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# QEX

January/February 2018

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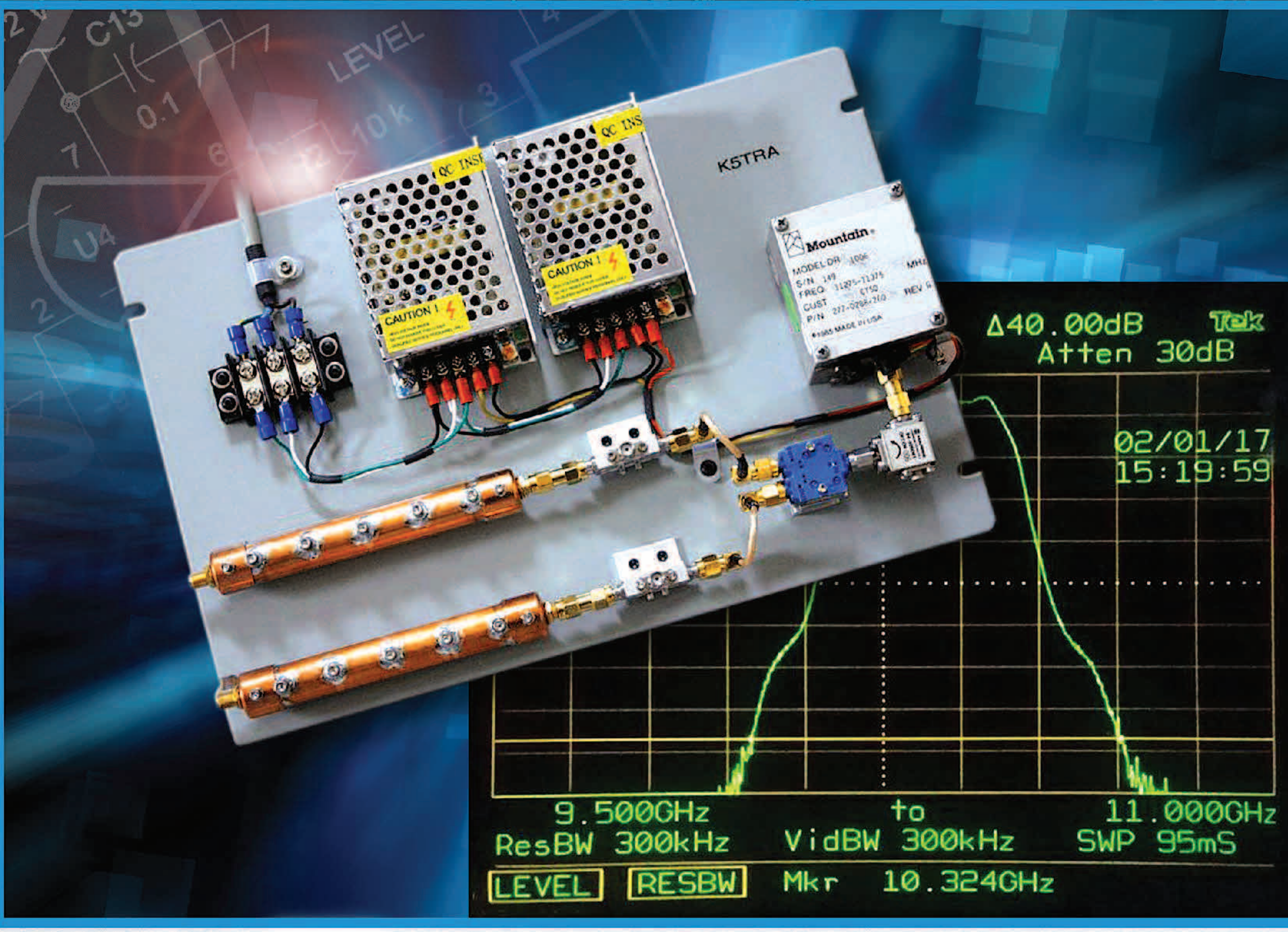
## A Forum for Communications Experimenters

Issue No. 306




**K5TRA** designs evanescent mode circular waveguide filters for his 10 GHz amateur band test set.

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Kazimierz "Kai" Siwiak, KE4PT
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**Advertising Information Contact:**

Janet L. Rocco, W1JLR
*Business Services*
860-594-0203 – Direct
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**Circulation Department**

Cathy Stepina, *QEX Circulation*

**Offices**

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Fax: 860-594-0259 (24 hour direct line)
e-mail: **qex@arrl.org**

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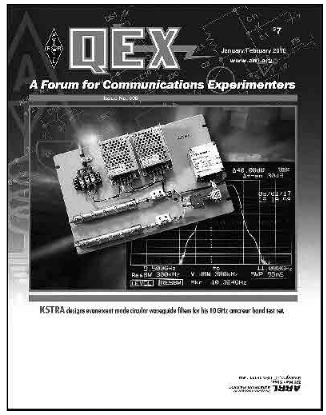
## About the Cover

Tom Apel, K5TRA, describes the development of evanescent mode circular waveguide filters for operation in his 10 GHz amateur band synchronous up/down conversion test set. The filter design is base on a software design approach that handles filter bandwidths up to 5%.




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

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**Kazimierz “Kai” Siwiak, KE4PT**

# Perspectives

## Constraints

It should not surprise you Dear Reader, that your Editor does not write the all of the articles that appear on these pages. You do, or at least a fraction of your colleagues do! The Editor and a capable Support Staff quite simply edit the content. That means we tend to the technical accuracy of articles, and we set the text, equations and figures to the *QEX* template. It is not unlike the bi-monthly tile-matching puzzle video game Tetris, with game over on the publishing deadline. Articles, ads, and announcements must fit within the publisher’s page constraints. Page count must be a multiple of 4 pages, or better yet, a multiple of 8 pages.

I bring this up because we’ve received requests for *limiting* the technical content and for *expanding* technical content, as well as for *covering* and *avoiding* some topics, and for *limiting the length* articles. In an ideal world we would conjure up a perfect balance of content. In reality the supply of articles is quite limited. In May 2017 we added a *Technical Notes* column to help fill the queue with short-length technical articles and observations that would be of interest to the communications experimenters. We’ve run several, and several more are queued up.

You, Dear Reader/Author, not the Editor, are in charge of the *QEX* content, by submitting articles and short technical exchanges. Review the handiwork of your colleagues in the 2017 Annual Index, herein, and please, join them with your contributed article.

## In This Issue

Our *QEX* authors contributed to the communication’s experimental arts in a wide variety of Amateur Radio topics.

George R. Steber, WB9LVI, uses variable capacitance diodes to replace expensive mechanically adjustable capacitors in an RF project.

Dr. Sam Green, WØPCE, relates the technical challenges in modifying coupler and higher frequency logarithmic detectors to extend frequency range of his SWR meter up to 4 GHz.

Michelle Thompson, W5NYV, and Howie DeFelice, AB2S, discuss the challenges of uncertain launch schedules, development schedules and justification for the projects that face next generation AMSAT satellites.

Tom Apel, K5TRA, applies his designs of evanescent mode circular waveguide 10 GHz filters with bandwidths up to 5% t his microwave test set.

Rick Campbell, KK7B, designs square four-element antennas for VHF and UHF.

Keep the full-length *QEX* articles flowing in, but if brevity is your forte, share a brief **Technical Note** of perhaps several hundred words in length plus a figure or two. Expand on another author’s work and add to the Amateur Radio *institutional memory* with your technical observation. Let us know that your submission is intended as a **Note**.

*QEX* is edited by Kazimierz “Kai” Siwiak, KE4PT, (**ksiwiak@arrl.org**) and is published bimonthly. *QEX* is a forum for the free exchange of ideas among communications experimenters. The content is driven by you, the reader and prospective author. The subscription rate (6 issues per year) in the United States is $29. First Class delivery in the US is available at an annual rate of $40. For international subscribers, including those in Canada and Mexico, *QEX* can be delivered by airmail for $35 annually. Subscribe today at **www.arrl.org/qex**.

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Very best regards,

Kazimierz “Kai” Siwiak, KE4PT

**Tom Apel, K5TRA**

7221 Covered Bridge Dr., Austin, TX 78736; **tom@k5tra.net**

# Evanescent Mode Circular Waveguide 10 GHz Filters

*Filters for operation in the 10 GHz ham band are based on a software design approach that handles filter bandwidths up to 5%.*

[A version of this article appeared in *Microwave Product Digest* June 2017.]

Motivation for this project began when I decided to extend my ham radio operation to 10 GHz. I have previously[1] published the LNA design details that was part of the station. Before any hardware could be tested, I needed to extend my measurement capability to 10 GHz. I have a vector network analyzer (VNA), spectrum analyzer and service monitor that provide metrology through L-band, but nothing supporting 10 GHz measurement capability. I decided to build a synchronous up/down conversion test-set to extend my bench equipment to 10 GHz. Figures 1 and 2 illustrate the block diagram and photo of that test-set. This enabled measurement of insertion gain or loss as well as return loss with a directional coupler. I needed image reject filters to build this test-set. A pair of the filters can be seen in the photo. This article presents the development of those filters. I used AWR Microwave Office software to design filters for a range of bandwidths up to 5%. The ability to optimize the design, while simultaneously watching resonator spacing, insertion loss and return loss was powerful.

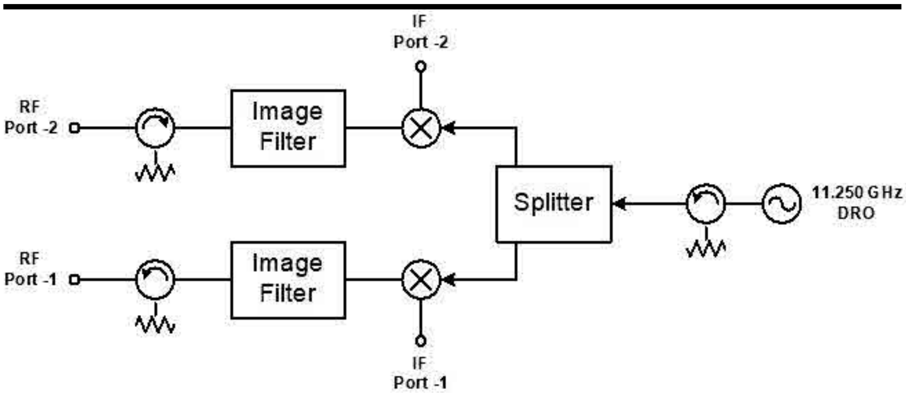


**Figure 1 — Test set block diagram.**

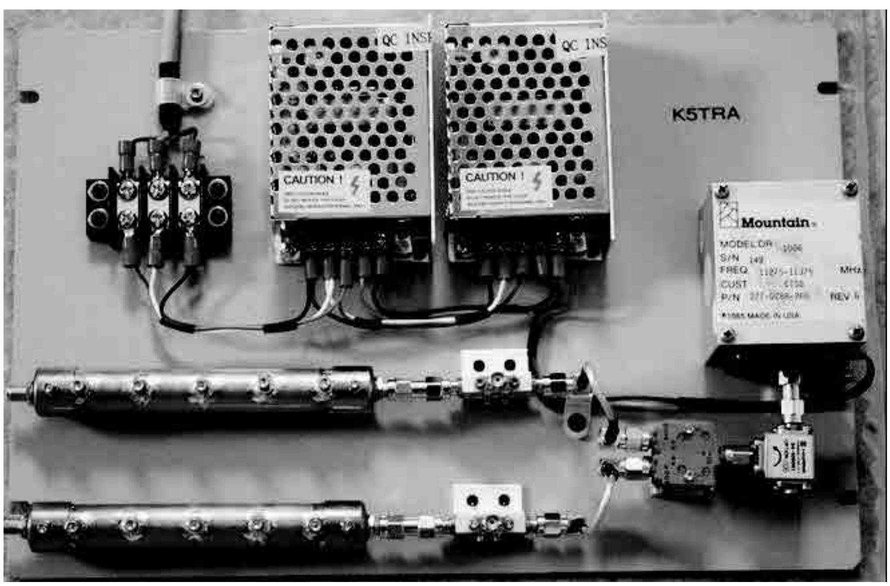


**Figure 2 — 10 GHz test set. [Tom Apel, K5TRA, photo.]**

## Band-pass Filters and Immittance Inverters

Half-inch copper tubing segments with end caps were used to realize tunable band-pass filters at 10 GHz. A bit of basic filter theory is necessary to understand how this type of filter works.

Consider first, the simple two branch

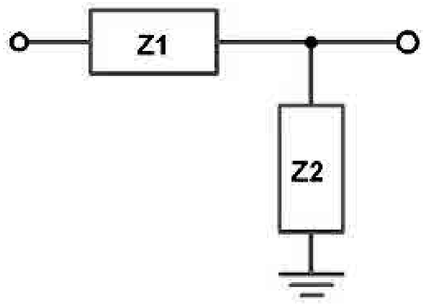


Figure 3 — Basic ladder building block.

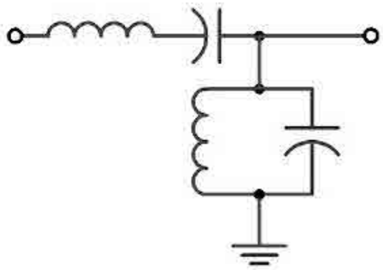

Figure 4 — Band-pass canonic ladder block.

ladder circuit shown in Figure 3. Many filters can be represented as cascades of series and shunt branches. In the passband frequency range, series branches are ideally short-circuits and shunt branches are open. This yields 0 dB insertion loss. Alternatively, in the stop band, series branches are ideally open and shunt branches are short-circuits, to yield no transmission and 100% reflection. For example, series inductor and shunt capacitor cascades form low-pass (LP) filters. Band-pass (BP) ladder canonic structures are formed from ladder cascades of series resonators and shunt resonators. This can be seen in Figure 4. Band-pass filters obtained from LP to BP transformation[2, 3] produce these structures with all resonators synchronously tuned to the center of the pass band.

You may be thinking that LC ladders comprised of alternating series and shunt resonators is quite far removed from waveguide filters operating at 10 GHz. Two conceptual steps are necessary to bridge the gap. The first, is the introduction of impedance (K) or admittance (J) inverters. Both filter structures in Figure 5 can be viewed as equivalent to the series-shunt LC cascade of Figure 4. An impedance inverter connected to a series resonator behaves as a shunt resonator when viewed from the opposite side of the inverter. Similarly, an admittance inverter connected to a shunt resonator behaves as a series resonator when viewed from the opposite side of the inverter. Admittance or impedance inverters (sometimes called immittance inverters) are symmetric networks where shorts map to opens and opens to shorts, series capacitors map to shunt inductors and series inductors map to shunt capacitors.

The use of immittance inverters allows reuse of the same type of resonator in an overall band-pass filter design, as illustrated in Figure 6. For example, comb-line filters can be viewed as cascades of shunt resonators and admittance inverters, where the inter-resonator coupling is set by the particular admittance of the inverter between a pair of resonators. Filters with the same input and output port impedances are usually symmetric. This is shown in Figure 6. We will make use of this property to simplify the design parameters.

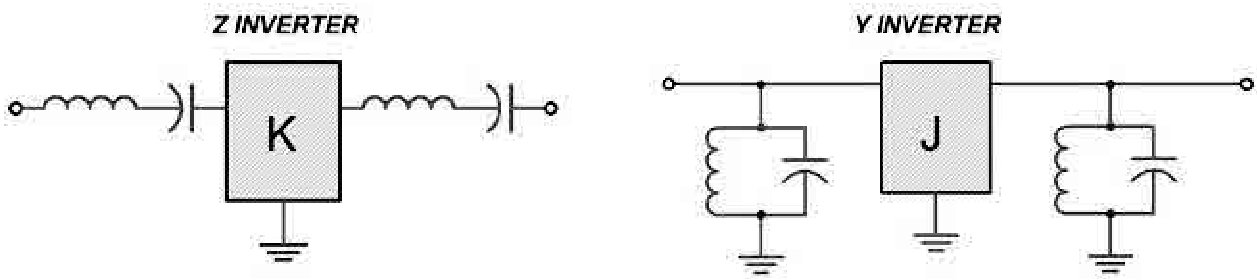


Figure 5 — K and J inverters in band-pass filter segments.

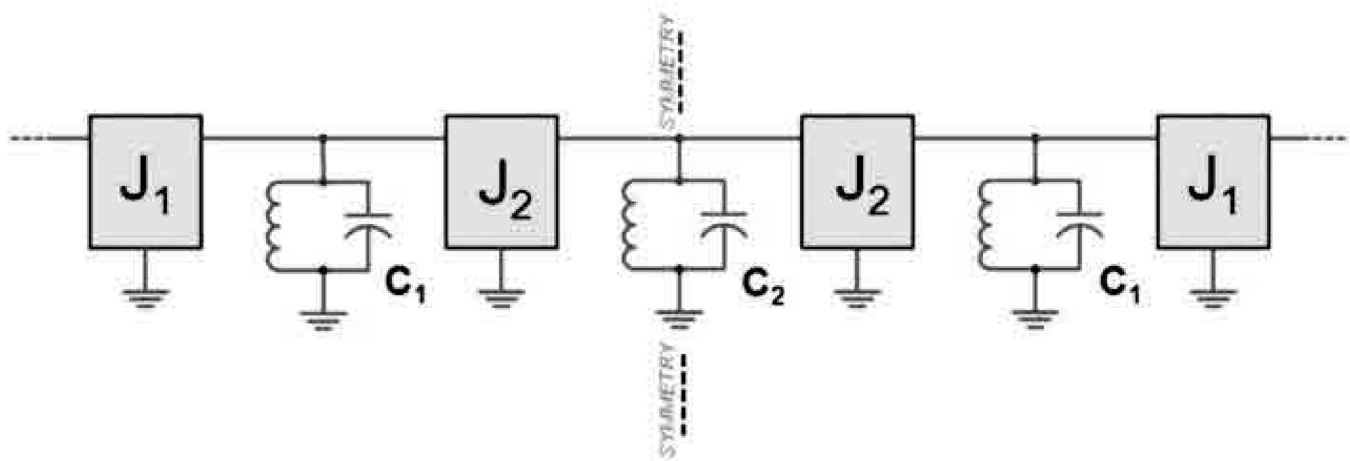


Figure 6 — Three resonator band-pass filters with J inverters and shunt resonators.

J and K inverters can be realized in many ways. Quarter-wave transmission lines are a common example of impedance (or admittance) transformer placed between similar resonators. Inverter forms commonly used in LC filter designs are shown in Figure 7, where the T networks (upper and lower left) are the impedance inverter representations and the π networks (upper and lower right) are the admittance forms. Note that a T to π transformation easily shows that they are equivalent. The T inverter network forms are better suited to interface with series resonators and the π representation is better suited to interface with shunt resonators. The obvious question now arises. Where does one find negative valued inductors or capacitors?

From Figure 6, it is clear that inverters will be placed between shunt resonators. Consider the left side of Figure 8. A pair of shunt resonators are coupled with an inductive admittance inverter. The inverter negative shunt inductance can be absorbed into the resonator representation. This results in a parallel equivalent inductance larger than the original. This is represented as $L''$ in the equivalent circuit on the right side of Figure 8.

You might be inclined to think that the resonant frequency has been lowered. It is important to note that when the adjacent resonator is short circuited, the resonant frequency is the same as the original resonator. When one side of an inverter is shorted, the other side is open. This is also true for interior resonators with inverters on both sides, as in the center resonator in Figure 6. When the adjacent resonators on both sides are short circuited, the interior resonator tunes the center frequency of the passband. In this way, all resonators are seen to be tuned to the center frequency. Filters of this type are called synchronously tuned BP filters. The relationship is a direct result of canonic BP ladder structures from Figure 4.

Recognition of the resonant frequency behavior of each resonator when adjacent resonators are shorted will be important later in greatly simplifying the design process. It is worth mentioning also that this behavior is the basis for Dishal's tuning method[4].

## Circular Waveguide

Next, we will consider the waveguide equivalent circuit. Metal pipes of various cross-section shapes have been used for many years as low loss transmission media at microwave frequencies. Circular cross-section copper pipe is the medium of interest here. As long as the RF frequency is above a critical frequency related to the pipe radius, RF will propagate. A half-inch copper pipe is actually 0.565" inside diameter. The corresponding TE11 cutoff frequency is 12.2516 GHz. Below the cutoff frequency RF field amplitude falls exponentially with distance and the transmission path becomes reactive. This is called evanescent mode propagation.

The equivalent circuit for a length of waveguide below the cutoff frequency can be seen in Figure 9. The possibilities for using below-cutoff waveguide to form admittance inverters now becomes apparent. Consider the equivalent circuit redrawn in the left side of Figure 10. Since inductive admittance inverters with series branch inductance of $L_S$ also have $-L_S$ shunt branches, additional shunt branches of $+L_S$ and $-L_S$ can be added with no net change (they cancel each other). This trick allows us to recognize that the segment of copper pipe can be viewed as an inductive admittance inverter with shunt inductors on both ends. Tuning screws can be added in the E-plane to introduce shunt capacitance. The

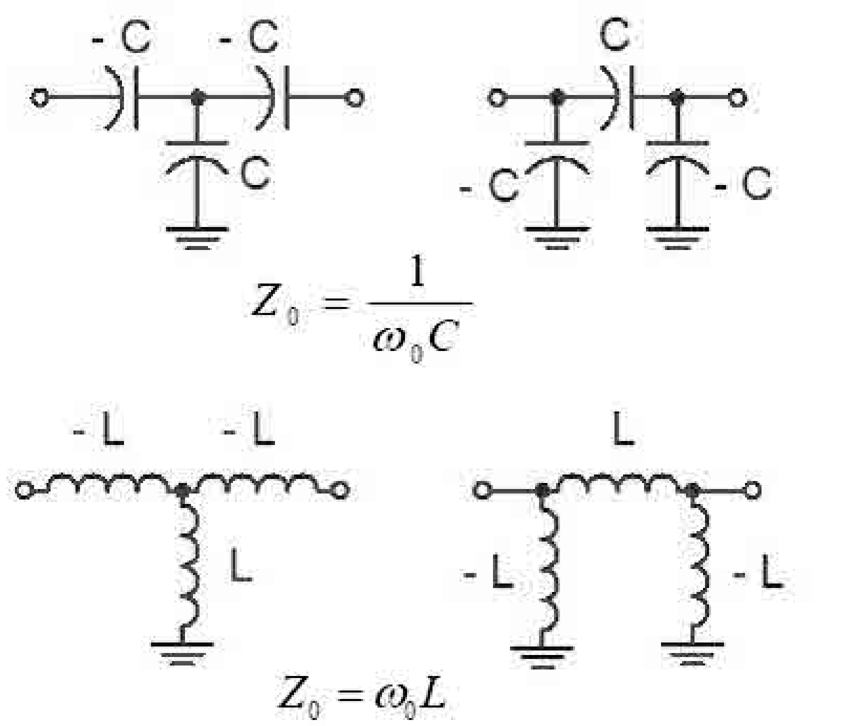


Figure 7 — LC realizations of K and J inverters.

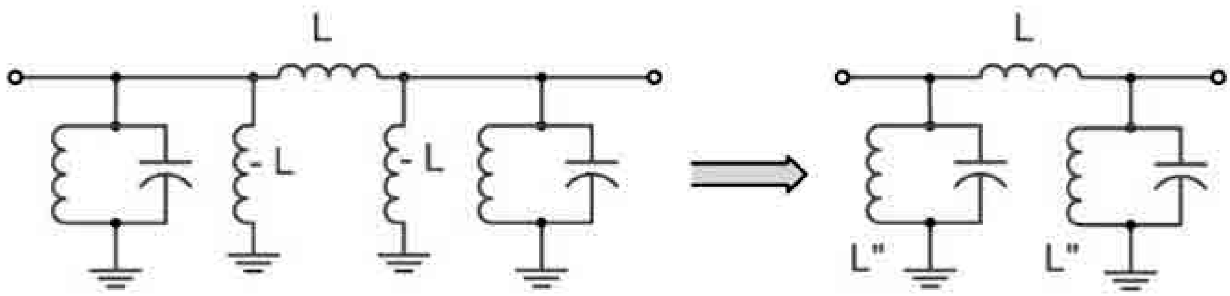


Figure 8 — Absorption of J inverter negative elements into resonator tuning.

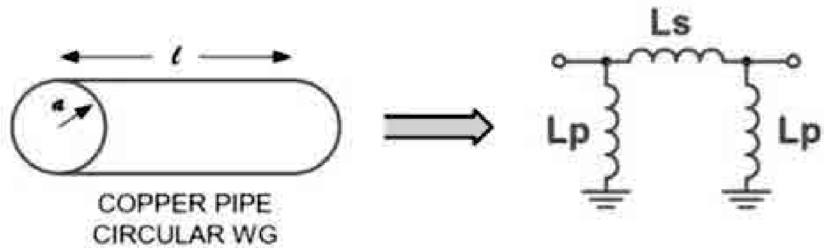


**Figure 9 — Below cutoff circular waveguide equivalent circuit.**

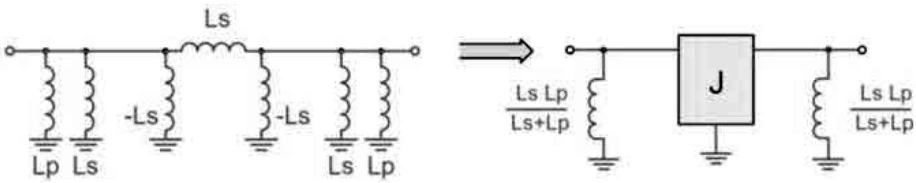


**Figure 10 — Below cutoff circular waveguide equivalent circuit, alternate interpretation.**

Zn=50 (port reference Z)
Qo=200

a=7.1755 (radius mm)
Fc=1.8412*300/6.2832/a
Fc: 12.25

Fo=10.4 (GHz)
b=(6.2832*Fo/300)*sqrt((Fc/Fo)^2 -1)
b: 0.1356

X0=377/sqrt((Fc/Fo)^2 -1)
X0: 605.4

| | |
|---|---|
| l1=5.47674781483291 | l2=21.9933981090275 |
| b: 0.1356 | b: 0.1356 |
| XLs1=X0*sinh(b*l1) | XLs2=X0*sinh(b*l2) |
| XLs1: 492.3 | XLs2: 5963 |
| Ls1=XLs1/6.28/Fo | Ls2=XLs2/6.28/Fo |
| Ls1: 7.537 | Ls2: 91.3 |
| XLp1=X0*cosh(b*l1*0.50)/sinh(b*l1*0.50) | XLp2=X0*cosh(b*l2*0.50)/sinh(b*l2*0.50) |
| XLp1: 1704 | XLp2: 670 |
| Lp1=XLp1/6.28/Fo | Lp2=XLp2/6.28/Fo |
| Lp1: 26.1 | Lp2: 10.26 |
| Lc1=1/(1/Ls1+1/Lp1+1/Ls2+1/Lp2) | Lc2=1/(2/Ls2+2/Lp2) |
| C1p=1000/(4269.9856*Lc1) | C2p=1000/(4269.9856*Lc2) |
| C1p: 0.06544 | C2p: 0.05079 |
| x1=l1/25.4 | x2=l2/25.4 |
| x1: 0.2156 | x2: 0.8659 |

**Figure 11 — Algebraic description in AWR. "l1" and "l2" are tune variables in mm; "*x*1" and "*x*2" are converted to inches.**

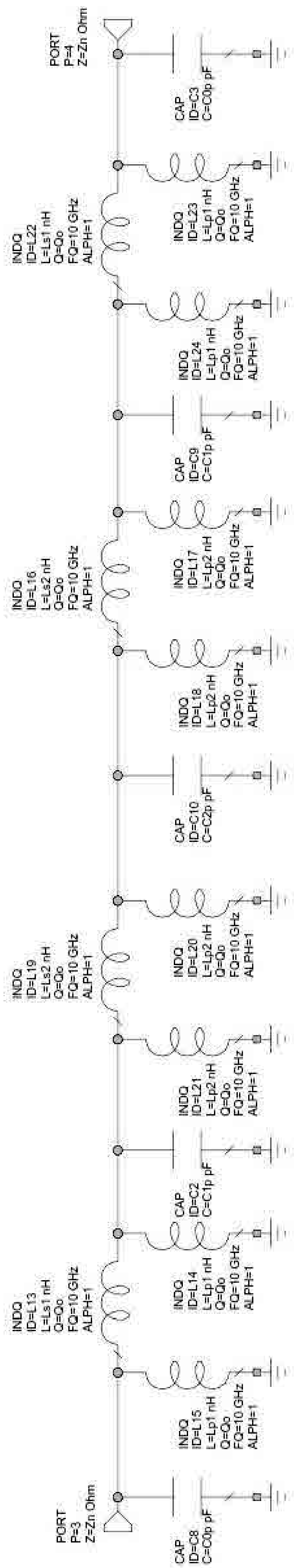

**Figure 12 — Three resonator filter with tuned port transition.**

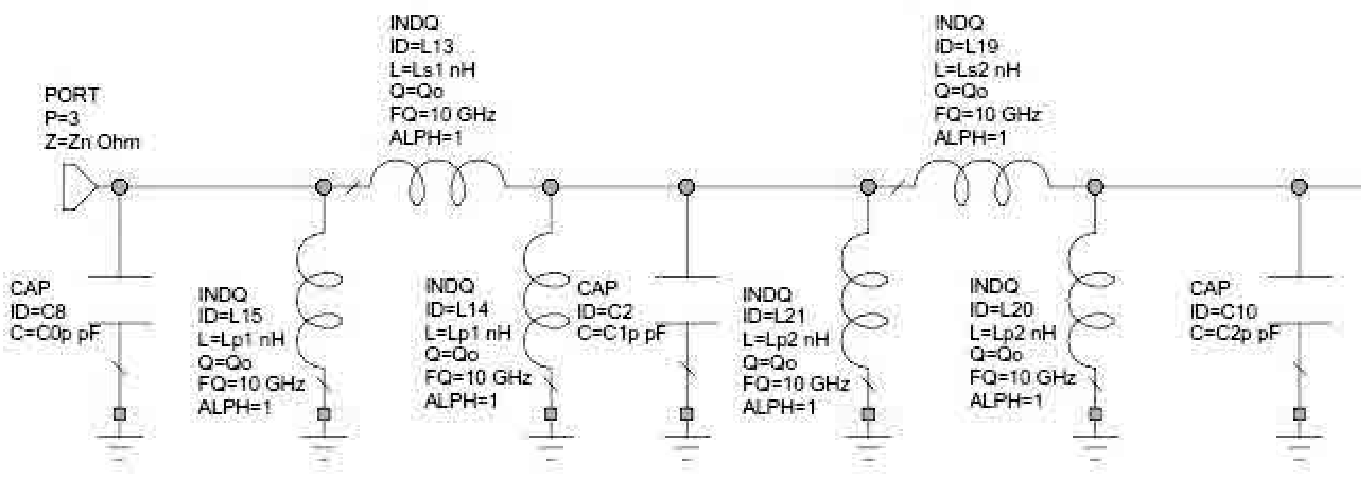


**Figure 13 — Half of the symmetric filter from Figure 11.**

shunt resonators are thereby created. For less inter-resonator coupling, the length of the pipe segment between tuning screws can be increased (LS is increased).

The following equations allow us to describe the circuit behavior of a segment of half-inch diameter copper pipe. The cutoff frequency, $F_c$, of a circular wave guide is a function of the wave guide inside radius, $a$. The measured inside diameter of a half-inch copper pipe actually is 0.565 inches, so, $a$=7.1755 mm. Thus,

Cutoff Frequency:

$$F_c = \frac{1.8412c}{2\pi a} = \frac{(1.8412)(299.79)}{2\pi(7.1755)} \approx 12.252 \text{ GHz}$$

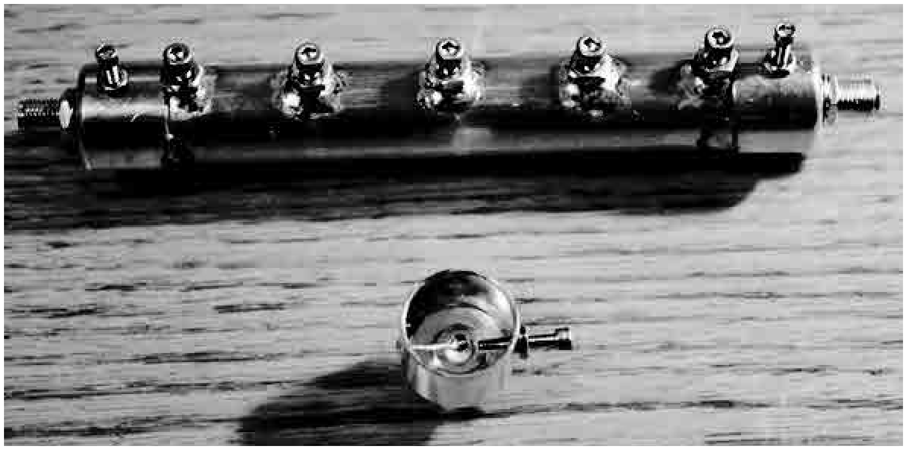

**Figure 14 — Five resonator band-pass filter with tuned H-plane launchers. [Tom Apel, K5TRA, photo.]**

Center Frequency:

$$F_0 = 10.4 \text{ GHz}$$

Evanescent mode propagation constant:

$$\gamma = \frac{\omega_0}{c}\sqrt{\frac{F_c}{F_0} - 1} \approx 0.1356$$

Wave impedance:

$$X_0 = \frac{120\pi}{\sqrt{\frac{F_c}{F_0} - 1}} \approx 605\ \Omega\,.$$

The equivalent circuit for a length, $l$, of half-inch copper pipe can be calculated as follows:

Series inductance:

$$L_s = \frac{X_0 \sinh(\gamma l)}{2\pi F_0}$$

Shunt inductance:

$$L_p = \frac{X_0 \coth\left(\frac{\gamma l}{2}\right)}{2\pi F_0}\,.$$

These are numeric values for the circuit representations found in Figures 9 and 10. Clearly, by changing the length, one can set the series inductance and thereby the coupling (admittance of inverter). The excess shunt inductance of Figure 10,

$$\frac{L_S L_p}{L_S + L_p}$$

can be resonated by adding a tuning screw in the E-plane. This provides a variable capacitance to form the shunt resonator.

## Design Procedure

The preceding paragraphs provide a procedure for analyzing a length of circular waveguide operating below cutoff. To design a filter in this media, one could start from LP prototype tables, perform a LP to BP transformation and map the series resonators (see Figure 4) into shunt resonators cascaded with inverters. This is the classic approach found in Craven[5] and in Howard[6]. This often involves some iterative optimization.

Another approach might be to fully describe a physical filter with inter-resonator lengths and tuning capacitors as variables in a circuit analysis environment and simply use optimization to obtain a solution. This is a bit 'brute force' in approach and might not yield an optimum solution.

If some knowledge of desirable filter solutions is used to simplify and constrain the structure, circuit analysis with optimization can efficiently obtain a solution that is well conditioned to converge to a desired solution. Recall the symmetry shown in Figure 6.

If the physical structure and thereby the circuit description is explicitly defined to be symmetrical, then the number of variables is cut in half. Also, recall the mention of Dishal's tuning method for synchronously tuned filters. Each resonator should be tuned to the center frequency when all adjacent resonators are short circuited. This constraint provides a closed form solution for all tuning capacitors as a function of surrounding inductors. This eliminates the capacitors as variables and provides a significant advantage in preconditioning the optimizer for a desired solution. This is the approach used in this project.

The AWR Microwave Office suite of circuit analysis and optimization was used. Circuit values were all described algebraically within the AWR environment. This feature enables easy implementation of symmetry and synchronous tuning variable elimination. The only independent variables were the distances between tuning screws. These are "i1" and "i2" in the example

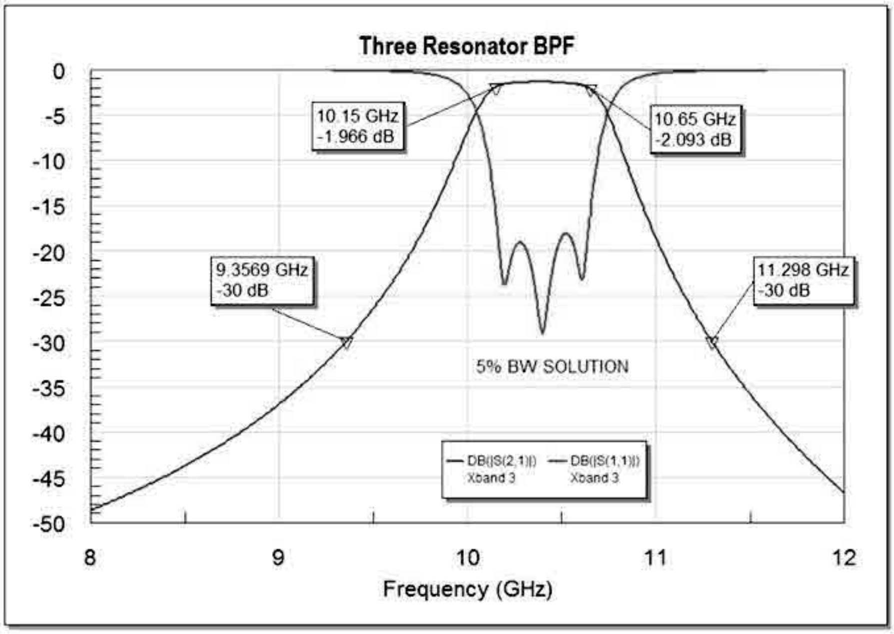


**Figure 15 — Optimized response as displayed within AWR design environment.**

N=6

| %BW | BW(MHz) | l1(mil) | l2(mil) | l3(mil) | l4(mil) | C1(pF) | C2(pF) | C3(pF) | Loss(dB) | RetLoss(dB) | F30L(GHz) | F30H(GHz) |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 5 | 500 | 218 | 887 | 1020 | 1038 | 0.06517 | 0.05069 | 0.05062 | 6.32 | 20.21 | 9.98 | 10.81 |
| 3 | 300 | 221 | 950 | 1116 | 1142 | 0.06470 | 0.05062 | 0.05057 | 6.72 | 29.45 | 10.10 | 10.70 |
| 2 | 200 | 223 | 983 | 1172 | 1204 | 0.06441 | 0.05060 | 0.05060 | 6.89 | 39.20 | 10.15 | 10.65 |
| 1 | 100 | 235 | 1035 | 1245 | 1285 | 0.06302 | 0.05058 | 0.05054 | 7.68 | 48.89 | 10.20 | 10.59 |
| 0.5 | 50 | 279 | 1162 | 1385 | 1436 | 0.05919 | 0.05055 | 0.05053 | 11.83 | 48.44 | 10.28 | 10.51 |

N=5

| %BW | BW(MHz) | l1(mil) | l2(mil) | l3(mil) | C1(pF) | C2(pF) | C3(pF) | Loss(dB) | RetLoss(dB) | F30L(GHz) | F30H(GHz) |
|---|---|---|---|---|---|---|---|---|---|---|---|
| 5 | 500 | 171 | 793 | 961 | 0.07322 | ,05081 | 0.05088 | 3.10 | 27.41 | 9.77 | 10.99 |
| 3 | 300 | 218 | 940 | 1098 | 0.06514 | 0.05063 | 0.05058 | 4.63 | 28.00 | 10.10 | 10.77 |
| 2 | 200 | 256 | 1050 | 1209 | 0.06103 | 0.05058 | 0.05055 | 6.43 | 29.20 | 10.14 | 10.66 |
| 1 | 100 | 279 | 1151 | 1344 | 0.05922 | 0.05055 | 0.05054 | 8.64 | 30.03 | 10.23 | 10.57 |
| 0.5 | 50 | 300 | 1225 | 1452 | 0.05788 | 0.05054 | 0.05053 | 11.13 | 31.48 | 10.28 | 10.52 |

N=4

| %BW | BW(MHz) | l1(mil) | l2(mil) | l3(mil) | C1(pF) | C2(pF) | Loss(dB) | RetLoss(dB) | F30L(GHz) | F30H(GHz) |
|---|---|---|---|---|---|---|---|---|---|---|
| 5 | 500 | 194 | 841 | 966 | 0.06865 | 0.05075 | 2.66 | 21.25 | 9.66 | 11.07 |
| 3 | 300 | 219 | 938 | 1070 | 0.06491 | 0.05064 | 3.17 | 25.46 | 9.87 | 10.90 |
| 2 | 200 | 259 | 1051 | 1180 | 0.06077 | 0.05058 | 4.41 | 25.12 | 10.04 | 10.75 |
| 1 | 100 | 298 | 1173 | 1314 | 0.05797 | 0.05055 | 6.08 | 30.26 | 10.17 | 10.63 |
| 0.5 | 50 | 308 | 1215 | 1369 | 0.05743 | 0.05054 | 6.64 | 40.98 | 10.20 | 10.60 |

N=3

| %BW | BW(MHz) | l1(mil) | l2(mil) | C1(pF) | C2(pF) | Loss(dB) | RetLoss(dB) | F30L(GHz) | F30H(GHz) |
|---|---|---|---|---|---|---|---|---|---|
| 5 | 500 | 216 | 866 | 0.06500 | 0.05079 | 2.10 | 15.69 | 9.36 | 11.30 |
| 3 | 300 | 240 | 961 | 0.06259 | 0.05066 | 2.31 | 20.38 | 9.63 | 11.09 |
| 2 | 200 | 280 | 1072 | 0.05913 | 0.05059 | 3.13 | 20.51 | 9.87 | 10.89 |
| 1 | 100 | 320 | 1196 | 0.05680 | 0.05055 | 4.22 | 23.45 | 10.05 | 10.74 |
| 0.5 | 50 | 361 | 1310 | 0.05513 | 0.05054 | 5.80 | 25.83 | 10.16 | 10.63 |

N=2

| %BW | BW(MHz) | l1(mil) | l2(mil) | C1(pF) | Loss(dB) | RetLoss(dB) | F30L(GHz) | F30H(GHz) |
|---|---|---|---|---|---|---|---|---|
| 5 | 500 | 214 | 753 | 0.06576 | 1.22 | 13.26 | 7.33 | 12.26 |
| 3 | 300 | 253 | 880 | 0.06136 | 1.43 | 15.70 | 8.53 | 11.77 |
| 2 | 200 | 272 | 947 | 0.05974 | 1.54 | 18.72 | 8.90 | 11.56 |
| 1 | 100 | 319 | 1079 | 0.05686 | 2.10 | 21.34 | 9.44 | 11.21 |
| 0.5 | 50 | 352 | 1176 | 0.05545 | 2.69 | 23.01 | 9.69 | 11.02 |

**Figure 16 — Table of design solutions for a 10 GHz center frequency.**

AWR algebraic description in Figure 11. All inductors and capacitors are calculated in closed form from these two values.

Figure 12 shows a corresponding circuit description. This is a three resonator filter with tuned port transitions. It is a bit of an ‘eye chart’. Since it is constructed to be symmetric about the center, the details can be more easily seen in Figure 13.

Please note that loss has been introduced through the inductor unloaded Q parameter, $Q_0$. After measuring a single unloaded resonator, the unloaded Q was empirically determined to be approximately 200.

The SMA transition can be seen in the lower part of Figure 14. A shorted loop forms an H plane transition. This launcher is tuned with a shunt capacitance from a tuning screw. For the simulation, the port transition tuning was expressed as C0p. It was calculated using Dishal’s method[4]. This can be seen algebraically,

Lc0 = 1/(1/Ls1+1/Lp1)

C0p = 1000/(4269.9856*Lc0)

C0p: 0.04005

Practically, the end transition tuning will differ slightly due to the H plane transition inductance.

The optimized swept response for this example filter is illustrated in Figure 15. This is a 5% bandwidth solution centered on 10.4 GHz.

## Results

A set of solutions were run over bandwidth steps for 2, 3, 4, and 5 resonator filters. These results are tabulated in Figure 16. Each case is for a center frequency of 10.4 GHz. Tuning screw locations, passband loss and reflection and -30 dB frequency points are displayed.

Photos of the five-resonator filters that were used in the 10 GHz test set are shown in Figures 14 and 17. Brass nuts that have been soldered to the copper tube provide solid support and grounding for resonator tuning screws. Additional lock nuts are on each tuning screw. Resonator screws are 4-40 and SMA transition tuning screws are 2-56 thread. Mounting stand-offs are 4-40 and are attached to the filter body with ‘button head’ screws.

The measured response of fabricated filters is very good. Figure 18 shows the measured swept response from one of the image reject filters in the 10 GHz test set of Figure 2. The measured passband is 10.18 GHz to 10.48 GHz with nominal insertion loss of 2.9 dB and ripple of less than 1 dB peak to peak. This 300 MHz bandwidth was obtained from a 3% bandwidth design. The 40 dB bandwidth is seen in Figure 19 to be 720 MHz.

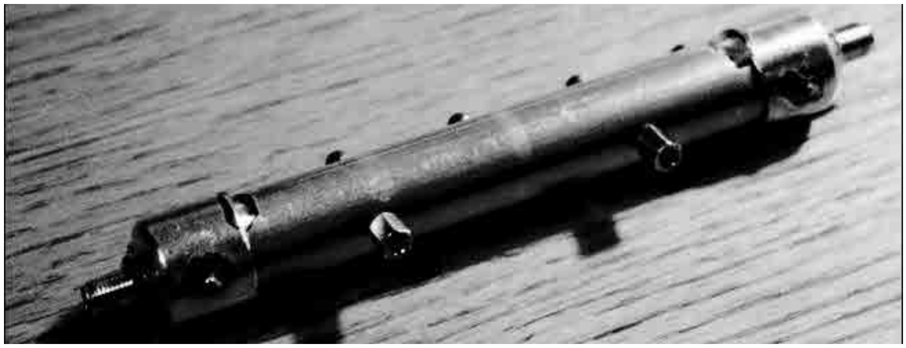

**Figure 17 — Stand-off mounting hardware. [Tom Apel, K5TRA, photo.]**

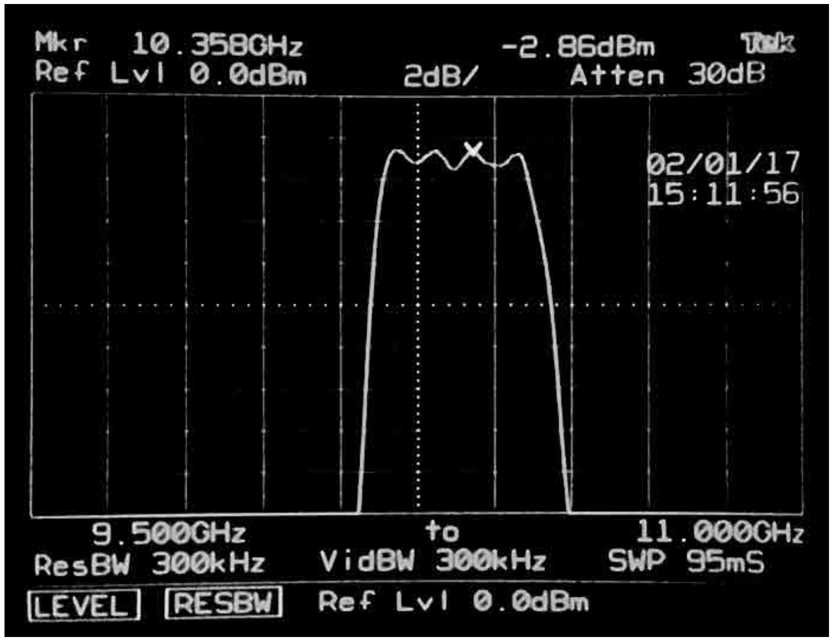


**Figure 18 — Measured response of 3% bandwidth filter. Vertical scale is 2 dB per division.**

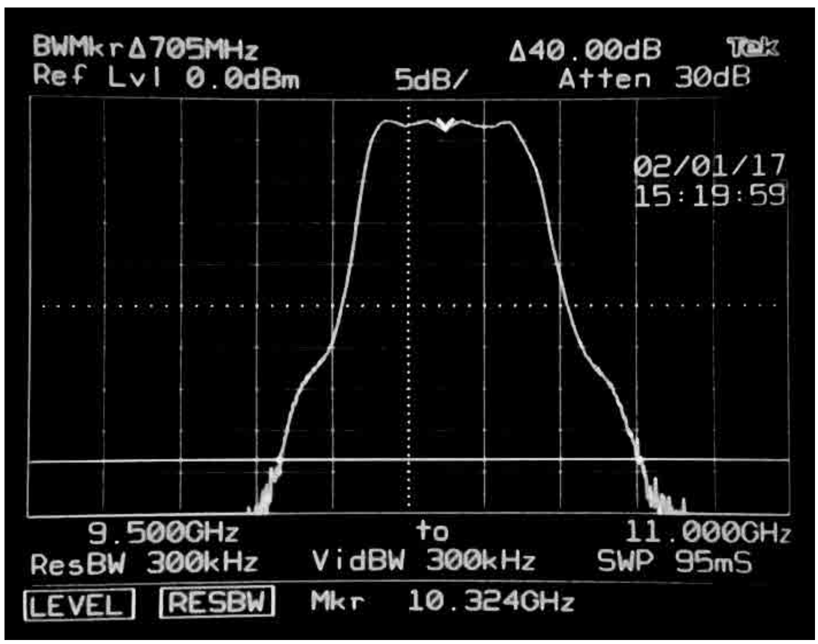


**Figure 19 — Measured 40 dB bandwidth response. Vertical scale is 5 dB per division.**

For more information on waveguides and filters search **k5tra.net/tech%20library.html**.

I again wish to acknowledge the AWR Microwave Office software that was used in the design of these filters.

*Tom Apel, K5TRA, is an electrical engineer, currently serving as president of the Roadrunners Microwave Group, an ARRL affiliated club. Tom chaired the technical program for the 2014 Central States VHF Society conference. In 2010 he retired from Triquint Semiconductor as Senior Engineering Fellow where he managed advanced component development. He has 33 years in microwave and RF component design at VHF through Ka band. He developed the first 6-18 GHz 2 W power amplifier MMIC to achieve volume production. More recently, his work has resulted in many power amplifier products for handset applications. During his career he was inventor on 35 US patents. Tom earned a BS Physics and BS Mathematics from Loras College, and MSEE from University of Wisconsin, Madison. Tom was first licensed in 1963 and has been home building since then.*

## Notes

[1]T. Apel, "Design of a10 GHz LNA for Amateur Radio Operation", *Microwave Product Digest*, Feb 2016, p. 8.

[2]G. Matthaei, L. Young and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*, McGraw-Hill, New York 1964, Sections 9.02 and 8.03, pp. 427-438.

[3]A. Zverev, *Handbook of Filter Synthesis*, Wiley, NY, 1967, Chapter 6.

[4]M. Dishal, "Alignment and Adjustment of Synchronously Tuned Multiple Resonant-Circuit Filters", *Proc IRE* 39, Nov 1951, pp.1448-1455.

[5]G. Craven and C. K. Mok, "The Design of Evanescent Mode Waveguide Bandpass Prescribed Insertion Loss Characteristic", IEEE MTT-19, Mar 1971 pp.295-308.

[6]J. Howard and M. Lavey, "Simplified Method Eases the Design of Bandpass Filters", *RF and Microwaves*, Dec 2000 p. 97.

[7]S. Mandev and R. Kwok, "Evanescent Mode Filter", SJSU EE172 Final Project, May 2013.

[8]P. Wade, "Waveguide Filters You Can Build and Tune, Part 3 Evanescent Mode Waveguide Filters", *QEX*, Mar 2011, pp. 23-29.

[9]C. Balanis, "Circular Waveguides", *Encyclopedia of RF and Microwave*, Wiley 2005, pp.641-653.

**George R. Steber, WB9LVI**

11105 N. Riverland Ct., Mequon, WI 53092; **steber@execpc.com**

# A Tunable RF Preamplifier Using a Variable Capacitance Diode

*Variable capacitance diodes can replace expensive mechanically adjustable capacitors in an RF project.*

Substituting a variable capacitance diode for an expensive mechanically adjustable capacitor can be done in many RF projects. But there are a few tricks that need to be learned. Take a look at how it's done with this versatile RF preamplifier that you can build yourself.

Working with home brew ham receivers, inexpensive commercial short-wave sets and software defined receiver (SDR) based "dongle" radios can be fun and challenging. In some cases, however, the benefits of good front-end RF selectivity may have been overlooked. For these cases adding a RF filter between the antenna and receiver as shown in Figure 1 can help. Filtering signals from the antenna provides rejection of strong out of band signals, which otherwise might overload the input. The filter can also prove to be highly beneficial in situations where there is a lot of noise or interference.

If your application requires just a single frequency or narrow band of frequencies then a fixed input filter can be used. But if your application requires a number of bands, as in short wave listening, you will need an input filter that is tunable to the desired band of interest.

Presented here is an easy to build tunable preamplifier that should find use in many interference situations. It has good performance and can be built for a fraction of the cost of a commercial unit. An interesting aspect of this project is that it uses a variable capacitance diode (VCD), sometimes referred to as a varicap or varactor in the literature.

The design offered here covers the frequency range of 6 MHz to 23 MHz. But we'll show you how to change some component values to cover the frequency band of your own interest. To ease construction, through-hole components are used exclusively — no surface mount parts. It runs on a 12 V dc supply and requires less than 100 mA. And, it doesn't require any fancy equipment except for your antenna and receiver to verify its operation.

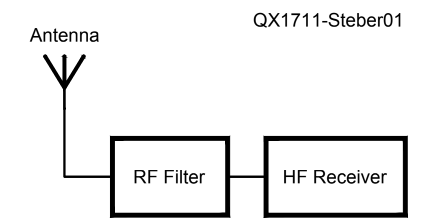


**Figure 1 — An RF filter between the antenna and a low cost radio or experimental HF SDR "dongle" receiver can significantly reduce noise and interference problems**

So, if you want to get started experimenting with a tunable RF preamp, tame your interference problems and at the same time learn a bit about the application of a varicap diode, read on. Home brewing your own tunable preamp might be the way to go.

## RF Tuner Background

This project began when I found that the low cost short-wave radio I used for receiving WSPR and JT9 signals on various bands was suffering from severe interference because of its wide input RF stage. My dongle-based SDR receiver was also suffering the same fate, being overloaded from strong nearby stations. WSPR and JT9 signals by definition are low power signals and this interference was making it more difficult to copy them with my computer software. The interference from the computer wasn't helping either. Adding this filter made it a lot easier to reduce this noise — but it does require turning a knob to peak the filter on the desired frequency.

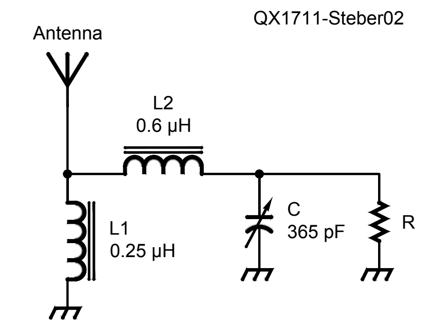


**Figure 2 — Classic tunable RF band-pass filter with variable capacitor *C*. Resistor *R* represents input to next stage, usually a FET amplifier.**

Figure 2 shows a classic tunable circuit that is often used to provide RF selectivity. It is not the same as an antenna tuner, which is used to match the impedance of your antenna/feedline to your receiver or transmitter. The tuner shown here is basically an adjustable band-pass filter. The inductors *L*1 and *L*2 provide matching to the antenna in the range of 25 to 100 Ω and form a tuned circuit with *C*. Resistor *R* is shown to represent the high impedance input of the next stage. *R* must be very high — in the megohm range — to avoid loading down the tuned *LC* circuit. Normally a FET is used here. The peak in the RF response is moved over the frequency range by adjusting the variable capacitor *C*. Figure 3 shows response curves for various values of *C*. They were taken from a SPICE simulation and were found to closely follow those seen with a spectrum analyzer on the actual circuit.

Finding a mechanically adjustable capacitor to use in the tuner would have been a piece of cake a few years ago. This once common part is now hard to find — even at hamfests! When found, the cost is skyrocketing. I found a few on an auction site but the prices were quite high. This could be a special hardship to those in clubs or outreach programs that wish to build your project using this component.

Fortunately there is another way to do the tuning. Use a varicap! In the next section we'll talk about the varicap and how it can be used in this application.

## Using A Varicap

A varicap is essentially a voltage-controlled capacitor. They have been around since the 1960s and are commonly used in voltage-controlled oscillators, parametric amplifiers and frequency multipliers.

Here's how it works. When a diode is operated in a reverse-biased state very little current flows in the device. The effect of applying the reverse bias voltage is to control the thickness of the depletion zone and therefore its p-n junction capacitance. The greater the applied voltage, the greater is the depletion zone and the smaller the capacitance. Most diodes exhibit this characteristic to some extent but varicaps are manufactured to exploit this effect and increase the capacitance over a larger range.

The varicap used in this project (1SV149) has a very large capacitance variation and was designed to replace the tuning capacitor in AM radios. It has high Q (at least 200), a small package, high capacitance ratio, and low voltage operation. It can be found on the internet for under a dollar! The varicap capacitance variation for the 1SV149 with respect to applied dc voltage is shown in Figure 4.

Substituting a varicap for a mechanical capacitor requires a bit of planning. We need to know the capacitance range covered and the tuning voltage required. The tuning curves shown in Figure 3 covered a range of 30 pF to 365 pF. So we would like to find a varicap that covers that range with a reasonable bias voltage range. The 1SV149 varicap covers 25 pF to 500 pF with a voltage range of less than 10 V.

When the 1SV149 varicap is substituted for the mechanical capacitor using an

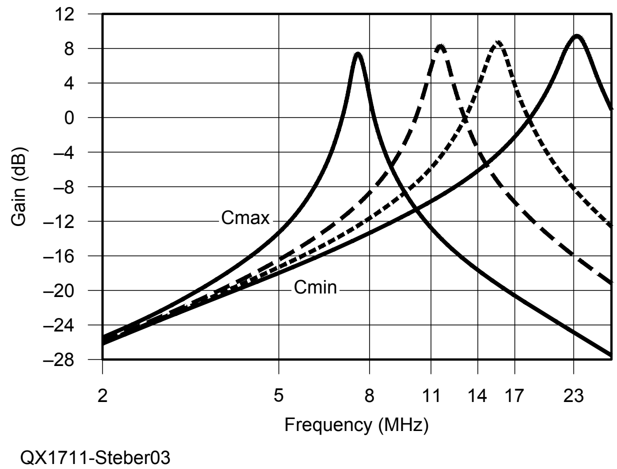


**Figure 3 — Four frequency response curves of the tuned preamplifier as capacitor *C* is varied.**

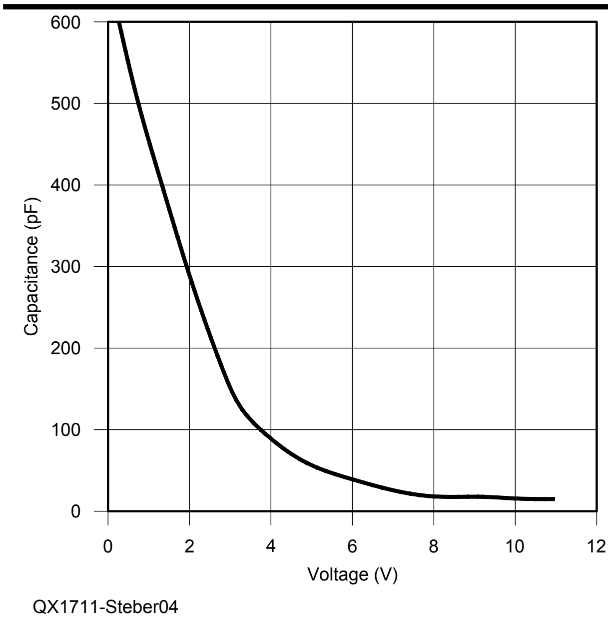


**Figure 4 — Varicap capacitance *C* versus applied dc voltage for the 1SV149 diode used in the project. Data was taken from the manufacturer's data sheet.**

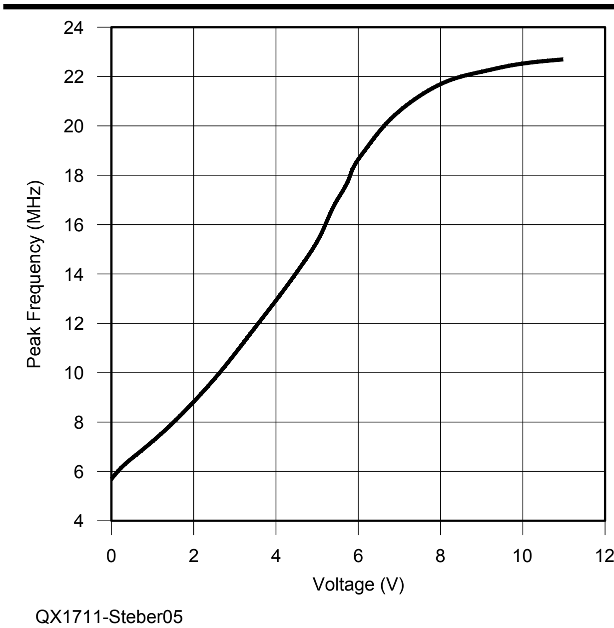


**Figure 5 — Variation of peak frequency with varicap voltage. Data is taken for the 1SV149 used in the project. Frequency range is about 6 to 23 MHz.**

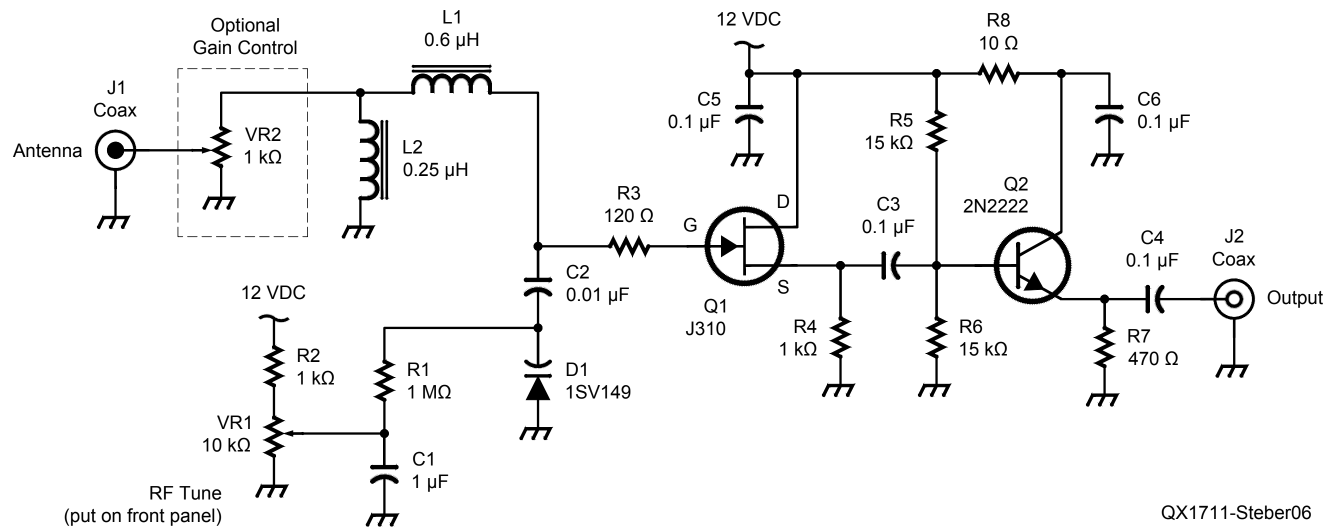


**Figure 6 — Schematic diagram of the tunable RF preamplifier. The component values are shown in Table 1.**

appropriate circuit, the tuning law shown in Figure 5 is obtained. As is seen, the varicap easily covers the same range as the mechanical capacitor and more — in this case 6 to nearly 23 MHz.

## Tunable RF Preamplifier Design Notes

Figure 6 shows the complete schematic of the RF preamp. This circuit was designed using a SPICE simulator, LTspice.[1] After the breadboard circuit was built, it was analyzed using a spectrum analyzer with tracking generator. Very close agreement was found between the Spice simulation and the actual hardware.

This tunable RF preamp can be built for about US $25, depending on your parts inventory. It does not require extraordinary skill with RF circuits. In fact it makes a nice weekend project once the materials are on hand. It is unique in some aspects, such as manual tuning the peak frequency. It will not compete with an expensive transceiver. But it does produce good results in many situations. It made my radios perform significantly better.

Here's how the circuit works. Starting at the antenna there is an optional gain control *VR*2. Its purpose is to reduce the signal level to the preamp if necessary. The preamp does provide RF gain of 6 to 9 dB, which helps with digging out weak signals. But if the signals are too strong, just reduce them with this control.

Next is the tuned circuit consisting of *L*1, *L*2, *C*2 and varicap *D*1. It looks very much like that of Figure 2 and operates the same way. Capacitor *C*2 blocks the dc that is on the varicap *D*1. If *C*2 is made much higher than the capacitance of varicap, the varicap capacitance dominates since they are in series. *R*1 feeds the dc voltage to *D*1 and is very large to avoid loading the *LC* circuit. *VR*1, *R*2 and *C*1 are there to provide a filtered, adjustable dc voltage to control the varicap. *VR*1 should be mounted on the front panel of your unit. Try to obtain a *VR*1 that rotates over a large angle (320 degrees) as this will provide finer control over your tuning.

*Q*1 (n-channel JFET J310) is used as a high impedance buffer for the *LC* circuit. It has a resistor *R*3 at the gate to reduce possible instability. It must have a bypass capacitor *C*5 close to the drain *D* of the device. The last stage is a standard bipolar transistor *Q*2 that provides a low impedance output. *Q*2 is a 2N2222 type, but if a higher noise figure can be tolerated a 2N3904 may be used. For a lower noise figure try a 2N5109 — but this one may be hard to find. Also be careful of the P2N2222, which has reversed pin order.

**Table 1.**
**Components values for the RF tunable preamplifier.**

| *Qty* | *label* | *value* | *description* |
|---|---|---|---|
| 1 | C1 | 1 µF | electrolytic capacitor |
| 1 | C2 | 0.01 µF | ceramic capacitor |
| 4 | C3, C4, C5, C6 | 0.1 µF | ceramic capacitor |
| 1 | L1 | 0.60 µH | inductor, 11 turns T50-6 core |
| 1 | L2 | 0.25 µH | inductor, 5 turns T50-6 core |
| 1 | R1 | 1 MΩ | 1/4 W 5% resistor |
| 2 | R2, R4 | 1 kΩ | 1/4 W 5% resistor |
| 1 | R3 | 120 Ω | 1/4 W 5% resistor |
| 2 | R5, R6 | 15 kΩ | 1/4 W 5% resistor |
| 1 | R7 | 470 Ω | 1/4 W 5% resistor |
| 1 | R8 | 10 Ω | 1/4 W 5% resistor |
| 1 | VR1 | 10 kΩ | potentiometer |
| 1 | VR2 | 1 kΩ | potentiometer (optional) |
| 1 | D1 | 1SV149 | varicap diode |
| 1 | Q1 | J310 | FET |
| 1 | Q2 | 2N2222 | bipolar transistor |
| 2 | J1, J2 | connector | coax connector |

BNC connectors were used for *J*1 (antenna) and *J*2 (output) as they matched my equipment. Other connector types may be used. A standard 12 V dc power supply is used. If there is noise from the power supply, add an electrolytic capacitor between the 12 V dc line and ground.

Various techniques may be used to build this preamp. Use your own ingenuity! Some good construction methods may be found in previous *QST* articles or one of the ARRL Handbooks.[2] I constructed a breadboard on a small solderable copper-pad perf-board. A simple PCB inside a metal case would probably be the best approach.

## Tunable Preamplifier Modifications

The tuning range measured for this circuit was 6 to 23 MHz. This range can be changed easily. It is mainly dependent on the inductance of *L*1, *L*2 and capacitance of *C*. It is difficult to predict the range accurately. This is because inductors may be constructed slightly differently, the stray capacitance of the circuit is unknown, and the varicaps have a large range of tolerance.

First, lets take a look at the problem theoretically. The peak frequency *f* occurs at resonance

$$f = \frac{1}{2\pi\sqrt{(L1+L2)C}}.$$

The tuning range for (*L*1+*L*2) equal to 0.85 µH and two values of *C* (30 pF and 500 pF) yields a range of 7.7 MHz to 31.5 MHz. This is quite different from the measured values — most likely because the inductor values are slightly different and there is stray capacitance.

To accurately calibrate the tuner, you must measure the response after building it. A suitable RF generator would prove useful in this case. With that caveat, here are some ranges for my circuit that I obtained experimentally.

Choosing *L*1 = 2 µH and *L*2 = 0.68 µH covers 4.3 MHz to 11.9 MHz. Going even lower choose *L*1=10 µH and *L*2=2 µH which covers 2.1 MHz to 8.0 MHz. You can play with inductor values to get many ranges. You may want to consider putting in a range selector switch to cover the ranges you want.

Generally speaking, *L*2 is chosen to provide 25 to 100 Ω reactance over the frequencies of interest to match the antenna. But that is not a strict rule as the tuner can be used with many different antennas, even long wires.

## Final Tuning

The RF tunable preamplifier can prove helpful in a variety of interference situations. Even though its design is simple, the varicap works well and is a good substitute for the mechanical capacitor. One nice benefit of using the varicap is that the variable resistor *VR*1, controlling the dc voltage, has a larger angle of rotation (320 degrees) than the mechanical capacitor (180 degrees) and therefore has finer control. The varicap also has a larger range.

Using the preamp is straightforward. Simply adjust the front panel-tuning resistor *VR*1 until the received signal is the loudest. If you are using decoding software such as JT9 or Fldigi that have a spectrum analyzer display you will see a pronounced increase in the signal as you approach the peak. Remember to tune slowly.

## Conclusion

I enjoyed constructing and using this preamp. Hopefully you will find that to be the case as well. While building projects like this is fun, they can also be educational. Consider demonstrating this application of a varicap in a teaching situation such as a presentation at your local ham club or possibly as an outreach tool at your local high school science class.

When you get your RF tuner working, please let me know about it. With the radio interference reduced, you might now be able to copy those weak signals from the far distant regions of the world.

*George R. Steber, Ph.D., is Emeritus Professor of Electrical Engineering and Computer Science at the University Of Wisconsin-Milwaukee. He is now semi-retired having served over 35 years. George, WB9LVI, has an Advanced class license, is a life member of ARRL and IEEE and is a Professional Engineer. His last article for QST was "An Easy WSPR 30 Meter Transmitter" in the January 2015 issue. George has worked for NASA and the USAF and keeps busy working and lecturing on various subjects at the University. He is currently involved in cosmic ray research and is designing equipment to study them on a global basis. In his spare time he enjoys WSPR-X mode JT9, racquetball, astronomy, and jazz.*

## Notes

[1]LTspice, **www.linear.com/designtools/software/**.

[2]*The ARRL Handbook for Radio Communications,* 2017 Edition. Available from your ARRL dealer or the ARRL Bookstore, ARRL item no. 0628. Telephone 860-594-0355, or toll-free in the US 888-277-5289; **www.arrl.org/shop**; **pubsales@arrl.org**.

**Dr. Sam Green, WØPCE**

10951 Pem Rd., Saint Louis, MO 63146; **w0pce@arrl.net**

# Microwave Version of Wideband QRP SWR Meter

*Modified experimental couplers and higher frequency logarithmic detectors extend the frequency range of the low power SWR meter up to 4 GHz.*

Four years ago, I published "An Extremely Wideband QRP SWR Meter" in this journal.[1] That SWR meter worked over the frequency range of 1 to 500 MHz at unprecedentedly low power (less than 100 mW). The paper also suggested that my colleague Lee Johnson and I were on the verge of making the meter or the coupler into a product by fabricating the coupler in stripline instead of embedded microstrip. That wasn't as simple as we thought. Four years later, we still haven't invested adequate money or effort into developing the microstrip coupler for various reasons. You just can't get stripline from ExpressPCB.

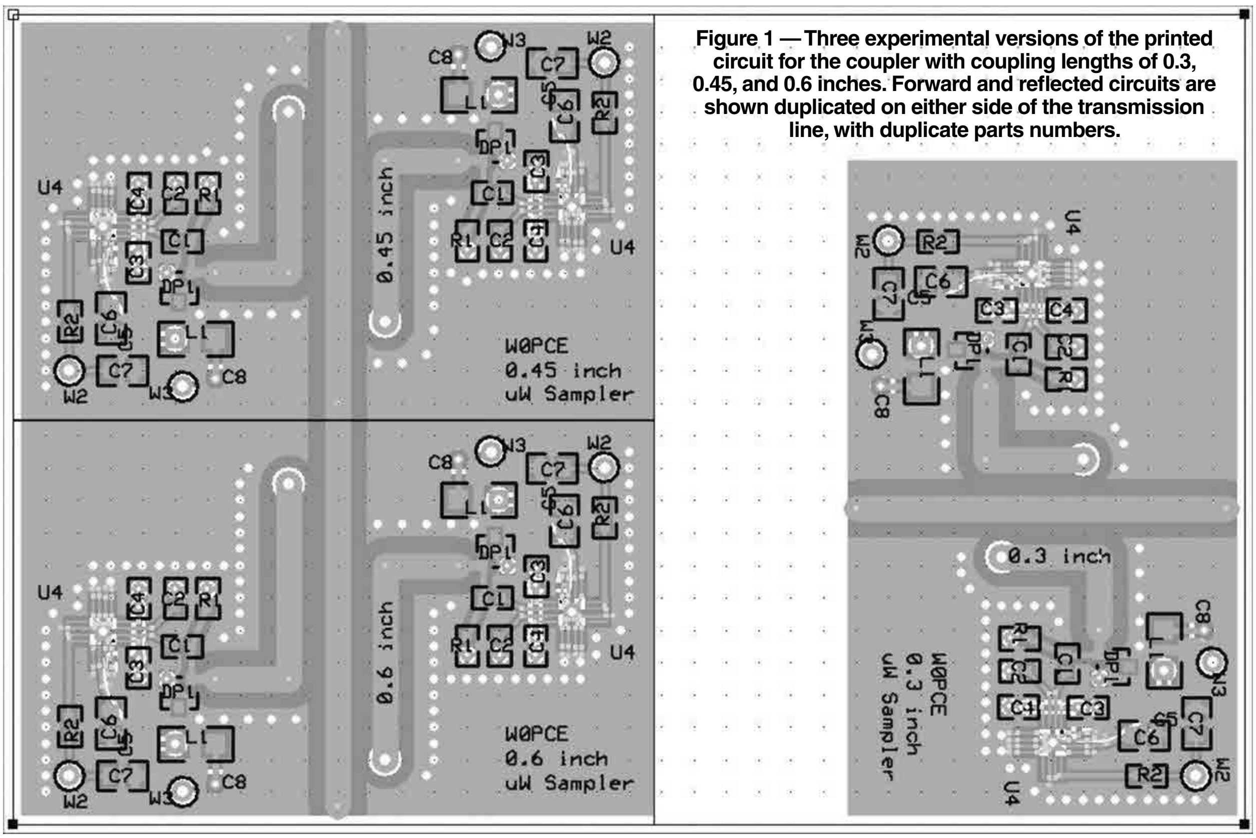


**Figure 1 — Three experimental versions of the printed circuit for the coupler with coupling lengths of 0.3, 0.45, and 0.6 inches. Forward and reflected circuits are shown duplicated on either side of the transmission line, with duplicate parts numbers.**

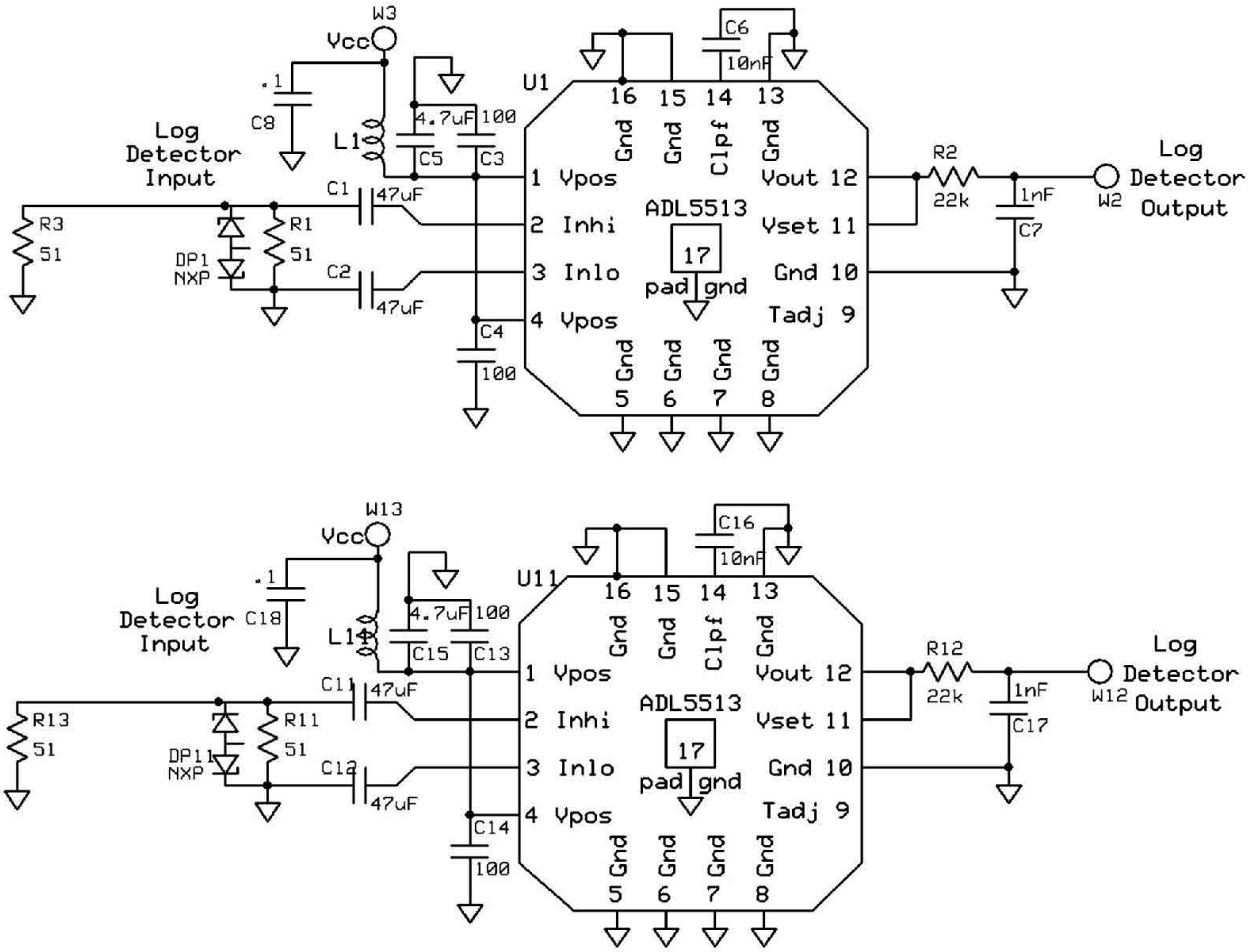


**Figure 2 — Diagram of the microwave circuit. L1 and L11 are the surface mount equivalent of shield beads, and the forward and reflected inputs are across R3 and R13 respectively. U1 and U11 are the "U4" ICs in Figure 1. Compare this to the 500 MHz version in Note 1.**

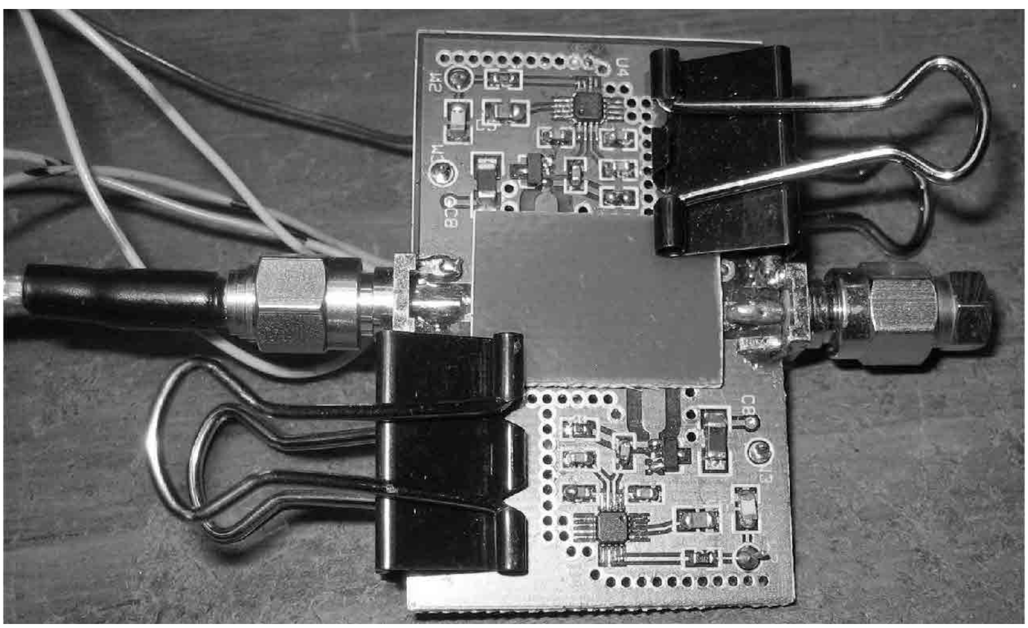

**Figure 3 — The assembled experimental coupler with the needed dielectric layer held in place by binder clips above the printed circuit board.**

Still, the embedded microstrip approach worked well enough, and I had a set of design parameters that I could transfer to a new printed circuit board layout with good confidence that I could again reproduce the optimal 50 Ω impedance transmission lines. I never purused that thought much further until I decided to try to make a microwave version with the 4 GHz logarithmic detectors that I had used in another project.[2]

## The Microwave Design

I designed new couplers using the same embedded microstrip transmission line dimensions and spacing, and incorporated a pair of 4 GHz ADL5513 logarithmic detectors [3], in place of the original 500 MHz AD8307 logarithmic detectors. Figure 1 shows three versions of the printed circuit for the coupler with coupling lengths of 0.3, 0.45, and 0.6 inches. Figure 2 shows the circuit of the couplers. Figure 3 shows a photo of the assembled coupler with the necessary dielectric layer held in place above the printed circuit board by binder clips to form the upper dielectric region necessary for embedded microstrip. Please refer to the original paper for a more thorough discussion of microstrip, embedded microstrip, and stripline. The following are highlights.

Wikipedia explains,[4] "The λ/4 coupled line design is good for coaxial and stripline implementations but does not work so well in the now popular microstrip format, although designs do exist. The reason for this is that microstrip is not a homogeneous medium —there are two different mediums above and below the transmission strip. This leads to transmission modes other than the usual TEM mode found in conductive circuits."

Another reference[5] reports, "Planar structures (unless they are stripline) have notoriously bad directivity. Directivity (isolation minus coupling) is determined in these types of structures by the difference between the even and odd mode phase velocities with the best coming when these are equal. In microstrip, the odd mode is mostly in dielectric, and the even mode is mostly in air. To equalize the phase velocities, you need to slow down the even mode which can be accomplished using a dielectric overlay over the lines (microstrip case)."

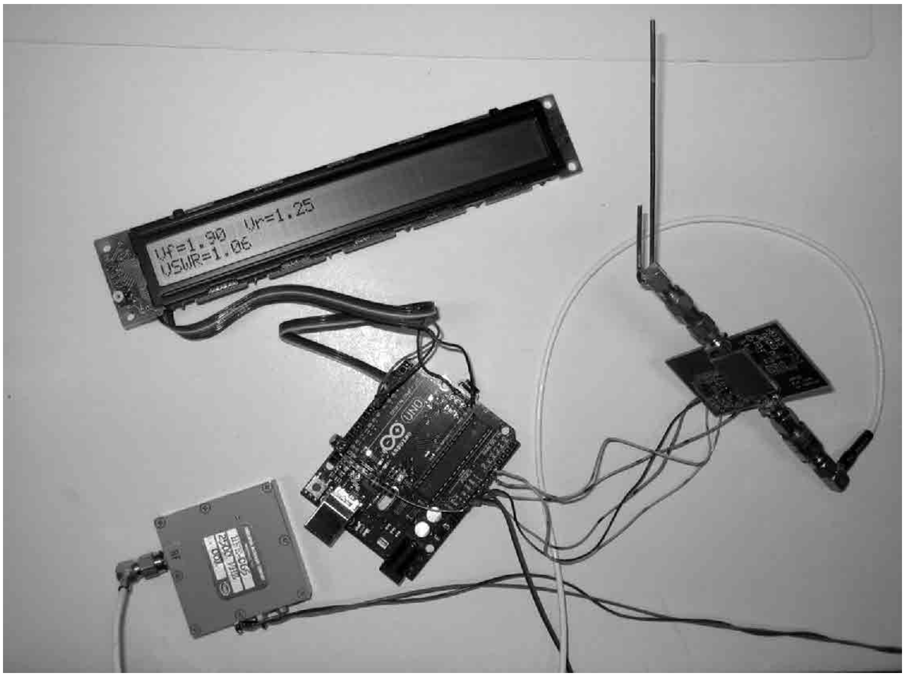


**Figure 4 — The new coupler is shown connected between a 2500 MHz VFO and an experimental 2500 MHz J-pole antenna.**

## Results

The new coupler connects to an Arduino as in the original paper to make the two analog measurements, perform the calculations, and drive the liquid crystal display. With a 2500 MHz VFO as a test source, the instrument indicates 1.0:1 VSWR with a good 50 Ω termination (load).

Figure 4 shows the new coupler connected between a 2500 MHz VFO and an antenna. Yes, that's a 2500 MHz J-Pole! Using this instrument, I moved the SMA connector up and down the J-Pole, and re-soldered it at various points until I achieved the low indicated VSWR. Success!

## Discussion

I would not recommend that you copy this coupler design. It was very difficult to get both ADL5513 logarithmic detectors to work at the same time. The 16-lead Lead Frame Chip Scale Package [LFCSP_VQ] has contact pads underneath instead of leads and is very difficult to solder with the techniques I then used. Since I made this one microwave coupler almost 3 years ago, Lee taught me to use solder paste and a toaster oven successfully on several other projects. Still, I am not anxious to build any more of these couplers with ADL5513 logarithmic detectors onto the remaining boards.

Thanks to Analog Devices for providing free samples of their parts.

*Dr. Sam Green, WØPCE, is a retired aerospace engineer living in Saint Louis, Missouri. He holds undergraduate and graduate degrees in Electronic Engineering from Northwestern University and the University of Illinois at Urbana respectively. Sam specialized in free space optical and fiber optical data communications and photonics. Sam is currently designing a prototyping innovation target for guns with laser bullets. Sam became KN9KEQ and K9KEQ in 1957, while a high school freshman in Skokie, Illinois, where he was a Skokie Six Meter Indian. Sam held a Technician class license for 36 years before finally upgrading to Amateur Extra class in 1993. He is a member of ARRL, a member of the Boeing Employees Amateur Radio Society (BEARS), a member of the Saint Louis QRP Society (SLQS), and a member of the Bi-State Amateur Radio Society. Sam is a Registered Professional Engineer in Missouri and a Life Senior Member of IEEE. Sam holds eighteen patents.*

## Notes

[1] Sam Green, WØPCE, "An Extremely Wideband QRP SWR Meter", *QEX*, Jan/Feb 2014, pp. 15-23.

[2] Sam Green, WØPCE, "A Fully Automated Sweep Generator Measurement System - Take 3", *QEX*, Jul/Aug 2014, pp. 7-15.

[3] **www.analog.com/media/en/technical-documentation/data-sheets/ADL5513.pdf**.

[4] See **https://en.wikipedia.org/wiki/Power_dividers_and_directional_couplers**

[5] The cited reference was from message board that is no longer active.

**Michelle D. Thompson, W5NYV**

5379 Carmel Kolls Dr, San Diego, CA 92130;
**mountain.michelle@gmail.com**

**Howie DeFelice, AB2S**

804 Fall Pl., Herndon, VA 20170;
**howied231@hotmail.com**

# Evolutionary Engineering for Revolutionary Satellites — AMSAT Next Generation

*Next generation AMSAT satellites are faced with new challenges in uncertain launch schedules, development schedules and justification for the project.*

AMSAT must shift from a reactive to a proactive engineering philosophy. In general, AMSAT waits for a launch opportunity and then starts recruiting volunteers to design satellites and payloads. This strategy is no longer viable due to three fundamental changes in the way the space industry works: (1) uncertainty in launch schedules, (2) lead time between launch availability and satellite development, and (3) justification for launching a satellite.

## 1 — Launch Uncertainty

There is increased uncertainty in launch schedules. Launch opportunities have fundamentally changed over the history of the organization. The impression is that launches were easier to obtain and less expensive in the past because Amateur Radio was familiar to the more military-minded launch authorities that approved the inclusion of Amateur Radio payloads. Launches were described as more experimental, less commercial, and more controllable within a hierarchy that was less sensitive to cost and more familiar with Amateur Radio. With the increasing industrialization and commercialization of satellite launches, amateur payloads were no longer privileged. They were expected to pay their way just like any other payload. Quotes were prohibitively high, ranging into the millions of USD. Amateur Radio's uncertain role in the commercial space industry makes it harder to recruit teams and raise funds.

There is an impression that lead times for launches have shortened, further challenging the AMSAT volunteer corps with shortened engineering schedules. However, there are some counterexamples. SumbandilaSat started in 2005 and did not launch until late 2009. FUNcube began around 2007 and didn't launch until 2013. Having a long launch window can be as negative as one that is too short, since it may be very difficult to keep a volunteer team together and motivated for 4 to 6 years, especially if the launch dates are unspecified for the majority of the design effort.

Launch opportunities have been offered with very short windows. Launch opportunities currently accepted and pursued by AMSAT have exhibited high probabilities of being delayed or cancelled or transformed. The modern launch opportunity landscape is dominated by uncertainty. Schedules might be short or long. Launches might be a sure thing or they might be very tentative. Coupling work to a launch means work stops and starts, or never starts in the first place because the launch isn't a "sure thing". Making payloads the priority instead of waiting on a launch to organize the work around is the right way to respond to increased schedule uncertainty. Suppressing publicity, marketing, recruitment, and fundraising because modern launches aren't a "sure thing" is detrimental.

## 2 — Design Time and Launch Lead Time

We can no longer afford to wait for a launch and still hope to finish modern designs. The lead time for development of highly desirable digital designs is unavoidably lengthy when compared to lead times for established analog designs. The increased functionality and flexibility of digital design comes at the direct cost of increased complexity. Complex things take longer to engineer, especially the first time through. The promise of re-usability through modular digital design and software is real. That promise is delivered through solid systems engineering and clear documentation.

AMSAT is working to build up a volunteer corps with expertise in digital design. The number of people with this expertise is still small compared to the industry demand. As a result, the number of people with the spare time to donate to AMSAT is limited. However, even if AMSAT was fully staffed and in possession of a working library of digital designs that could be efficiently adapted to the specifics of individual payloads, the increased complexity of digital design may still mean longer schedules when compared to simpler analog designs. Software is most often the biggest schedule risk in digital design.

## 3 — Justifying the Need for a Launch

Communications services alone are no longer a justification for launching a satellite. The continuing commodification of communications is something AMSAT, and all of Amateur Radio, must adapt to and confront so as to remain relevant. This means that if we want to continue concentrating almost exclusively on satellite communications, then we must pick truly challenging space-based communications projects and provide designs, justification, and documentation. We must be prepared to opportunistically lobby for inclusion in nontraditional niches.

An example of this is AMSAT's participation in the NASA CubeQuest Challenge. AMSAT volunteers devised a method to determine satellite range and range rate using modest antennas out to a distance of over 750,000 km. This is challenging and engaging work that needs full recruitment publicity support from AMSAT.

What are the things that we are already doing within AMSAT that would be the best fit for a proactive strategy?

## Proactive Strategy

The "Five and Dime" digital microwave approach is a good candidate. The 4A, 4B, and CubeQuest payloads, terrestrial Groundsats, and Phase 4 Ground radios are all reconfigurable by design. The idea of building to a published air interface that relies on a well-known open standard (DVB) serves a proactive payload principal very well. If this approach proves to be as successful as we think it will, other international AMSAT groups may adopt this proven standard making larger numbers of satellites available for use with a common ground station.

The FOX designs are excellent candidates. The FOX LEO design has shown to be reliable and robust.

What Amateur Radio centric designs are unquestionably attractive as payloads, besides the successful FOX program? What could we as amateur experimenters build and demonstrate that would be worth the cost to launch?

## Engineering Challenges

An engineering challenge that could be worth a subsidized launch would be to build a practical HEO CubeSat in a 1U form factor. This would be an extremely rewarding endeavor requiring creativity in mechanical, RF, analog and digital design. AMSAT could be the record holder for the smallest HEO ever orbited. This is the kind of engineering challenges that will attract creative thinkers with the possibility of seeing their ideas actually placed into orbit.

**Table 1.**
**Maximum expected power that can be generated from various size CubeSats.**

| *Size* | *Fixed solar panels* | *Fixed plus deployable panels* |
|---|---|---|
| 1U | 2 watts | 10 watts |
| 2U | 5 watts | 12-23 watts |
| 3U | 7 watts | 21-30 watts |
| 6U | 14 watts | 40-60 watts |

AMSAT has not had a high altitude satellite since AO-40 stopped functioning in 2004. The economics and politics of replacing AO-40 with *any* sort of HEO satellite has relegated AMSAT to LEO satellites of the CubeSat variety in the form of the Fox series of satellites. The commercial success of the CubeSat platform has created opportunities to reach higher orbits as a ride share. Creating a practical Amateur Radio satellite that is operable from HEO altitudes of over 40,000 km is extremely challenging. The biggest problem is generating sufficient power from the available satellite real estate.

The Table 1 summarizes the maximum expected power that can be generated from various size CubeSats based on data from the Clyde Space web site (**https://www.clyde.space/**).

To put this in perspective, AO-13 generated 50 W of solar power and had a 50 W VHF downlink linear transponder. AO-40 could have generated in excess of 600 W of solar power. The 2.4 GHz downlink had 50 W of RF power. Both satellites used omni directional antennas. Anyone who operated those satellites remembers the difference in antenna size required to make a contact. AO-13 required long boom or stacked VHF/UHF beams where AO-40 could be worked with a small (0.6 meter) dish for downlink and a single 10 element WA5VJB "cheap Yagi" (**www.wa5vjb.com/references.html**). For the same level of RF power, higher frequencies mean smaller antennas on the ground.

## Ground Station Antennas

Smaller antennas on the ground are another reality that has become increasingly important over the past 13 years since AO-40. More Amateur Radio operators and potential Amateur Radio operators live in antenna restricted areas. A practical Amateur Radio ground antenna needs to be not more than 1 meter in diameter. This establishes another criteria for our HEO CubeSat. It must be able to complete a downlink to a 1 meter antenna without exceeding the dc power limits of the satellite. The RF power of the transmitter is the largest consumer of power but it's not the only power consumer on the spacecraft. Power must be allocated for satellite monitor and control, on board signal processing, environmental control, attitude control and any potential scientific experiments that may be required to qualify for a launch.

## Conservation of Satellite Power

Conservation of power is critical for a HEO CubeSat so the best place to start is with the transponder. General guidelines for an efficient and effective downlink are as follows.

1.— Use a single carrier for downlink with the SSPA operating close to saturation.

2.— The downlink should be PSK digitally modulated with robust FEC.

3.— Directional antenna with gain should be used to increase EIRP.

The desire to use directional antennas impose other requirements on the satellite, attitude control. The satellite attitude must be controlled with enough accuracy to keep the Earth within the beamwidth of the antenna. This should not be too difficult for the size and gain of the typical CubeSat antenna. Another advantage of attitude control is the possibility to maximize Sun angle to increase solar panel output.

All previous Phase 3 satellites used attitude control. AO-10 (see Figure 1) and AO-13 were spin stabilized. AO-40 had the ability to be three axis controlled using momentum wheels but due to other issues it too was spin stabilized. Spin stabilization is an effective means of controlling satellite attitude when used in conjunction with *magnetorqueing*. For spin stabilization to work, the satellite frame must be constructed with a high width to height ratio, with the satellite weight equally distributed so that the satellite is balanced about the axis of rotation.

This type of structural symmetry is difficult to achieve with a CubeSat. Other forms of stabilization such as gravity gradient boom, miniature reaction wheels and relocatable mass need to evaluated for effectiveness.

## Small Footprint, More Capable Satellites

The orbital parameters of the launch could

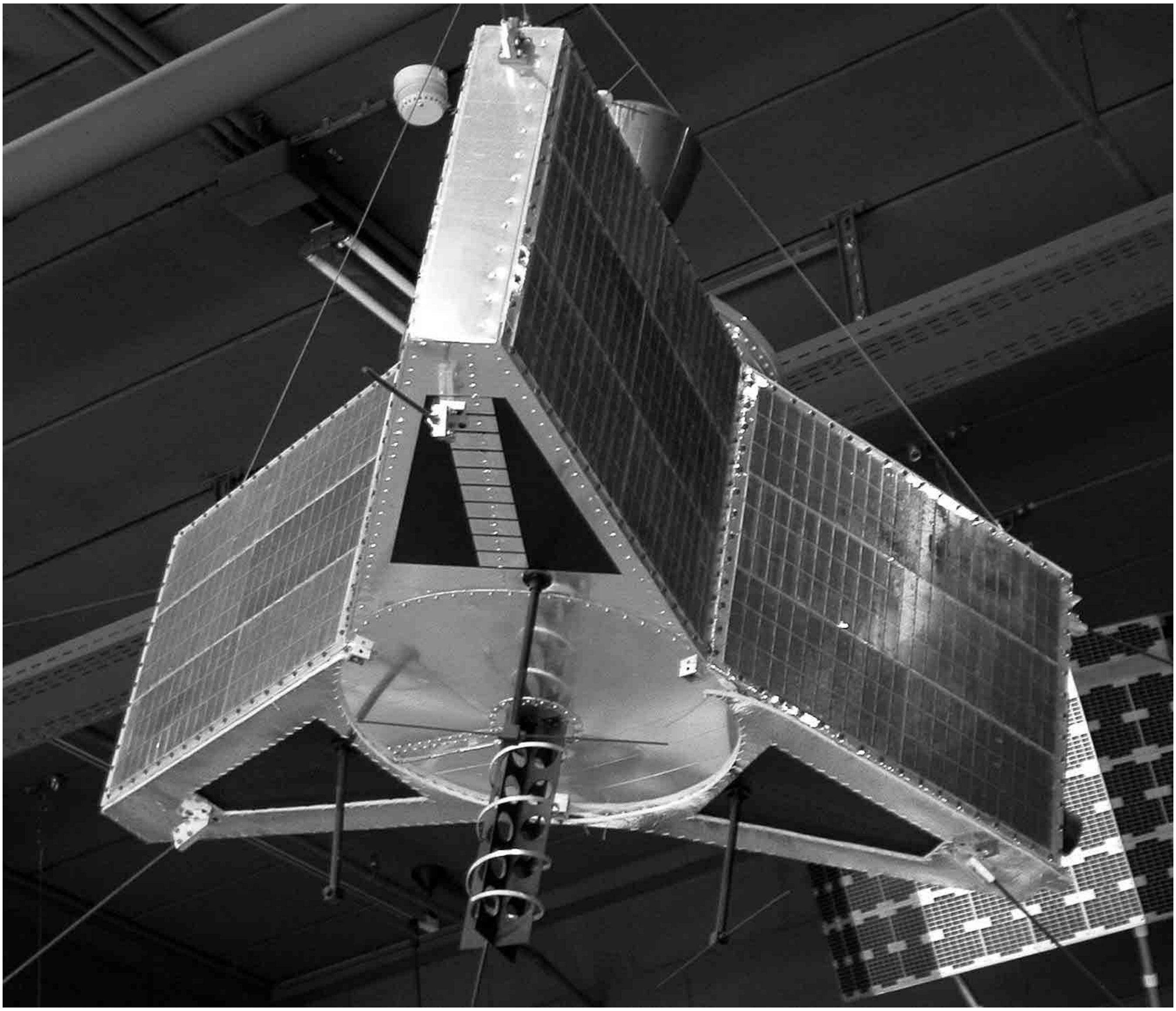

**Figure 1 — A model of AMSAT OSCAR 10.**

have an impact on the satellite design as well. If launched as a secondary payload on a GTO (Geosynchronous Transfer Orbit), the perigee may be too low to prevent premature orbit decay. Thrusters using volatile fuels are usually prohibited on secondary payloads so some form of non-volatile propellant needs to be investigated.

While we have been focused on improving our expertise in digital design, an over-arching challenge to AMSAT is to build more capable satellite in smaller footprints. The smaller the satellite, the greater the challenges. A HEO CubeSat will require thinking outside the cube.

P4 is an unusual exception in that the most common launches we can expect which will most likely be 6U or 3U CubeSats to HEO. Therefore, AMSAT should declare work on a family of HEOs that included 1U, 3U, and 6U packages. Each of these projects would be expected to go above and beyond the current status quo. The vision would be to exceed expectations and attract launches by demonstrating successful designs worth launching. The payloads must be more than commodity communications or they must provide value-added service to a scientific package. Identifying synergy between a science package or experiment and the radio communications requires being able to communicate with a wide variety of scientists and engineers.

## AMSAT Needs

AMSAT must have a volunteer corps of digital and analog designers that are willing to work, independent of launches. AMSAT must identify a family of payload projects, independent of launches. AMSAT must find the money to build and test those payloads, independent of launches. In parallel, AMSAT must opportunistically and aggressively pursue launches and missions where these projects can be deployed.

Culturally, AMSAT has biased itself towards appealing to operators instead of technical experimenters. This was fine in the early days when a large part of the membership were highly technical, that got on the air just to see how well their new modification or home-brew antenna worked. Chasing awards was an anomaly. The recruitment of operators was a way to grow membership beyond the highly technical crowd. However, in the opinion of AMSAT leaders such as Howie DeFelice, AB2S, we "went too far."

The typical AMSAT member is portrayed as an FM LEO operator chasing grid squares. AMSAT emphasizes and celebrates the operating skills and dedication of the people who enjoy this aspect of satellite communications. Is this a large and diverse enough community to ensure the viability of AMSAT moving forward?

AMSAT's newest satellite is an FM repeater. There are two or three more FM repeaters scheduled to follow. More linear LEOs are not the answer either. There are now at least four linear LEOs in orbit and the biggest complaint among the people who operate them is that there are not enough people to talk to.

## AMSAT, Maker Movement, and ITAR/EAR

Howie and others feel that AMSAT has a unique opportunity with the resurgence of the Maker Movement. There is a renewed interest in experiencing the satisfaction that comes with using something you built yourself. AMSAT provides a conduit that the Maker Movement can't get anywhere else. This requires AMSAT to have ongoing engineering projects with lots of people working in small teams as either part of a bigger project or as a research team for a particular goal, similar to the way Phase 4 Ground is constructed to serve multiple payloads.

Decoupling the work from launch schedules and increasing our presence in communities such as the Maker Movement will require a major change in the way AMSAT thinks about engineering communication and where it spends volunteer time, energy, effort, and funds. The ITAR/EAR (International Traffic in Arms Regulations / the Export Administration Regulations) problem has to be clearly and publicly resolved. The lack of a clear, consistent, public policy from AMSAT on ITAR/EAR has been a major ongoing impediment to amateur satellite engineering communications in the United States. Silence in the face of need stands in very stark contrast to the open innovative nature of Amateur Radio in general and the open source Maker Movement in particular. Being afraid to communicate comes from the perception of enormous legal risk if one guesses wrong on what to say or not say. Any policy must come from AMSAT's board of directors. The continuing damage done to amateur satellite service engineering by ITAR/EAR is an existential crisis to AMSAT and should be treated as such.

One way forward is to continue and expand the practice of teaming up with universities and other research institutions. These organizations often demand clear guidelines on what communications are legally allowed, and very often have departments or human resources tasked with managing ITAR/EAR constraints. This spreads the risk and increases collaboration at the cost of having a centralized and consistent AMSAT policy. Leadership in establishing a clear pro-volunteer ITAR/EAR policy provides substantial long-term benefits to Amateur Radio and the larger science, engineering, and technical fields.

## Summary Remarks

Launch uncertainties cannot be used as a gating item to halt or delay or suppress work on digital designs. There's just too much work that needs to be done. If work stopped and waited for a rock-solid launch opportunity, then the work would literally never be completed. This is the reality of large complex digital designs and the challenge of smaller footprints. If we accept the conventional AMSAT wisdom that people won't volunteer without solid launch schedules, then the obvious implication is that AMSAT can't do challenging and complex designs. If AMSAT can't execute challenging and complex designs, then we cannot participate in current, let alone future, space-based communications projects. If we want to keep Amateur Radio in space, then we have to execute challenging and complex designs. In order to engineer these designs, we have to decouple launches from our schedule and design optimistically, speculatively, and flexibly. This means a big step up in volunteer corps expertise and a change in communications, fundraising, and recruiting focus.

## How to Help

Join the membership organizations that help achieve the goals you want to see in Amateur Radio. Join AMSAT, ARRL, TAPR, ATN, your local or regional amateur technical organization. Support efforts that advance the state of the art. Have expectations. Ask questions. Active members of a community set the agenda. Become active.

*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, 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.*

*Howie DeFelice, AB2S, was employed in wireless communications continuously since 1974, starting as a test technician at Communications Associates, Inc., working with HF-SSB commercial marine transceivers. This led to employment in land mobile radio, then as a field engineer for Magnavox Marine and Survey Division, and then various positions in California Microwave designing and building INMARSAT transportable earth stations, fixed earth stations and designing and delivering multi user VSAT networks. He attended SUNY Farmingdale and received an Electrical Engineering Technology degree, and currently works with wireless security. Howie has a Commercial Radiotelephone license, earned the Novice license in 1976, upgraded to Advanced class in 1977, and to Amateur Extra in 1979. He joined AMSAT around the time AO-13 was launched. Howie enjoys building and experimenting more than operating and looks forward to help make the next generation of HEO satellites a reality.*

**Rick Campbell, KK7B**

Dept. of Electrical and Computer Engineering, Portland State University, P.O. Box 751, Portland, OR 97201; **RLC3@pdx.edu**

# Square Four Aerials

*This four-element Yagi with the boom length nearly equal to the reflector length is an elegant VHF/UHF gain antenna solution that has an exceptionally clean pattern.*

Yagi-Uda antennas with driven and parasitic wire elements come in many shapes and sizes. Historically, radio amateurs focused on maximum gain that would fit the space and budget, until they began aiming their antennas into cold space and discovered that receive noise floor was often limited by terrestrial noise sources in the side and back lobes. In the 21st century, gain and clean pattern are the criteria. The antennas described here are compact, small, lightweight, easily taken apart and reassembled, and they are a convenient size for strapping to a backpack for hiking into portable locations. Gain is modest and the radiation patterns are exceptionally clean. In addition, they embrace an aesthetic that appeals to the technical senses, as did the *Ariel Square Four*.[1]

## Square Four Aerial for Portable Use

The local 21st century noise environment, dominated by a plethora of digital devices and their switching power supplies, makes urban residential VHF-UHF operation an exercise in frustration. A drive out of town and hike up some hill with a view offers relief from both the cares of daily life and the awful EM noise environment in which we live. Events like Summits on the Air (SOTA) encourage lightweight portable VHF operation. A younger generation is discovering the joys of the oldest radio experiment: "can you hear me now?" Often no, when the antenna is a rubber duck on a small portable transceiver.

The ubiquitous rubber duck is an exceptionally modest antenna that receives equally well in all directions, and local hills tend to be line-of-site to another hill with a forest of high power transmitting facilities. So a clean pattern may be as necessary for portable work as for EME. The initial "Square Four" Yagi was developed years ago for 2 meters. It has four elements, and the boom length is nearly equal to the reflector length. It fits in a square. The three parasitic elements poke through a lightweight wooden boom, held by friction, and the driven element is a design puzzle.

## Feeding and Matching

A clean pattern, and indeed antenna performance in general, requires decoupling from the feed line. The simplicity of J-driven elements is appealing for impedance matching, but appalling from an electromagnetic point of view. I confess to have used them, but my antenna professor D. K. Reynolds literally gagged at the concept.

A half-wave dipole in free space presents an impedance near 70 Ω at its center. Parasitic elements nearby couple electromagnetically, and the arrangement of elements becomes an array with all the far-field radiation from each element adding in-phase in some directions and cancelling in others. Since the parasitic elements take energy from the driven element, the drive point impedance goes down. Historically, small Yagi antennas have been designed for a drive point impedance near 12.5 Ω, and that is stepped up to 50 Ω with a folded dipole driven element. This is good practice. The resulting currents in the driven element and parasitic elements are symmetrical, and if the balanced feed line follows a right angle path away from the driven element, no antenna currents are induced. A few ferrite beads may be slipped over the coax as a balun, but measurements have shown that they might not be necessary. Symmetry does most of the job.

## Common Mode Coupling

Electromagnetically, coax feed line has two paths: a TEM travelling wave inside, that carries the signal, and the outside of the outer conductor, electromagnetically separate, just a length of wire. That's why you can grab the outer conductor of aluminum hard line with a kilowatt inside and not feel a thing. When the outer conductor of a coax line is in an asymmetrical EM field, for example if it leads away from a Yagi parallel with the elements or if the Yagi antenna currents are asymmetrical, the coax becomes part of the antenna. [A balun, or common mode choke, at that point is useful because it is difficult in practice to achieve a truly electrically and electromagnetically balanced load. The balanced to unbalanced transition upsets the needed symmetry and balance[2]. — *Ed*.]. A simple test for feed line decoupling is to tune in a weak signal and grab the outside of the coax a small distance away from the antenna. Nothing should change. If it does, then the coax is an unintentional antenna element. The effects can be quite dramatic with J poles and J-driven element Yagi antennas. [See the John Stanley vertical antenna study.[3] — *Ed*.].

## Folded Dipole Driven Element

Folded dipole driven elements are elegant and work well, but present two challenges to the builder of a few antennas: they are hard to mount, and they are difficult to tune. The mounting becomes even more of a challenge when the antenna needs to be easily assembled and disassembled for portable work. Figure 1 show the present solution for the antenna experiments at my station. The 144 MHz Square Four Aerial is tied to a lightweight mast with a sailor knot. The mast and all the elements disassemble quickly and all slip inside the thin hollow fiberglass mast for transport. The mast is in four telescoping sections that are less than 4-feet long. This remains a work in progress.

A PC board center insulator clamps to the boom with a screw. U-shaped folded dipole halves are soldered in place, and the length trimmed by de-soldering and making fine adjustments. The feed line is soldered in place as in Figure 2, or attached with screws and nuts.

Figure 3 shows the 222.1 MHz and 144.2 MHz optimized square Yagi antennas. Construction is ultra light, with 1/8-inch diameter parasitic elements, #12 AWG copper wire folded dipoles, and wood booms. The 222.1 MHz boom is 3/4-inch clear fir, and the 144.2 MHz boom is 5/8 inch laminated from two pieces of fir, glued with Titebond® III.

**Figure 1 — The 144 MHz Square Four Aerial is shown tied to lightweight mast with a sailor knot. The mast and all the elements disassemble quickly and all slip inside the thin hollow fiberglass mast for transport. The mast is in four telescoping sections, each less than 4 feet long.**

## Design by Modeling

A feature of Yagi-Uda antenna design is that the parasitic element lengths and spacing from the driven element set the pattern. The amplitude and phase of the current in each parasitic element is relative to the driven element, so you can adjust the length of the driven element last, and it has no effect on the pattern. The traditional approach to Yagi design is to adjust the elements for pattern, and then adjust the driven element and its feed network for a 50 Ω match. A transmatch needs two knobs, because the Smith Chart is a target with a bulls eye around the center, and not a line. However, a simple folded dipole has a length adjustment, but that's it. Fortunately, EZNEC[4] facilitates getting the real part close to 50 Ω by adjusting parasitic lengths and spacings, and then tuning out reactance by adjusting the folded dipole length.

In practice, the antennas are designed using EZNEC with the constraint that the boom length is nearly equal to the reflector length. Lengths of the elements are adjusted for pattern and a real part of the drive impedance near 50 Ω. With only four elements and a fixed boom length, just the driven element and first director positions are variable, and manual optimization of the E and H patterns while watching the drive impedance goes quickly.

## Optimizing with EZNEC

EZNEC is now mature enough that the simulated antenna may be constructed with as much care as possible. The software allows easy exploration of a 1/16-inch error in the first director length, for example. These antennas all use wood booms, so the conducting boom effect is zero. For different parasitic element diameters, change the diameter in the simulator, observe the pattern, and then change the design frequency up and down until the pattern looks good again. For example, if the original design was optimized at 144 MHz and with a new element diameter it looks better at 142 MHz, then the lengths are a bit long — by about 1.5%. Trim the element lengths by 1.5% and check the pattern again. The design will need to be tweaked a bit more to simulate an impedance near $50 + j\,0$ Ω, but after a few designs with the limited number of variables in a four element square Yagi, your design intuition starts to kick in. EZNEC is the best video game ever, and time spent at the basic levels pays off with longer Yagi designs.

The 4-element Yagi is useful on its own, but it is also a good starting point as the launch for a long Yagi or as the feed for a shallow dish. They may be combined in arrays, but one more element on a significantly longer

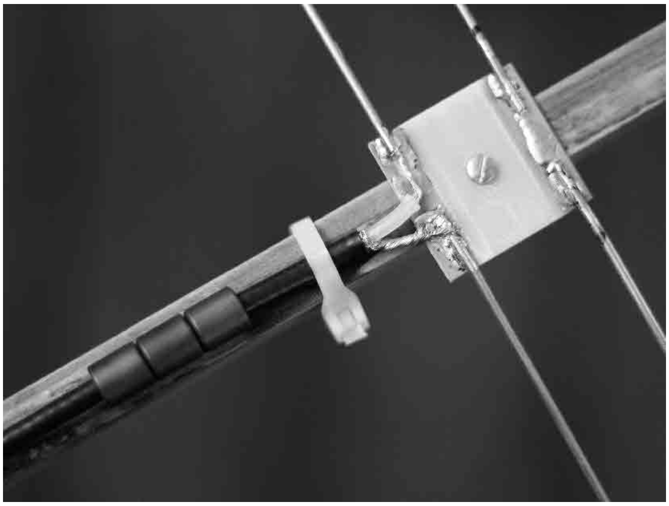

Figure 2 — Detail of a prototype 2 m square Yagi folded dipole feed, shows how length is adjusted by unsoldering half sections of the folded dipole and trimming. After the optimum length is determined, a one piece folded dipole is cut to length so that this adjustable feed may be used for the next prototype. Note the common mode choke ferrites.

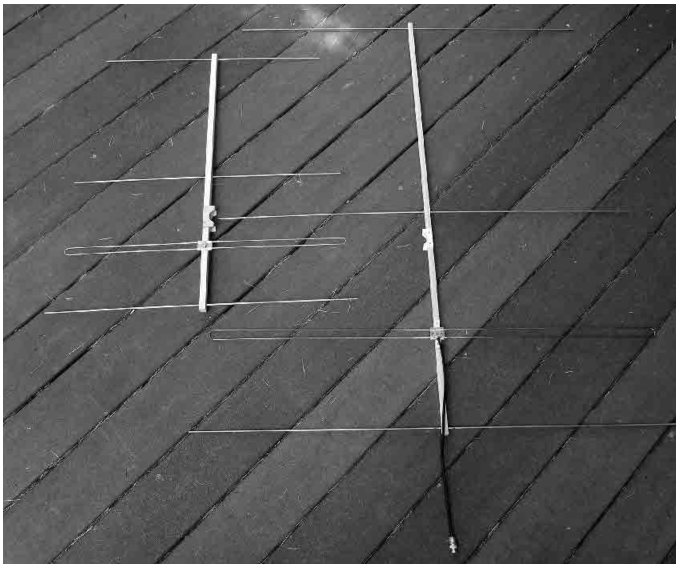

Figure 3 — The 222.1 MHz and 144.2 MHz optimized square Yagi antennas. Construction is ultra-light, with 1/8-inch parasitic elements, #12 AWG copper wire folded dipoles, and wood booms. The 222.1 MHz boom is 3/4-inch clear fir, and the 144.2 MHz boom is 5/8-inch laminated from two pieces of fir, glued with Titebond® III.

boom achieves nearly 2 dB more gain.

The three 4-element Yagi examples at 144, 222 and 432 MHz may be constructed using the dimensions in Tables 1, 2 and 3 respectively without further adjustments or simulation. Experiments with insulated #12 AWG house wire instead of bare #12 AWG copper wire for the folded dipoles suggest that insulated elements need to be about 1% shorter than bare elements, nearly 1/2 inch at 144 MHz. The dimensions in the tables are for bare #12 AWG copper wire folded dipoles.

If you have a convenient way to measure SWR or return loss at the weak signal calling frequency, it is useful to attach the folded dipole driven element to the boom with a cable tie so it may be slid back and forth a bit on the boom. If sliding it toward the 1st director improves the match, then the driven element needs to be shorter, and vice versa.

## Antenna Patterns

Since sliding the folded dipole back and forth on the boom is much easier than trimming it to length, it is useful to know what happens to the antenna pattern when the return loss is adjusted by moving the folded dipole rather than adjusting its length. The table has dimensions adjusted for an E plane pattern (Figure 4) with low and equal side and back side lobes, typically 28 dB below the main lobe. Moving the folded dipole back toward the reflector typically nulls the back lobe, but the back side lobes increase. Moving toward the 1st director increases the back lobe. Neither has much impact on forward gain. Most of the change is in the imaginary part of the drive impedance, so adjusting the length of the folded dipole will achieve the same result without changing the clean E and H patterns shown in Figures 4 and 5. All three antennas have nearly the same patterns.

Tables 1 to 3 show the element lengths and spacing along boom, with the reflector position as the "0.000 inch" reference point. These dimensions are taken directly from EZNEC simulations, hence the number of significant digits. In the simulations, 1/32-inch change in element length makes a difference at 144 MHz, so careful cutting is expected. The numbers for the folded driven element "Distance Along Boom"

**Table – 1.**
**Square Four Yagi element dimensions in inches for 144.2 MHz.**

| *Element* | *Reflector* | *Folded Driven* | *Director* | *Director* |
|---|---|---|---|---|
| Distance along boom | 0.000 | 7.500; 8.500 | 20.000 | 42.000 |
| Length | 41.000 | 38.250 | 38.125 | 35.000 |

**Table – 2.**
**Square Four Yagi element dimensions in inches for 222.1 MHz.**

| *Element* | *Reflector* | *Folded Driven* | *Director* | *Director* |
|---|---|---|---|---|
| Distance along boom | 0.000 | 5.250; 5.875 | 13.000 | 27.250 |
| Length | 26.625 | 24.750 | 24.5625 | 21.500 |

**Table – 3.**
**Square Four Yagi element dimensions in inches for 432.1 MHz.**

| *Element* | *Reflector* | *Folded Driven* | *Director* | *Director* |
|---|---|---|---|---|
| Distance along boom | 0.000 | 2.6833; 2.9833 | 6.666 | 14.000 |
| Length | 13.666 | 12.750 | 12.440 | 11.000 |

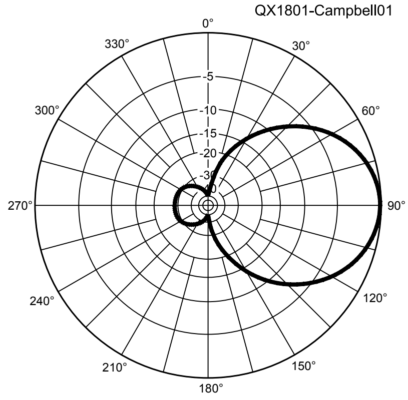


**Figure 4 — EZNEC simulated square Yagi E-plane azimuth far field pattern.**

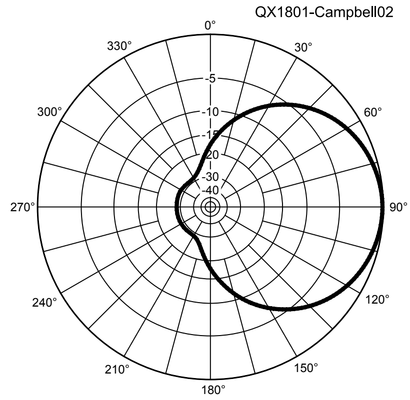


**Figure 5 — EZNEC simulated square Yagi H-plane azimuth far field pattern.**

are for the two sides of the folded dipole in the simulation. "Distances along boom" are less critical. Experiments in EZNEC are encouraged, with these lengths and positions as a starting point.

These antennas are optimized for the weak signal calling frequencies. The 50 Ω match is narrow band at the weak signal calling frequency on each of the bands, and they don't perform as well in the FM portions of the 2 m and 70 cm bands.

A version of this article appeared in the *Proceedings of the 51st Conference of the Central States VHF Society*.

*Rick Campbell, KK7B, began building radios as a young boy and was licensed at age 14. He followed his interests into advanced degrees and a long career in experimental physics, radio science and electronics. Rick has worked as a US Navy Radioman, surface physicist at Bell Labs, RFIC designer at TriQuint and Analog Devices, and developed THz instrumentation at Cascade Microtech. From 1983 to 1996 he served on the faculty of Michigan Tech University, and since 2012 he has been a faculty member at Portland State University. He is co-author of the reissued classic book Experimental Methods in RF Design with Wes Hayward and Bob Larkin. He published a series of No-Tune Microwave Transverters with Jim Davey, and developed and published the seminal R2 series of direct conversion receivers and transmitters. The antennas described here were initially part of a new course in antenna theory and design Rick will be teaching again this Spring. He also enjoys performing and teaching alternative styles on alternative musical instruments, and simply messing about in boats.*

## Notes

[1]The 1952 Vincent Black Lightning (motorcycle) represents a pinnacle of engineering, immortalized by Richard Thompson in a song describing an all-too-mortal young man who is carried away by "Angels on Ariels," leaving young Molly his Vincent to ride. The Ariels ridden by the angels are *Ariel Square Fours*, true engineering marvels and evidence that the arts, creativity, and engineering were perhaps more closely entwined in the mid-20th century. So also the Johnson Ranger front panel...Clegg color schemes...I rest my case.

[2]R. Quick, W4RQ and K. Siwiak, KE4PT, "Does Your Antenna Need a Choke or a Balun?", *QST*, Mar 2017, pp 30-33.

[3]J. Stanley, K4ERO, "Controlling Unwanted Feed Line Resonance in VHF Vertical Antennas", *QST*, Nov 2016, pp 33-36.

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

# Upcoming Conferences

## 2018 SARA Western Conference

Palo Alto, California
March 23 – 25, 2018
**www.radio-astronomy.org**

The 2018 SARA Western Conference will be held at Stanford University in Palo Alto, California on Friday, Saturday and Sunday, March 23 – 25, 2018. The meeting will include a visit the Kavli Institute for Particle Physics and Cosmology (KIPAC).

**Call for papers**: Papers are welcome on subjects directly related to radio astronomy including hardware, software, education and tutorials, research strategies, observations and data collection and philosophy. If you wish to present a paper please email a letter of intent, including a proposed title and abstract to the conference coordinator at **westernconf_at_radio-astronomy.org** no later than December 31, 2017. Full copies of the presentations should be submitted by March 1, 2018, for inclusion in the proceedings.

Be sure to include your full name, affiliation, postal address, and email address, and indicate your willingness to attend the conference to present your paper. Submitters will receive an email response, typically within one week.

**Presentations and proceedings**: In addition to presentations by SARA members, we plan to have speakers from the Stanford University faculty, and possibly KIPAC. Papers and presentations on radio astronomy hardware, software, education, research strategies, philosophy, and observing efforts and methods are welcome. Formal proceedings will be published for this conference. If presenters want to submit a paper or a copy of their presentation, we will make them available to attendees on CD.

**Basic schedule**: Our first day will include a visit to the KIPAC facilities at the Stanford Linear Accelerator Center (SLAC). The next two days' meetings will take place on the Stanford University campus and will include presentations by members and guest speakers. A board meeting for the Society will also be held during the conference.

**Getting there**: Fly into the San Jose or San Francisco airport and rent a car to drive to Palo Alto. It is also possible to use CALTRAIN to get from the San Jose or San Francisco airport to Palo Alto, but you would still need a car to get from the hotel to the meeting site at Stanford University.

**Registration**: Registration for the 2018 Western Conference is just $60.00 US. This includes snacks and lunch on Saturday and Sunday. Breakfast should be eaten at the hotel. Payment can be made through PayPal, **www.paypal.com** by sending payment to **treas_at_radio-astronomy.org**. Please include in comments that the payment is for the 2018 Western Conference. You may also mail a check payable to SARA Treasurer, c/o Bill Dean, 2946 Montclair Ave., Cincinnati, OH 45211. Please include an e-mail address so a confirmation can be sent to you when we receive your payment.

**Hotel reservations**: Marriott Courtyard Palo Alto Los Altos, 4320 El Camino Real, Los Altos, CA 94022. Tel.,(650) 941-9900. (Group Rate $139 per night plus taxes = $154.56 per night). Last day to book is March 2, 2018.

**Additional Information**: Additional details will be published online. Please contact conference coordinator David Westman if you have any questions or if you would like to help with the conference: **westernconf_@_radio-astronomy.org**.

## The 2nd Annual Utah Digital Communications Conference

Sandy, Utah
March 28, 2018
**utah-dcc.org**

The 2nd Annual Utah Digital Communications Conference will be held March 24, 2018.The conference will be a fusion of Amateur Radio communications and Maker topics. Amateur radio is the pioneer of digital modes. This conference will focus on the Amateur Radio hobby that surrounds utilizing digital modes. Current emerging topics such as digital modes for emergency communications and building your own components. If you have questions please email **UtahDCC@gmail.com**

**Registration**: Registration will open in the fall (2017) using the EventBrite system. Cost is $15 per person pre-registration (before February 24, 2018, $20 after).

Prefer Cash/Check/PayPal? Email **utahdcc@gmail.com** for instructions. Check website for further information.

## 2018 Southeastern VHF Conference

Valdosta, Georgia
April 26 – 29, 2018
**svhfs.org/wp/**

The 2018 Southeastern VHF Society Conference will be in Valdosta, Georgia, Hosted by the Suwannee Amateur Radio Club and Down East Microwave Inc, the festivities will start on Thursday afternoon, April 26 and continue through Sunday morning, April 29. The main conference will be held in the Holiday Inn Hotel and Conference Center located only seconds from I-75 in Valdosta, Georgia.

As usual, the conference will offer presentations, antenna and equipment testing, along with a chance to get together with other Society members through the weekend.

We expect to conduct the Best Paper and Presentation competition, and will offer some great prizes along with the prizes at the Banquet. There will be a family program this year that will visit sites in Southern Georgia and Northern Florida.

Conference and Banquet pricing will be announced at a later date, along with the opening of registration and any other special events that may become available throughout the process. A call for papers will be announced in late fall. Check the website for details.

## 2018 Central States VHFS Society, Inc. Conference

Airport DoubleTree Hotel,
Wichita, KS
July 26-29, 2018
**www.2018.CSVHFS.org**

**Call for papers**: Papers are being solicited for publishing in the *Proceedings of the 2018 Central States VHF Conference* on all weak-signal VHF and above Amateur radio topics, including: antennas: including modeling, design, arrays, and control; test equipment: including homebrew, commercial, and measurement techniques and tips; construction of equipment such as transmitters, receivers, and transverters; operating, including contesting, roving, and DXpeditions; RF power amps, including single and multi-band vacuum tubes, solid-state, and TWTAs; propagation, including ducting, sporadic E, tropospheric, meteor scatter, etc.; Pre-amplifiers (low noise); digital modes, such as WSJT, JT65, FT8, JT6M, ISCAT, etc.; regulatory topics; moon bounce (EME); software-defined radio (SDR); and digital signal processing (DSP).

Topics such as FM, repeaters, packet radio, etc., are generally considered outside of the scope of papers being sought. However, there are always exceptions. If you have any questions about the suitability of a particular topic, contact **wa2voi@mninter.net**.

You do not need to attend the conference nor present your paper to have it published in the *Proceedings*.

**Deadline for receipt of papers for** inclusion in the Proceedings is Tuesday, May 15, 2018.

Complete information, including a style guide, can be found on the Central States VHF Society, Inc. website.

# 2017 *QEX* Index

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