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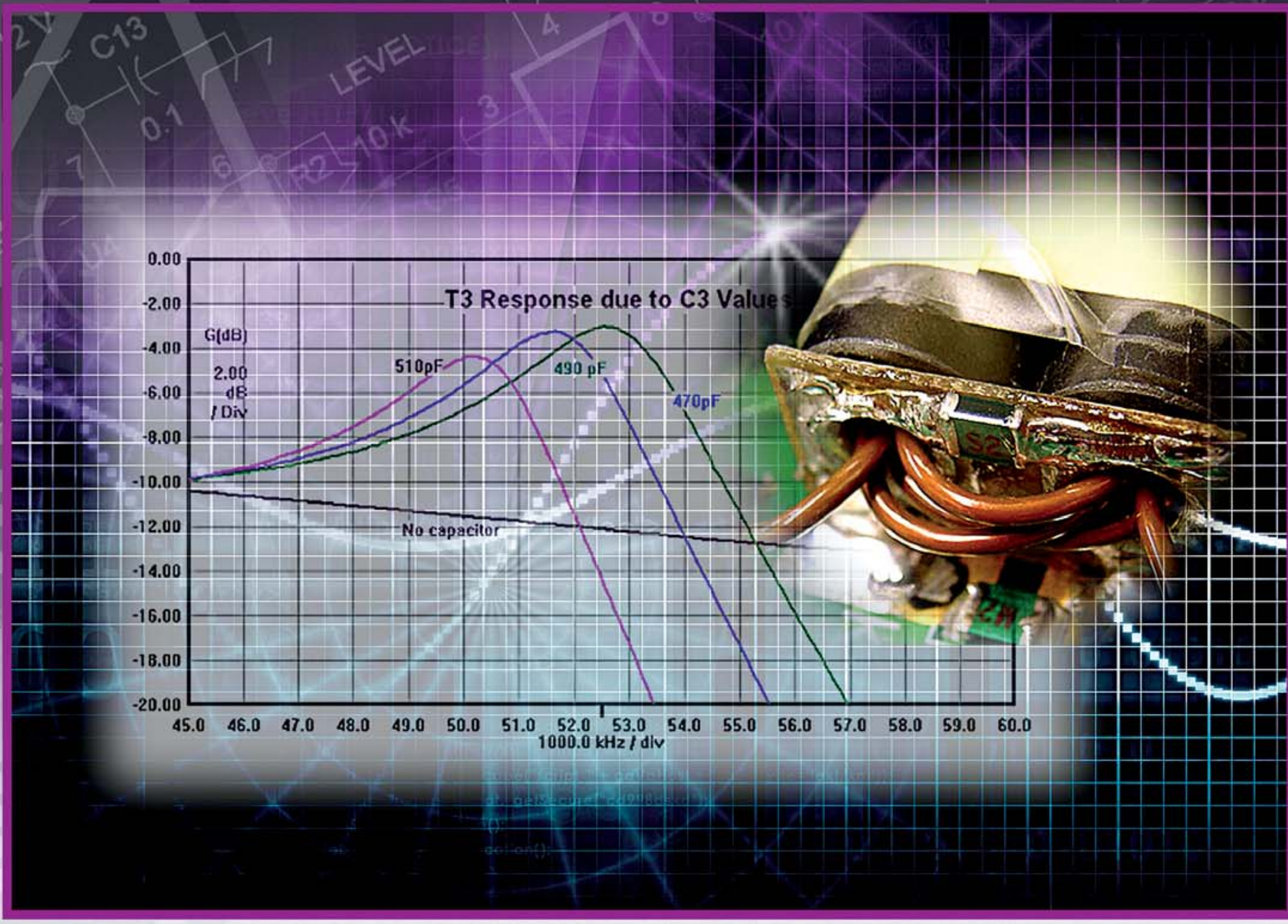

W6WO describes the output transformers he built and tested for a fixed-tuned, solid-state power amplifier. The output transformer may be the most critical component in obtaining good performance for HF through 6 meters.

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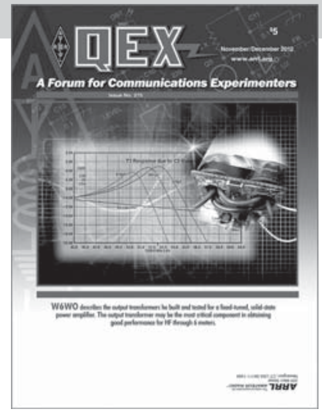
About the Cover

Ron Skelton, W6WO, describes the construction and testing of several output transformers for a fixed-tuna solid-state power amplifier. Ron explains that these output transformers may be the most critical components of an amplifier design, for obtaining good performance from such an amplifier on the HF bands, up through 6 meters.



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The purpose of *QEX* is to:

1) provide a medium for the exchange of ideas and information among Amateur Radio experimenters,

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

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

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and letters for publication in *QEX* should be marked Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted in word-processor format, if possible. We can redraw any figures as long as their content is clear. Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX* or high-resolution digital images (300 dots per inch or higher at the printed size). Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

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Larry Wolfgang, WR1B

lwolfgang@arrl.org

Empirical Outlook

On the weekend of September 21-23 I attended the ARRL/TAPR Digital Communications Conference in Atlanta. This conference rates as one of, if not *the* top technical conference in Amateur Radio every year. The presentations this year were once again outstanding. The attendees have such a wealth of knowledge and experience to share, whether they are giving a presentation or continuing the discussions in the hallways and demo room long after the talk is over.

One of the first presenters on Friday morning was Steve Bible, N7HPR. Steve gave a report on ARISSat-1 and the results of an on-line survey. ARISSat-1 provided a fantastic opportunity to share the excitement of Amateur Radio and the space program with a wide audience of students from elementary age through college engineering programs. Although it is hard to tell how the word about the survey spread outside of the AMSAT/Amateur Radio community, it was quite disappointing to see that less than 5% of respondents were educators on any level, with only about 2% being students.

We can hope that many more teachers took this opportunity to share lessons about science, math, radio, communications, space and engineering with their students, but there is little solid evidence that the program had the wide impact it was designed to create. Getting the word out about some of the neat things we do in Amateur Radio, which can be used in educational programs to spark interest in students, continues to be a challenge. There must be more we can do to reach out to students and help them become excited about science, technology, math and engineering topics and careers! Do you have any ideas?

Along those same lines, Bdale Garbee, KBØG, gave a presentation about AMSAT Fox. AMSAT's latest idea for obtaining launch opportunities for Amateur Radio satellites is to partner with various universities, where students are building cubesats. AMSAT can help by providing expertise with the RF engineering aspects. The partnerships will provide a convenient platform to include some Amateur Radio equipment on board. Fox-1 is a partnership with Penn State, where they are building a MEMS Gyro experiment. This satellite has been accepted by NASA for a free launch, with a target launch date of the second half of 2013.

I'm sure there was a lot more information and discussion about this program at the AMSAT Symposium in Orlando, FL at the end of October, but I did not attend that conference. This may be one example of a way for Amateur Radio organizations to form stronger partnerships with educational institutions. Most of us would probably agree that we, as the Amateur Radio Service, need to reach out more effectively to younger students, so that there would be more college engineering students interested in studying RF.

Bdale was also the Saturday evening banquet speaker. The theme of this talk was "Sharing the Joy of Making." Those who are acquainted with Bdale will know that he is an excellent story teller, and he kept us spellbound throughout his presentation. It's a theme that many hams can relate to. We enjoy building things, and perhaps taking something designed for one application and adapting it to something else entirely. Bdale has taken this to an extreme, and he certainly exudes the joy he feels when he builds something useful. He serves as an excellent example of how any of us should reach out to others and share the excitement we find in Amateur Radio!

Chris Testa, KD2BMH, became interested in Amateur Radio through a *Make Magazine* article. He has been licensed for a little more than 6 months, and was showing the circuit board he has produced, which will be an all-mode handheld SDR 2 m and 70 cm transceiver. Chris's radio has receive capability in the 100 MHz to 1 GHz range. He has the hardware working, and is looking for some others to help write software to enhance the radio. He is moving ahead with the manufacture of a limited number of "kits." Chris expressed interest in writing about his project for *QEX*, so I hope you will be hearing more about this in the near future.

I also had the great honor of presenting the 2011 Doug DeMaw, W1FB, ARRL Technical Excellence Award to James Ahlstrom, N2ADR, for his January/February 2011 *QEX* article, "An All-Digital Transceiver for HF." Jim is a regular at the DCC, and we have another of his excellent articles in the works for next year.

Gary Pearce, KN4AQ, has been videotaping all of the DCC presentations for a number of years, and he provided this service again at the 2012 Conference. Rather than producing and selling a set of DVDs with the entire Conference, Gary is now making those videos available on his website, **www.HAMRADIONOW.tv**. By the time you read this, Gary may have some, if not all of the recordings from this year's DCC posted. The videos don't take the place of actually being there to meet and talk with so many active experimenters, but there is a lot to learn just by watching the presentations. Enjoy!

Ron Skelton, W6WO

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Managing the Response of HF Transformers at 50 MHz

The author explores ways to build and test HF amplifier output transformers.

The design of a fixed-tuned, solid-state power amplifier for the whole range from 2 to 54 MHz is a complex, challenging task. Relatively few detailed designs extend beyond 30 MHz. Some choices and techniques for improving performance at 50 MHz in a 100 W Class AB amplifier are described in this article. Provided the active devices and the board layout are up to the task, the output transformer will be the most critical component. The focus here is evaluating transformer options prior to committing to board and circuit design.

Conventional and Transmission Line Transformers

The pros and cons of two classes of transformer were considered. One (termed "conventional") relies on magnetic-flux-coupled windings on a magnetic core and the other relies on transmission-line principles. The work of Jerry Sevick, W2FMI, (SK) published in his book, *Transmission Line Transformers* is the primary reference of choice on the subject.[1] The book provides precise details on construction and performance, but does not include a 1:16 balun. The pros and cons of the two classes are also well covered in Motorola application notes that come together in the book entitled *Radio Frequency Transistors*, and the Phillips Semiconductors application note *Design of HF Wideband Power Transformers*.[2,3]

My focus on conventional designs is principally because I did not achieve good results from using transmission lines. A transmission line transformer for matching the typical 3 Ω balanced output of a pair of transistors to a 50 Ω unbalanced load seems a daunting task; for example, note the special construction requirements described in Jerry's book in section 8-30. There is little doubt in my mind that conventional designs are simpler for covering the whole range from 2 to 50 MHz.

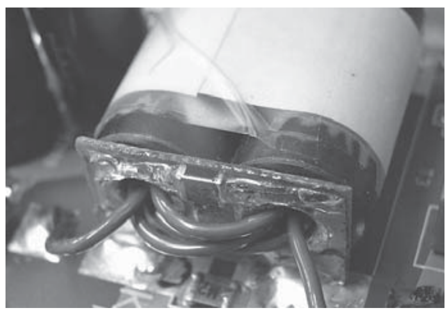

Figure 1 — Here you can see a 16 gauge, 4 turn winding in a binocular core.

The principal advantage of transmission line transformers is their immunity to forms of stray reactance that limits conventional designs. What follows are some results of dealing with these limitations in a conventional transformer.

Stray Reactance

An ideal conventional transformer would involve only inductive reactance, and all the flux will couple equally with all turns of the windings. In practice, some parts of the windings will not be coupled completely and they will behave as isolated inductive components, called leakage inductance. The upper frequency limit will be reduced by this leakage inductance, and conversely will be extended if turns are tightly wound to minimize leakage. The first clue is to observe the space between turns and between primary and secondary — open space makes it more likely there will be reduced output at the high end of 6 meters. There is a catch, however — tight turns also increase intra-winding capacitance that will tend to reduce output at the highest frequencies.

The output impedance of a pair of transistors of a few ohms will require a 1:9 or 1:16 transformer to match 50 Ω. The higher the turns ratio the more difficult it becomes to control leakage and 1:16 is probably a practical limit; this is the only type being considered here.

The binocular ferrite core seen in Figure 1 is a standard type with metallic tube construction forming a single turn. These can be very effective, but available impedance ratios would be $1:(N)^2$. Where these options are unsuitable, multi-ratio windings on a torroid could be used, however the limitations due to leakage reactance apply in either case. Figure 1 illustrates moderate, loosely coupled turns with a heavy gauge wire.

The connections to the board also contribute to leakage inductance. There will be some capacitance in any transformer, from turn-to-turn in a winding, and from inter-winding coupling primary to secondary. Capacitive reactance combines with leakage reactance to ultimately limit response at high frequencies.

Any dielectric material within the transformer naturally increases capacitance, and this makes the use of coax, shielded or jacketed wire generally undesirable.

Magnet wire is a logical choice for higher impedance windings and there is a bewildering variety to choose from. Wire intended for environments with fast rise times seem an appropriate choice. Specifications of this standard of wire will include details of how much scraping is tolerated, temperature and electrical properties. Some forms have internal lubrication to make tighter turns possible. REA is a magnet wire specialist and their website is worth studying. See

[1]Notes appear on page 7.

www.reawire.com. I would not recommend wire without provenance, as found in shopping malls.

Unfortunately, magnet wire is sold on 10 lb reels, but I found a local motor-winding shop that gave me odd ends of a wire (type TAIHSD) designed for severe duty with improved insulation against transient spikes and high frequencies. The REA type Nysol® also looks appropriate, and claims it can be soldered without prior insulation removal. It is conceivable that the dielectric factor and loss of insulation materials could influence high frequency performance at 50 MHz, but the makers cannot answer such questions. I found references to properties of some materials at 100 MHz, but these values do not seem to apply to thin films. My guess is that dielectric loss would be minor, so this line of enquiry ended.

A few words about core material seem in order. Low frequencies require a material with sufficient permeability to keep the shunt inductive reactance high. At 50 MHz we are more concerned about eddy current losses in the core material, however, such losses are beyond the scope of this article. Suffice it to say that nickel-zinc ferrites are preferred, and there is not much difference between mix 43 and mix 61 at 50 MHz.

Transformer Equivalent Circuits

It is valuable to compare one transformer with another before committing to a specific transformer design or board layout, and to do so leakage parameters need to be evaluated. Most, if not all, texts on the subject start with a complicated model followed by moderately complex calculations and simplifications. Philips Semiconductors and Reuben Lee of Westinghouse Electric provide some excellent examples of this approach in Note 3 and Note 4, and both are highly recommended reading. A simpler method for defining and measuring leakage parameters will be described next.

Leakage Measurement and Impacts

Leakage inductance (identified here as L_l) comes principally from the highest impedance winding and in the case of the single tubular turn, this is often all we need to be concerned with. A low-impedance short across the single turn will be coupled across the transformer as a short on the higher impedance turns. By definition, leakage inductance is not coupled or shorted, therefore L_l is the inductance measured between the ends of the high impedance winding. I used a VNA designed by Paul Kiciak, N2PK, to measure the L_l of a variety of windings and the powerful software by Dave Roberts, G8KBB, called *MyVNA* produced the results.

As mentioned above, inductive and capacitive leakages combine to limit frequency response, so C_l must be determined. Direct measurement of intra-winding leakage capacitance is difficult, however. A simpler technique is to first measure the self-resonance of L_l and C_l in the following manner, as described by Chris Trask, N7ZWY.[5] You will need a VHF signal generator and a 50 Ω detector with a bandwidth of at least 100 MHz. A spectrum analyzer with a tracking generator is ideal. Alternatively, a scope shunted by 50 Ω will be suitable. A power meter, such as the home-brew instrument described by Wes Hayward, W7ZOI, and Bob Larkin, W7PUA, in the June 2001 issue of *QST* is ideal for this purpose.[6]

Use a section of braid to short circuit the single tube and connect one end of the high impedance winding to it. Extend a section of coax from the detector and connect only the center conductor to the braid. Connect the output of a signal generator to the other end of the winding and connect the ground sides of the generator and detector cables together, without any additional connection to ground.

The L_l and C_l form a parallel circuit and act as a trap in series with the transformer. By sweeping the generator, the detector will find a sharp dip at the resonant frequency, f_r. Knowing L_l, calculate the reactance XL_l and since at resonance $XL_l = XC_l$ we calculate C_l. To see the impact, build a model in a filter design program. (The obvious choice is Jim Tonne's (W4ENE) wonderful *Elsie* program that produced Figure 2. *Elsie* is included on the CD that comes with any recent *ARRL Handbook*.)

This self-resonance technique provides a simple HF model and is a handy metric for rapidly comparing one transformer to another prior to placement on the circuit board. The graph of Figure 2 illustrates how the self-resonance of two different 4 turn 16:1 transformers affects the transformer at HF.

Trace 1 has a much higher self-resonance and ideally this should be at least 50% higher than the highest frequency of interest. What

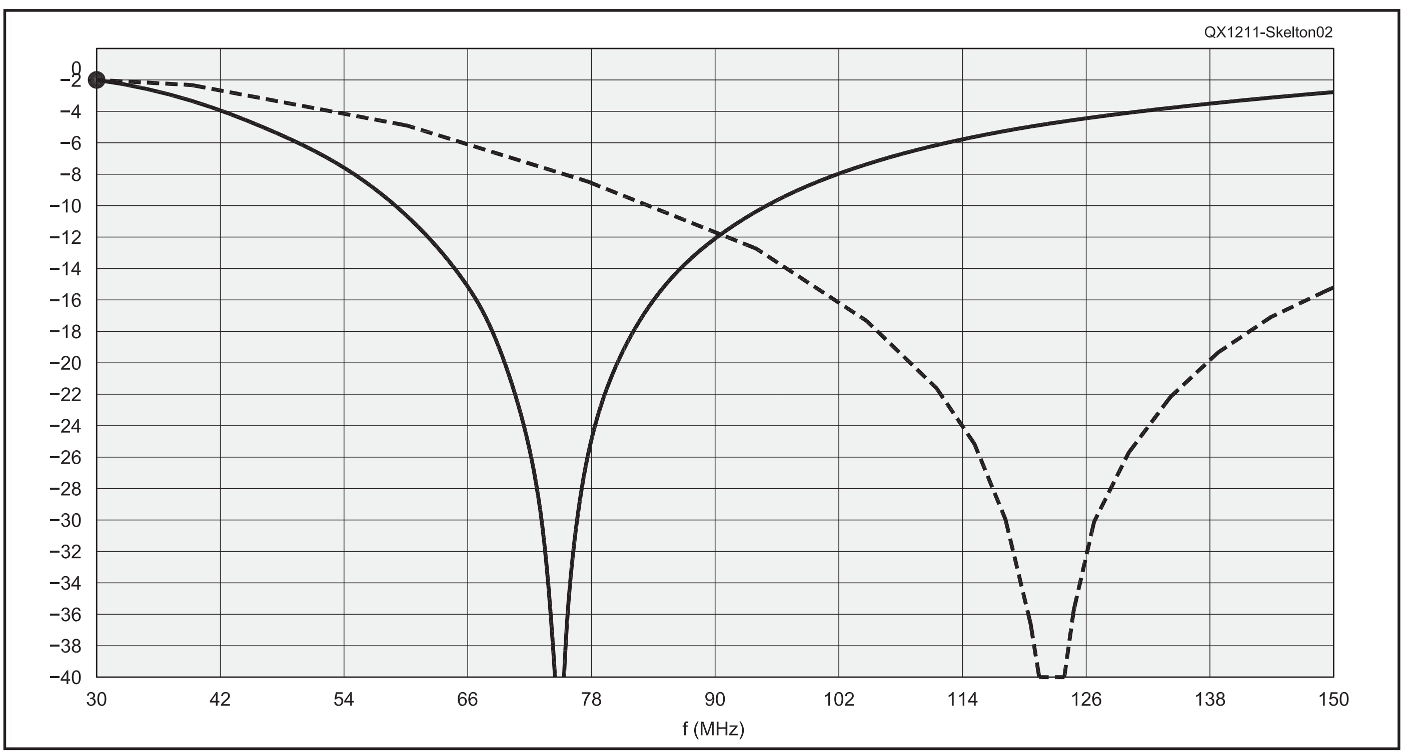

Figure 2 — This graph illustrates the affect of self-resonance on the HF response.

follows are details of some alternative windings.

The Prospect of Turns Wound with Multiple Strands

Recall that Litz wire lowers resistive losses by using bundles of many fine wires in parallel but insulated from one another. Litz wire is particularly helpful in achieving high Q in coils with the many turns required in LF tuned circuits. Stray capacitance limits the use of Litz wire for tuned circuits above 1 MHz so it is not obvious that the technique could be helpful at 50 MHz. Reducing leakage reactance using multiple strands is, however, advocated in texts such as *Design of HF Wideband Power Transformers*. See Note 3. The transformer seen in Figure 3 has three 20 gauge enamel insulated wires soldered together at the ends to form a bundle. Note the piece of braid used to short circuit to primary to enable measuring resonance by the method described earlier. The turns were not deliberately twisted but this occurred in the process of inserting one strand at a time.

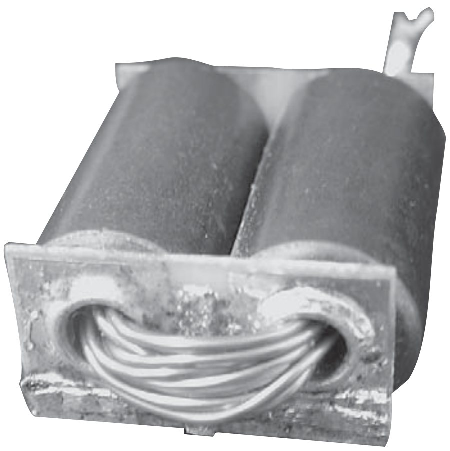

Figure 3 — Here is a 4-turn winding, with each turn being triple 20 gauge wire.

Measurements on many 1:N transformers with a few turns of multiple conductors often showed that stray reactance and self resonance were either superior comparable to single-strand turns.

Resistive losses are important at 50 MHz because they result in heat that has to be dissipated, potentially lower gain, increased drive power and increased intermodulation distortion. Resistive losses at HF are primarily caused by skin effect and proximity effect, which are both related to surface area rather than cross-sectional area. Consider a binocular core having tubes with 0.2 inch ID. The 4 turns required for a 1:16 ratio can be easily wound as a single conductor of 16 gauge, two of 18 gauge or three of 20 gauge wire.

Table 1 shows that a length of a twin pair of 18 gauge wire provides 58% more area than the single 16 gauge winding and the triple 20 gauge almost 90 % more.

The triple 20 gauge winding in Figure 3 had low leakage inductance but increased capacitive reactance, consequently the self-resonance was just 75 MHz and lower than the other windings. Winding three strands through metal tubes took more care to avoid bare copper, therefore overall either single 16 gauge or twin 18 gauge would be preferred to the triple strands.

Transformer Leakage Compensation

Compensation is the art and science of

Table 1

Dimensions of Single, Twin and Triple Conductor Windings

Conductors	*Gauge*	*Diameter (Inches)*	*Circumference (Inches)*	*Cross sectional area (Inch²)*
1	16	0.0508	0.16	0.002
1	18	0.0403	0.127	0.0013
1	20	0.032	0.101	0.0008
2	18	0.0403	0.254	0.0026
2	20	0.032	0.202	0.0016
3	20	0.032	0.303	0.0024

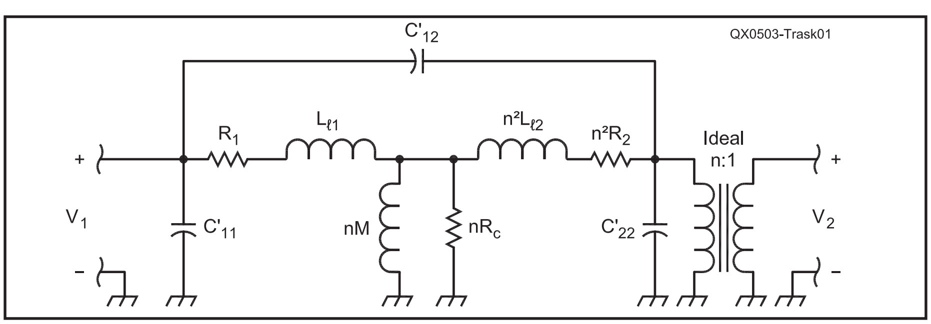

Figure 4 — This is the complete wideband transformer model presented by Chris Trask, W7ZWY, in his Mar/Apr 2005 *QEX* article.

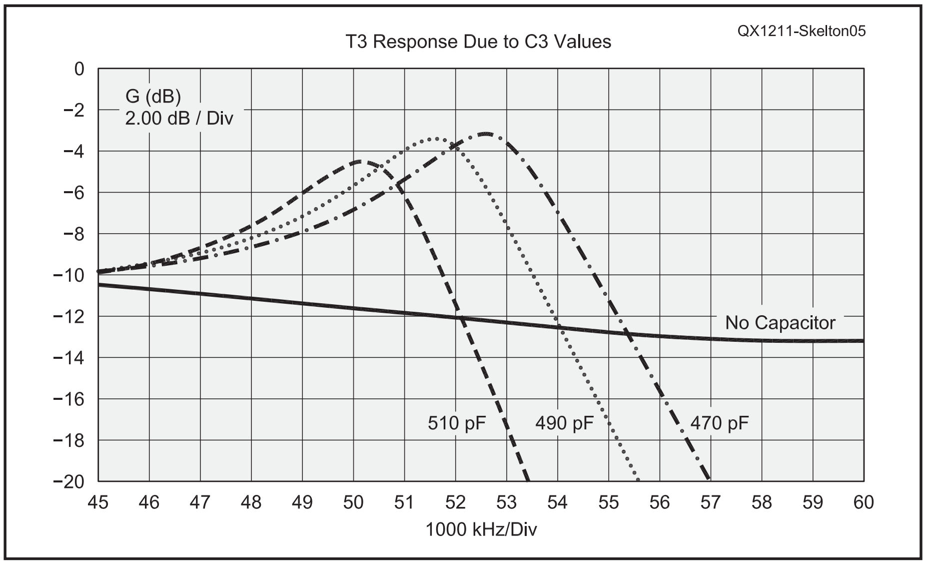

Figure 5 — This graph shows how a single compensation capacitor tunes the transformer frequency response.

negating leakage reactances with added components. The focus here is the frequency range between 28 and 54 MHz and the Mar/Apr 2005 issue of *QEX*, contains an article by Chris Trask, N7ZWY, that mentions this topic.[7] His Figure 1 (shown here as Figure 4) is very helpful in exploring aspects of compensation.

I ignored all resistances and assumed the high impedance of the shunt inductance can be treated as an open circuit at the frequencies being considered. The two leakage inductance values can be combined and the net result has the form of a low-pass filter and can be modeled as such.

Jerry Sevick, in his book describes a step-by-step method to calculate values for an upper limit of 30 MHz and a limited range of leakage inductance. (See Note 1.) Compensation may require the addition of capacitive reactance in the order of 1000 pF and it is essential to use capacitors with very low self-inductance. This practically mandates multiple SMD components in parallel and short, wide, balanced traces from transistor outputs to the transformer input. These capacitors must have low ESR and a voltage rating suited to the output power; 500 V is advisable for a 100 W amplifier. For example, see the specifications listed in the Digi-Key catalog.[8]

One such capacitor can be seen on the board in Figure 1. It is across the single turn and is the final one of several like it that may be required across the primary output path. There is another across the single turn at the upper level. At dc they would be in parallel but this circuit is so sensitive to inductance that they behave as individuals. Figure 5 illustrates adjustments to the upper capacitor on a transformer with leakage inductance of 0.25 µH. This illustrates an on-board measurement and how the value of the upper level acted to adjust response within the 6 meter band.

Figure 5 also illustrates that the tolerance of this specific compensation capacitor needs to be 5% or better. Where no components are available at the required value it is possible to add 1% or 2% silver mica capacitors in parallel, provided they have essentially no lead length. Compensation will require trial and error and one might say no amount of calculation is worth an ounce of persistence and luck.

Accurate measurements combined with the software equivalent of cut-and-try will be required to efficiently determine the value of the compensating components. The best program that I know of for this purpose is Ward Harriman's *SimSmith*, so aptly described on his web site along with short video tutorials. As the name implies the program conveniently combines component simulation with a computer-aided Smith chart. Figure 6 is an example of a *SimSmith* screen of a model optimized for the HF bands. With one

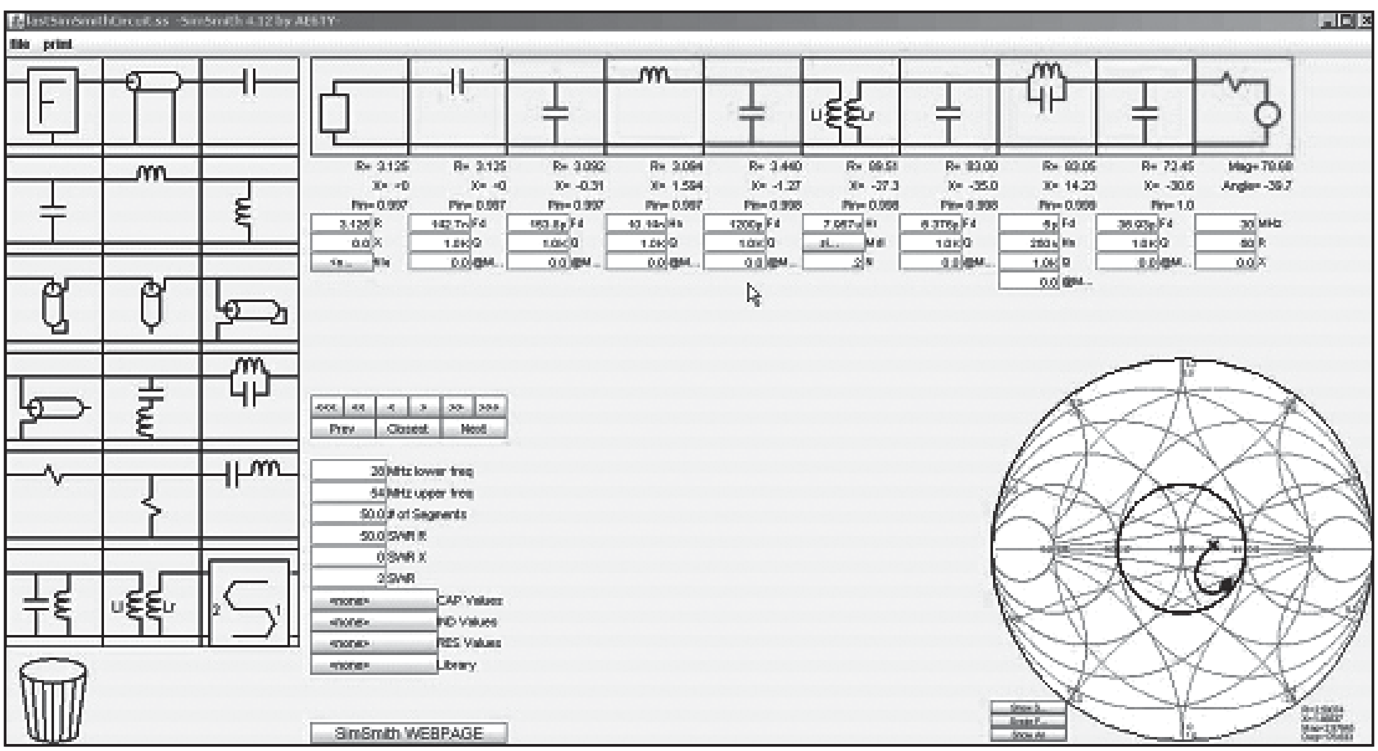

Figure 6 — This *SimSmith* screenshot shows a transformer model optimized for the HF bands.

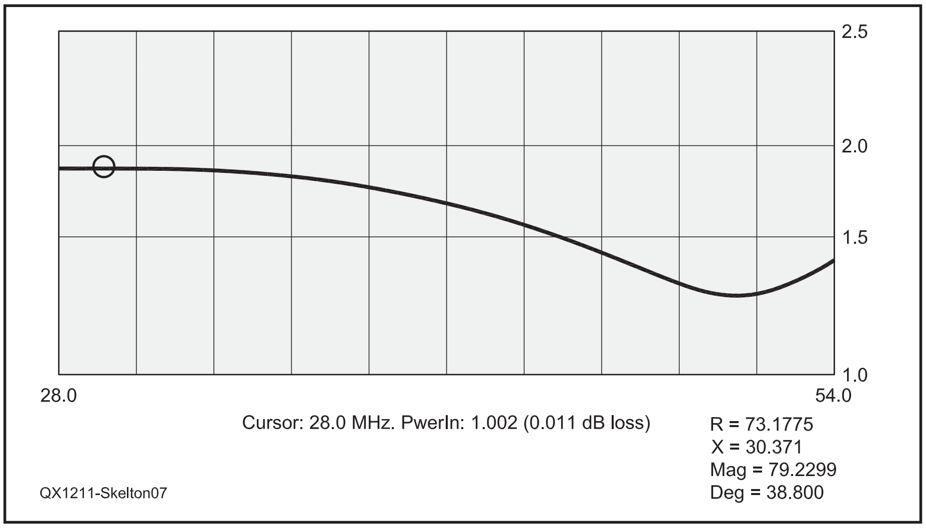

Figure 7 — *SimSmith* can generate this familiar SWR curve with a single mouse click.

mouse click, the familiar SWR graph shown in Figure 7 can be viewed. Without compensation, the SWR at 50 MHz was off the chart.

The transformer in this model initially had 4 turns to suit the end impedances of 3.1 and 50 Ω. I discovered, however, that a much better match came from a transformer with 5 turns. This is possibly caused by additional components modifying the return loss characteristic as seen by the generator. Incidentally it seems that Smith chart models expect the source to be on the far right of the circuit.

An unexpected feature of *SimSmith* allows the use of negative component values. With an existing design, negative values can be inserted in a model to tune existing components without the aid of a soldering iron.

Amplifier Efficiency

The efficiency of a Class AB amplifier is expected to fall in the range of 45 to 65% but section 5.5.3 of *The 2012 ARRL Handbook* makes the following statement "Class AB amplifiers are capable of higher efficiency, although the wideband circuits popular in HF transceivers typically offer only 30% at full power." Using transformers described in this article, the measured efficiency of 100 W amplifiers at 50 MHz was 38 to 40%.

In Conclusion

The HF performance of alternative transformers has been compared using the value of self-resonant frequency as a figure of merit. Self-resonance well above 50 MHz doesn't guarantee high on-board bandwidth but resonance close to 50 MHz does guarantee bandwidth will be very low. Providing the desired output in the 6 m band is quite possible given that active devices have generous margins, and special attention is paid to board layout and output stage components. In practice, some fine tuning of compensating components must be expected because tolerances of many circuit elements are critical.

Ron Skelton, W6WO, was first licensed in the UK as G3IHP in 1950. He worked for many years in Commonwealth countries in North America, Africa, Asia and the Caribbean. He moved to the US in 1975 and retired following a varied career in communications and information systems. He is an ARRL Member, a Fellow of the Institution of Engineering and Technology in the UK and is a Life Member of the IEEE.

Notes

[1] Jerry Sevick, N2FMI (SK), *Transmission Line Transformers*, 4th Edition, 1996-2001, SciTech Publishing. Available from your local ARRL dealer, or from the ARRL Bookstore. ARRL Order no. TLT4. 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] Norman Dye and Helge Granberg, Radio Frequency Transistors: Principles and Practical Applications, EDN Series for Design Engineers, Butterworth Heinemann, 2001.

[3] Phillips Semiconductors. Application Notes ECO 6907, *Design of HF Wideband Power Transformers. (transmission line transformers*) and ECO7213 (conventional transformers). These app notes are available for download from **www.datasheet4u.net/html/E/C/O/ECO6907_Philips.pdf.html** and **www.datasheet4u.net/datasheet/E/C/O/ECO7213_Philips.pdf.html.**

[4] Reuben Lee, Advisory Engineer, *Electronic Transformers and Circuits*, Second Edition, Westinghouse Electric Corp, Pages 150-153.

[5] Chris Trask, N7ZWY, "Wideband Tranformer Models: Measurement and Calculation of Reactive Elements," 10 October 2008: **http://home.earthlink.net/~christrask/Wideband Transformer Models.pdf**.

[6] Wes Hayward, W7ZOI and Bob Larkin, W7PUA "Simple RF-Power Measurement," *QST* June 2001, pp 38-43.

[7] Chris Trask, N7ZWY, "Designing Wideband Transformers for HF and VHF Power Amplifiers," *QEX*, Mar/Apr 2005, pp 3-15.

[8] DigiKey SMD 1812 Mica chip capacitors, MCFD221J-F.

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An All-Band Antenna

In this third article on surface wave transmission line theory and applications for use by radio amateurs, the authors describe a single antenna for the 160 m through 3 cm bands.

This article describes the construction and operation of an antenna that can be used on all of the amateur bands from 160 m through 3 cm. This extremely broad range is made possible by combining two structures; a traditional HF vertical antenna and an extended discone antenna. The HF vertical is used from 160 m through 6 m. An antenna for VHF and above is placed at the top and uses the HF vertical conductor as a surface wave transmission line (SWTL) feed line. This is done by means of an integrated SWTL launcher at the top antenna and a second launcher located at the bottom of the HF vertical. The VHF portion of the antenna can operate from the 2 m amateur band through the 3 cm band. Figure 1 shows this combination of HF and VHF+ as a single All-Band Antenna.

In many ways this article is a combination of previous presentations given at a number of hamfests, conventions and local club meetings. It uses an extended discone antenna at its top and it uses a SWTL similar to that presented in the May/June 2012 issue of *QEX* as a feed line.[1, 2, 3] One difference is that the SWTL for this antenna is made from a relatively large conductor. Instead of using no. 24 AWG magnet wire, it uses the ⅜ inch to 1¼ inch aluminum tubing of the HF vertical.

The addition of the VHF+ antenna to a conventional vertical antenna made from aluminum tubing causes some shift in electrical length of the vertical at HF, but otherwise it functions in these two regions as fairly independent antennas, and with appropriate filtering and matching, can even operate at the same time. If suitable radios are available, with this antenna it is possible to transmit or receive on the HF amateur bands while also transmitting or receiving on 144 MHz and higher frequencies.

The HF vertical is operated in the commonly accepted manner with the possible exception of the use of a base-located antenna tuner and matching transformer. These provide good match virtually everywhere within HF.[4] There is no compelling reason to run a vertical only on bands where it is an odd number of quarter-wavelengths long and near a low impedance resonance, that is, where it has a feed point impedance near 50 Ω. Good matching technique bypasses this restriction and allows the antenna to function well, even in the absence of an extensive grounding system. This is because where it is not an odd number of quarter-wavelengths, and particularly where it is an even number of quarter wavelengths, the impedance is high and the corresponding feed point currents are relatively low. This reduces the power losses in the ground or radial system that is used to provide the ground reference (image plane). Over all of the amateur HF bands, this antenna has relatively low ground current and matching losses.

The antenna does not need to operate at or near a resonance since efficient matching between a 50 Ω radio and the higher antenna impedance is possible down to below 2 MHz without it. In fact, the vertical need not be any particular length, though longer is usually better for the VHF operation of the All-Band antenna, since it places the radiator higher and generally improves communications.

The mode of operation at VHF and above may be less obvious. While it is essentially a vertical SWTL connected to a modified discone antenna, there are practical details that must be addressed in order to get this arrangement to work well. The first of these is the method of coupling to the SWTL. Both at the bottom and at the top of the vertical, the SWTL is made from a conductor considerably larger than that shown in the previous articles. Since the launcher's function is to transform 50 Ω to near 377 Ω while converting the TEM (transverse electromagnetic) mode in coax to the TM (transverse magnetic) mode on the SWTL conductor, increasing the conductor size forces the wide end of the launcher to be larger. This is because in order to reach this high impedance in a coaxial structure, the outer/inner conductor diameter ratio must be more than 500:1. To prevent this dimension from becoming impractical, at the ends of the vertical where the launchers attach, the tubing tapers to a relatively small diameter, ¼ inch or ⅜ inch. Also, at both ends of the HF vertical, the outer conductor of this Klopfenstein taper transformer is fabricated from ¹⁄₁₆ inch brazing rod rather than solid sheeting. This has the effect of further reducing the influence of the outer shielding conductor and produces higher impedance from a smaller outer/inner ratio.

As a practical matter, a launcher with a mouth hundreds of times ¼ inch is still many feet in diameter and not structurally viable. Fortunately, because the impedance of coaxial line in the launcher varies as the logarithm of the outer/inner ratio, it is possible to compromise a little on the high impedance end and greatly reduce this size without too much loss of performance. For the integrated launcher/discone at the top of the SWTL, 330 Ω was targeted instead of 377 Ω. This produced a little additional mismatch error but fortuitously, at the high impedance end of the launcher, most of the

[1]Notes appear on page 18.

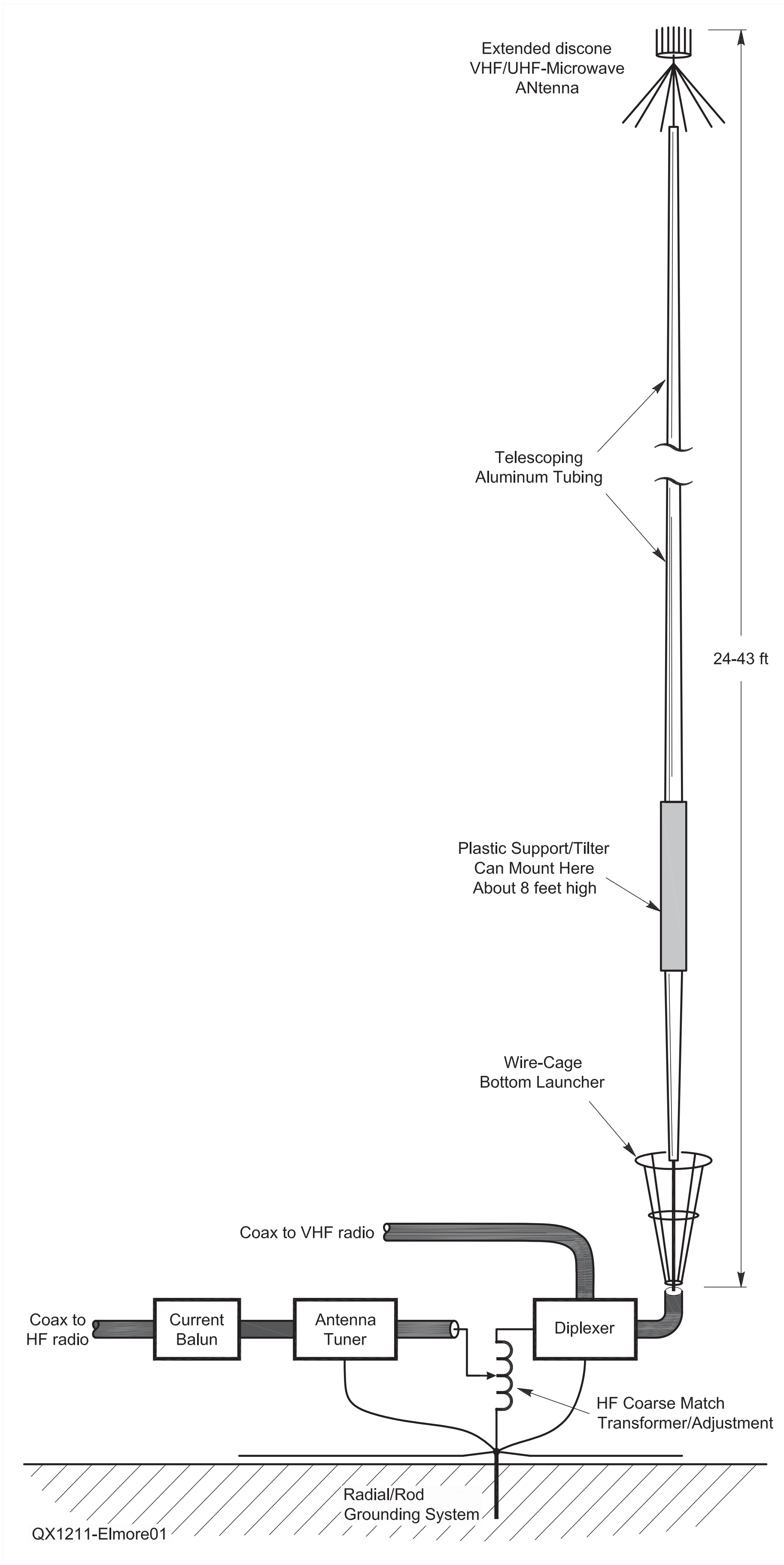


Figure 1 — A combination of an HF vertical and a VHF+ extended discone are used to make an effective antenna that can be used on all amateur frequencies from 1.8 MHz through 10 GHz. The VHF+ antenna is fed by a surface wave feed line, which uses the HF vertical aluminum tubing and special launchers at the top and bottom. The top launcher is integrated into the extended discone itself.

energy has already been converted to the TM mode, so impedance mismatches have less influence on the overall performance. By limiting the wide end of the extended discone to 36 inches and using ¼ inch as the center conductor diameter, a practical compromise was reached.

The bottom end of the SWTL has an identical problem. Here, even more deviation from ideal was made in the interest of practicality. The launcher shown is actually one that was designed as an all-weather version of the SWTL launcher shown in the first *QEX* article. It was originally designed for use with no. 24 AWG copper wire. For expedience, we simply used this launcher and inserted it into the ⅜ inch aluminum tubing until the tubing inner diameter matched the taper of the center. This is a fairly serious compromise but it avoids a larger structure at the bottom of the antenna, which was not visually acceptable on the backyard lawn at the N6GN station QTH. Using the launcher this way effectively truncated its transformation function and ability to generate 377 Ω impedance at the wide end.

Because of these compromises, the resultant SWTL impedance match is worse than the 1.22:1 SWR (20 dB return loss) target for a properly built and applied Klopfenstein taper launcher shown in the first article. That greater SWR along with the approximately 2:1 SWR of the extended discone results in higher overall SWR for the finished antenna at 2 m and above, but the actual impact of this higher SWR is not as severe as might generally be thought. Even an SWR of 4:1, which equates to a return loss of about 4.4 dB, only results in about 2 dB of mismatch loss. In practice, the degradation due to such a mismatch is barely perceptible and this antenna has SWR much better than this over almost the entire VHF/UHF range.

A somewhat better VHF+ match would be obtained by substituting a 36 inch diameter launcher at the bottom that is similar to the cone portion of the extended discone at the top, but even without this improvement, the version we built shows a final VHF+ SWR as plotted in Figure 4. The antenna works well and is really quite acceptable on all of the amateur bands below 2.4 GHz.

Operation on even higher frequency microwave bands is possible, but good performance requires considerable precision. The launchers and SWTL can easily operate past 10 GHz if care is taken to avoid sudden discontinuities. This can best be done by replacing the wire cage with solid sheeting for a few inches near the coaxial ends of each launcher. Similarly, the discone at the top needs to have the region near the apex of the cone carefully constructed. As a practical matter, however, an omni-directional antenna above 2.4 GHz may not be too useful for DX communications. Because the physical aperture of fixed-gain antennas falls as the square of wavelength, communications links using low gain antennas like dipoles or this discone exhibit high path loss at higher frequencies such as the amateur microwave or millimeter wave bands.[5]

In the final analysis, this antenna is "just a vertical" and behaves like one — neither dramatically better nor worse than a vertically polarized dipole at the same location and having the same far field environment. But a vertical can be a very useful antenna, particularly if the regions of radiation are well situated. This antenna offers a relatively low visual profile, can provide excellent results and is truly an All-Band Antenna.

Vertical Construction

The HF vertical portion of the All-Band Antenna is constructed from 6 foot sections of telescoping aluminum tubing of multiple diameters. Each section nests snugly and overlaps with adjacent sections providing a way to taper from the ⅜ inch diameter at the bottom where the bottom launcher attaches, up to 1¼ inch diameter at the plastic support and back down to ⅜ inch diameter at the top, where the ¼ inch threaded rod of the integrated launcher/discone attaches. This tapering provides adequate mechanical strength along with good SWTL performance.

In order to both support and to access the extended discone for changes while we were developing it, we constructed a "tilter" from plastic pipe and fittings in order to hold the vertical at its wide, stronger midsection

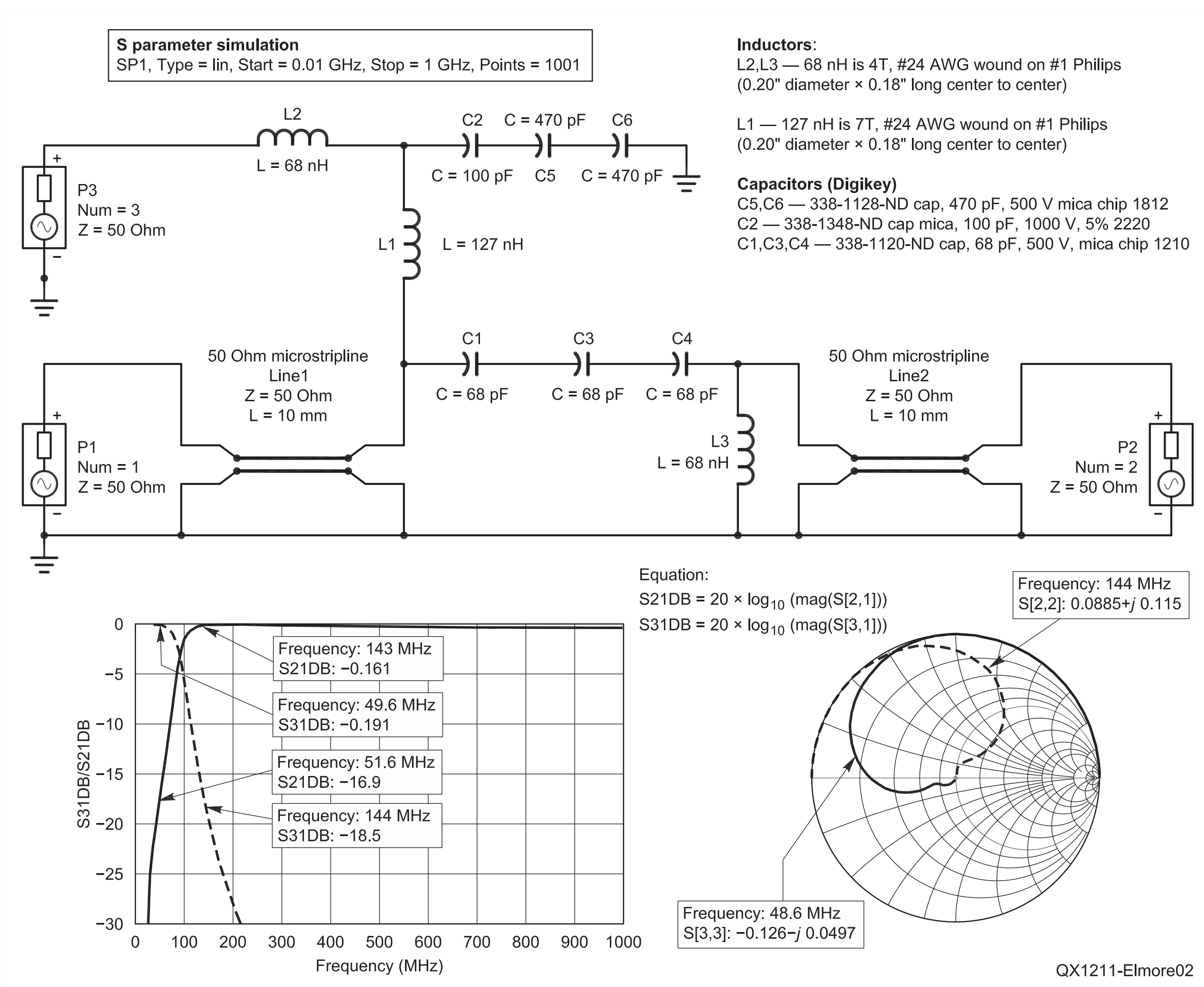


Figure 2 — Schematic and response of a HF/VHF+ diplexer that may be used with the All-Band Antenna.

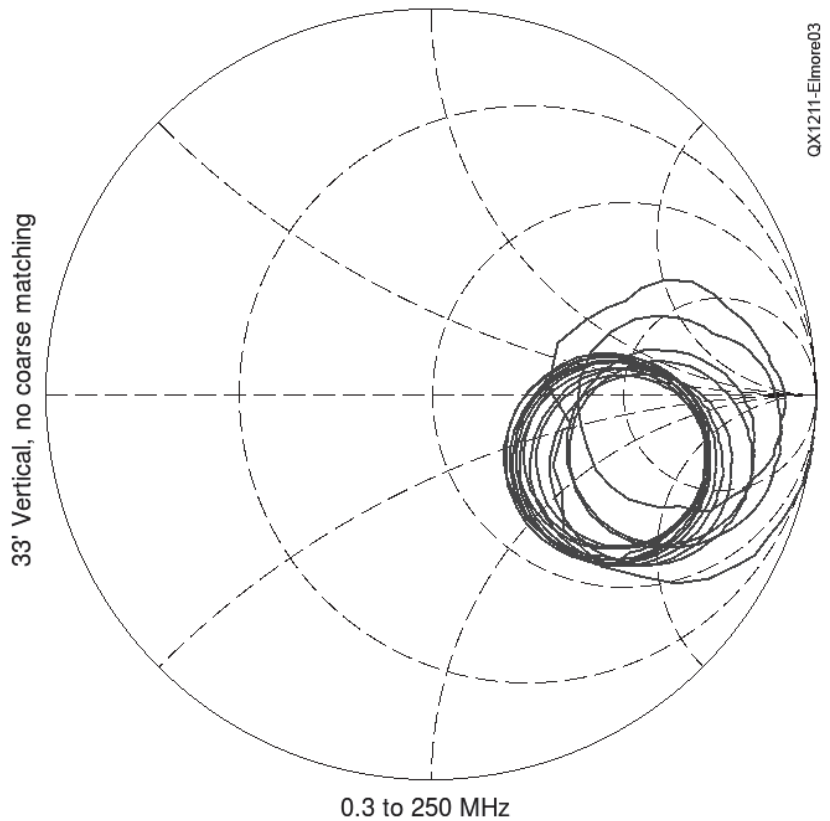


Figure 3 — Measured S11 (50 Ω reference) from 0.3 to 250 MHz of 33 foot vertical (without SWTL) with 24 inch metal disk improving a sod and ground rod image plane. Improvement by this disk is evidenced by the cleaner, lower impedance circles at higher frequencies. Note that in operation the antenna actually uses a transformer to shift the circle centers to nearer 50 Ω.

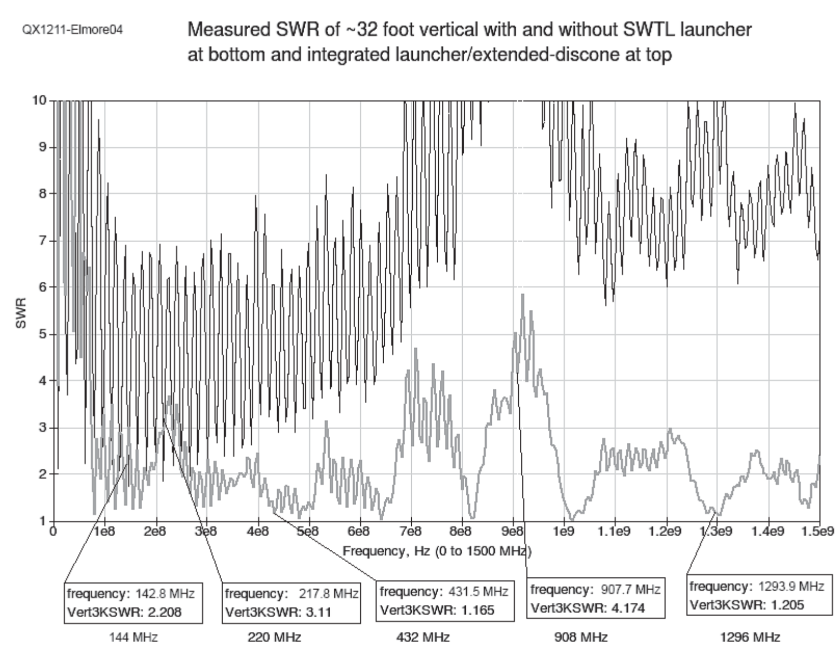


Figure 4 — Measured UHF SWR of All-Band vertical at the bottom launcher N connector without and with the SWTL launchers and extended discone.

and allow lowering the entire assembly to horizontal. This plastic pipe does have some negative impact on SWTL performance but it isn't too severe. A different method of support that doesn't involve any metal or plastic near the aluminum tubing would probably be better but hasn't yet been built. It is likely that a permanent antenna constructed without the tilter but with dielectric guys, perhaps of Dacron, could be simpler and might perform even better than the design we show here.

The tilter supports the antenna at the 7½ foot point so it is necessary to taper rapidly from the ⅜ inch tubing at the bottom to the 1¼ inch diameter tubing at the support point. We used a taper of approximately equal length sections. On the upper part of the vertical, where it is necessary to taper back down from wide to narrow, we used the full 6 foot lengths of the largest three diameters and then tapered linearly down to the ⅜ inch in order to gain as much rigidity as possible. The tilter can be constructed from PVC pipe fittings as shown in Photo 1. Other than keeping the majority of the plastic as far away as possible from the aluminum tubing, there are no special requirements.

Photo 1 — N6GN holding a bottom launcher and pointing toward a second launcher installed at the base of the All-Band HF /VHF/UHF antenna. The PVC tilter can be seen in this Photo. Styrofoam doughnuts have been added to keep the center conductor concentric with the outer cone.

Table 1

Wire-Cage 144+ MHz SWTL launcher for use with No. 24 AWGwire

Dimensions for bottom launcher inner and outer conductors and spacers. Though designed for #24 magnet wire, it is being used to drive a ¼" to 1¼" tapered aluminum conductor.

This is a wire substitute for the original K&S tubing and Paper Cone SWTL Launcher that was detailed in the first *QEX* article. It is also used as the bottom launcher for the All-Band Antenna.

The tapered center conductor "cage" is made from 6 (hexagonal) 1⁄16" brazing rods. The outer conductor is a conical "cage."

A female bulkhead N connector is mounted on a ½ inch copper water pipe flange and receives the 1/16 inch rods of the outer conductor and also a single connection from the tapered centered conductor to the N center pin.

At about 27 inches, the inner conductor/hex-cage rods end in a 3⁄16 inch OD tube that traps them inside.

For use as an all-weather SWTL launch, the taper continues down to no. 24 AWG wire at the mouth of the launcher.

When used as the bottom launcher on the All-Band antenna, ⅜" aluminum tubing from the vertical slides over the center conductor from about 23 inches forward — truncating the tapered region.

Position, Inch	*Desired Z_0 (Ω)*	*Center-center Outer Wire Spacing Flange*	*Inner Wire Spacing Pin*	*Inner Wire Spacers N Connector Here*
0	61	0.87	0.34	six 1⁄16 inch rods around ¼ inch tube w/ hole for N conn pin
1	62.5	1.05	0.44	
2	64.2	1.23	0.53	
3	66.1	1.41	0.63	0.625 @ 2.95
4	68.3	1.59	0.72	
5	70.8	1.77	0.81	
6	73.5	1.95	0.86	0.875 @ 6.25
7	76.5	2.13	0.92	
8	79.8	2.31	0.96	
9	83.6	2.5	1	1 @ 9
10	87.7	2.68	1	
11	92.2	2.86	1	
12	97	3.04	1	1@ 12
13	102.4	3.22	0.97	
14	108.3	3.4	0.93	
15	114.7	3.58	0.88	0.875 @ 14.9
16	121.6	3.77	0.83	
17	129	3.85	0.74	
18	137	4.13	0.64	
19	145	4.31	0.62	0.625 @ 18.75
20	155	4.49	0.53	
21	165	4.67	0.46	
22	175	4.85	0.4	0.375@ 22.36
23	186	5.04	0.33	11⁄32 inch joiner. For All-Bander, the vertical attaches about here.and smaller indductor conductor diameters aren't used.
24	197	5.21	0.28	
25	208	5.4	0.23	
26	220	5.58	0.18	
27	233	5.76	0.14	3⁄16 inch clump around 1⁄16 inch tube (All-Band antenna)
28	245	5.94	0.11	
29	258	6.12	0.08	
30	271	6.3	0.05	
31	284	6.48	0.03	
32	296	6.61	0.02	No. 24 AWG wire from here to mouth (All-weather SWTL launcher)
33	309	6.85		
34	321	7.03		
35	333	7.21		
36	344	7.39		
37	355	7.57		
38	365	7.75		
39	374	7.93		
39.34	377	8		

Since neither the HF nor VHF modes of operation rely on resonance, there is really no reason that the antenna can't be a length different from the 33 feet we used. If it is possible to go longer and thus higher, both HF and VHF/UHF performance will likely improve.

At this relatively short length, 160 m matching is a bit more challenging and probably not quite as efficient as a longer antenna would be. If you have the possibility of making the antenna longer, using additional sections of larger diameter tubing is perfectly acceptable and probably worthwhile. This will likely require the addition of insulating guy lines placed at one or more points, however, to withstand normal winds.

As we were developing it, we first used plastic hose clamps to capture the vertical within the plastic tilter and to adjust the lengths of the sections, but once we were happy with the mechanical strength and electrical performance we replaced all the clamps, except those at the tilter, with sheet metal screws close to each section end to tie the whole structure together both electrically and mechanically. We avoided using metal hose clamps since they produce discontinuities to the surface wave that can negatively impact the UHF/microwave performance.

Bottom Launcher Construction

The bottom launcher is made mostly from 1⁄16 inch brazing rod, for both the inner and outer conductors of the Klopfenstein tapered coaxial transformer. The result is a sort of wire cage that provides function without being quite as unsightly as sheet metal and solid tubing would be. This launcher should also be able to serve as an all-weather substitute for the metalized paper SWTL launcher shown in the first *QEX* article. It's heavier and more difficult to construct, but is a much better fit for continuous outdoor duty.

Target dimensions and impedances versus length for the coaxial line formed are shown in Table 1. The last column in this table also gives locations and center to center dimension for the metal spacers used to maintain the inner conductor shape. These spacers are made by drilling six holes, laid out in a hexagon with diameters and longitudinal locations shown in the Table 1, in 1½ inch square pieces of 0.01 inch thick brass shim stock. The rods are threaded through these holes and everything set squarely into position on a flat surface. Once everything is correct the spacers can be first tack-soldered and then completely soldered into place. At this point, the square spacers have done their job and can be trimmed and sanded or filed down to smaller circles as shown in Photos 1 and 2A.

Extended Discone/Launcher Construction

The combined SWTL launcher and extended discone antenna is made mainly from 1⁄16 inch brazing rod in generally the same way as the bottom launcher. Table 2 provides the dimensions.

The top disc is 5 inches in diameter and made from copper sheeting, pre-drilled at the center to clear the ¼ inch threaded center support. Four 24 inch pieces of brazing rod are bent into U shapes and soldered to the top of this disc to form a 9 inch high, 6 inch diameter cylinder. Prior to soldering, these can be held in place by drilling two pairs of holes for each of the 8 resulting upright rods, one pair near the outer edge of the disk and the other an inch or two from the center hole. A short piece of bare wire can be stitched around the rod at each location and twisted to hold things tight while soldering. Once soldered, the center of each U is snipped away, to leave room for a ¼ inch brass nut and washer, which attach to the central supporting rod. A 1 inch PVC plastic reducer is attached to the disc with short sheet metal screws. Photo 4 shows the bottom part of the finished cylinder joined with the cone. A setscrew holds the PVC pipe and reducer together and provides a way to easily disassemble the cone and cylinder parts when necessary.

The cone portion, which does double duty as the outer conductor of the top Klopfenstein tapered SWTL launcher, is built on a very short section of 1 inch PVC plastic water pipe. The ends of eight 36 inch long brazing rods were bent and inserted into one of eight equally spaced holes in the pipe. Circular copper wire rings were then soldered at the

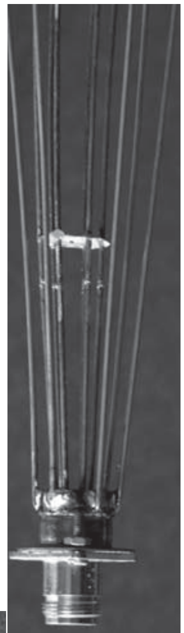
Table 2

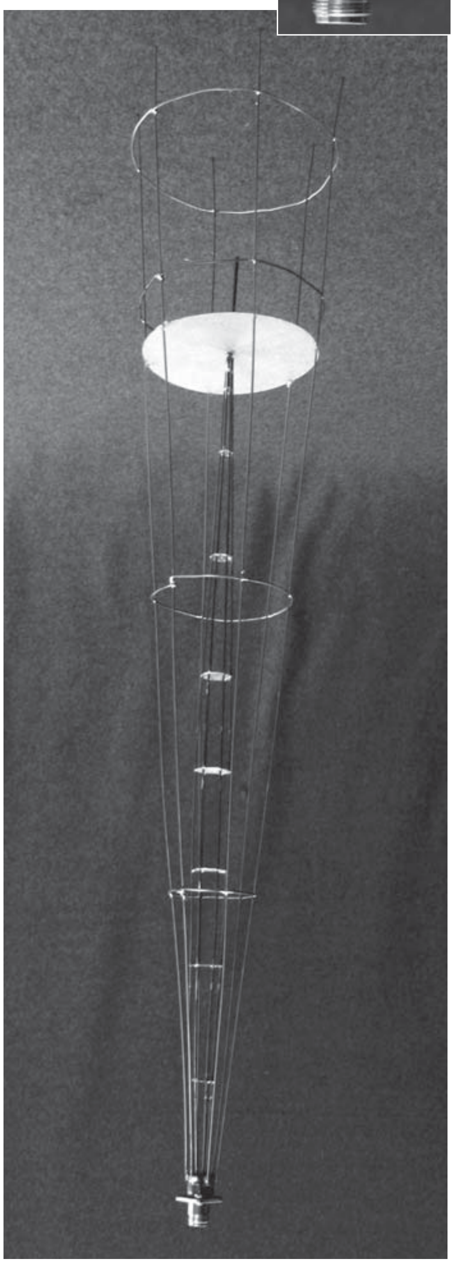
Dimensions for top launcher inner & outer conductors along with target impedance for the Klopfenstein taper. The outer conductor of this launcher doubles as the bottom cone of the integrated extended discone antenna. Although this outer conductor/cone makes no electrical connection to any other part of the SWTL or antenna, its presence is vital to the proper operation of both.

Integrated launcher used in VHF+ extended discone antenna. As for the bottom launcher, this launcher is constructed from 1⁄16 inch brazing rod. The inner conductor is a hexagon of 6 rods, separated by plastic spacers. The outer conductor, which doubles as the cone portion of the antenna, is made from eight brass rods, each 36 inches long. The plastic spacers hold the inner conductor rods in position while copper wire circles help maintain the outer conductor/cone dimensions.

Position from narrow end of cone	*Desired Impedance*	*Outer Diameter of 8 rods in Discone*	*Inner Conductor of 6 rods,| center to center*
0	60	2	1
2	66	4.5	2.7
4	70	6.5	4.2
6	77	8.5	5.2
8	86	11	6.2
10	97	13	6.2
12	111	15.5	6
14	127	17.5	5
16	147	19.5	4
18	168	21.2	3
20	193	23.5	2
22	218	26	1.4
24	247	28.5	0.9
26	273	30.5	0.58
28	298	33	0.38
30	320	35	0.26
32	330	36	0.25

Photo 2 — Part A shows the detail of the coaxial end of the bottom launcher, with an N connector mounted on a copper flange. Part B shows the complete bottom launcher with both the inner and outer assemblies constructed from 1⁄16 inch brazing rod. A sheet plastic disc is used to maintain alignment, spacing and shape between the inner and outer conductors. The outer conductor is held in shape by copper wires formed into circles and soldered around the outer rods.





pipe, as seen in Photo 4, and also at a point about ⅔ of the way to the bottom of the cone. The resulting cone apex angle is about 60°.

The inner conductor of the launcher is specially tapered to provide impedance matching between the high impedance of the SWTL line and the lower impedance of the extended discone. As already described, to keep the structure size down, the target transformation was from 50 to 330 Ω rather than to 377 Ω. This was made with eight 1⁄16 inch brazing rods equally spaced around a central ¼ inch threaded rod, which runs almost to the top. Near the top we extended the steel rod with a section of ¼ inch threaded brass. Because the steel is everywhere inside brass or aluminum, no significant RF current flows on it. The conductor taper and shape is set by four plastic and one metal spacer, as seen in Photo 3. Metal or plastic are equally acceptable as spacers but plastic spacers were used at the wider portions for reason of mechanical strength and reduced weight. At the narrow end of the center conductor the eight rod ends are captured inside a length of ⅜ inch brass tubing around the threaded rod and the center conductor from that point on is made either from tubing or ¼ inch rod. When assembled, the threaded rod will slip inside the top vertical aluminum tubing, which is ⅜ inch OD and a little more than ¼ inch ID. The top inch or so of the aluminum rod can be slotted and clamped around the threaded rod with a plastic hose clamp to guarantee good electrical contact or, as an alternative; a short set screw can be used. If you use a set screw, pick the length so that no unnecessary extra length protrudes from the tubing because this can produce unwanted reflections of the surface wave.

The completed SWTL/extended discone mounted on top of the aluminum vertical can be seen in Photo 5. In this picture, the plastic spacers of the center conductor look dark instead of clear only because the protective paper had not yet been removed.

HF/VHF Diplexer Construction

It isn't necessary to build an HF/VHF diplexer in order to use the All-Band Antenna, but when used use with a radio that has separate HF and VHF connections or with separate HF and VHF radios, it can allow all-band operation without requiring any switching. It provides a means of connecting to the 50 Ω VHF-microwave connection from the bottom launcher at the same time a suitable matching section is being used for HF operation. By separating these connections, the antenna is always ready for either or both HF and VHF+ operation. We use the All-Band Antenna with ICOM IC-706 MKIIG transceivers, which cover 160 m through 6 m using one coax connection and 2 m and 70 cm on a second. Of course, a pair of these diplexers could also be used to double-up on a single piece of coax with this or with other types of antennas. For example, a very wide range spectrum analyzer, receiver or transceiver that covers HF through microwave could be fed from a single

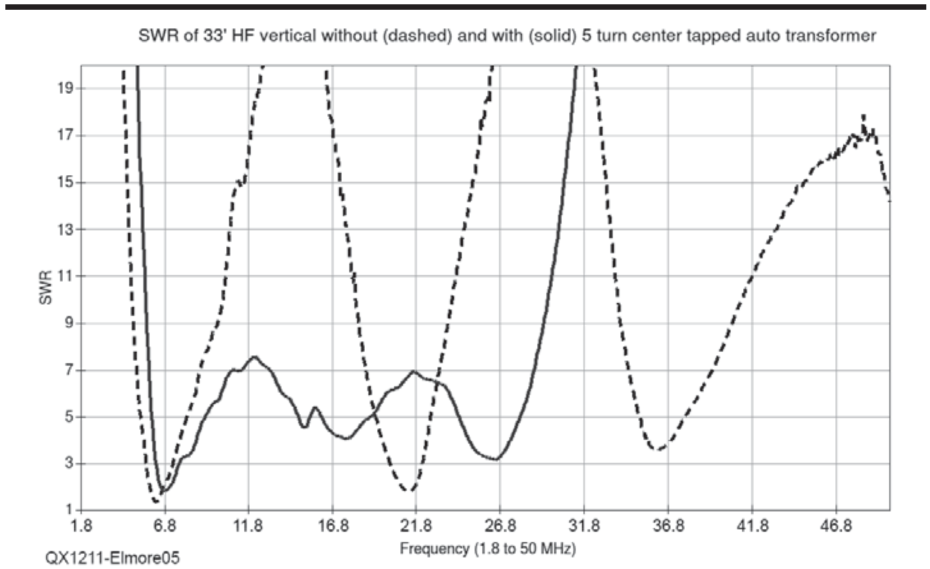


Figure 5 — Measured HF SWR of the 33 foot vertical (with bottom SWTL launcher but without top launcher/discone) without (dashed line) and with (solid line) the auto-transformer coarse matching inductor. The reference impedance is 50 Ω. The inductor improves the SWR presented to the antenna tuner over most of the HF range, to the extent that most automatic antenna tuners can achieve good match on the HF amateur bands. Operation on the 160 m and 6 m band may require different coarse matching inductance or to be operated without any at all in order to achieve 1:1 SWR with some antenna tuners.

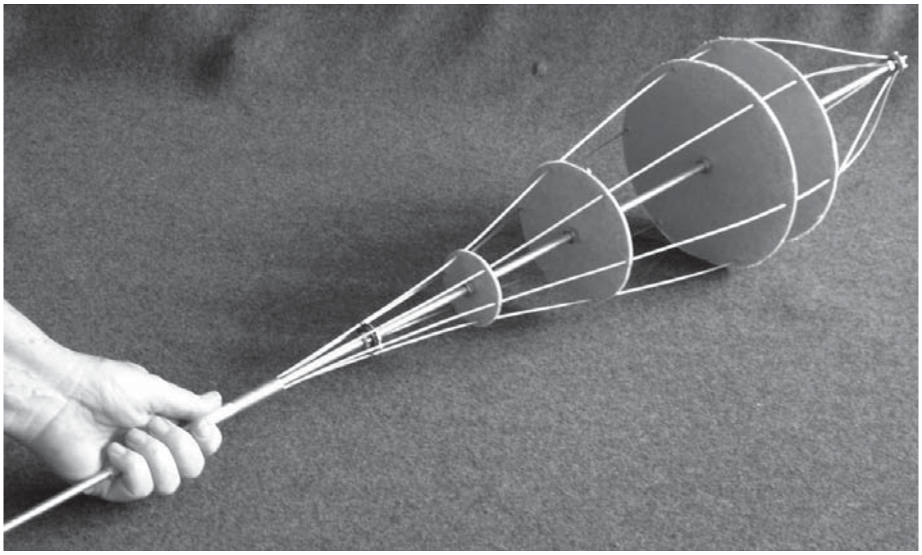


Photo 3 — In conjunction with a conical outer conductor the tapered center conductor of the integrated top launcher uses a combination of metal and plastic spacers to conform it's shape to produce the correct TEM impedance to produce a broadband Klopfenstein coaxial transformer. The conductor is assembled around a central ¼ inch threaded steel rod, which Is extended by a short section of ¼ inch brass rod near the wide end. This brass rod extends through a hole in a PVC plastic cap that the top disc/cylinder is built around (not shown).

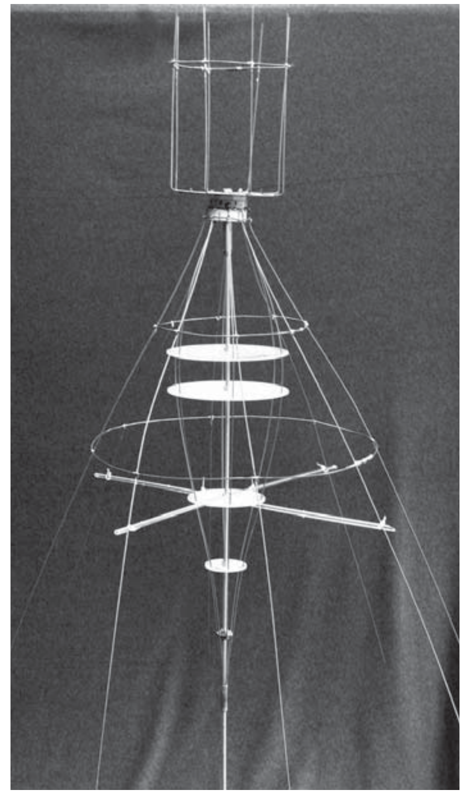


Photo 4 — The top wire cage cylinder is built on a PVC plastic reducer that mates with a short section of plastic pipe on which the cone is constructed. The cone is insulated from both the center conductor and the top cylinder. The central threaded rod from the center conductor of the launcher attaches to a copper disk that, along with the brazing rod, makes up the top cylinder. A single set screw is enough to secure the cone to the reducer on the cylinder and two screws attach the disc to the reducer.

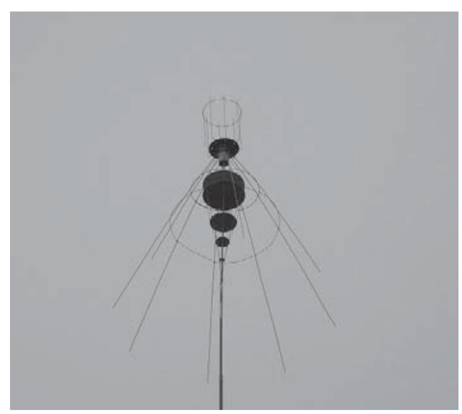


Photo 5 — The extended discone antenna with integrated SWTL launcher mounted atop the HF vertical.

coaxial connection this way.

The goal of this diplexer design was just to protect a second radio from RF energy being transmitted by the first. It was not intended as a low-pass filter for HF or a high-pass filter for VHF/UHF. If this additional functionality is desired, more filtering can still be placed between the radio and the diplexer. By using this diplexer and an automatic antenna tuner for HF with the ICOM transceivers, complete all-band and even automated operation over the entire range of the radio is possible. We can run frequency-hopping WSPR on 160 m through 432 MHz completely automatically this way.

It may also be desirable to insert a high-pass filter in the HF side of the radio. At N6GN, a vertical such as this can deliver a significant portion of one watt from a local AM broadcast station and this causes problems with160 m operation. An additional high-pass filter to protect the receiver from this sort of problem can be inserted between the radio and the diplexer's HF input.

Construction of the diplexer isn't particularly difficult but in order to get good UHF performance, good connectors and a micro-strip transmission line is important. We used surplus Mini-Circuits bias tees as a starting point because they provided a nice package with good sturdy connectors and exactly the internal circuit board micro-strip transmission line we wanted. The bias tee components were removed and replaced with the inductors and capacitors shown in Figure 2. If you don't use the Mini-Circuits package as seen in Photo 6, you should be able to build your own package from scratch by mounting appropriate connectors on package walls made from double side PC board and cutting a piece of the same double-clad material to shape so that the connector grounds can be soldered to the bottom side and the center pins laid and soldered directly on the board trace. For common $\frac{1}{16}$ inch epoxy board, 50 Ω micro-strip will be a trace about 0.110 inch wide. Really, only the UHF diplexer path needs to be made in this manner and normal lumped techniques and point-to-point wiring can be used on the HF side.

Because the impedance at the base of the All-Band Antenna can be high, so can the RF voltage, even when not driven by a kilowatt transmitter. To withstand this, we used multiple surface mount mica capacitors connected in series. Otherwise, there's nothing special about the components. As shown, the diplexer should be able to easily handle 200 W, even after an antenna tuner and the 1:4 transformer have transformed a 50 Ω transmitter to the impedance required to match the load presented by the antenna. If you contemplate higher power operation, you should calculate or measure to be sure that you won't exceed the ratings of any of the matching components in the antenna tuner, 1:4 transformer or diplexer. You should also verify that the diplexer limits the unwanted power at the other output to an acceptable level.

Wind the inductors exactly as indicated in Figure 2 and you'll obtain the indicated inductance. Mount them with minimum lead length to the surface of the 50 Ω micro-strip.

Lay the capacitors across gaps cut in the 50 Ω micro-strip line.

Measured results of the finished diplexer made this way are so close to the modeled results shown in Figure 2 that we haven't bothered to show them separately.

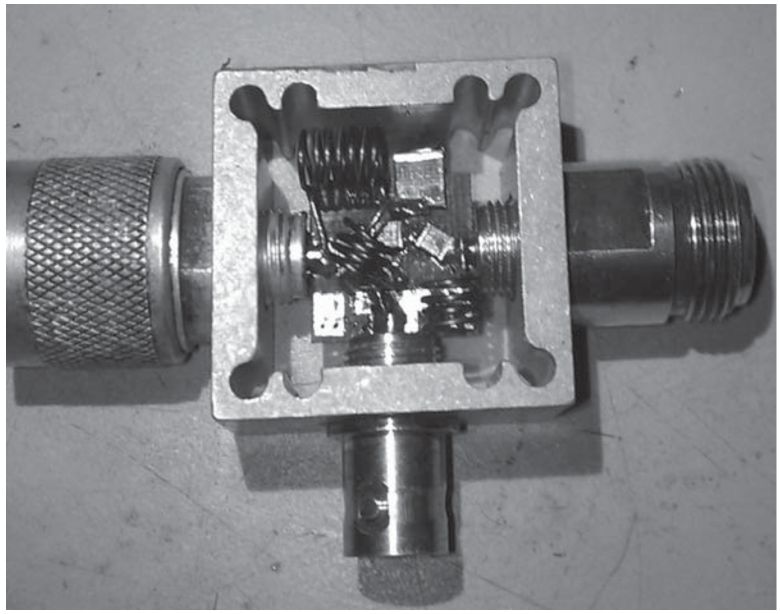


Photo 6 — The HF/VHF+ diplexer seen here was constructed from a surplus Mini-circuits ZFBT-2G-1 bias tee package which already had connectors and a circuit board with 50 Ω micro-strip transmission line.

Impedance Matching

The setup shown in Photo 7 was used to measure the antenna. Various SWR measurements with a 50 Ω reference impedance are plotted in Figures 3, 4 and 5.

Figure 3 shows the feed-point impedance of the HF vertical in the absence of any launchers. This is just a simple monopole-over-ground operated with an 8 foot ground rod, but the image plane (ground) has been further improved for higher frequencies by adding a 2 foot diameter disk at the base, as shown in Photo 7. The transmission line nature of a monopole is particularly obvious where the ground is good. At lower frequencies the impedance is increased due to imperfect conductivity. A larger disk or radial system could improve this. In use, the coarse tuning coil or transformer shifts the center of these circles to more nearly coincide with the 50 Ω impedance of coaxial cable. This measurement was shown in the previous article.[4]

The All-Band Antenna, including the effect of the SWTL at frequencies above the launcher low frequency cut-off is easy to see in Figure 4 and a relatively good match is available for all of VHF and above. The compromises made to achieve acceptable launcher dimensions have hurt the match slightly but the impact on communications is minimal.

Above the launcher cutoff frequency the antenna ceases to act like a monopole. In the transition region between 60 and 100 MHz, however, the All-Band structure is operating partially as a normal vertical and partially as an extended discone fed with an SWTL. In operation, this transition can be observed by the relative strength of FM broadcast stations at 88 MHz as compared to those near 108 MHz, with the higher end stations somewhat stronger as the discone takes over. Had our 1:4 transformer worked better near 100 MHz this difference might have been reduced.

On HF, we used a coarse matching inductor as a 1:4 auto-transformer between the low pass output of the diplexer and the automatic antenna tuner, to transform the HF antenna impedance nearer to 50 Ω. With no transformation, the impedance rotates about a central point on the order of 120 to 200 Ω. By providing the 4:1 impedance step down, SWR and variation of SWR can both be reduced over the entire HF range of the antenna.

The transformer we used is simply a center-tapped air coil with a 1:2 turns ratio and provides a 1:4 impedance transformation along with some leakage inductance. At low HF, this transformation and inductance is helpful to move the nominal center of impedance of the vertical nearer to 50 Ω and thereby reduces the SWR that the antenna tuner needs to accommodate. The best inductor value is somewhat a trade-off between being large enough to improve the 160 m and 80 m impedance match, while not having so much leakage inductance that it worsens the match at 6 m. Our coil was made by winding five turns of no. 14 AWG bare copper wire on a 4 inch diameter Styrofoam form. It was 1¼ inches long and center tapped. This gives a total inductance of a little less than 4 µH, and with the center tap it acts as a 2:1 transformer with a K factor of about 0.8.

Near the low impedance quarter wave resonances, ground resistance may influence the SWR, but with the 33 foot length "top loaded" by the extended discone, none of the amateur bands should show much of this effect. As already described, one advantage of the All-Band antenna is that it need not be operated near an odd quarter wave resonance, and so it can be set to present a higher impedance in the higher HF bands so that even with poor grounds a simple ground rod is adequate to achieve efficient match.

Figure 5 shows the SWR of the vertical before and after addition of the air core auto-transformer. For 160 m operation, a larger inductance might be needed with some tuners and for 6 m operation the auto-transformer could probably be removed entirely, depending upon the capability of the antenna tuner used. But using only the air core auto-transformer and an LDG IT-100 automatic antenna tuner, we were able to achieve a match at least as good as 1.5:1 on all amateur bands from 160 m to 6 m.

Generally the tuner should be placed as close to the auto-transformer and diplexer as possible. Cable length between the tuner and the N connector adds capacitance, which is not what is needed to improve the 160 m match, which is high impedance and already quite capacitive. Coaxial cable between the tuner and the transformer is somewhat less of a problem than between the diplexer and N connector, but length should still be minimized, even though this will no doubt require a weatherproof enclosure for the tuner. At N6GN, after Photo 8 was taken, the tuner was mounted along with the 1:4 transformer in a plastic NEMA enclosure right at the base of the antenna.

Performance and Use

This antenna performs well on both HF and VHF. Although we've only used it for a short time, it is a pleasure to operate WSPR and span bands from low HF all the way through 432 MHz — the full range of our ICOM IC-706 MKIIG transceivers.

Signal reports on HF appear to be typical for a ground mounted vertical. For low angle communications, which can provide particularly long DX at the MUF just as a band is opening or closing, this antenna consistently beats a horizontal dipole by a wide margin. For stateside QSOs from California it doesn't have the high angle component of a low horizontal antenna but we have no trouble working US stations with it. It is not uncommon to have WSPR spots from all seven continents in a 24 hour period with it.

As with any HF antenna where the ground characteristics in the far-field affect the take-off angle, its pattern may vary seasonally due to changes in the ground surrounding the QTH. The lower take-off angle can also make it much more susceptible to local suburban neighborhood QRN, which generally seems to come from near the horizon. Because of this low angle response, in a suburban QTH, this antenna can be noisier on receive than other antenna types at the same time that it works better on transmit.

Performance of the All-Band Antenna on VHF and above has been excellent. It appears to act very nearly the same as a ground plane, dipole or discone located at the same height above ground. On 432 MHz, signals are about 10 dB stronger on both transmit and receive when compared to an omnidirectional vertical antenna mounted at roof height. The attenuation of the bottom launcher, vertical/SWTL and integrated top launcher appears to be no more than a few dB and competitive with a similar length of common coaxial cable.

Photo 7 — The All-Band Antenna measurement setup is shown In these photographs. On the left it is being operated only as a simple vertical with good grounding. For the photograph on the right the SWTL launchers and VHF+ antenna are present. In addition to the hexagonal aluminum foil disk, which provides a good VHF ground, an 8 foot ground rod near the wooden post is also connected to the disk. This arrangement was left in place for all measurements even though the disk is not necessary when the launcher is in place.

Modifications and Improvements

It seems that any good project always inspires ideas for changes to make it better, more useful and more fun. We think this project is no exception. While we are pleased with the results we've experienced building and using the All-Band Antenna as shown, other Amateur Radio operators will certainly have different needs and desires and, we hope, will want to experiment. The following are a few possible changes we've thought of so far:

1) Weatherize.

Although the Photos show the antenna free-standing, for all-weather use in most climates this antenna will probably need to be guyed. As a guy line, don't use anything that is conductive within a few feet of the attachment point on the vertical, because doing so could interfere with the SWTL operation of the antenna — if not with the HF performance as well. If you have to use a conductive guy beyond this region, break its length up in a non-periodic fashion as you would for any HF vertical by using suitable insulators. Except for fair weather, we'd recommend that you use Dacron or similar non-conductive twine. We also found it necessary to drill holes in the copper flange at the N connector on the bottom launcher to keep rain water from accumulating at the feed point.

2) Use Noise Cancellation.

Particularly in suburban locations, the noise level on receive is sometimes higher with a vertical than with a horizontal antenna, but signals seem bigger too. Because of the lower take-off angle, WSPR often reports little improvement in S/N ratio on distant stations in the presence of the increased noise — indicating that the vertical is actually capturing more signal power. A future project is to add noise cancellation to further improve the reception of weak signals. With the improved performance of the vertical and if the noise floor can be reduced, the All-Band Antenna may be one of the best multi-band solutions possible — short of multiple large and highly directive arrays.

3) Improved 1:4 Transformer.

A toroidal ferrite 2:1 voltage transformer might be an alternative to the air core inductor we used. A W2AU 2:1 current balun works fine at 160 m and 80 m but not as well at the high end of HF and 6 m. With some antenna tuners it may still be possible to match using it. A better solution might be a toroidal transformer with fewer turns wound on a lower permeability core. This could reduce leakage inductance yet still be adequate for 160 m operation while not operating above self-resonance at 6 m. Band-

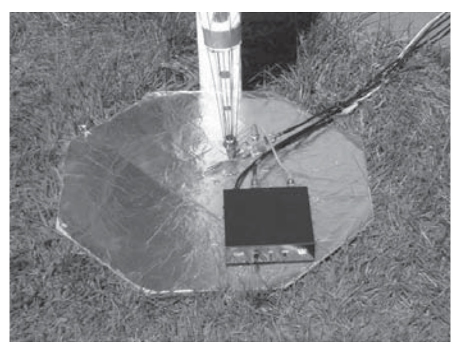


Photo 8 — Bottom of the completed All-Band Antenna with diplexer and antenna tuner. Two coaxial cables and one control cable go to the ham shack. The metal foil hexagon is not necessary for normal operation but was left from the measurements shown in Photo 7 and Figures 3 and 4. For normal operation the foil disk is replaced with a simple coax-braid jumper since the launcher provides ground reference for VHF and only a relatively high-impedance connection is required for HF as long as the antenna is not operated near an odd quarter wave resonance.

switched matching transformers are another possibility.

4) Make it Taller.

Because operation at resonance is not necessary, there is no particular length required for the HF section. K6PZB has been experimenting with a 43 foot design. Making the HF section longer reduces the burden on the matching network at the low frequency end. Generally speaking, the impedance gets higher and capacitive at frequencies below one quarter-wavelength. Extra length has negligible effect on VHF+ operation except for the great benefit from the "height gain."

5) Make it Shorter.

If low frequency operation can be sacrificed, the matching over most of the amateur HF bands, certainly 30 m and higher frequency is easy even with a vertical considerably shorter than 33 feet. On both HF and VHF, height is useful to get the radiating regions above surrounding clutter but in some situations this may be less of a problem and a short antenna may be a desirable alternative.

6) Operate HF-Only.

The SWTL and VHF portion of the antenna and the diplexer can be eliminated and the result operated only as a conventional vertical, but with broadband matching. If a really good planar ground is used near the base, like the one shown in Photo 7, but perhaps ten feet in diameter, and with short, direct connections between a flange N connector mounted in its center and the bottom of the antenna tubing, above about 90% of the quarter-wave length frequency the impedance can be transformed down such that the SWR is almost constant and nowhere much greater than 3:1. This is within easy range of almost any antenna tuner.

7) Move the Transition Frequency.

We chose to make the HF/SWTL transition between 6 m and 2 m. This transition could have been placed elsewhere, however. Pushing it higher reduces the effect of the SWTL launchers on HF operation and reduces negative visual impact and wind loading. Pushing it lower gives broader VHF-and-above operation with the top antenna.

8) Replace the Discone with Single-Band Antenna(s).

We have built single band antennas with built-in SWTL launchers and successfully used them with SWTL feed. One of the first narrow-band antennas we made was a metalized paper halo antenna for 432.1 MHz horizontally polarized operation on SSB and CW. This was to match the polarization of the UHF DX and terrestrial weak signal operations in our area. A 50 Ω connector can be placed where the discone cylinder attaches. and used to allow easy VHF-and-above antenna changes.

9) Integrate Coarse and Fine Matching.

Operate the antenna on HF only, as described above plus automatically switch coarse tune inductors and 2 to 3 transformers to cover 137 kHz to 144 MHz.

10) Build 2 or 3 HF-only antennas spaced by their height and add an automatic phasing network to produce 6 to 9 dB of wide band gain along with electronic steering.

These are just a few of the alterations that can be considered. Hopefully you will build and use this antenna or one similar to it and think of more changes for yourself. If you do, please contact us and let us know. We would like to learn from both your successes and failures.

Permission to Use

The surface wave transmission line technology described here is patented and requires licensing agreements to build or use. Corridor Systems Inc, the patent holder, is permitting licensed Amateur Radio operators worldwide to build and deploy devices and systems that use it for their personal, non-commercial use, under the terms of their Amateur Radio licenses. Any other use requires licensing from Corridor Systems Inc, 3800 Rolling Oaks Road, Santa Rosa, California 95404, USA.[6]

Glenn Elmore, N6GN, has been a licensed Radio Amateur for the past 51 years, and has held call signs of WV6STS, WA6STS and now N6GN. He has held an Amateur Extra class license since 1972. For most of his working career, Glenn has been an electrical engineer involved with the design of RF and microwave test and measurement equipment, notably scalar, vector network and spectrum analyzers.

Glenn's Amateur Radio interests have included weak signal VHF/microwave operation including meteor scatter, EME, terrestrial DX as well as higher speed Amateur TCP/IP radios and networks. He has recently been active on WSPR, the weak signal reporting network. Glenn is an ARRL Member.

John Watrous, K6PZB, is an ARRL Member who was first licensed in 1956. Several times he has won the San Francisco VHF Contest in the QRP category. He is active in WSPR and has been working with Glenn Elmore on radio projects for over 20 years. For 34 years he taught people to behave more creatively in an Art Department at Santa Rosa Junior College, where he first used computers in art in 1983. He initiated the college's first on-line class in 1995. Retiring in 2007, John has focused his energy toward brainstorming ideas with Glenn, and building models in his shop. John holds a Masters Degree in sculpture and has always been interested in art and technology.

Notes

[1]Glenn Elmore, N6GN, and John Watrous, K6PZB, "The Mercury Capsule, A Light Weight Broadband Antenna," ARRL Pacificon, 2011, San Jose, CA.

[2]Glenn Elmore, N6GN, and John Watrous, K6PZB, "The Flying Antenna". **www.youtube.com/watch?v=-VWBUDJv2n0**. [This video shows operation of a balloon supported antenna, which is a potentially dangerous and even lethal activity that *QEX* suggests should not be attempted. — Ed.]

[3]Glenn Elmore, N6GN, and John Watrous, K6PZB, "A Surface Wave Transmission Line," *QEX*, May/Jun 2012, pp 3-9.

[4]Glenn Elmore, N6GN, "A New Antenna Model," *QEX*, Jul/Aug 2012, pp 8-18.

[5]Glenn Elmore, N6GN, "Physical Layer Considerations in Building an Amateur Radio Network," 9th ARRL Computer Networking Conference Proceedings.

[6]You can find more information about the patent and this license for Amateur Radio operators at Corridor Systems' website: **www.corridorsystems.com**.

QEX

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APRS Unveiled

All the sneaky bit-level details of APRS messages...with an example packet.

Anyone attempting to create APRS equipment "from scratch" has immediately confronted the lack of a complete, detailed summary of all the APRS message requirements. Finding these details requires considerable effort, more than a little "luck," and a vocabulary that an APRS "novice" simply won't have. The lack of a simple, concise (and complete) summary has probably thwarted many attempts to create APRS technology.

This article (hopefully) addresses that problem. It provides a detailed description of a typical APRS message string, byte by byte, with a complete example provided. Basically, there is enough "message protocol" information to enable the creation of a circuit that can plug into the MIC jack of a non-APRS radio, and generate APRS packets that will successfully propagate across an APRS network.

The guidelines described here have been tested and proven. I used them to develop a 2 m APRS beacon transmitter design that has been in service for 2 years, and which has been used by several people. You can find more information about the beacon itself on my website: **www.silcom.com/~pelican2/PicoDopp/XDOPP.htm#MBCN**.

The Bell 202 Modulation Method

APRS data is transmitted at a 1200 baud data rate, typically on the (US) national APRS channel of 144.390 MHz. The data is frequency-modulated onto the RF carrier with two audio tones (1200 / 2200 Hz) that comply with a modified version of the Bell 202 modem standard. At the moment of data bit transition, a logic "zero" data bit is signified by "flipping" between tones, (for example, 1200 to 2200, or vice versa) whereas a logic "one" data bit is signified by no "flip" (steady frequency, either 1200 or 2200 Hz)

The Bell 202 standard specifies that the tone "flip" must be "phase contiguous," which basically means that the transition between tones must be as smooth as possible. The phase angle of the audio waveform (at the instant of tone switching) must be preserved and used as the "starting" phase angle for the new tone waveform. This minimizes switching transients and reduces the required bandwidth of the signal, resulting in improved signal to noise ratio (SNR). An example modulation waveform is shown in Figure 1.

Octets Versus Bytes

APRS message bytes are transmitted with no START or STOP bits, (each byte is called an octet) which is different from regular RS232 bytes. Most of the message data is encoded using ordinary ASCII characters, but there are some exceptions to this rule, described later. In all cases, octet (byte) data bits are transmitted least significant bit (LSB) first. (The bit order is B0 to B7.)

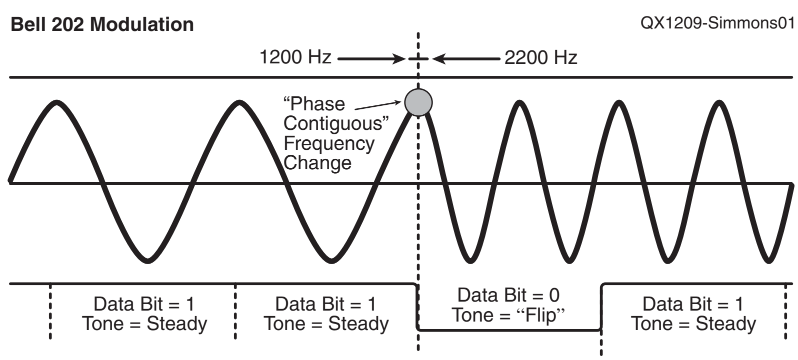

Figure 1 — This graph illustrates the Bell 202 Modem modulation specification for the transition between tone frequencies to represent a 0 data bit.

APRS Position Reports: The Message Format

The APRS "example" message provided in this article complies with the format identified on page 33 of the APRS specification document, version 1.0.1, dated 29 August 2000.[1] In that document, (page 33) this message format is identified as: "Lat/Long Position Report Format with Data Extension (no Timestamp)" Byte by byte, the transmitted message has this format:

FDDDDDDdSSSSSSsVVVVVVvVVVVVVvCPsLLLLLLLHsLLLLLLLLHsCCCsSSSCcFF

With each byte defined as shown below:
(begin transmission)
(NOTE: Tone modulation begins IMMEDIATELY when transmission begins… no "dead carrier" time is provided.)

Starting Flag Bytes	(1 byte minimum, usually several identical bytes)	
F	Flag byte, always = 0x7E	
Preamble Bytes		
DDDDDDd	Destination addr	(7 bytes)("APDF00"+ SSID = "0" in this example)
SSSSSSs		
Source addr	(7 bytes = 6 call sign bytes + 1 SSID byte)	
VIA addr (0 - 8 addr = 0 - 56 bytes)	(7 bytes each = 6 call sign + 1 SSID byte)	
	(NOTE: this example uses only 2 VIAs, shown below)	
VVVVVVv	VIA 1 address + SSID (optional, 6 + 1 bytes)	
VVVVVVv	VIA 2 address + SSID (optional, 6 + 1 bytes)	
(end of preamble bytes)		
Control and Protocol Bytes		
C	Control field	(1 byte, always = 0x03)
P	Protocol field	(1 byte, always = 0xF0)
(end of control and protocol bytes)		
Information Field	(APRS Position report, no timestamp, no messaging)	
s	Symbol	(1 byte)(APRS message type identifier)
	(! = exclamation mark = no APRS messaging, no time stamp)	
LLLLLLLH	Latitude	(8 bytes, XXxx.xxH)
		(XX = degrees latitude, 00 to 89)
		(xx.xx = minutes + dp + decimal minutes latitude)
		(H = hemisphere, N or S)
s	Symbol	(1 byte)(= primary or alternate APRS symbol table)
	(This identifies the type of APRS map icon to be displayed)	
LLLLLLLLH	Longitude	(9 bytes, XXXxx.xxH)
		(XXX = degrees longitude, 000 to 179)
		(xx.xx minutes + dp + decimal minutes longitude)
		(H = hemisphere, E or W)
s	Symbol	(1 byte)(= map symbol displayed on APRS screens)
(data extension begins here: COURSE and SPEED)		
CCC	Course	(3 bytes, xxx = 001-359, true degrees, 000 = stationary)
s	Symbol	(1 byte)(delimiter = "/")
SSS	Speed	(3 bytes, xxx = 000-999 knots)
(end of data extension)		
C	Comment	(0 - 36 bytes, 1 byte shown here)
(end of information field)		
Frame Checksum Bytes		
cc	FCS field	(2 bytes, CRC checksum, sent low byte / high byte)
(end of frame checksum bytes)		
Ending Flag Bytes		
FF	Flag	(2 bytes minimum)
(end of message…. end transmission)		

NOTE: As a courtesy to the receiving decoder (to make its job easier) it is not unusual to send several bytes of 0x00 data before sending the first FLAG byte. The pattern of several successive "0" bits causes the Bell 202 tones to constantly flip between tones, which simplifies the detection of the boundary between successive data bits, at the receiving decoder. For similar reasons, it is not unusual to send several FLAG bytes at the beginning (and end) of an APRS message, even though the APRS spec states only one FLAG byte is required.

Bear in mind that the receiver "at the other end" probably is an ordinary voice radio, with ordinary squelch circuits that will require 50 to 100 milliseconds (or more) of time to detect the presence of a signal on the channel, before any speaker audio is generated. At 1200 baud, 100 milliseconds of time equates to 15 transmitted bytes of message data… so the typical "courtesy" practice of transmitting several starting flag bytes is (probably) more important than the APRS specification indicates.

[1] Ian Wade, G3NRW, Editor, *Automatic Position Reporting System APRS Protocol Reference*, TAPR, 2000, p 33: **www.aprs.org/doc/APRS101.PDF**. [Yes, that should be "Automatic Packet Reporting System," but the title of the document was not changed. — Ed.]

Flag Bytes and Bit Stuffing

The lack of START and STOP bits in APRS messages means that some other method must be provided for an APRS decoder to "synchronize" itself with the bitstream of arriving messages.

The boundary between successive data bits can be identified by observing the 1200 / 2200 Hz tones, but the boundary between successive BYTES (end of one byte and start of next byte) must be identified by some other means. This is accomplished with special octets called FLAG bytes, consisting of a bit pattern of "01111110." (hex 0x7e = ASCII "tilde" character: ~)

This pattern (six consecutive "one" data bits) is reserved EXCLUSIVELY for FLAG bytes in the APRS specification. Therefore, any "accidental" occurrence of the same pattern (in the transmitted data) must be detected and prevented, but the data itself must (somehow) be preserved and recovered at the receiver. To accomplish this, a method is employed called "bit stuffing."

With "bit stuffing," each transmitted message is examined (bit by bit) as it is transmitted, to detect any (accidental) occurrence of five consecutive "1" bits. If such an event is detected, the sixth data bit (which might be either "1" or "0") is delayed, and a "0" bit is sent, ("stuffed" into the data stream) followed immediately by the (delayed) sixth data bit.

At the receiving end of the message, detection of 5 consecutive "1" data bits will alert the software that the following (6th) bit will determine if the data is a FLAG byte, or simply part of a regular message byte...if the 6th bit is a "1," the byte is judged to be a FLAG byte... otherwise, the 6th bit (which is a "0") will be ignored and discarded from the bitstream.

The Preamble: General Description

The message preamble includes the source, destination and VIA address bytes, and their associated SSID bytes. According to the APRS specification, the number of VIAs (which are user specified) can vary from zero to eight, but in the example provided in this article, the number is limited to two VIASs.

The SOURCE address is actually the FCC call sign of the transmitter operator, (6 characters, always spelled with CAPITOL letters) with an SSID byte appended to the end (7 bytes total...more info about SSID bytes later). If the call sign is less than 6 characters long, it is left-justified and padded with trailing ASCII "blank" characters, (0x20) followed by the SSID byte.

The DESTINATION address is not actually used in APRS… it is a legacy of packet communications, but APRS is a specialized subset of packet that does not employ this data field. Instead, it is filled with a fixed string of characters that identifies the type of software used to generate the APRS message. The string provided in this article's example was assigned to the author by Bob Bruninga, (developer of APRS) and consists of the text string "APDF00," with an SSID character of zero. This "assignment" is a matter of social courtesy, so that any problems in the resulting APRS messages can be traced back to the software author, and corrected. Anyone creating their own software should therefore contact Bob Bruninga for a similar assignment.

The VIA call signs (and their SSID characters) are optional (two are provided in this article's example). These are supplemental identifiers that provide information about the preferred signal path or direction for the message to take, and/or the preferred recipients for the message.

Typically these two VIAs are "WIDE1" (with SSID = 1) and "WIDE2." (with SSID = 2) These particular VIAs are actually requests for automatic "message relays" by any digipeater station that hears the messages.

The Preamble: SSID Bytes

SSID stands for "Secondary Station ID" (secondary station identification) SSID is encoded as a single byte that can express a number ranging from 0 to 15. Various (somewhat complex) "rules" for selection of SSID numbers are included in the APRS specification, but their actual values do not seem to be critical to message detection / propagation through the APRS network. In this article's example, the SSID for the DESTINATION station (= APDF00) is zero. For a WIDE1 VIA, this SSID should be one, and for a WIDE2 VIA, this SSID should be 2.

It is important to mention that the SSID values shown in ordinary computer displays (and in published articles) always include a hyphen character, so that "W6XYZ-0" indicates station W6XYZ with an SSID of zero, but in the actual transmitted message, no hyphen character is transmitted.

Furthermore, the SSID character is *not* an ASCII character; the SSID number is a 4-bit BINARY number, encoded into an 8-bit byte. The remaining bits are employed for other purposes and a description of the bits is provided below:

SSID bit 0 = extension bit (= 1 for last PREAMBLE field, = 0 otherwise)

SSID bits 1 to 4 = secondary station identification number (0 to 15, = "SSID" number)

SSID bits 5 and 6 = reserved, always = 1

SSID bit 7 = "control info," (C-bit) always = 0

The Preamble: Extension Bits and Byte Rotation

The APRS specification allows zero to eight VIA stations to be identified in a message, so some method must be provided to indicate how many VIAs are actually contained in any specific message. (This signals the end of the preamble block of data.) This is accomplished with the least significant bit (LSB) in ALL the preamble bytes. If the LSB (= bit 0) in a preamble SSID byte equals zero, then more preamble bytes remain in the message. If this bit equals one, no more preamble bytes remain. This bit is called the "extension bit"

This bit is often used in ordinary ASCII codes, and therefore it is not normally available for this purpose. To deal with this conflict, the ASCII codes used in the preamble (but *not* in the main message body) are limited to the 7-bit ASCII codes only (high order bit = always zero). This includes all "printable" ASCII characters, which are a subset of the entire ASCII set.

Furthermore, each ASCII byte (in the preamble only) is "rotated left" by one bit position, which is (arithmetically) equivalent to multiplying the character's binary value by two. This can be done without loss of information because the top bit of all 7-bit ASCII codes always equals zero. As a result of this "rotation," the LSB in each preamble byte is "liberated" for use as an APRS "extension bit."

For example, ASCII character "3" would normally be expressed as hex number 0x33, but in an APRS preamble, (due to the byte rotation) this would be transmitted as hex number 0x66, (if the extension bit = 0) or as hex number 0x67 (if the extension bit = 1). A table of ASCII characters with their regular and "rotated" values is in included in the APRS specification, in Appendix 3, Part 2.

ASCII "3" character	= 0x33	= 00110011	
Rotated character	= 0x66	= 01100110	(if extension bit = 0)
	= 0x67	= 01100111	(if extension bit = 1)

This "byte rotation" method is *only* applied to the preamble bytes — *not* to the entire contents of the APRS transmission. The first "rotated" byte is the first byte of the destination address, and the last "rotated" byte is the last byte of the last VIA address (SSID byte of the last VIA). If no VIAs are used, then the last "rotated" byte would be the SSID byte of the source address.

Summarizing, the extension bit in all PREAMBLE characters must be zero, EXCEPT for the VERY LAST character in the PREAMBLE, in which the extension bit must equal one.

Control and Protocol Characters

The control and protocol characters consist of two octets trans-

Table 1
Sample APRS Example Message

NAME	*VALUE*	*HEX DATA TRANSMITTED*								
(begin transmission)										
NULLS	(5X <nul>)	0x00	0x00	0x00	0x00	0x00				
FLAGS	(5X <tilde>)	0x7e	0x7e	0x7e	0x7e	0x7e				
(begin CRC calculation here)										
(begin bit stuffing here)										
(NOTE: The following bytes are left-rotated one bit position to provide bit 0 = extension bit)										
DESTINATION	APDF00	0x82	0xa0	0x88	0x8c	0x60	0x60			
DEST SSID	<SSID = 0>	0x60								
SOURCE	W6XYZ<sp>	0xae	0x6c	0xb0	0xb2	0xb4	0x40			
SRC SSID	<SSID=15>	0x7e								
VIA1		WIDE1<sp>		0xae	0x92	0x88	0x8a	0x62	0x40	
VIA1 SSID	<SSID=1>	0x62								
VIA2		WIDE2<sp>		0xae	0x92	0x88	0x8a	0x64	0x40	
VIA2 SSID	<SSID=2>	0x65								
(end left-rotation)										
CTRL CHAR	<control>	0x03								
PROTO CHAR	<protocol>	0xf0								
MSG TYPE	<msg type>	0x21								
LATITUDE	3426.22N	0x33	0x34	0x32	0x36	0x2e	0x32	0x32	0x4e	
SYMB TABLE	<primary>	0x2f								
LONGITUDE	11943.57W	0x31	0x31	0x39	0x34	0x33	0x2e	0x35	0x37	0x57
SYMB CODE	<car>	0x3e								
COURSE	264	0x32	0x36	0x34						
DELIMITER	/	0x2f								
SPEED	000	0x30	0x30	0x30						
COMMENT	COMMENT	0x43	0x4f	0x4d	0x4d	0x35	0x4e	0x54		
(end CRC calculation)										
CRC LSB	<CRC lo byte>	0xf9								
CRC MSB	<CRC hi byte>	0x3c								
(end bit stuffing)										
FLAGS	(5X < tilde>)	0x7e	0x7e	0x7e	0x7e	0x7e				
(end of transmission)										
(total = 81 bytes = 540 ms at 1200 baud)										

mitted immediately after the preamble. The control octet is transmitted first, and always consists of 0x0f. The protocol octet is transmitted next, always consisting of 0xf0. (No explanation is offered here for their purpose.)

Message Body (Information Field)

The message body (called the "information field" in the APRS spec) has various forms, depending on the type of APRS message being transmitted. The format of the data contained in this field is identified by the very first character, ("symbol") and different APRS messages use different characters for this field. (Refer to the APRS specification.)

APRS Map Symbol

The APRS map symbol is identified with two ASCII bytes located in the message body. One is located immediately after the latitude data field, the other immediately after the longitude data field. These two bytes are defined in the APRS specification, in Appendix 2.

The first character identifies one of two "symbol tables" in the appendix, (PRIMARY or ALTERNATE) each containing 93 "symbols" that will be shown on a map display when the message is received. The second character identifies one of the 93 symbols in the associated table.

Data Extensions

Data extensions are optional 7-byte fields that (if employed) express additional information, as described in Chapter 7 of the APRS specification. In this example, a data extension is employed to express the COURSE and SPEED of the reporting station.

Comment Field

The COMMENT field is optional. The maximum allowed length of the COMMENT field varies depending on the type of APRS message being sent. (See the details in the APRS specification for a particular message type.)

Frame Checksum

The frame checksum is calculated using a CRC calculation method. CRC refers to "Cyclic Redundancy Check," which consists of a special two byte "checksum" that allows the integrity of the message data to be tested, after it is received. The CRC checksum (= frame checksum) is generated when each message is transmitted, and evaluated at the destination, when the message is received.

The CRC checksum generation is performed by examining each byte in the transmitted message, using a special "formula" that is applied to each bit in the message. The result of this special "formula" is a two byte number that expresses the CRC (frame) checksum.

Bits that are added to the bitstream as a result of "bit stuffing" are *not* included in the calculation of the CRC checksum. The two-byte checksum itself is also excluded from the calculation.

CRC checksum calculation begins with the first byte in the PREAMBLE block, (immediately after the last starting FLAG) and ends with the last byte in the COMMENT block. The two CRC bytes are then transmitted LSB / MSB (low byte first, then high byte).

Rather than re-explaining it here in the author's own words, I defer to the source where I learned of it myself — many thanks to Scott Miller, N1VG, for posting this simple and concise explanation of the CRC checksum calculation method on his website:

Frame Check Sequence

One detail of the AX.25 format that deserves attention is the Frame Check Sequence (FCS) checksum. This is a two-byte checksum added to the end of every frame. It's generated using the CRC-CCITT polynomial, and is sent low-byte first.

The CRC-CCITT algorithm has plenty of published code examples, but the one I needed, and had trouble finding, was the algorithm for calculating the FCS one bit at a time, rather than a byte at a time. That algorithm is as follows:

Start with the 16-bit FCS set to 0xffff. For each data bit sent, shift the FCS value right one bit. If the bit that was shifted off (formerly bit 1) was not equal to the bit being sent, exclusive-OR the FCS value with 0x8408. After the last data bit, take the ones complement (inverse) of the FCS value and send it low-byte first.

NOTE: this text (and more useful information) can be found at Scott Miller's website, at: **http://n1vg.net/packet/index.php**

Those who choose to double-check this information against the AX.25 protocol specification, AX.25.2.2, dated July 1998, will find in section 3.8 that the order of bit transmission for the FCS data bits is opposite to that for the rest of the packet data, that is, the FCS bits (in the spec) should be transmitted most significant bit first (bit order B15 to B0 for the two FCS bytes) whereas the bit order for all other packet bytes should be sent least significant bit first (bit order = B0 to B7).

This contradicts the author's experience, in which successful on-air tests (and iGate postings) of the beacon transmitter's APRS packets used FCS data transmitted LSB first, just like the rest of the APRS packet data. There also is no mention of this "reversed" bit order in Scott Miller's comments on the topic, so it seems that the AX.25 spec is "suspect," on this point.

A Message Example

The message example given in Table 1 expresses a complete APRS message generated in compliance with the guidelines described in this article. For clarity, bits added as a result of "bit stuffing" are not shown in this data. Because a few of the bytes consist of unprintable ASCII characters, the data here is expressed in hexadecimal notation.

Bob Simmons, WB6EYV, was first licensed as a novice in 1964 at age 13, and remained licensed (more or less) constantly ever since. He also earned a commercial FCC license in 1967. He served Naval Reserve duty as a radar technician (ETR2) with about 6 months of total sea time. He spent several years of civilian work in nautical and marine electronics in Los Angeles harbor, as well as doing some land mobile radio work, followed by 5 years in flight line avionics, working on business jets. He moved to Santa Barbara, CA in 1992 and worked on vacuum deposition systems for 5 years, and held assorted odd engineering jobs at other times.

Presently, Bob is self employed and runs a website making and selling radio direction finding equipment and modules, with a majority of his "new" work spent creating embedded software / hardware and developing technologies to enable Internet-linked remote DF stations. His primary interest is developing and applying new technologies to old problems, and pushing the DF "art" forward.

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Tall Vertical Arrays

The author presents his EZNEC *analysis of a variety of vertical antenna arrays for 80 m.*

Many hams use phased-vertical arrays, such as the two-element cardioid or the popular 4 Square, for communications on the low bands from 40 through 160 meters. Designs for larger, more complex arrays (employing from 5 to 9 elements) are available if improved performance is desired.[1] Typically, these antennas use ¼ λ monopoles in conjunction with ground screens composed of ¼ λ radials. I was curious to see what would happen if taller (⅝ λ) monopoles and/or radials were substituted. This article discusses the results of that investigation.

Computer Simulations

The computer analysis was done on the 80 meter band at a frequency of 3650 kHz. Each vertical monopole is built from no. 12 AWG copper wire, and the ground screen includes 60 radials made of no. 16 AWG copper wire. These radials were buried to a depth of just 3 inches, in "average" soil having a conductivity of 0.005 Siemens per meter and a dielectric constant of 13. I simulated all of the antennas described in this article using the *EZNEC* software package, which is available from Roy Lewallen, W7EL.[2]

[1]Notes appear on page 35.

Results for a Single Vertical Element

Table 1 shows what happens when an isolated monopole, whose height is either ¼ λ or ⅝ λ, is installed over a buried ground screen composed of 60 radials, whose length may also be either ¼ λ or ⅝ λ. In free space, an actual quarter wavelength is 67.37 feet, while ⅝ λ amounts to 168.42 feet, so those lengths were used for the buried radials. For the vertical element, however, the height of the "¼ λ" monopole was adjusted to resonate the antenna (input reactance equal to or close to zero) at 3650 kHz. In a similar fashion, the height of the "⅝ λ" element was trimmed to produce maximum gain at the lowest elevation angle, in combination with the shortest possible monopole. (Throughout this analysis, radial lengths and radiator heights were always varied in increments of 0.01 foot.)

Table 1 displays some interesting results. First, we see that the resonant height of a nominal "¼ λ" element is slightly shorter when longer radials are employed. Also, the height of the "⅝ λ" monopole that generates maximum gain is significantly reduced when it is placed over a ground screen using longer radials.

When the nominal element height is fixed, the installation of a larger ground screen yields more gain: from 0.39 dBi to 1.00 dBi for the ¼ λ monopole, and from 0.73 dBi to 1.00 dBi for the ⅝ λ radiator. Figures 1 and 2 show the elevation plane radiation patterns for the two cases.

When the radius of the ground screen is fixed at 0.25 λ, then upgrading to a taller ele-

Table 1

Performance of a single vertical monopole antenna with ground screen, as a function of radiator height and radial length. Each antenna is designed to operate at 3650 kHz. Antennas with ¼ λ monopoles have their element height adjusted for resonance at 3650 kHz, while those with ⅝ λ elements are adjusted for maximum gain and lowest take-off angle at the same frequency. Each monopole is built from no. 12 AWG copper wire, while the ground system is composed of 60 no. 16 AWG wire radials. The soil is "average" (conductivity = 0.005 Siemens/meter and dielectric constant = 13).

	¼ λ Monopole ¼ λ Radials	*¼ λ Monopole ⅝ λ Radials*	*⅝ λ Monopole ¼ λ Radials*	*⅝ λ Monopole ⅝ λ Radials*
Radiator Height (ft)	65.46	65.38	151.32	142.48
Radial Length (ft)	67.37	168.42	67.37	168.42
Resonant Frequency (kHz)	3650	3650	1586	1679
Maximum Gain (dBi)	0.39	1.00	0.73	1.00
Take-off Angle (°)	24.9	26.3	15.2	16.5
Z_{input} (Ω)	38.1	40.56	281.6 – *j* 990.5	769.3 – *j* 1479

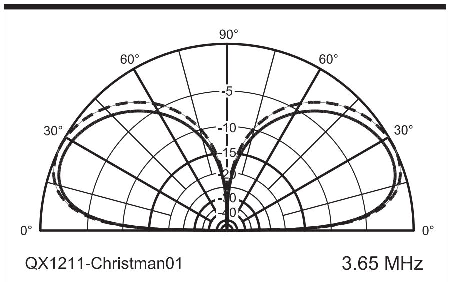

Figure 1 — Elevation-plane radiation patterns for a resonant ¼ λ vertical-monopole antenna, when placed over a ground screen composed of 60 buried radials. Solid trace = ¼ λ radials (L = 67.37 ft at 3650 kHz), peak gain = 0.39 dBi at 24.9°take-off angle. Dashed trace = ⅝ λ radials (L = 168.42 ft). Peak gain = 1.00 dBi at 26.3° take-off angle.

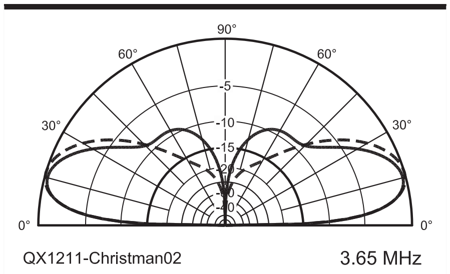

Figure 2 — Elevation-plane radiation patterns for a nominal ⅝ λ vertical-monopole antenna, when placed over a ground screen composed of 60 buried radials. The height of each element was adjusted for maximum gain. Solid trace = ¼ λ radials (L = 67.37 ft at 3650 kHz), peak gain = 0.73 dBi at 15.2° take-off angle. Dashed trace = ⅝ λ radials (L = 168.42 ft), peak gain = 1.00 dBi at 16.5° take-off angle.

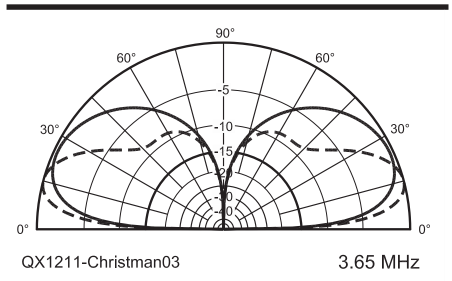

Figure 3 — Elevation-plane radiation patterns for nominal ¼ λ and ⅝ λ vertical-monopole antennas, when placed over a ground screen composed of 60 buried ¼ λ radials (L = 67.37 feet at 3650 kHz). The ¼ λ element was tuned to resonance at 3650 kHz, while the height of the 5/8 λ radiator was adjusted for maximum gain. Solid trace = ¼ λ monopole (H = 65.46 feet), peak gain = 0.39 dBi at 24.9° take-off angle. Dashed trace = ⅝ λ monopole (H = 151.32 feet), peak gain = 0.73 dBi at 15.2° take-off angle.

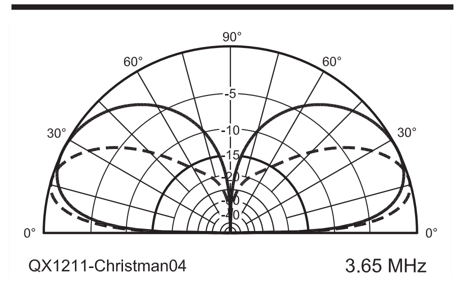

Figure 4 — Elevation-plane radiation patterns for nominal ¼ λ and ⅝ λ vertical-monopole antennas, when placed over a ground screen composed of 60 buried ⅝ λ radials (L = 168.42 feet at 3650 kHz). The ¼ λ element was tuned to resonance at 3650 kHz, while the height of the ⅝ λ radiator was adjusted for maximum gain. Solid trace = ¼ λ monopole (H = 65.46 feet), peak gain = 1.00 dBi at 26.3° take-off angle. Dashed trace = ⅝ λ monopole (H = 151.32 feet), peak gain = 1.00 dBi at 16.5° take-off angle.

ment gives us both an increase in peak gain and a lower take off angle (from 0.39 dBi at 24.9° to 0.73 dBi at 15.2°), as revealed in Figure 3. If a larger ground screen (radius = 0.625 λ) is present, installing a taller monopole still provides a lower elevation angle (from 26.3° to 16.5°), but the peak gain remains unchanged at exactly 1.0 dBi (see Figure 4).

Results for a Two element Cardioid Array with 90° Current Phasing

Figure 5 is a plan view of the ground screen for a 2 element array with ¼ λ spacing (67.37 feet) between the radiators. Each monopole has 60 radials in its ground screen, and their maximum length is also ¼ λ. Notice that this antenna uses a "broadcast style" ground screen, where none of the radials overlap one another. Instead, the radials are truncated and bonded together at those locations where they intersect, and a "common bus" links these points together, as shown in the drawing.

Table 2 lists the outcome for each combination of element height and radial length,

Table 2

Performance of a 2 element cardioid array with 90° spacing and phasing, as a function of radiator height and radial length. Each antenna is designed to operate at 3650 kHz. Arrays with ¼ λ monopoles have their element heights adjusted for resonance at 3650 kHz, while those with ⅝ λ elements are adjusted for maximum gain and lowest take-off angle at the same frequency. Each monopole is built from no. 12 AWG copper wire, while the ground system is composed of 60 no. 16 AWG wire radials. The soil is "average" (conductivity = 0.005 Siemens/meter and dielectric constant = 13).

	¼ λ Monopole ¼ λ Radials	*¼ λ Monopole ⅝ λ Radials*	*⅝ λ Monopole ¼ λ Radials*	*⅝ λ Monopole ⅝ λ Radials*
Radiator Height (ft)	65.45	65.42	158.73	158.03
Resonant Frequency (kHz)	3650	3650	1512	1514
Radial Length (ft)	67.37	168.42	67.37	168.42
Endfire Mode:	*$I_{front} = 1<-90°$ and $I_{back} = 1<0°$)*			
Maximum Gain (dBi)	3.49	3.87	3.26	3.45
Take-off Angle (°)	24.8	25.5	14.5	14.6
Elevation Plane F/B Ratio (dB)	13.60	12.26	12.88	12.84
Azimuth Plane F/B Ratio (dB)	22.92	22.76	13.18	13.09
Azimuth Plane Half Power Beamwidth (°)	177.2	182.0	198.2	197.4
Z_{input} (Ω) (Front Element)	55.04 + *j* 17.49	57.58 + *j* 18.85	172.7 – *j* 604.8	185.9 – *j* 621.6
Z_{input} (Ω) (Back Element)	21.99 – *j* 17.68	22.55 – *j* 18.96	148.3 – *j* 799.8	157.2 – *j* 828.7
Broadside Mode:	*($I_{front} = I_{back} = 1<0°$)*			
Maximum Gain (dBi)	1.72	2.31	2.12	2.31
Take-off Angle (°)	24.9	26.3	14.8	14.8
Azimuth Plane F/S Ratio (dB)	2.27	2.51	2.75	2.74
Z_{input} (Ω) (Both Elements)	56.09 – *j* 16.62	58.97 – *j* 17.57	258.0 – *j* 714.5	275.1 – *j* 739.5

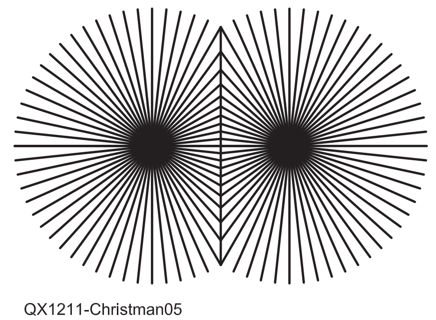

Figure 5 — Plan view of the ground screen for a 2 element vertical array. Spacing between the elements is ¼ λ, and each element has 60 buried radials, whose maximum length is also ¼ λ (67.37 feet at 3650 kHz).

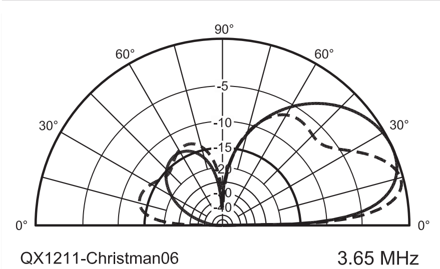

Figure 6 — Elevation-plane radiation patterns for 2 element cardioid arrays with quadrature phasing (I_{front} = 1<–90°, I_{back} = 1<0°), when using either ¼ λ or ⅝ λ monopoles. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz). The ¼ λ monopoles were tuned to resonance at 3650 kHz, while the height of the ⅝ λ radiators was adjusted for maximum gain. Solid trace = ¼ λ monopoles (H = 65.45 feet), peak gain = 3.49 dBi at 24.8° take-off angle. Dashed trace = ⅝ λ monopoles (H = 158.73 feet), peak gain = 3.26 dBi at 14.5° take-off angle.

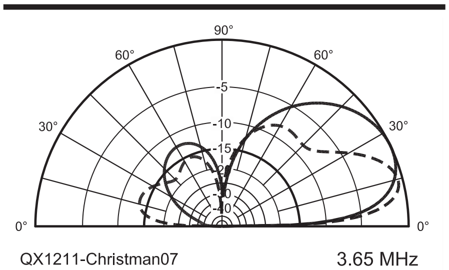

Figure 7 — Elevation plane radiation patterns for 2 element cardioid arrays with quadrature phasing (I_{front} = 1<–90°, I_{back} = 1<0°, when using either ¼ λ or 5/8 λ monopoles. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz). The ¼ λ monopoles were tuned to resonance at 3650 kHz, while the height of the ⅝ λ radiators was adjusted for maximum gain. Solid trace = ¼ λ monopoles (H = 65.42 feet), peak gain = 3.87 dBi at 25.5° take off angle. Dashed trace = ⅝ λ monopoles (H = 158.03 feet), peak gain = 3.45 dBi at 14.6° take off angle.

when the array is driven in the classic cardioid fashion, utilizing equal magnitude currents that are 90° apart in phase angle. Again we find that ¼ λ resonance occurs with a somewhat shorter monopole height (from 65.45 ft to 65.42 ft) when a larger ground screen is employed. Note that resonance was achieved by placing a *single* (isolated) radiator over the *entire* ground screen, and then adjusting its height to minimize the input reactance at 3650 kHz. For a "⅝ λ" element, maximum gain occurs at a slightly lower height (from 158.73 feet to 158.03 ft) if longer radials are used.

Surprisingly, we find that, for *either* ground screen, switching to the much taller "⅝ λ" radiator actually generates *less* gain (from 3.49 dBi to 3.26 dBi when using ¼ λ radials, and from 3.87 dBi to 3.45 dBi for ⅝ λ radials), although the take off angle still falls by roughly 10°. (As we shall see in a moment, a phase lag of 90° is far too small to generate very much gain from such tall elements.) See Figures 6 and 7 for the plots.

As expected, installing longer radials under either monopole leads to an increase in the peak forward gain: from 3.49 dBi to 3.87 dBi for the ¼ λ radiator, and from 3.26 dBi to 3.45 dBi for the ⅝ λ element. The corresponding elevation-plane patterns are displayed in Figures 8 and 9, respectively.

Table 2 also provides information for the case where the two monopoles are driven with equal-amplitude in-phase currents, which generates a *broadside* (rather than an *endfire*) radiation pattern. Using longer radials generates more gain from either the ¼ λ or the ⅝ λ element, while substituting a taller radiator (over a ground screen of either size) reduces the take-off angle by 10° or more.

Two-element Cardioid Array with Modified Current Phasing

One way to improve the performance of the traditional cardioid array is to increase the phase-lag of the current delivered to the front element. When optimized for maximum end-fire gain at the lowest-possible take-off angle, the results are as shown in Table 3. For resonant ¼ λ vertical monopoles, changing the phase-lag from 90° to 136° yields about 0.8 dB of extra gain. The taller "⅝ λ" elements generate around 1.4 dB of additional gain, although the phase-lag needs to be further increased, to around 144°. As usual, there is no "free lunch." These significant improvements in forward gain are achieved at the expense of great reductions in the front-to-back ratio, which falls to less than 10 dB in both the elevation and azimuth planes.

Using a larger ground screen under an array with elements of a fixed height leads to more end-fire gain, but at a slightly higher take-off angle (see Figures 10 and 11). Employing a taller monopole over a ground screen whose radius is held constant produces both more gain and a lower take-off angle, which can be seen in Figures 12 and 13.

As before, a *broadside* radiation pattern can be created when the two vertical elements are driven with equal-amplitude in-phase currents. (Note that the lower portion of Table 3 is nearly identical to that of Table 2, since only minor changes were made in the height of the "⅝ λ" monopoles to achieve maximum gain.) Once again, the use of longer radials generates more gain from either the ¼ λ or the ⅝ λ element, while substitut-

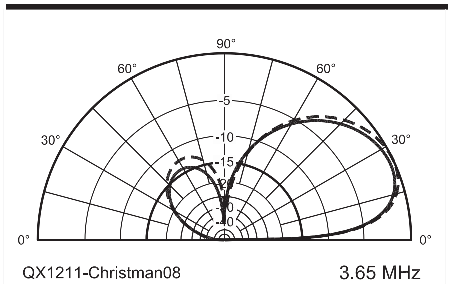

Figure 8 — Elevation-plane radiation patterns for cardioid arrays using two ¼ λ elements with quadrature phasing (I_{front} = 1<–90°, I_{back} = 1<0°), when the ground screens are composed of 60 buried radials whose maximum length is either ¼ λ or 5/8 λ. The monopoles were tuned to resonance at 3650 kHz. Solid trace = ¼ λ radials (L = 67.37 ft at 3650 kHz), peak gain = 3.49 dBi at 24.8° take-off angle. Dashed trace = ⅝ λ radials (L = 168.42 ft), peak gain = 3.87 dBi at 25.5° take-off angle.

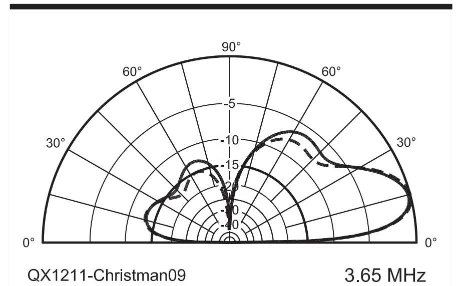

Figure 9 — Elevation-plane radiation patterns for cardioid arrays using two ⅝ λ elements with quadrature phasing (I_{front} = 1<–90°, I_{back} = 1<0°), when the ground screens are composed of 60 buried radials whose maximum length is either ¼ λ or ⅝ λ. The height of the monopoles was adjusted for maximum gain. Solid trace = ¼ λ radials (L = 67.37 ft at 3650 kHz), peak gain = 3.26 dBi at 14.5° take-off angle. Dashed trace = ⅝ λ radials (L = 168.42 ft), peak gain = 3.45 dBi at 14.6° take-off angle.

ing a taller radiator (over a ground screen of either size) reduces the take-off angle by 10° or more.

Results for a 4-Square Array with ¼ λ Elements and ¼ λ Radials

The classic four-square phased-vertical array utilizes ¼ λ elements spaced ¼ λ apart, with a large number of ¼ λ radials in the ground screen. In this article I will examine 4-squares with monopole heights of approximately ¼ and ⅝ λ, installed over ground screens composed of radials whose maximum length is either ¼ λ or ⅝ λ (60 radials per radiator). Figure 14 is a plan view of a ground screen made from ¼ λ (max) radials. Traditionally, the antenna is designed to fire along the diagonals of the square (through the corners), but it can also be configured to beam through the sides.

The first 4-square design to be reviewed is a typical array with ¼ λ monopoles and ¼ λ (max) radials. Table 4 lists the performance parameters for this antenna, when we vary the phase-angles of the equal-amplitude currents which are driven into the bases of the radiators. To begin the analysis, a single (isolated) element was placed over the entire ground screen, and its length was adjusted for resonance at 3650 kHz, which required an overall height of 65.43 feet.

The left column in Table 4 is for a normal corner-fire feed system, which employs

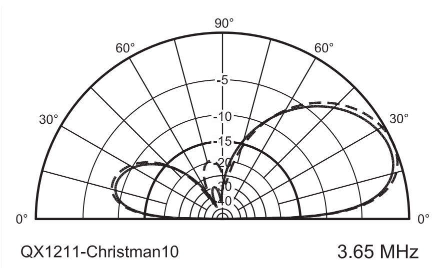

Figure 10 — Elevation-plane radiation patterns for cardioid arrays using two ¼ λ elements with phasing adjusted for maximum end-fire gain (I_{front} = 1<–136°, I_{back} = 1<0°), when the ground screens are composed of 60 buried radials whose maximum length is either ¼ λ or ⅝ λ. The monopoles were tuned to resonance at 3650 kHz.
Solid trace = ¼ λ radials (L = 67.37 ft at 3650 kHz), peak gain = 4.25 dBi at 23.1° take-off angle.
Dashed trace = ⅝ λ radials (L = 168.42 ft), peak gain = 4.71 dBi at 23.6° take-off angle.

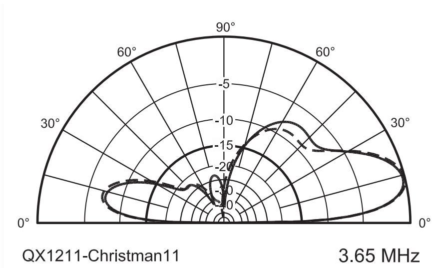

Figure 11 — Elevation-plane radiation patterns for cardioid arrays using two ⅝ λ elements with phasing adjusted for maximum end-fire gain (I_{front} = 1<–143° or 1<–144°, I_{back} = 1<0°), when the ground screens are composed of 60 buried radials whose maximum length is either ¼ λ or ⅝ λ. The height of the monopoles was adjusted for maximum gain.
Solid trace = ¼ λ radials (L = 67.37 ft at 3650 kHz), peak gain = 4.63 dBi at 13.9° take-off angle.
Dashed trace = ⅝ λ radials (L = 168.42 ft), peak gain = 4.81 dBi at 14.1° take-off angle.

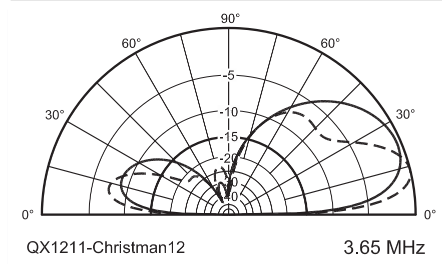

Figure 12 — Elevation-plane radiation patterns for 2 element cardioid arrays with phasing adjusted for maximum end fire gain (I_{front} = 1<–136° or 1<–143°, I_{back} = 1<0°), when using either ¼ λ or ⅝ λ monopoles. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz). The ¼ λ monopoles were tuned to resonance at 3650 kHz, while the height of the ⅝ λ radiators was adjusted for maximum gain.
Solid trace = ¼ λ monopoles (H = 65.45 feet), peak gain = 4.25 dBi at 23.1° take-off angle.
Dashed trace = ⅝ λ monopoles (H = 158.73 feet), peak gain = 4.63 dBi at 13.9° take-off angle.

Table 3

Performance of a 2 element cardioid array, as a function of radiator height and radial length. Again, the monopole spacing is 90°, but this time the current phase angles are adjusted for maximum gain at the lowest-possible elevation angle. Each antenna is designed to operate at 3650 kHz. Arrays with ¼ λ monopoles have their element heights adjusted for resonance at 3650 kHz, while those with ⅝ λ elements are adjusted for maximum gain and lowest take-off angle at the same frequency. Each monopole is built from no. 12 AWG copper wire, while the ground system is composed of 60 no. 16 AWG wire radials. The soil is "average" (conductivity = 0.005 Siemens/meter and dielectric constant = 13).

	¼ λ Monopole ¼ λ Radials	*¼ λ Monopole ⅝ λ Radials*	*⅝ λ Monopole ¼ λ Radials*	*⅝ λ Monopole ⅝ λ Radials*
Radiator Height (ft)	65.45	65.42	159.17	157.97
Resonant Frequency (kHz)	3650	3650	1508	1515
Radial Length (ft)	67.37	168.42	67.37	168.42
Endfire Mode:	*(I_{back} = 1<0°)*			
I_{front}	1<–136°	1<–136°	1<–143°	1<–144°
Maximum Gain (dBi)	4.25	4.71	4.63	4.81
Take-off Angle (°)	23.1	23.6	13.9	14.1
Elevation Plane F/B Ratio (dB)	8.32	8.21	7.19	6.91
Azimuth Plane F/B Ratio (dB)	8.58	8.50	7.24	6.95
Azimuth Plane Half Power Beamwidth (°)	129.6	132.4	130.6	130.0
Z_{input} (Ω) (Front Element)	37.34 + *j* 24.0	38.64 + *j* 25.68	87.79 – *j* 623.0	96.56 – *j* 654.2
Z_{input} (Ω) (Back Element)	14.39 – *j* 0.42	14.3 – *j* 0.58	72.33 – *j* 736.7	79.79 – *j* 776.4
Broadside Mode:	*(I_{front} = I_{back} = 1<0°)*			
Maximum Gain (dBi)	1.72	2.31	2.12	2.31
Take-off Angle (°)	24.9	26.3	14.8	14.9
Azimuth Plane F/S Ratio (dB)	2.27	2.51	2.75	2.74
Z_{input} (Ω) (Both Elements)	56.09 – *j* 16.62	58.97 – *j* 17.57	250.1 – *j* 702.9	276.3 – *j* 741.1

Table 4

Performance of a 4 Square array using ¼ λ monopoles and ¼ λ radials, as a function of the phase angles of the base currents. The height of the elements was adjusted to 65.43 feet, for resonance at 3650 kHz, while the radials have a length of 67.37 feet. Each monopole is built from no. 12 AWG copper wire, while the ground system is composed of 60 no. 16 AWG wire radials per element. The soil is "average" (conductivity = 0.005 Siemens/meter and dielectric constant = 13).

	Traditional Phasing	*W8JI Phasing*	*Maximum Gain Phasing*
Firing through the corners of the square:			
I_{front}	1<–180°	1<–240°	1<–250°
I_{sides}	1<–90°	1<–120°	1<–125°
I_{back}	1<0°	1<0°	1<0°
Maximum Gain (dBi)	5.89	6.36	6.36
Take-off Angle (°)	23.7	22.2	22.0
Elevation Plane F/B Ratio (dB)	17.99	30.3	33.96
Azimuth Plane F/B Ratio (dB)	23.35	36.99	38.35
Azimuth Plane F/S Ratio (dB)	N.A.	15.9	14.1
Azimuth Plane Half Power Beamwidth (°)	100.6	80.0	76.8
Z_{input} (Ω) (Front Element)	67.17 + *j* 54.22	30.97 + *j* 52.63	26.14 + *j* 50.08
Z_{input} (Ω) (Side Elements)	43.19 – *j* 18.82	25.54 – *j* 2.48	22.94 – j 0.08
Z_{input} (Ω)(Back Element)	1.82 – *j* 16.38	7.06 – *j* 0.99	8.08 + *j* 0.41
Firing through the sides of the square:			
I_{front}	1<–90°	1 <–105°	1<–131°
I_{back}	1<0°	1<0°	1<0°
Maximum Gain (dBi)	4.90	5.22	5.42
Take off Angle (°)	25.0	24.4	23.4
Elevation Plane F/B Ratio (dB)	13.93	17.71	9.41
Azimuth Plane F/B Ratio (dB)	23.6	21.65	9.76
Azimuth Plane Half Power Beamwidth (°)	130.0	118.2	104.1
Z_{input} (Ω) (Front Elements)	91.7 + *j* 5.71	84.81 + *j* 14.07	68.64 + *j* 23.41
Z_{input} (Ω) (Back Elements)	21.28 – *j* 38.29	16.79 – *j* 28.42	15.49 – *j* 9.79

progressive 90° phase-shifts between the back, side, and front monopoles (I_{back} = 1<0°, I_{sides} = 1<–90°, I_{front} = 1<–180°). The middle column is for the design suggested by Tom Rauch W8JI, which incorporates larger current phase-angles (I_{back} = 1<0°, I_{sides} = 1<–120°, and I_{front} = 1<–240°).[3] Finally, the right-hand column is for the situation where the phase-shifts have been optimized for *maximum* forward gain at the lowest-possible take-off angle. The outcome for this trial-and-error solution was: I_{back} = 1<0°, I_{sides} = 1<–125°, and I_{front} = 1<–250°. Notice that the W8JI phase-angles actually yielded the *same* amount of gain as the "max gain" set, although the take-off angle and front-to-back ratio obtained in the W8JI case could be improved just a bit by including a few additional degrees of phase-lag.

See Figures 15 and 16 for a comparison of the principal-plane radiation patterns which are produced by the three different sets of current phase-angles. Tom's recommended values produce noticeably smaller side lobes (in the azimuth plane) than the "max gain" set, while generating exactly the same amount of peak gain, along with front-to-back ratios that are nearly as good, in both the elevation and azimuth planes.

One advantage of utilizing the traditional 0°/–90°/–90°/–180° phase-angles is the fact that the resulting half-power beamwidth in the azimuth plane is more than 100°, which allows good coverage of all points of the compass with only four directions of fire. Employing larger phase-shift values (such as those shown in Table 4) provides more gain in the bore-sight direction, but narrows the beamwidth by more than 20°. So, it may be desirable to include a provision to allow such an array to beam through the *sides* of the square, as well as through the *corners*.

The latter portion of Table 4 covers this option. In the left column, the front pair of monopoles are fed in quadrature with those in the back (I_{back} = 1<0°, I_{front} = 1<–90°).

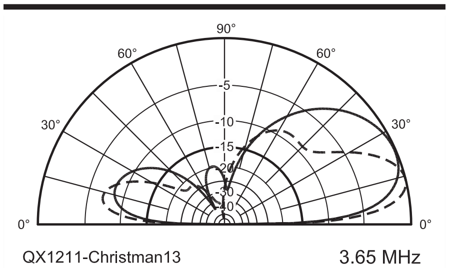

Figure 13 — Elevation-plane radiation patterns for 2 element cardioid arrays with phasing adjusted for maximum end fire gain (I_{front} = 1<–136° or 1<–144°, I_{back} = 1<0°), when using either ¼ λ or ⅝ λ monopoles. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz). The ¼ λ monopoles were tuned to resonance at 3650 kHz, while the height of the ⅝ λ radiators was adjusted for maximum gain. Solid trace = ¼ λ monopoles (H = 65.42 feet), peak gain = 4.71 dBi at 23.6° take-off angle. Dashed trace = ⅝ λ monopoles (H = 158.03 feet), peak gain = 4.81 dBi at 14.1° take-off angle.

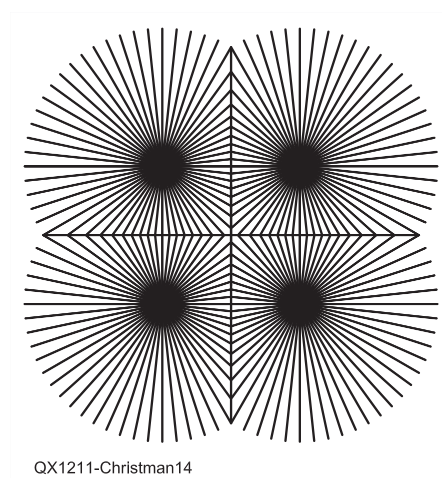

Figure 14 — Plan view of the ground screen for a 4-Square vertical array. Each side of the square has a dimension of ¼ λ, and each element has 60 buried radials, whose maximum length is also ¼ λ (67.37 feet at 3650 kHz).

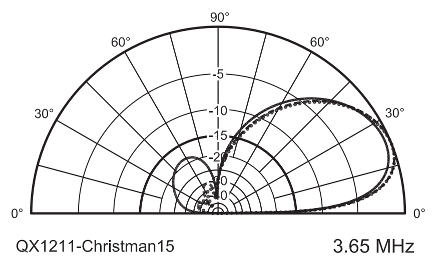

Figure 15 — Elevation-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the corners of the square, using ¼ λ monopoles that were tuned to resonance at 3650 kHz. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase-angles (0°, –90°, –90°, –180°), peak gain = 5.89 dBi at 23.7° take-off angle.
Dashed trace = "W8JI" angles (0°, –120°, –120°, –240°), peak gain = 6.36 dBi at 22.2° take-off angle.
Dotted trace = "max gain" angles (0°, –125°, –125°, –250°), peak gain = 6.36 dBi at 22.0° take-off angle.

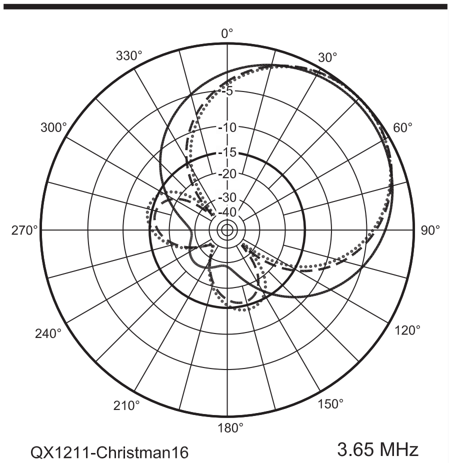

Figure 16 — Azimuth-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the corners of the square, using ¼ λ monopoles that were tuned to resonance at 3650 kHz. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, –90°, –90°, –180°), peak gain = 5.89 dBi at 23.7° take-off angle.
Dashed trace = "W8JI" angles (0°, –120°, –120°, –240°), peak gain = 6.36 dBi at 22.2° take-off angle.
Dotted trace = "max gain" angles (0°, –125°, –125°, –250°), peak gain = 6.36 dBi at 22.0° take-off angle.

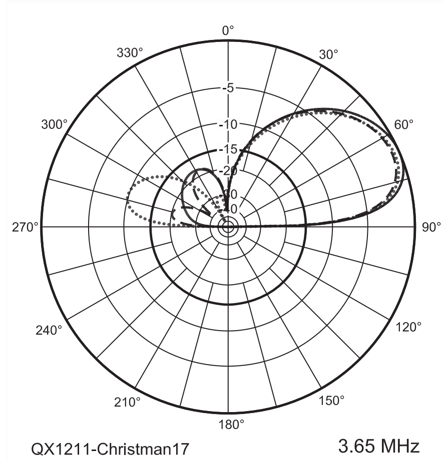

Figure 17 — Elevation-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the sides of the square, using ¼ λ monopoles that were tuned to resonance at 3650 kHz. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, 0°, –90°, –90°), peak gain = 4.90 dBi at 25.0° take-off angle.
Dashed trace = arbitrary angles (0°, 0°, –105°, –105°), peak gain = 5.22 dBi at 22.4° take-off angle.
Dotted trace = "max gain" angles (0°, 0°, –131°, –131°), peak gain = 5.42 dBi at 23.4° take-off angle.

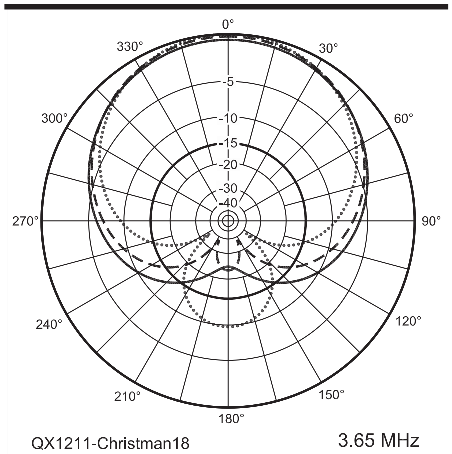

Figure 18 — Azimuth-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the sides of the square, using ¼ λ monopoles that were tuned to resonance at 3650 kHz. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, 0°, –90°, –90°), peak gain = 4.90 dBi at 25.0° take-off angle.
Dashed trace = arbitrary angles (0°, 0°, –105°, –105°), peak gain = 5.22 dBi at 24.4° take-off angle.
Dotted trace = "max gain" angles (0°, 0°, –131°, –131°), peak gain = 5.42 dBi at 23.4° take-off angle.

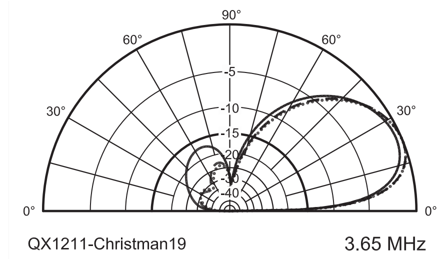

Figure 19 — Elevation-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the corners of the square, using ¼ λ monopoles that were tuned to resonance at 3650 kHz. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, –90°, –90°, –180°), peak gain = 6.34 dBi at 24.1° take-off angle.
Dashed trace = "W8JI" angles (0°, –120°, –120°, –240°), peak gain = 6.81 dBi at 22.5° take-off angle.
Dotted trace = "max gain" angles (0°, –125°, –125°, –250°), peak gain = 6.81 dBi at 22.3° take-off angle.

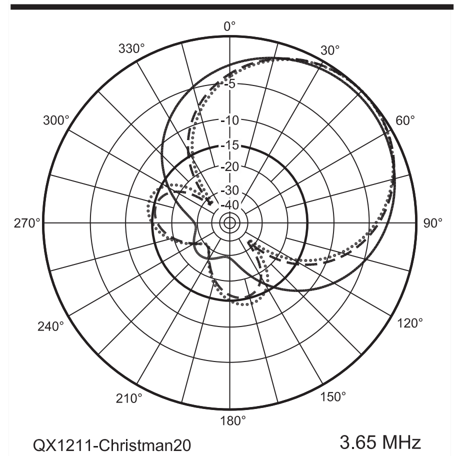

Figure 20 — Azimuth-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the corners of the square, using ¼ λ monopoles that were tuned to resonance at 3650 kHz. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz).
Solid trace = "Normal" phase-angles (0°, –90°, –90°, –180°), peak gain = 6.34 dBi at 24.1° take-off angle.
Dashed trace = "W8JI" angles (0°, –120°, –120°, –240°), peak gain = 6.81 dBi at 22.5° take-off angle.
Dotted trace = "max gain" angles (0°, –125°, –125°, –250°), peak gain = 6.81 dBi at 22.3° take-off angle.

Table 5

Performance of a 4 Square array using ¼ λ monopoles and ⅝ λ radials, as a function of the phase angles of the base currents. The height of the elements was adjusted to 65.42 feet, for resonance at 3650 kHz, while the radials have a length of 168.42 feet. Each monopole is built from no. 12 AWG copper wire, and the ground system is composed of 60 no. 16 AWG wire radials per element. The soil is "average" (conductivity = 0.005 Siemens/meter and dielectric constant = 13).

	Traditional Phasing	*W8JI Phasing*	*Maximum Gain Phasing*
Firing through the corners of the square:			
I_{front}	1<–180°	1<–240°	1<–250°
I_{sides}	1<–90°	1<–120°	1<–125°
I_{back}	1<0°	1<0°	1<0°
Maximum Gain (dBi)	6.34	6.81	6.81
Take-off Angle (°)	24.1	22.5	22.3
Elevation Plane F/B Ratio (dB)	16.48	22.76	23.85
Azimuth Plane F/B Ratio (dB)	24.26	30.12	30.63
Azimuth Plane F/S Ratio (dB)	N.A.	15.51	13.85
Azimuth Plane Half Power Beamwidth (°)	100.8	80.2	77.2
Z_{input} (Ω) (Front Element)	69.65 + *j* 56.56	31.52 + *j* 55.1	26.42 + *j* 52.42
Z_{input} (Ω) (Side Elements)	43.75 – *j* 20.11	25.66 – *j* 2.99	23.0 – *j* 0.47
Z_{input} (Ω) (Back Element)	1.16 – *j* 15.76	7.29 – *j* 0.31	8.39 + *j* 1.01
Firing through the sides of the square:			
I_{front}	1<–90°	1<–105°	1<–131°
I_{back}	1<0°	1<0°	1<0°
Maximum Gain (dBi)	5.34	5.67	5.89
Take-off Angle (°)	25.5	24.9	23.9
Elevation Plane F/B Ratio (dB)	12.94	15.49	9.07
Azimuth Plane F/B Ratio (dB)	24.57	20.95	9.41
Azimuth Plane Half Power Beamwidth (°)	129.4	118.6	103.8
Z_{input} (Ω) (Front Elements)	95.03 + *j* 5.28	88.0 + *j* 14.19	71.27 + *j* 24.34
Z_{input} (Ω) (Back Elements)	20.28 – *j* 39.23	15.79 – *j* 28.8	14.84 – *j* 9.25

Table 6

Performance of a 4 Square array using ⅝ λ monopoles and ¼ λ radials, as a function of the phase-angles of the base currents. Maximum gain at the lowest take-off angle always occurred at an element height of about 165.6 feet (where the monopole was resonant at 1448 kHz), and the radials have a length of 67.37 feet. Each monopole is built from no. 12 AWG copper wire, and the ground system is composed of 60 no. 16 AWG wire radials per element. The soil is "average" (conductivity = 0.005 Siemens/meter and dielectric constant = 13).

	Traditional Phasing	*W8JI Phasing*	*Maximum Gain Phasing*
Firing through the corners of the square:			
I_{front}	1<–180°	1<–240°	1<–270°
I_{sides}	1<–90°	1<–120°	1<–135°
I_{back}	1<0°	1<0°	1<0°
Maximum Gain (dBi)	5.45	6.57	6.82
Take-off Angle (°)	14.0	13.4	13.1
Elevation Plane F/B Ratio (dB)	11.1	17.11	20.1
Azimuth Plane F/B Ratio (dB)	11.27	17.33	20.18
Azimuth Plane F/S Ratio (dB)	N.A.	13.43	10.7
Azimuth Plane Half Power Beamwidth (°)	109.6	82.8	72.0
Z_{input} (Ω) (Front Elements)	63.65 – *j* 360.9	6.71 – *j* 426.3	– 0.78 – *j* 460.3
Z_{input} (Ω) (Side Elements)	157.6 – *j* 524.7	85.88 – *j* 524.6	56.06 – *j* 524.4
Z_{input} (Ω) (Back Elements)	61.02 – *j* 648.3	21.93 – *j* 593.0	17.5 – *j* 568.5
Firing through the sides of the square:			
I_{front}	1<–90°	1<–105°	1<–147°
I_{back}	1<0°	1<0°	1<0°
Maximum Gain (dBi)	4.25	4.72	5.56
Take off Angle (°)	14.5	14.3	13.6
Elevation Plane F/B Ratio (dB)	9.31	11.25	6.76
Azimuth Plane F/B Ratio (dB)	9.44	11.27	6.84
Azimuth Plane Half Power Beamwidth (°)	149.2	133.0	99.6
Z_{input} (Ω) (Front Elements)	192.5 – *j* 395.8	161.4 – *j* 397.5	87.52 – *j* 441.8
Z_{input} (Ω) (Back Elements)	171.1 – *j* 634.8	140.9 – *j* 628.7	75.98 – *j* 572.1

W8JI's website does not reveal the phase-angle he uses in this application, so I have selected a value of –105° (I_{back} = 1<0°, I_{front} = 1<–105°).

As before, the right-hand column is where the phase-shifts have been optimized for *maximum* forward gain at the *lowest* take-off angle. This time, the outcome was: I_{back} = 1<0° and I_{front} = 1<–131°. The "max gain" phasing yields an extra 0.2 dB of forward gain (in comparison to using a phase-lag of 105°), but the front-to-back ratios fall considerably. A study of the elevation- and azimuth-pattern plots (Figures 17 and 18) indicates that choosing an intermediate phase-shift value, such as 105° or thereabouts, may be a good compromise between maximum gain and low side-lobe levels.

Results for a 4-Square Array with ¼ λ Elements and ⅝ λ Radials

If we have installed ⅝ λ radials beneath our 4-square, but the monopoles themselves are trimmed to a height which yields quarter-wave resonance (at 3650 kHz), what kind of performance can we expect? Refer to Table 5 for the answers. It appears that the classic (0°/–90°/–90°/–180°) current phase-angles will yield well over 6 dBi of forward gain when beaming along the diagonals of the square. Larger phase-lags, such as the W8JI and "max gain" values, can produce nearly half a decibel of additional gain. The key radiation-pattern plots appear in Figures 19 and 20.

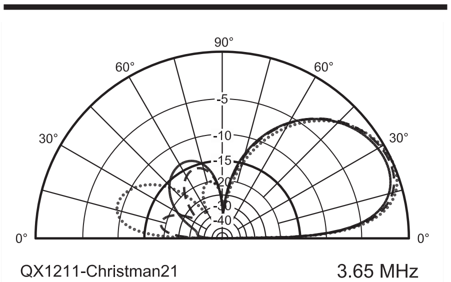

Figure 21 — Elevation-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the sides of the square, using ¼ λ monopoles that were tuned to resonance at 3650 kHz. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, 0°, –90°, –90°), peak gain = 5.34 dBi at 25.5° take-off angle.
Dashed trace = arbitrary angles (0°, 0°, –105°, –105°), peak gain = 5.67 dBi at 24.9° take-off angle.
Dotted trace = "max gain" angles (0°, 0°, –131°, –131°), peak gain = 5.89 dBi at 23.9° take-off angle.

When adjusted to fire through the sides of the square, the traditional quadrature-fed (0°/–90°) array generates more than 5 dBi of gain, while larger phase-lags providing an extra ⅓ to ½ dB in the favored direction. Figures 21 and 22 display the elevation and azimuth-plane patterns.

If we compare the data in Tables 4 and 5, we can determine how the performance of the array will change if we modify a conventional 4-square by simply *extending* the maximum length of its radials from ¼ λ to ⅝ λ. When firing through the diagonals of the square, the peak forward gain rises by 0.45 dB, no matter which set of current phase-angles we choose. On the down side, the front-to-back ratio also deteriorates in most cases. When firing through the sides of the square, the gain increases once again, by about 0.44 and 0.47 dB (depending upon the current phase-angles), but the front-to-back ratios don't suffer this time. So, we can pick up nearly half a decibel of forward gain by increasing the maximum length of the radials from ¼ λ to ⅝ λ, no matter what set of current phase-angles we pick.

Results for a 4-Square Array with ⅝ λ Elements and ¼ λ Radials

On the other hand, what if we don't have enough room to put in a larger ground screen? Is it worthwhile to make the radiators themselves taller? Let's examine Table 6, which lists the performance parameters for an array of "⅝ λ" monopoles working in conjunction with a normal ground screen whose

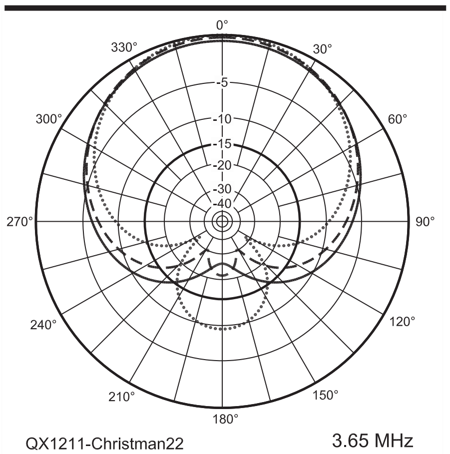

Figure 22 — Azimuth-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the sides of the square, using ¼ λ monopoles that were tuned to resonance at 3650 kHz. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, 0°, –90°, –90°), peak gain = 4.90 dBi at 25.0° take-off angle.
Dashed trace = arbitrary angles (0°, 0°, –105°, –105°), peak gain = 5.22 dBi at 22.4° take-off angle.
Dotted trace = "max gain" angles (0°, 0°, –131°, –131°), peak gain = 5.42 dBi at 23.4° take-off angle.

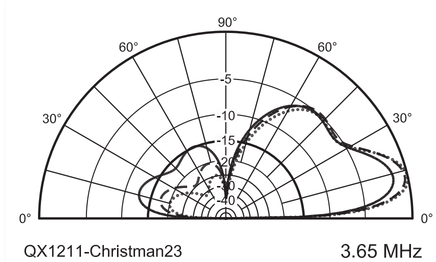

Figure 23 — Elevation-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the corners of the square, using ⅝ λ monopoles whose height was adjusted for maximum gain. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, –90°, –90°, –180°), peak gain = 5.45 dBi at 14.0° take-off angle.
Dashed trace = "W8JI" angles (0°, –120°, –120°, –240°), peak gain = 6.57 dBi at 13.4° take-off angle.
Dotted trace = "max gain" angles (0°, –135°, –135°, –270°), peak gain = 6.82 dBi at 13.1° take-off angle.

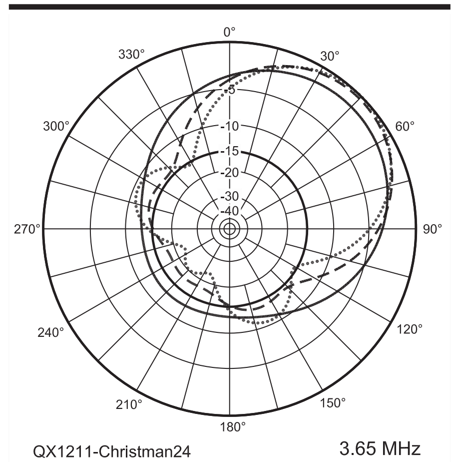

Figure 24 — Azimuth-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the corners of the square, using ⅝ λ monopoles whose height was adjusted for maximum gain. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, –90°, –90°, –180°), peak gain = 5.45 dBi at 14.0° take-off angle.
Dashed trace = "W8JI" angles (0°, –120°, –120°, –240°), peak gain = 6.57 dBi at 13.4° take-off angle.
Dotted trace = "max gain" angles (0°, –135°, –135°, –270°), peak gain = 6.82 dBi at 13.1° take-off angle.

radials are (at most) ¼ λ long.

The first thing we notice is that the normal drive-current phase-angles don't work very well, if applied to the bases of "⅝ λ" elements. Adjusting the current phase-angles from (0°/–90°/–90°/–180°) to something on the order of (0°/–120°/–120°/–240°), or even (0°/–135°/–135°/–270°), allows us to easily obtain more than a full decibel of additional gain, as well as increasing the front-to-back ratio, when beaming through the diagonals of the square. Figures 23 and 24 illustrate the patterns. Utilizing larger-than-normal current phase-angles (–105° to –147°) for the front elements can also be advantageous when firing through the sides of the square (see Figures 25 and 26 for the plots).

By comparing Tables 4 and 6, we can find out if it makes sense to extend the height of existing ¼ λ monopoles to ⅝ λ, when the radials in the ground-screen are no longer than ¼ λ. With ⅝ λ radiators, the chief area of improvement is a reduction (by roughly 9°) in the take-off angle of the main lobe. If we insist upon keeping the traditional current phase-angles, then the peak forward gain *decreases* if we switch to taller elements. But, there is an incremental *increase* in gain of 0.21 dB with W8JI phase-angles, and 0.46 dB with the "max gain" values, when beaming through the corners of the square. While firing through the sides of the square, a significant phase-lag in the front monopoles (about 147°) is needed in order to produce an improvement. Thus, a decision to employ ⅝ λ radiators in this situation should definitely be accompanied by a change in the phase-angles of the base currents supplied to the elements.

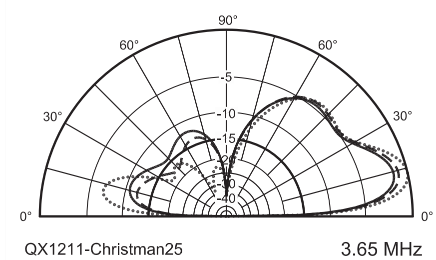

Figure 25 — Elevation-plane radiation patterns for 4-Square arrays with 3 different sets of current phase angles, when firing through the sides of the square, using ⅝ λ monopoles whose height was adjusted for maximum gain. Each element has a ground screen composed of 60 buried radials, whose maximum length is ¼ λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase angles (0°, 0°, –90°, –90°), peak gain = 4.25 dBi at 14.5° take-off angle.
Dashed trace = arbitrary angles (0°, 0°, –105°, –105°), peak gain = 4.72 dBi at 14.3° take-off angle.
Dotted trace = "maxgain" angles (0°, 0°, –147°, –147°), peak gain = 5.56 dBi at 13.6° take-off angle.

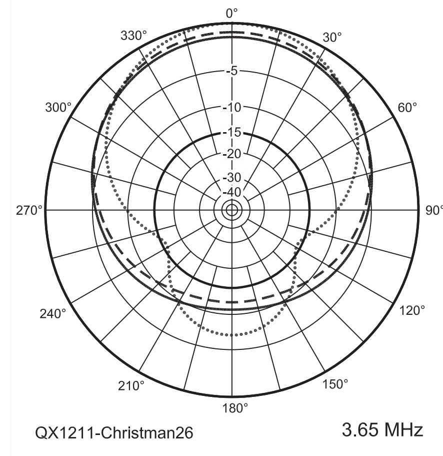

Figure 26 — Azimuth-plane radiation patterns for 4-square arrays with 3 different sets of current phase-angles, when firing through the sides of the square, using ⅝ λ monopoles whose height was adjusted for maximum gain. Each element has a ground screen composed of 60 buried radials, whose maximum length is 1/4 λ (L = 67.37 feet at 3650 kHz).
Solid trace = "Normal" phase-angles (0°,0°,-90°,-90°)
Peak gain = 4.25 dBi at 14.5° take-off angle
Dashed trace = arbitrary angles (0°,0°,-105°,-105°)
Peak gain = 4.72 dBi at 14.3° take-off angle
Dotted trace = "max-gain" angles (0°,0°,-147°,-147°)
Peak gain = 5.56 dBi at 13.6° take-off angle

Table 7

Performance of a 4 Square array using ⅝ λ monopoles and ⅝ λ radials, as a function of the phase-angles of the base currents. Maximum gain at the lowest take-off angle always occurred at an element height of about 165.6 feet (where the monopole was resonant at 1445 kHz), and the radials have a length of 168.42 feet. Each monopole is built from no. 12 AWG copper wire, and the ground system is composed of 60 no. 16 AWG wire radials per element. The soil is "average" (conductivity = 0.005 Siemens/meter and dielectric constant = 13).

	Traditional Phasing	*W8JI Phasing*	*Maximum Gain Phasing*
Firing through the corners of the square:			
I_{front}	1<–180°	1<–240°	1<–272°
I_{sides}	1<–90°	1<–120°	1<–136°
I_{back}	1<0°	1<0°	1<0°
Maximum Gain (dBi)	5.62	6.73	6.97
Take-off Angle (°)	13.9	13.4	13.1
Elevation Plane F/B Ratio (dB)	11.13	16.95	19.75
Azimuth Plane F/B Ratio (dB)	11.29	17.16	19.83
Azimuth Plane F/S Ratio (dB)	N.A.	13.56	10.56
Azimuth Plane Half Power Beamwidth (°)	109.4	82.6	71.2
Z_{input} (Ω) (Front Element)	68.1 – *j* 361.5	9.32 – *j* 425.5	0.72 – *j* 461.5
Z_{input} (Ω) (Side Elements)	157.7 – *j* 527.9	86.0 – *j* 526.2	54.41 – *j* 525.2
Z_{input} (Ω) (Back Element)	58.61 – *j* 649.0	20.88 – *j* 593.0	16.89 – *j* 567.0
Firing through the sides of the square:			
I_{front}	1<–90°	1<–105°	1<–146°
I_{back}	1<0°	1<0°	1<0°
Maximum Gain (dBi)	4.42	4.89	5.71
Take-off Angle (°)	14.4	14.2	13.6
Elevation Plane F/B Ratio (dB)	9.31	11.22	6.97
Azimuth Plane F/B Ratio (dB)	9.43	11.24	7.05
Azimuth Plane Half Power Beamwidth (°)	149.0	132.6	100.0
Z_{input} (Ω) (Front Elements)	196.1 – *j* 399.9	165.0 – *j* 400.8	91.48 – *j* 441.5
Z_{input} (Ω) (Back Elements)	168.7 – *j* 638.0	138.7 – *j* 631.2	76.22 – *j* 574.9

Results for a 4-Square Array with ⅝ λ Elements and ⅝ λ Radials

The final configuration incorporates monopoles whose height is approximately ⅝ λ, in combination with buried radials with a maximum length of ⅝ λ. This array would be the most expensive to construct, requiring the largest amount of land as well as the tallest radiators. Table 7 provides the critical data we need. Notice that the application of larger-than-normal phase-shifts (either the W8JI or "max-gain" values) to the drive currents can generate at least a full decibel of extra gain, along with more front-to-back ratio, when compared to the typical (0°/–90°/–90°/–180°) angles. The key radiation-pattern plots are shown in Figures 27 and 28. Greater current phase-shifts are also beneficial when the array is firing through the sides of the square (see Figures 29 and 30).

Reviewing Table 4 together with Table 7 permits us to see the "margin of superiority" that the biggest 4-square array (⅝ λ elements and radials) enjoys over the smallest one (¼ λ elements and radials). With normal current phasing (0°/–90°/–90°/–180°) the results are disastrous when we look at forward gain — the smaller array works better! With larger phase-lags, however, we calculate 0.37 dB of extra gain for W8JI phasing, and 0.61 dB with "max gain" phasing. However, the taller monopoles in the big array will always give us a main-lobe take-off angle that is lower by about 9°. If we are beaming through the sides of the square, only the "max gain" phase-angles provide additional gain, versus the small array.

Keeping the Current Phase-angles Constant

In this section, we will examine what happens if the phase-angles of the base currents are held fixed, while the length of the radials and the height of the radiators is varied. The results for "traditional" phase angles (0°/–90°/–90°/–180°) are presented in Table 8. The highest gain is achieved when ¼ λ elements are placed over a ground screen composed of radials whose maximum length is ⅝ λ. This set of current phase-angles doesn't work too well when combined with ⅝ λ monopoles, but the taller radiators *do* generate a much-lower main lobe, which may be preferable. In that case, an array using ⅝ λ elements and ⅝ λ radials would be the one to pick. The elevation-plane patterns for both of these alternatives are given in Figure 31.

If the "W8JI" current phase-angles are applied to the input terminals of the radiators, the outcome will be as displayed in Table 9. Once again, a system of ¼ λ elements combined with ⅝ λ (max) radials produces the most gain. If a lower take-off angle is

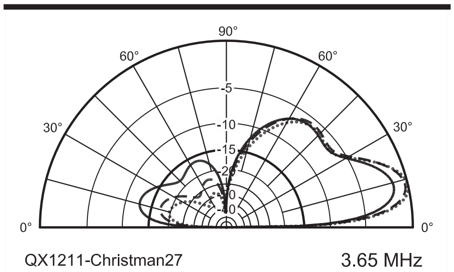

Figure 27 — Elevation-plane radiation patterns for 4-square arrays with 3 different sets of current phase-angles, when firing through the corners of the square, using ⅝ λ monopoles whose height was adjusted for maximum gain. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz).
Solid trace = "Normal" phase-angles (0°,-90°,-90°,-180°)
Peak gain = 5.62 dBi at 13.9° take-off angle
Dashed trace = "W8JI" angles (0°,-120°,-120°,-240°)
Peak gain = 6.73 dBi at 13.4° take-off angle
Dotted trace = "max-gain" angles (0°,-136°,-136°,-272°)
Peak gain = 6.97 dBi at 13.1° take-off angle

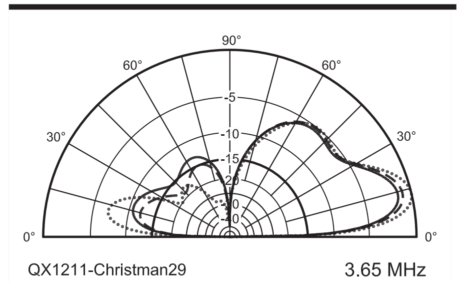

Figure 29 — Elevation-plane radiation patterns for 4-square arrays with 3 different sets of current phase-angles, when firing through the sides of the square, using ⅝ λ monopoles whose height was adjusted for maximum gain. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz).
Solid trace = "Normal" phase-angles (0°,0°,-90°,-90°)
Peak gain = 4.42 dBi at 14.4° take-off angle
Dashed trace = arbitrary angles (0°,0°,-105°,-105°)
Peak gain = 4.89 dBi at 14.2° take-off angle
Dotted trace = "max-gain" angles (0°,0°,-146°,-146°)
Peak gain = 5.71 dBi at 13.6° take-off angle

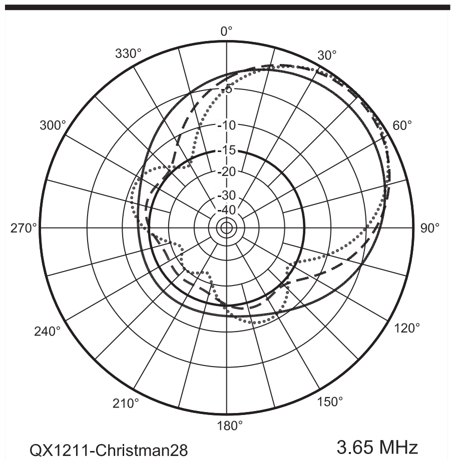

Figure 28 — Azimuth-plane radiation patterns for 4-square arrays with 3 different sets of current phase-angles, when firing through the corners of the square, using ⅝ λ monopoles whose height was adjusted for maximum gain. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz).
Solid trace = "Normal" phase-angles (0°,-90°,-90°,-180°)
Peak gain = 5.62 dBi at 13.9° take-off angle
Dashed trace = "W8JI" angles (0°,-120°,-120°,-240°)
Peak gain = 6.73 dBi at 13.4° take-off angle
Dotted trace = "max-gain" angles (0°,-136°,-136°,-272°)
Peak gain = 6.97 dBi at 13.1° take-off angle

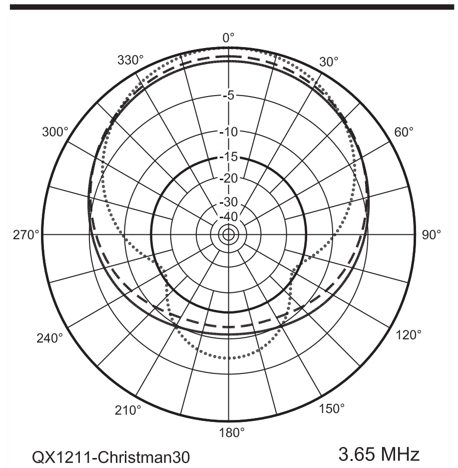

Figure 30. Azimuth-plane radiation patterns for 4-square arrays with 3 different sets of current phase-angles, when firing through the sides of the square, using ⅝ λ monopoles whose height was adjusted for maximum gain. Each element has a ground screen composed of 60 buried radials, whose maximum length is ⅝ λ (L = 168.42 feet at 3650 kHz).
Solid trace = "Normal" phase-angles (0°,0°,-90°,-90°)
Peak gain = 4.42 dBi at 14.4° take-off angle
Dashed trace = arbitrary angles (0°,0°,-105°,-105°)
Peak gain = 4.89 dBi at 14.2° take-off angle
Dotted trace = "max-gain" angles (0°,0°,-146°,-146°)
Peak gain = 5.71 dBi at 13.6° take-off angle

Table 8

Performance of a 4 Square array when the elements are driven with traditional current phase angles (0°/–90°/–90°/–180°), as a function of radiator height and radial length. Each antenna is designed to operate at 3650 kHz. Arrays with ¼ λ monopoles have their element heights adjusted for resonance at 3650 kHz, while those with ⅝ λ elements are adjusted for maximum gain and lowest take-off angle at the same frequency. Each monopole is built from no. 12 AWG copper wire, while the ground system is composed of 60 no. 16 AWG wire radials. The soil is “average” (conductivity = 0.005 Siemens per meter and dielectric constant = 13).

	¼ λ Monopole ¼ λ Radials	*¼ λ Monopole ⅝ λ Radials*	*⅝ λ Monopole ¼ λ Radials*	*⅝ λ Monopole ⅝ λ Radials*
Radiator Height (ft)	65.43	65.42	165.62	165.62
Resonant Frequency (kHz)	3650	3650	1448	1445
Radial Length (ft)	67.37	168.42	67.37	168.42
Firing through the corners of the square:				
	(I_{back} = 1<0°, I_{sides} = 1<–90°, and I_{front} = 1<–180°)			
Maximum Gain (dBi)	5.89	6.34	5.45	5.62
Take-off Angle (°)	23.7	24.1	14.0	13.9
Elevation Plane F/B Ratio (dB)	17.99	16.48	11.1	11.13
Azimuth Plane F/B Ratio (dB)	23.35	24.26	11.27	11.29
Azimuth Plane Half Power Beamwidth (°)	100.6	100.8	109.6	109.4
Firing through the sides of the square:				
	(I_{back} = 1<0° and I_{front} = 1<–90°)			
Maximum Gain (dBi)	4.90	5.34	4.25	4.42
Take-off Angle (°)	25.0	25.5	14.5	14.4
Elevation Plane F/B Ratio (dB)	13.93	12.94	9.31	9.31
Azimuth Plane F/B Ratio (dB)	23.60	24.57	9.44	9.43
Azimuth Plane Half Power Beamwidth (°)	130.0	129.4	149.2	149.0

Table 9

Performance of a 4 Square array when the elements are driven with W8JI current phase angles (0°/–120°/–120°/–240°), as a function of radiator height and radial length. Each antenna is designed to operate at 3650 kHz. Arrays with ¼ λ monopoles have their element heights adjusted for resonance at 3650 kHz, while those with ⅝ λ elements are adjusted for maximum gain and lowest take-off angle at the same frequency. Each monopole is built from no. 12 AWG copper wire, while the ground system is composed of 60 no. 16 AWG wire radials. The soil is “average” (conductivity = 0.005 Siemens per meter and dielectric constant = 13).

	¼ λ Monopole ¼ λ Radials	*¼ λ Monopole ⅝ λ Radials*	*⅝ λ Monopole ¼ λ Radials*	*⅝ λ Monopole ⅝ λ Radials*
Radiator Height (ft)	65.43	65.42	165.59	165.59
Resonant Frequency (kHz)	3650	3650	1448	1445
Radial Length (ft)	67.37	168.42	67.37	168.42
Firing through the corners of the square:				
	(I_{back} = 1<0°, I_{sides} = 1<–120°, and I_{front} = 1<–240°)			
Maximum Gain (dBi)	6.36	6.81	6.57	6.73
Take-off Angle (°)	22.2	22.5	13.4	13.4
Elevation- Plane F/B Ratio (dB)	30.30	22.76	17.11	16.95
Azimuth-Plane F/B Ratio (dB)	36.99	30.12	17.33	17.16
Azimuth-Plane F/S Ratio (dB)	15.90	15.51	13.43	13.56
Azimuth-Plane Half Power Beamwidth (°)	80.0	80.2	82.8	82.6
Firing through the sides of the square:				
	(I_{back} = 1<0°and I_{front} = 1<–105°)			
Maximum Gain (dBi)	5.22	5.67	4.72	4.89
Take-off Angle (°)	22.4	24.9	14.3	14.2
Elevation-Plane F/B Ratio (dB)	17.71	15.49	11.25	11.22
Azimuth-Plane F/B Ratio (dB)	21.65	20.95	11.27	11.24
Azimuth-Plane Half Power Beamwidth (°)	118.2	118.6	133.0	132.6

our goal, then an antenna utilizing both ⅝ λ monopoles and ⅝ λ radials works best. (See Figure 32 for the elevation-plane plots.)

Table 10 provides the performance data for 4-square arrays in which the current phase-angles have been adjusted to supply the maximum-possible amount of forward gain at the lowest-attainable take-off angle. If we must limit ourselves to relatively-short (¼ λ) radiators, then they should be combined with ⅝ λ radials, as usual. However, a system of ⅝ λ radiators and ⅝ λ radials is the overall winner, yielding almost 7 dBi of gain at an elevation angle of just over 13°. The radiation-pattern plots are shown in Figure 33.

Conclusions

This article has discussed the use of extended-height radiators and extended-length radials in vertical antenna systems. Designs composed of a single element have been reviewed, along with both 2- and 4-element arrays. Computer analysis reveals that in many cases, normal ¼ λ monopoles can be combined with longer radials and modified current phase-angles to provide better performance. If lower take-off angles are important, then taller radiators must be employed.

Notes

[1]John Devoldere, ON4UN, *ON4UN's Low-Band DXing* (*4th edition)*, ARRL, Newington, CT, 2005; see Chapter 11 for details.

[2]Roy Lewallen, W7EL, *EZNEC* antenna-simulation software; available from Roy Lewallen, W7EL, PO Box 6658, Beaverton OR 97007.

[3]Tom Rauch, W8JI, "Four-Square," **w8ji.com/tx_four_square.htm**.

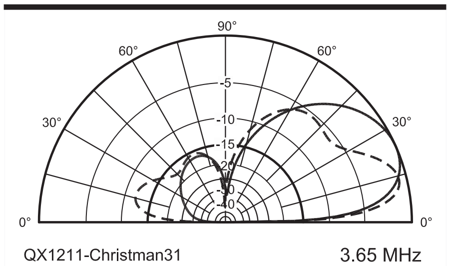

Figure 31 — Elevation-plane radiation patterns for the two best- performing 4-square arrays, when firing through the corners of the square, using traditional current phase-angles (0°,-90°,-90°,-180°).
Solid trace = 1/4 λ elements and ⅝ λ radials
Peak gain = 6.34 dBi at 24.1° take-off angle
Dashed trace = ⅝ λ elements and ⅝ λ radials
Peak gain = 5.62 dBi at 13.9° take-off angle

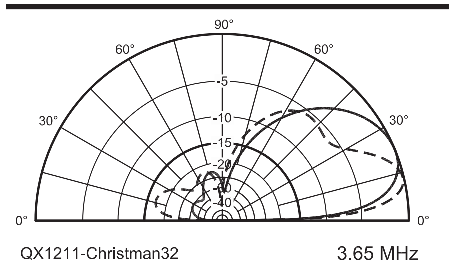

Figure 32 — Elevation-plane radiation patterns for the two best performing 4-square arrays, when firing through the corners of the square, using "W8JI" current phase angles (0°,-120°,-120°,-240°).
Solid trace = 1/4 λ elements and ⅝ λ radials
Peak gain = 6.81 dBi at 22.5° take-off angle
Dashed trace = ⅝ λ elements and ⅝ λ radials
Peak gain = 6.73 dBi at 13.4° take-off angle

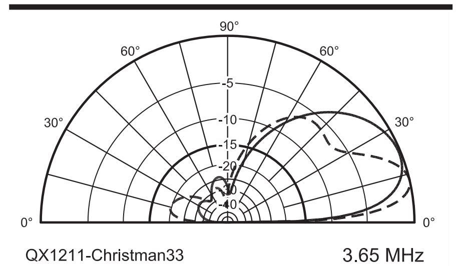

Figure 33 — Elevation-plane radiation patterns for the two best- performing 4-square arrays, when firing through the corners of the square, using "max-gain" current phase- angles, (0°,-125°,-125°,-250°) or (0°,-136°,-136°,-272°)
Solid trace = 1/4 λ elements and ⅝ λ radials
Peak gain = 6.81 dBi at 22.3° take-off angle
Dashed trace = ⅝ λ elements and ⅝ λ radials
Peak gain = 6.97 dBi at 13.1° take-off angle

Table 10

Performance of a 4 Square array when the phase angles of the element currents are selected to produce maximum forward gain, as a function of radiator height and radial length. Each antenna is designed to operate at 3650 kHz. Arrays with ¼ λ monopoles have their element heights adjusted for resonance at 3650 kHz, while those with ⅝ λ elements are adjusted for maximum gain and lowest take-off angle at the same frequency. Each monopole is built from no. 12 AWG copper wire, while the ground system is composed of 60 no. 16 AWG wire radials. The soil is "average" (conductivity = 0.005 Siemens per meter and dielectric constant = 13).

	¼ λ Monopole ¼ λ Radials	*¼ λ Monopole ⅝ λ Radials*	*⅝ λ Monopole ¼ λ Radials*	*⅝ λ Monopole ⅝ λ Radials*
Radiator Height (ft)	65.43	65.42	165.59	165.59
Resonant Frequency (kHz)	3650	3650	1448	1445
Radial Length (ft)	67.37	168.42	67.37	168.42
Firing through the corners of the square:				
		(I_{back} = 1<0°)		
I_{sides}	1<–125°	1<–125°	1<–135°	1<–136°
I_{front}	1<–250°	1<–250°	1<–270°	1<–272°
Maximum Gain (dBi)	6.36	6.81	6.82	6.97
Take-off Angle (°)	22.0	22.3	13.1	13.1
Elevation Plane F/B Ratio (dB)	33.96	23.85	20.10	19.75
Azimuth Plane F/B Ratio (dB)	38.35	30.63	20.18	19.83
Azimuth Plane F/S Ratio (dB)	14.10	13.85	10.70	10.56
Azimuth Plane Half Power Beamwidth (°)	76.8	77.2	72.0	71.2
Firing through the sides of the square:				
		(I_{back} = 1<0°)		
I_{front}	1<–131°	1<–131°	1<–147°	1<–146°
Maximum Gain (dBi)	5.42	5.89	5.56	5.71
Take-off Angle (°)	23.4	23.9	13.6	13.6
Elevation Plane F/B Ratio (dB)	9.41	9.07	6.76	6.97
Azimuth Plane F/B Ratio (dB)	9.76	9.41	6.84	7.05
Azimuth Plane Half Power Beamwidth (°)	104.1	103.8	99.6	100.0

QEX

Ray Mack, W5IFS

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SDR: Simplified

Danger — Math Ahead

The Fourier Transform and Variations

Every DSP based radio of which I am aware has some sort of spectrum display. These come in panadapter and waterfall flavors. These are all "real time" displays of energy versus frequency. The heart of the spectrum display in all of these cases is the Discrete Fourier Transform (DFT). In this column we will look at the DFT and a special case called the Fast Fourier Transform (FFT).

The general form of the continuous Fourier Transform is:

$$X(f) = \int_{-\infty}^{\infty} x(t) e^{-j2\pi ft} dt \qquad \text{[Eq 1]}$$

This yields a complex frequency function (contains both sine and cosine terms from the $e^{-j2\pi ft}$ term) that is continuous in frequency from a frequency of –∞ to ∞. The transform is derived from a continuous time function that is not periodic (in general) and extends from time –∞ to ∞. The Fourier Series we use to describe a periodic signal such as a square wave or triangle wave is a special case of the general Fourier Transform. If you plot a periodic function in time, it is continuous from –∞ to ∞, but the spectrum is composed of discrete frequency elements. There is also an inverse function that will transform a continuous frequency function into a corresponding continuous time function (Inverse Fourier Transform):

$$x(t) = \int_{-\infty}^{\infty} X(f) e^{j2\pi ft} df \qquad \text{[Eq 2]}$$

Notice that the only difference is the sign of the complex exponent. Likewise, there is an inverse of the Fourier Series that takes all of the component sine and cosine waves and adds the frequencies from –∞ to ∞. The integral is replaced by a summation of the discrete frequencies from –∞ to ∞. These are not terribly useful in the real world since we cannot really go forward or backward in time to plus or minus infinity.

The math so far has been ugly, with imaginary exponents, infinities, and integrals, but the math gets really ugly when you do the transformation from the continuous Fourier and Inverse Fourier forms to the Discrete Fourier and inverse Discrete Fourier forms. To see just how ugly, go to the Wikipedia page for Discrete Fourier Transform — **http://en.wikipedia.org/wiki/Discrete_Fourier_transform**. Quite a few folks did the really hard math over the past 250 years. Oddly, the first Fourier work was actually for the discrete forms rather than the continuous forms! Nyquist detailed part of the theory of practical DFT in 1928, and his work was expanded by Shannon and others later in the 20th century (all before practical DSP computers). When all the math is done, they showed that you can sample a continuous function at equally spaced intervals for a finite time, convert that sequence of samples into a limited discrete frequency set (the DFT), and then convert that limited frequency set back into a continuous time function. All of those conversions come with some specific constraints in order to make the conversions work in both directions to produce the output the same as the input.

The first constraint is the Nyquist limit, which requires that we sample at a rate *greater than* twice the highest frequency component. A discrete transform always acts on a limited number of samples. The discrete transform is much like a Fourier Series because the transform only contains a sequence of discrete frequencies. The Fourier Series is a consequence of the input signal being periodic. The limited number of discrete frequencies of the DFT forces the math to look like we have applied a rectangular window to a periodic signal. The second constraint is that a periodic function must be an exact sub-multiple of the sample frequency in order to do an exact transform or inverse transform.

The definition of the DFT is an operation that transforms a finite sequence of time samples into a sequence of the same number of complex frequency samples. In math form:

$$X[f] = \sum_{n=0}^{N-1} x[t] W_N^{nf} \qquad \text{[Eq 3]}$$

where:
$X[f]$ is the set of frequency samples,
$x[t]$ is the set of time samples,
N is the number of samples in each set, and
W is the transform operator. Each of the elements (X, x, and W) can be represented as an array of values in a computer, so this is perfect for implementation in software.

Consequences of Signals in the Real World

Let's look at some examples to see what happens with respect to the constraints. We will look at a spectrum for a system sampled at 8 kHz and look at the signal every 12.5 ms. This means we update the spectrum 80 times per second. The 12.5 ms will result in 100 samples of the signal. The first example is for a 240 Hz signal. The update rate of 80 corresponds to each frequency sample mapping to exactly 80 Hz. The 80 Hz comes about because we can fit exactly one cycle of 80 Hz in 100 samples at 8000 samples per second. This is a discrete transform, so mathematically there is no energy at 5 Hz or 60 Hz for example, just energy at dc, 80 Hz, 160 Hz, … 4 kHz. Each of those exact frequencies is called a "bin." Figure 1 shows the continuous and sampled 240 Hz signal. Notice that it holds exactly three cycles of the 240 Hz waveform but it has 45° negative offset from a sine wave. Figure 2 shows the DFT of the signal. It is important to note that the DFT is a complex function that has both a cosine term (real) and a sine term (imaginary). In our example, the cosine term at 240 Hz is positive and the sine term is negative because of that –45° offset. If the samples had started at 1 for a cosine wave, the real part would have been one and the imaginary would have been zero. If it had started at zero for a sine wave, the real part would have been zero and the imaginary part would have been one.

Now, let's look at a cosine wave at 280 Hz. Figure 3A shows the 100 samples of our waveform that we use in our computer. Figure 3B shows what the math "sees." The math assumes that the 3½ cycles of the 280 Hz repeats with that discontinuity occurring every 12.5 ms from –∞ to ∞. Figure 4 shows the DFT of this new signal. Now, we have energy in multiple bins because the real energy falls between adjacent bins for the fundamental sine wave. There is also additional energy in other bins because of the discontinuity at the end of the set of samples.

More Math

A DFT is really a conversion from a complex (real and imaginary) set of samples in time to a set of complex samples in frequency. Our first two examples implement what is called a "real" transform. These examples placed all of the time samples in the real part of the input set and loaded all of the imaginary samples with zero. Many SDR systems do all of their work with I and Q channels, so we have both a real set of samples (I) and an imaginary set of samples (Q). For spectral analysis, there is no real advantage for complex or real transforms. The Nyquist rate is required for a real time sequence in order

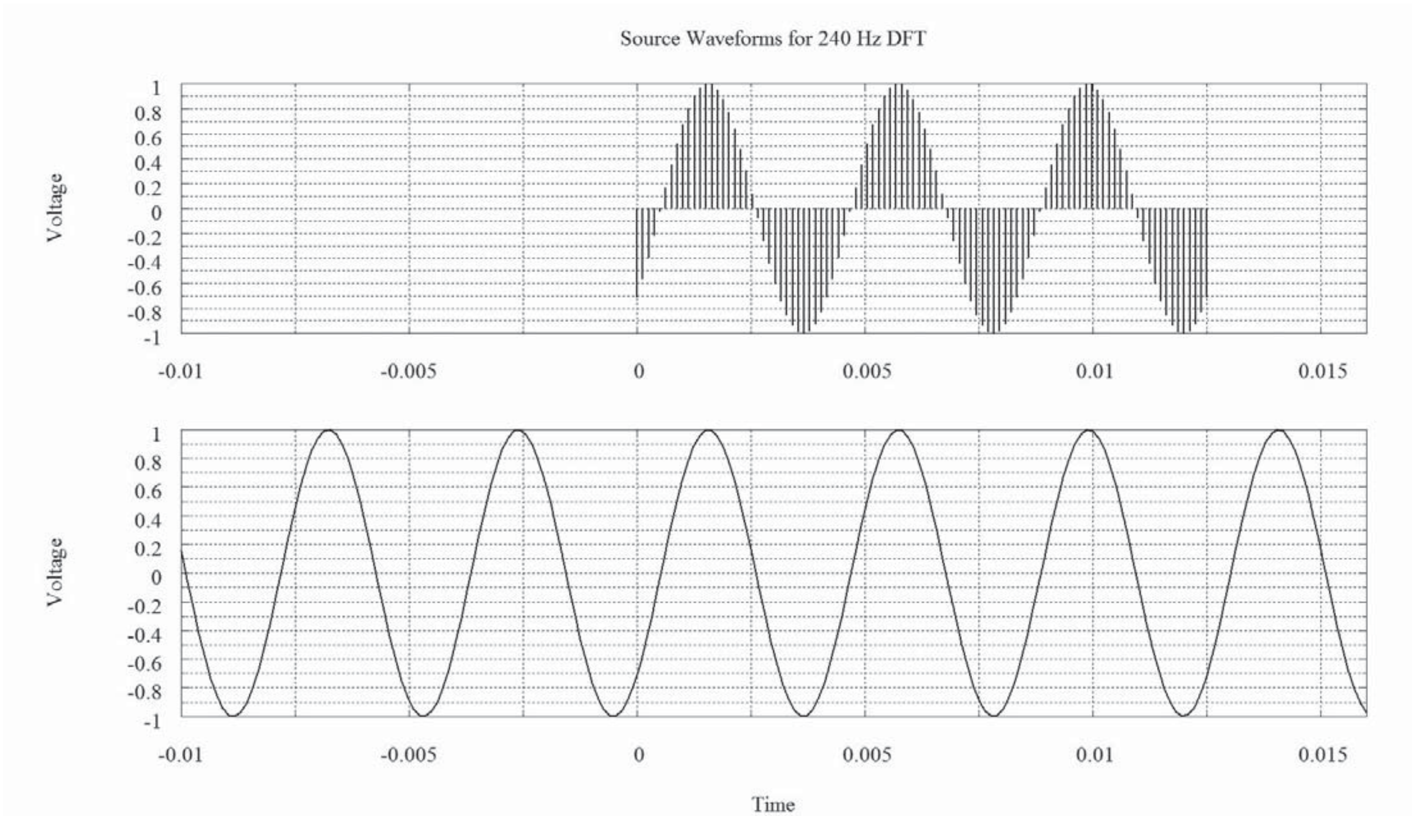


Figure 1 — Sampled and continuous versions of a 240 Hz sinusoid offset by 45°.

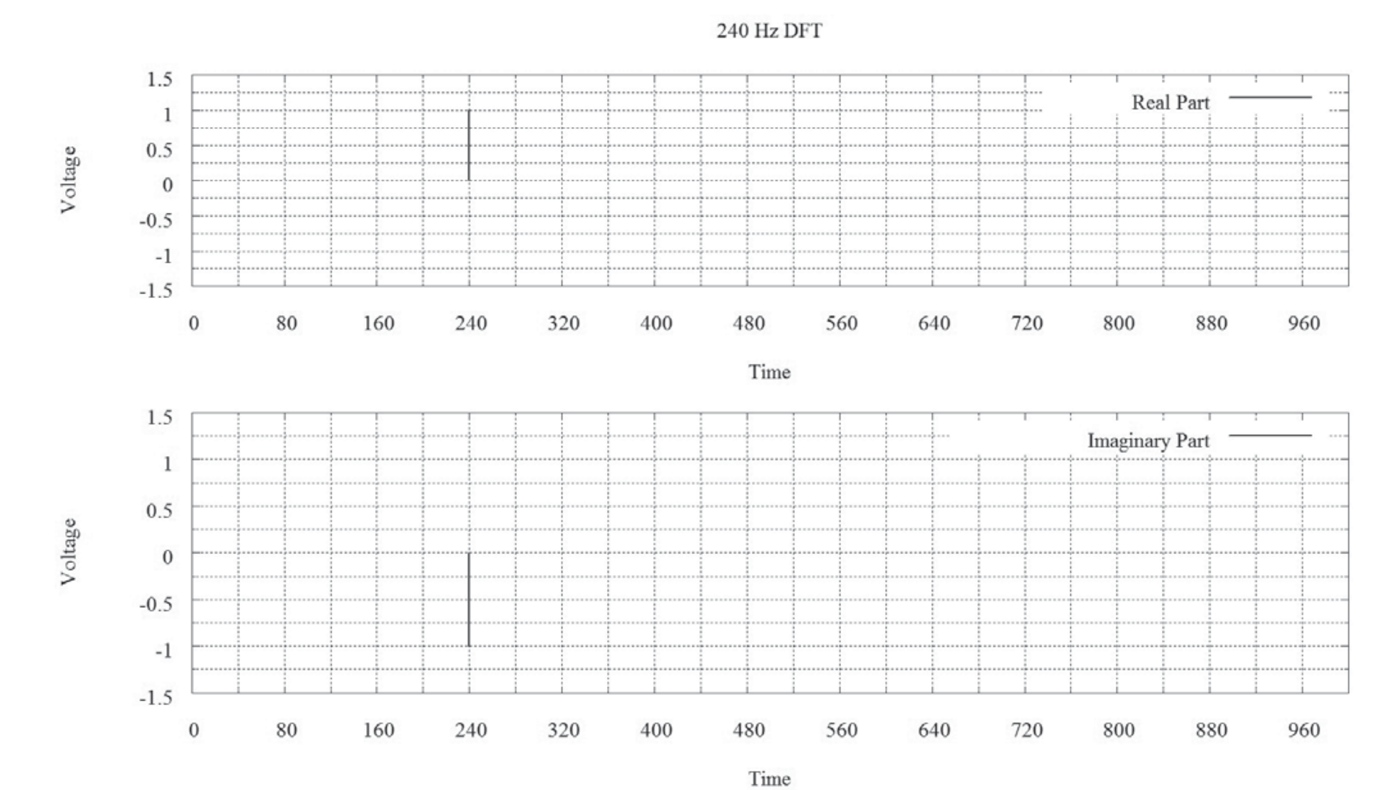


Figure 2 — The Discrete Fourier Transform (DFT) of the sinusoid in Figure 1.

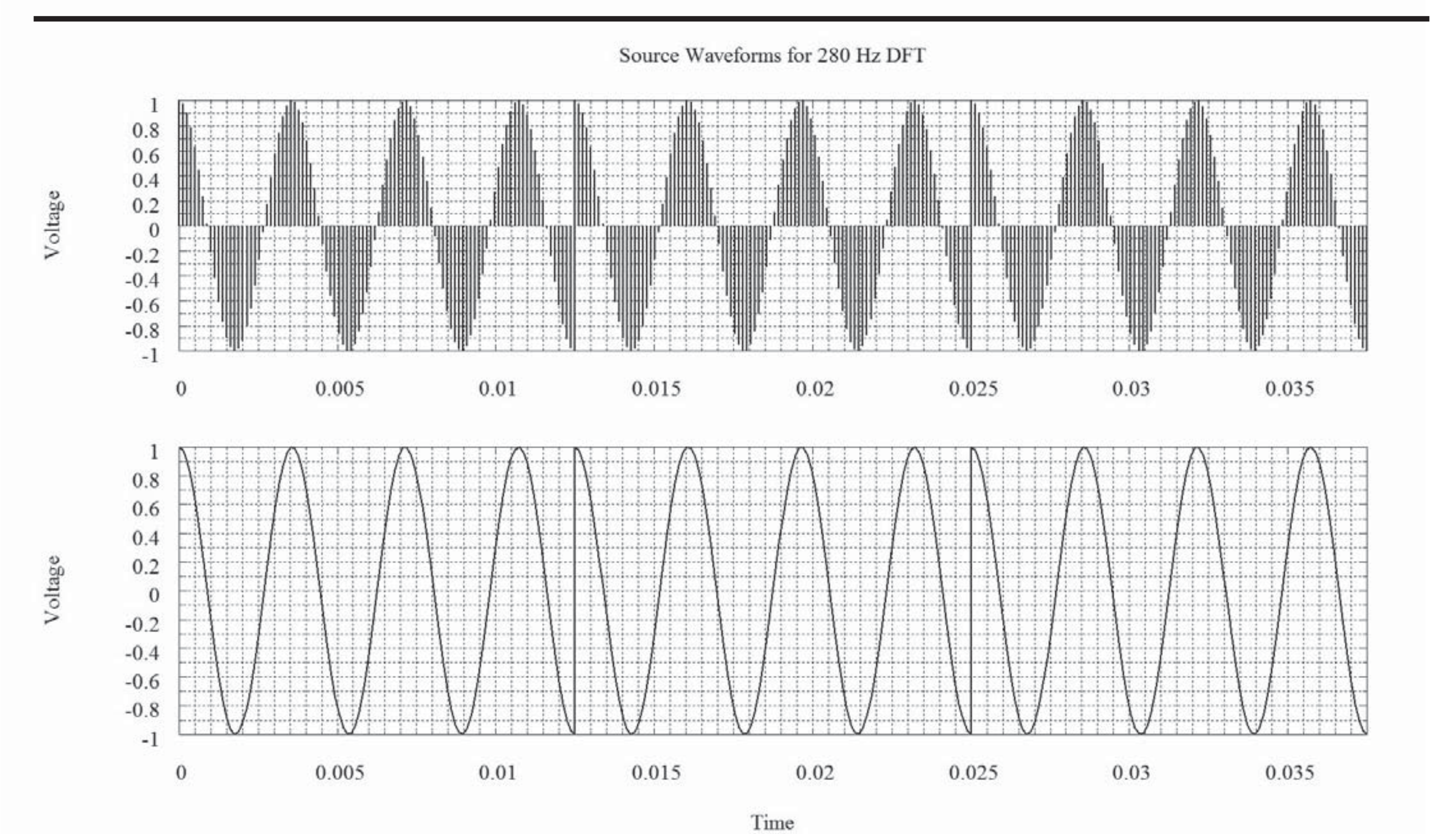


Figure 3 — Part A shows the sampled waveform for a 280 Hz sinusoid. Part B shows the effective signal that is transformed by the DFT.

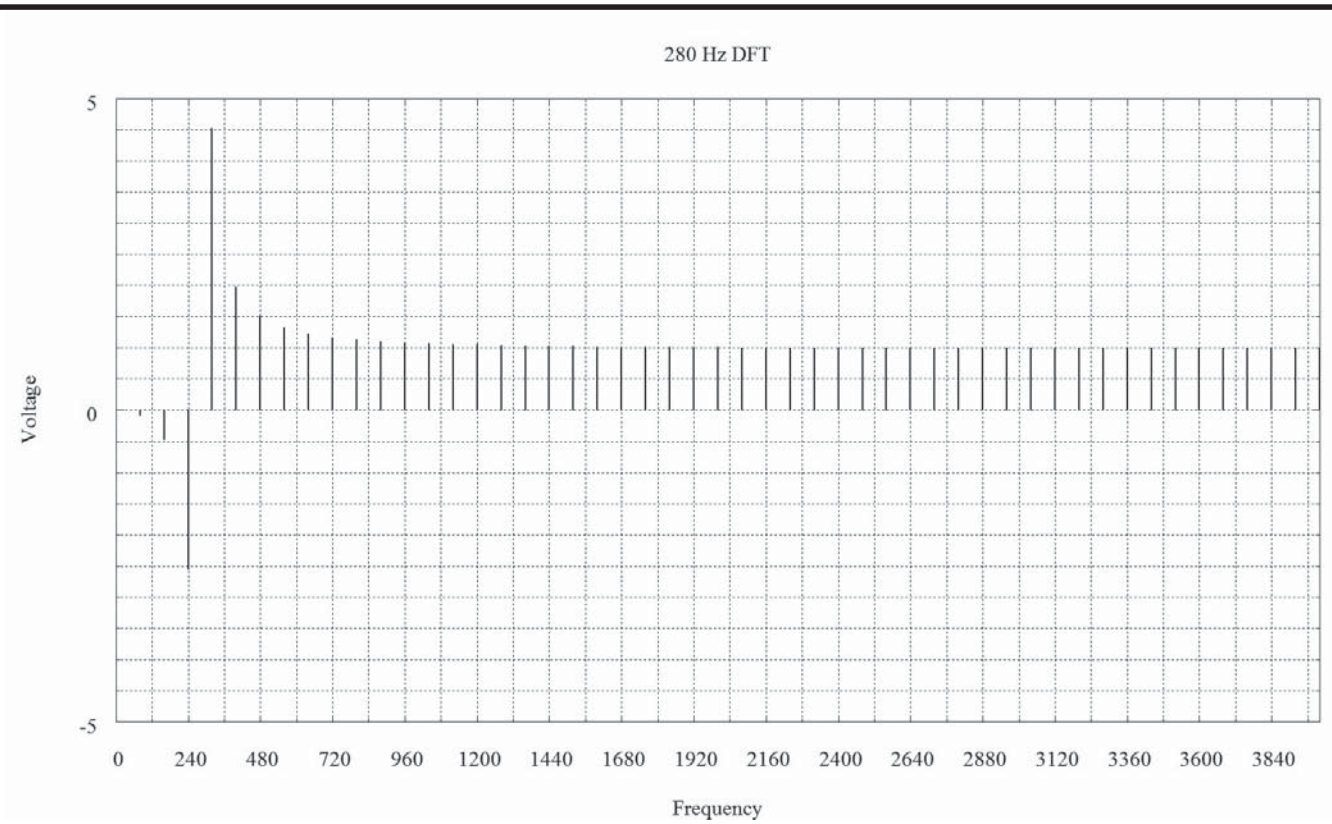


Figure 4 — The DFT of the signal of Figure 3.

to be able to determine both phase and frequency of the component signals, but only for positive frequencies. If we use an I-Q system we essentially double the sample rate and obtain exact phase and frequency for both positive and negative frequencies. I-Q systems double the sample rate, the computational load, and the covered output spectrum. The cost is twice the number of pieces of hardware for conversion.

The DFT is a general case function. J. W. Cooley and J.W. Tukey developed a computer algorithm in 1965 that takes advantage of the symmetry of the DFT if the number of samples is a power of two. This algorithm is called a Fast Fourier Transform (FFT) because it has significantly fewer arithmetic operations than a general DFT. The Cooley-Tukey FFT is just one of many possible fast transforms, but it is the one usually implemented. A DFT requires approximately N^2 operations, where an FFT requires approximately $N \log_2 N$ operations (where N is the number of samples). The software for this article contains an implementation of the FFT from *Numerical Recipes in C*.[1, 2]

The examples so far have looked at the real or complex input and complex output of the DFT. The complex output tells us both phase and frequency relative to a cosine wave that would begin at time sample zero. Phase is arbitrary since we almost never have a real reference between the signal we see and the place it was generated. For that reason, it is normal practice to convert the rectangular I-Q presentation to polar form with magnitude and phase (which we discard or simply choose not to calculate). This is called the power spectrum, since it is formed from squaring the two rectangular components. If we want the results in decibels, we save computation by doing just a log calculation rather than doing both a square root and log function. The log function allows us to multiply the result by 10 instead of 20 to get the magnitude in dB.

MATLAB and *Octave*

MATLAB and *Octave* contain functions for calculating the FFT and inverse FFT of data sets. Of course, *Octave* is a more useful tool for experimenters since it is free compared to hundreds of dollars for *MATLAB*. I find that the book *Computer Based Exercises for Signal Processing Using MATLAB 5* is a useful reference for understanding DSP concepts and using the computer tools for experiments.[3] It has a significant amount of engineering level math, though.

The FFT Algorithm

The heart of most FFT algorithms is a data flow pattern called a butterfly. Figure 5 shows a two element butterfly. The input contains two elements $x(0)$ and $x(1)$, and the output has two elements $y(0)$ and $y(1)$. The equations for the butterfly are:

$$y(0) = x(0) + x(1)W^k \quad \text{[Eq 4]}$$

$$y(1) = x(0) - x(1)W^k \quad \text{[Eq 5]}$$

where W^k is the appropriate element

$[\cos(2 \times k) + j\sin(2 \times k)]$.

Figure 6 shows the data flow diagram of the scrambled in/natural out (the output is ordered from lowest to highest element). The number in the circle at each stage indicates the value of k for W^k. It is interesting that the values of k are even for all stages except the final stage. A dashed line indicates an element that is added and a solid line indicates the element is multiplied by W^k. Notice that the first stage of calculation performs four 2×2 butterfly operations, the second is two 4×4 butterfly operations, and the final is one 8×8 butterfly. Each stage has fewer butterfly operations so it is called a decimation in time algorithm. The other thing to notice is that the input is grouped with the even time elements in the top group and the odd time elements in the bottom group.

This example requires just 24 complex multiplication operations. One other large advantage of the FFT is that the algorithm is a "multiply accumulate" type operation, where each horizontal position in Figure 6 has a multiply in place at each stage.

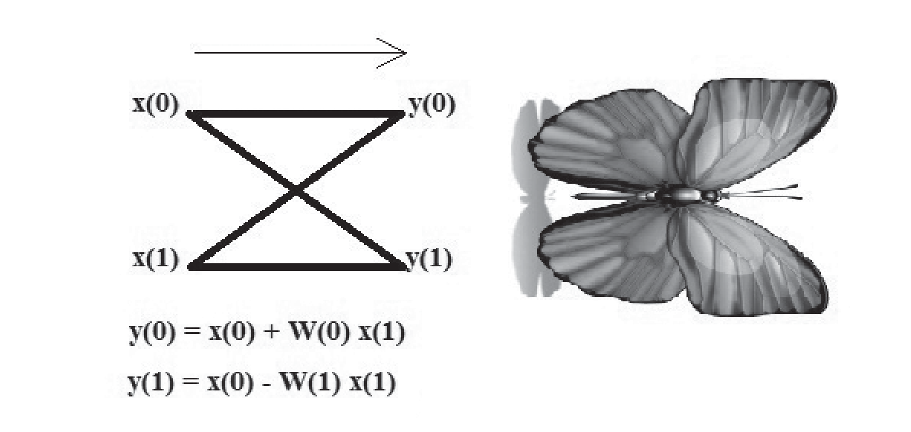


Figure 5 — A two element "butterfly" signal flow diagram. The crossing of the solid and dashed lines make the diagram look like stylized butterfly wings.

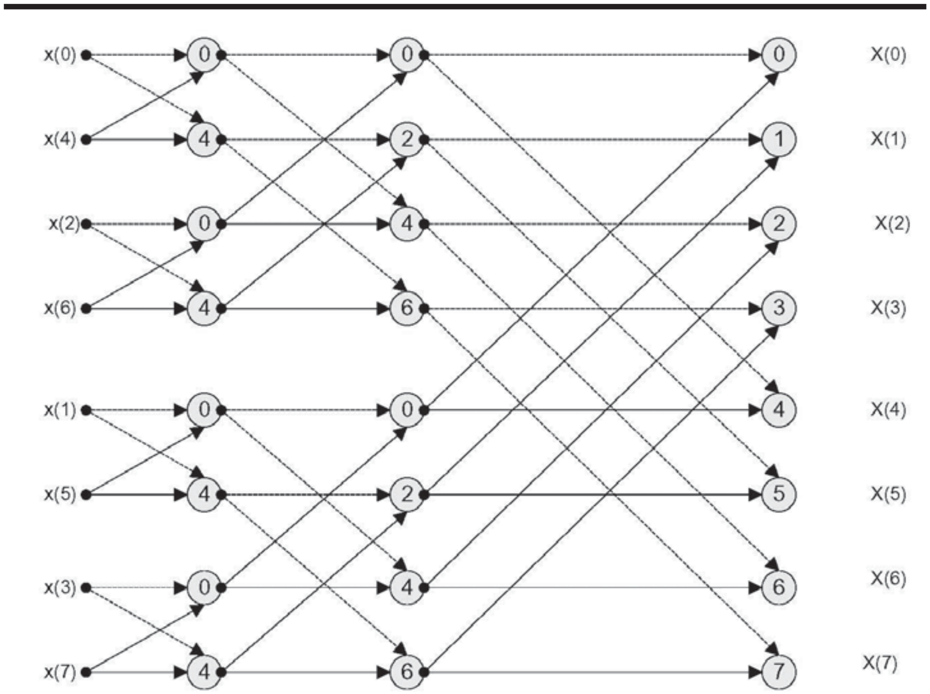


Figure 6 — A complete flow diagram for an 8 element Fast Fourier Transform (FFT) calculation. Each solid line represents a direct contribution of the element to the left. Each dashed line represents a contribution that is multiplied by the complex operator. Each number in a circle represents the *k* value for the W^k complex operator that is used for the multiplication.

[1]Notes appear on page 00.

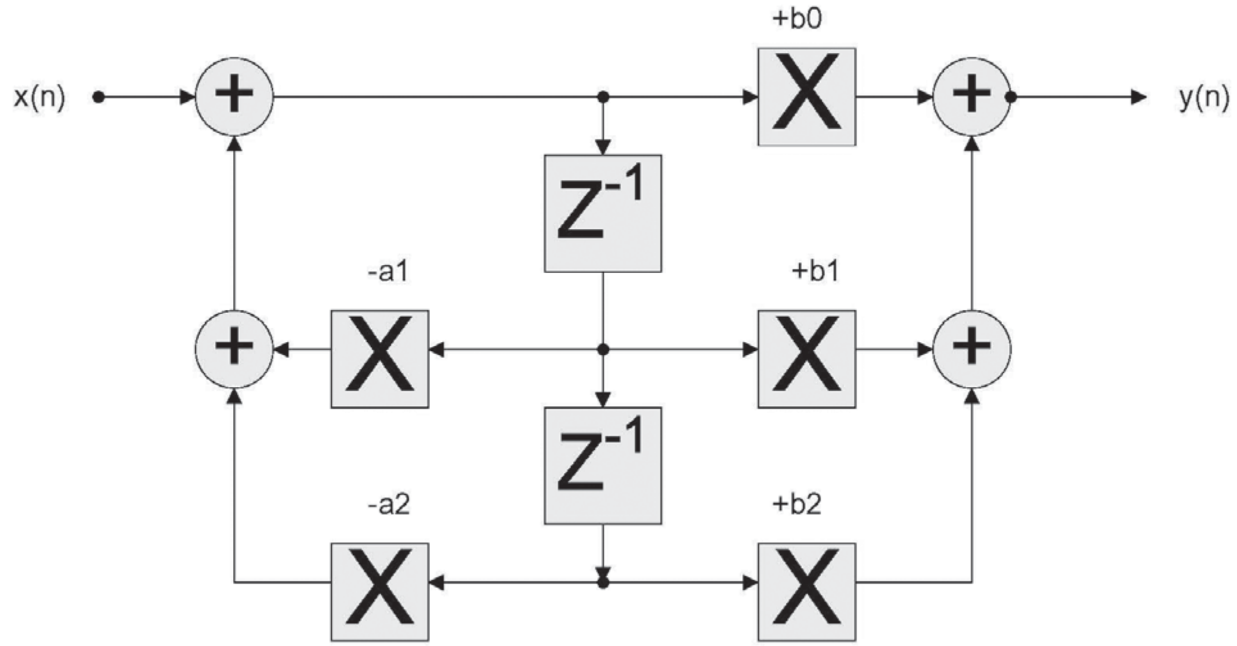


Figure 7 — A *z* space representation of an Infinite Impulse Response (IIR) filter. Each "b" value is a forward time contribution to the output, as in a Finite Impulse Response (FIR) filter, and each "a" value is a feedback element.

The result is that an *N* element FFT only requires an array of *N* registers to hold the entire data. The fixed sizes allow hardware implementation of an FFT with a very modest cost in resources. Our eight element FFT requires eight complex registers, logic to scramble the input as the data arrives, just eight *W* values (because the sine and cosine values will all use the same array from $-\pi$ to π), and eight multipliers. For this reason it is very easy for a processor such as the TMS320C5535 to include a fixed length (1024 in this case) hardware FFT block.

IIR Filters

We looked recently at Finite Impulse Response filters (FIR), which are non-recursive. They have the advantage that one can easily (more or less) calculate the filter tap values from a basic filter shape and apply a window function to minimize the Gibbs phenomenon. They also have the advantage of linear phase response and constant group delay. The biggest disadvantage to FIR filters is the large number of taps necessary for filters with maximum flatness and best transition band performance.

Infinite Impulse Response filters are a special case of a recursive digital filter. A recursive filter is built so that some of the output is fed back to the input. In an IIR filter, the output will continue to change forever even after the input signal is removed. It is possible to build recursive filters that do not have infinite impulse response but they are not very interesting.

I have been asked if there are programs to design IIR filters. Fortunately, the answer is yes. A group called *Octave*-forge has a large array of signal processing functions that work with *Octave* to perform an array of DSP design and analysis functions. We are interested in four functions: butter, cheby1, cheby2, and ellip. These let us design Butterworth, Chebyshev with pass band ripple, Chebyshev with stop band ripple, and elliptical filters. Each of these will return a filter design in a number of different formats. We will be interested in the z coefficient form. *Octave* tries to be as compatible with *MATLAB* as possible. I have found that it is frequently better to look at the *MATLAB* documentation on line rather than the documents from *Octave*-forge. The *MATLAB* documentation is professionally written and is easily understood. Frequently, the *Octave* documentation does not even exist.

The continuous time analysis uses the LaPlace transform *s* space to define a frequency response. In general, the response looks like:

$$H(s) = \frac{1 + a_0 s + a_1 s^2 + a_2 s^3 + \ldots}{1 + b_0 s + b_1 s^2 + b_2 s^3 + \ldots} \quad \text{[Eq 6]}$$

We factor the numerator polynomial to get the zeros of the response and factor the denominator polynomial to get the poles of the response. The response for a Butterworth filter is:

$$H(s) = \frac{1}{1 + s^{2N}} \quad \text{[Eq 7]}$$

where *N* is the order of the filter. The math allows us to convert from the polynomial form for an analog filter to an equivalent IIR digital filter. The equation follows the form:

$$H(z) = \frac{a_0 z^{-1} + a_1 z^{-2} + a_2 z^{-3} + \ldots}{b_0 z^{-1} + b_1 z^{-2} + b_2 z^{-3} + \ldots} \quad \text{[Eq 8]}$$

The filter design functions in *Octave* return two arrays containing the coefficients of "a" and "b" for the filter being designed. The filter design follows the method of Figure 7. The "a" coefficients appear on the left side of the system and apply to the feedback operation of the filter. The "b" coefficients appear on the right side of the system and apply to the forward operating part of the filter. Most of the filter design methods for IIR filters attempt to use polynomial representations from analog filters and adapt them to the *z* transform polynomials. Designing an analog filter using polynomial synthesis is a non-trivial process that involves a lot of math to achieve the desired features of a filter. The same is true of designing an IIR filter using those methods. That is one of the main reasons you won't see a lot of IIR filters in the literature; it is just a whole lot easier to do an impulse response and a DFT to generate an FIR filter from a desired frequency response.

Notes

[1]William Press, Saul Teukolsky, William Vetterling and Brian Flannery, *Numerical Recipes in* C*; The Art of Scientific Computing*, Cambridge University Press, 1992. This publication is available free on line at: **apps.nrbook.com/c/index.html**.

[2]The software for this column is available for download from the ARRL *QEX* files website. Go to **www.arrl.org/qexfiles** and look for the file **11x12_Mack_SDR.zip**.

[3]McClellan, Burrus, Oppenheim, Parks, Schafer, Schuessler , Computer Based Exercises for Signal Processing Using MATLAB 5, Prentice-Hall, 1997

QEX

Letters to the Editor

Stabilizing Your Transceiver Frequency Using GPS and Rubidium Reference Sources (Mar/Apr 2012)

Hi Larry,

Figure 1 of my article, on page 10, has an error on the schematic diagram of the overall system. On the right side, at the Elgin clock module, there is a 300 Ω resistor shown connected between ground at the bottom and the ground connection of the clock module. In reality, that resistor should be shown as a 3 kΩ resistor, and the top lead should go to a connection point between the 150 Ω resistor and the 0.1 µF capacitor. Builders should also change the 150 Ω resistor to a 1 kΩ resistor. I have experimented more with the circuit, and found that the original resistor values created a voltage divider that was nearly overloading the PLL output with a total load of 450 Ω.

On Figure 2, the schematic of the PLL board, I found a missing connection between the common ends of R1 and R2 to V_{CC}. These two resistors make "pull-ups" for the clear/preset pins (pins 1, 4 and 13,10) of U6A and U6B. There should be a connection shown to V_{CC} on pin 14 of U6A.

U5 is a 74HC393 IC, and not a 74HC390 as shown. Also, C14 has a value of 10 µF, not 10 pF as would be implied by no unit or multiplier shown on the schematic diagram.

I also have some information to pass along to interested readers. I contacted the eBay dealer (l.fluke), who was selling hundreds of LCD monitors for GPS. He informed me that all monitors are sold out and he is not expecting more.

Fortunately, we can follow the nice project described by Didier Juges, KO4BB, about building such a monitor for the Thunderbolt GPS. Didier has provided all of the details about the very simple hardware and software. Please go to **www.ko4bb.com/Timing/GPSMonitor**.

Thanks.

— Best regards, Eugene Woloszcuk, W6EAW, 494 Curie, Dr, San Jose, CA 95123; **ea114w42@hotmail.com**

Hi Eugene,

Thank you very much for pointing out these schematic diagram corrections for our readers. I apologize for those drawing errors. Thanks also for the additional information about the LCD monitor.

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

Hi Larry,

I was a little bit concerned by some aspects of the article "Stabilizing Your Transceiver Frequency Using GPS and Rubidium Reference Sources" in the March/April 2012 issue of *QEX*.

My main concern is that the specified FEI 5680A Rubidium "brick," comes in a variety of "flavors," and the majority of those currently on eBay do not have the programmable DDS output that the author of the article has, and can only be adjusted to give an output near 10 MHz. As such, they will be of no use for deriving the oddball frequency that the author required for his Yaesu radio. This might in any case be more easily derived from one of the Chinese DDS modules widely found on eBay for under $10. I did find one of the 5680A modules that can be programmed on offer, but it was considerably more than $40. If you go for the cheap ones you will probably be stuck with a unit that is indeed useful in some ways, but which will not be adaptable in the way the author describes.

It is quite hard to tell from the eBay ads exactly which variety of 5680A is being offered. FEI apparently made many different versions of this unit, and have not been forthcoming with much information about them. The ones marked FEI P/N 217400-30352-1 seem to be the ones that don't have the programmable DDS output, but only the 10 MHz one. These also seem to require a 5 V supply to one pin in addition to the 15 V main supply. The ones that have the DDS, and therefore are programmable, seem to be more difficult to find. When I checked right after I read this article, I saw three of the programmable units, all for around $120 including shipping. It seems as though the price may be the best guide, although I did see one of the non-programmable ones on offer for $140. It seems to be a bit of a minefield!

The Thunderbolt GPS LCD monitor that Eugene specifies seems to be a copy of a design by Didier Juges, KO4BB, details of which are on his website. The KO4BB monitor can be set to display correct UTC or GPS time directly so there is really no need to have a complicated method of displaying UTC by way of a commercial digital clock as W6EAW describes.

There has been endless discussion of the LCD monitors on the "Time Nuts" list (**www.leapsecond.com/time-nuts.htm**). I would refer readers to Didier's wiki, which has pages on these units: **www.ko4bb.com/dokuwiki/doku.php?id=precision_timing:fe5680a_faq**. Here is also a link to another page that apparently identifies four different varieties of 5680A: **http://vk2xv.djirra.com/tech_rubidium.htm**.

It might also be worth pointing out that the Thunderbolts can be commanded via the Tbolt Mon program to output either GPS time or UTC and date directly to the ko4bb monitor which it then displays continuously, so there is not really any need for an elaborate extra external device to display time. Obviously, that is a personal choice.

— 73, Dan Rae, AC6AO / G3NCR, 1618 Wellesley Dr, Santa Monica, CA, 90405; **danrae@verizon.net**.

Hi Dan,

Thanks for taking the time to write about this article, and for passing along quite a bit of additional information.

—73, Larry Wolfgang, WR1B

New Results on Shortening Beverage Antennas (Jul/Aug 2012)

Dear Larry,

In the July/August issue, on page 31, the two 100 µF electrolytic capacitors in the K9AY control box are shown incorrectly. Their polarities need to be reversed. That is also true for the 100 µF electrolytic capacitor near the middle of the page, for the relay coil.

*—73, Kenneth Hansen, KB2SSE, 10 Maple Rd, Ringwood, NJ, 07456;***kb2sse@arrl.net**.

Hi Kenneth,

You are absolutely correct. We drew those capacitors with the wrong polarity. In each case that you mentioned, the capacitors should be shown with the straight and curved sides reversed. The + sign, which indicates that it is an electrolytic capacitor is always on the straight line side.

—73, Larry Wolfgang, WR1B

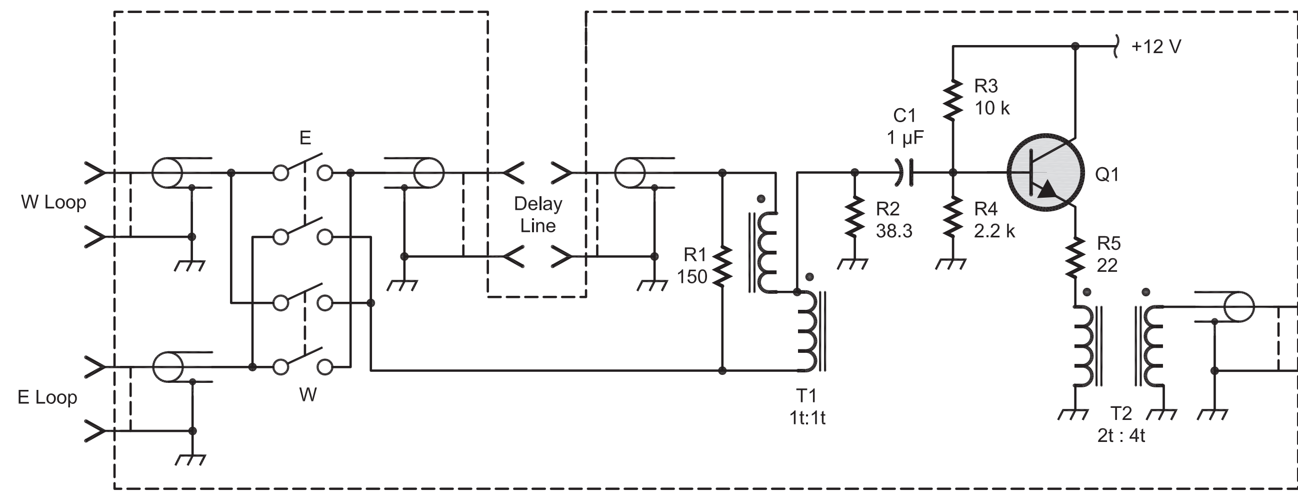
Introducing the Shared Apex Loop Array (Sep/Oct 2012)

Hi Larry,

Thanks for publishing the *QEX* article for the Shared Apex Array. It has generated a lot of interest, and some good questions from readers.

There is one significant error in the schematic diagram of Figure 2, where the transistor is shown upside down. The collector should connect to the +12V supply and the emitter to R5.

In addition, some people have asked for further clarification on the construction of T1 and T2. The transformers are wound on Laird Technologies LFB095051-000 cores. These ⅜ inch diameter by ¾ inch long ferrite beads are available from Digi Key, catalog no. 240-2277-ND. T1 is six bifilar turns, QX1209-Bauman02



wound on a single bead. T2 has a 2 turn primary and a 4 turn secondary wound on a single bead.

— Mark Bauman, KB7GF, 1910 Sunflower Court, College Place, WA 99324; **kb7gf@arrl.net**

Hi Mark,

Thanks for being the first to call my attention to that schematic error. I should have noticed that when I looked at the schematic diagram. We are including a copy of the corrected diagram here, and I have also posted a copy on the ARRL *QEX* website (**www.arrl.org/this-month-in-qex**).

—73, Larry Wolfgang, WR1B

F_a: Measurement and an Application to Receive Antenna Design (Sep/Oct 2012)

Dear Editor,

Robye Lahlum's article on "Radio Noise" in the Sep/Oct 2012 issue of *QEX* is very interesting. The reference to ITU-R Rec. P372-7, however, is way out of date. This edition was in 2001: the current version, dating from 2010, is ITU-R Rec. P.372 -11. Versions 7, 8 and 9 have been superseded.

ITU-R Recommendations can be downloaded for free from the ITU-R website: **www.itu.int**. For a direct link to the Recommendation, go to **www.itu.int/rec/R-REC-P.372/en**.

I do not think it will make any difference to the calculations in the article. What is now questionable is how accurate the figures are for the urban cases, in view of the spread of in-home BPL, especially in Europe. Certain papers delivered to ITU-R SG1 WP1A suggest there could be quite an increase because of aggregisation. I am an IARU Technical Consultant who attends Study Group 1 and ITE Working Parties, so I have seen these papers.

Some work has been carried out by CEPT countries on measurements of radio noise, although I understand little has been done at HF. The RSGB Propagation Studies Committee are, I believe, attempting to get some measurements made, while I am told the ITU-R figures are hopelessly over pessimistic in certain areas of the world — South Georgia, South Shetlands and so on! The measurements that I made ("HF Receiver Dynamic Range: How Much Do We Need?" May/Jun 2002 *QEX*, "Receiver Parameters for Contesters," Mar/Apr 2008 *NCJ*) suggest that at least at my QTH (rural), the noise levels have stayed relatively constant over the years, at least as far as 7 MHz is concerned.

— 73, Peter Chadwick, G3RZP, Three Oaks, Braydon, Swindon SN5 0AD, England; **g8on@fsmail.net**

Hi Peter,

Thanks for the information and the URL for the newer ITU Recommendation. At that address I did find Recommendation 372-10, but not 372-11. 372-11 may exist but I do not seem to be able to access it.

I did, however, review the numbers shown for manmade noise in the last three issues of p.372; 372-7, 372-9, and 372-10. The c and d values in section 5, table 1 of 372-10 (which is the latest version that I can access) are the same values that I used in the *QEX* article (Table 2 of my article.)

Location Type	*c*	*d*
City	76.8	27.7
Residential	72.5	27.7
Rural	67.2	27.7
Quiet Rural	53.6	28.6

F_a was calculated for Table 2 in my article from the *c* and *d* values by:

$F_a = c - d \log f$ (per P.372)

As pointed out in 372-10, the F_a values for manmade noise are based on measurements made in the 1970s, and that these values may change with time. Looking over the manmade noise data from 372-7, 372-9, and 372-10, I notice that the values for *c* and *d* are identical for all three issues of P.372, and appear to be all based on the 1970 measurements. If P.372-11 has any new data or new values for *c* and *d*, it would be of interest to see what that information is.

Thanks

— 73, Robye Lahlum, W1MK, 45 Boxford, MA 01921; **w1mk@arrl.net**

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