HF vertical antennas, published in the Jan/Feb, Mar/Apr and May/Jun 2009 issues of *QEX*. I had earlier written three articles on the same subject, published in the Winter and Spring 1998, and the Spring 1999 issues of *Communications Quarterly*. In particular, comparing what Rudy has written in his Mar/Apr, 2009 article with the Winter 1998 article I wrote, a number of the figures are very similar. And, we both referred to an article by Dick Weber, K5IU, published in the Spring 1997 issue of *Communications Quarterly*.

In fact it was the article by K5IU that prompted the paper I wrote, published in the Winter 1998 issue of *Communications Quarterly*. See in particular Appendix 1 of that paper, in which I have rigorously numerically modeled the K5IU antenna system. His elevated radial wire antenna was suspended by wires attached to his 140 foot metal tower, which I assumed supported a Yagi antenna. All wires, the support wires, his elevated radial wire antenna, the tower, and a 20 m Yagi antenna were included in my numerical analysis. All conductors carry induced currents, which K5IU did not consider, and it was (in my view) the antenna in its operating environment that resulted in the directivity that he observed. I have written many papers on the performance of antennas in their operating environment.

Let me continue on the topic under discussion. Most of what I have written on elevated radials was done in collaboration with MF broadcast consultants, since MF broadcasters have lots and lots of measured data — I was in that time period concerned with validating NEC.[1] This overview paper also compared theory with measured performance for a vertical monopole with elevated radials, and we (Steven White and I) measured not only impedance and ground wave field strengths, but the field strengths for the E and H fields beneath the elevated radial wires. This is a matter of concern for MF broadcasters, since many stations employ 50 kW of power. My Winter 1998 paper in *Communications Quarterly* (referenced above) compared prediction and measurement for a 2-element MF Phased Array. Again, the result was a good comparison between measured and predicted performance. Broadcasters in that time period were very skeptical that 3 or 4 elevated radials could produce performance equivalent to 120 buried radial wires! Perhaps they still are!

Continuing, for the interest of the radio amateur, let me refer to Figure 5 in the N6LF paper (Mar/Apr 2009 issue of *QEX*). See my Figure 1. This is an antenna system I devised for military tactical communications in the early 1980s.[2] The antenna system was intended to be tree supported, and the antenna wire was wound on bobbins, so the lengths of the vertical and elevated radial wires could be changed for resonance, since frequency agility was a requirement. This antenna system has to be fed by a balun — to isolate it from a transmitter ground connection. The advantage of such as an antenna is that it is a simple wire antenna, and if the radial wire is not too close to ground, the antenna radiates a mix of vertical and horizontal polarization, as required for both distant and not so distant communications. See Figure 2, which shows the computed radiation patterns for 3.8 MHz, with an elevated radial wire height of 4.6 m.

The antenna wire wound on bobbins, for resonating the wires, was home constructed. The antenna wire was AWG no. 14 or no. 16 Teflon covered, for protection of the stranded wire, which might be dragged over rough ground, and for extra strength.

Rudy, in reference to my studies (my *Communications Quarterly* Winter 1998 article), has wondered why the radial length should be resonant, particularly important when the elevated radial is close to ground. This is not surprising for several reasons: "resonant" because this maximizes current on the radial and hence the interaction of the antenna with the ground — recall that a vertical monopole antenna is a ground plane antenna; a resonant length introduces no reactance at the feed point; and for low heights above the ground the ground

[1]Notes appear on page 35.

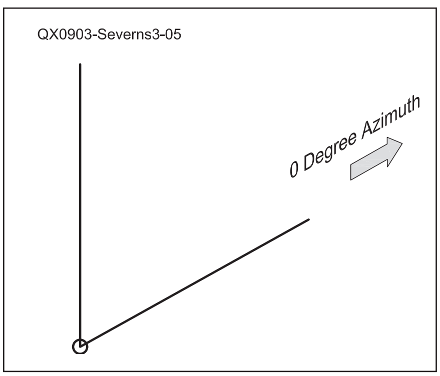

Figure 5 from Severns, Mar/Apr 2009 *QEX*, reproduced here for our readers' convenience — One elevated radial, 48 inches above 0.015/30 soil.

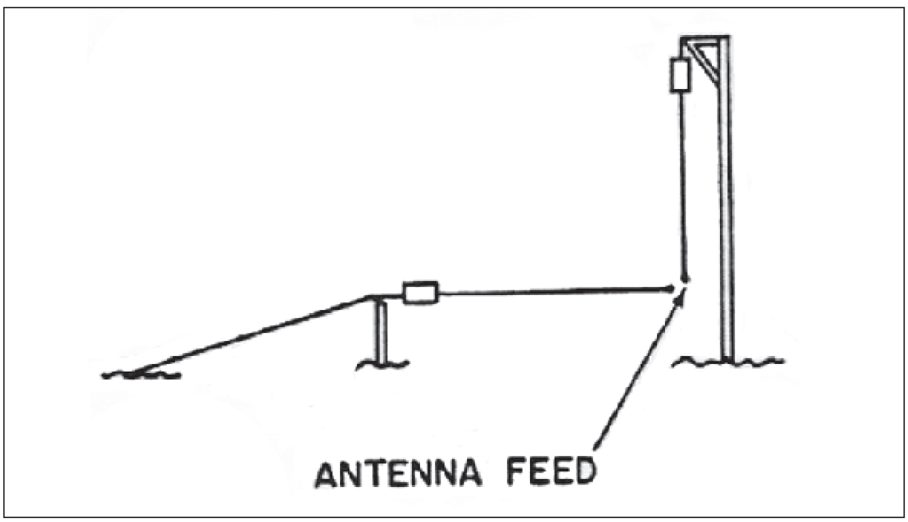

Figure 1 — A vertical monopole with 1 radial (dubbed a *VE2CV FD Special*). Performance details were published in AGARD Lecture Series 165, October 1989 (see Note 2). The rectangular boxes are bobbins, with extra wire for length adjustment.

has a dominant influence on the resonant length of the radial (significantly shortening its length for resonance).

To further clarify my line of thought, recall that the dominant field radiated by a horizontal "resonant" dipole at low height above ground is the vertically polarized field off the ends of the dipole. The resonant radial is no different.

In fact, a horizontal dipole at a low height above ground is a rather (perhaps surprisingly) efficient LF antenna system — radio amateur experimenters at 137 kHz and 500 kHz take heed.

If a near vertical incidence skywave (NVIS) link is a requirement, the antenna of Figure 1 can be reconfigured as a drooping dipole, so indeed, a 2-arm wire antenna, in which the arm lengths can be changed, is indeed a very versatile antenna, for tactical communications, emergency communications, and for trail-and-remote radio communications (a requirement for communications from hunting and fishing camps used by Indian and Inuit people in the north).

The antenna of Figure 1 is an antenna system that I have used for Field Day operation. It has a directive radiation pattern that is just what is needed for a Field Day station located to the north of the USA. In the time period I was using this antenna (1980s), I was in correspondence with Al Christman, K3LC, who has also written on elevated radials, and Al, in one of his papers, dubbed it a ***VE2CV Field Day Special!***

Elevated radials is a topic I have written and re-written about, in particular using elevated radial wires with "half" diamond loop antennas. "Half" because when the antenna system is used on the 160 m, 80 m and 40 m bands, the antenna is rather electrically close to the ground beneath it, and the other half of the loop is its virtual image in the ground plane. Interested readers can see, for example, Figure 13 in my Spring 1998 *Communications Quarterly* article, which shows the radiation patterns for an 80 m half diamond loop system, for one loop, and for two in-plane loops with a common feed point, which is in fact a phased array. It is a phased array with no problem in getting the currents right (phase and amplitude) on the 2-elements of the antenna system, since there is only one feed-point. And, the directivity of the antenna system can be easily changed, from unidirectional (elevated radials pointing in the desired directions), to bidirectional (elevated radials pointing in the two directions, forward and backward) — in fact with switching relays, one could change the radiation pattern from forward, to backward, to bi-directional.

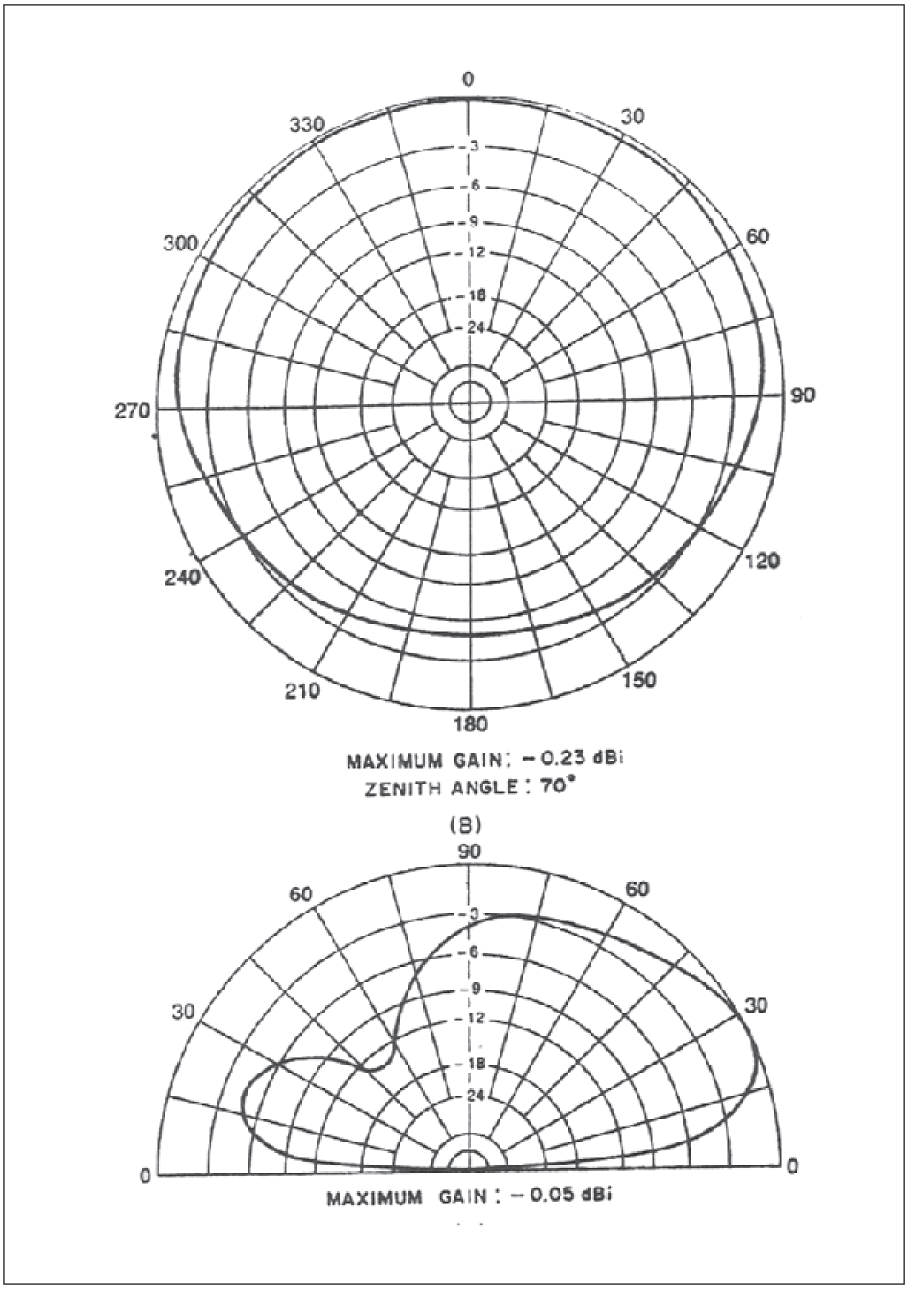

Figure 2 — The predicted azimuthal and elevation radiation patterns for a quarter wave monopole with one resonant radial, frequency 3.8 MHz. Note: this is the space-wave radiation pattern. The field strength at 0° elevation angle (not shown) is not zero. The signal on the horizon is the ground-wave. The antenna has a gain equivalent to120 buried radial wires, but only in the direction of maximum gain.

Notes

[1]John S. Belrose and Steven White, "On Validation of NEC-MoM: A Useful Tool for MF Antenna Design," IEEE Antennas and Propagation Symposium, Monterey, CA, 20-26 June 2004.

[2]John S. Belrose, G. M. Royer and L. E. Petrie, "HF Wire Antennas over Real Ground: Computer Simulation and Measurement," *Modern Antenna Design and Measurement: Application to Problems of Military Interest,* AGARD Lecture Series No. 165, October 1989, pp 6-1 to 6-30.

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SDR: Simplified

SDR: Simplified

We are going to talk a lot about classic amplitude modulation in this installment and build a radio that will tune one AM broadcast band station. Our first step is to look at how sampling is exactly the same as a general purpose amplitude modulator and how that lets us "cheat" Nyquist by sub sampling. Figure 1 shows the block diagrams of two receivers. One is a superheterodyne with IF filtering and the other is a basic "crystal set" done in digital signal processing (DSP).

Figure 7 shows a schematic for the digital to analog converter (DAC) output board that we need to get audio out of our system. It uses an inexpensive 8 bit DAC in a dual in-line pin (DIP) package, so construction is inexpensive and easy.

Hardware Update

There must be a lot of interest in this column, because we keep depleting the stock of items. First we depleted DigiKey and now Analog Devices. The last report I had from readers was that the AD7476-DBRD for the 12 bit analog to digital converter (ADC) is at "lifetime buy" status. I have contacted Analog Devices to ask that they continue to supply the board, but have not heard back. If there is still a need, we can duplicate the circuit board and make a group order. Depending upon quantities needed the empty boards can probably be as cheap as $4 or $5 each if someone with a University address will volunteer to serve as intermediary. Contact me about your interest in a circuit board.

An RF Explanation for Sampling

A classic AM modulator is composed of a generator that has zero average voltage, with both positive and negative values. That generator is multiplied by a signal that has a constant positive average (dc) and a varying modulating signal. The simplest form is a sine wave carrier modulated by a sine wave information signal. This is the equation for this modulator:

$V_m = 1 \sin(x)$

for the modulating information, and:

$V_c = 1 \sin(y)$

for the carrier signal.

$$V_{out} = 1 \sin(y) \times (1 + 1 \sin(x)) = 1 \sin(y) + \tfrac{1}{2} \cos(x-y) - \tfrac{1}{2} \cos(x+y) \quad \text{[Eq 1]}$$

The result is the full carrier plus the two sidebands. Figure 2A shows the (time domain) waveform and Figure 2B is the (frequency domain) spectrum of the result. The important point here is that an amplitude modulator is implemented as a circuit that multiplies two signals. In DSP,

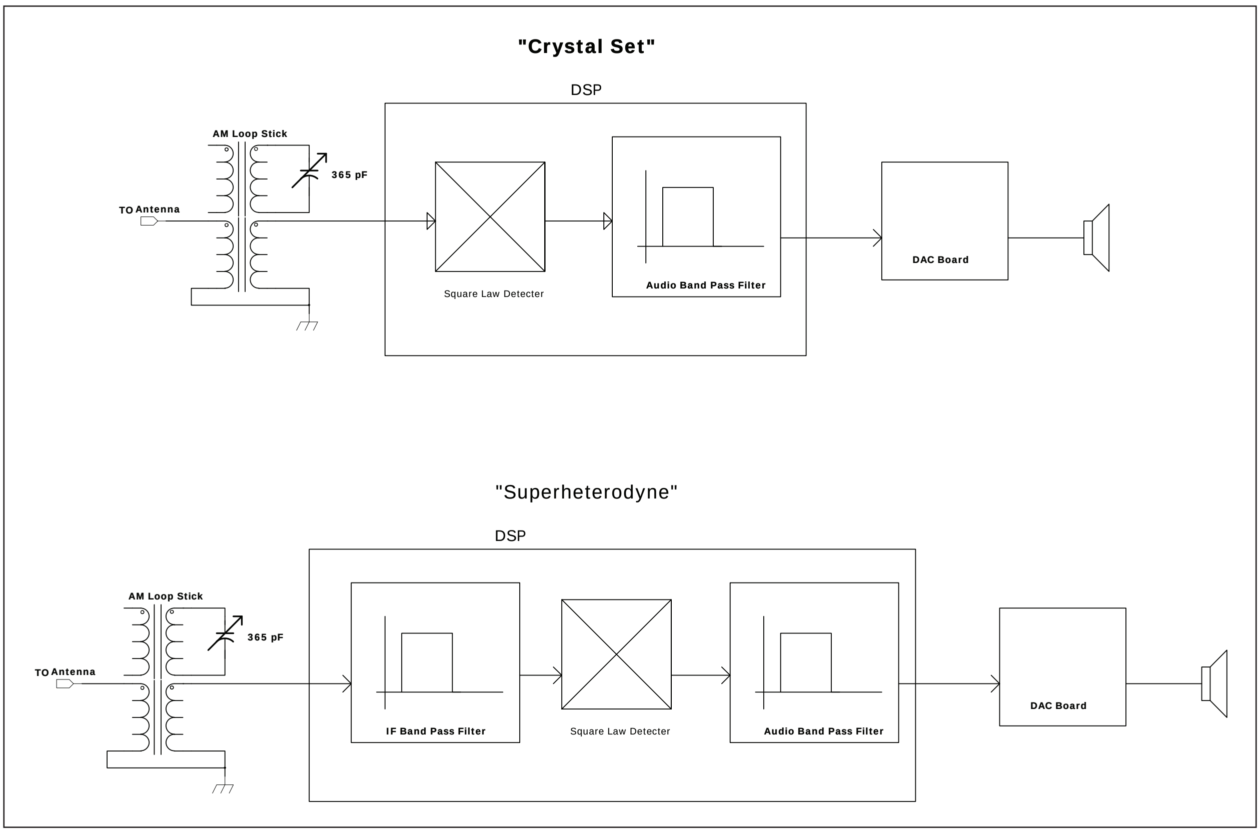

Figure 1 — Part A is the block diagram of a DSP "crystal set." The germanium or point contact diode is replaced with the DSP block. Part B is the block diagram of a DSP super heterodyne receiver. The DSP performs the mixing, IF filtering, demodulation, and baseband filtering.

we can do that multiplication using the computer.

Instead of adding one to our modulating signal, let's add one to our carrier frequency. The new set of equations becomes:

Vout = 1 sin(x) × (1 + 1 sin(y)) = 1 sin(x) + ½ cos(x–y) – ½ cos(x+y) [Eq 2]

This result is similar but not identical. Let's put real numbers to the equations. In Equation 1, let the information be 5 kHz and the carrier be 100 kHz. The resulting output frequencies are 100 kHz, 95 kHz, and 105 kHz. In Equation 2, we still have 95 kHz and 105 kHz, but the third signal is our original 5 kHz information signal. We actually have a double sideband suppressed carrier signal, plus our original information. Figures 2C and 2D show the waveform and spectrum.

Instead of using a sine wave carrier, let's use a square wave carrier. We know from the Fourier series that a square wave is composed of the fundamental and all the odd harmonics. When we multiply our sine wave information by the square wave, we get a series of sidebands around each of the harmonics. Figure 3 shows the modulated waveform and the spectrum of the result. (*Gnuplot* doesn't have a square wave function, so Figure 3 uses a Fourier series to approximate the square wave.) This operation is very similar to what we do in sampling for our DSP. The difference is that instead of using a square wave, we use the equivalent of a 100 kHz pulse signal, with very narrow pulses.

Now, we can take the next step. I hope you are familiar with a double balanced mixer being used as either a modulator or a frequency mixer. The equations we looked at above are just as relevant for the frequency mixer case. Let's look at a CW signal at 105 kHz, and multiply it by our 100 kHz sine wave signal. In this case x is 105 kHz and y is 100 kHz, so we get frequencies of 5 kHz, 105 kHz, and 205 kHz. If we make the oscillator a square wave instead of a sine wave, we again have signals spaced around the harmonics of the 100 kHz oscillator.

Let's look at what happens when we replace the square wave with a 100 kHz pulse signal that has a 1% duty cycle. Figure 4 shows that we see the envelope of a 5 kHz signal. Notice that the spectrum of the signal has components associated with all of the harmonics of the pulse signal. This is exactly what happens in our analog to digital converter. We get a sequence of numbers that correspond to "one" times the value of the incoming

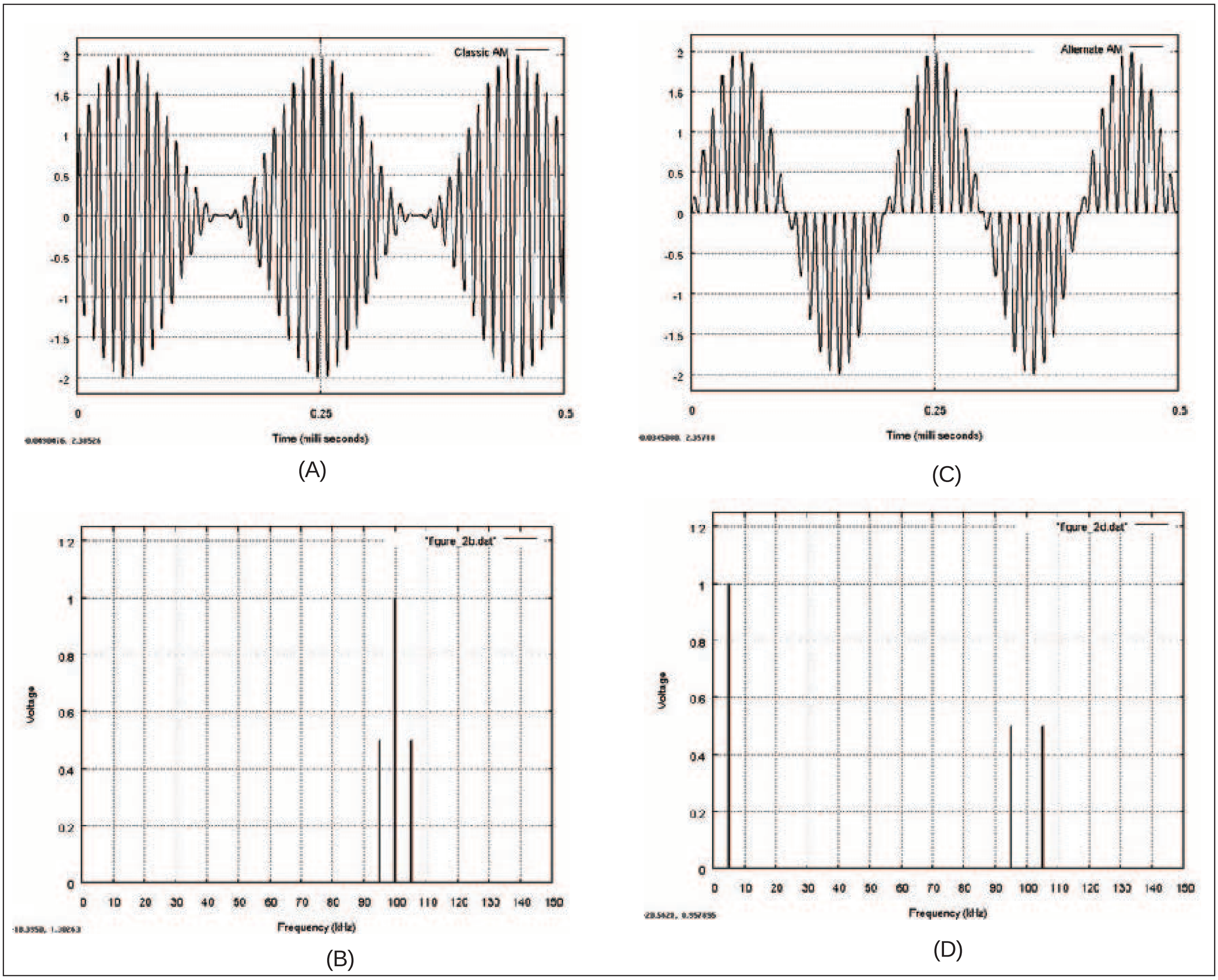

Figure 2 — Part A shows the waveform of classic amplitude modulation, where the envelope follows the modulating waveform. Part B shows the spectrum of the classic AM waveform, showing the original carrier and the two sidebands. Part C is a modification of AM, in which the carrier is offset by a dc value. Part D shows the spectrum of the modified AM signal, showing the original modulating waveform and the two sidebands.

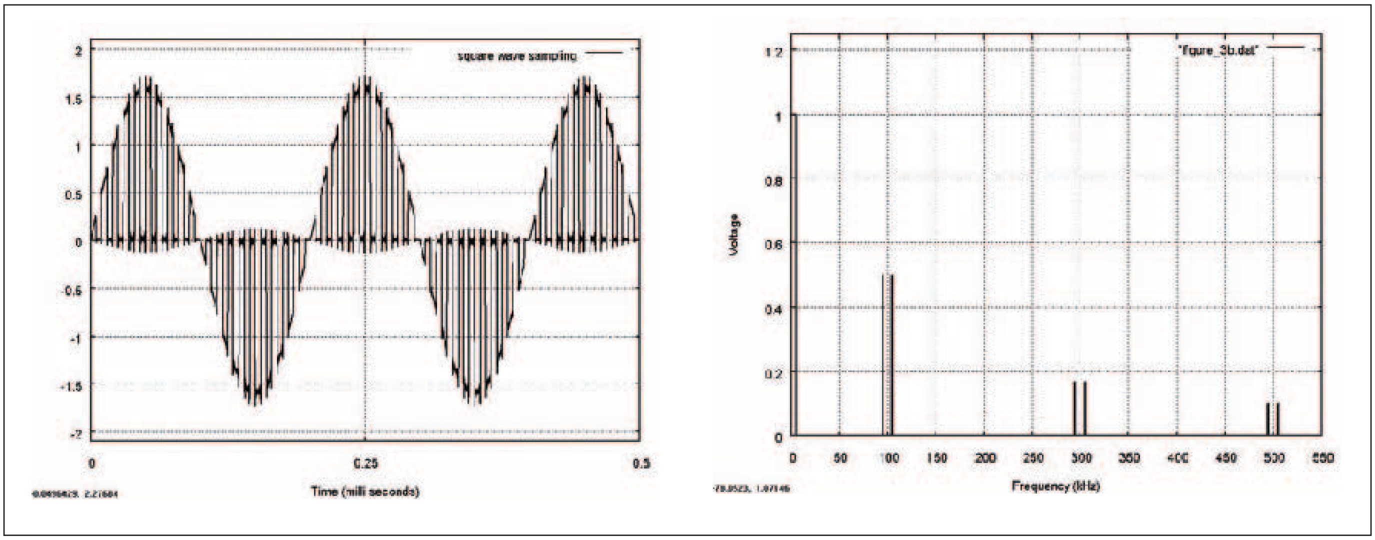

Figure 3 — Part A is an example of the modified amplitude modulator, in which the carrier is a square wave. Part B shows the spectrum of the square wave modified amplitude modulator.

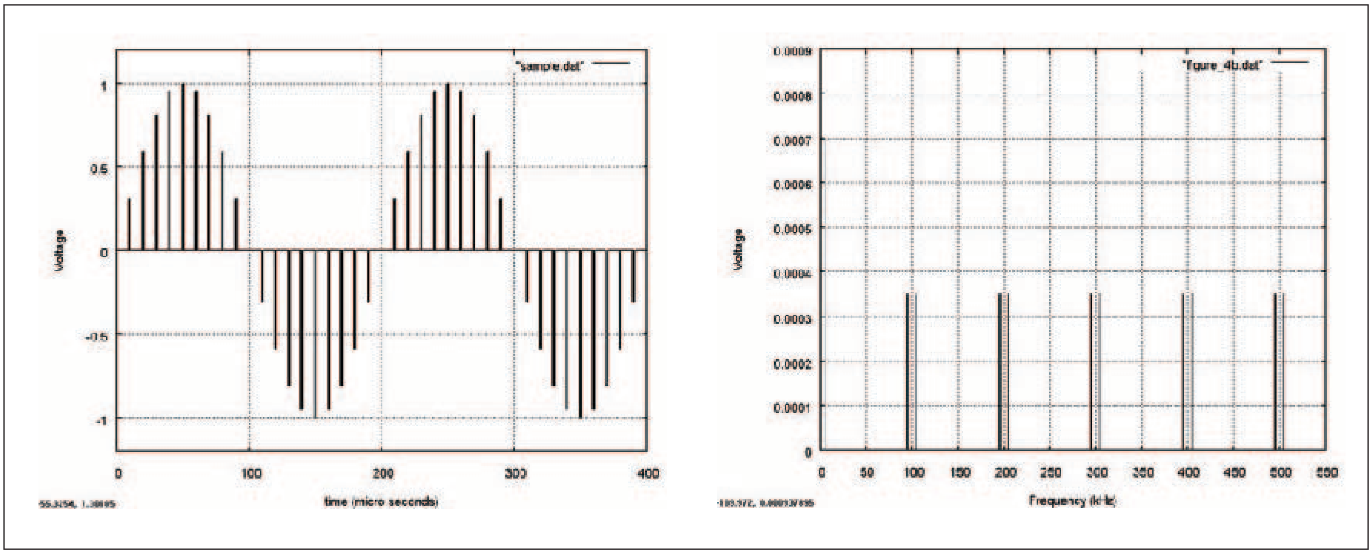

Figure 4 — The waveform from the modified amplitude modulator is shown in Part A, when an impulse is used as the carrier. This waveform is the basis for our sampling of the incoming signal. The height of each impulse corresponds to the sampled ADC value. Part B shows the spectrum of the sampled signal.

signal at the time of our sample. The effect we see where our 105 kHz signal (which is certainly above the Nyquist limit) translated down to a 5 kHz sampled signal is called "aliasing" by DSP folks. However, we have seen that aliasing is just a mathematical way to describe what we know in the RF world as mixing. The DSP world also refers to the intentional use of aliasing as "subsampling." This property will be very useful for our "DSP crystal set."

I am going to design a radio to receive the strongest signal on the AM Broadcast band here in Austin, which is KLBJ on 590 kHz. In order to reduce the burden on our system we will filter the incoming signal with a single tuned circuit just as you would in a crystal set. We will set the tuning to peak at 590 kHz, and sample at 500 kHz. Just as in a superheterodyne receiver, we have to worry about IF leakage at 90 kHz and image rejection at 410 kHz. This means we want to keep the Q of our single tuned circuit pretty high. We can build two different Software Defined Radios at this point. We can implement a true superheterodyne and further filter the signal at 90 kHz before doing the detection. The option we will use is to send the signal directly from the input tuned circuit to the detector, as in a crystal set.

The IF or Base Band filter

There don't seem to be many books that are easy to understand with respect to how one designs a digital filter. Probably the best I have seen is Chapter 10 of ***Experimental Methods in RF Design***, published by ARRL.[1] I have the 2004 first edition, but the authors have completely updated the text, so I am sure the new revised first edition is even better. We will take a very detailed look at filters in later columns, but for now, we will look at the most common ones for communications work.

It turns out that a digital low pass filter, a digital high pass filter and a digital band pass filter all have exactly the same shape: two rectangles. This is true because in the digital world we have a bounded system. In the case of our 500 kHz sampled system, our view of the universe begins at exactly –250 kHz and goes up to exactly +250 kHz (the Nyquist region). In our digital view, no other frequencies exist. So, Figure 5 shows

[1]Notes appear on page 41.

what the three filter types look like. It is obvious that the high pass filter and band pass filters have two rectangular regions with some amount of space between them. The low pass filter is actually two rectangular areas, too, but there is zero space between them.

We describe filters by their properties in the frequency domain. That is, we plot amplitude versus frequency. We implement filters, however, by describing their operation in the time domain. We take the incoming samples that are equally spaced in time, operate on them, and then send the modified equally spaced samples to an output device.

It turns out that you can filter just about any kind of digital data and, we RF guys are not the only ones that do it. A good example of a digital filter that isn't electronic is used by financial people to analyze the stock market. A 50 day moving average of stock prices takes the closing price of a stock for the last 50 trading days (our digital samples), adds them all together, and then divides them by 50. Now on the next day, the average throws away the price from 51 days ago and adds the 50 last samples together and divides by 50. This process repeats each new trading day. Note that we could just as easily multiply each price by 1/50 and then added all of the scaled prices one by one to get the same effect. So in this case our sample rate is once per day, our filter length is 50, and the coefficient for each sample is 1/50. This actually creates a very rudimentary low pass filter.

When we filter electronic signals, we do exactly the same set of operations. First we decide how big to make the filter. I am selecting a filter size of 50 (a number admittedly pulled out of the air). When we run the filter, we will multiply each of the last 50 samples from our ADC by its associated coefficient and add it to the result of all the previous multiplications and additions. This is called a multiply accumulate operation. We will then repeat the process after dropping the oldest sample and adding the newest sample and continue forever (or until we turn off the power).

A moving average is usually a useless electronic filter. Useful filters have coefficients that are different for each sample. We really want a band pass filter that has a response of "one" inside the pass band and "zero" in the stop band. In the DSP world, we call this a brick wall filter because the sides are straight up and down just like a brick wall. We can come pretty close to real brick wall filters in DSP where such a filter in the analog world is extremely difficult to produce.

We talked last column about the Fourier transform. It is a mathematical tool that lets us look at time varying data in the frequency domain. The Inverse Fourier Transform lets us convert data in the frequency domain into a time varying signal in the time domain. Once again the math will make your eyes glaze over, but the

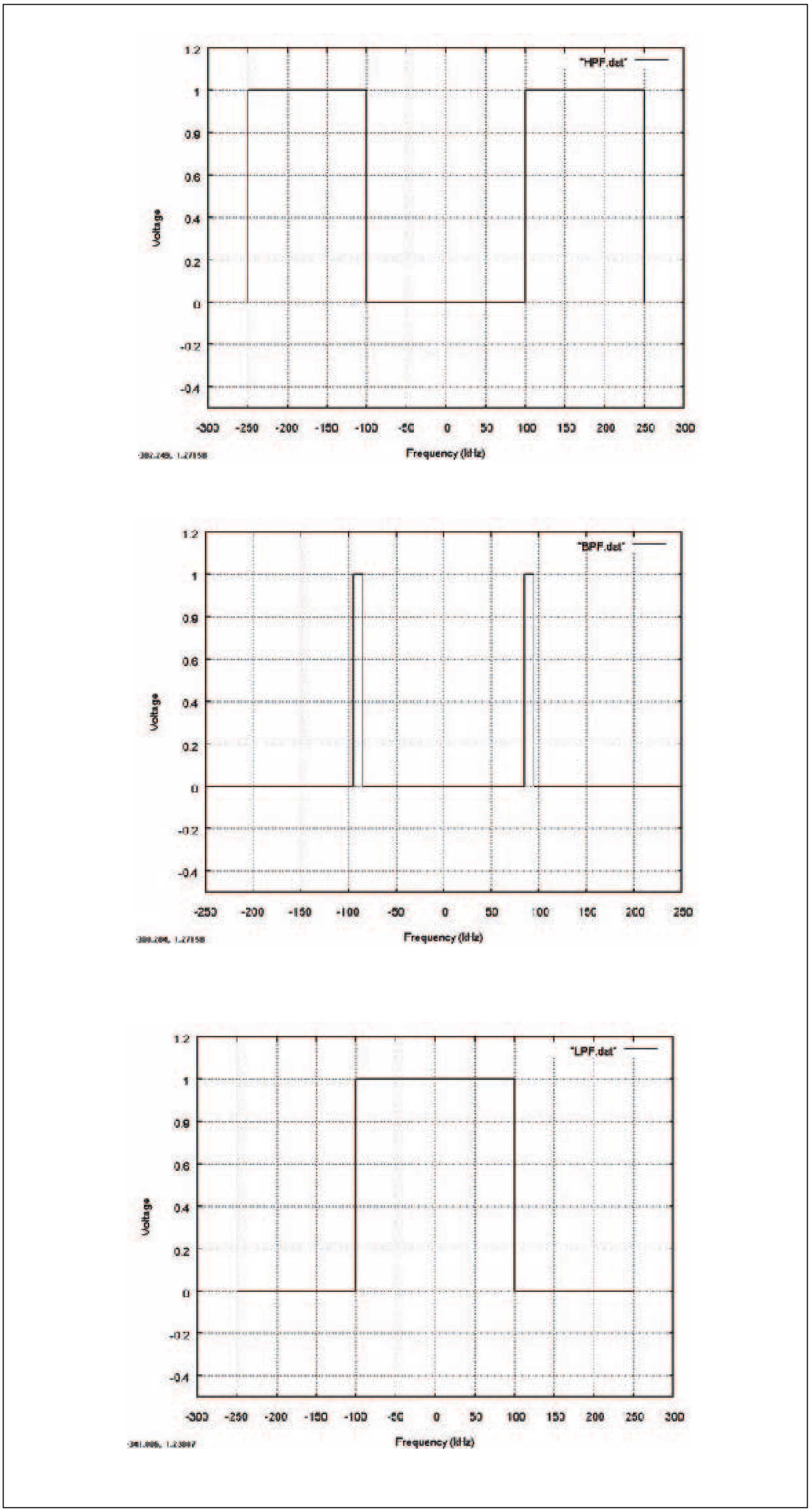

Figure 5 — The spectrum of a digital high pass filter is shown in Part A. The spectrum of a digital bandpass filter is given in Part B. Part C shows the spectrum of a digital low pass filter. Notice that all three spectra contain two rectangles with space between them. Part C does not appear to have space, but the size of the "logical" space is exactly zero. Also notice that the graph does not extend beyond –250 kHz or +250 kHz.

Fourier Transform and Inverse Fourier Transform are exactly the same thing! The difference is the data rather than the transform itself. We use this property to convert the process of filtering which occurs in the frequency domain into a time domain operation.

Let's look at how this works. The night time spectrum you might see in Austin is shown in Figure 6A. It shows WFLF 540 kHz in Orlando, KLBJ 590 kHz in Austin, KMJ 580 kHz in Fresno and KFI 640 kHz in Los Angeles. Figure 6B shows the filter response. The operation we want to perform in the frequency domain is to multiply our filter response times the input spectrum. The output is the spectrum shown in Figure 6C.

We can perform the exact same operation described above in the time domain by taking the time domain response of our filter and multiplying it by the time domain input signals. We need to somehow convert the frequency domain representation of our filter into its corresponding time domain representation. That's where the Inverse Fourier transform comes in. Very short pulses in one domain will give us the response in the other domain when we do a Fourier transform. This is called the impulse response.

Fortunately, we do not have to calculate the inverse transform. It has already been done for us for rectangular shaped filters. Equation 3 is an expression of the Fourier series that comes from a discrete Fourier transform of our rectangular filters. Each C_k is one of the coefficients in the filter. We are going to have a filter with 50 coefficients, but we will only calculate 25. The reason is that the filter response is symmetric about time zero. Remember from last column that negative time corresponds to negative frequency. There is a BASIC program that comes with *Experimental Methods in RF Design* that will calculate the 25 coefficients. That program, FIRDSN3.BAS, is available for download in the *QEX* files section of the ARRL Web site. Thanks to Bob Larkin, W7PUA, for writing that program and for making it available for download by *QEX* readers.[2]

$$C_k = 1/\pi k \ ((\sin ((2 \pi k f_H)/f_S)) - (\sin ((2 \pi k f_L)/f_S))) \qquad \text{[Eq 3]}$$

f_H is the upper cutoff frequency
f_L is the lower cutoff frequency
f_S is the sample frequency

The Detector

We need to convert the full carrier double sideband signal into the baseband information. One of the classic AM demodulation methods is a square law detector. Very few devices have a true square law response over a large range of signals. The JFET comes close, but has limited range. With DSP we can implement a square law detector with range that is

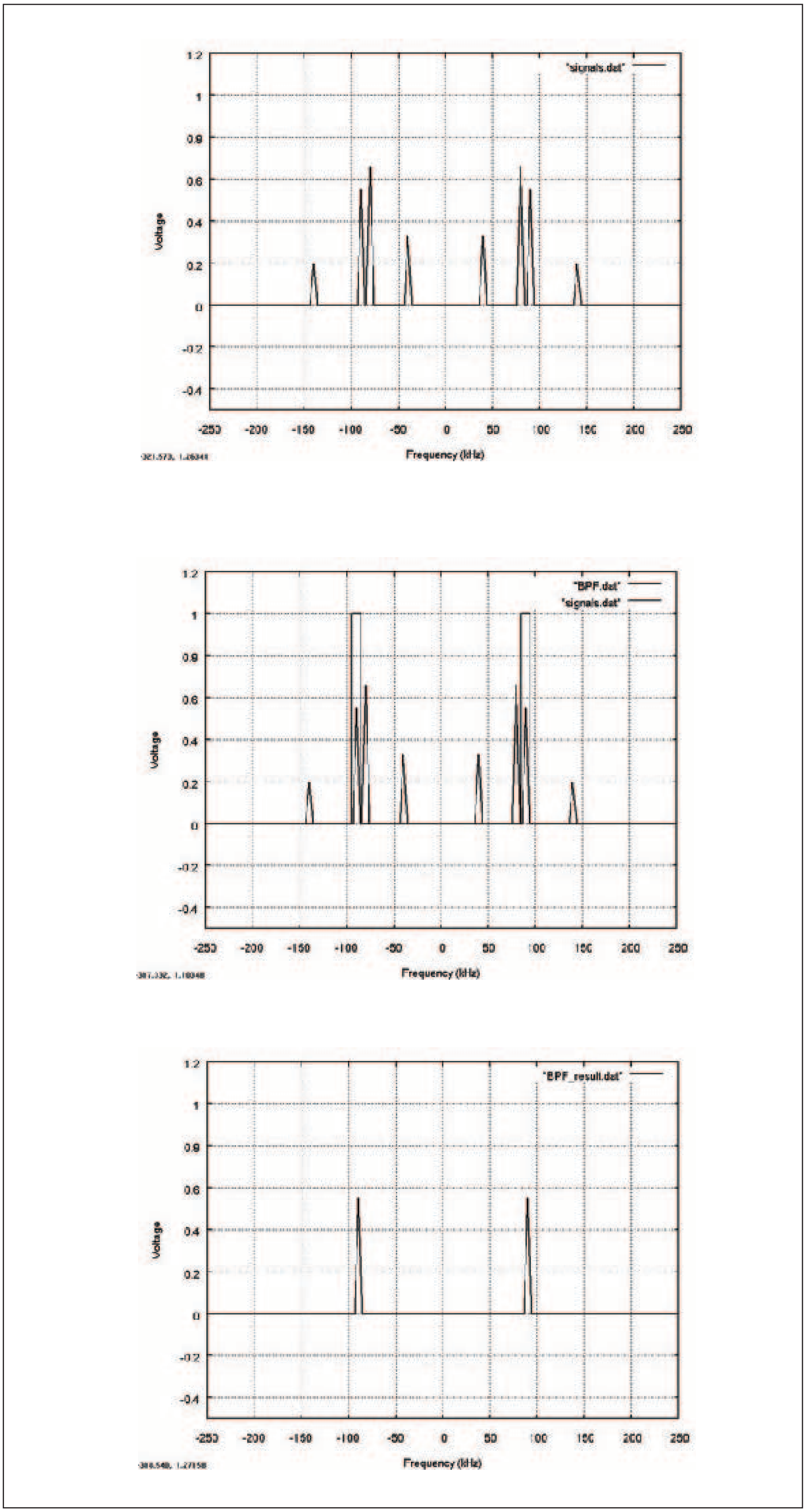

Figure 6 — The spectrum of a hypothetical AM receiver at night (with the low-frequency end clear channel signals) is shown at A. Part B shows the band pass filter response overlayed with the spectrum. In Part C, we see the result of multiplying the band pass response and the spectrum.

limited only by the size of the data that the DSP can handle. If we use 16 bit data for the input, our multiplier must have a 32 bit result. This is a very simple and fast operation for the DSP. Equation 4 shows how and why a square law detector works to retrieve the original modulation.

$$V_{out} = (1\sin(y) + \tfrac{1}{2}\cos(x-y) - \tfrac{1}{2}\cos(x+y))^2$$

$$V_{out} = \sin(y)^2 + \sin(y)\cos(x-y) - \sin(y)\cos(x+y) + \tfrac{1}{4}\cos(x+y)^2 + \tfrac{1}{4}\cos(x-y)^2$$

$$V_{out} = \tfrac{3}{4} - \tfrac{1}{2}\sin(2y) + \tfrac{1}{2}\sin(x) + \tfrac{1}{2}\sin(2y-x) - \tfrac{1}{2}\sin(-x) - \tfrac{1}{2}\sin(2y+x) + \tfrac{1}{8}\cos(2y+2x) + \tfrac{1}{8}\cos(2y-2x)$$

$$V_{out} = \tfrac{3}{4} + 1\sin(x) - \tfrac{1}{2}\sin(2y) + \tfrac{1}{2}\sin(2y-x) - \tfrac{1}{2}\sin(2y+x) + \tfrac{1}{8}\cos(2y+2x) + \tfrac{1}{8}\cos(2y-2x) \qquad \text{[Eq 4]}$$

This result will have 32 bits, but our filter math should only be 16 bits. We can divide the result by 65536 to get the signal back into range for further operations. In software, this just means we throw away the bottom 16 bits of data before we store the result.

The signal we want is sin(x), but we have a significant dc component and a whole bunch of junk at twice the carrier frequency. The only real problem is that dc component, since it is really close to the desired audio signal. We need to use a band pass filter to remove the dc and the RF components. Since all but the dc are far removed from the desired audio, we can use a relatively simple band pass filter with only 10 coefficients.

The Output

The last step is to take our detected signal and send it to the real world again. The SPI system is tied up doing the input of our signal, so we can hook a parallel DAC to one of the parallel ports. All we have to do is take every sample as it comes out of the audio band pass filter and send it directly to our 8 bit DAC. The fidelity won't be great, but it will be okay for talk radio.

The Software

We can do all of the software for this first project by writing a user level program in *C* to run on the Stamp with the AD7476-DBRD and an extra DAC board for output.

We can use the device drivers that the Stamp development team wrote for the SPI system to handle input. We just do a normal read of an I/O device to get our data. Then we can square the data and filter it before writing the data directly to our output DAC. See the sidebar for the *C* program listing. The top level design will look like the simplified code in the sidebar. The actual program is stored in the *QEX* section of the ARRL Web site.[3]

We open the SPI as if it were simply a file on a hard disk. The same is true for the parallel port used for the 8 bit output values. We read eight 12 bit words from the ADC and store them. The code squares each value and then scales it back to 16 bits. Next, we send all 8 values to the baseband filter to be cleaned up. Last, we write the 8 values to the parallel output "file" to go to a speaker. This process repeats until we end the program.

Next Issue

In the next issue, we will look again at more "theoretical" topics. The IF filter version of our radio very possibly won't work at the 500 kHz sample frequency, so we will look at the next "trick" that we can use to sample close to our incoming RF and still use a reasonable DSP IC.

Notes

[1]Wes Hayward, W7ZOI, Rick Campbell, KK7B, and Bob Larkin, W7PUA, *Experimental Methods in RF Design*, Revised First Edition, ARRL, 2009. ARRL Order No. 9239, $49.95. ARRL publications are available from your local ARRL dealer or from the ARRL Bookstore. Telephone toll free in the US 888-277-5289, or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

[2]You can download the program file FIRDSN3.BAS from the ARRL *QEX* Web site. Go to **www.arrl.org/qexfiles** and look for the file **7x09_Mack_SDR.zip**.

[3]The *C* program file is included in the file **7x09_Mack_SDR.zip** given in Note 2.

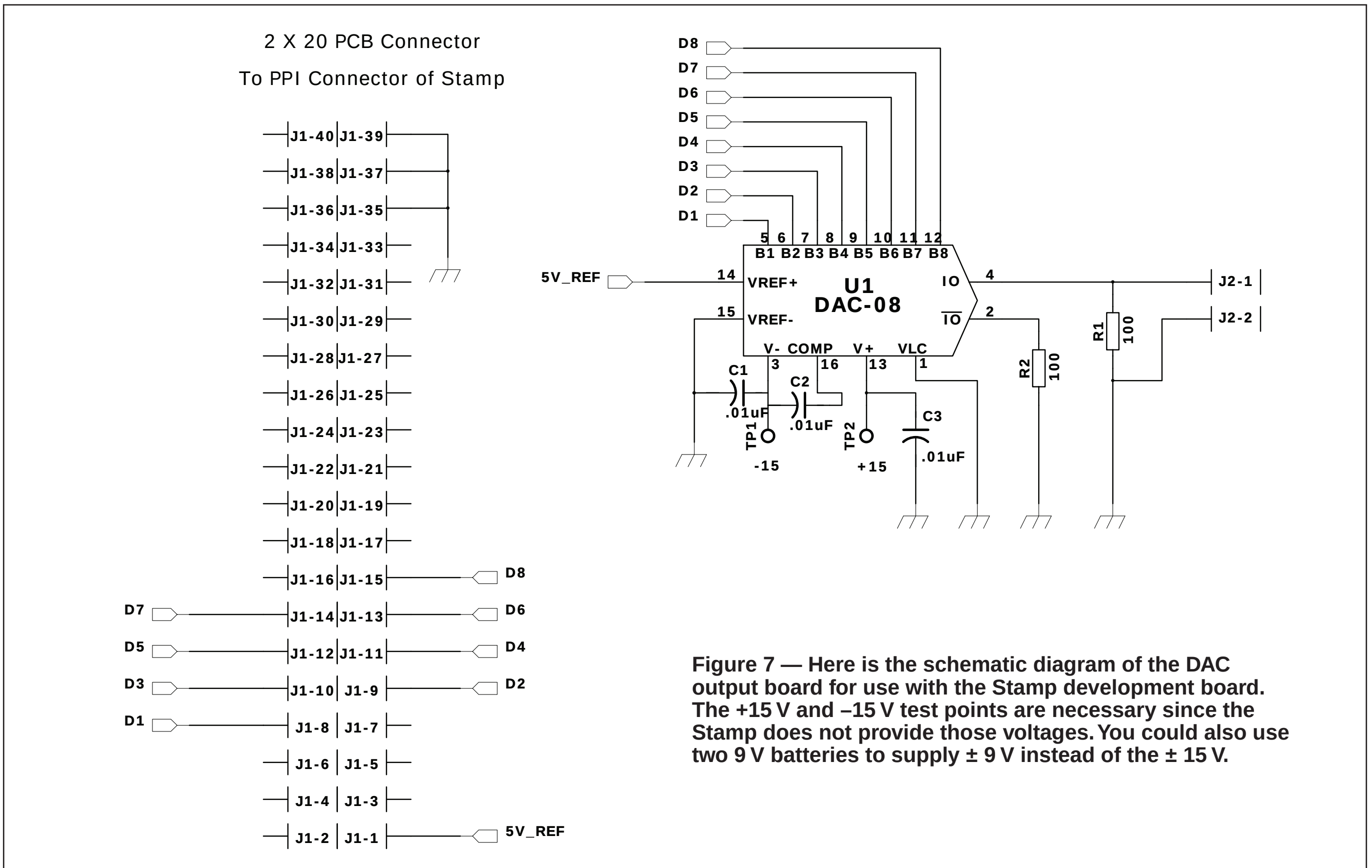

Figure 7 — Here is the schematic diagram of the DAC output board for use with the Stamp development board. The +15 V and –15 V test points are necessary since the Stamp does not provide those voltages. You could also use two 9 V batteries to supply ± 9 V instead of the ± 15 V.

C Program Listing

```
#include <io.sys>

// Note that this is a simplified version of the actual
// software.  You can get the real software from the ARRL
// Web site at www.arrl.org/qexfiles. Look for the file 7x09_Mack_SDR.zip.

int  input_descriptor;
int  output_descriptor;
long temp;
int  input_data[8];
int  filter_data[8];
char output_data[8];

Void main(void)
{
       // open the two IO files
       input_descriptor = open("/dev/spi", O_RDWR);
       output_descriptor= open("/dev/ppi", O_WR);

       // this loop runs forever until we kill the process
       while(1)
       {
          // SPI requires that you write before you read
          write(fd, input_data, 8);
        read(fd, input_data, 8);

           // (long) changes our 12 bit input data to 32 bit    // data.  Here we square the
           data and scale it back // to 16 bits before storing it in the filter array

          temp = (long) input_data[0] * (long) input_data[0];
          temp = temp / 65536;
          filter_data[0] = temp;
          temp = (long) input_data[1] * (long) input_data[1];
          temp = temp / 65536;
          filter_data[1] = temp;
          temp = (long) input_data[2] * (long) input_data[2];
          temp = temp / 65536;
          filter_data[2] = temp;
          temp = (long) input_data[3] * (long) input_data[3];
          temp = temp / 65536;
          filter_data[3] = temp;
          temp = (long) input_data[4] * (long) input_data[4];
          temp = temp / 65536;
          filter_data[4] = temp;
          temp = (long) input_data[5] * (long) input_data[5];
          temp = temp / 65536;
          filter_data[5] = temp;
          temp = (long) input_data[6] * (long) input_data[6];
          temp = temp / 65536;
          filter_data[6] = temp;
          temp = (long) input_data[7] * (long) input_data[7];
          temp = temp / 65536;

          filter_data[7] = temp;
          // we send the data to the filtering function here
          Band_pass_filter(filter_data, output_data);
          // output the data here
          write(fd, output_data, 8);
       }
}
```

QEX

Letters to the Editor

Exploring Near-End-Fed Wire Antennas (Mar/Apr 2009)

Dear Larry,

An error has been pointed out in my Near-End-Fed Wire Antennas article. On page 34, the first full paragraph on the page says, "The results shown in Figure 3, from modeling a 40 m NEF half wave dipole, are typical of those obtained for other bands. This model was 68 ft long at 40 ft elevation, with the feed point 10.5 ft from one end."

The distance from one end should be about 10.5% instead of 10.5 ft. This makes the position about 7.25 ft from the end.

This was my error, and I am sorry for the confusion this may have caused readers.

— 73, Ron Skelton, W6WO, 4221 Gull Cove Way, Capitola, CA 95010; **w6wo@k6bj.org**

Hi Ron,

Thank you for sending along that correction.

— 73, Larry Wolfgang, WR1B, QEX *Editor;* **lwolfgang@arrl.org**

A Versatile Two-Tone Audio Generator for SSB Testing (Mar/Apr 2009)

Hi Ken,

I read with great interest your article about a versatile two-tone audio generator in the Mar/Apr issue of *QEX*. You did a great job and the finished unit looks very professional. The article took me back 40 years, to when I did quite a bit of work on two-tone testing while building various SSB amplifiers.

The one potential problem I noticed upon studying your circuit was that the balance adjustment range is likely to be too low for many rigs. Your summing circuit has only about a ±1 dB range. It has been my experience that a ±5 dB range is often needed due to irregular frequency response of many rigs. In fact, if you run tones as low as 300 Hz or as high as 2700 Hz, an even bigger range might be needed due to filter roll-offs in various transmitters. I'd increase R10 to about 50 kΩ to provide more range on the balance control.

I question the need to have the frequencies of the two tones variable. It is my experience that for two-tone testing, frequencies of around 800 Hz and around 1800 Hz are satisfactory. Using fixed capacitors simplifies the circuit and makes it much more compact. Also, for those builders without access to variable capacitors (or those wishing to keep the size smaller), the frequency can also be varied using ganged potentiometers.

For those readers interested in building a two-tone test generator, I would like to remind them that there is a much simplified way to do it, which I described in the Aug 1971 issue of *QST* (pp 17-21). The basic idea was to power an audio oscillator with half-wave rectified ac and produce an 8.3 ms audio burst followed by 8.4 ms of silence. By carefully adjusting the feedback in the oscillator, a constant amplitude perfect sine wave is produced during the burst. Obviously, building two such oscillators will produce a burst two-tone generator with a 50% duty cycle, which is better suited to driving an amplifier to its peak without overheating it. Although my circuit of 40 years ago used 709 op-amps and 741s when they came out, I was curious to see if the same pulsed power supply would work with modern JFET op amps. To that end, I built the front end (U1 and U2) of your *QEX* circuit of Figure 6 on page 27, but substituted 270 Ω resistors for the lamps DS1 and DS2. I then adjusted R3 and R7 of each oscillator to get a constant amplitude audio burst at the output of the buffers U1B and U2B. It worked well except that there was a transient during the first and last half cycles of the burst. This was created by the turn on/off of the buffers, so looking at the waveform at the output of each oscillator (pin 1 of U1A and U2A) gave a cleaner waveform. The outputs of the op amps are of low enough impedance so that the buffers are not really needed. For the same reason, U3 is also not needed for a very simple circuit.

If the reader wants a very simple two-tone generator, I would recommend the following simplifications:

1. Use fixed capacitors to determine each frequency by eliminating C1A/B and C2A/B and increase C3, C4, C5 and C6 by about 100 pF.

2. Replace DS1 and DS2 with 270 Ω resistors.

3. Use a single TL082 for U1A and U2A. U1B, U2B, and U3 are all not needed.

4. Eliminate S2 so it is two-tone only.

5. Connect the summing network directly to the outputs of the oscillators. I would change R9 and R11 to 47 kΩ and make R10 a 100 kΩ potentiometer.

6. Connect the top end of R14 (Output Level) to the wiper of the balance pot and the wiper of R14 directly to the output.

7. Modify the power supply by eliminating all capacitors and regulators. Replace the bridge rectifier with oppositely poled diodes at each end of the winding and connect the

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3CX400U7	4CX250FG	4X150A	872A-SS
3CX800A7	4CX250R	YC-130	5867A
3CX1200A7	4CX350A	YU-106	5868
3CX1200D7	4CX350F	YU-108	6146B
3CX1200Z7	4CX400A	YU-148	7092
3CX1500A7	4CX800A	YU-157	3-500ZG
3CX2500A3	4CX1000A	572B	4-400A
3CX2500F3	4CX1500A	807	M328/TH328
3CX3000A7	4CX1500B	810	M338/TH338
3CX6000A7	4CX3000A	811A	M347/TH347
3CX10000A7	4CX3500A	812A	M382

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cathode of one diode to +12 V and the anode of the other diode to –12 V supply. A 10 or 12 V CT transformer will work as well as the 20 V one specified. Don't go over 20 V on the transformer, because the op amp specification is 15 V max.

8. Make a test point at the output of each oscillator so it is easy to connect a scope to adjust R3 and R7 for a constant amplitude burst.

That's it! The result is a very compact two-tone test oscillator.

There is a lot more I could go into about burst oscillators producing low-distortion sine waves. If there is reader interest in the methods of two-tone testing, I would be glad to write-up some of my observations and experiences over the years.

— 73, Bob Buus, W2OD, 8 Donner St, Holmdel, NJ 07733; **w2od@aol.com**

Hi Bob,

Thanks for your comments regarding my article.

As the title says, I wanted it to be a device whose primary purpose was to be SSB testing, but which was also versatile enough to be used elsewhere in the ham shack or on the experimenter's bench.

I had come by a pair of old vacuum-tube HP200-AB audio generators in very good condition, which would have sufficed, but they took up half of my bench. Anyway, they were the inspiration.

All of the literature and service information that I've seen calls for equal amplitude tones. Hence, the balance trimpot was designed to be set-and-forget. While the circuit was being developed, I came to realize the wisdom of using ganged variable capacitors rather than ganged potentiometers. The potentiometers (commercially-made and from a well-known manufacturer) didn't track well at all, causing the oscillator to drop out and / or change amplitude.

The lamp in the negative feedback loop of each oscillator is not operating as a lamp but rather as a positive temperature coefficient resistor, which keeps the oscillator amplitude very stable. The oscillator buffers may not be needed, but they're the other half of the IC package anyway and were included as a matter of good practice.

I look forward to reading any of your future contributions to *QEX*.

— 73, Ken Grant, VE3FIT, 5 Windrush Trail, West Hill, Ontario, Canada M1C 3Y5; **ve3fit@arrl.net**

Hi Ken,

I received your reply to my comments on your article, and read it carefully. You make some good points.

With regard to the balance trimpot, it is not sufficient to have the amplitudes of the two audio frequencies exactly equal at the microphone input to the transmitter, although it is a good starting point. If the audio gain in the transmitter is not the same at each of the two frequencies (due to equalization, high frequency boost, and so on), then unequal tone amplitudes will be presented to the balanced modulator and you will not get the clean Xs on the cat's-eye pattern. Furthermore, if the sideband filter does not have equal response to the frequency-shifted RF two tones (possibly due to ripple in the filter response), again the Xs will be lost and you will not observe a clean cat's-eye pattern. To observe the clean Xs on the cat's-eye pattern, it is necessary that the amplitude of the two tones at the point of measurement (usually at the output of the transmitter) be of exactly equal amplitude. This is easily accomplished by adjusting the balance control for clean Xs as you observe the transmitter output. It has been my experience in testing many transmitters that you need a balance range of several dB and it must be adjustable for each rig you might be testing.

You are absolutely correct in observing that variable capacitors can give you much better tracking than obtained with run of the mill potentiometers. They are also quieter and, if you use variables with a non-symmetrical pivot on the plates (as used in broadcast radios), you get the high frequency end of the range to be spread out a little (that's why they were used in broadcast radios).

I suggested using dual pots instead of variable capacitors for a number of reasons:

1. Many hams today do not have variable capacitors in their junk box and they are expensive to buy new.

2. Variable capacitors are much larger than potentiometers.

3. Isolating the frame and rotor from ground is difficult while minimizing stray capacitance.

4. You only get 180° of dial instead of the 270° or more with potentiometers.

5. It is difficult to get an oscillator below a few hundred hertz using variable capacitors. When the resistors get much above 1 MΩ, stray capacitances start causing problems.

6. My first Wien bridge oscillator that I built in 1960 (and still use) used dual pots and covered 20 Hz to 200 kHz in four ranges (a decade per range). It's hard to beat past successes.

When I read of your tracking problems with potentiometers, I was puzzled because I hadn't observed such problems. So, I pulled some dual 50 kΩ pots out of my junk box and measured them with a digital ohmmeter. Although the overall resistance was low by about 10% to 15%, the dual pairs were within 1% or 2% of each other. Also, when I put the dual pot at mid-range and measured the resistance of each section, they were still within a few percent of each other. These pots, which are garden variety brand X units, would not result in output level variations like you observed.

I then thought that maybe your use of the 327 lamp as a regulator might not be optimum. So, I measured the voltage-resistance characteristic of the lamp and found that its most sensitive range (highest change in resistance to smallest change in voltage) was right around 1.2 V RMS, which is reached when the oscillator output is 10 V p-p. Thinking I might be missing something, I built your circuit but substituted dual 50 kΩ pots for the resistors and used fixed capacitors. When I varied the frequency from 200 Hz to 3000 Hz, the amplitude variations in the output were at most a few tenths of a dB (a few percent). I then raised the value of one of the capacitors by 15% and still couldn't see an output variation greater than a few percent. Your circuit seems very robust! I see no reason why the use of potentiometers would give you so much trouble.

The one thing I did observe about varying the frequency with the dual pots was that noise is introduced as you rotate the pots. It takes the 327 lamp regulator a fraction of a second to calm the output to its regulated output amplitude after changing the frequency. I haven't yet tried squirting some cleaner into my potentiometers to see if that would reduce the noise. Is it possible that when you tried potentiometers, the problem was noise rather than a tracking problem?

In looking at the feasibility of using variable capacitors to get a large (15:1) frequency range, I looked carefully at the minimum capacitance in your circuit. In your article, you attributed 18 pF of stray capacitance to the input capacitance of the FET op amps. I doubt if the FET inputs are even close to this value. Instead, the stray capacitance, which should be compensated as you have done, is due to the stray capacitance to ground of the variable capacitor frame and shaft. Without care, this could easily exceed 20 pF. This stray is a little tricky to measure accurately and I am wondering how you found it to be 18 pF?

Anyway, I'm still playing with Wien bridge oscillators and the jury is still out as to whether I'll use dual pots or dual section variable capacitors to vary the output frequency. There are a lot of factors to be considered on both sides but I think either one would work okay for amateur purposes.

— 73, Bob Buus, W2OD

Larry

Your readers should be aware that a truly inexpensive two or more tone generator is available free on the Internet if they are using Windows computers. The program, NCH Tone, generates sine and other waveforms using the computer sound card. Running the program twice simultaneously will allow two tone generation, with frequency control over each tone. Both tones will have equal amplitude, and the combination can be adjusted with the computer volume control. See the URL, **www.world-voices.com/nchtone.HTML**.

The purchased version of the program only requires one program execution, and allows many tones, each of which is independently adjustable in frequency and volume. The price of a license for home use is less than $20, still a pretty good deal for what is a flexible function generator from 1 Hz to 20 kHz ,with sine, square, triangle, sawtooth, impulse and white and pink noise outputs.

For most two tone applications, the free version is satisfactory since equal tones are generally desired anyway and it also includes other waveforms.

— Bob Hicks W5TX, 1737 Midcrest Dr, Plano, TX 75075; **w5tx@verizon.net**

NimbleSig III — A Dual Output DDS RF Generator and Low Level RF Power Meter — Part 3 (May/Jun 2009)

Dear Thomas,

One point about the calibration of the power meter should be born in mind. Years ago, when Plessey Semiconductors were the world's leading supplier of log amp ICs, we found that it was very easy to have a log amp that had better linearity than the signal generator attenuators. We ended up with specially calibrated H-P attenuators external to the generator to ensure we had monotonicity and linearity in terms of the dB changes. Before we did that, we had a lot of anomalous results. The checks against a known power meter that Thomas Aldread, VA7TA, used is the way to go, but for anyone without a calibrated power meter, relying on a signal generator, I would recommend caution, especially if there is a "step" in the calibration. If that happens, try another generator or an external attenuator.

— 73, Peter E. Chadwick, G3RZP; **peter.chadwick@Zarlink.Com**

Greetings Peter:

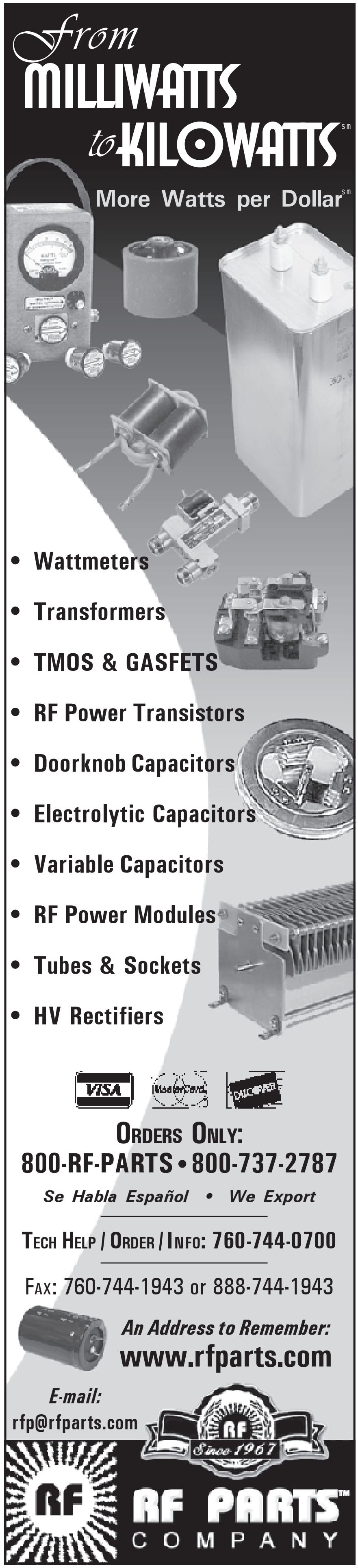
I agree with your comments and maybe I should have expanded on possible calibration pitfalls within my NimbleSig III (NS3) article. The use of a step attenuator with known good accuracy for controlling the generator output level is probably the best method for calibrating the NS3 detector tracking. My 1990s vintage 1 GHz class commercial signal generator happens to have a pair of electronically controlled step attenuators built into it that use hermetically sealed relays for switching. Fortunately it seems to have retained good accuracy.

The AD8307 specification provides a typical 80 dB dynamic range logarithmic conformance spec of ± 0.5 dB, which certainly sets a high standard for a variable attenuator to improve upon. In contrast the AD8307 specification indicates the logarithmic intercept point, which directly affects the absolute level calibration, can vary by as much as 10 dB. Although typically I have found this variation between devices to be less than a couple of dB it justifies the need for the detector calibration routine. The calibration procedure also provides assurance that the power detector is functioning properly.

As you point out there are probably many RF generators around that do not track the dynamic range as well as a typical AD8307 detector chip without any response calibration correction. Some legacy generators used analog mechanisms for the output attenuator design where tracking errors can occur as the dials are not resettable to fine accuracy and/or the log conformance of the mechanical mechanisms may not have been as good as that provided by modern day logarithmic amplifiers to start with. I have also encountered step attenuators with inaccurate step sizes or step sizes with frequency response issues. In some cases the switch/relay contacts had become resistive or in other cases the attenuator had changed value (possibly damaged from exposure to too much power). The availability of an RF signal source with accurate absolute level that can cover the dynamic range and frequency spectrum of interest is needed to achieve improved calibration. If the level accuracy of the RF signal source is questionable then possibly the NS3 power meter should be left with the default calibration data until confidence in the reference standard can be established.

One can usually get an indication of attenuator step size inaccuracy by reviewing the NS3 calibration data prior to saving it to non-volatile memory. The calibration data should closely follow the AD8307 nominal step size of 25 mV/dB throughout the calibration range. If there is a significant deviation from 25 mV/dB between calibration points which follows the insertion or bypass of a particular pad the attenuator accuracy should be considered suspect. Additionally multiple attenuator sections of similar loss can be compared to detect inconsistencies.

Of interest is that the NS3 DDS generator output, which can be adjusted in 0.1 dB steps over a 10 dB range, can be used to confirm the accuracy of attenuator sections up to 200 MHz. Within NS3 the level steps are defined digitally and are then created by the DDS digital to analog converter thus step size relative accuracy limited only by the DAC resolution is assured. The NS3 DDS DAC resolution with 10 dB of level reduction is reduced from the full output



level of 1024 steps to 323. This results in a remaining single step voltage scale accuracy of 0.31% or < 0.03 dB when operating with a 10 dB output level reduction.

DDS technology provides a useful tool for checking the loss accuracy of attenuators. NS3 can be used to check the loss of pad values up to 10 dB directly by simply comparing the loss of the attenuator to the level shift provided from the DDS. Once a 10 dB pad value is established the loss of 20 dB sections can also be checked by comparing to the level shift from the DDS plus the loss of the previously measured 10 dB section.

Note however that the 10 bit resolution of the NS3 power meter is about 0.13 dB thus worst case relative measurement accuracy, when using the built in NS3 power detector, would be limited to about 0.2 dB. There is room for improvement in this resolution limitation.

Also note that should one wish to check the attenuator under test for inaccuracies caused by possible stray RF coupling the output level of the NS3 generator might need to be amplified to extend the measurement dynamic range. Stray coupling is most significant when the attenuator is set to high overall insertion loss values with all attenuator sections switched in. Quality of test lead shielding can become an important factor during high insertion loss tests.

— 73, Tom Alldread, VA7TA, 7056 Railway Ave, Coutenay, BC V9J 1N4, Canada; **va7ta@telus.net**

Appalachian Golden Packet Attempt

On Sunday, 26 July some hams in thirteen eastern states will attempt the first annual Appalachian Golden Packet attempt. The objective is to locate 15 drive-up or hike-up sites where simple APRS packet mobiles or hand-held radios can be used as emergency digipeaters to demonstrate emergency communications capabilities. The APRS stations and other hikers along the 2000 mile area will be able to exchange APRS text messages live during the 6 hour event. The temporary RF links zig-zag across the trail to take advantage of the height above average terrain across valleys. Fifteen excellent sites have been identified so that all messages can be accomplished using a dual 7-hop digipeater path.

Although hikers along the trail can already get into the global APRS system from anywhere on the high parts of the trail using the national 144.39 MHz frequency, such routine operation depends on Internet gateways and is highly dependent on the established infrastructure. This golden packet event will use all portable equipment and an alternate frequency for 6 hours, to attempt the 2000 mile all-RF packets. For details and current information see WB4APR's Web page: **www.aprs.org/ at-golden-packet.html**.

— 73, Bob Bruninga, WB4APR, 115 Old Farm Ct, Glen Burnie, MD 21401; **410-293-6417 days; bruninga@usna.edu**.

Hi Bob,

Thanks for sending us this information. Perhaps some *QEX* readers will be interested in participating in the effort.

— 73, Larry, WR1B

Upcoming Conferences

43rd Annual Central States VHF Society Conference

July 23-25, 2009
Elk Grove Village, IL

The Central States VHF Society 43rd Annual CSVHFS Conference will be held in Elk Grove Village, IL, near Chicago on July 23-25, 2009. Presentations and posters on all aspects of weak-signal VHF and above Amateur Radio are requested.

Topics for papers and presentations at the conference may include:

Antennas, including modeling/design, arrays, and control.

Construction of equipment, such as transmitters, receivers, and transverters.

Digital Modes, such as WSJT, JT65 and other modes.

RF power amps, including single and multi-band vacuum tube and solid-state.

Propagation, including ducting, sporadic E, tropospheric and meteor scatter.

Pre-amplifiers (low noise).

Software-defined Radio (SDR).

Regulatory topics.

EME.

Digital Signal Processing (DSP).

Test Equipment including Homebrew, Using, and making measurements.

Operating, including contesting, roving, and DXpeditions

Non-weak signal topics, such as FM, Repeaters, packet radio, and so on, are generally not considered acceptable. There are always exceptions, however. Please contact K9XA, as listed below, if you have any questions about the suitability of a topic.

Submission Deadlines:

For Presentations delivered at the conference: Monday, 29 June 2009

For notifying us you will have a Poster to display at the conference: Monday, 29 June 2009. Bring your poster with you on July 23/24.

Contact: Kermit Carlson, W9XA, 1150 McKee St, Batavia IL 60510; **w9xa@yahoo.com**

Please see the Web site at **www.csvhfs.org** for more information.

The 28th Annual ARRL and TAPR Digital Communications Conference

September 25-27, 2009
Chicago, Illinois

Mark your calendar and start making plans to attend the premier technical conference of the year, the 28th Annual ARRL and TAPR Digital Communications Conference to be held September 25-27, 2009, in Chicago, Illinois. The conference location is the Holiday Inn Elk Grove Village Hotel, Elk Grove Village, IL. This is the same location as last year's DCC.

The ARRL and TAPR Digital Communications Conference is an international forum for radio amateurs to meet, publish their work, and present new ideas and techniques. Presenters and attendees will have the opportunity to exchange ideas and learn about recent hardware and software advances, theories, experimental results, and practical applications.

Topics include, but are not limited to: Software defined radio (SDR), digital voice (D-Star, P25, WinDRM, FDMDV, G4GUO), digital satellite communications, Global Position System (GPS), precision timing, Automatic Position Reporting System (APRS), short messaging (a mode of APRS), Digital Signal Processing (DSP), HF digital modes, Internet interoperability with Amateur Radio networks, spread spectrum, IEEE 802.11 and other Part 15 license-exempt systems adaptable for Amateur Radio, using TCP/IP networking over amateur radio, mesh and peer to peer wireless networking, emergency and Homeland Defense backup digital communications, using Linux in Amateur Radio, updates on AX.25 and other wireless networking protocols.

This is a three-Day Conference (Friday, Saturday, Sunday). Technical and introductory sessions will be presented all day Friday and Saturday.

Join others at the conference for a Friday evening social get together. A Saturday evening banquet features an invited speaker and concludes with award presentations and prize drawings.

The ever-popular Sunday Seminar focuses on a topic and provides an in-depth four-hour presentation by an expert in the field. Check the TAPR Web site for more information: **www.tapr.org**.

Call for Papers

Technical papers are solicited for presentation and publication in the *Digital Communications Conference Proceedings*. Annual conference proceedings are published by the ARRL. Presentation at the conference is not required for publication. Submission of papers are due by 31 July 2009 and should be submitted to:

Maty Weinberg, ARRL, 225 Main Street, Newington, CT 06111, or via the Internet to **maty@arrl.org**.

AMSAT Satellite Space Symposium and Annual Meeting

October 9-11, 2009
Four Points Sheraton Hotel
Baltimore-Washington Airport

There is a free shuttle bus from the airport to the hotel. You can use public transportation to visit Washington and the inner harbor area in Baltimore. Annapolis will require a car. Within 5 minutes of the hotel is the National Electronics Museum.

Call for Papers

This is the first call for papers for the 2009 AMSAT Space Symposium and Annual Meeting to be held October 9 - 11, 2009.

Proposals for papers, symposium presentations and poster presentations are invited on any topic of interest to the amateur satellite community. We request a tentative title of your presentation as soon as possible, with final copy submitted by September 1, 2009 for inclusion in the printed proceedings. Abstracts and papers should be sent to Dan Schultz, N8FGV, at **n8fgv@amsat.org**.

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Call for Papers

Microwave Update 2009 will be held on Friday, October 23 through Saturday, October 24 in Dallas, Texas. Technical papers are currently being solicited for publication in the ARRL *Microwave Update 2009 Conference Proceedings*. You do not need to attend the conference nor present your paper to have it published.

Strong preference will be given to original work and to those papers that are written and formatted specifically for publication rather than as a visual presentation aid. As this is a microwave conference papers must be on topics for frequencies above 900 MHz. Examples of such topics include microwave theory, construction, communication, deployment, propagation, antennas, activity, transmitters, receivers, components, amplifiers, communication modes, LASER, software design tools, and practical experiences.

Authors retain the basic copyright, and by submission, consent to publication in the *Proceedings* and possible publication of the *Proceedings* in CD/DVD format. If you are interested in submitting a paper please contact Kent Britain, WA5VJB at **wa5vjb@flash.net** for additional information. The deadline for papers is Monday, August 31, 2009.

Please refer to the Proceedings Style Guideline on the conference Web site at **www.microwaveupdate.org/index.php**.

If you are interested in presenting at Microwave Update please contact Al Ward, W5LUA, at **w5lua@sbcglobal.net**.

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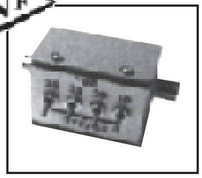

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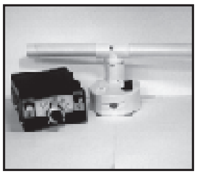

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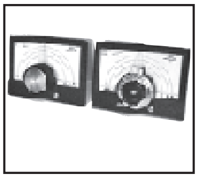

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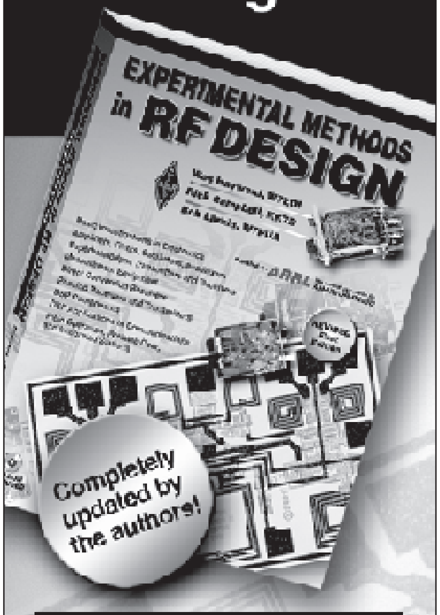

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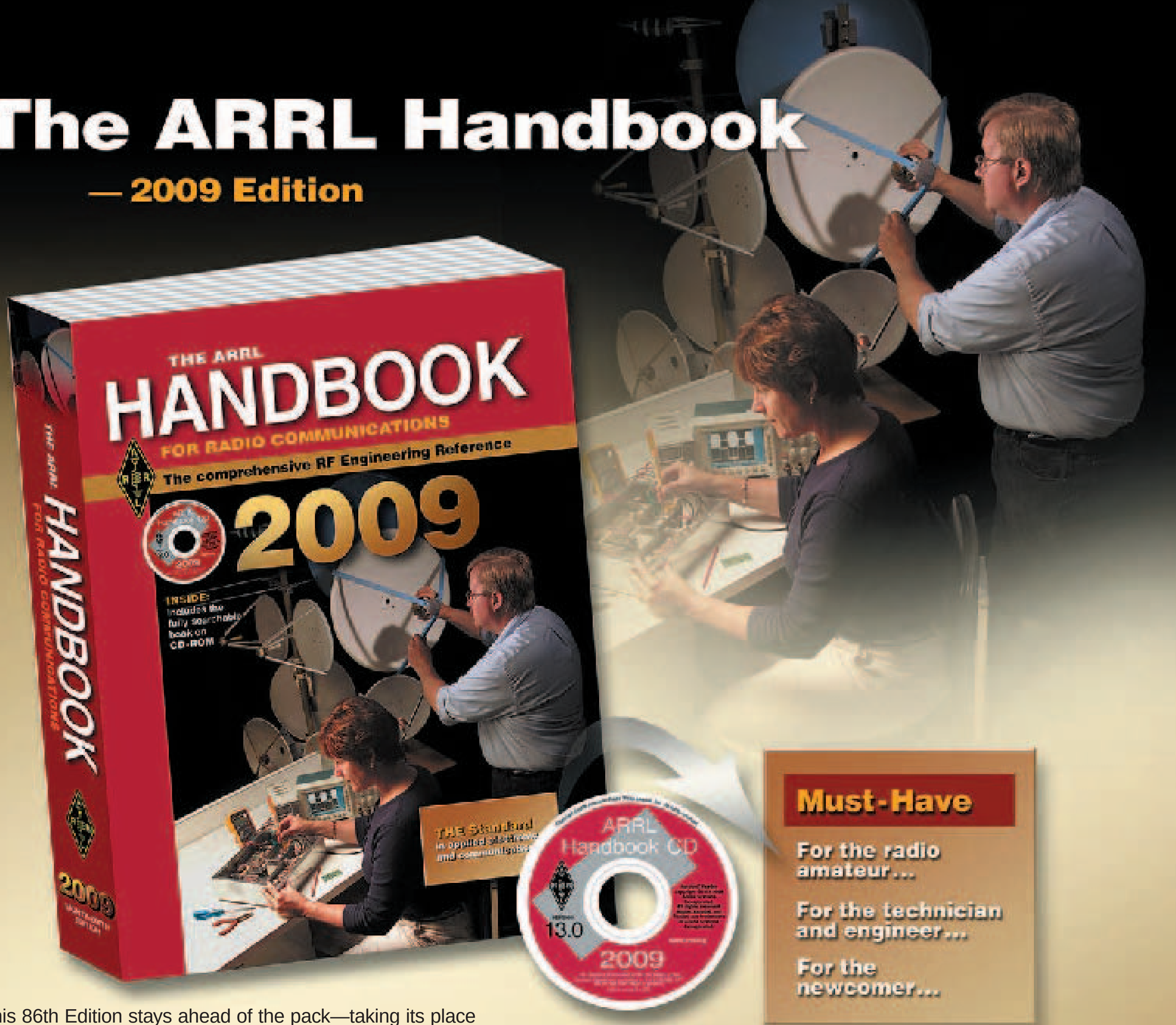

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