PVRC Newsletter April 2017

President's Letter – Bud W3LL

Hello PVRC,

it sure looks like we're heading into the last of the prime time contest season. As I look at my calendar I see a dearth of major league contests. I do see a note on the calendar for this month to put the 6M antenna back on one of the 80M 4-SQ towers. Is it wishful thinking that I can get the 6M country count up from 70 to 100 this year as the sunspots go into their waning years? Yes, avid contesters are also DXers as a spinoff of contesting.

So what have we been doing this month to appease the membership?

The 20 FOR 20 Award is rapidly drawing to a close. For me I've been procrastinating and now find that all remaining 5M contests must be entered or I'm out of the running. How about you? Are you keeping track on the PVRC website to see where you stand?

It'll be a big letdown if nothing follows the 20 FOR 20 Award. But, lament not. Seriously, do you think this productive group of officers would have nothing up our sleeves. Fear not. I'm super excited about what follows in the shadow of the 20 FOR 20. As you know all our creations have a genesis. All it takes is a creative person to apply it. Just like engineers do in turning interesting discoveries into useful items for the benefit of mankind. In this case, it's for the benefit and amusement of we contesting hams. I'm sworn to secrecy but you know how that goes – can't trust anyone to keep a secret these days. But I can give a few hints. The PVRC Activity Award – OOPS, might have multiple tiers of difficulty. It might be just be like winning the Olympics and getting to wear a medallion around your neck or some other place. Maybe it's gold or silver or it could be bronze. Only thing missing is a pedestal and someone playing the national anthem – well maybe we can do that too. Who knows with this group? Too bad many of our ingenious award programs have a twilight ending. How about a perpetual award which starts annually each year just like the CQ Marathon. Could be. We'll just have to wait and see. Or, is this all just an April Fool's joke that I'm playing?

Many of us have been spending a lot of time for the past two months working on another big project. This includes your officers, trustees and chapter chairmen. It's our biggest undertaking yet. We're cashing in all our "atta boy" chips in hopes that you'll help us make the functioning of the club better. Seriously, that's all I can say about it. Stay tuned to the PVRC reflector in the coming weeks.

Lastly, we're getting closer to the PVRC reunion date of Saturday 05 June. We'll be using the Reunion to celebrate our 70th anniversary. Secretary Tim N3QE has been working very hard to integrate the format into the N1MM SS format. More to follow from Tim.

The starting gun for WPX SSB is approaching as I write this column. So it'll be a less verbose column this month.

Now onto the Sports Pages.

Last month's sports pages went to press before everyone reported their CQ 160 SSB and NAQP RTTY scores. First, for the CQ 160 SSB addendum: Let the record show that K3ZM put up the top US score in SO HP by a huge margin with 945 QSO's for 367,000 points. K3ZO made the top 10 in that category with 527 for 104K. The crew at WA3EKL (WA3EKL, K0OO, KB3VQC, N3DPB) made 363 QSO in CQ 160 SSB for 50K points, and also put in a full-time effort in NAQP RTTY with 541 QSO and 68K points. Quite a weekend at WA3EKL! PVRC activity in NAQP RTTY was up over last February with 56 operators, including several M/2 entries. NA3DX also did a real M/2 from NA1DX.

Lots of PVRC players in ARRL DX SSB, despite some very difficult conditions. Team W3LPL finished second this time around, continuing the recent trends to victory by W3LPL in the CW contests and K3LR winning in SSB. W4RM looks like the NA winner in M/2 HP with an impressive 2404 QSO/2.7M effort. WA3EKL also put up a good score in M/2 with 1391/1.4M. K3ZU took the top spot in NA M/S HP with 2075/2.2M. The top PVRC SO U HP score was by N4RV (with N4RA and WB2ZAB not far behind) and K3ZO led the PVRC pack in SO HP. PVRC member N3KS's TI5W station, operated by M0DXR took third overall in SO HP in the battle of the dual CQ'ing SO2R operators. KQ4LA made the top ten in NA in SOAB LP. AA4NC down at 5K0N made some 753 stations very happy to get the San Andres QSO on 15M after he had operated at 5J0NA on Saturday – all with 100W to an OCF dipole that wasn't supposed to work on 15M. W3LL appears to have taken the No. 1 NA spot on SOSB 80 HP with 388 QSO and 132 mults.

The NC QSO party brought out some PVRC types in the southern end of the circle. Team NC4KW (at N1LN) put up the top score overall at 423K, making 1,450 QSO (that's a LOT of QSO's in a state QSO party!). N4CW made an impressive 751 QSO in Mobile Mixed LP, and W4FS (at N3YDU) made 304 QSO in SO mixed HP. PVRC'ers W4WWQ, N3KN and N4GU also reported for duty in this one.

The Spring Stew Perry brought out some good efforts. K2AV made the top ten in SOHP and K7SV sits in third place among SOLP entries. N3HEE, N1LN and AA3S all made more than 100 QSO in SOHP. N3HEE and AA3S both operate with antenna spaces of about a quarter acre, showing that you can be effective on 160 in a limited space. And K4FTO made 66 QSO operating LP with a 31-ft. vertical. AA4XX has a commanding lead in SO QRP with 123 QSO for a score of 843. That score would have been top 10 in both SOHP and SOLP and was the top reported PVRC score. How did Paul do it? Could be a good topic for a newsletter article (hint, hint).

There were 25 participants in the Russian DX Contest, up from 18 last year. Big scores from N4AF (1218 QSO/2.6M in SO CW HP) and W3LL (958 QSO/1.2M in SO SSB HP) were good for second and first places in the US in their categories, respectively. Others with over 500 QSO's included K0OO M/2 at WA3EKL (first place in US), N3QE, N4CW, NR4M, K7SV (first place in US in SO CW LP), K3ZU (second place in US SO Mixed HP).

The PVRC RTTY die-hards showed up for the BARTG RTTY Contest. W4TMO (at W4AAW with PVRC'ers W4TMO, W3UL, W4AAW and W6IHG) put up a fabulous score of 2.4M with almost 1600 QSO. K4GMH took second in SOAB LP with a 700 QSO effort. Others playing in this one were K0OO (at WA3EKL), W1IE (451 QSO in SOAB HP), W7HJ/4, KS0CW, N4CW, NN3RP and W3UL.

That's a wrap for this month. Coming attractions: Next month look for the stories from WPX SSB!

73, Bud W3LL

The Editor's Last Word (Special Up Front Edition) – John K3TN

This is a looong newsletter - Jeff KØZR's filter design piece was so good, I couldn't bring myself to shorten it. You can skip the math and jump to construction and it is still a great article.

Thanks also to Frank W3LPL and Mark N2QT for submissions for this month's newsletter. If you have any good Sweepstakes stories or photos, or anything else – send to jpescatore at aol dot com.

W3LPL's High QSO/$ Amplifiers

Recently on the PVRC email reflector, Phil KT3Y noted: "*After reading Frank's recent CW DX score and station description, I am ready to assert that his homebrew monoband 3-1000Z amps at W3LPL hold the following world records:*

- *most QSOs by band*
- *most QSOs in total across the HF bands*
- *lowest failure rate per QSO*
- *least expensive per QSO by a huge factor*
- *least frills*

If I recall correctly, many of them have been in service for over 40 years as I first tried them at his MD QTH in Crownsville in the mid-1970s. Back then, the power limit was 1KW INPUT and the ARRL DX Test was two full weekends per mode."

Frank replied: "*Interesting observations! My first 3-1000Z amp was built in 1966 as a band switching amp then converted to a single band amp as I built the remaining amps from about 1975-1978.*

Over one million QSOs have been made with those amps. You could count the total number of failures on the fingers of one hand. All of the repairs were completed in less than an hour each (mostly the time required to open up and re-close the amplifier).

The tubes have been replaced two or three times as their thoriated tungsten filament emission weakened after thousands of hours of use. They would have been quite usable on CW for many more years, but low emission tubes produce unacceptably high SSB inter-modulation products."

The 10 meter amp at W3LPL, showing state of the art front panel labeling techniques…

L – The power supply that feeds all 6 bands, and taught WB3JRU (now W3UR) that high voltages and human skin don't do well together.
R – Inside view of the 160M amp warming Frank's basement

(K3TN) My favorite stories about Frank's amps:

- Back in the late 1980s in an SSB contest at Frank's, the power supply started smoking, the big transformer had started to melt. Roger K1DQV and I drove back to Roger's house, dragged one of Roger's big transformers into his truck and drove back to Frank's. Total time to repair and restore high power operation was about 90 minutes.
- In a CW contest during one of the first few years at Frank's current location, I was operating 40 CW overnight in the CQ WW CW during a horrendous rain storm. There was a leak in Frank's roof that came down right on top of the 40M amp. While the amp really didn't seem to mind the cooling flow, I went over to the 160M position and grabbed Frank and we rigged up a plastic sheet to redirect the water onto the floor and I went right back to running EU.

PVRC Mega Meeting in Richmond

Bud W3LL presenting to Bob W3IDT.

The PVRC crowd at the gathering.

Filter Design: Theory and Real World Construction – Jeff KØZR

Since designing my first 2 KW filter for 40m, I have continued to research different circuit and filter techniques. In so doing I have come across, or should I say "remembered", some very powerful circuit transformations which are invaluable to the filter designer. My intention, expressed over the following pages, is to lay the ground work enabling a non-engineer to champion these circuit techniques in their own designs. While I will attempt to provide sufficient background with some examples here and there, it is unlikely I can cover everything to the extent needed. Therefore, making use of some of the cited references and other information available will likely be necessary.

Before getting into the details of the remaining filter designs for use at KØZR, I will develop the necessary background in the following areas:

- Transformer winding ratios and impact on voltage, current, and impedance
- Develop background on definition and use of ABCD parameters
- Use ABCD parameters to derive the first of Norton's impedance transforms (very powerful)
- Summarize many of the transformations that are available to the designer
- Exhibit, step-by-step, the use of some of these transforms in an actual filter design for 20m
- Outline the details for my 80, 20, and 15m bandpass filters
- Provide other helpful insights in an Appendix

Transformer Relationships

Pictured to the right is a typical transformer, wherein the primary has associated with it an impedance Z_1, and the secondary has an impedance, Z_2. The transformer serves to make a "transformation" in impedance as will be described shortly.

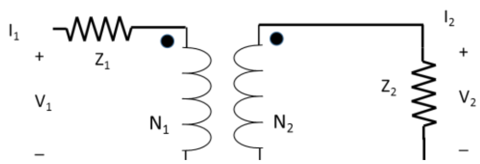


Figure 1

The voltage on the transformer secondary is governed by the number of turns on the primary, N_1, compared to the number of turns on the secondary, N_2. If the primary has more turns, $N_1 > N_2$, the transformer will have a lower voltage on the secondary compared to the primary. Conversely, if $N_1 < N_2$, the secondary voltage is stepped-up, or higher, than that applied to the primary.

$$N_1 > N_2 \qquad \text{Voltage Step-Down at Secondary}$$

$$N_1 < N_2 \qquad \text{Voltage Step-Up at Secondary}$$

There is a direct relationship between number of turns and the voltage observed. In simple arithmetic, this leads to:

$$\frac{V_1}{V_2} = \frac{N_1}{N_2} \quad [1]$$

Turns Ratio and Transformer Currents

The turns ratio also has an impact on the currents observed in the primary and secondary. If we consider the transformer as "ideal", the power input to the transformer must equal the power appearing at the output of the transformer. In terms of the mathematics of voltage and current,

$$v_1 i_1 = v_2 i_2 \quad [2]$$

where the "1" and "2" relate to the primary or secondary voltage or current of the transformer.

We can divide both sides of equation [2] by i_1, then divide both sides of that result by v_2, and obtain a voltage-current relationship for the transformer.

$$v_1 i_1 = v_2 i_2 \qquad \frac{v_1 i_1}{i_1} = \frac{v_2 i_2}{i_1} \qquad \Rightarrow \quad v_1 = v_2 \frac{i_2}{i_1}$$

$$\frac{v_1}{v_2} = \frac{v_2}{v_2}\frac{i_2}{i_1} \qquad \Rightarrow \qquad \frac{v_1}{v_2} = \frac{i_2}{i_1}$$

Principle: If the voltage is increased at the secondary, the secondary current decreases.

[3]

Turns Ratio and Impedance

To understand how the transformer impacts primary and secondary impedances, we begin with the principles just introduced. As already mentioned, transformers are often described in terms of "turns-ratio", that is N_1 compared to N_2. This is commonly written as $N_1 : N_2$. Generally, both sides of this relationship are divided by N_1 or both sides are divided by N_2, giving these modifications:

r:1 where $r = \frac{N_1}{N_2}$ [4]

$1 : r$ where $r = \frac{N_2}{N_1}$ [5]

Both ideologies are found in literature, and therefore one must be prepared to work with either.

A given value of "r" will represent an impedance step-up for one definition of "r" while for the other, an impedance step-down would result.

Form 1: r:1 Ratio

$i_1 v_1 = i_2 v_2$ Divide both sides of this equation by i_1^2 giving $\frac{v_1 i_1}{i_1^2} = \frac{v_2 i_2}{i_1^2}$

Let's look at the left hand and right hand sides of this equation separately.

Left Hand Side: $\frac{v_1 i_1}{i_1^2} = \frac{v_1}{i_1} = Z_1$

Right Hand Side: $\frac{v_2 i_2}{i_1^2}$ Multiply numerator and denominator by i_2: $\frac{v_2 i_2}{i_1^2}\left(\frac{i_2}{i_2}\right) = \frac{v_2}{i_2}\left(\frac{i_2^2}{i_1^2}\right) = Z_2\left(\frac{i_2}{i_1}\right)^2$

The ratio of the currents can be rewritten in terms of the turns ratio, which in this case is:

$$for\ \frac{i_1}{i_2} = \frac{1}{r} \qquad \frac{i_2}{i_1} = r \qquad \boxed{Z_1 = Z_2 r^2 \qquad Z_2 = Z_1\left(\frac{1}{r^2}\right)} \quad [6]$$

Form 2: 1:r Ratio

For this alternate definition of transformer ratio, the voltage and current relationships are flipped. The left hand side of the equation remains the same as before, but the right hand side involving the current ratio of i_2 to i_1 is impacted.

$\frac{v_2 i_2}{i_1^2}\left(\frac{i_2}{i_2}\right) = \frac{v_2}{i_2}\left(\frac{i_2^2}{i_1^2}\right) = Z_2\left(\frac{i_2}{i_1}\right)^2$ Here, for the 1:r definition, $\frac{i_2}{i_1} = \frac{1}{r}$ giving

$$\boxed{Z_1 = Z_2 \frac{1}{r^2} \qquad Z_2 = r^2 Z_1} \quad [7]$$

Summary Table

Transformer Ratio	**Voltages**	**Currents**	**Impedance**
r:1	$V_1/V_2 = r$	$I_1/I_2 = 1/r$	$Z_2 = Z_1/r^2$
1:r	$V_1/V_2 = 1/r$	$I_1/I_2 = r$	$Z_2 = Z_1 r^2$

Table 1

Note: in either case, r can take on the following values: $0 < r < \infty$
{Obviously, 0 and ∞ are impractical values….} See ***Appendix***

ABCD Matrices and Transformations

Two-Port Parameters in General

Throughout electrical engineering it is common practice to use what are termed “two-port parameters” in circuit design. There are several parameter sets in use, each with an accompanying list of pros and cons. For example, z-parameters are used to combine impedances which occur in series, while y-parameters, often called admittance parameters, find use in dealing with parallel or “shunt” components. The general area of RF and microwave design relies heavily on two-port s-parameters, known as “scattering parameters”. Table 2 introduces several of these parameters. We will make use of ABCD parameters.

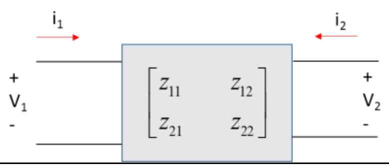
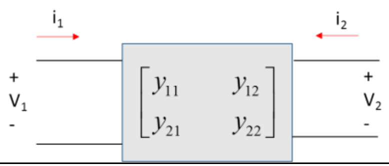
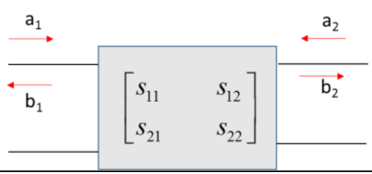
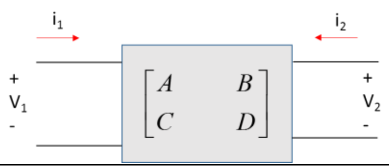
Parameter Set	When Used
 $\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} z_{11} & z_{12} \\ z_{21} & z_{22} \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \end{bmatrix}$	Impedance or “Z” Parameters Used at lower frequencies where impedances are more easily measured. Suited for combining R, L, and C when in series
 $\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} \begin{bmatrix} v_1 \\ v_2 \end{bmatrix}$	Admittance or “Y” Parameters Used up to several hundred MHz In deriving each element, is easier to get good “shorts” to ground rather than “opens” as required with z-parameters
 $\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$	Scattering or “S” Parameters Used from tens of MHz to greater than 50 GHz. Based on incident and reflected voltage waves
 $\begin{bmatrix} v_1 \\ i_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} v_2 \\ -i_2 \end{bmatrix}$	ABCD Parameters Provides for cascading elements. One multiplies, in order, the ABCD matrix for each successive element in the chain

Table 2

As mentioned, we will make use of ABCD parameters to derive a powerful transformation known as the Norton transformation. You may have heard of Norton and Thevenin equivalents, wherein a current or voltage source with its accompanying resistor may be changed into the other format. This is different than what will be discussed here.

The Norton transforms I will be developing are used to change sections of a filter to higher or lower impedance levels, or introduce additional capacitance at nodes in a filter where stray capacitance could damage the response of the filter. In this latter case, it is better to have incorporated into a design a known, deliberate amount of capacitance rather than be at the mercy of whatever stray capacitances are in the design due to the physical layout. This is not a significant problem at HF, but increasing frequency elevates this potential area of concern.

The ABCD Parameter Set

Why ABCD parameters? ABCD parameters allow one to directly matrix-multiply the ABCD matrix of each component that is cascaded in series, arriving at the ABCD matrix for the total cascade. This is far simpler than had a different two-port set been used.

As you will soon see in developing the Norton transform, a shunt capacitor, series capacitor, and additional shunt capacitor are all cascaded together and equated to a single series capacitor and transformer. The same analogy for inductors applies. The ABCD parameters, sometimes called "transmission parameters", are ideally suited for this cascading.

Figure 2 below shows schematically what we wish to develop. It is desired to develop the element values which will make a series capacitor and ideal transformer electrically equivalent to a "pi" section of capacitors. In so doing, the impedance level within a circuit, in our case a filter, can be increased or decreased, or a shunt capacitor introduced to deal directly with the stray capacitance issue.

The reason one may want to increase or decrease the impedance within a filter is that in so doing, the inductor or capacitor values can be made more achievable. From a manufacturing sense, these transforms also allow one to synthesize a filter with more uniform inductor values throughout, as an example, thus lessening the unique parts count.

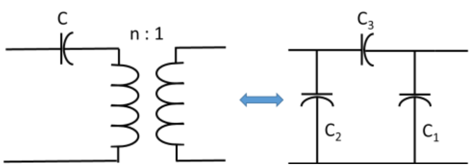


Figure 2

Our objective is to derive the mathematical relationships for these two circuits to the left, making them equivalent over all frequencies. Note the phrase "all frequencies" rather than an "approximation".

This is where we begin. $\begin{bmatrix} v_1 \\ i_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} v_2 \\ -i_2 \end{bmatrix}$ or written out: $\begin{matrix} v_1 = A v_2 - B i_2 \\ i_1 = C v_2 - D i_2 \end{matrix}$. In developing the values for A, B, C, and D, v_2 is set to zero or i_2 to zero and the reduced equations solved. For example, to determine the value for "A", one open-circuits the second port making i_2 equal to zero. When this is done the first equation becomes

$$v_1 = A v_2 + B \times 0 \qquad A = \frac{v_1}{v_2} \quad [8] \quad \text{for the circuit under consideration}$$

This simple procedure is applied, in turn, for each A, B, C, and D for each particular circuit element. This procedure was applied to a shunt capacitor, series capacitor, and ideal transformer resulting in the following ABCD parameters for each of these simple elements.

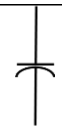
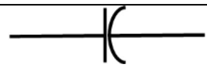
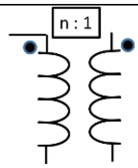
Lumped Element Component	Describing ABCD Matrix
 Shunt "C"	$\begin{bmatrix} 1 & 0 \\ j\omega C & 1 \end{bmatrix}$
 Series "C"	$\begin{bmatrix} 1 & 1/j\omega C \\ 0 & 1 \end{bmatrix}$
 Ideal Transformer	$\begin{bmatrix} n & 0 \\ 0 & 1/n \end{bmatrix}$

Table 3

The beauty of the ABCD parameter set allows us to multiply each of these matrices in succession, setting the two representations equal. Solving the resulting simultaneous equations will give us the Norton transformation for a single series capacitor and ideal transformer. The figures below show pictorially the next steps.

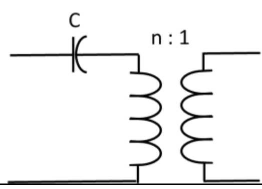
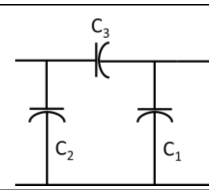
	
$\begin{bmatrix} 1 & 1/j\omega C_S \\ 0 & 1 \end{bmatrix}\begin{bmatrix} n & 0 \\ 0 & 1/n \end{bmatrix}$	$\begin{bmatrix} 1 & 0 \\ j\omega C_2 & 1 \end{bmatrix}\begin{bmatrix} 1 & 1/j\omega C_3 \\ 0 & 1 \end{bmatrix}\begin{bmatrix} 1 & 0 \\ j\omega C_1 & 1 \end{bmatrix}$

Table 4

The left column above is the matrix-multiply of the ABCD matrices for the series capacitor and ideal transformer. The right column, in similar manner, is the ABCD matrix multiplication of the three capacitors in cascade.

Multiplying the matrices for the series capacitor and ideal transformer gives:

$$\begin{bmatrix} n & \dfrac{1}{j\omega C n} \\ 0 & \dfrac{1}{n} \end{bmatrix} \qquad [9]$$

Multiplying the three ABCD matrices together for the pi section of capacitors gives:

$$[10] \qquad \begin{bmatrix} 1+\dfrac{C_1}{C_3} & \dfrac{1}{j\omega C_3} \\ j\omega C_2 + j\omega C_1\left(1+\dfrac{C_2}{C_3}\right) & 1+\dfrac{C_2}{C_3} \end{bmatrix}$$

In methodical order, we now equate each of the four terms in the first matrix to the second matrix.

(1,1)	$n = 1 + \frac{C_1}{C_3}$ [11]
(1,2)	$\frac{1}{j\omega C_S n} = \frac{1}{j\omega C_3} \rightarrow C_3 = nC_S$ [12]
(2,1)	$C_2 + C_1\left(1 + \frac{C_2}{C_3}\right) = 0 \rightarrow C_2C_3 + C_1C_3 + C_1C_2 = 0$ [13]
(2,2)	$\frac{1}{n} = 1 + \frac{C_2}{C_3}$ invert and sub for C_3 from (1,2): $n = \frac{nC_S}{nC_S + C_2}$ [14] $n^2C_S + nC_2 = nC_S \qquad C_2 = C_S(1-n)$

Table 5

So far we have relationships for C_3 and C_2. Substituting (1,2) into (1,1) gives the final needed relationship for C_1.

$$n = 1 + \frac{C_1}{C_3} \qquad n = 1 + \frac{C_1}{nC_S} \qquad n = \frac{nC_S + C_1}{nC_S} \quad n^2C_S - nC_S = C_1 \qquad [15]$$

$$C_1 = n(n-1)C_S$$

Norton Transformation for a Single Series Capacitor and Ideal Transformer

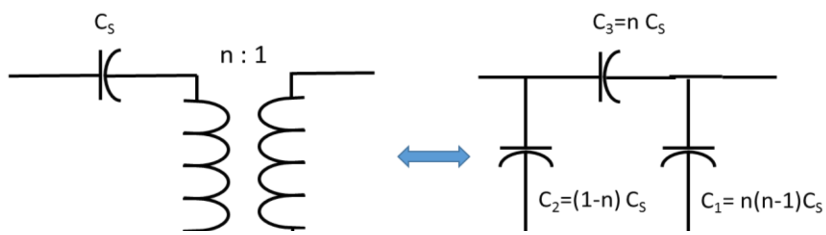
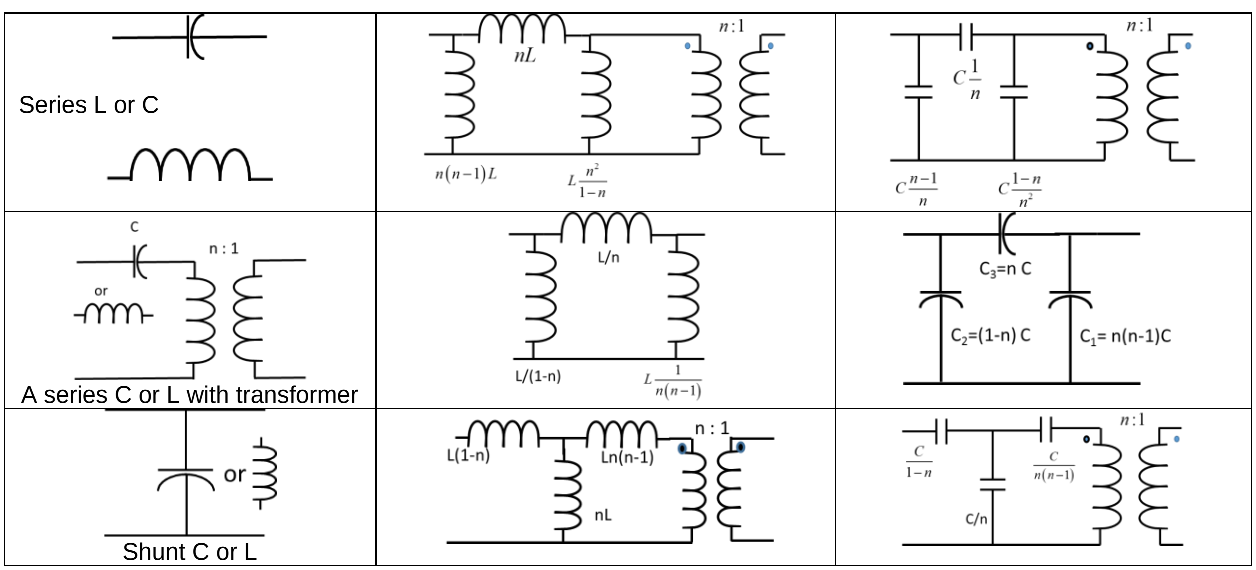


Figure 3

A set of Norton Transformations is nicely summarized[i] below. Other useful transformations are found in the references. In the *Appendix* I address a turns ratio 1:n and its impact on the formulae.



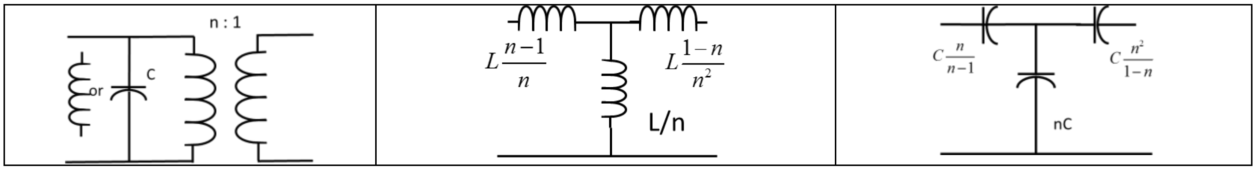


Table 6

The left-hand column in Table 6 depicts the transformation for L or C instances. Looking at the first row (previous page), if one has a series C you use the transformation containing the C values in the third column. If you were working with a series L, one would use the transformation containing L values in the second column.

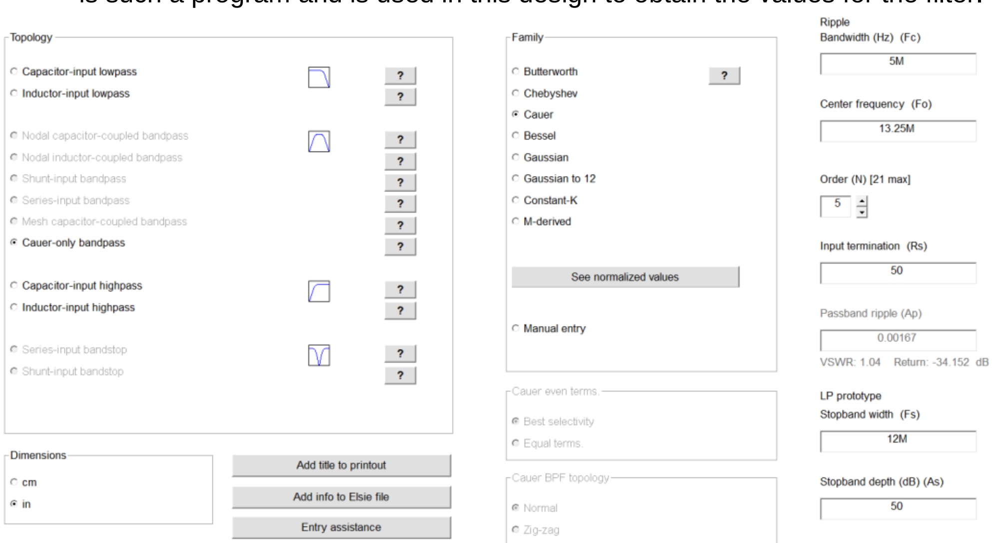
An important point is the following: performing these transformations results in negative component values. These will occur <u>on the high Z side of n</u>. The negative value is combined with other Ls and Cs in the circuit (in other words, other Ls or Cs as appropriate must be present to make this combination; if they are not you cannot use the transform in this way). Obviously, one must perform the proper transformation so the resulting negative values can be absorbed with a positive component value. {This will limit you sometimes in whether you can do a step-up or step-down in impedance}

Filter Design in General

A number of parameters are available to the filter designer. The different filter families, i.e. Butterworth, Chebyshev, and Cauer (elliptic) to name a few, have their individual characteristics. Of great importance is selection of the proper passband width (for bandpass filters), the width of the stopband, and in the case of Cauer types, what ultimate attenuation is desired in the stopband. Additionally, the filter's characteristic impedance may be designed for lower or higher than 50 Ω, subsequently employing impedance matching networks at the input and output of the filter if necessary. While this increases total part count, sometimes the benefit in reduced voltages or currents makes the additional impedance transformations worthwhile. And of course, there is the application of available transforms, such as the Norton transforms[ii], albeit this is considerably more complex than aforementioned methods.

20m Elliptic Bandpass Filter Design

Objective: Attain passband insertion loss ~ 0.2 dB while placing notches as close as possible at 7 and 21 MHz. Stopband attenuation of 50 dB was selected, resulting in a 5th order requirement. A multitude of software tools are available to the filter designer, with some being completely free or "close to it". ELSIE is such a program and is used in this design to obtain the values for the filter.



In selecting these design parameters, a considerable amount of time was spent trading f_c, stopband bandwidth, passband bandwidth, etc in order to strategically place the notches of the elliptic response. Furthermore, it is known that the narrower the passband the higher some of the currents will be in parts of the filter; one needs to minimize this.

Figure 4 ELSIE Design Window

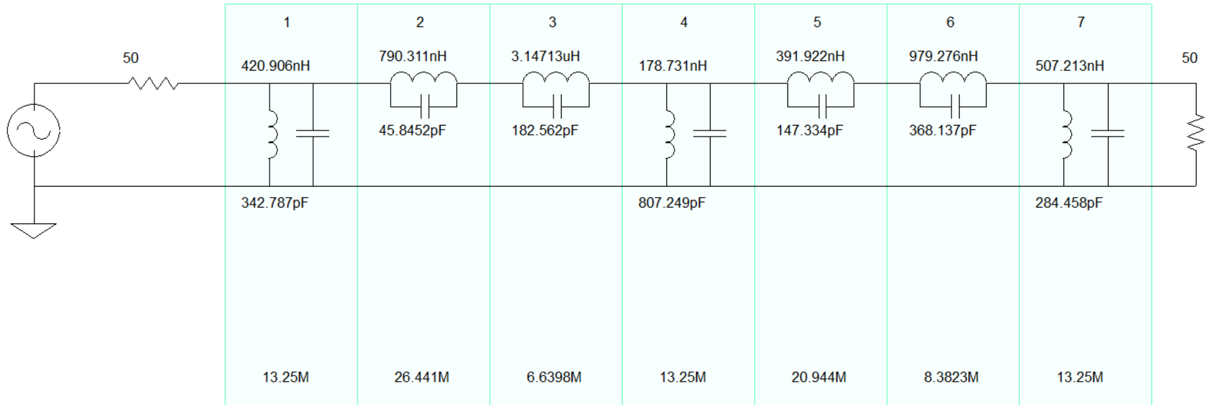


Figure 5 Output from ELSIE

The circuit for this filter was augmented with coil resistance netting Qs of 400. A SPICE-like program, Simetrix, was then used to analyze the filter for voltage and current when subjected to 1500 watts at the input. For the seven different resonators above, the accompanying currents were determined as shown below in Table 7.

L_1	11.5	L_5	14
C1	9.7	C_5	6.25
L_2	11.1	L_6	3.6
C_2	3.0	C_6	11.3
L_3	1.9	L_7	8.1
C_3	9.9	C_7	9.6
L_4	21.2		
C_4	25		

Table 7 RF Currents in ELSIE 20m Filter

Clearly some of these RF currents are noteworthy and require special attention if one does not want a catastrophic failure when operating at high power.

To withstand these high currents, multiple paralleled capacitors are required. More will be said on this topic later, but the CDV series of silver mica capacitors is documented as handling under continuous conditions, 5-6 amps at HF. My original designs had planned on using the thought-to-be less expensive route of MLCCs (multi-layer ceramic chip capacitors), however further research into the minimal available data indicated a higher risk than I was willing to accept. This design will use CDV16 and CDV19 capacitors purchased from Mouser Electronics.

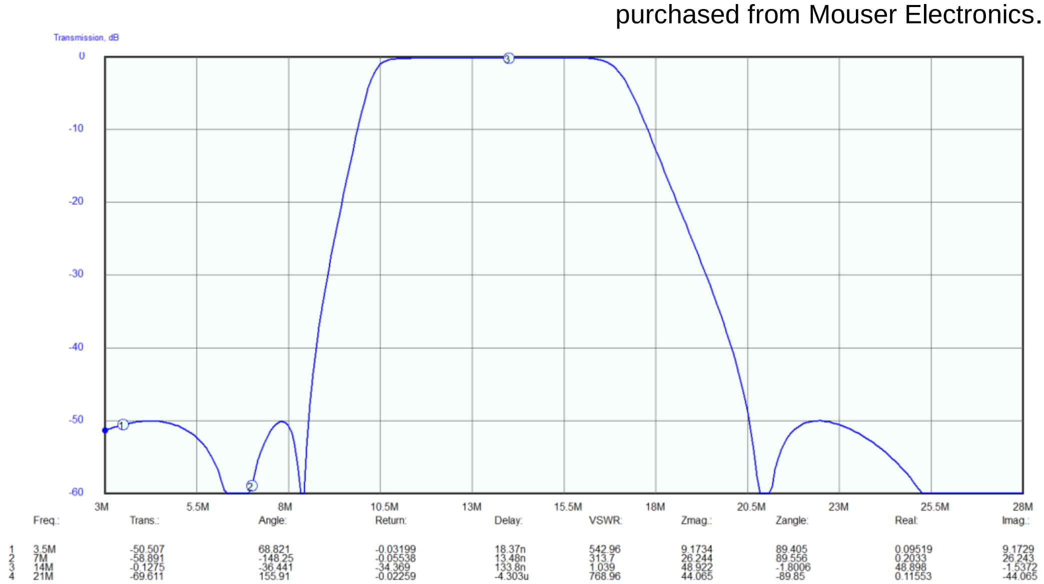


Figure 6 ELSIE Simulated Filter

Due to the low frequencies in use here, as well as the air-suspended nature of filter construction, I do not believe stray capacitance occurring at two of the floating nodes is a serious concern. Of greater concern is the high current in L_4-C_4. In an attempt to reduce this current, and simultaneously diminish any potential stray capacitance occurring at the two high impedance points (between L2||C2 and L3||C3 as well as between L5||C5 and L6||C6 of Figure 5), Norton transformation techniques were applied to the ELSIE synthesized filter.

In use of the transform, it is the case that negative values of capacitance and inductance result. This is expected and used to our advantage. The direction of the transform (impedance step-up or step-down) must be done in such a direction that where the negative component values occur, other positive components are present to absorb them. More will be said about this shortly.

The transform is first applied to the second series parallel resonator composed of L_3 and C_3, values L=3.147u and C = 182.6p. An additional transform directly attributable to the work already discussed, is used as it applies directly to the case of a parallel L-C followed by a transformer. The details of the derivation follow a similar venue as used in the earlier development for the first Norton transform.

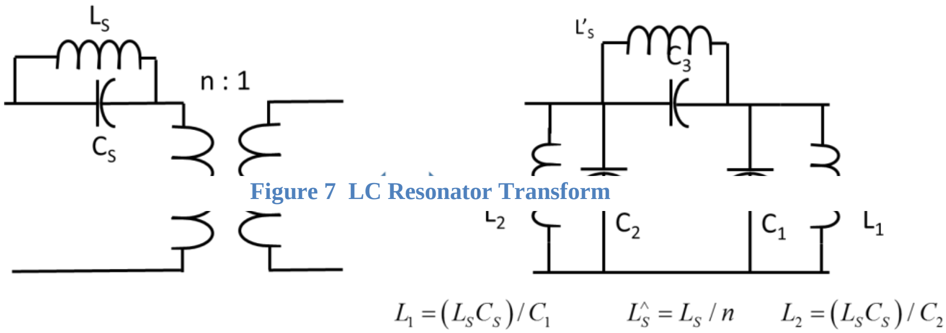


Figure 7 LC Resonator Transform

To eliminate any possibility for ambiguity, the beginning filter schematic is shown below in Figure 9.

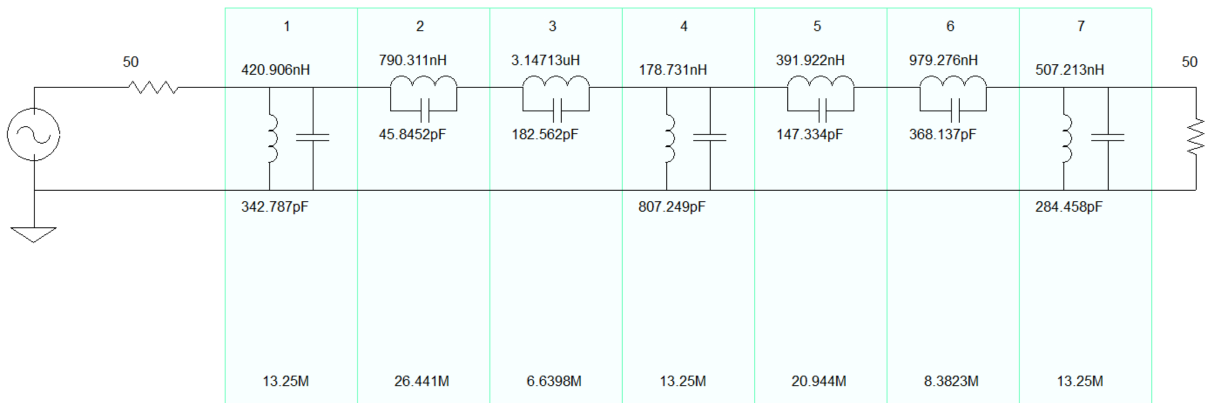


Figure 8 Beginning Filter - Same as Figure 5

The transform just presented (Figure 7) will be applied to resonator #3. In so doing the impedance at that point (and beyond to the right) is elevated to 100 ohms. Therefore, the 4^{th} and 5^{th} resonators must be impedance scaled by a factor of 2 as well. Resonators 6 and 7 remain unchanged because an impedance step-down transformer will be applied next at resonator 5. More discussion on this follows. The equations presented with the transform are programmed into a simple Excel sheet. The results of those calculations are the following:

Ls	3147.00	
Cs	182.56	
n	0.71	
Cap1	-37.8094	
Cs_prime	129.0894	pF
Cap2	53.4706	
L1	-15195.0602	
Ls_prime	4450.5301	nH
L2	10744.5301	

A quick review of the equations shows that units cancel out, thus one can enter the inductance of L_S in terms of uH or nH, and similarly for the capacitors. We see immediately that C_1 and L_1 have negative values, and as such must occur to the right side of the transform so they can be later combined with resonator 4. (This also tells one a step-down is not realizable) As a result of this first transform application, raising the impedance in this area to 100 Ω, the following schematic of Figure 10 applies. Included in this schematic is the impedance scaling of the shunt resonator #4 occurring in Figure 8.

Figure 9 Transform Calculations

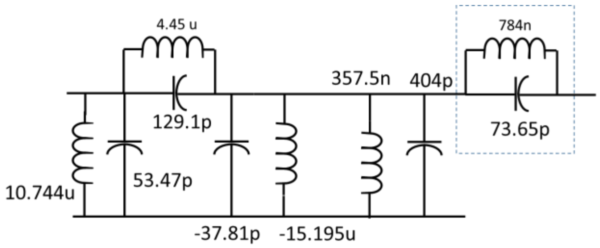


Figure 10 Application of First Transform

To the left in Figure 10 are the results of:

- Applying the transform to the 3.147 uH and 45.8 p cap
- Doubling the impedance of the original shunt 807pF and 178.7 nH inductor
- Impedance scaling by X2 the next parallel resonator, originally 391.9 nH and 147.3 pF (before the scaling), outlined in a "dotted" box in Figure 10

It is extremely easy to forget that before applying the second transform (yet to come), one must impedance scale those Ls and Cs to the right of the first transformed area. Also, note that no combination of negative elements has been done yet because additional shunt elements will appear when the reverse transform (going from 100 Ω back to 50Ω) is applied to the 784n||73.65p combination in Figure 10.

As you can see, there is a lot going on here. The only way you will fully understand this process is by drudging through the calculations yourself.

Now we are ready to apply the second transform. The Excel spreadsheet for this is shown below in Figure 11.

Ls	7.84E+02	
Cs	7.37E+01	
n	1.414213562	
Cap1	43.15	
Cs_prime	104.19	pF
Cap2	-30.52	
L1	1338.10	
Ls_prime	554.26	nH

Figure 11 Second Transform

The second transform is now applied, reverting from 100 Ω back to 50Ω. As shown to the left, there are negative values of L and C which occur on the left side of the transformed result. These additional elements are now added to those just presented schematically in Figure 10.

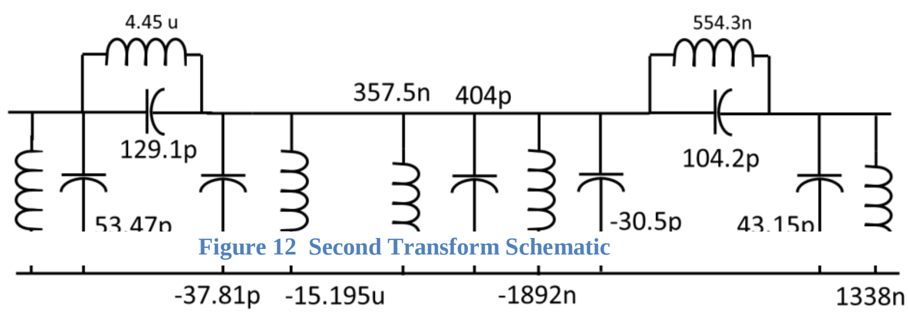
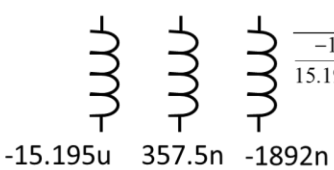


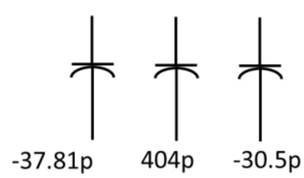
Figure 12 Second Transform Schematic

The finish line is in sight! All that is left to do is combine the inductors, yes both the positive and negative ones, and the same for the capacitors. The inductors combine as resistors in parallel and the capacitors like resistors in series.



$$\frac{1}{\frac{-1}{15.195}+\frac{1}{0.3575}-\frac{1}{1.892}} = 454n$$

Figure 13



$$-37.81 + 404 - 30.5 = 335.7n$$

Figure 14

Transformed 20m Filter

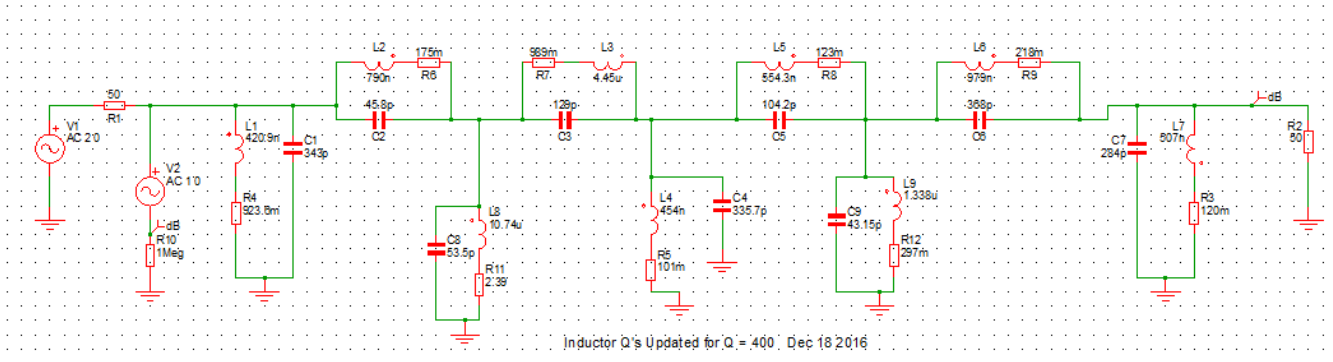


Figure 15 Results of Double-Transformed 20m Filter

Transformed 20m Filter Response

Over 65 dB of loss at 21 MHz; 58 dB at 7 MHz Insertion Loss ~ 0.25 dB

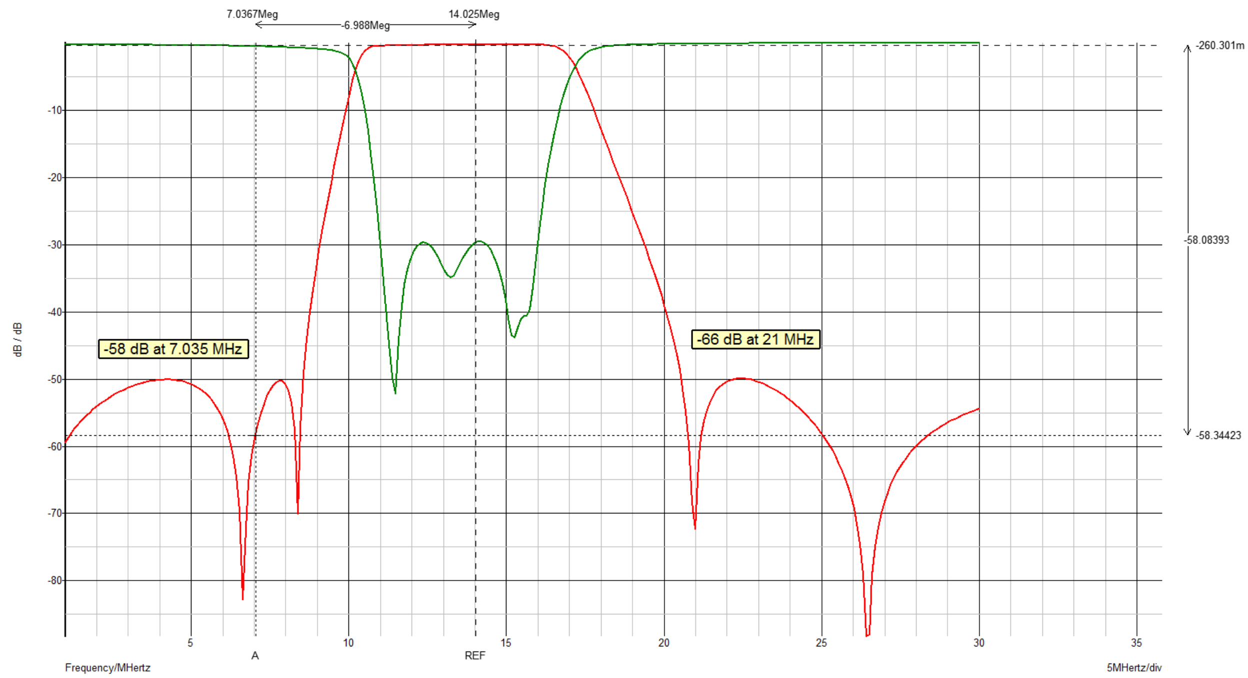


Figure 16 Simetrix Simulation With Finite Inductor Qs

L_1	11.5 (10.0)	L_5	14 (11.5)
C1	9.7 (11.5)	C_5	6.25 (5.4)
L_2	11.1 (11.5)	L_6	3.6 (4.3)
C_2	3.0 (3.4)	C_6	11.3 (11.9)
L_3	1.9	L_7	8.1
C_3	9.9	C_7	9.6
L_4	21.2 (13.5)		
C_4	25 (16.4)		

Table 8 Comparison of Currents in Two Designs

While there is not a one-to-one correspondence in part labeling between the original and transformed filters, some of the "like parts" currents are shown to the left in parenthesis. Of particular note is the troublesome 25 amps in the former design. It is now ~ 16 amps after the transformation – a huge improvement. Those current values within parentheses in Table 8 represent the improved design values.

Don't Plug in the Soldering Iron Yet

While the filter just designed has improvements, primarily in the high currents within the middle parallel resonator, all is not well when one considers the component values required. For example, the second shunt resonator requires almost 11 μH and one of the series inductors requires over 4 μH. While one might get away with this, the physical size may become unacceptable, or possibly the interwinding capacitance may lead to an unacceptably low self-resonance for the 11 μH inductor; a real possibility. Therefore another design attempt was made on the filter beginning with the schematic of Figure 15. The Norton transform is now performed on the #2 and #8 resonators, first transforming down to 25 ohms then back up to 50 ohms. The inductor sizes are hence reduced and the parts spread is diminished. No inductors are larger than 1.6 uH, giving added confidence that inductor self-resonances will not be problem. However, with these positive attributes comes a huge negative of some 30 amps of current circulating in new resonator #4, the middle resonator shunted to ground. This occurs because the impedances were reduced by 50% in doing the transformation, resulting in a current doubling.

No combinations of transforms worked to remove this problem that I could identify. I considered quite a number of options far beyond what is discussed herein, and invariably negative valued components would end up in the wrong place where they could not be absorbed by other components in the vicinity. I decided to apply another solution.

One of the virtues obtained by the transformation between resonators #2 and #6 is, I believe, a diminished voltage level across the circuit nodes compared to the original ELSIE design. This allows one to consider a powdered-iron transformer where the center shunt resonator is located. SPICE evaluates this node as being 216 V rms for a 1500 watt input. HOWEVER, this 216 V rms is across each winding, so the voltage appearing across C_4 in Figure 17 is closer to 1.2 KV. The 216 V rms is used in the formulas for toroid B_{MAX} available from Amidon (and repeated in the index for completeness) and found to be very compatible with our design goals. The lower V_{RMS} per winding, higher frequency, and use of a quadrifilar winding is the ticket and solves the high current issue.

The use of the quadrifilar winding elevates the impedance 16 times (to 400 ohms), thus the current through the 100 pF capacitor is now only some ~ 7.4 amps rms; a fantastic improvement of 4X over the original design of Figure 5 that had C_4 current max of 25 amps.

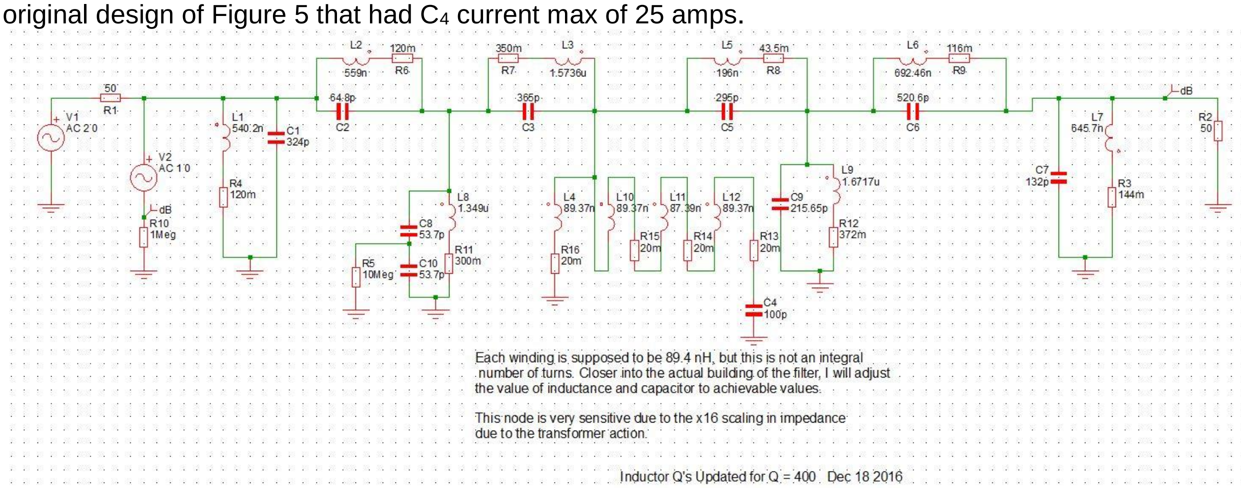


Figure 17 Second 20m BPF Filter Attempt

In SPICE, every node must have a DC path to ground, even if through a 10 megohm resistor. This, in fact, occurs between the two series capacitors above in Figure 17 and is a requirement of the SPICE algorithms.

For completeness and further documentation of this second design attempt, the input and output transformations as calculated in a simple Excel spreadsheet are included below.

Input Transform

	n:1 definition								
Ls	790.31					64.83462077			
Cs	45.85					Cs_prime			
n	1.41421								
						558.8343		Zo =	0.5
						Ls_prime			
Cap1	26.8554								
Cs_prime	64.8346	pF		-1907.9795	-18.9896		26.85538	1349.1453	
Cap2	-18.9896			L2	C2		C1	L1	
L1	1349.1453								

Figure 18 First Transform

Output Transform

Cs	736.27						520.62			
n	0.707107						Cs_prime			
			Zo =	0.5			692.46			
Cap1	-152.49						Ls_prime			
Cs_prime	520.62	pF								
Cap2	215.65				1671.74	215.65		-152.49	-2364.19	
					L2	C2		C1	L1	
L1	-2364.19									
Ls_prime	692.46	nH								
L2	1671.74									

Figure 19 Second Transform

The filter response is shown below in Figure 20. It meets most of the design parameters established at the beginning of this process.

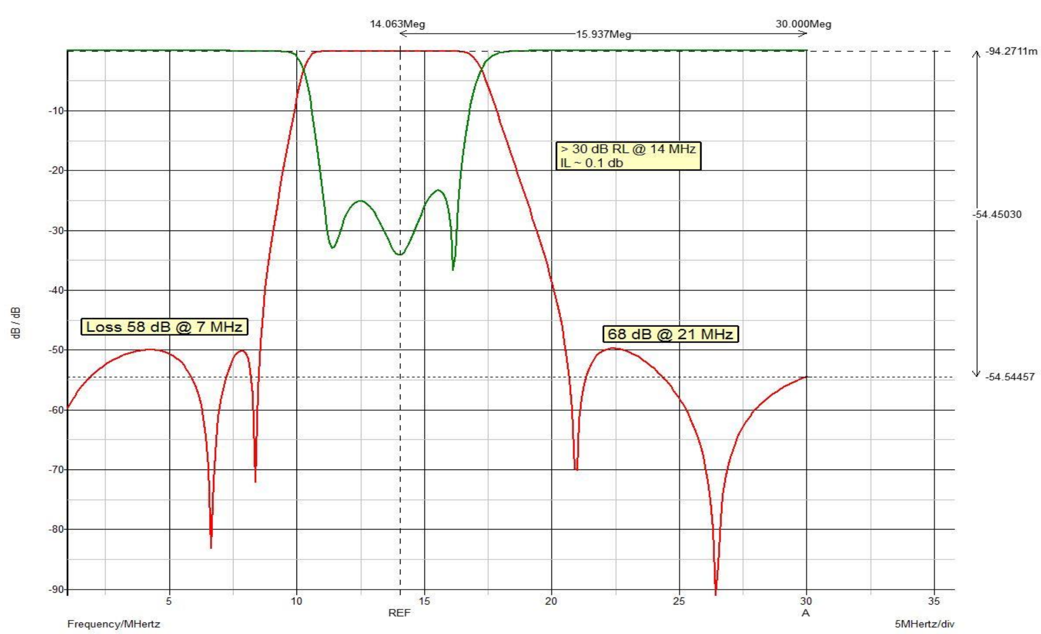


Figure 20   Second 20m BPF Design Simulated in Simetrix

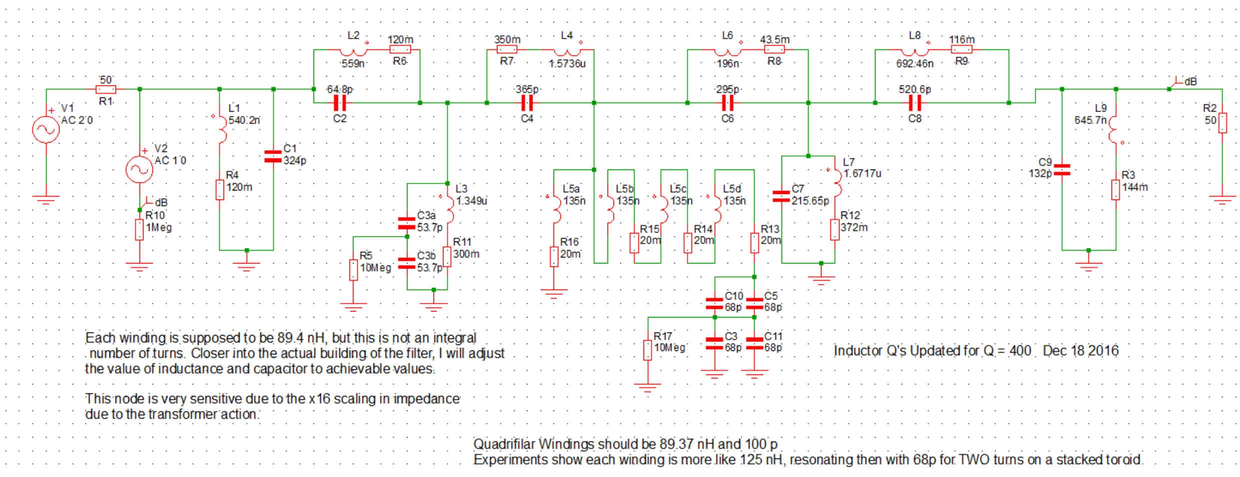


Figure 21 Finalized Schematic of 20m BPF

Figure 21 is essentially a repeat of Figure 17 with two differences. First, the toroid and its accompanying capacitor C_5 have been adjusted to what I believe will be closer to the actual implementation in terms of amount of inductance. Secondly, all the L and C designators have been relabeled, starting from the left, to improve readability. The component peak currents and individual L-C resonant frequencies are shown in Tables 9 and 11. The "Res #" is for the L-C pair, beginning from the left and proceeding to the far right. Critical node voltages as a function of load impedance are shown in Table 10.

L_1	8.4	L_6	20.6
C_1	11.8	C_6	9.6
L_2	11.8	L_7	2.7
C_2	3.4	C_7	7.8
L_3	4.6	L_8	4.5
C_3	1.3	C_8	12.5
L_4	3.0	L_9	6.7
C_4	14	C_9	4.5
L_5	6.7		
C_5	6.7		

Table 9 Peak RF Currents by Component

Load Z	L_3	L_5	C_5	L_7
25	510	220	850	390
50	485	270	1.1 K	400
75	520	320	1.3 K	430
100	560	360	1.43 K	460
100 @ 2 KW	645		1.6 K	

Table 10 Critical Node Voltages as Loads

$$I_{Pow\ Lev} = I_{1500W}\sqrt{\frac{Power}{1500}} \quad [16]$$

Res #	Res Freq
1**	12.03
2	26.44 ++
3	26.44 ++
4	6.641 ++
5**	13.309
6	20.93 ++
7	8.382 ++
8	8.382 ++
9**	17.239

Table 11 Resonator and Resonant

In tuning the filter, the transmission zeros, designed with a double "++" sign Table 11, should be adjusted first. They control the deep notches in the stopband of Figure 20. To the extent possible, further tuning of the filter should attempt to leave these untouched once set. Those resonators with a double asterisk ** in Table 11 are adjusted last and will have a strong impact on the passband return loss.

Construction Considerations

The 40m bpf constructed a few weeks ago was done wholly with MLCCs (multi-layer ceramic chip capacitors). MLCCs are more generally available, are considerably less in cost, but do come with some challenges. I have listed what I see as the major ones below, which include:

1) ESR (equivalent series resistance) and RF current capacity details are difficult to find, especially at RF
2) Difficult to really know how much margin one has in current capacity due to lack of information
3) Current capacity is generally < 1 amp per MLCC, thus requiring many capacitors in parallel, driving up capacitor costs

4) Most MLCCs, except the most expensive ones (from American Technical Ceramics), stipulate more exotic soldering methods, such as solder reflow, with very tailored and highly controlled, short duration temperature exposures, etc. In lieu of this type of manufacturing, i.e. just using a soldering iron, one subjects unknown thermal stresses on the MLCCs, perhaps leading to premature catastrophic failures. Some things can be done to ameliorate this concern, such as heating the PCB before mounting, putting the MLCCs on a hotplate prior to mounting to lessen the thermal shock, etc. In my mind today, the MLCC route comes with too many process control issues which I do not care to solve, at least right now.
5) Another process related concern. The MLCCs should be mounted with solder paste. This has a shelf life of approximately six months (you might find a product with longer shelf life). The most problematic to me is the concern that the resistivity of the solder paste may be higher than that of a soldered connection. If this were true, in the presence of some of the high RF current, elevated heating at the junction between the MLCC and PWB could occur, adding further thermal stresses to the MLCC during actual operation. Some PMP (processes, manufacturing, parts) expert could better address this. In recent weeks I have found that CDV16 and CDV19 silver mica capacitors have HF RF current capacities around 5-6 amps, and while more expensive than most MLCCs, in the long run are likely better than the MLCC route, and less expensive, in my mind. So, for this design, CDV16 capacitors are used almost exclusively.

Construction Details and Measured Results

Below in Figure 22 is the 20m BPF populated PCB. It is not yet in a chassis; that will be the final step. In the center you are able to see a stacked pair of T-130 Mix 17 powdered iron toroids. Two turns was "too much" but that was about the minimum I could reasonably do. Had I had some more suitable toroids available (without ordering more), I could have likely gotten closer to the design value of 1.43 μH. Unfortunately a higher inductance closer to 1.8 μH or 1.9 μH was realized (resonant frequency measurements with a known capacitance). Placing this higher inductance value in the Simetrix model decreased the stopband attenuation some (47 dB instead of the as-designed 50 dB) but I elected to take this "hit". Recall that had the original design value of 89.37 nH been used, some 30+ amps would have been circulated in this parallel resonator rather than ~ 7 amps.

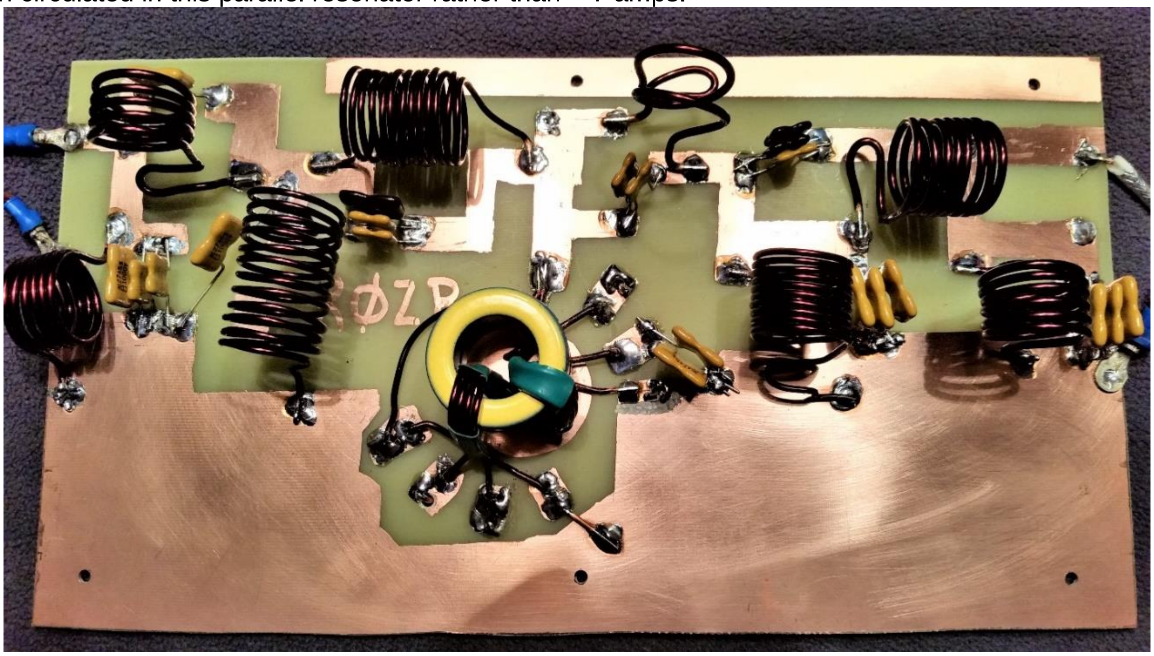


Figure 22 Populated 20m PCB

The inductors in Figure 22 which carry the higher currents are formed from AWG 12 permaleze wire while the others use AWG 14. The "loose arrangement" of the toroid and its own self resonance is believed responsible for eroded stopband performance above 30 MHz; more on that shortly. My greatest concern is performance on 40m and 15m since in SO2R operation, the neighboring bands will likely find the greatest utilization. Such a caveat is not applicable in a multi-multi station, in which case all stopbands are important.

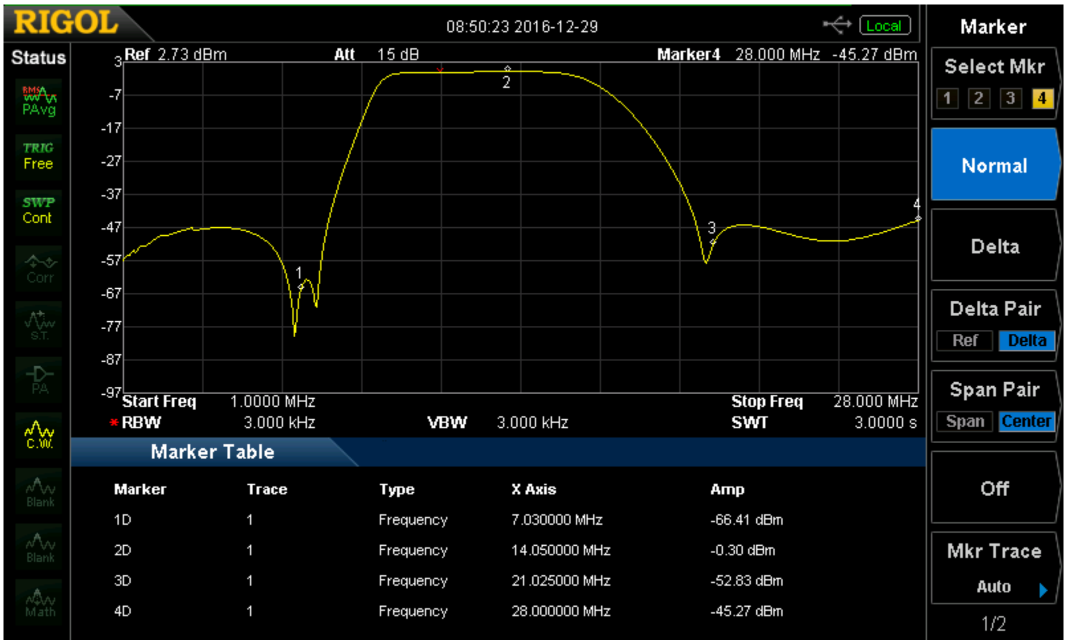


Figure 23 MHz to 28 MHz Sweep With Marker Table – Unmounted PCB

In Figure 23 it is noticeable that the stopband is edging up higher at 28 MHz. That is shown further in Figure 24 below. Take note, however, that this is without the PCB mounted to its aluminum plate.

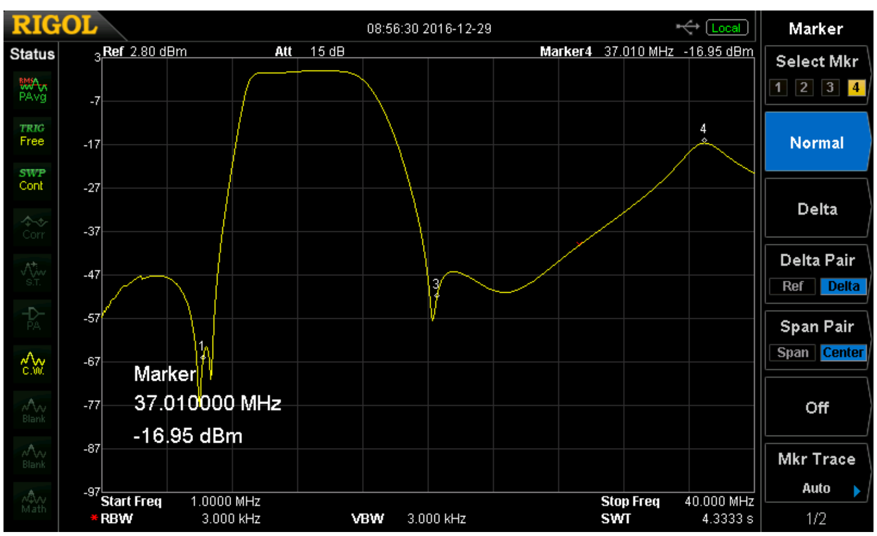


Figure 24 Full Bandwidth Span – Unmounted PCB

While I will be the first to admit I do not like this "second return" around 37 MHz, in actuality it is far from the frequencies of operation and I want to preserve the 4X reduction in current afforded by use of the toroidal transformer.

During adjustments on the filter, I rather “double dipped” in that the pole normally occurring at 8.38 MHz was moved down in frequency to add to the attenuation normally afforded to the 40m band. Figure 24 shows > 65 dB attenuation on 40 m!

Insertion loss has been somewhat difficult to measure, accurately, because it is so low. Several attempts have garnered measurements between 0.1 and 0.15 dB loss. The return loss over the 20m band is also very good, as one would expect for such low insertion loss; return loss exceeds 25 dB across the band.

Final Tweaks Required

Somewhat surprising to me, given my experience on my first 40m BPF (which had no toroids, however), was the fact that when mounting the PCB on an aluminum plate caused some shifts in VSWR and passband. The plate will eventually be enclosed by an aluminum box. The inductance of the toroid increased still further (likely due to “reflected inductance” in close proximity to the aluminum ground plane and interconnecting wires), thus requiring another change in the capacitor to ground off the fourth winding of the cores.

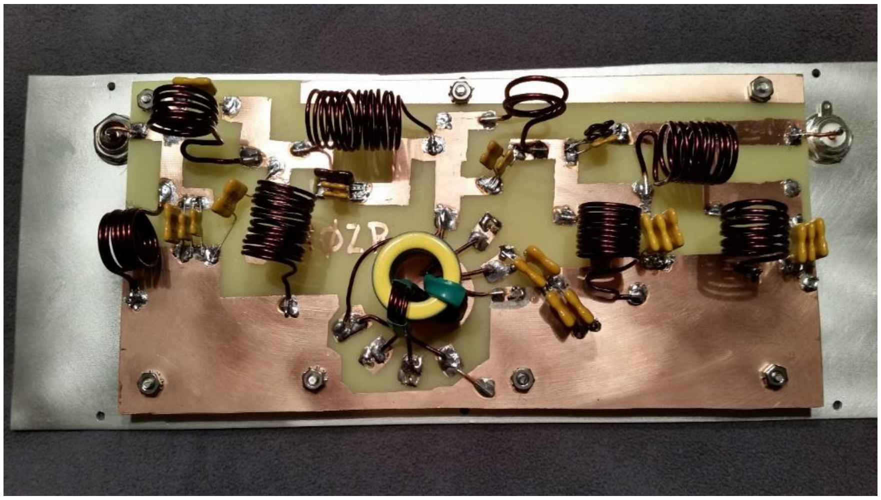


Figure 25 20m BPF Mounted on Aluminum Plate

The overall results of the filter once placed on the aluminum plate are more favorable than presented earlier in Figures 23 and 24; see Figures 26-28. The aluminum plate brings ground closer to the entire PCB rather than just through the copper ground plane of the board itself.

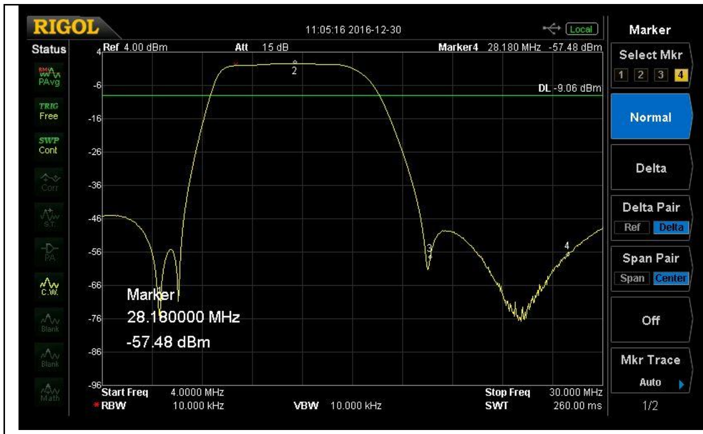


Figure 26 4 MHz to 30 MHz Response

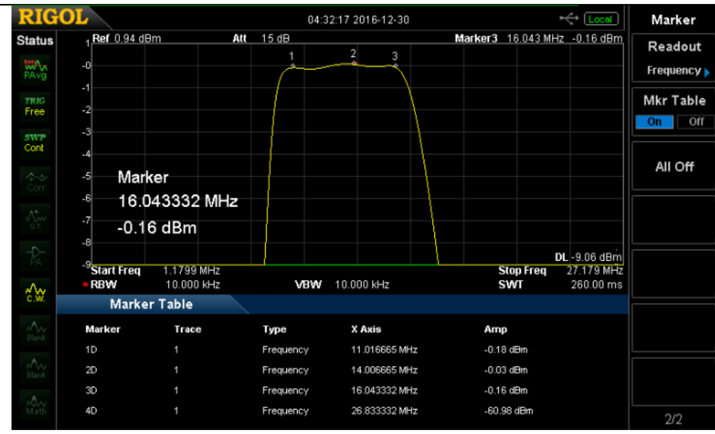
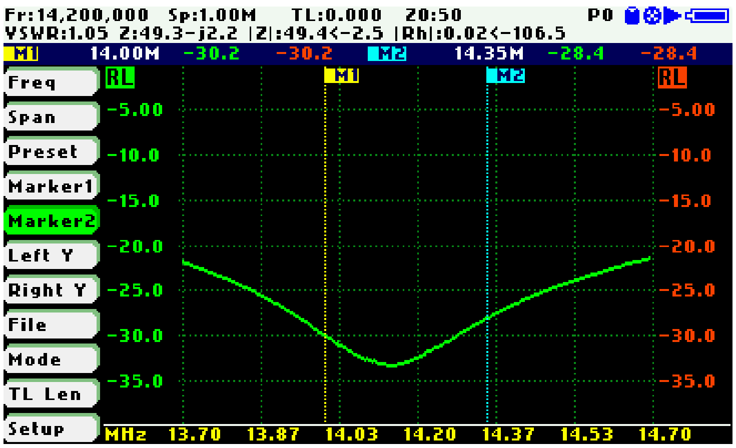


Figure 27 Passband Response on 1-dB Scale



Frequency	Insertion Loss, dB
3.5	~ 50
7.0	65
14.1	0.15
21	58
28	57
	Return Loss
14.0	30 dB
14.35	28.4 dB

Table 12 Electrical Parameters

Figure 28 Passband Return Loss

Other Construction Details and the Smoke Test”

As one would expect, I was unable to get the “exact” capacitor values, thus opting for as close as possible with the CDV capacitors I had on hand. An updated schematic is shown in Figure 29. In preparing for the powered test, I discovered an oversight in the design of the PCB layout. While I have been vigilant regarding each components’ voltage and current requirements, I overlooked the fact that the capacitor on the fourth winding of the toroid actually receives 4X the voltage I was expecting, so instead of the anticipated 216 V predicted by Simetrix, the voltage appearing across the capacitors Figure 21, L5d, is over 1.2 KV. Therefore, in Figure 25 one can see four capacitors to ground off one of the toroidal windings.

The filter was placed immediately after a Tentec 1.2 KW amplifier, operating into a 2 KW dummy load. Operation at several hundred watts revealed no problems. I gradually elevated the power to ~ 1.2 KW and transmitted CW as though I were in “contest mode” for some 5 minutes. I detected only the smallest heating in coils L_1 and L_2. The stacked toroids were as cool as if no power was going through the filter. I will probably install a low-volume fan on the filter just to be on the safe side for the big contests.
In the layout of the filter I was conscious about field interactions between the multiple inductors. Some is unavoidable in a design such as this, but you must attempt to minimize this impact as much as possible. A nice reference on this aspect of the design is that provided by W0QE, “Coupling Between Coils or Coefficient of Coupling”.

The final test was performed with the complete the enclosure for the filter, operating approximately an hour at 1000 watts. Operation with my Alpha 8410 amplifier at the 1500 watt level has revealed no problems whatsoever.

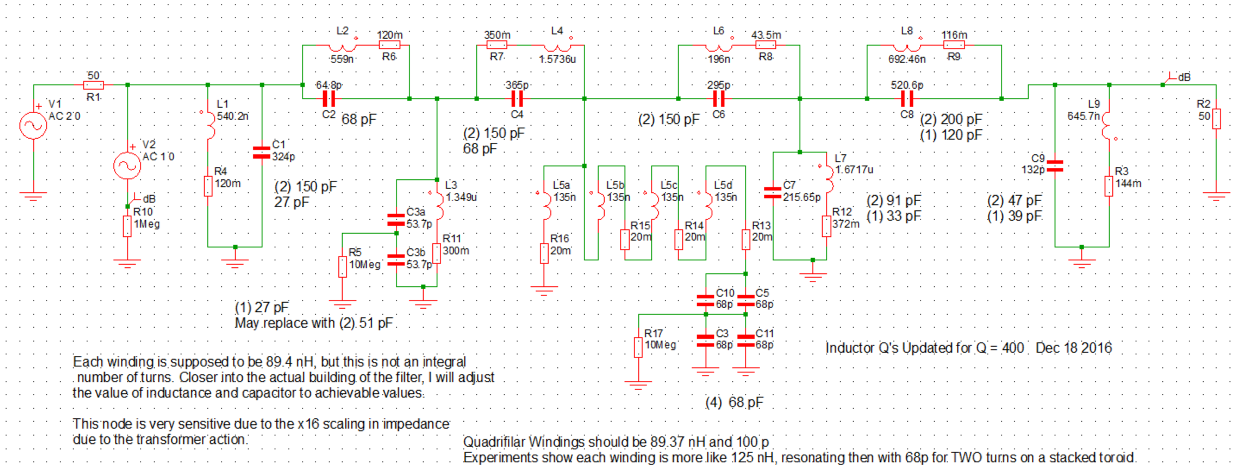


Figure 29 As Constructed Schematic

All capacitors are CDV16 or CDV19 available through Mouser Electronics. The PCB is 1 ounce copper.

Figure 30 BPF in Homebrew Aluminum Housing

When placing the filter in the housing shown in Figure 30, there was no detectable change in passband insertion loss or return loss, nor in the stopbands. The enclosure measures (LWH) 15”x6”x4”.

Other Filters to Design

At this juncture, filters for 160, 80, 15, and 10 remain to be designed and constructed. I am in the midst of building the 80m filter. This filter and likely that for 160m, will be unique from the others in that I believe more toroids will be used, with possibly no toroids in the other filter designs. Tuning is greatly facilitated if one can eliminate toroids from use as they are quite inflexible, physically and electrically, once wound. While I could possibly build the 80m with only one toroid (the resonator to ground in the filter’s center), I plan to use a bifilar T-200 Mix 2 toroid to realize one of the inductors that was rather large; ~ 11 uH. I am striving to keep all filters in the same (LWH) 15"x6"x4" footprint.

I am also planning to make the 160 and 80m filters "brute force capable" of 2 KW or more. It is likely that VSWR may enter in more during a contest where one is stretching the limits, and as we know, amplifiers put out more power lower in frequency and I want to ensure sufficient headroom is available in general.

In Closing
If you encounter mistakes, have helpful suggestions, etc., I would like to hear about it so that I may update this document and hopefully not be a source of confusion for future readers. Currently my email address is kzerozr@gmail.com, but in all likelihood the email associated with QRZ.com will always be current.

Explaining bandpass filters to non-hams:
Separating Good from Bad and Ugly

Membership News – Tim N3QE

PVRC added several members in the few months – please welcome:

- In January: Chip KC3ICH and Susan N3DPB, both in the Annapolis Chapter.
- In February: Steve N3PMF in the Colonial Chapter.
- In March: Don W4BBT in the North Carolina East Chapter, and Chuck K4QS in the Rappahannock Chapter.

Chapter leaders please remember to complete the Meeting Attendance Report. Members can check and update their roster details via the Roster Lookup.

Upcoming Contests and Log Due Dates

Contests This Month

- **Apr 1 – SP DX**
- **Apr 2 - NA SSB Sprint**
- **Apr 8 – Yuri Gagarin DX**
- **Apr 8 – JIDX**
- **Apr 15 – YU DX**
- **Apr 22 – Helvetia DX**

Logs Due This Month

- **Apr 2 – RU DX**
- **Apr 4 – ARRL DX SSB**
- **Apr 9 – NA SSB Sprint**

See WA7BNM's Contest Calendar for more detail and the latest information.

From the PVRC Treasurer – Tom K3AJ

PVRC has chosen not to implement an annual Dues requirement. We depend on the generosity of all our club members to finance our annual budget. In addition, active PVRC members are expected to participate and submit logs for at least two PVRC Club Competition contests per year.

When contemplating your donation to PVRC, each member should consider the benefit you are receiving from PVRC and its many opportunities for your personal growth in our wonderful hobby, then donate accordingly.

Direct donations to PVRC via Credit Card or PayPal may be made by clicking this "Donate" button and clicking the next Donate button that appears on your screen:

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Eyeball QSO Directions

The latest info on local club meetings and get together will always be sent out on the PVRC reflector and posted on the PVRC web site.

NW Region: Meetings are generally held on the third Tuesday of each month at: Chef Lin, 417 S Jefferson St. Frederick, MD 21701 Phone #: 301-620-0664(2675) The meeting begins at 7:00 PM.

Contact: Jim WX3B

DC Metro: Meets monthly the second Monday of each month, except June, July & August). The location alternates between the below MD and VA locations. Pre-meeting dinners start at 6:00 pm and meetings start at 7:30 pm.

VA LOCATION: Anita's, 521 E. Maple Ave, Vienna, VA. Tel: 703-255-1001. Meets at this location during the months of February, April and October.
Contact: Rich NN3W

MD LOCATION: Max's Café. 2319 University Blvd W, Wheaton MD 20902. Tel: 301-949-6297 People usually begin arriving at the restaurant around 6:00. Meets at this location during the months of January, March, May, September and November. Contact: Art K3KU

The Laurel, MD Region: Bill N3XL The PVRC get-together is held at the first LARC meeting each quarter at the clubhouse.

The Annapolis Crew: Dan K2YWE Meetings are held on the 4th Wednesday of each month at Broadneck Grill in Annapolis. We gather at about 5:30 PM and order dinner about 6. We break up usually before 8 PM. E-Mail K2YWE to be put on the e-mail reminder list.

PVRC-NC: The **PVRC-NC East** chapter meetings are held at Manchester's Bar and Grill on the 9100 block of Leesville Rd. in North Raleigh, with "QRM" beginning at 6:00pm and the dinner meeting following shortly thereafter. The meeting is held monthly on the 1st Thursday of most months, cancellations or changes usually announced on the PVRC-NC website.

The **PVRC-NC West** chapter meets the 3rd Monday of each month (except December) at about 7:00 PM at Hams Restaurant, 414 S. Stratford Rd., Winston-Salem on the south end of the Thruway Shopping Center. We meet in the front meeting room of the restaurant. A wide variety of cold 801s and Sports bar menu available. Contact Henry Heidtmann W2DZO, full info at http://www.w4nc.com

Over the Hill Bunch: The group meets for lunch at noon alternately in Maryland at the Sir Walter Raleigh Inn 6323 Greenbelt Rd, Berwyn Heights, MD or in Virginia at Anthony's restaurant in Falls Church. Meetings generally are held on the last Wednesday of the month and are subject to change. Meetings are announced by email.

All PVRC members, non-members interested in membership and guests are welcome. For information contact Roger Stephens, K5VRX, 703-658-3991 for Virginia meetings; or Cliff Bedore W3CB or get on 147.00 for Maryland meetings.

Downtown Lunch Group: Meets: At 12 o'clock noon on the first Wednesday of every month in Downtown DC.

Location: R.F.D. Washington, 810 7th St., N.W. Washington - Metro (Green Line) at Gallery Place Station

Meeting since April 2005, PVRC'ers, contesters, DX'ers, and others coming through this area have been meeting for lunch in the Metro Center area of downtown Washington, DC. If you give him enough notice, regular attendee Fred, K3ZO, may bring you your 3-land QSL bureau cards! We invite and encourage you to join us every month for good food, good company, and good conversation. If you have any questions feel free to email or call W3DQ or our co-convener, Rich NN3W. Hope you can join us!

Southwest VA Chapter: The Southwest VA group meets each Wednesday at about 8:30 AM at Hardees at 20265 Timberlake Road in Lynchburg, VA. This is an informal gathering, but normally has about 10-12 attendees. Contact Mark Sihlanick N2QT, Tel: 434-525-2921

Eastern Shore Chapter: Meets every three months, on the second Saturday of April, July, October and January at noon. In keeping with the tradition established by SK Dallas W3PP we will also meet at the contest station of Eric WG3J one hour before the start of most major contests. Contact Eric Hudson WG3J

Location: Delmar Pizza, north west corner of the intersection of highways 13 and 54 in Delmar, DE

Southern Maryland Chapter: Currently on hiatus, if interested in meeting contact the Chapter Chair, Tom Shelton, ND3N via email or (240) 434-3811

Colonial Capital Chapter: Meets the 2nd Thursday of each month at 8:30 am
Location: Hot Stacks Restaurant, 6495 Richmond Rd, Williamsburg, VA 23188
757-565-1105

Contact: Bill Conkling NR4C

The Tidewater Chapter meets the 3rd Tuesday of every month at Frankie's Place for Ribs located in the Fairfield Shopping Center on the corner of Kempsville Road and Providence Road in Virginia Beach. The meeting starts at 7:00 PM. All amateurs are invited.

Contact either Chapter Chair: Don Lynch, W4YZT, or Ron Young, W8RJL

If you'd like to add or correct a listing, contact K3TN for inclusion in the Newsletter!

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The IC-7300's new RF Direct Sampling System borrows SDR tech to replace the conventional superhetrodyne design with an RF Direct system. The result is a versatile and budget-friendly radio that delivers incredible receiver performance, easy filter adjustments and awesome audio clarity. The radio features an intuitive multi-dial knob and a built-in antenna tuner. It works on SSB, CW, RTTY, AM and FM modes.
More importantly, Icom's engineers were still able to cram it all in a rugged and compact package—perfect for your portable or EMCOMM station. The IC-7300 provides the performance and features of an SDR, yet you still get the familiar tactile "feel" of radio knobs and buttons.
DX ENGINEERING
Wire Antenna Kits
When it comes to Field Day operation, it's tough to beat the portability, versatility and performance of a wire antenna. DX Engineering has put together EZ-BUILD UWA Center T and End Insulator Kits that let you build virtually any wire antenna type—folded dipole, inverted-vee, off-center fed, Windom, Zepp, loops and more. Multi-band operation is also possible. Search keyword EZ-BUILD at DXEngineering.com to see your options.
LBS
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14
21
PerfoBox
HF Multiplexers and Band-Pass Filters
Use Low Band Systems' Multiplexers to connect multiple radios to a single multi-band antenna, so you can use each radio to operate on a different band simultaneously. This reduces equipment installation hassles and saves money since there's no need for extra antennas and coax cables, making them a great solution if you're bringing several radios to Field Day.
Available in duplexer, triplexer and quadplexer configurations, these units work with Low Band Systems' multi-stage band-pass filters to limit the RF to a single band, effectively eliminating most RF interference issues. Sold separately, the filters deliver the isolation demanded by the multiplexer's band input.
DX ENGINEERING
Coaxial Cable Assemblies
Don't wait until the week before Field Day to ensure you have enough cable to build a temporary station. These low-loss cable assemblies are available in standard lengths with DX Engineering's revolutionary new PL-259 connector, featuring the best qualities of both crimp-on and solder-on connectors. Use the online Custom Cable Builder at DXEngineering.com to build assemblies made to your exact specs. DX Engineering's coaxial cable is also available by the foot or in bulk spools.
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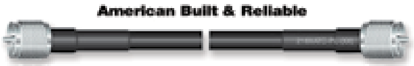

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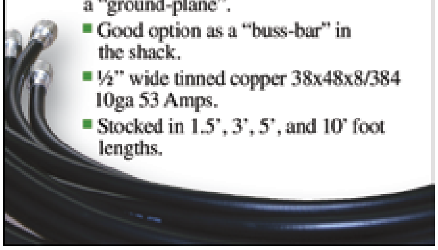

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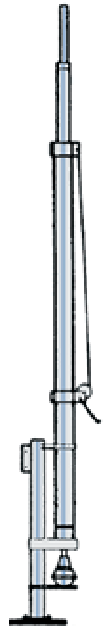

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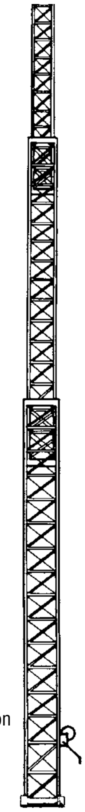

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