QEX

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September/October 2016

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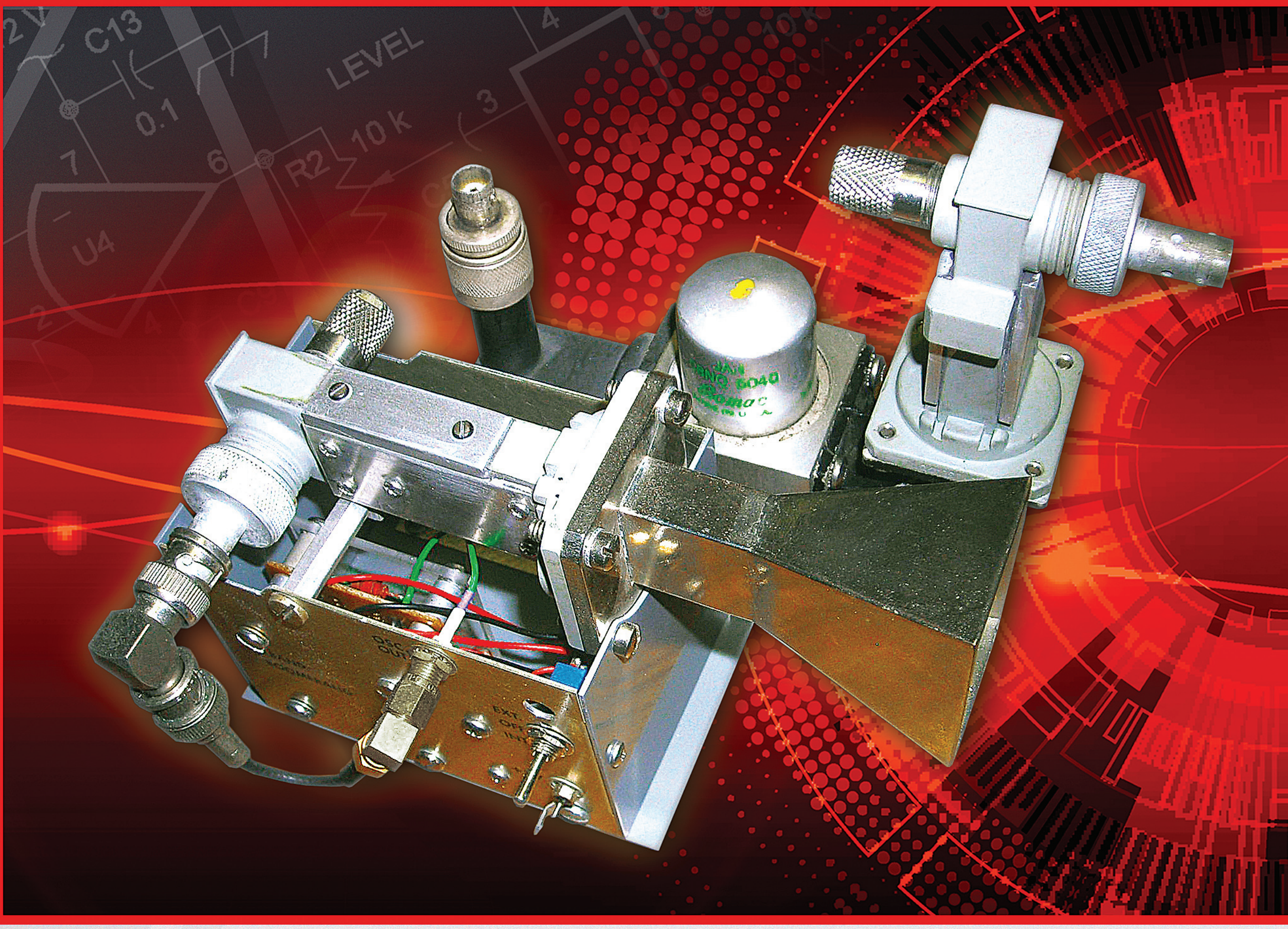
A Forum for Communications Experimenters

Issue No. 298



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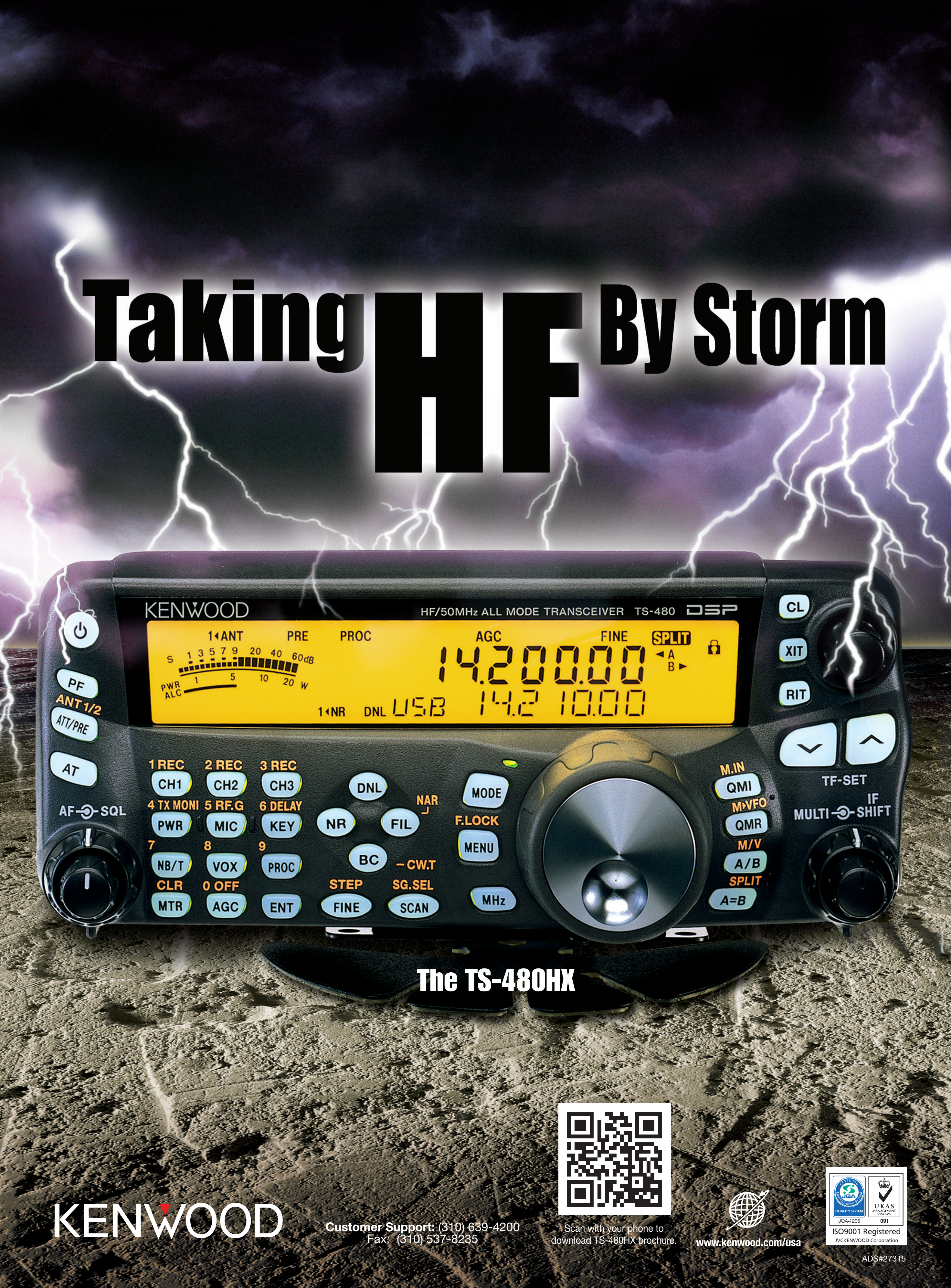

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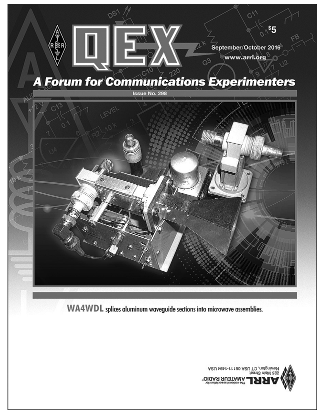
September/October 2016

About the Cover

John M. Franke, WA4WDL shows how parts of aluminum waveguide components can be spiced together and reinforced to make needed waveguide assemblies.



In This Issue

Features

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

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

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Kazimierz "Kai" Siwiak, KE4PT

Perspectives

621.384 Then and Now

As a high school student during the early 1960s, my "search engine" was the card catalog of my neighborhood Brooklyn Public Library. The Dewey Decimal System 621.384 classification — "Radio" — was my search parameter. I borrowed and absorbed books on radio topics from a small treasure trove of titles in that classification. I also began to absorb a somewhat limited collection of old issues of *QST* and other Amateur Radio magazines available for reading in the reference room of the library. I photo copied negative images of likely ham projects at a then-precious nickel a page. So began my ham radio career, and my professional engineering career.

Times change. By 1990 we were introduced to a variety of pre-web and web search engines, including *Archie*, *Gopher*, *Altavista*, and *Ask Jeeves* among many others. Sophistication and convenience increased with time. Now we have the Internet, with graphical interface search engines, and whole libraries of materials online. The entire online universe is now my neighborhood library. My personal radio library still includes book classics by Kraus, Jordan and Balmain, Terman, Jakes, and many ARRL titles. But my searches are now online. *QST* is no longer confined to the reference room of the library; it is available online in its entirety.

On the professional side, vast libraries of professional journals and reports can be searched and downloaded from online portals like the IEEE® Xplore Digital Library. My 621.384 classification has expanded to the boundaries of the Internet. Google "621.384" and you will find a library of ARRL book titles.

What of the future? What will our *QEX* authors produce for future hams and prospective hams to explore?

In This Issue

You, the reader and prospective author determine the content of *QEX*; the character and content is driven by you. Please put your favorite topic or innovative measurement on paper, and share it on these pages. Just follow the details on the **www.arrl.org/qex-author-guide** web page, and contact us at **qex@arrl.org**. We value your feedback, comments and opinions about these pages.

In this issue, our *QEX* authors describe making microwave assemblies and printing horn antennas, crystal test oscillators, low frequency receiving antennas, and RF filters. Fred Brown, W6HPH, built universal oscillators to test a wide range of fundamental and overtone crystals. Gary A. Appel, WAØTFB, shows how to modify staggered LC resonators to implement the filter poles and simplify the tuning of the staggered filter, removing the need for high bandwidth operational amplifiers. John Franke, WA4WDL, splices sections of aluminum WR90 waveguide to make microwave assemblies. Michelle Thompson, W5NYV, and Kerry Banke, N6IZW, use a 3-D printer to make microwave horn antennas. Rudy Severns, N6LF, presents a study of a receiving array for 160 m through 2200 m bands. Rosser B. Melton, AD5MI, presents a practical approach to inductance, based on physical principles.

Please continue to support *QEX*, and help it remain a strong technical publication.

73,

Kazimierz "Kai" Siwiak, KE4PT

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Staggered Resonator Filters using LC Resonators

A modified staggered resonator filter using LC resonators to implement the filter poles simplifies the tuning of the filter and removes the need for high-bandwidth operational amplifiers.

I needed a band-pass filter, centered at 16 kHz to limit the signal energy into a digital signal processing (DSP) chip. Several different implementations are available. The two options that I immediately considered were a coupled resonator filter, and a staggered resonator filter implemented with operational amplifiers. In the end, neither option was suitable. Instead I selected a modified staggered resonator filter using *LC* resonators to implement the filter poles. This implementation simplified the tuning of the staggered filter, and removed the requirement for high bandwidth operational amplifiers. Before looking at the chosen implementation, let's take a brief look at the strengths and shortcomings of the coupled resonator filter.

The Coupled Resonator Filter

The first option that I considered was the coupled resonator filter. The coupled resonator filter is well documented and easy to tune, with many options for determining the component values. One feature of the coupled resonator filter, that is not usually mentioned, is the ability to choose the coupling elements to provide additional attenuation in either the lower or the upper filter skirt. Figure 1 shows the schematic diagram for a four resonator coupled resonator filter designed to have a nearly symmetrical response. The filter was designed using equations presented in *Reference Data for Radio Engineers*, with some additional manual tweaking required to optimize the passband ripple.[1] Note the use of inductors L2 and L4 for coupling between the first and second, and third and fourth resonators, to improve the high frequency rejection of the filter. The more common implementation, using all capacitors as coupling elements, will result in a very soft upper skirt on the filter.

The calculated response of the filter, assuming lossless elements, is shown in Figure 2. The symmetry of the filter is quite good. Unfortunately, lossless elements are hard to find. Figure 3 shows the calculated response of the coupled resonator filter with inductor *Q* values of 70, along with the lossless response. The loss is nearly 2 dB in the passband, which is probably not a problem, but the inductor loss also distorts the passband, with increasing loss as the signal approaches the passband limits. This again may not be a problem, but other crystal and *LC* filters in the system already exhibit this characteristic, and we would prefer a flat passband to avoid compounding the passband distortion. Another problem would be the realization of the coupling inductors. Obtaining a 12.55 mH inductor with a good *Q* in a small space, might be quite a challenge. Given the obstacles to realizing a 16 kHz coupled resonator filter with good performance, I decided to look at other options.

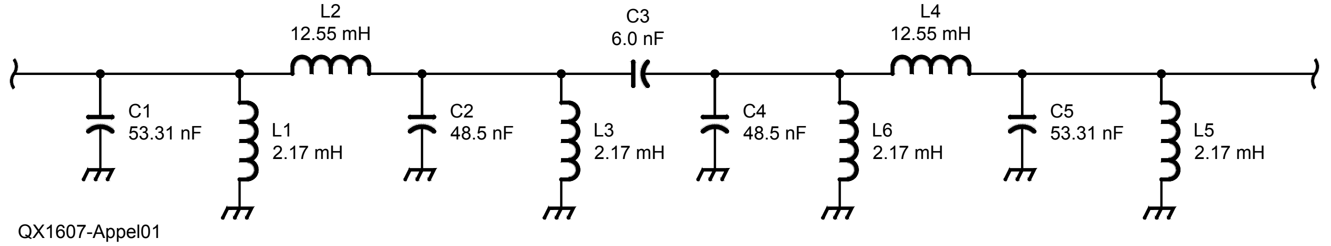


Figure 1 — Mixed coupling elements have been chosen in this coupled resonator filter to achieve a more symmetrical filter response.

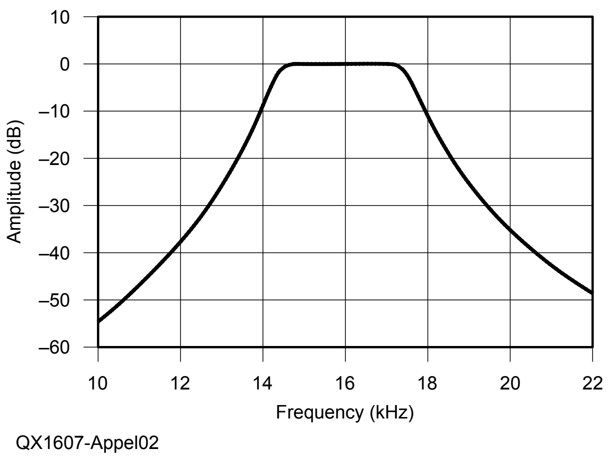


Figure 2 — The mixed coupling provides a filter response that is nearly symmetrical in the skirt response.

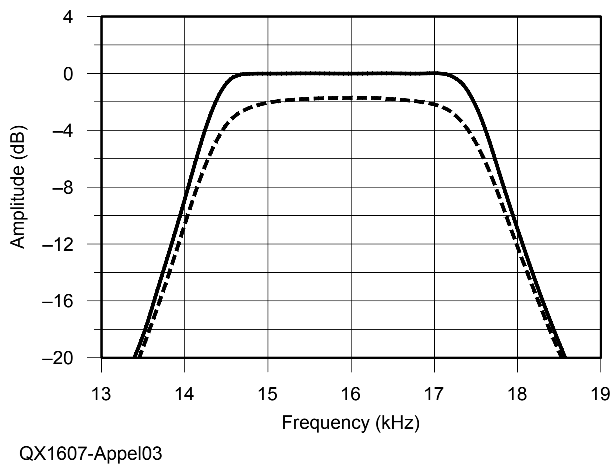


Figure 3 — The finite *Q* of the inductors adds loss to the filter, and also distorts the passband response.

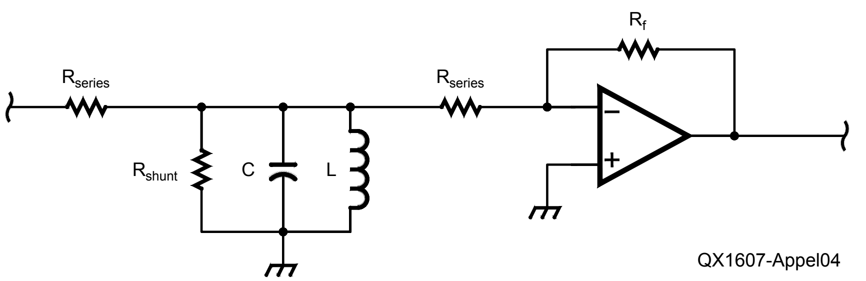


Figure 4 — The staggered resonator filter can be implemented using *LC* resonators, with low frequency operational amplifiers to provide gain and isolation between the stages.

The Staggered Resonator Filter

The staggered resonator filter using operational amplifiers to simulate *LC* resonators is a popular implementation for low frequency audio filters. No inductors are required for this implementation, and the passband response is not degraded by loss. The passband response of a properly designed and tuned staggered resonator filter will follow the theoretical response. One disadvantage of the staggered resonator filter is that we do not have the option of tailoring the response to optimize the high side or low side filter skirt — the filter will display geometric symmetry. The response below the passband will fall off faster than the response above the passband. As the percentage bandwidth increases, this characteristic will become more pronounced. An initial design for a four resonator filter similar to the coupled resonator design was accomplished using *FilterPro Desktop*, a free application available from Texas Instruments.[2]

This filter requires four operational amplifiers, with gain bandwidth requirements up to almost 40 MHz. Each stage requires two capacitors and at least two resistors to set the center frequency and *Q* of the simulated resonator. Each of the components must be a precision part, or must be tuned to optimize the filter performance. Given the number of precision components required, and requirement for high-bandwidth operational amplifiers — which I did not have on hand — and the difficulty of tuning, I looked for an alternative solution.

My next thought was to use the staggered resonator approach by implementing the resonators with *LC* resonators rather than operational amplifiers. I would still require op amps to provide gain and isolation, but they could be lower frequency amplifiers, capable of providing gain in the passband of the filter. Resonator tuning would be fairly simple. Adjust either the inductance or capacitance to set the center frequency of the resonator, and adjust the resistive loading on the resonator to obtain the required *Q*. While we saw that the finite inductor *Q* values will degrade the response of the coupled resonator filter, in the case of the staggered resonator filter, the loss needs only to be taken into account in establishing the resonator *Q* values. I decided to give this modified filter implementation a try.

The Alternative Staggered Resonator Design

Figure 4 shows a single stage of the alternative circuit. The resonator is implemented as a parallel *LC* resonator, with the *Q* determined by the shunt resistance which each of the two R_{series} resistors appear in parallel with the equivalent shunt resistance of the inductor due to the inductor loss, and shunt resistor R_{shunt} that has been included to allow fine adjustment of the resonator *Q*. Note that the R_{series} resistor to the right of the resonator is tied to the virtual ground of the operational amplifier, and the R_{series} resistor to the left of the resonator must be driven by a low impedance source. Since each stage delivers the output signal from an operational amplifier, the stages can be cascaded to provide the required low impedance source to the following stage. An operational amplifier would normally be required to drive the input of the first stage as well.

For my initial design I again used the *FilterPro Desktop* application to obtain the center frequency and *Q* for each of the resonators.[3] These resonator parameters are provided along with the component values for the conventional staggered resonator filter design. The poles in the filter appear in pairs, plus a single resonator if the filter order is odd. Each pair contains one resonator below the center frequency, and one above the center frequency, with identical *Q* values. My design required one pair of resonators with a Q of 13.988, and another pair of resonators with a Q of 5.776. The required resonator Q will influence the selection of the core material for the inductors. A higher Q requirement for the resonator will require a higher Q material to maintain gain in the resonator stage, as well as a more temperature stable material. Temperature drift in the higher Q resonators will cause more passband distortion than temperature drift in the lower Q resonators. For my design I selected powdered iron core toroids for the higher Q resonators and ferrite cores for the lower Q resonators. While the ferrite cores have proven acceptable in this application, they would probably not be appropriate in a location that is subject to significant temperature changes.

Calculating the Element Values

The first resonator in my design requires a resonant frequency of 14.673 kHz, and a Q of 13.988. Because I had a number of 100 nF film capacitors, I chose to start the design with that value for the capacitor, resulting in a reactance of about 108.5 Ω at the resonant frequency, and requiring an inductance value of 1.177 mH to realize the required resonant frequency. Since we will need to tune the resonator to bring it on to frequency, it's best to begin with an inductor value that is just a bit low, allowing the resonator to be tuned by adding in fixed value capacitors to bring the resonator on to frequency. If an inductor is being fabricated it might be best to first put on a few extra turns, then remove turns until the resonator frequency exceeds the design value. We can then add the fixed value capacitors tor bring the resonator down to the required frequency.

Along with tuning the resonator frequency, we will also need to adjust the resonator Q. Given that the reactances of the inductor and capacitor are 108.5 Ω at the resonant frequency, we can calculate the required shunt resistance as the reactance times the Q, giving a total shunt resistance value of about 1.52 kΩ to achieve the required resonator Q. The iron core inductor Q measured about 70 at the center frequency, which will result in an equivalent shunt resistance of 7.59 kΩ that is due to the inductor loss. Taking the resonator loss into account, an external shunt resistance of 1.90 kΩ will be required to establish a total shunt resistance of 1.52 kΩ. Since R_{series} appears in shunt with the resonator twice, each of these resistors must be at least twice this value, at least 3.8 kΩ. The closest 5% resistor exceeding this value is 3.9 kΩ, which we will select for the two series resistors, providing an external shunt resistance of 1.95 kΩ. To bring this resistance down to the required 1.9 kΩ will require an additional 74 kΩ for the resistor at R_{shunt}. We might start by choosing 75 kΩ knowing that some adjustment of R_{shunt} will probably be required to establish the required resonator Q. As a check, the parallel equivalent of the 7.59 kΩ resistance due to the resonator loss, the two 3.9 kΩ R_{series} resistors, and the 75 kΩ R_{shunt} resistor is 1.52 kΩ, validating the selected resistor values. Finally, I chose a value of 10 kΩ as the starting point for R_f, which can be adjusted as required to establish the gain of the stage.

Tuning the Resonator

If we want the completed filter to match the theoretical response, we need to make sure that the resonators are all tuned properly. This requires setting the center frequency, and the Q of each resonator. To set the center frequency, a signal generator is swept over the resonator frequency. The signal generator must be buffered to assure a low impedance drive source. Since each stage in the filter is likely preceded by an operational amplifier, we can just inject the signal generator into the operation amplifier preceding each stage. For intermediate stages this can be accomplished by lifting the second (rightmost) R_{series} resistor on the previous stage and injecting the test signal into it. The first resonator should also be driven by an operational amplifier to provide the required low impedance source.

While the signal is swept over the resonant frequency of the resonator, the peak signal voltage at the output of the resonator stage is noted. The test signal is then swept up and down from the center frequency, noting the frequencies at which the response has dropped to 0.707 times the peak voltage — the 3 dB bandwidth points. The center frequency f_0 is then the geometric mean of these two frequencies.

$$f_0 = \sqrt{f_L f_H}$$

The Q is given by the center frequency of the resonator divided by the 3 dB bandwidth

$$Q = f_0 / \left(f_H - f_L \right)$$

where f_H is the upper 3 dB frequency and f_L is the lower 3 dB frequency. Because the reactance of the inductor and capacitor is a function of the center frequency, the Q is also a function of the center frequency. On the other hand, the center frequency is not a function of the resistive loading, so the center frequency of the resonator should be adjusted first. The center frequency varies inversely as the square root of the capacitance, or equivalently, the required capacitance varies inversely as the square of the frequency. Given the known value of capacitance, the resulting resonant frequency, and the desired resonant frequency, we can calculate the value of capacitance C' needed to bring the resonator on frequency from the expression,

$$C' = C \times \left(f_{meas} / f_{res} \right)^2$$

where f_{meas} is the measured center frequency, f_{res} is the desired center frequency of the resonator, and C is the existing resonator capacitance. If we have set the resonance a bit high in frequency, the required capacitance C' should be somewhat greater than the existing capacitance. We can then select a fixed capacitor value equal to the difference between the required capacitance and the existing capacitance, and place it in parallel with the existing capacitor to bring the resonator to the desired frequency. Because of errors in measurement, and tolerance in the capacitors, this measurement and correction will likely need to be repeated to fine tune the resonator.

Once the resonator has been set on frequency, the Q can be adjusted. We would normally like to set the Q just a bit high, so we can reduce the Q by adding resistance in shunt with the resonator. We measure the resonator Q, as discussed above, and use that Q to determine the actual total shunt resistance,

$$R_{actual} = QX$$

where X is the reactance of either the inductor or the capacitor at the resonant frequency. This total shunt resistance will include the two R_{series} resistors, the R_{shunt} resistor, and the equivalent shunt resistance due to the inductor loss. We already know what value we wanted for the total shunt resistance $R_{desired}$, as determined from the desired resonator Q and the reactance of the inductor and capacitor at the resonant frequency, as discussed above. If the existing value of R_{actual} is greater than $R_{desired}$, which was achieved by setting the Q higher than the design value, we can reduce the resonator Q by placing an additional resistor in shunt with the resonator. The shunt resistance R_{trim} required to lower the Q can be determined by calculating the parallel equivalent of $R_{desired}$ and the negative of R_{actual},

$$R_{trim} = \left(R_{desired}^{-1} - R_{actual}^{-1} \right)^{-1}$$

Again, $R_{desired}$ is our desired total shunt resistance and R_{actual} is the actual total shunt resistance determined from the measured bandwidth. This R_{trim} will be in parallel with R_{shunt}. The value of R_{shunt} can be adjusted, or an additional shunt resistor placed in parallel with R_{shunt} to reduce the Q. This adjustment may also need to be repeated to set the Q adequately.

Normally we will want to cascade the two resonators displaying the same Q so that we can measure the response of the two resonators in cascade. If the resonator Q is high enough, the cascade will result in two peaks. If the Q values are matched, then the two peaks will be at the same level. If the Q values are not matched, one peak will be higher than the other, which will result in a slope in the passband of the completed filter. This then allows us to match the Q values of the two resonator stages. The resonator that exhibits the higher peak, has the higher Q.

With each of the resonators tuned properly, the filter should display the desired frequency response. Inter-stage measurements should be made in order to determine the individual stage gains and to verify that the cascaded filter response at that inter stage is correct. It is likely that the value of R_f in each stage will require adjustment in order to maintain a desirable gain profile through the filter. If the gain is excessive at some point, then we need to be concerned about signal levels, and make sure that the amplifier does not saturate with a signal at any frequency. If the gain is too low, we might need to worry about the introduction of noise in that stage. If the total filter gain is too small, we may want to add a buffer between two of the stages to increase the filter gain.

An Easier Way

Calculation of the required component values to obtain the desired filter performance is a time consuming adventure. After going through the calculations multiple times, it seemed like a worthwhile effort to write a software program to perform the calculations. Figure 5 shows the opening window for the *Cascade Resonator Filter Designer*. As the application opens it displays the design for a four section Butterworth filter at a center frequency of 16 kHz, with a 3 dB bandwidth of 3 kHz. The bottom panel of the window lists the four resonators required to obtain the desired response. The two resonators labeled A form a complementary pair, with resonant frequencies of approximately 14.6 kHz, and 17.4 kHz. The Q of each resonator is 13.928.The two resonators labeled B form another complementary pair with resonant frequencies of approximately 15.4 kHz, and 16.5 kHz. The Q of each resonator is 5.751. The filter specification above the resonator list can be modified, and the new resonator requirements displayed. For example, we can change the desired response to a 0.1 dB Chebyshev response, with a ripple bandwidth of 3 kHz, as shown in Figure 6. Note that the Chebyshev response has resulted in a significant increase in the resonator Q values. Although not obvious, the increased resonator Q values will also impact the cascade gain — higher resonator Q values will require more gain in each stage in order to maintain the same total gain in the filter.

By double-clicking on one of the resonators we bring up a window displaying the element values for the selected resonator stage. Double-clicking on the first resonator brings up the *ResonatorForm* window

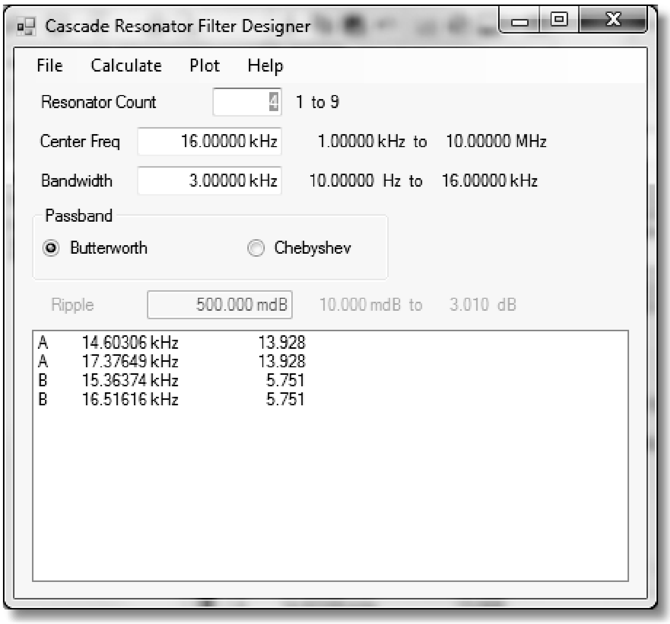


Figure 5 — The *Cascade Resonator Filter Designer* will display the resonator characteristics required to obtain the specified filter response.

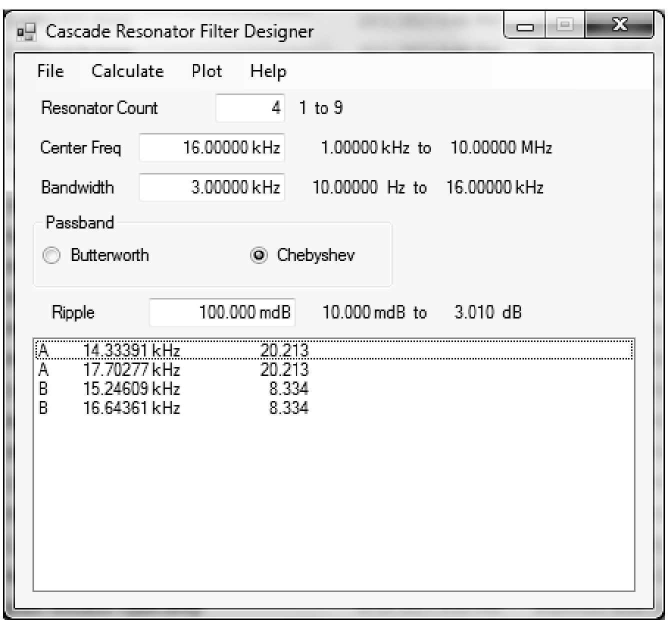


Figure 6 — Changes in the filter specification are immediately reflected in the resonator characteristics.

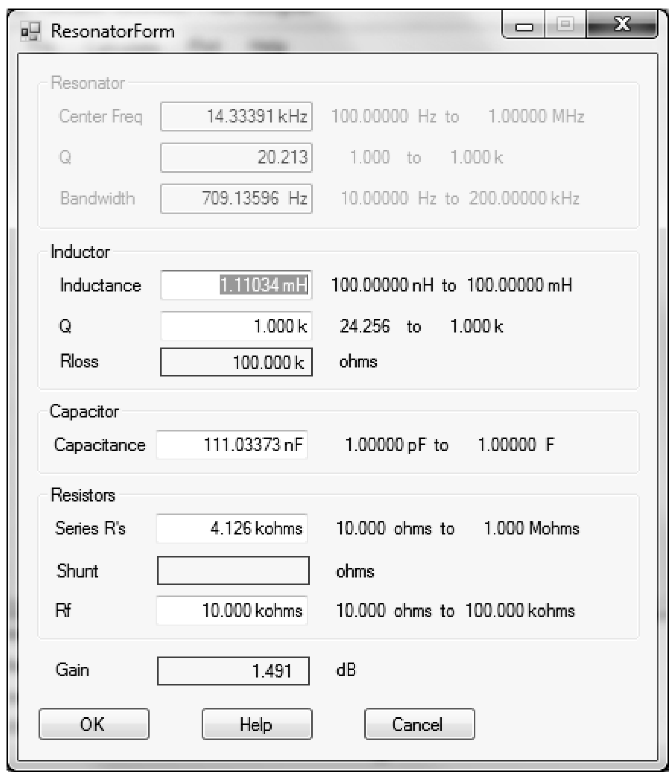


Figure 7 — Double-clicking on a resonator will bring up a window displaying the initial element values for the resonator stage.

shown in Figure 7. The first modification would normally be to change either the inductor value, or the capacitor value. We will start here by setting the resonator capacitance to 100 nF. Note that the default inductor Q is a very unreasonable value of 1000. We'll change that value to 70. With these changes a new value for the series resistors is calculated as 6.311 Ω, as shown in Figure 8. We could use a precision resistor value here. The closest 1% value greater than the displayed value is 6.34 kΩ. But even with a 1% tolerance, we'd still likely need to include an additional shunt resistor to set the Q accurately, so we'll just select a standard 5% value of 6.8 kΩ for the two series resistors. In Figure 9 we see that an additional shunt resistor with a value of 43.88 kΩ is required to maintain the required resonator Q. We could put in a resistor with a value of 43 kΩ, or we might want to set the resistance value just a bit higher, knowing that tweaking will likely be required during the tuning process. We have also changed the value of R_f to establish a peak gain in this stage of just over a 0.5 dB, as displayed at the bottom of the window. Note that the gain of the cascaded stages is not the sum of the stage gains. The displayed gain is the gain at the resonator peak frequency, and the resonator peaks are not aligned. Clicking on OK will close the window, retaining the new values.

Similarly, we can modify the element values for each of the stages. One possible solution for this Chebyshev filter is as shown in Table 1. The value of R_f for each stage has been set to 22 kΩ. The filter response shown in Figure 10 can now be displayed by selecting the Plot | Filter Response item from the main menu. With the selected values, the cascade gain is about -3.5 dB. If a unity gain is desired, it can probably be obtained by a slight adjustment in the value of R_f for one or more stages. If the filter gain is too low for correction by adjusting the values of R_f, a buffer can be inserted into the cascade by selecting one of the resonators, then right clicking and selecting Insert Buffer from the pop-up menu. A buffer can be appended to the cascade by selecting Append Buffer from the pop-up menu. The buffer gain can be changed by double clicking on the buffer stage.

The cumulative gain at the output of any of the stages can be displayed by selecting a stage, then right clicking on the Plot Cumulative Response item from the pop-up menu. An example plotting the filter gain through the first three stages is shown in Figure 11. This plot is useful for managing the gain profile through the filter cascade during the design process. It is also helpful in verifying the cascade performance during the tuning process. The plot of the cumulative gain at the last stage will be identical to the filter response shown in Figure 10.

Table 1

One possible set of component values to achieve the response of the 0.1 dB Chebyshev filter. All inductors are assumed to exhibit a Q of 70.

Frequency	*Q*	*L*	*C*	*Rseries, Ω*	*Rshunt Ω*	*Rf, Ω*	*Stage Gain*
14.334 kHz	20.21	1.23285 mH	100 nF	6.8 k	43.88 k	22 k	0.57 dB
17.703 kHz	20.21	808.27 µH	100 nF	5.6 k	29.20 k	22 k	2.11 dB
15.246 kHz	8.33	1.08974 mH	100 nF	2.2 k	9.66 k	22 k	11.94 dB
16.644 kHz	8.33	914.42 µH	100 nF	2.2 k	5.09 k	22 k	11.18 dB

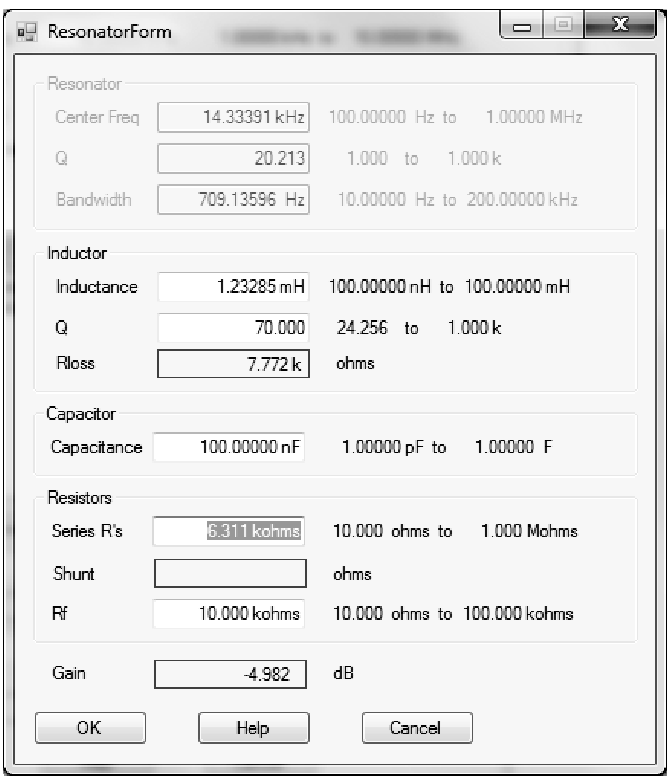


Figure 8 — Changing the capacitance value and inductor *Q* will return a new value for the two series resistors.

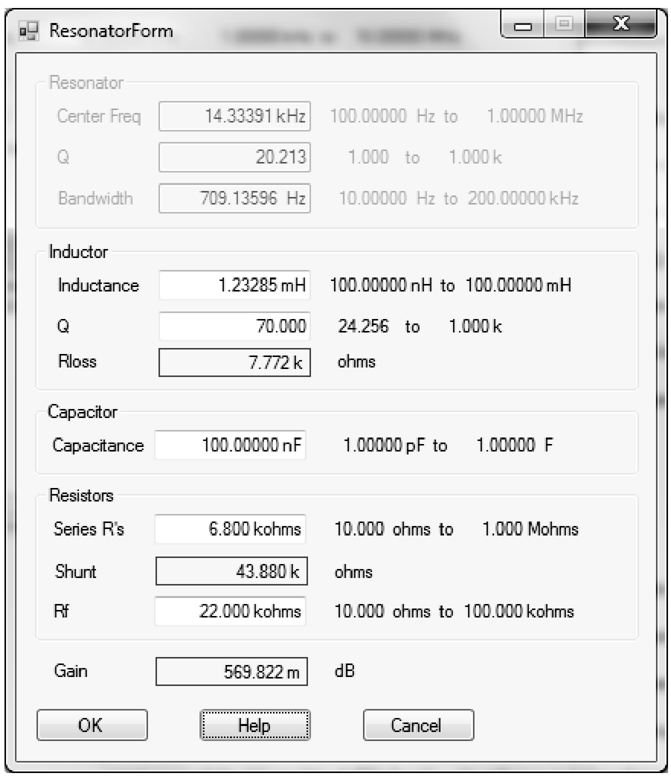


Figure 9 — Choosing a standard value for the series resistors will result in a calculation of the additional shunt resistance required to establish the desired resonator *Q*.

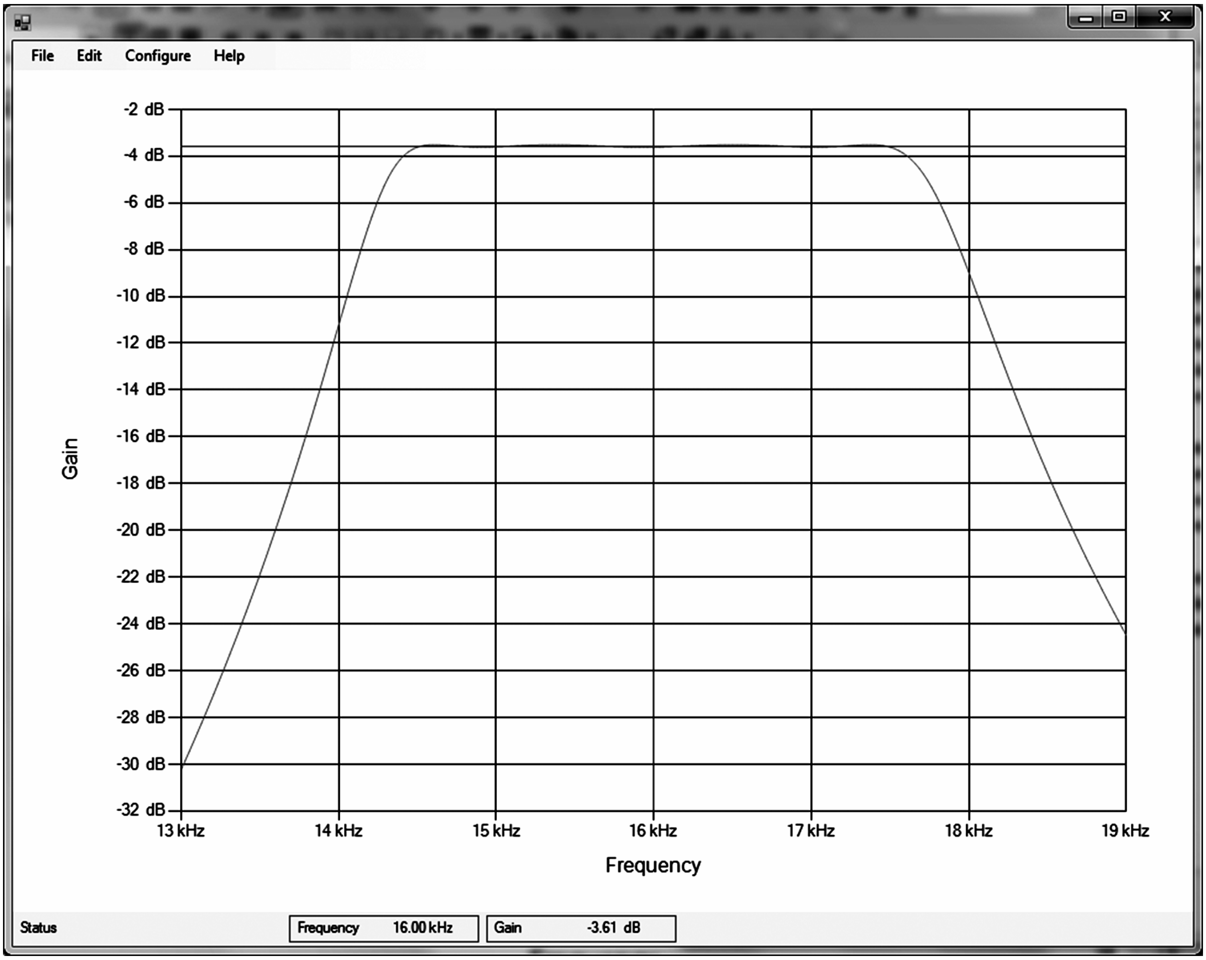
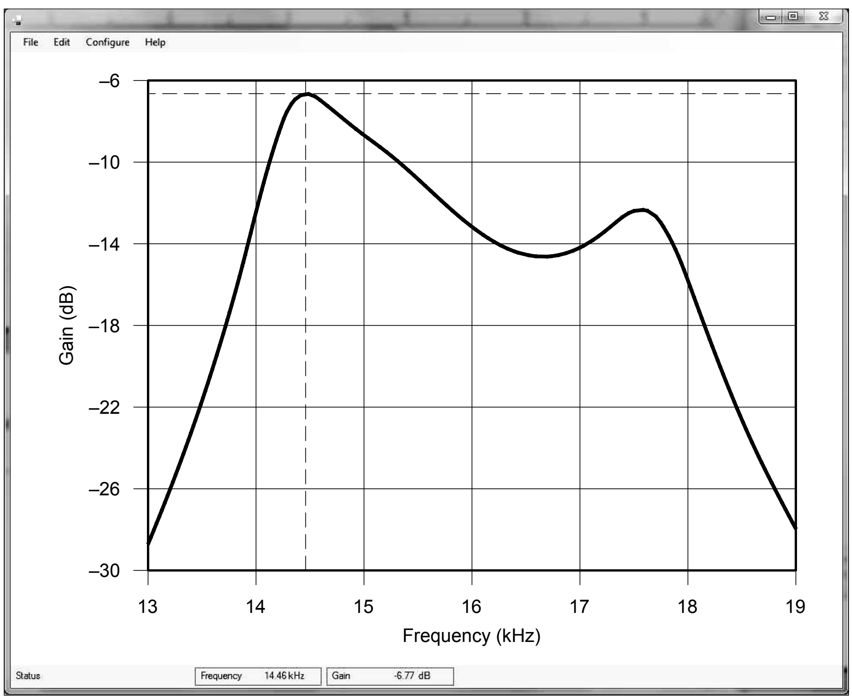


Figure 10 — The cascaded filter displays a loss of about 3.5 dB.



QX1607-Appel11

Figure 11 — A plot of the cumulative response can help verify the filter response as the signal progresses through the resonator stages.

Figure 13 shows the measured response of the filter employed to establish the channel bandwidth prior to sampling for the DSP. The filter does exhibit the flat passband expected with the staggered resonator design. The 3 dB bandwidth is just over 3.0 kHz. This filter did incorporate a buffer between the second and third stages to maintain a gain of approximately unity through the filter.

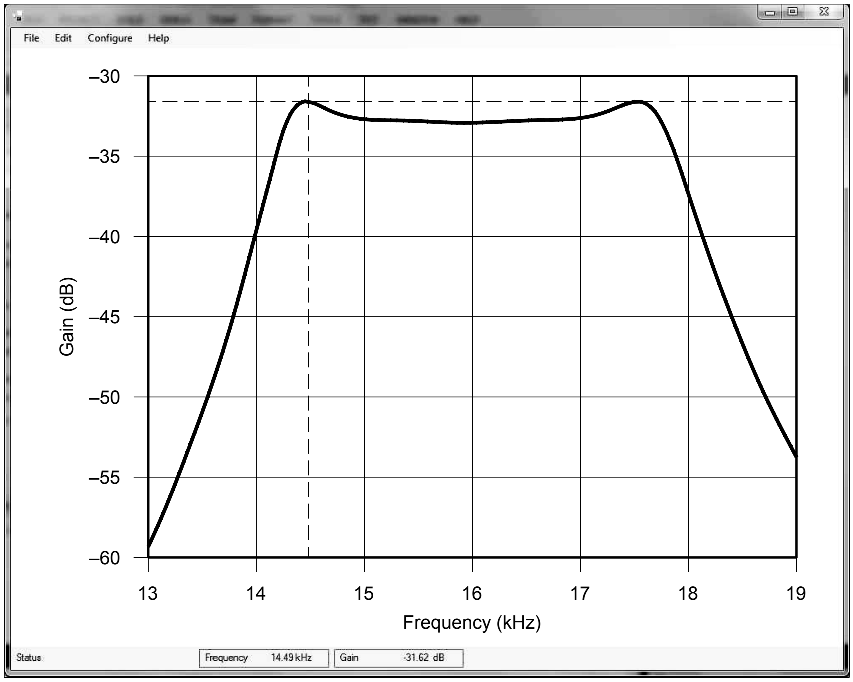
Adjusting Resonator Parameters

If a resonator is selected in the resonator list, we can change the resonator parameters by selecting a resonator, right-clicking and selecting the `Edit Resonator` item in the pop-up menu. This will bring up the same window as double-clicking on the resonator, but now the resonator parameters are highlighted. The resonator frequency or *Q* can be modified. Rather than modifying the *Q*, the 3 dB bandwidth of the resonator can be modified instead. Modifying these values will, of course, change the filter response. Why would we want to do that? In Figure 12 we see the filter response if the two resonators with the higher *Q* have their *Q* values increased to 20. We now have a dip in the center of the passband. By adjusting these *Q* values, the dip can be adjusted to compensate for the lossy passband response of other filters in the system. By adjusting the *Q* values of all resonators, a variety of responses can be obtained for compensating for other filters in the system. We already noted that, if a complementary pair of resonators do not have equal *Q* values, the result will be a slope in the response of the filter. This might be desired to compensate for a slope in the passband of other filters. Partial compensation of passband loss is relatively simple, more accurate compensation requires a series of adjustments, seeking an optimum response using trial and error.

The software program shown here was developed using *Visual Studio Express 2013*, and is written in *C#*. It was developed on a computer running *Windows 7 Professional*. The installation has been tested on another computer also running *Windows 7 Professional*.

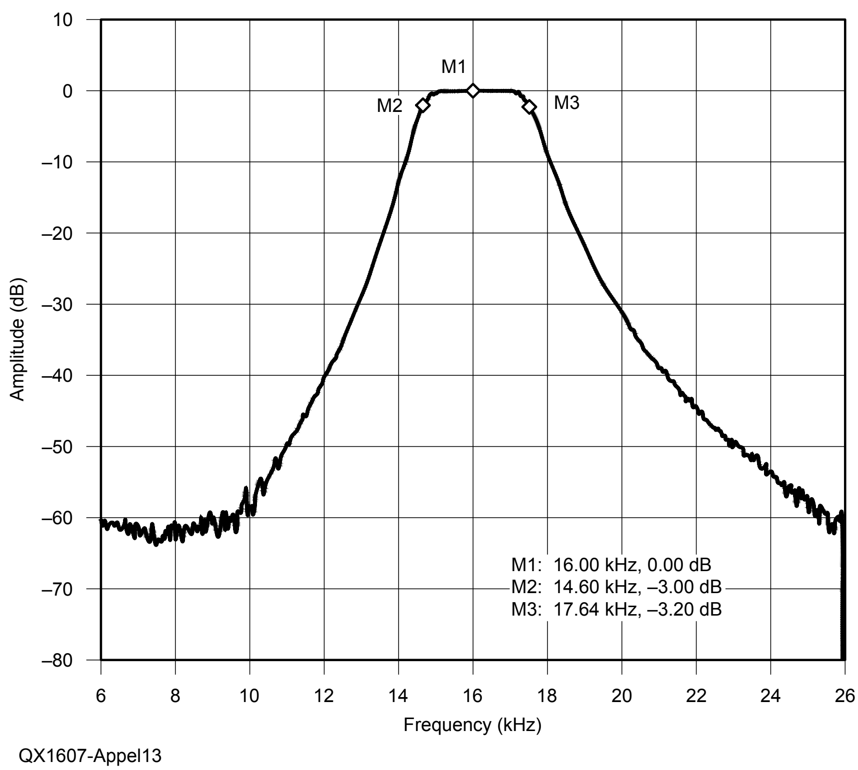
Conclusion

We have looked at an alternative to the standard staggered resonator filter implemented using cascaded operational amplifiers to simulate individual resonators. Our alternative filter displays the same response as the traditional implementation by cascading lower frequency operational amplifiers along with *LC* resonators. Advantages to this implementation are the ability to employ lower frequency operational amplifiers, and directly tuning the frequency



QX1607-Appel12

Figure 12 — Modification of the resonator parameters can compensate for other filters in the system.



QX1607-Appel13

Figure 13 — The completed filter displays the flat passband of an ideal bandpass filter.

and Q of each of the resonators. We've presented an application that will perform the required calculations, and display the filter response. The desired filter was fabricated, and the frequency response satisfied our requirement of filtering the input signal prior to applying that signal to the digital processing circuitry.

Gary Appel, WAØTFB, has been involved in the design of radio frequency equipment for over 30 years, most recently as an RF design consultant in the Silicon Valley. Gary has been fascinated with radios since his first crystal set, and was first licensed as WNØTFB in November of 1967 at the age of 14. He is a member of the ARRL and holds a BSEE degree from Washington University in St Louis. Gary has been retired since 2008 and enjoys the opportunities that his retirement has provided for working on homebrew projects and pursuing other technical areas of interest.

Notes

[1] *Reference Data for Radio Engineers, Fifth Edition*, Howard W. Sams & Co, Inc, 1968, pp 8-24 through 8-28.

[2] Equations for calculating the resonator parameters for the staggered resonator filter are also presented in: Arthur B. Williams, *Electronic Filter Design Book*, McGraw-Hill Book Company, 1981, pp 5 – 39, and in: Jack Porter, "Stagger-Tuned Bandpass Active Filters," *RF Design*, Mar 1988, pp 39 – 45.

[3] If using *FilterPro* to calculate filter parameters, the filter specification must be based on the geometric center frequency, not the arithmetic center frequency.

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Crystal Test Oscillators

These oscillators can test a wide range of fundamental and overtone crystal frequencies, and can measure the crystal activity.

Many experimenters have accumulated an assortment of oscillator crystals, and some will be unmarked or marked with a channel number or the output frequency of a transmitter rather than the crystal frequency. A test oscillator for determining the condition of a crystal and its actual frequency is therefore a worthwhile addition to the amateur radio workshop. My own collection of more than 2,000 crystals allows me to almost always find one very close to the desired frequency.

The UCO

The ideal universal crystal oscillator (UCO) should oscillate with any quartz crystal of any frequency and should give sufficient output to drive a frequency counter. It should also give an indication of the crystal's inclination to oscillate — what in the old days was called "activity" of a crystal.

An exhaustive search for such an ideal UCO has evolved into the circuit shown in Figure l. It will oscillate with any crystal from below 25 kHz to above the 25 MHz upper frequency limit of fundamental-mode crystals. In essence, the circuit is an FET Pierce oscillator with switchable capacitance between drain and ground. Strength of oscillation (activity), which corresponds to rectified drain voltage, is indicated on the meter. The 1N34 diode across the meter prevents over-deflection and gives a somewhat logarithmic response. Output for a frequency counter is also taken from the FET drain. Oscillation frequency will be close to the parallel resonant combination of the crystal in combination with a shunt capacitance of about 50 pF. This frequency is typically less than 200 parts per million above the series resonant frequency.

The UCO will oscillate with overtone crystals, but the frequency will be the crystal's fundamental, not the overtone. Although the overtone frequency does not bear an exact numerical relationship to the fundamental, it will always be very close, within 0.1%, to an odd integer (3, 5, 7, ...) multiple of the crystal's fundamental.

When investigating unknown crystals, all positions of switch S1 should be checked. This is because of different crystal modes. I have a 100 kHz crystal that oscillates at 100 kHz in the LF position of Sl, as it should, but in the MF position it oscillates at 500 kHz, and in the HF position at 3.639 MHz! Of course, the MF position is the fifth overtone. However the HF position is not an overtone but some completely different mode of vibration.

The UCO, along with a frequency counter, can also be used as an inductance meter where an unknown inductance replaces the crystal. Inductance will be inversely proportional to the square of the frequency. Table 1 gives the frequencies for different inductance values that were measured with my UCO. The Table also gives the calculated capacitance that would resonate with those inductance values at the measured frequency. I was astonished at how consistent the capacitance was over an inductance range of more than ten million to one.

Because of the consistency in capacitance, it is possible to derive simple formulas for approximate inductance *vs*. frequency. In the VLF and LF ranges,

$$L = \frac{500}{F^2}$$

where L is in H and F is in kHz.

For the MF and HF ranges,

$$L = \frac{900}{F^2}$$

where L is in µH and F is in MHz.

Of course, these will be ballpark values. If you want precision you should use an impedance bridge operating at 1000 Hz. Since the UCO works at a much higher frequency, inductance will be affected by distributed capacitance of the coil. Remember that apparent inductance tends toward infinity at the self-resonant frequency of the inductor. Even at half the self-resonant frequency the apparent inductance is one-third larger than the low frequency value. If you are not interested in inductance, the 0.01 µF blocking capacitor on the drain of Ql can be omitted. If you are interested only in crystals above 1 MHz, the range switch S1 can be omitted.

The UOO

Above about 20 MHz crystals are always overtone types. A Universal Overtone

Table 1
Frequencies for different inductance values measured by the UCO.

Range	*Inductance*	*Frequency*	*Capacitance*
VLF	2.85 H	14.1 kHz	44.7 pF
LF	116 mH	63 kHz	55 pF
MF	8.6 mH	317 kHz	29.5 pF
HF	10 µH	10.1MHz	24.8 pF
HF	0.25 µH	60 MHz	28.1pF

Oscillator (UOO) must be tunable to select the correct overtone. The circuit shown in Figure 2 does this well. It will tune in any overtone between 18 and 160 MHz. Overtones other than the intended one are often just as usable. By using unintended overtones you can double or triple the number of frequencies available from your crystal collection.

For example, I have a fifth overtone crystal at a fundamental frequency of 14.09 MHz that will oscillate at 42.3 MHz on its third overtone,70.5 MHz on its fifth (the marked frequency), 98.7 MHz on its seventh,126.9 MHz on its ninth, and 155.1 MHz on its eleventh overtone. I get six frequencies from one crystal! I even have one crystal that can produce a nineteenth overtone.

The UOO of Figure 2 tunes from 18 to 160 MHz in 3 overlapping ranges, 18 to 67 MHz, 31 to 97 MHz, and 75 to 160 MHz. It shares the 100 µA activity meter and power supply with the UCO circuit. A center tapped RF transformer Tl is used to neutralize the crystal capacitance, sometimes called the "holder" capacitance. The 5 pF neutralizing capacitor could be made variable but the

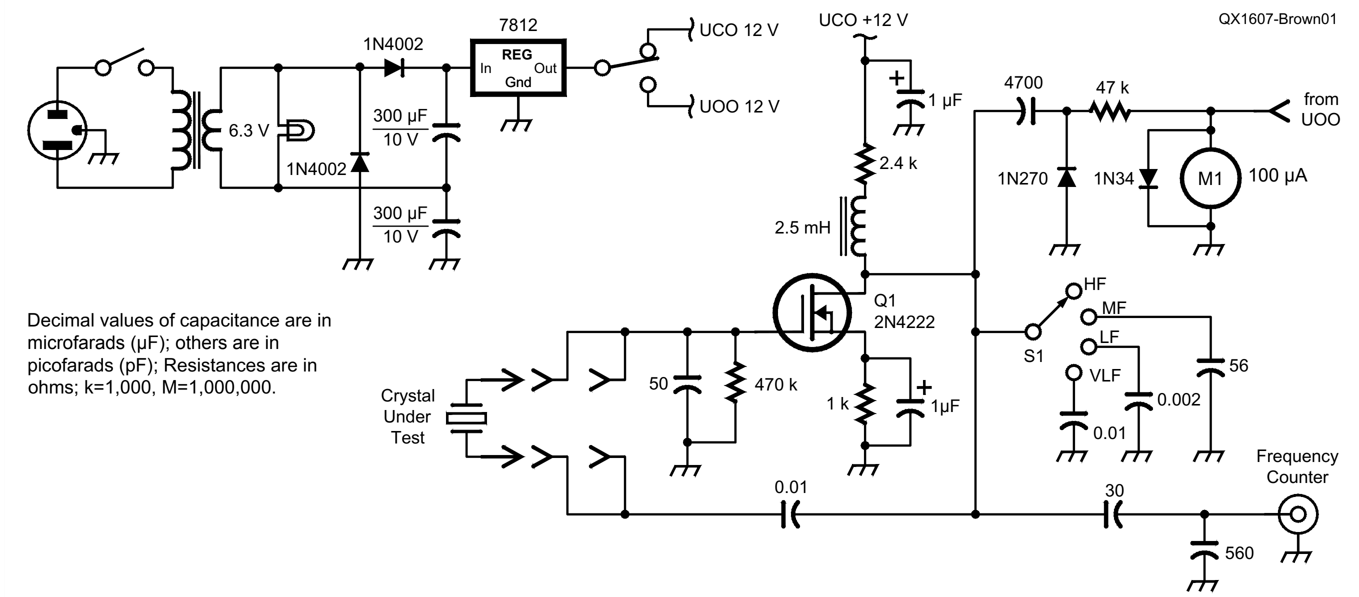

Figure 1 — The UCO, and its 12 V dc power supply, covers four frequency ranges selected by switch S1.

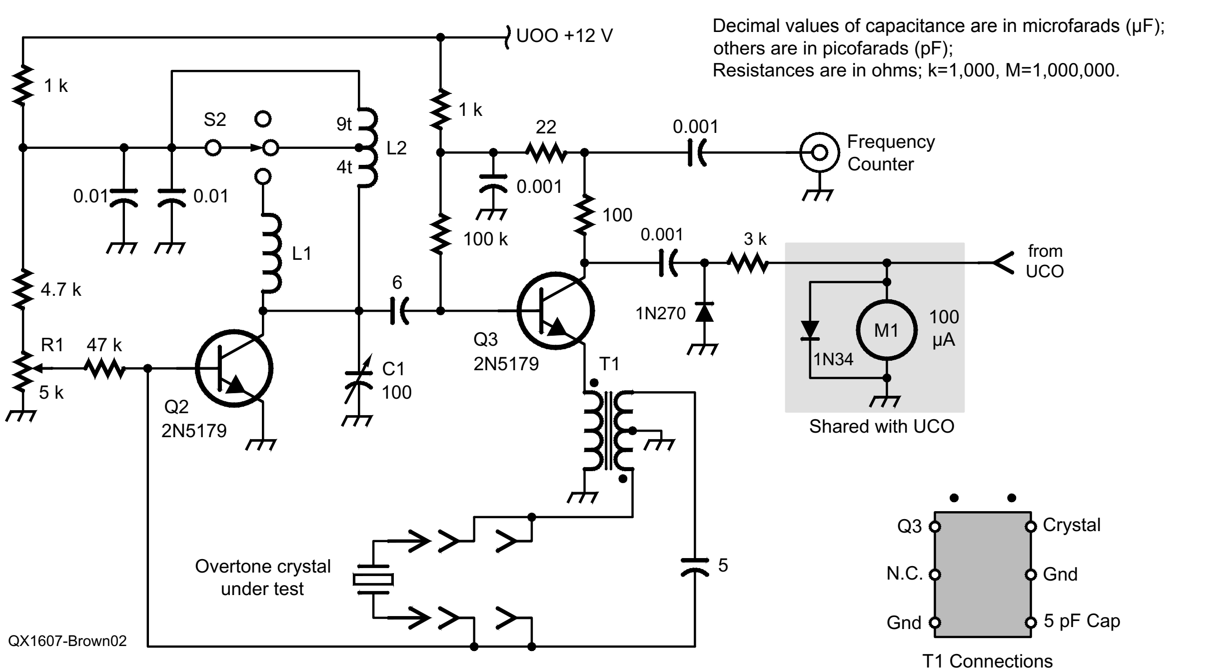

Figure 2 — The UOO shares the same meter and power supply with the UCO. RF transformer T1 is a Mini-Circuits T4-1H-X65. L1 is 1.5 inches of #22 AWG running between stator of C1 and S2. L2 is 13 turns of #18 AWG, 0.4 inch inner diameter,0.9 inches long, tapped 4 turns from the C1 end.

fixed value shown has proven to work well. Tl has a turns ratio of 1:1 from primary to each half of the secondary. Be sure to observe the phasing.

A gain control R1 is included because if the gain is set too high the UOO will free-run with some crystals. That is, it will self-excite. You can always recognize this condition because the frequency counter will not be consistent from one count to the next. When crystal-controlled, the frequency counter will be stable, plus or minus one count.

Construction

Construction is not critical, and each builder will have a personal preference. I built both oscillators and power supply on a 7.6 by 4.2 inch aluminum panel that fits on a plastic box. The power supply, the UCO, and the UOO are each built on separate rectangles of tin cut from a tin can that is tin plated on both sides.

Q2 and Q3 are on opposite sides of the tin, which forms a shield between them. Each plate is mounted upright on the panel and I used old fashioned point-to-point wiring. It is important to keep RF leads short, especially on the UOO, which must work up to 160 MHz. RF transformer Tl is a Mini-Circuits T4-1H-X65. Figure 2 shows the connections for proper phasing.

Fred Brown, W6HPH, has held his call sign since 1949. He earned a BS in Electronics Engineering from Cal Poly and an MSEE from the University of Illinois. He has worked as an engineer and has taught electronics in college. He has authored more than 100 technical articles in amateur and professional journals.

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Splicing Sections of Aluminum WR90 Waveguide

Parts of aluminum waveguide components can be spiced together and reinforced to make needed waveguide assemblies, WA4WDL shows how.

It seems I never have the correct sections of X-band waveguide on hand. Digging in the junk box, I can usually find some pieces of WR90 waveguide, which could be modified to work for 10 GHz projects. However, more often than not, the pieces are aluminum rather than brass or copper. I recently needed two mounts for 1N23 style diodes — one for making a boomerang tester and one for a YIG oscillator assembly. I found a four-port magic-tee with fatigue cracks in the junction. The two arms ending with diode detector mounts were salvageable and there were also two useable short straight sections, each having an attached flange.

Figure 1 — Using gusset plates to reinforce a waveguide splice.

Salvaging Needed Parts

It did not take long to separate the parts from the magic-tee using a hack saw. The cut ends were trued up using a disc sander, hand file, and hand surface sanding. Next came the hard part, joining the pieces together. I do not have a TIG or MIG welder and have not had good experience with aluminum solder. I could, and did, butt-join the parts together using J-B Weld™ epoxy. The result looks good, but is structurally weak. The joints can withstand reasonably strong steady forces, but a sharp whack will cause the epoxy to fracture and fail. The joint needs to be reinforced.

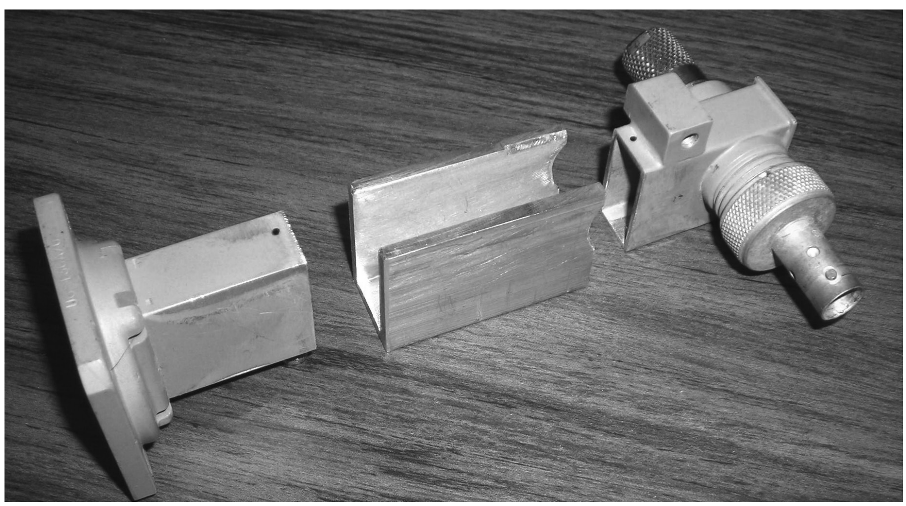

Figure 2 — Aluminum channel can reinforce a waveguide splice.

Reinforcing the Joints

I use two methods for reinforcing the butt joints. For each method, I butt-join the parts together and leave the assembly lightly clamped in a small bench vise overnight to cure. By so doing, I am more easily able to prevent any epoxy from entering the inside walls of the waveguide through the seams, and to hold everything together during the final assembly. Until it cures, the epoxy can act as a thick grease allowing parts to move out of position. I file and sand all of the parts before applying epoxy for the final assembly.

Using small aluminum gusset plates

The first reinforcing method uses small aluminum gusset plates, see Figure 1. Two plates, one on each of two adjacent waveguide faces are all that is needed for each joint. The plates are easily fabricated by hand from pieces of scrap 1/16 inch thick sheet aluminum. The number and placement of the screws is not critical. Drill 2-56 tap holes through the plates. Then lightly clamp the plates on the waveguide. Use the plates as guides to drill matching tap-drill holes in the waveguide. Next release the clamps and enlarge holes in the gusset plates to form clearance holes. Tap the holes in the waveguide for 2-56 screws. The gusset plates are screwed and fastened in place with epoxy. Add thin washers under the screw heads as needed to allow the screws to be threaded as deeply as possible into the waveguide walls without protruding into the waveguide. This method works fine but takes a lot of time and requires finding screws of the proper length.

Using a U-shaped channel

The second method involves the use of U-shaped aluminum channels. The outer cross sectional dimensions of the most common examples of WR90 waveguide are 1.0 by 0.5 inches. A search for aluminum channel with matching inner dimensions was unsuccessful until I ran across "6063-T52 Aluminum Channel, Architectural" from **www.onlinemetals.com**. It comes in two channel heights, 3/4 inch and 1-1/2 inch. The outer channel width is 3/4 inch. The channel wall thickness is 1/8 inch so the channel depths are 1/8 inch less than the heights and the internal channel width is 1/2 inch. The architectural channel has sharp inner corners which allow the waveguide pieces to sit deep within the channel. I chose to use the deeper version, knowing I could trim or shape the height as needed, and the measured channel width for the deeper version turns out to be a bit wider: ~0.505 inches versus ~0.496 inches. Figure 2 shows the pieces of waveguide and U channel section prior to assembly.

Figure 3 shows completed diode mounts using both methods of splicing prior to being painted. Both methods are successful and yield strong joints. Figure 4 shows the assembled pieces in use.

Summary

I no longer turn away from odd-shaped pieces of aluminum WR90 waveguide at ham fests and flea markets. Instead, I purchase them as potential stock for making assemblies. Now, I need to find someone who sells the same size channel made of brass, which could be used for brass or copper waveguide.

Photos courtesy of the author.

John M. Franke, WA4WDL, was first licensed as a Novice in the early sixties. He currently holds an Amateur Extra class license. John also holds a First Class (now General) Radiotelephone Operator Certificate with Ship's Radar Endorsement. The licenses enabled him to work through college as a transmitter engineer at two AM broadcast stations. His degrees include AAS, BSEE, and MS in physics. John retired from NASA in 2005 after more than 31 years of service. His duties included the design, construction and operation of optical and electronic instrumentation supporting wind tunnel research and supporting the licensing and commercialization of NASA technology. He concurrently served as a radar operator onboard US Navy E-2B Hawkeye aircraft, and aircrew on CH-53 helicopters in the Naval Reserves. He was a member of the Association of Old Crows for over 25 years. Time permitting, he has served as a docent at the Virginia Air and Space Center. His interests include electronic warfare, microwaves, VLF, and precision timing. John is the inventor or co-inventor on three US Patents and has authored or co-authored 130 professional and amateur radio articles.

Notes

[1]6063-T52 Architectural aluminum channel, **onlinemetals.com**.

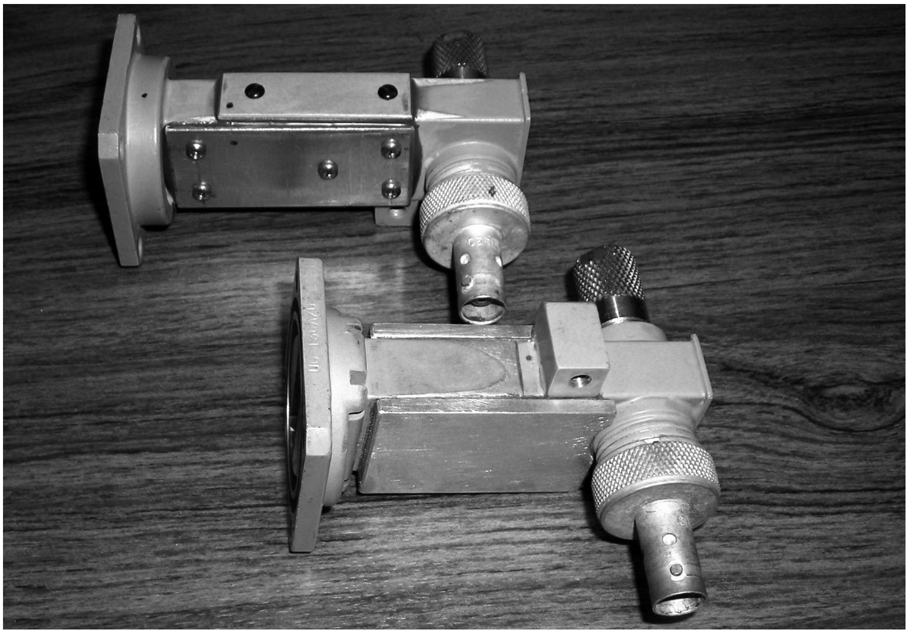

Figure 3 —Completed reinforced a waveguide splices.

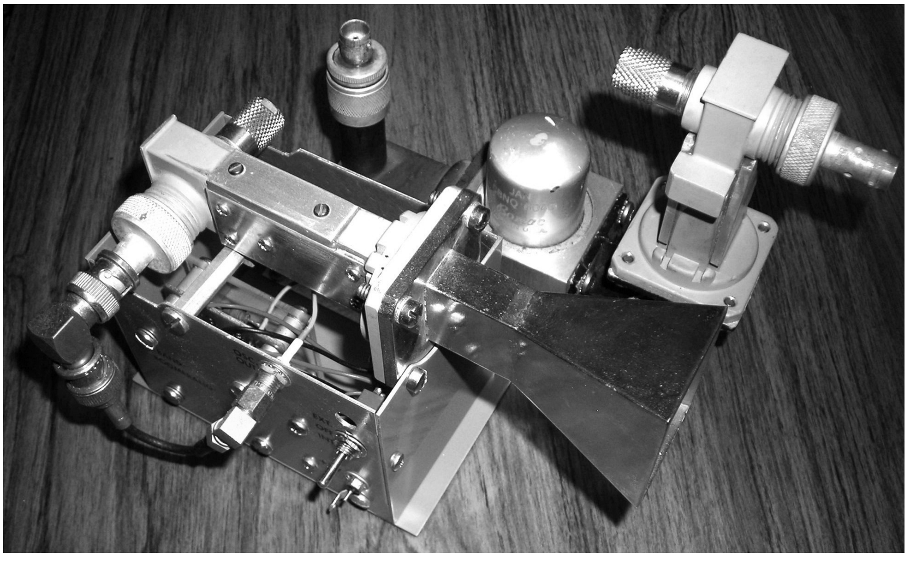

Figure 4 — Spiced aluminum waveguide sections used in assemblies.

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3-D Printed Horn Antennas

3-D printing allows the specifications in software of arbitrary shapes and complex curves for horn antennas, then printing and applying a conductive coating. This allows expermentation with almost any shape — and raises interesting intellectual property issues.

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3-D printing is a set of tools and techniques that allows the creation of custom objects with a 3-D printer. The 3-D printer technology described in this paper is heated filament deposition printing. This is analogous to having a computerized numerically controlled (CNC) hot glue gun. Plastic filament is extruded through a nozzle in a heated head. Our printer uses two stepper motors to move the heated head for print length and width, one stepper motor to raise and lower the heated bed for print height, and a stepper motor that moves the filament into the heated head. These motors provide three-dimensional control of the print space as well as control over the rate and direction of filament feed. The question considered here was whether new horn antenna designs could be successfully printed that would allow experimentation with complex tapers.

3-D Modeling and Printing

Melted filament forms the layers of a 3-D printed object. After each layer is printed, the bed is lowered, and the next layer is printed on top of the previous one. Each layer adheres to the previous layer due to heat fusion. In general, the z-axis (up and down) is perpendicular to the deposited layers, which are on the xy-plane. The usable thickness in the z-direction of the melted filament determines the resolution along the z-axis. If the next layer is started too high above the previous layer, there will not be sufficient adhesion. If the next layer is started too low in relation to the previous layer, then the previous layer will be damaged or disturbed by the heated head as it attempts to cram new melted filament on top of older already-cooled filament.

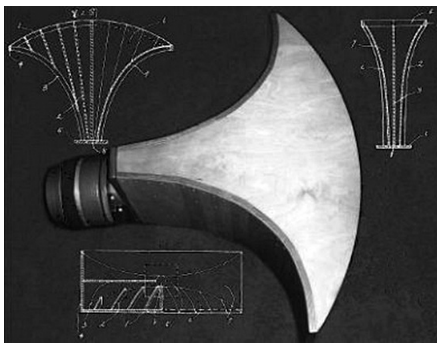

Figure 1 — An example of an audio horn with tapered sides.

3-D models of objects are created in software by either scanning or specification. The 3-D model is then sliced into layers that correspond to the thickness of the layer of heated filament that the 3-D printer produces. The process of taking a concept or drawing all the way from a sketch to a set of instructions that the printer will understand is generally referred to as 3-D modeling.

There are two main types of plastic used in 3-D printers, PLA (polylactic acid) and ABS (acrylonitrile butadiene styrene). The plastic filament is generally purchased on reels, and is on the order of 2 mm in diameter. Matching the filament size to the size of the nozzle in the heated head is important. Matching the material to the purpose and characteristics of the type of object printed is also important.

Similar to machining, sewing, software programming, and other crafts, there is an ensemble of skills involved in successfully producing a quality 3-D print. Troubleshooting, materials selection, experimentally determining the right settings for any particular job, cleaning, and maintenance are all very much part of the 3-D printing process.

The main reward of 3-D printing is the ability to make shapes that are difficult to manufacture otherwise. For the case of microwave horn antennas, experimenting with non-rectangular horns means more difficult fabrication techniques. The ease of cutting shapes out of sheet metal and bending them into rectangular horns is undeniable, especially compared to making horns with curved sides.

Instead of the straight sides of a rectangular horn, the sides of a microwave horn antenna can be curved into a taper. Some examples are elliptically or exponentially tapered sides. An example of an audio horn with tapered sides can be seen in Figure 1. To control the taper, either a form or some sort of press would be required to create a specified curve. When the cost and hassle of making equipment to make equipment is substantially more time and trouble than an easy-to-make alternative, then the easier alternative will be chosen, even if the performance is compromised.

3-D printing allows the specification in software of arbitrary shapes, like elliptically tapered sides, or other complex curves. Printing a horn or other part, then applying a conductive coating, allows experimentation with almost any shape. This opens up tremendous possibilities for microwave enthusiasts to try all sorts of ideas. The question under consideration for this paper was whether new horn designs could be successfully printed that would allow experimentation with complex tapers.

Printing a Rectangular Horn Antenna

A rectangular horn design by Kodera2t was obtained from *Thingiverse*[1] and printed on my Ultimaker2 3-D printer. See Figure 2 for an image of the horn as printed by Kodera2t in Japan, and Figure 3 for initial results. One purpose of printing this design was to confirm the ability to print, metalize,

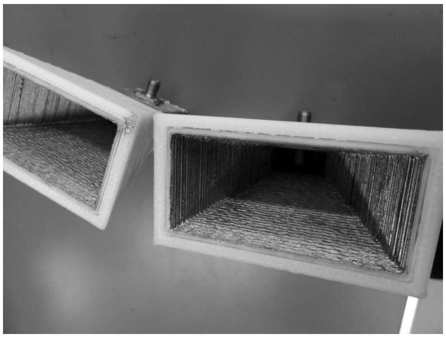

Figure 2 — Image of the horn as printed by Kodera2t in Japan.

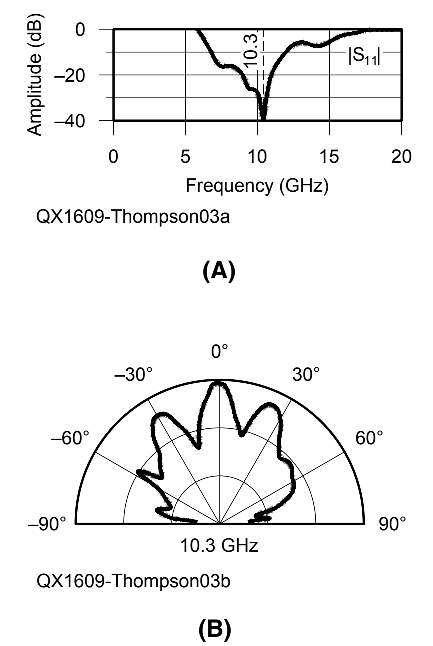

Figure 3 — (A) Initial S_{11} performance of the horn of Figure 2. (B) Initial radiation pattern of the horn of Figure 2.

and characterize a "known good" horn antenna. Any problems inherent in the printing, metallization, connectorization, and measurement stages could be addressed with some confidence that they weren't due to the horn design itself. Horn antennas are very popular in the microwave band. Rectangular horns provide high gain, low SWR, and relatively wide bandwidth, and they are not difficult to make.

What About Intellectual Property?

Ability to share 3-D models makes experimentation with objects much easier. Kodera2t speaks more English than I speak Japanese, but if we weren't able to simply share the 3-D model, collaboration would be much more difficult. Freely sharing 2- and 3-D models on sites such as *Thingiverse*[2] directly supports the open source movement, where work output is given away to the public domain, and others are encouraged to use, modify, and republish the work for their own applications and needs. This is not mandatory. Many 3-D models are unpublished or controlled, for all sorts of reasons. This brings up the question of where 3-D printed objects fall in the universe of patents, trademarks, and copyrights. A good starting point for understanding these issues is the white paper by Michael Weinberg.[3]

Multiple patents cover a wide variety of horn antennas. Almost any creative image of a horn antenna has a copyright. A horn antenna as part of a logo of a specific manufacturer or seller could be a trademarked image. All parts of the process of 3-D printing largely fall into the existing framework of intellectual property, but there are new challenges and novel legal questions that will have to be addressed in the coming years as 3-D printing becomes more and more widely available. The ability to make 3-D scans of objects, and then recreate them with a high degree of precision, means that the market for some useful manufactured objects might decrease whenever a 3-D model becomes available. This is similar to the challenges the music industry believes that they are facing with people being able to easily copy and share music files. Another specific example is war-gaming models such as ones from *Games Workshop*.[4] These small gaming models are expensive and available only from one manufacturer. If one could scan a completed model and then print out an entire army, it would save hundreds or thousands of dollars.

While 3-D printers are not yet, and may never be, a good solution for mass manufacture. They are still very expensive. High-resolution printers can cost thousands of dollars. The filament is expensive at about US$40 per kilogram. Many prints

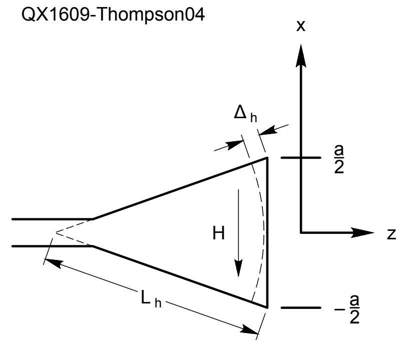

Figure 4 — Phase error in a pyramidal horn antenna.

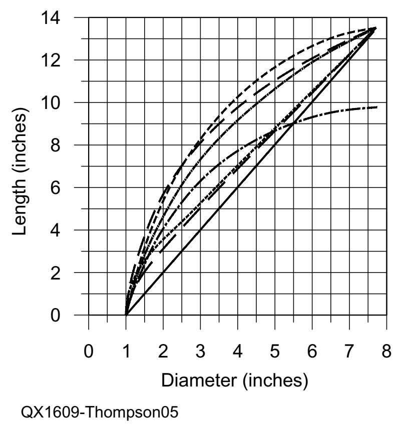

Figure 5 — Different tapers investigated by the audio horn community.

fail for a variety of reasons. Home 3-D printing is not in any way as cheap and easy as making a copy of a CD or DVD or MP3. Injection molding is still superior in terms of resolution. For mass manufacture, even a very expensive mold for an injection machine is the most profitable way to create objects for sale.

Elliptical Taper Horn Design

The next step is to attempt to create a 3-D software model of a more complex horn. There are many different of tapers to choose from. The exponential taper horn shape minimizes reflected power.[5] This means that this taper is the most efficient way to get the signal from the wire or waveguide into the air. Tapers also affect phase error. Spherical waves leaving the antenna encounter straight sides causing reflections at slightly different delay times (Figure 4), introducing phase error. Tapers that conform with the spherical wave front introduce less phase error. The contours in Figure 5 show comparisons of many different tapers, from right to left at the

4 inch length axis, Conical, Oblate Spheroid, Hughes (Peavey), Spherical, Exponential, Tractrix, and Hyperbolic.

Horns with an elliptical taper are described in the literature as having less phase error than an equivalent straight-sided horn. Using *OpenSCAD*[6] I began to describe the shapes I wanted. See Figure 6 for an example of what the 3-D modeling workspace looks like. For the elliptical horn, I created the horn using four solid ellipses having surfaces that would be the inside, or throat, of the horn. I then used the difference function in *OpenSCAD* to subtract a slightly smaller ellipse to turn the solid ellipses into a set of curved walls. This process is somewhat similar to using one layer of an onion, instead of the entire onion.

Simply scaling down the ellipse that I wanted to subtract meant that the wall thickness was not constant. I had to scale the major and minor axes differently in order to have constant wall thickness. After that I attached the SMA connector. If the wall where the connector is attached isn't controlled in the model, it can end up being too thin to attach the connector. Screws might protrude into the throat of the horn, or the antenna might end up being too short or too long. Part of good 3-D design, especially when the model is parameterized, is controlling the repercussions of changing the parameters. Code for all of the horns can be found online.[7]

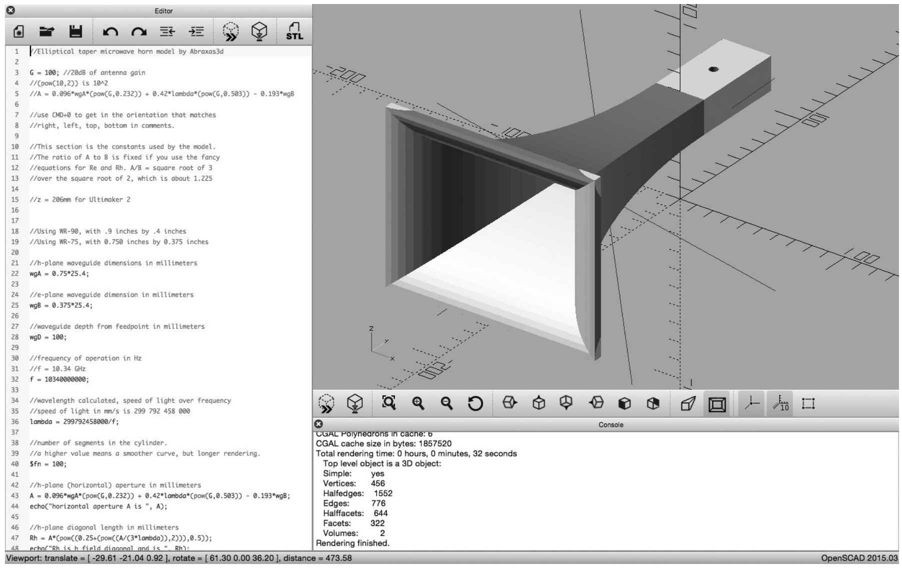

Figure 6 — An example of the *OpenSCAD* scripting language 3-D modeling workspace.

Printing the Elliptical Horn

I printed each elliptical horn in two pieces because the dimensions of the elliptical horn were greater than the available print dimensions on the Ultimaker2. In Figure 6, you can see the transition from the waveguide/SMA portion of the antenna, to the tapered part of the antenna leading to the aperture. I separated the model in *OpenSCAD* where the gray scale density changes. I wanted to print the horn with the aperture facing up in order to make the inner surface as smooth as possible. I decided not to use support material for the outside surface, which would overhang to near horizontal at the aperture.

Support material is a lacework of 3-D printed filament that allows the printing of overhangs. Wherever the solid object has a horizontal part projecting into the air, support material is "grown" up from the platform so that the overhanging structure somewhere up above the platform has something to sit upon. It's somewhat like scaffolding when building a construction project. The decision to skip support material was somewhat risky because there is an overhang at the top of the print.

Figure 7 shows the print at two points in time. The pencil shows scale. The print is at about the halfway point on the left and is nearly complete on the right. Note some of the filament is loose on the overhang on the right-hand image. Support material prevents this. However, printing the required amount of support material for a tall print like this is risky as well. The lacework would have to print perfectly all the way up to meet the relatively small amount of overhang. This print, without support material, took approximately 40 hours. The driver software estimated 55 hours if support material had been included.

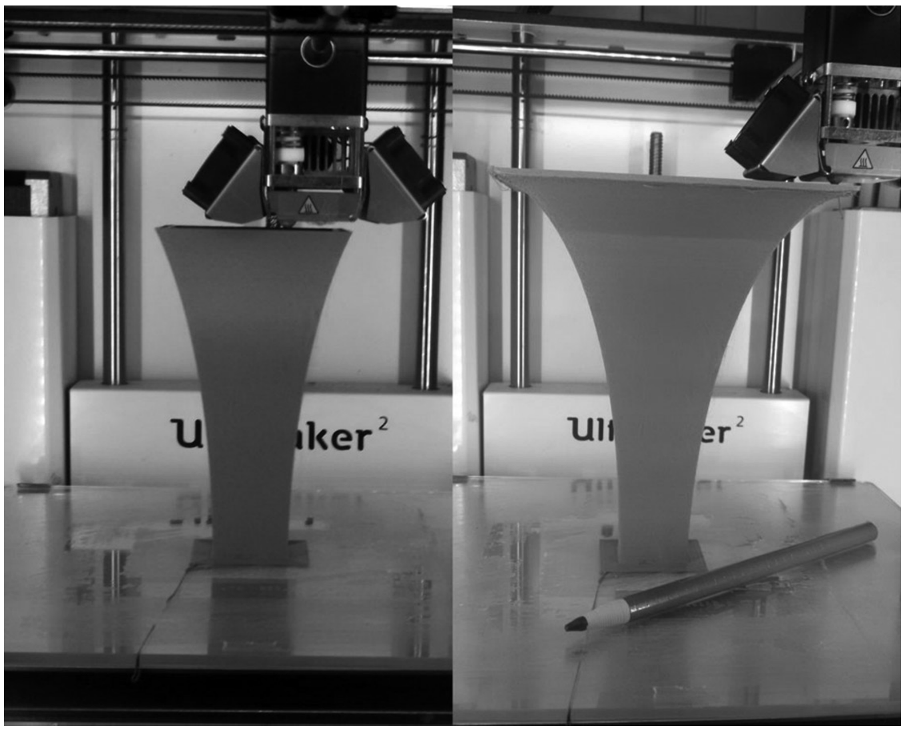

Figure 7 — The print at two points in time. The pencil shows scale.

Two horns were successfully printed out of two attempts. The four parts (Figure 8) were metalized with conductive spray paint (MG Chemicals 843-340G Super Shield Silver Coated Copper Conductive Coating, 5-Ounce Aerosol, at a US$40 cost). The toy dinosaur is for scale. The waveguide/SMA sections were then super-glued to the tapered sections.

Figure 8 — The four parts were metalized with conductive spray paint, the toy dinosaur shows scale.

Initial Testing of Rectangular and Elliptical Horns

I obtained SMA connectors and advice from RF Parts, San Marcos, CA. Kerry Banke, N6IZW, assisted with installing SMA connectors for the 3-D-printed rectangular horns and the 3-D printed elliptically tapered horns. We made initial measurements in his lab of both the rectangular and elliptically tapered horns. The first rectangular horn tested was one that didn't print well. I selected one that had voids and other problems with the print in order to test whether the conductive paint would work as a reflective surface at 10 GHz. While confidence was high, and measured resistance was very low (1 Ω across the horn). I didn't want to use one of the nicer prints if conductive paint would fail to properly metalize the printed shapes.

The backup plan was to apply copper foil to the surfaces of the horns, which would be much more painstaking to apply. Copper foil would definitely work according to advice from Professor Nuno Borges Carvalho, of the Instituto de Telecomunicacoes, Universidade de Portugal. He presented extensively about 2-D printed circuits and 3-D printed horn antennas at the 3-D printing workshop at IEEE Radio Wireless Week conference held in January 2015. He and his graduate students used copper foil to metalize their 3-D printed 10 GHz horn antennas. They are advocating for a two-part 3-D printer that would metalize as part of the printing process.

I had neglected to paint the rear surface of the waveguide-shaped section of the rectangular antenna. A whole lot of signal was blasting out of both the front and back of the antenna. This gave us confidence that the conductive paint actually worked as a reflective surface. We added some foil to the back of the horn, and achieved 30 dB front-to-back ratio. We removed the foil, and painted the rear surface with conductive paint and retested. The painted surface was now "closed" and reflected RF to the same level as the copper foil.

This particular rectangular antenna had several large voids in the layering of the print. Sniffing around with a probe revealed RF leaking through the void. Since we could literally see through this void to the other side, this result was not surprising.

The elliptical horns were painted with conductive paint only on the inside, with the SMA connector hole painted as well. This turned out to not be enough for them to work. I then coated the outside of the horns as well.

Kerry, N6IZW, attached one horn with an SMA connector to a network analyzer. It had a return loss of 15 dB at 10 GHz with large dips at 7.5 GHz and 9 GHz. He removed the SMA connector and found that the conductive coating had not been sufficiently applied. Additional conductive material was added below the SMA, and the dips were substantially reduced. Gain was measured in comparison to a reference antenna and found to be at least 12 dBi.

The second elliptical horn had 0 dBi gain and was returned for more conductive coating.

The waveguide of the horn is designed to be WR-75, with inner dimensions of 0.750 inches by 0.375 inches. The dimensions of the waveguide affected the size of the horn, with WR-90 making the horn large enough in some dimensions to not fit as desired on the print surface. We chose WR-75 because it was the smallest waveguide that would work at 10 GHz and also allow the horn to print completely within desired printer dimensions. WR-75 works from 10 to 15 GHz. The next size up, WR-90, works from 8.2 to 12.4 GHz. WR-90 puts 10 GHz much more comfortably in the middle of the range at the cost of making the print slightly larger. Some of the results that we were seeing could be due to choosing WR-75 over WR-90 for the prototype horns.

Antenna Range Party Results

An elliptical horn antenna was tested at the San Diego Microwave Group range party on 27 July 2015. Fourteen operators attended with gear covering 10 – 47 GHz. The range tests include measuring output power and minimum discernible signal.

Figure 9 shows Kerry, N6IZW, holding up the elliptical taper 3-D printed horn in operation at the range test. This horn was directional and had at least 12 dBi of gain. The second elliptical taper horn was re-painted and the SMA connecter re-seated. In late August 2015, this horn underwent further tests. Performance was not in line with the first horn, so the tapered part was separated from the waveguide and tested with a known good SMA to WR-75 transition held firmly in place. The antenna worked well, achieving a gain of about 20 dBi. The transition from taper to waveguide was determined to be problematic, even after rework. However, the antenna shape, surface texture and conductive coating all performed very well.

Figure 9 — Kerry, N6IZW, operates the elliptical taper 3-D printed horn at the test range.

Suggested Improvements

Several improvements were suggested based on experiences with handling, printing, and testing the 3-D printed 10 GHz horns. First, we should consider adding a radius to the corners in the model. The edges are all sharp. It would improve paint adherence to radius the corners without much cost in terms of gain or phase performance. This can be achieved with a relatively simple function in *OpenSCAD*. Second, we could print the model so that the horn is assembled around the SMA connector. The horn could be printed in pieces where the connector can be captured by the sides, instead of fitting through an SMA hole. With the current design, screws to hold the SMA connector are going into the surface of the 3-D print. There is a layer of solid PLA material for the outside wall of the print. However, the inside of these prints is a honeycomb. It's not solid plastic all the way through.

The decision as to the amount of material used for a print is made at print time. In general, the outer walls are a few layers thick, and the interior is a honeycomb of about 20% material and 80% air. Rectangular and hexagonal honeycomb are the most popular. Printing a large object such as this horn in solid plastic would take a very large amount of additional print time and filament. I chose 25% fill for the honeycomb for these prototype horns.

For an antenna with a connector, reinforcing the area where the holes for the SMA connectors go seems to be a necessary improvement. The area immediately around the SMA connector can be solid plastic all the way through without costing much additional printing time. This would improve the seating of the connector and reduce unreliability of this particular interface. However, we came to believe that printing horn antennas that directly connect to a waveguide with a flange would be superior to attempting to incorporate an SMA connector. This leads to the next suggested improvement.

We explored the idea of building in an RF choke flange into the design wherever a transition was required. This would improve the reliability of any interface, whether the transitions were due to having a multi-part print or when the horn was designed to attach to a waveguide.

Conclusion

We designed, printed, metalized, and tested rectangular and elliptical taper 3-D-printed 10 GHz horn antennas. 3-D printing technology can be used to create complex tapers for 10 GHz horn antenna experimentation. We believe that the reliability of the horns can be improved by making the improvements discussed in this paper.

Photos courtesy of the authors.

Michelle D. Thompson, W5NYV, enjoys thinking and doing. Not necessarily in that order! Book learning includes BSEET, BSCET, math minor, MSEE Information Theory. Actual doing includes engineering at Qualcomm Incorporated, engineering at Optimized Tomfoolery, Amateur Extra class license, AMSAT Phase 4 Ground lead, Organ Donor Pipe Organ lead, DEFCON, IEEE, Burning Man, and community symphony.

Kerry Banke, N6IZW, was first licensed in 1961. He retired as an Electrical Engineer from Qualcomm in San Diego where he worked on the development of cellular and satellite communications equipment. Kerry holds a BSEE degree. He has been hosting the San Diego Microwave Group since 1986. He is currently part of a team developing the new power supply system for the improved Amateur Radio equipment headed for the International Space Station.

Notes

[1]**www.thingiverse.com/thing:87574**.

[2]**www.thingiverse.com**.

[3]Michael Weinberg, "What's the Deal with Copyright and 3-D Printing?" **https://www.publicknowledge.org/files/What's%20the%20Deal%20with%20Copyright_%20Final%20version2.pdf**.

[4]**www.games-workshop.com**.

[5]U. A. Bakshi; A. V. Bakshi, K. A. Bakshi, *Antenna and Wave Propagation*, Technical Publications, 2009, pp. 6.1–6.3, ISBN 81-8431-278-4.

[6]**www.openscad.org**.

[7]**https://github.com/Abraxas3D**.

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A Receiving Array for 160 m Through 2200 m

N6LF presents study of an antenna with low back lobes and the ability to switch the pattern direction and shape from the shack in a simple structure with no phasing networks.

For the past ten years I've participated in the ARRL 600 m experimental license group, WD2XSH, and tried a variety of receiving antennas from phased verticals (*E*-probes) to BOG's (Beverage on the ground) to terminated loops. I've also used regular Beverages on 160 m but at 475 kHz a 1.5 λ Beverage would be ≈ 3000 ft long and at 137 kHz over 10,000 ft, not very practical for most of us.

With the imminent authorization of the 2200 m and 630 m bands I needed an LF-MF receiving antenna with good performance from 100 kHz through 2 MHz. What I wanted was an antenna with low side lobes off the back (azimuths 90° through 270°) and the ability to switch the pattern direction and shape from the shack. All this of course is in a simple structure with no phasing networks.

Comments on Terminated Loops

Resistively terminated loops have many names: flags, pennants, EWEs, and so on. These antennas are usually electrically small — loop perimeters smaller than 0.1 λ — where λ is a wavelength at the operating frequency. Given the long wavelengths this will be the case for any practical antenna at 630 m or 2200 m. Because of the small size the current amplitude will be almost the same along the wire. The small variation in current magnitude translates into an insensitivity to the shape of the loop. Round, square or triangular makes little difference. This encourages us to use shapes that fit the available space and supports. Changing the size (area) of the loops has little effect on the pattern, it mostly affects the amplitude of the received signal. The greater the area of the loop, the greater the signal voltage *V* amplitude at a given frequency. It's just Faraday's law,

$$V = n\frac{d\phi}{dt}$$

where ϕ is the total flux and n is the number of turns. As we go down in frequency, for the same physical size, the signal decreases.

An essential feature of terminated loops is the use of a resistive termination somewhere in the loop. The value of the terminating resistor is typically in the range of 200 – 1200 Ω , which is much greater than the self-impedance of a small loop without the termination. The result is a feed-point impedance dominated by the fixed termination resistance. The feed-point impedance changes little as the frequency and/or loop size are changed. Another effect

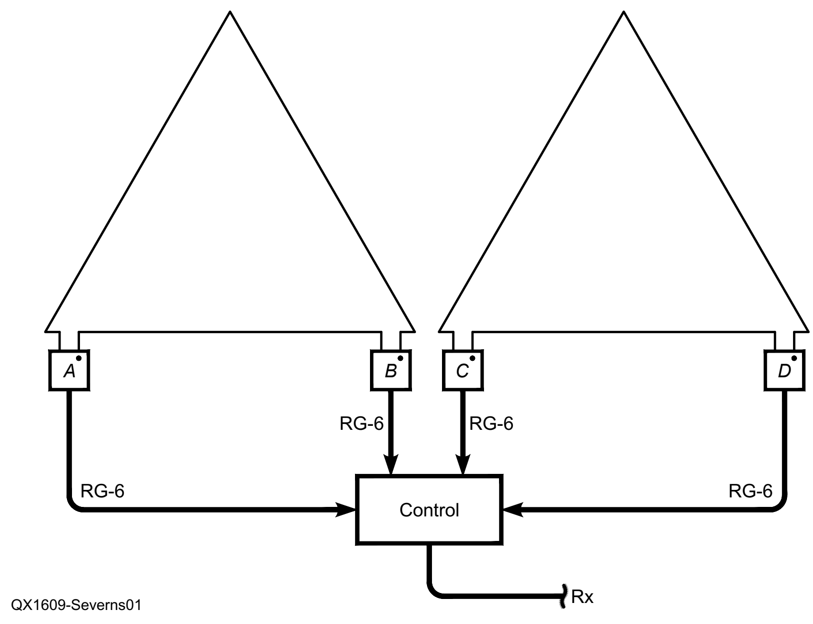

Figure 1 — *EZNEC* model for the receiving antenna.

of using a termination is to swamp out the mutual impedance due to coupling between loops. Changing the phase differences or the spacing between the loops has little effect on the feed-point impedances, which simplifies feed network design. This reduction in mutual coupling is exactly the same effect seen in phased arrays using short vertical elements (*E*-probes).

The properties of terminated loops lead me to think about combining them in an array. About that time the March 2015 issue of *QST* arrived with an article by Chris Kunze, DK6ED, on a his version of a double loop antenna.[1] This antenna is basically two triangular terminated loops in a line, fed 180° out of phase. What attracted my attention was the good pattern off the back of the antenna, sharp broadside nulls and the simplicity of the phasing scheme, which might allow the antenna work from 100 kHz to 2 MHz if it could be made large enough to have sufficient received signal on 2200 m but still be small enough to behave like a "small" loop on 160 m.

A bit of modeling with *EZNEC* was very encouraging so I built and tested an antenna.[2] This note describes that antenna in some detail. However, the reader should keep in mind this is just one example that happens to fit my particular location.

These antennas can be scaled up or down in size to suit a particular situation. The primary effect of scaling is to change the received signal strength. The directive patterns change very little.

The Antenna

The antenna is shown in Figure 1. I have two ≈80 ft poles, spaced 150 ft in my pasture from which I could suspend the antenna. Each loop is an equilateral triangle 73 ft on a side. The bottom wires are 8 ft above ground and the corners at the mid-point are 2 ft apart. At each end of each of the bottom wires (points *A*, *B*, *C* and *D*) there is a 1 kΩ to 75 Ω impedance transformer with a common-mode choke for isolation (Figure 2). Each choke is connected to a length of 75 Ω RG-6 leading back to the control box in the shack. The control box determines how the feed points are driven — which are terminated, which are driven and what the phase relationship will be between the two loops. The cables back to the control box can be of any length but all four cables must be the same electrical length! It's best if all four cables are cut to the same physical length from the same roll of cable.

The 100 kΩ resistor in Figure 2 is for

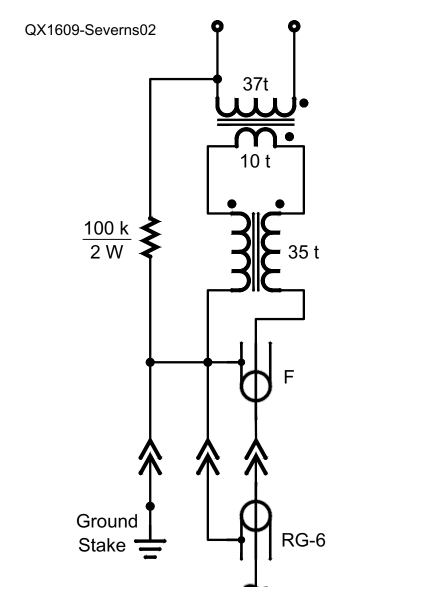

Figure 2 — Impedance transformer and common mode choke. RG-6 with F-connectors runs to the control box.

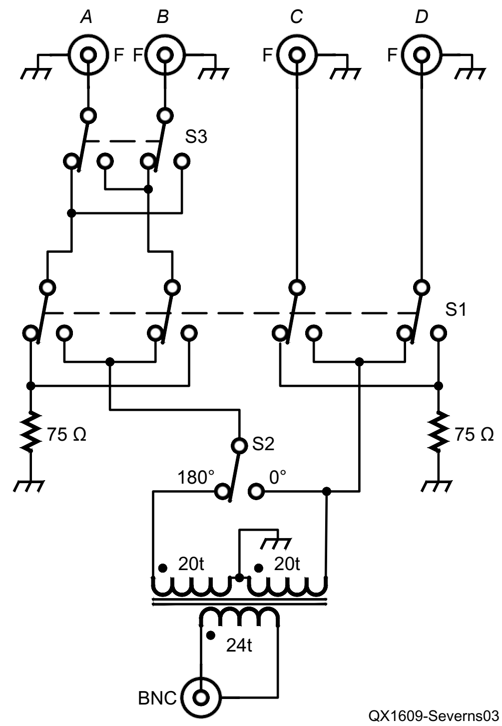

Figure 3 — Control unit schematic. F-connectors are used at *A*, *B*, *C* and *D* in the 75 Ω portion of the system, and a BNC connector is used at the 50 Ω connector to the receiver.

Table 1
Source and termination locations.

Configuration	*Left source*	*Right source*	*Left termination*	*Right termination*	*Relative phasing*
1	*B*	*D*	*A*	*C*	0
2	*A*	*C*	*B*	*D*	0
3	*B*	*D*	*A*	*C*	180°
4	*A*	*C*	*B*	*D*	180°
5	*A*	*D*	*B*	*C*	0
6	*B*	*C*	*A*	*D*	0
7	*A*	*D*	*B*	*C*	180°
8	*B*	*C*	*A*	*D*	180°

static discharge, these are large wire antennas that could accumulate a charge under some weather conditions. Construction details for the transformer-chokes and the control box are in the last section of this article. The control box contains only three switches and a phase-inversion transformer as shown in Figure 3.

The terminations are 75 Ω resistors placed in the control box. The 75 Ω is transformed to 1 kΩ at the antenna with the transformers at *A*, *B*, *C* and *D*. Whether a cable is acting as a source or as a termination is determined in the control box. If *A* and *C* are terminated and *B* and *D* are sources, the radiation maximum is to the right, from the terminations towards the sources. The transformer provides 180° phase inversion and, with the turns ratios shown, also transforms the 75 Ω impedances to 50 Ω at the receiver output.

There are eight different combinations of sources, terminations and relative phasing (0° or 180°). These combinations are summarized in Table 1.

Each combination has a specific pattern although configurations 5 and 6 have the same pattern as do 7 and 8. The result is four different patterns, two of which are reversible, that can be selected from the control box in the shack.

Figures 4 through 7 are for 475 kHz but the patterns at 1.83 MHz and 137 kHz are very similar except for differences in peak gain. This is illustrated in Figures 8 through 11, which compare the directivity patterns for 160 m and 630 m. The outer (higher gain) patterns are configuration 1, the loops are driven in-phase. The inner patterns are for configuration 3, loops driven 180° out of phase.

At 160 m, Figures 8 and 9 illustrate significantly improved directivity going from the loops in-phase to 180° out of phase, it also shows the significant reduction in peak gain (≈ -5 dBi). Figures 10 and 11 are for 630 m and again we see a significant improvement in directivity with 180° phasing, but an even larger reduction in peak gain (≈ -16 dBi). The patterns for 2200 m are very similar to 630 m except that there is another 20 dB of

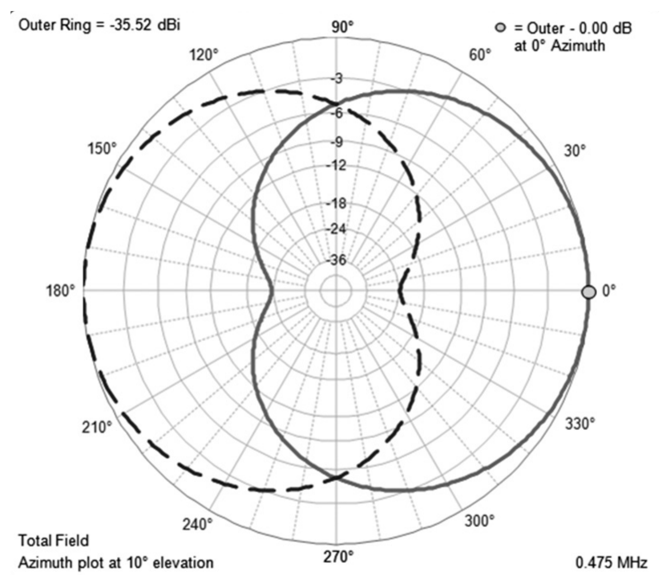

Figure 4 — Pattern for configurations 1 and 2.

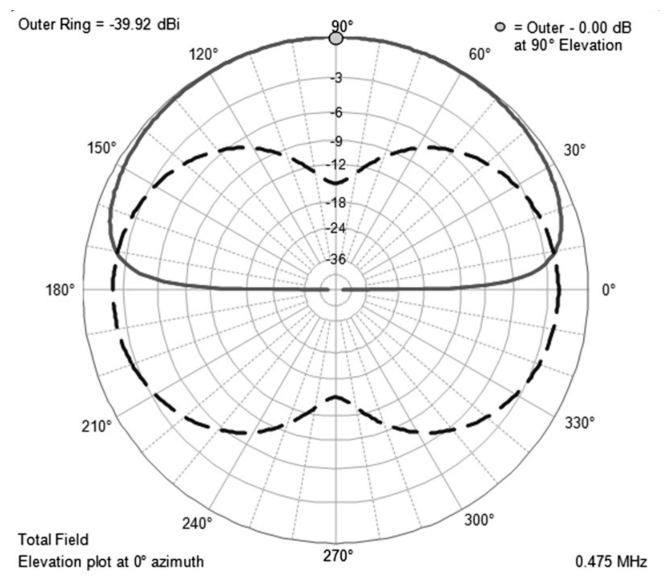

Figure 6 — Pattern for configurations 5 and 6.

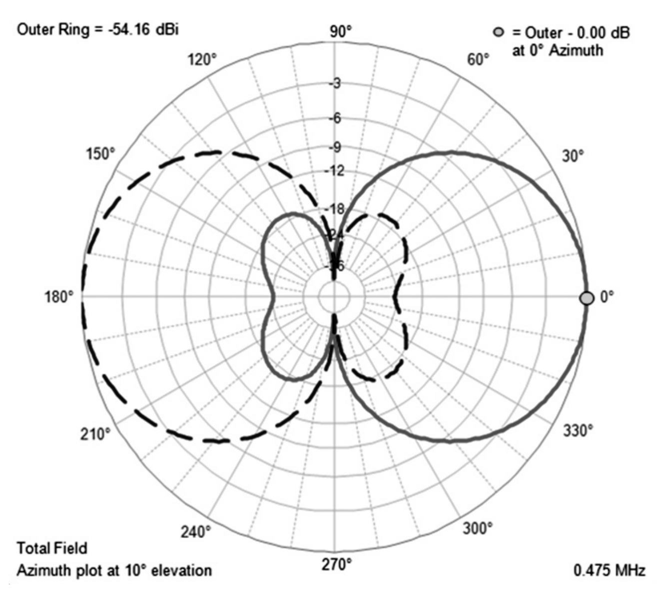

Figure 5 — Pattern for configurations 3 and 4.

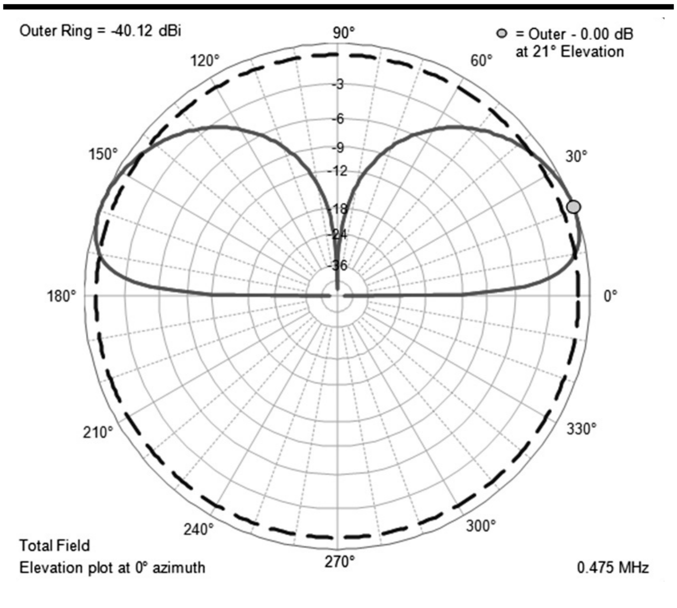

Figure 7 — Pattern for configurations 7 and 8.

gain reduction. The signal levels on 160 m and 630 m are not alarming low and on-the-air testing has shown that an amplifier is not needed. However, on 2200 m a preamp would be helpful — between 20 to 40 dB would be adequate — although I have been using my antenna successfully on 137 kHz for WSPR signals without additional receiver gain.

The predicted performance on 160 m, 630 m and 2200 m for different configurations is summarized in Table 2.

Near-field Patterns

All of the directivity patterns shown to this point have been for the far-field — many wavelengths from the antenna. At 475 kHz λ is $\approx$ 2,000 ft and at 137 kHz λ is $\approx$ 7,200 ft. The directivity pattern for any noise source — like a utility line or neighbors TV — within that distance will be the near-field pattern, which can be very different from the far-field pattern. Figures 12 and 13 show a comparison between near and far-field patterns with the noise source at a distance of 400 ft at 475 kHz for the near-field pattern.

The solid lines represent the far-field patterns and the dashed lines the near-field patterns. Note the scale is in m*V*/m not dB. When the loops are both driven in phase (configuration 1) there is some degradation in the near-field pattern compared to the far-field but it's not too severe. However, the difference between the near and far-field patterns with 180° phase difference (configurations 3 and 4) is very great. This is a very important observation for locations in congested urban environments. Although the far-field pattern with 180° phase difference is much more directive, the local noise rejection is grossly inferior. Configurations with 180° phase difference may not be usable in these situations.

Sensitivity to Shape

The configurations listed in Table 2 assume two symmetric triangles. To illustrate how insensitive to loop shape the antenna is, I modeled the variation shown in Figure 14, and show a performance comparison in Table 3. The first entry is Figure 1 and the second Figure 14.

The differences are very small. This implies that the primary driver for loop shape will be the available supports.

An Extended Version

I happen to have another 80 ft pole in line with the first two, again spaced 150 ft. I've considered duplicating the present antenna and extending it to four loops as shown in Figure 15. Figures 16 – 18 show patterns associated with Figure 15. Receive directional factor (RDF) is 13.6 dBi at 475 kHz with an antenna that is only 300 ft long! A comparable Beverage would be almost a mile long. However, the Beverage would have a lot more signal coming out of it.

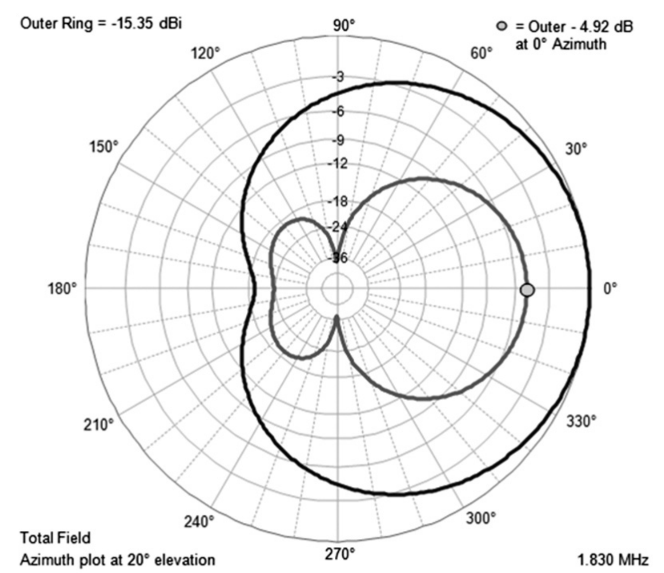

Figure 8 — 1.83 MHz azimuth plot at 20°.

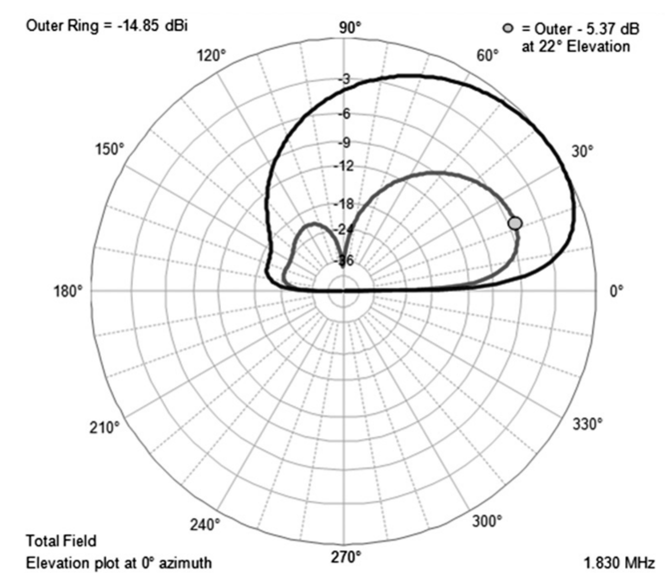

Figure 9 — 1.83 MHz elevation plot.

Table 2
Performance summary.

Band	*Configuration*	*F/B [dB],10° elev.*	*F/R [dB],10° elev.*	*RDF*	*Max gain [dBi]*	*at Az°*	*at El°*
160 m	1 & 2	18.39	3.91	7.13	-12.48	0	38
160 m	3 & 4	18.07	15.07	11.22	-20.12	0	22
160 m	5 & 6	0.00	0.00	6.33	-15.81	0	90
160 m	7 & 8	0.00	0.00	5.01	-17.40	0	26
630 m	1 & 2	23.49	5.22	7.71	-34.44	0	26
630 m	3 & 4	24.43	16.73	11.52	-53.55	0	18
630 m	5 & 6	0.00	0.00	5.47	-39.92	0	90
630 m	7 & 8	0.00	0.00	4.77	-40.12	0	20
2200 m	1 & 2	23.63	5.33	7.71	-55.46	0	20
2200 m	3 & 4	14.63	14.63	11.08	-85.18	0	14
2200 m	5 & 6	0.00	0.00	5.22	-61.38	0	90
2200 m	7 & 8	0.00	0.00	4.71	-61.08	0	16

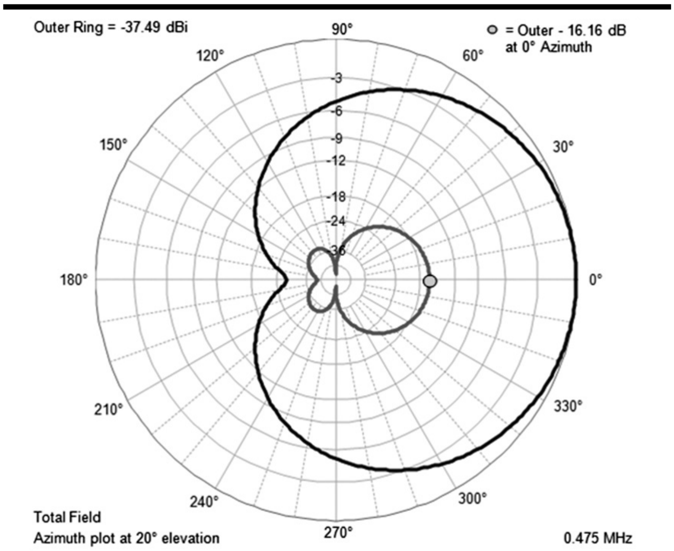

Figure 10 — 475 kHz azimuth plot.

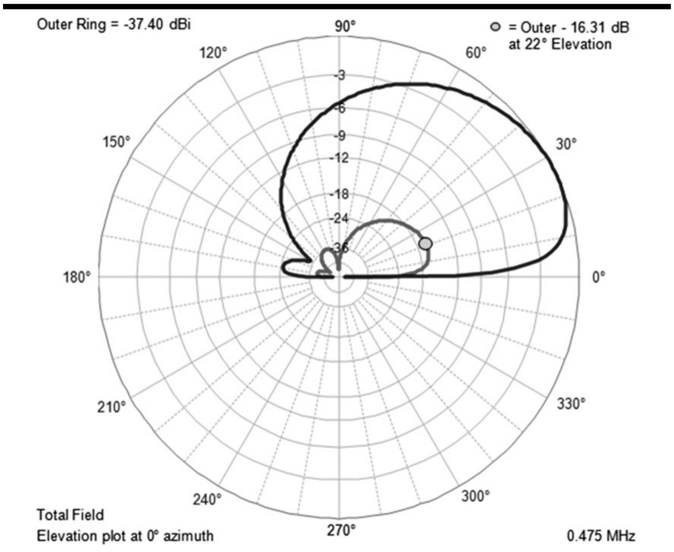

Figure 11 — 475 kHz elevation plot.

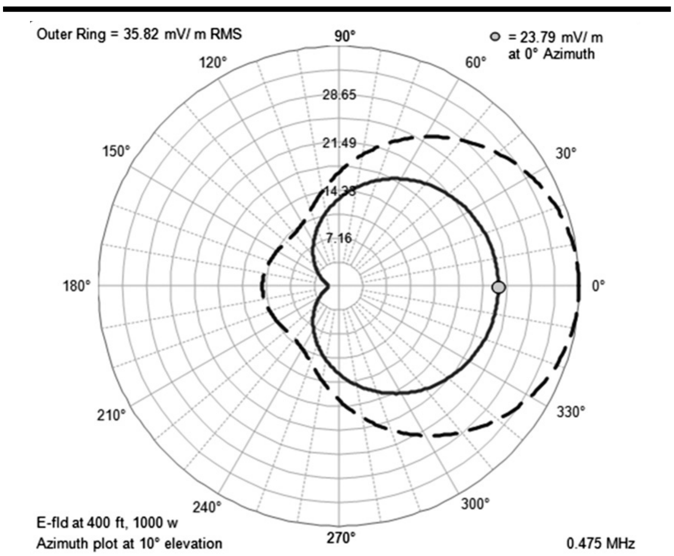

Figure 12 — Comparison between near and far-field patterns for zero phase difference.

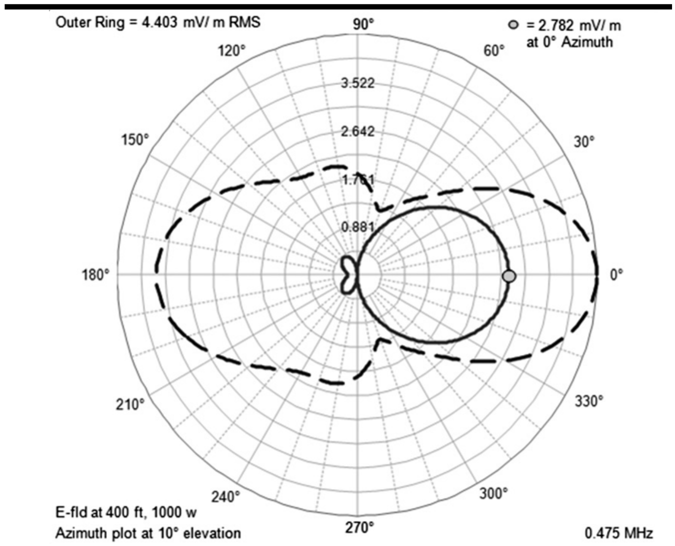

Figure 13 — Comparison between near and far-field patterns for 180° phase difference.

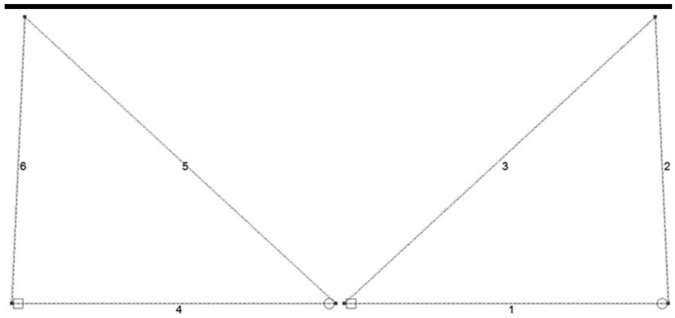

Figure 14 — An alternate loop shape.

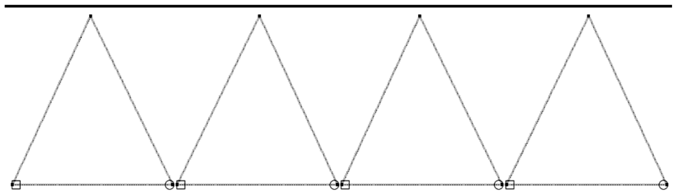

Figure 15 — Four loop version.

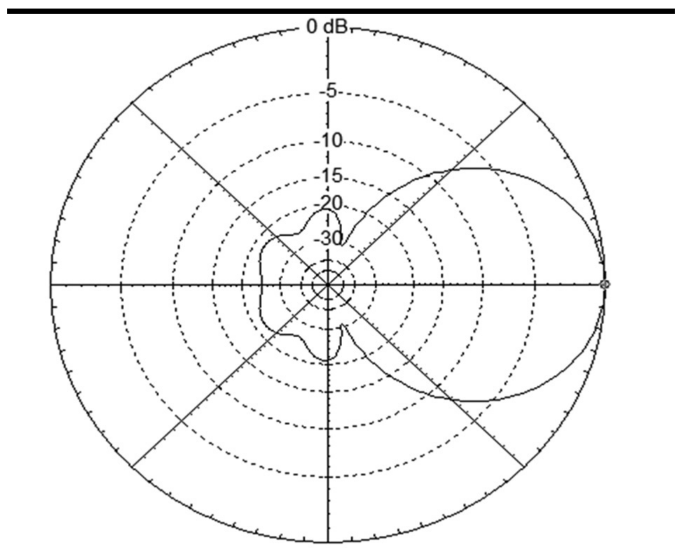

Figure 16 — Four loop azimuth pattern.

Table 3
Performance comparison.

Band	*Configuration*	*F/B [dB]at 10° elev.*	*F/R [dB] at10° elev.*	*RDF*	*Max gain [dBi]*	*at Az°*	*at El°*
630 m	1 & 2	23.49	5.22	7.71	-34.44	0	26
630 m	1 & 2	21.92	5.10	7.68	-34.36	0	26

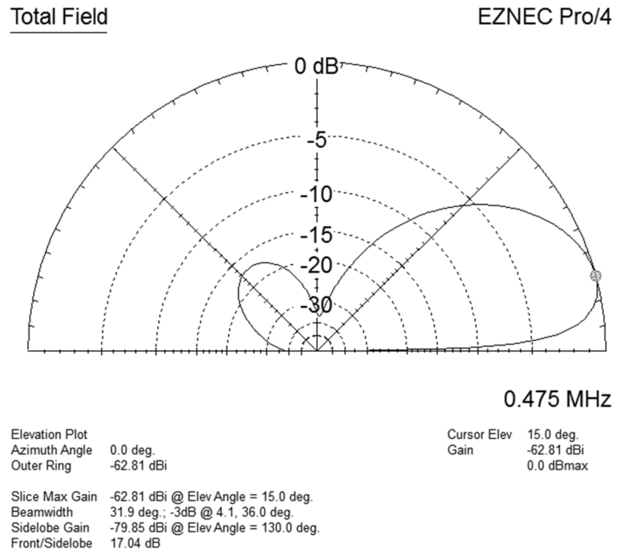

Figure 17 — Four loop elevation pattern.

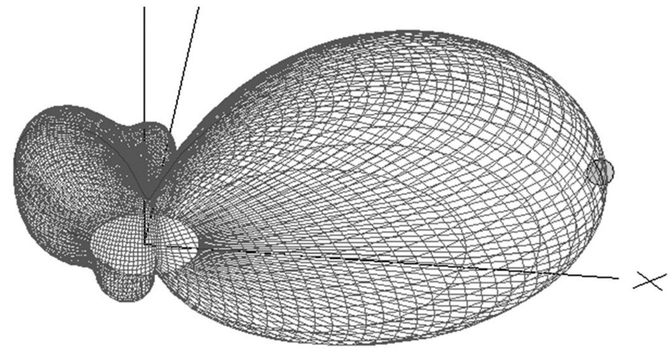

Figure 18 — 3-D pattern for four loops.

Verification

Modeling is a great tool, providing reliable predictions, but in the end it's necessary to verify the predictions and that the antenna is correctly assembled. Does this contraption actually work? After a careful visual check that all the electrical connections are correct, and that all of the transformer/chokes are correctly connected to provide proper phasing. Figures 1 and 2 have prominent phasing dots to indicate the proper connections. Even with careful assembly it is possible to switch one or more of the connections. There are a couple of ways to quickly check the polarity of the transformers. First, set the control to 0° phasing (configuration 1), then switch the direction (configuration 2). There should be no significant change in signal level for the background noise. If there is a large change then at least one of the transformers is reversed. Next change the phasing to 180° (configuration 3). There should be a substantial drop in signal level but the new level should not change much when the pattern is reversed (configuration 4). Finally, select a strong signal with a known direction, more or less in line with the main lobe, then reverse the pattern. This should show the F/B of the array and confirm the directions are correct. If all these are as expected then you probably have the phasing correct.

You can also make some impedance measurements. The feed system is designed for 75 Ω up to the control box, and the impedances within the feed system should be close to this over the entire frequency range. Using a VNA2180 vector network analyzer I measured the impedances at several points from 100 kHz to 2 MHz as I switched the control box through the various configurations. The first point was the output port to the receiver. The impedance was close to 50 Ω as designed. Tthe phase inversion transformer converts the 75 Ω impedance of the feed system to 50 Ω for the receiver. I next measured the impedances at the control box end of the feed cables one at a time while switching between configurations. Each of these measurements was a sweep over the frequency range. All of the graph plots were very similar with an SWR < 1.5:1, indicating there were no major errors. The antenna impedances agreed with predictions.

That was the easy part! The next step was to verify that the antenna had the predicted directivity patterns associated with each configuration. The ideal procedure would be to place a signal source well beyond the Fresnel zone, that is, more than 10 λ distant at various azimuths and measure signal strengths as the pattern was switched. At 137 kHz or even 475 kHz the distances to the sources would have to be many miles

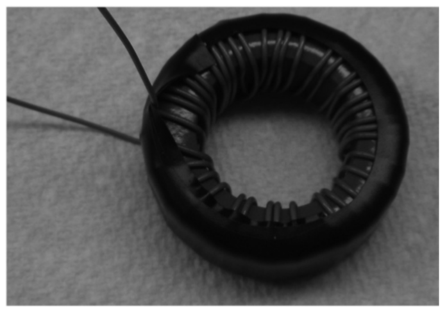

Figure 19 — Secondary winding on the impedance transformer.

although at 1.8 MHz the distances are not so great. My location is in a small valley surrounded in most directions by hills so this approach did not seem practical except perhaps for checking the depth of a null in a particular direction on 160 m. I needed to be a bit more crafty! Because the patterns are basically the same from 100 kHz to 2 MHz, I realized I could use signals anywhere in that range. There are a large number of well defined signals in this range, most prominently AM broadcast stations. There are also aeronautical and coastal navigation beacons and the WSPR transmissions by Amateur Radio experimental stations. From long experience with Yagis and other arrays we know that the null depth and location is much more sensitive than the details of the main lobe. In general if the nulls are where they should be and the null depth anywhere near what it should be, then we can have confidence that the pattern is close to its predicted form. Locating and measuring pattern nulls can take us a long way towards verifying the actual pattern.

To identify and measure signals I have an old HP3585A spectrum analyzer. This allowed me to see the station signals and measure their amplitudes. The instrument displays the amplitude to 0.01 dB but that's deceiving. Even strong local BC signals have several dB of variation (noise) even with very narrow scans, which makes resolution of the main lobe impractical but it's still possible to get a good estimate of null depths and locations by observing the signal while switching the pattern direction. Switching the pattern doesn't help however, with the nulls to the side (±90°, see Figure 5). I was able to find BC stations lying along the axis of the array which showed the predicted F/B ratios reasonably well. The preliminary measurements with BC and 630 m WSPR stations indicate the patterns are close to the NEC predictions, at least the nulls.

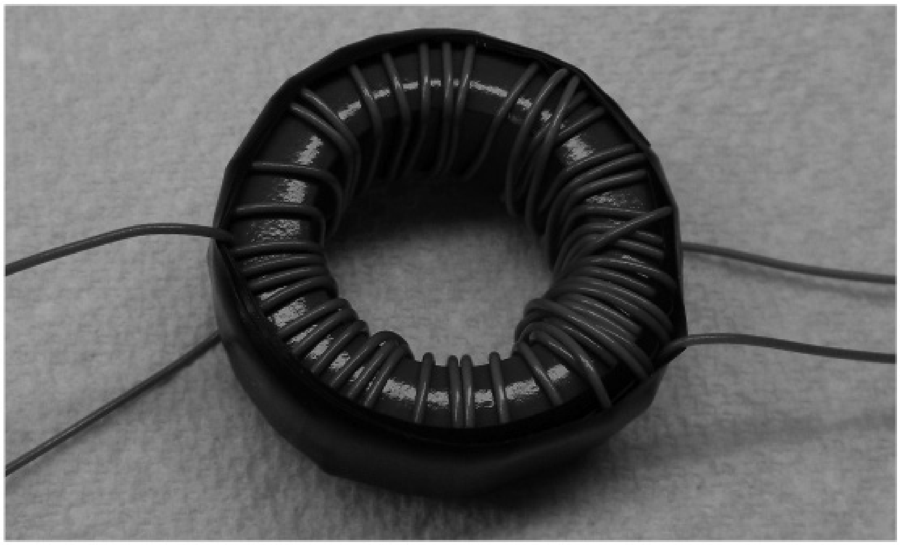

Figure 20 — Primary winding added to the impedance transformer.

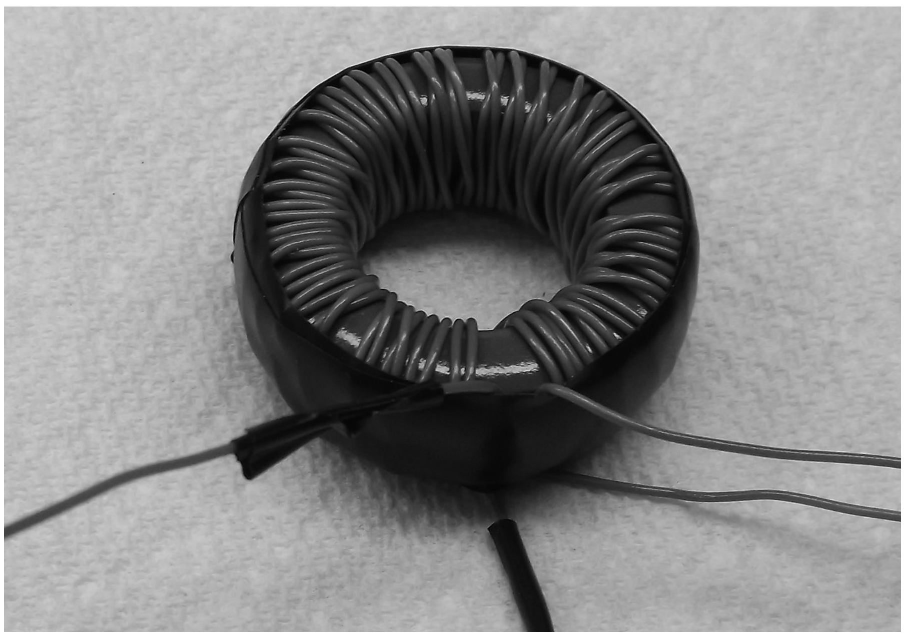

Figure 21 — Common mode choke.

Transformers and Control Unit Details

As indicated in Figure 1, the loops are fed or terminated at the lower corners. At each point (*A*, *B*, *C* and *D*) there is an isolated impedance transformer, 1000 Ω to 75 Ω like the one shown in Figure 2. To further isolate the transmission lines from the antenna, on the primary of the impedance transformer there is a common mode choke. Note the use of winding polarity dots in the transformer-choke schematic of Figure 2. Keeping track of the phasing is critical! When toroidal cores are used, two windings are in phase — the same dot — when both wires come out of the core in the same direction.

The impedance transformers, the common mode chokes, and the phase inversion transformer are all wound on the same toroidal ferrite core, Fair-Rite #5977002721. Nine cores are needed for this project. I obtained them from Mouser Electronics for about $3.75 US each.[3] These cores are type 77 ferrite, recommended for use in low flux applications below 3 MHz. All of the windings used #26 AWG insulated wire. Neither the wire size nor the insulation type is critical. I simply used what I had on hand. You have to use wire small enough for the windings to fit on the cores. The magnetic components must to work from 137 kHz through 1.9 MHz. The feed-point transformers are used to isolate the antenna from the feed system and to transform the to 75 Ω resistance on the primary to1000 Ω on the secondary to properly terminate the loops. The transformer shunt impedance has be significantly greater than1000 Ω to maintain proper termination. This has to be the case over the entire range of 137 kHz to 2 MHz. At the low frequency the issue is enough inductance with a reasonable number of turns. The type 77 ferrite has high permeability, about 2000, up to 1 MHz, above which it starts to decrease but is still adequate for this application at 2 MHz. We also have to maintain a sufficiently high self resonant frequency, f_r, so that there is sufficient shunt impedance, Z_s, at 2 MHz. Like the transformer, the choke also needs to have sufficient Z_s over the entire range. This becomes a bit of a balancing act, more turns give more low frequency impedance but lower f_r with reduced impedance at 2 MHz. 35 turns gave f_r=700 kHz, with Z_s=2.8 kΩ at 137 kHz, 20 kΩ at 475 kHz and 6.1 kΩ at 1.8 MHz. These values, while not ideal, are an acceptable compromise. Figures 19 – 21 show some of the winding details.

The common mode choke has 35 turns wound bifilar (two wires twisted together). Note the careful marking of one pair of wires, these allow us to indentify each of the windings. As shown in Figure 2, for correct phasing the center conductor of the feed line must be connected to the dotted end of the primary winding. As shown in Figure 21, I placed a small piece of tape on one winding. On the bottom of the choke I connected the taped winding to the center conductor of the input F connector. I then connected the other end of the taped winding to the dotted end of the transformer.

Note also that the ends of two windings come out on the same side of the toroid, the windings from the same side have the same polarity — they share the same "dot". This convention applies also to the impedance transformer.

The transformer-chokes were installed in insulated junction boxes (Figure 22) available at most hardware stores. The left box is for point *A* in Figure 1. Points *B* and *C* are combined in a common box (middle) and point *D* is in the box on the right. The cores are secured with some silicone caulk/adhesive. The terminals to which the antenna wire is attached were simple SS machine screws in holes through the sides of the boxes. The holes were tight and caulked with silicone.

The installation at point *B* – *C* at the center of the antenna is shown in Figure 23. Notice the careful markings on the box and the cables to keep track of proper phasing and cable connections. For the antenna to work as expected it is vital that all the connections are correct. To this end every cable was marked at both ends, *A*, *B*, *C*, etc. Every RF connector on the feed point boxes and the control unit was also carefully marked to avoid confusion during assembly. The antenna was made from #17 AWG aluminum electric fence wire.

Summary

The final version of my antenna is basically the same as DK6ED's, just scaled up and with some added switching to give additional patterns. There are four modes of operation, two of which are reversible. On several occasions while using the antenna I've found the pattern associated with 180° phase shift to be too narrow for general listening. The deep side nulls cut out stations north and south of me. In fact most of the time I leave the loops in-phase, switching to 180° phasing only when it seems to help. I have been using the antenna on 160 m, 630 m and 2200 m without an amplifier. This has worked very well, however, if the antenna were scaled down in size, an amplifier might be needed especially on 2200 m.

I spent a great deal of time trying to optimize this antenna, varying the shape, relative phasing, termination resistances and even exploring reactive terminations. I found all this made very little difference. The antenna seemed to work about the same no matter what I did to it. Even changing the soil characteristics under the antenna has only modest effect. The received signal amplitude is a function of the size of the loops. Bigger loop mean more signal, but that's about all that changes as the loop size is varied.

Rudy Severns, N6LF, was first licensed as WN7AWG in 1954. He is a retired electrical engineer, an IEEE Fellow and ARRL Life Member.

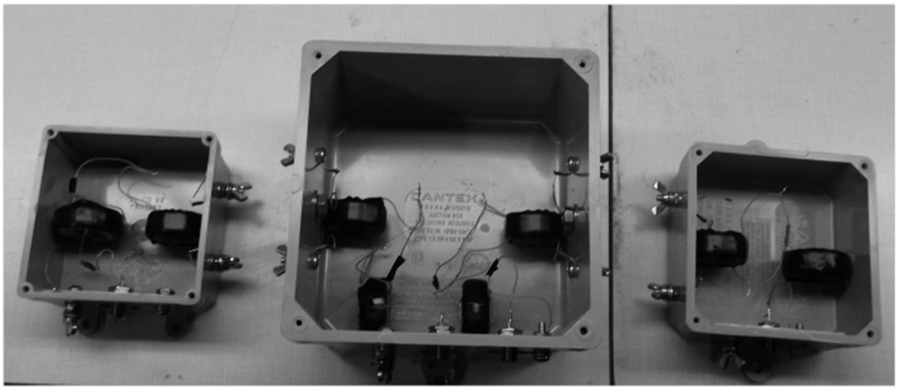

Figure 22 — Feed point boxes with transformer-chokes installed.

Figure 23 — Transformer box at the center of the array.

Notes

[1]Chris Kunze, DK6ED, "The DK6ED Double Loop", *QST* Mar 2015, pp. 34-37.

[2]Several versions of *EZNEC* antenna modeling software are available from developer Roy Lewallen, W7EL, at **www.eznec.com**.

[3]**www.mouser.com**.

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A Somewhat Different Conceptual Approach to Inductance

A practical approach based on physcial principles to inductance, capacitance, and impedance of wires and coils.

Inductance of a wire in air is, to a considerable degree, about 0.3 or 0.4 μH/ft of length for a wide range of wire size, and is not very much dependent on how it is coiled, which can be verified from well-known formulas. This insight lessens reliance on formulas and charts in coil design, and helps with an understanding of wiring, antennas, ground radials and ground leads, plus coils, chokes, and transformers. Furthermore, it's surprising but explainable why one can't exceed perhaps a few hundred ohms of relatively pure inductive reactance without approaching resonance. Air-core transformers can be made that work fairly well. Capacitance of a wire has its own similar rule, about 2 to 4 pF/ft. A long discussion is possible, but the approach here is to describe the end results first, then give theory. All these observations can be partially confirmed with something as basic, yet valuable, as a MFJ-259 (or similar) antenna analyzer.

The writer has long been inspired by the late Henry Martyn Paynter, Sc.D., Professor of Mechanical Engineering at Massachusetts Institute of Technology, back in the late 1960s, who seemed always to try to generalize practical and theoretical knowledge and explore its ramifications. Any errors and shortcomings are mine, not Prof. Paynter's. The initial goal was understanding the use of air-core coils as step-up autotransformers, but I had to go back to basics, like this article does.

Prior Art

A worker at the U.S. Bureau of Standards, around a hundred years ago, apparently came up with the basic formula for inductance of a single conductor, which is repeated in works like Terman's and the Radio Amateur's Handbook. Another person several decades ago — in perhaps *QST* — reportedly treated the matter of high-Q tank circuits working well as transformers. The torus/donut form for maximum inductance was determined long ago, and may even carry a person's name.

Because coils aren't very different from single conductors. Non-coiled conductors will be treated first. Since inductance is so simply stated, first consideration is given to impedance of wires, not inductance. Then, the effect of coiling is explained. Later on, the underlying physical theory is discussed, which also shows the limitations of this approach. Finally, a design example is given. The discussion follows this outline.

Part 1 — Impedance of wires
Part 2 — Inductance of coils
Part 3 — Impedance of coils
Part 4 — Air-core voltage dividers and transformers, also autotransformers.
Part 5 — Theory and limitations
Part 6 — Some design examples.

Part 1 — Impedance of Wires

Inductance of wires, to a fairly good approximation, is declared as roughly 0.3 or 0.4 μH/ft. However, what normally interests radio engineers is impedance, which involves frequency. A wire has about 8 Ω/ft of inductive reactance at 4 MHz, with this reactance being directly proportional to frequency, within limits. There also is capacitive reactance of about 5 to 10 thousand Ω/ft, again at 4 MHz, except that capacitive reactance of course varies inversely with frequency. Inductance never is "pure," particularly when its impedance is over perhaps a hundred ohms, or when the length of the wire approaches a quarter of an electrical wavelength. Wires function as antennas, whether this is desired or not.

Impedance of wires that function as antennas, ground leads, ground radials, and wide-spaced transmission two-wire line

An infinitely long wire in empty space has a so-called "surge impedance" of about 400 Ω, resistive, which actually is a property of space via the fundamental constants of electrostatics and electromagnetism. Two wires, fairly widely spaced, have an impedance of double this — about 800 Ωs between them; this is the practical upper limit for any two-wire line. This is a convenient starting-point for discussion. It follows that anytime one measures an impedance different from about 400 Ω resistive and not independent of frequency, it's not an infinitely long wire in empty space — it possibly has an end somewhere nearby, with or without something connected to that end, or something is beside or around the wire, like for twin lead or coaxial cable.

Quarter-wave Antenna

If the far end of a wire is "grounded" — often practically meaning looped around to the ground of one's measuring instrument, true grounding is hard to achieve — at low frequency the behavior is that of an inductance of about 0.3 or 0.4 μH/ft, as already described. Take the frequency up to the first resonance (at one fourth of a wavelength) and one gets an impedance of several thousand ohms, probably its radiation resistance. This is a quarter-wave (normally vertical) antenna, with impedance measured at the free end. There is a practical problem in how one would actually do this measurement. Zero radiation loss would result in a near-infinite measured impedance at resonance. Just above resonance, the impedance is that of a small capacitance. At frequencies other than resonant, there is substantial reactance. There are an infinite number of resonances, but only the first few are significant.

Half-wave Antenna

If the far end of the wire is free, at low frequency the principal reactance is capacitive. If frequency is raised to the point that the wire is an electrical quarter wavelength, the returning wave from the far end, reduced slightly by radiation, decreases the impedance to something like 36 Ω. This is a base-fed, quarter-wave vertical, or one side of a half-wave dipole antenna. Zero radiation and other loss would produce 0 Ω impedance at resonance. If the free-ended wire is an electrical half wavelength, the returning wave boosts the impedance also to several thousand ohms, again its radiation resistance. This is an end-fed, half-wave antenna.

Ground Radials

Ground radials can be understood in the same way as antennas. If they are extremely long, they will show roughly 400 Ω each, which tends to be unsatisfactory but quite normal. If they are exactly a quarter wavelength, they will have an impedance of about 36 Ω each, which is about the minimum achievable. Having *n* of these connected in parallel reduces the impedance to 36/*n* Ω and causes more power to go into the presumed quarter-wave vertical antenna. It probably takes about ten random-length radials to force half the power into the vertical portion of the antenna. If radials happen to be cut to a half wavelength, impedance would be several thousand ohms, the worst possible impedance for a grounding radial. This is for radials clear of the ground. If the wire is anywhere near soil of any kind, the situation is harder to analyze, but it clearly will be very difficult to achieve as low as 36 Ω because this is the impedance of about four or five feet of wire at 4 MHz!

It should be possible to tune out either inductive or capacitive reactance of any ground, or ground radial. An infinitely long, or very long, radial can't be tuned because there's no reflection from the far end to change the impedance. Short antennas have low radiation resistance at resonance. The drawback is that they have to be resonated at the frequency used, so frequency-hopping is made more difficult. If one can use a half-wavelength for a vertical, this gives a low (maybe zero) radiation angle and doesn't need nearly as effective a ground, by a factor of about 100 in impedance, relative to a quarter-wave radiating element. End-feed it with open-wire 800 Ω line (effectively a Windom vertical), or center-feed it with twin-lead or coax.

Easy, effective grounding seems nearly impossible to achieve. About all that can be done is to bond all the equipment together (including the operator), because of the impedance situation stated in the previous paragraph. Clearly, one shouldn't try to export RF power via a single wire. The worst imaginable case is that of trying to feed a quarter-wave vertical straight out of the transmitter, and the best possible "ground" probably would be to have numerous tuned, quarter-wave radials well above the ground. An alternative would be to have several separate, short, separately-ground-rod-connected wires with conductive soil, but this is not feasible much above 4 MHz. One should not have ground wires that are a quarter wavelength at the frequency used!

This is about all that I can say for the impedance of wires, but it should facilitate understanding of antennas, ground radials, feed lines and grounding, no small achievement. When coiled wires are considered, it will be found that things are not greatly different, though radiation resistance vanishes when size in electrical wavelengths is small.

Part 2 — Inductance of Coils.

Right off, one should forget about any idea that air-cored coils will have inductance proportional to number of turns squared; this is true only in very special circumstances. It's more like to the first power, thus proportional to wire length. For ferromagnetic cores, however, the turns-squared rule is very appropriate.

Here is the Low-down on Coils

For a one-turn loop or a multi-turn coil with well-spaced turns, inductance is about the same as for a straight wire of equal length. One can boost a single-layer coil inductance to about double by close-winding the turns, and one can get to about four times the straight-wire inductance by winding it as a compact torus (donut) shape. Curiously, one can actually *decrease* the inductance of a length of wire significantly, to about half, by coiling it then stretching it a lot. This is about the range of what's possible with wire — from one-half to about four times the nominal 0.3 to 0.4 μH/ft. I apologize for condensing such an important subject into a single paragraph, but that's the power of this approach. It's simpler than going to charts and formulas, if you don't mind perhaps being off by 10-20%.

A little reflection on the foregoing gives the rule for maximizing inductance — place each new turn of wire as close as possible to *all the other turns*. To really minimize inductance of a single-layer coil, space the turns about one winding diameter. This all is the result of magnetic flux being somewhat concentrated around a conductor. Although inductance of a coil is somewhat cut-and-dried with the above approach, there are some aspects that merit further attention.

First, turns spacing *is* quite important. The discussion above gave a range of about 0.5 up to about a maximum of 4 times the basic inductance of the wire used, depending on how the wire is coiled. While one gets the 0.5 figure by stretching a coil a whole lot, one gets from 1.0 to 2.0 by winding tightly in a single layer, or one gets to 4.0 by winding a tight torus or donut shape. A single-layer coil can be stretched from 2.0 down to around 1.0 within the elastic limit of the wire if the wire is somewhat thin relative to the winding diameter. Clever winding and stretching of the turns of a donut-shaped coil seemingly could vary the inductance, within the elastic limit of the wire, from about 4.0 down to around 1.0. Either way, this is enough range to employ these as tuning elements. Of course, VHF circuits had been routinely fine-tuned by "knifing the coil" [inserting a knife blade between closely spaced turns to widen the spacing. — *Ed*.].

Second, conductive objects near a coil don't change inductance much, which also seems a little surprising. Since inductance of a coil is mostly inductance of the wire, and is enhanced only by wires being very close together, for something to change the total inductance of a coil, that something must be quite close to *all* of the wire, not just be positioned somewhere in the vicinity of the coil itself. The rule of keeping metal at least one coil radius from the end of the coil is very conservative. One can place very large conductive items (like an aluminum soda can) actually inside an air-core coil without changing its inductance much; winding enameled wire directly on a section of copper pipe still leaves significant inductance, though it's not recommended!

Third, shorting just a few turns of an air-cored coil with spaced turns doesn't change inductance very much, another surprising result. It's as though the shorted turn or turns, plus part of another turn, were simply removed physically from the coil, so inductance is proportionately reduced. This is completely different from iron-cored coils, where a short drastically reduces inductance and the shorted turn sometimes overheats. This shorting technique works fine, even in broadcast transmitters at the hundreds of kilowatts level, because air-cored coil inductance is largely inductance of the conductor itself, with some mutual inductance from the immediately adjacent turns, but with very little inductive coupling to other, more remote, turns.

Last, a so-called "choke" is just a coil, an inductor. Sometimes these are wound with "pies" connected in series, as will be discussed in the next Part.

Part 3 — Impedance of Coils

Impedance is what radio engineers expect from inductors, and inductance is typically related to impedance by a formula that is deceptively simple. But that "surge impedance" of about 400 Ω, and "stray capacitance", again are involved. An explanation follows.

If one winds a coil that has a certain measured inductance at low frequency and then tries to use it at a frequency where its impedance should approach 400 Ω, it will have significantly higher measured impedance because it actually is approaching resonance, or starting to approach a quarter wavelength. An alternative explanation, just as valid, is that "stray" capacitance is canceling some of the inductive current and one measures a higher apparent inductive reactance. A "smart" measuring instrument such as the MFJ259A reports increasing "inductance" as frequency is raised. One should ignore (or almost ignore, see below) the part of the inductance-capacitance-reactance chart in the *Handbook* that shows thousands of ohms for inductive reactance. There is no such thing as inductance without capacitance, nor is there capacitance without inductance; it's a matter of their relative proportions, always, which involves size and frequency, thus wavelength. For careful work, one must always consider a coil as inductance with capacitance in parallel. Careful modeling of a capacitor also has to consider that series inductance is also always present.

There is one important situation where stray capacitance isn't significant, and that is where the inductor is paralleled with a capacitance that is much greater than the stray capacitance. This makes a sharply-resonant "tank" circuit that has to be used at a frequency much lower than the frequency at which the inductor is self-resonant. High power and large circulating currents are often involved. Such circuits are always used at the resonant frequency of the tank circuit. The presence of large circulating currents can make coils work like transformers when otherwise they wouldn't, as will be discussed below.

Part 4 — Air-core Voltage Dividers, Transformers and Autotransformers.

This is getting close to the understanding I was trying to attain. With an iron core, transformer operation is trivial in terms of the basic theory. One can grossly over-simplify and characterize these by just their turns ratio, operate them as voltage dividers and step-up autotransformers, and expect that maximum power output will be a large fraction of power input. With air-core coils it's another matter entirely, because all of the flux doesn't link all of the turns.

First, consider a single-layer coil with spaced turns used as a voltage divider. It works quite well, and is quite linear because there's little coupling between turns. Problem is, the maximum current that can be drawn is just the reactive current flowing in the coil, not a current that is inversely proportional to the step-down ratio, and the voltage droops when current is drawn. It's impossible to have anywhere near as much power out as power in. It's even worse if link-coupling is attempted. So, this works basically like a resistive voltage divider, not a step-down transformer. It's equivalent to a capacitive voltage divider.

A narrow-band air-core transformer is "old hat," having been used since the beginning of radio. These have greatly reduced inductance and are resonated with a capacitor to increase impedance at the resonant frequency. Because the circulating current (and power) is so high, it's possible for very poor coupling, or a tap at one end of the coil, to take away as much power as is being put into the "tank" circuit, so that impedance transformation and power transfer both work well and the device acts like a transformer at the resonant frequency. It can even be tapped and used as a quite linear voltage divider, and used at any fraction of a turn. This is just fine as a step-down autotransformer. The old-fashioned "link coupling" works this way too, except that the step-down ratio can't easily be calculated.

For use as a step-up autotransformer, still with the narrow-band tank circuit, one can just reverse the input and output connections. This leaves the capacitor connected from the output to the neutral, and the input tapped down. Another type of step-up autotransformer can be made by placing the step-up winding very close to the wire of the main coil and using somewhat more turns than would be calculated, to compensate for some flux loss between the two, essentially a "bifilar" winding on part of the main tank coil. If the extra turns aren't closely spaced, significantly more turns are needed, calculation of the step-up ratio is not easily done, and the output will "droop" drastically under increased load.

One can fairly successfully make a broadband, air-core, one-to-one ratio transformer by using the "bifilar" technique: two wires, parallel and close together, one as primary and one as secondary. This can be made two-to-one (or one-to-a-half) by

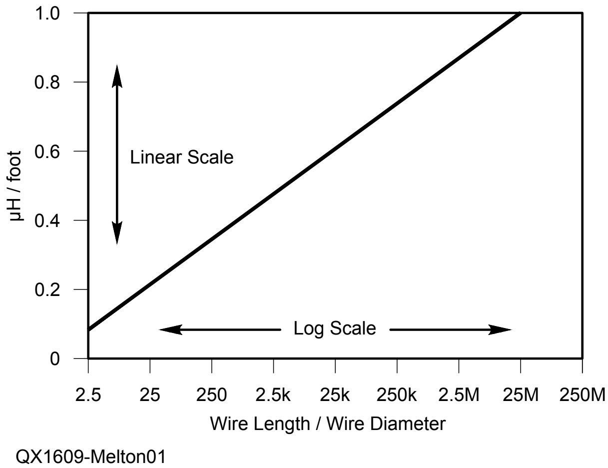

Figure 1 — Incremental inductance of straight wire *vs.* length to diameter ratio of the of wire.

connecting the wires as an autotransformer. This can be extended to trifilar, even quad or more. This is standard with high-power solid-state amplifiers, to get impedance up to 50 Ω. Ferrite cores are often used to reduce reactive current in the primary. Inter-turn capacitance is high in these, so if too much wire is used, they will self-resonate, which may not be a total disaster. They probably are usable above self-resonance. These must have an integer turns ratio, they have several wires around one wire of the low-impedance winding, and they must be long enough to give sufficiently low shunt reactance. Insulation is a consideration if voltages are high. One ideally wishes to have zero-thickness insulation for closest coupling.

Conclusion

An effort has been made to take some of the mystery out of coils and wires, as well as air-core transformers, by pointing out that in many cases, the coupling between turns isn't enough to change the inductance much compared to that of a lone wire, and that even tight coiling has its limits, which makes the length of the wire more significant than is apparent from the usual inductance charts. Electrical wavelength of the conductor is very important. Inductive reactance is never "pure," capacitance invariably is present to some extent.

Part 5 — Theory

The foregoing was simply declared with little explanation. A more careful technical explanation follows of the inductance and capacitance of wire conductors.

Around a wire, the strength of magnetic induction is inversely proportional to radial distance. Presuming skin-effect operation, if total magnetic field energy is calculated, it's a logarithmic function, which means that equal amounts of energy are in each zone, such as 1-2, 2-4, 4-8, 8-16 radii, or 1-10, 10-100, 100-1000, etc. This theoretically goes out to infinity, and an infinitely long wire computes as having infinite inductance per unit length. All practical wires are in some kind of loop or circuit, and the magnetic field at great distance is cancelled out by the field from the return side of the loop. A fairly accurate result is obtained by presuming zero energy beyond a distance equal to about the length of the wire, which yields the formula in handbooks. The formula seems to integrate to double the length, which actually doesn't matter much. A wire of radius 0.1 inches that is 100 inches long includes about three of those decade ranges (0.1-1, 1-10. 10-100), and if one changes the "wire" size to 1.0 inch radius the inductance will go down to 2/3. Going to 0.01 inches wire radius will raise the inductance to 4/3. Practically speaking, this is a lot of change in wire size without a very great change in inductance, which allows for the "microhenry per foot" approach. This is the direct result of inductance per unit length being proportional to the logarithm of the length-to-diameter ratio.

The unexpected reduction by about half from having turns spaced by the winding diameter results from the inductive coupling from a portion of the winding being positive in the part of the next turn on the same side, and negative on the part of the adjacent turn that is on the opposite side from the imagined portion. Thus, there is cancellation when the turns spacing is equal to coil diameter. This is instructive but not very useful.

This rule-of-thumb approach is completely in accord with the accepted formula for inductance of a straight wire, as found in references like Terman and the *ARRL Handbook*.[1] The figure of 0.3 to 0.4 µH/ft applies to wires that have a length-to-diameter ratio in the range of roughly 100 to 500. If your "wire" has a length-to-diameter ratio of 25, you'd best use 0.22 µH/ft; if the ratio is 2500, figure on 0.5 µH/ft. If you can get to a length-to-diameter ratio of 25 million, you can achieve 1.0 µH/ft, but this has to be very long or very fine — 2000 feet of 1-mil wire will satisfy this. The human hair is reportedly 3 mils thick! The relationship between inductance L in µH/ft and the logarithm of the length-to-diameter ratio x/d is shown in Figure 1, and expressed by

$$L = 0.06096 \left[2.303 \log\left(\frac{4x}{d}\right) - 1 \right]$$

[The leading constant is 0.20 for inductance per length in µH/m. A corresponding formula for capacitance of a straight wire is

$$C = \frac{17}{\left[2.303 \log\left(\frac{4x}{d}\right) - 1 \right]} \quad \text{pF/ft}$$

The constant in the numerator is 55.63 for pF/m. Formula from: Yu. Ya Iossel', E. S. Kochanov and M. G. Strunskiy, "The Calculation of Electrical Capacitance," (Translation) US Air Force Report, FTD-MT-24-269-70, 1969. — *Ed*.].

Cylindrical Geometry: Coaxial Capacitor or Inductor

If one considers two concentric cylinders of conductor material, one inside the other, for inductance and for capacitance per foot of the inner conductor, it turns out that the actual size isn't important, only the relative proportions. For the case where the inner conductor is a tenth the diameter of the outer conductor, inductance is about 0.14 µH/ft and capacitance is about 8 pF/ft. If the outer portion is 100 times the size of the inner conductor, inductance goes to about 0.28 µH/ft and capacitance drops to about 4 pF/ft. A thousand times gives an inductance of about 0.42 µH/ft and a capacitance of about 2.66 pF/ft. Each decade of increase gives another 0.14 µH/ft of inductance and inserts another 8 pF/ft in series with the existing capacitance. This is the same as stated in the previous paragraph, and represents logarithms at work, pure and simple.

Applying this rule-of-thumb in practice is straightforward. For capacitance, just estimate how wide the basic conductive environment is, and ratio this to the wire diameter. Ten to one gives 8 pF/ft, a hundred to one gives maybe 4 pF/ft, and a thousand to one, 2.67 pF/ft. Figure that the wire is a little longer than its physical length to allow for end fringing effects. We already did the example for the inductance of a wire, but here it is, slightly reworked: Figure the ratio of the diameter of the space the wire passes through to the diameter of the wire, if there's something that will limit the extent of the magnetic field, or if there's no limitation, use the wire length. If your ratio computes as ten to one, it's about 0.14 µH/ft, a hundred to one, 0.28 µH/ft, and a thousand to one, 0.42 µH/ft. All this becomes largely irrelevant if the frequency used results in the length that approaches a significant fraction of a quarter wavelength.

It superficially appears that compact, low-inductance capacitors can be made by stacking plates alternately, while compact, low capacitance inductors can't. This isn't quite accurate. The comparable situation for maximum inductive reactance with minimum capacitance probably is achieved by series-connecting "pies" — the ancient 2.5 mH choke was made this way since perhaps 100 years ago. This is the stacked inductor analog to the multiple-plate capacitor.

The foregoing is the "basic physics" explanation of the "microhenry per foot" approach to inductance of wires and coils. The limitations were likewise explained, namely that the length-to-diameter ratio of the conductor does change the inductance per foot, though not strongly. Electrical wavelength as a major consideration has been mentioned all through this article.

For a recap, it's been shown how one can dispense with formulas and charts and reckon inductance as well as impedance for wires and coils rather accurately. In the process, it was shown how voltage dividers and transformers made with wire, work and don't work. One can come away with the idea of 0.3 to 0.4 µH/ft, and reactance of 8 Ω/ft of length at 4 MHz, with this raised up to about double for close-wound, single-layer coils, or up to four times for coils wound into the donut or toroid shape, multilayer. Make adjustments for very high or very low length-to-diameter ratios. One could also think of perhaps 4 pF/ft, up to maybe 8 pF/ft or more in close quarters, which might be roughly 10,000 to 5,000 Ω/ft capacitive reactance at 4 MHz. Also if capacitive reactance approaches inductive reactance, that is the definition of resonance.

Part 6 — Some Design Examples

Next we will see how utterly easy it is to design a tank coil using the above approach. But first, two calculation shortcuts:

1 µH has a calculated inductive reactance of about 25 Ω at 4 MHz.

1 pF has a calculated capacitive reactance of about 40,000 Ω at 4 MHz.

To design a tank coil that will have a reactance of 80 Ω at 4 MHz, which would work for an 800 W linear amplifier, think of using wire, which when straight has an inductance of 0.33 µH/ft, as this article declares. Think of coiling it with somewhat spaced turns and getting 1.5 times that inductance, or 0.5 µH/ft. This would be about 12 Ω of reactance per foot, so that to get to 80 Ω one uses approximately 7 ft of wire. This is about 7 spaced turns on a form that has an outside diameter of 3.5 inches. That's all there is to it! If the length-to-diameter ratio of the wire is not inside the range 100 to 1000, the result will change slightly.

The capacitor to resonate this coil at 4 MHz must have the same 80 Ω, but of capacitive reactance. This would be calculated as 40,000/80 or 500 pF.

To have the same impedance tank coil at 40 MHz, one needs a tenth as much wire and a tenth as much capacitance. The length-to-diameter ratio of the coil wire probably will increase, so allowance for that might be necessary. The result would be about one foot of wire, wound as two or three turns on a much smaller form, and a tuning capacitor of 50 pF.

If this circuit is operated at 1600 V peak, the peak coil current will be 20 A, so something like #8 AWG wire is necessary. The rms current would be 14.14 A, the rms reactive power in the inductor would be about 16,000 VA — 50% of peak voltage times peak current.

To make a 1.0-to-0.9 step-down, or a 1.0-to-1.11 step-up autotransformer, just make a tap down 10% on the winding and connect it appropriately for step-down or step-up. Or, make a bifilar winding on slightly more than 10% of the coil and connect it to step up or step down. You can do the same thing at the ground end to have a 10:1 or 1:10 narrow-band transformer.

In the way described in this article, handbooks and formulas and charts may be dispensed with, and the same result obtained, as has hopefully been shown to both the practical and the theoretical-minded enthusiast.

Rosser B. Melton, Jr., AD5MI, became interested in radio after getting a crystal radio kit at age 10. He was licensed as novice in 1954 as WN5FEH. He operated mostly CW and earned he Amateur Extra class license in 1958. Throughout high school Rosser constructed various ham equipment. He received his Sc.D degree in Mechanical Engineering from M.I.T. in 1969. He was a teaching assistant, later an instructor, during his graduate years. After graduation, he worked at a research institution, and became inventor on seven patents. For a short while, he operated and maintained a 100 kW 5 MHz shortwave radio broadcasting station. Rosser is active on 80 to 20 meters CW, and also participates in the Sunday morning "Texas Kilowatt Net" on 3962 kHz SSB.

Notes

[1] *The ARRL Handbook Book,* 2016 Edition. ARRL item no. 0413, available from your ARRL dealer, or from the ARRL Store, Telephone toll-free in the US 888-277-5289, or 860-594-0355, fax 860-594-0303; **www.arrl.org/shop/**; **pubsales@arrl.org**.

Letters to the Editor

6 m Monoband Conversion for Heathkit SB-1000 Amplifier (Mar/Apr 2016)

Dear Editor,

There is a simpler way to deal with the tuning and loading capacitors for the conversion of an SP-1000 to 6 meters, if someone didn't want to replace them. Simply place an appropriately sized fixed capacitor in series with them. It even would accomplish the objective of increasing the breakdown voltage (mentioned in the article) because the voltage would be divided between the two series capacitors. The fixed capacitor would have to be rated for sufficient voltage and current handling. A silver mica or a doorknob capacitor would be a good choice. But it would still be much simpler and cheaper than replacing the variable capacitors. — *Wilton Helm, WT6C, Evergreen CO.*

[The author replies]

That seems unconventional but may have some merit. Although the cost of my plate capacitor was only US$50, and I would suspect you would spend more on many doorknob capacitors. This may also result on limited tuning range and not be practical. I would love to see someone experiment on this. — *Ron Berry, WB3LHD.*

Geodetic and Maidenhead Locater System Conversion (May/June 2016)

Dear Editor,

The Maidenhead locator system has been explicitly based on the WGS 84 geodetic datum since 1999, as set by IARU at Lillehammer. No grid locator is valid unless the correct geodetic datum is used. — *Kindly, E-P Mänd, OH2NFI, member ARRL and SRAL (Finland).*

Using a Wide-band Noise Generator with a Spectrum Analyzer (May/June 2016)

Dear Editor,

There is always a space between the number (value) and the following SI letter symbol. Concerning the article, on pages 25 and 26, "290K" should be "290 K". — *73, Lawrence Joy, 9V1MI/WN8P, ARRL LM and MI Section TS.*

Introducing AACTOR: A New Digital Mode (Jan/Feb 2016)

Dear Editor,

In Letters to the Editor of Jul/Aug *QEX*, author Mr. Roby does not seem to have defended portions of his original article fully. Mr. Phillips states that,

*"The entire concept of a binary number using a decimal point in it's representation is invalid, a detail that Mr. Roby himself asserts by stating "By using binary notation, and by disregarding the 0 to the left of the decimal point as well as the decimal point itself **f** simply becomes a stream of binary 1's and 0's." ... For over fifty years, the common practice for representation of fractional numbers in binary has been floating point notation. Not only is it's use common, it is the standard notation used by all computer platforms for fractional arithmetic operations. Throughout this time, computer scientists, programmers, engineers and academicians have understood and used floating point notation with absolute accuracy and confidence."*

There are a number of errors here. To begin, the concept of a binary number with a radix is indeed valid and well known, see **en.wikipedia.org/wiki/Radix_point**. Digits to the left of the radix represent whole number multipliers of the base to increasing positive integer powers, while the numbers to the right are increasing negative powers. For our binary example, to the left we have digit values 1, 2, 4, 8... corresponding to the 0, 1, 2, 3... powers of 2. To the right we have 0.5, 0.25, 0.125... corresponding to the -1, -2, -3... powers of 2.

Next, Mr. Phillips seems to imply that one cannot do anything useful with fixed-point binary fractional numbers. Embedded microcontroller systems and more particularly Digital Signal Processing (DSP) systems have used fixed-point fractional numeric representation for decades, with only recent migration to predominant use of floating point as the cost of the floating point hardware has come down. I personally have developed products on Analog Devices ADSP2100 family DSPs where all of the math is done in fractional fixed-point representation. Virtually every dial-up modem ever sold uses a fixed-point DSP and binary fractional math.

There are cases where fixed point is preferred, as for very low cost systems where floating point hardware cannot be justified. Microcontrollers using the ARM cortex M4 include a full fixed-point fractional math DSP instruction set. ARM provides a large library of DSP fixed-point fractional based functions called CMSIS-DSP. Fixed-point DSPs using fixed-point fractional math are widely available and widely used.

As such Mr. Phillips' statement that floating point *"is the standard notation for all computer platforms"* is simply not true. Fixed-point fractional math is entirely valid, quite useful and commonly found in many applications.

My only gripe with the original article is the premise of applying data compression schemes to a message, then sending that compressed content over RTTY, a rather inefficient modulation scheme. It would seem much more appropriate to pair this compression technique with a more modern narrow-band complex modulation such as PSK or similar. — *David Lundquist, N2DJE, Stony Brook, NY.*

QEX Availability Online? (June 19, 2016)

Dear Editor,

I recently noticed that I hadn't received my Issue #295 of *QEX*. I thought to remedy this by accessing the *QEX* library at ARRL. However, as best I can see, *QEX* is not available to subscribers online.

I thought this strange, because I can read *QST* online. After all, all of the ARRL publications are bundled on an annual DVD (which I buy and will continue to buy), so it can't be any extra effort to package each issue into a PDF. It would seem to me that, as each *QEX* goes to press, there would exist a PDF (or other suitable format) electronic file ready to be archived for subscriber online access. I would accept dropping paper delivery of both *QST* and *QEX*. I value *QEX* far more than *QST*, if that's a factor.

Indeed, online may be the only future for *QEX*. For now, I wonder why *QEX* is not available online? I am a long-time subscriber to *Nuts & Volts*, and they manage a subscribers-only online archive of issues going back 10 years or so, using a password-protected library. Why couldn't this work for *QEX*?

Subscriber Services promptly mailed me a copy of Issue #295. I would rather see the costs represented by the time and effort expended providing me with a second copy be directed to more useful *QEX* functions. — *Regards, Ed Price, WB6WSN, Chula Vista, CA.*

[*QEX* availability online has been and continues to be a source of debate at ARRL. I wonder what other subscribers to *QEX* think on this matter? — *Ed.]*

Send your *QEX* Letters to the Editor to, ARRL, 225 Main St., Newington, CT 06111, or by fax at 860-594-0259, or via e-mail to qex@arrl.org. We reserve the right to edit your letter for clarity, and to fit in the available page space. "Letters to the Editor" may also appear in other ARRL media. The publishers of *QEX* assume no responsibilities for statements made by correspondents.

Upcoming Conferences

The 35th Annual ARRL and TAPR Digital Communications Conference

September 16-18, 2016, St Petersburg, FL

www.tapr.org

Mark your calendar and start making plans to attend the premier technical conference of the year, the 35th Annual ARRL and TAPR Digital Communications Conference to be held September 16-18, 2016, in St Petersburg, FL. The conference location is the Hilton St Petersburg Bayfront.

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

Topics include, but are not limited to: Software Defined Radio (SDR), digital voice (D-Star, P25, Mototrbo, CODEC2, FreeDV), digital satellite communications, Global Position System (GPS), precision timing, Automatic Packet Reporting System™ (APRS), short messaging (a mode of APRS), Digital Signal Processing (DSP), HF digital modes, Internet interoperability with Amateur Radio networks, spread spectrum, IEEE 802.11 and other Part 15 license-exempt systems adaptable for Amateur Radio, using TCP/IP networking over Amateur Radio, mesh and peer to peer wireless networking, emergency and Homeland Defense backup digital communications, using Linux in Amateur Radio, updates on AX.25 and other wireless networking protocols.

Hotel

Conference presentations, meetings, and seminars will be held at the Hilton St Petersburg Bayfront. It is highly recommended that you book your room prior to arriving. A special DCC room rate of $109.00 single/double has been negotiated, and is until August 25, 2016; after that you will pay the regular room rate. Come early, stay late. The conference rate is good for 3 days before and 3 days after DCC.

To book your room, use the reservation link below, or call the hotel and mention the group code DCC when making reservations.

www.hilton.com/en/hi/groups/personalized/S/SPTSHHF-TAPR-20160914/index.jhtml.

Hilton St Petersburg Bayfront, 333 1st St S, St Petersburg, FL 33701; 1-800-HILTONS (1-800-445-8667) National reservation line; 1-800-944-5500 Hotel Direct.

Microwave Update 2016

October 13-15, 2016, St Louis, Missouri

www.microwaveupdate.org

St Louis, Missouri – Gateway to the West – is proud to host MUD 2016 on October 13-15, 2016 at Holiday Inn- Earth City, conveniently located minutes away from St Louis International Airport along I-70. In addition to technical presentations on Friday and Saturday, there will be preamp noise figure and antenna gain measurements, as well as, a "rover/dish row," the traditional MUD auction and banquet. Microwave Update is the only conference in the US devoted to amateur microwave bands starting at 900 MHz. Anyone with an interest in Amateur Radio microwave technology is invited to attend.

Conference fee, including a copy of the Proceedings is $75 per person, on or after September 15, 2016. Early Bird Special is $99 if you register before September 15, 2016. Fee includes registration, Proceedings, and Friday and Saturday lunch. See website for selections.

A special rate of $99.00 (plus tax) has been negotiated for the Holiday Inn Airport West, 3400 Rider Trail South, Earth City, MO 63045. The cut-off date for this rate is Thursday, September 22, 2016. Direct line is 314-475-3808, mention Block Code MWP for the special rate. There is also a reservation link available on the Microwave Update website.

For more information, go to **www.microwaveupdate.org** or send an email to **mud2016.info@gmail.com.**

Call for Papers

The Microwave Update 2016 program committee is calling for presentations and papers on the technical and operational aspects of amateur radio microwave communications. The deadline for proceedings paper submissions is September 1. (The deadline for the presentation version of selected papers is September 15, at the present time.)

Suggested topics include: antenna and dish feed design and construction; antenna modeling and testing methods; attracting "new blood" to the microwave bands; beacon design, construction and operation; best operating practices; broadband & mesh networks; commercial and surplus microwave components – adapting and using them; demonstration projects for schools, non-microwave Hams; designing for mobile and portable microwave operation; effectiveness of various digital weak-signal modes at microwave frequencies; EME station design and operation; homebrew construction methods and tips; microwave-band repeaters – voice, video, data; propagation characteristics and paths; use of test equipment to test, optimize and troubleshoot equipment; and weatherproofing and or ruggedizing outdoor fixed and portable installations. Format guidelines can be found at **www.ullmann.us/MUD2016/papers.htm.**

Please e-mail your papers and direct any questions or comments regarding the conference technical program to: **mud2016.papers@gmail.com**