

High-Power HF Band-Pass Filter Design

K0ZR covers some of the considerations essential to a successful high-power filter design in an example 20 m band pass filter.

The first step in a filter design for high power and contesting applications, in my opinion, is to leave nothing to chance or hope. At the 1,500 watt level there is little room for mistakes.

Let's consider a filter with a dissipative insertion loss of 0.3 dB. Simple mathematics shows that 100 W of that 1,500 W, will be dissipated in the filter. This heat will age the components more rapidly, may shift the filter's return loss as a function of duty cycle, and may unnecessarily cause thermal stresses to the filter, leading to possible premature failure. Heat is one of the enemies in high power filter design. The filter described here has about 0.1 dB insertion loss.

As a first step, one can carefully design a filter, paying special attention to notch

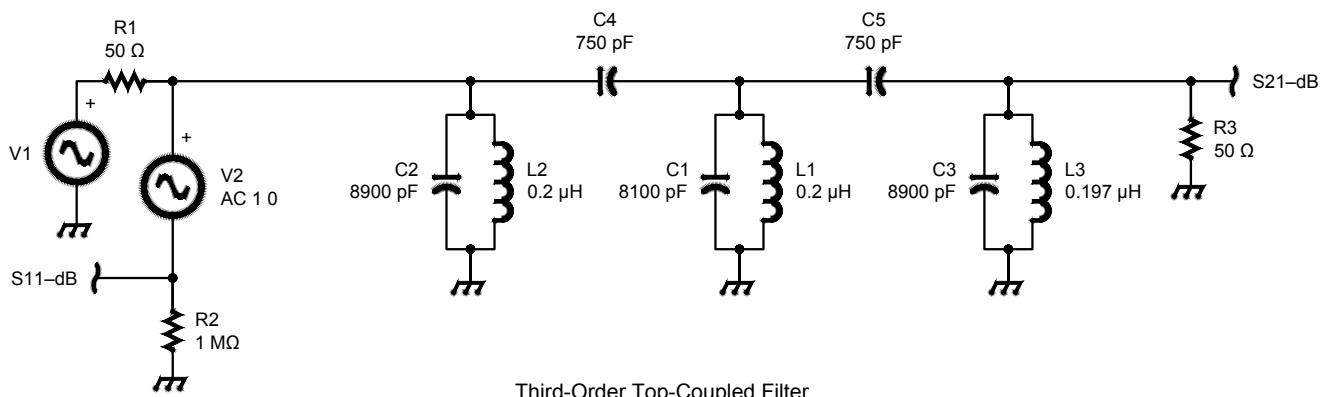
placement for the adjacent contest bands, achieve acceptably low insertion loss, great return loss, thus reaching the point it is ready to build. This is only the first step. Consider the filter in Figure 1, which was designed with the “*q k*” method^{1, 2}. Its response is shown in Figure 2. It has a rather clean passband, good return loss, and a smaller parts count than a comparable *N* = 5 elliptic filter. We discover, however, that at 1,500 W, some RF currents exceed a peak value of 100 A. I do not believe your printed circuit board will handle that current.

This example shows us that we need to consider alternate filter layouts. Knowledge of voltages, currents, parts values, core flux densities, and so on, are needed to avoid a possibly costly mishap. The intent over

these next pages is to cover some of the considerations essential to a successful high-power design, rather than a complete design of a 20 m band pass filter. There are a considerable number of design references available, some of which we will reference here to facilitate your design efforts.

Getting Ready

A certain minimum tool set is necessary to successfully design a filter. A filter design package such as *Elsie*³ is invaluable, and is available for free. A more traditional manual approach⁴ is possible as well. A circuit simulation program such as *LTspice*⁵ or *SIMetrix*⁶ is essential for performing design tradeoffs, especially in ascertaining expected



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Figure 1 — Top-coupled filter.

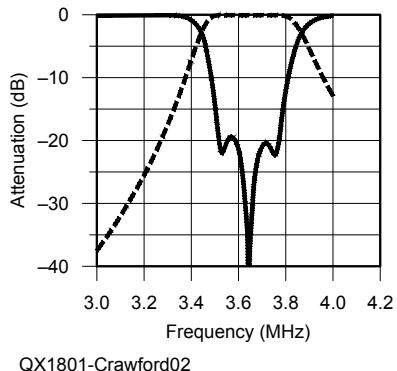


Figure 2 — Response of the filter in Figure 1.

operational voltages and currents. Test equipment to measure S_{21} and S_{11} is essential as well. Some short-cut methods can be used on simpler filter designs, however the filter of the complexity described here does not lend itself well to such an approach.

Discussion

The fundamentals covered here include filter loss, associated voltages and currents, and different design concepts such as impedance scaling of the filter, and use of powerful transform techniques, the Norton Transform in particular.

Minimizing Loss

As you delve deeper into general filter theory, you will encounter what are termed “elemental g -values” for Butterworth and Chebyshev filters. These “ g -values”, upon impedance and frequency scaling, evolve directly into the L and C values composing a low pass filter. The low pass filter can then be transformed into a band pass filter by resonating capacitors with inductors, and inductors with capacitors. There are multiple resources that describe how this is accomplished, some of which are cited^{7, 8, 9} herein.

Your intuition may lead you to believe that the smaller the number of filter components, the lower the loss. This is not necessarily true. It can be shown theoretically that,

$$L = \frac{4.34}{wQ} \sum_i g_i \quad \text{where } w = \frac{\omega_2 - \omega_1}{\omega_0} \quad (1)$$

L is the filter loss,

Q is the unloaded Q of resonators.

There are cases where a more complex filter, with higher filter order, actually has lower insertion loss than a simpler filter. It comes down to the calculated g -values for each implementation per Equation (1).

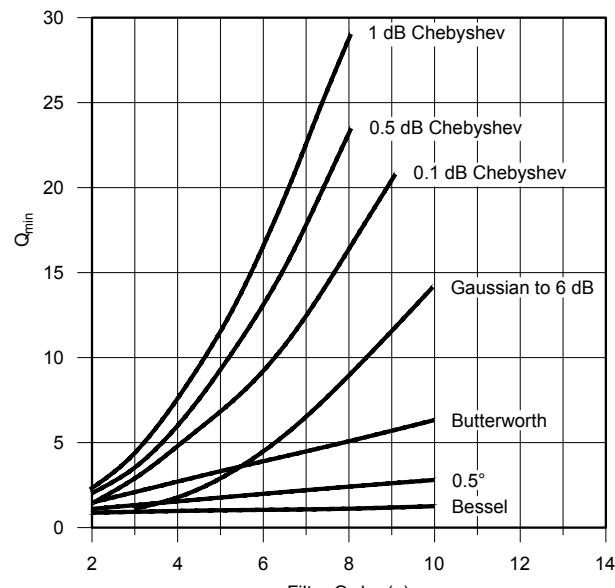


Figure 3 — Minimum Q s required in low pass filters.

Table 1
RF currents for 1,500 W and 50 Ω .

Component	Amps	Component	Amps	Component	Amps
L1	8.4	L5	6.7	C1	11.8
C5	6.7	L2	11.8	L6	20.6
C2	3.4	C6	9.6	L3	4.6
L7	2.7	C3	1.3	C7	7.8
L4	3.0	L8	4.5	C4	14
C8	12.5	L9	6.7	C9	4.5

Another important aspect of the design critical to loss is Q_{BP} , the Q of the pass band. In the case of the 20 m filter described here, ω_2 and ω_1 when multiplied by 2π , are 15.75 and 10.75 MHz, respectively, with a center frequency of 13.25 MHz. Q_{BP} is then,

$$Q_{BP} = \frac{13.25}{15.75 - 10.75} = 2.65 \quad (2)$$

Any given low-pass filter has minimum Q -values that each L and C must exceed to attain the desired passband shape, see Figure 3-8 in Williams¹⁰. For band pass filters, these minimum Q values are multiplied by Q_{BP} . Had the designer of this filter opted for a narrow passband, such as 13 to 15 MHz, Q_{BP} would have been ~ 7 making the inductors that much more difficult, if not physically impossible to build see Equation (3).

$$Q = Q_{LP\ Minimum} \times Q_{BP} \quad (3)$$

Q_{BP} in the 20 m case is made as low as possible to offset this effect while still attaining the desired rejection at 7 MHz and

21 MHz. Additionally, heightened Q s in resonators¹¹ will impact the accompanying voltages and currents, possibly further complicating your design and component selection. The needed Q is given in Equation (3).

If the low pass minimum required Q were 25 for example, the inductor Q would have to be higher,

$$Q = 25 \times 7 = 175 \quad (4)$$

This example shows why band pass filter component selection can be more difficult than for low pass or high pass filters because of the Q_{BP} multiplying effect. Figure 3 illustrates the increasing minimum Q of low pass elements as the filter order increases, with filter family as a parameter. The Q_{BP} impact is precisely why the passband for the 20 m filter is a full 5-MHz wide even though the 20 m band is 350 kHz in width.

Other Factors in Loss

A familiar and often used expression for air-core coil inductance is,

Table 2
RF voltage at 1,500 W input.

Load, •	<i>L</i> ₃ , V	<i>L</i> ₅ , V	<i>C</i> ₅ , V	<i>L</i> ₇ , V
25	510	220	850	390
50	485	270	1,100	400
75	520	320	1,300	430
100	560	360	1,430	460
100 at 2 kW	645	-	1,600	-

$$L = \frac{r^2 n^2}{9r + 10s}$$

where *r* is radius in inches, *n* is the number of turns and *s* is the coil length.

This is just a starting point and can be rather inaccurate as the length to diameter aspect ratio changes, frequency increases, and wire size is varied. There is an optimum range of coil aspect ratios which, when chosen, will heighten the available *Q* of the coils. An internet-based tool¹² employed in this design uses modified Bessel functions for wire loss, and considers the length of the coil as a function of wavelength at the frequency of operation. The coils used in the subject design have a diameter of 0.75 inches and theoretical *Q*s of approximately 400.

Voltages and Currents

The opening example (Figure 1) served to emphasize the all-important consideration of voltages and currents encountered in a filter. The best means to assess these conditions is through use of the circuit analysis tools previously mentioned. They are valuable for “what if analyses”, such as, “what happens to my RF currents if the VSWR were to be 2:1 instead of a nice, perfect 1:1?” Table 1 shows RF currents and Table 2 shows the RF voltage that result from this analysis for the 20 m filter shown in the schematic of Figure 4.

At different power levels the currents are,

$$I_{\text{Pow Lev}} = I_{1500\text{W}} \sqrt{\frac{\text{Power}}{1500}}$$

For different load conditions, simulations show approximately,

$$I_{25\Omega} = I_{50\Omega} \times 1.3$$

$$I_{100\Omega} = I_{50\Omega} \times 0.70$$

In constructing of this filter, capacitors are placed in series when higher breakdown voltages are required, and similarly, capacitors are placed in parallel to increase the net current capacity of a given capacitor. These capacitor combinations are annotated in Figure 4. Research revealed that CDV-16 capacitors can handle 5 A continuous current at HF. This serves as a guideline.

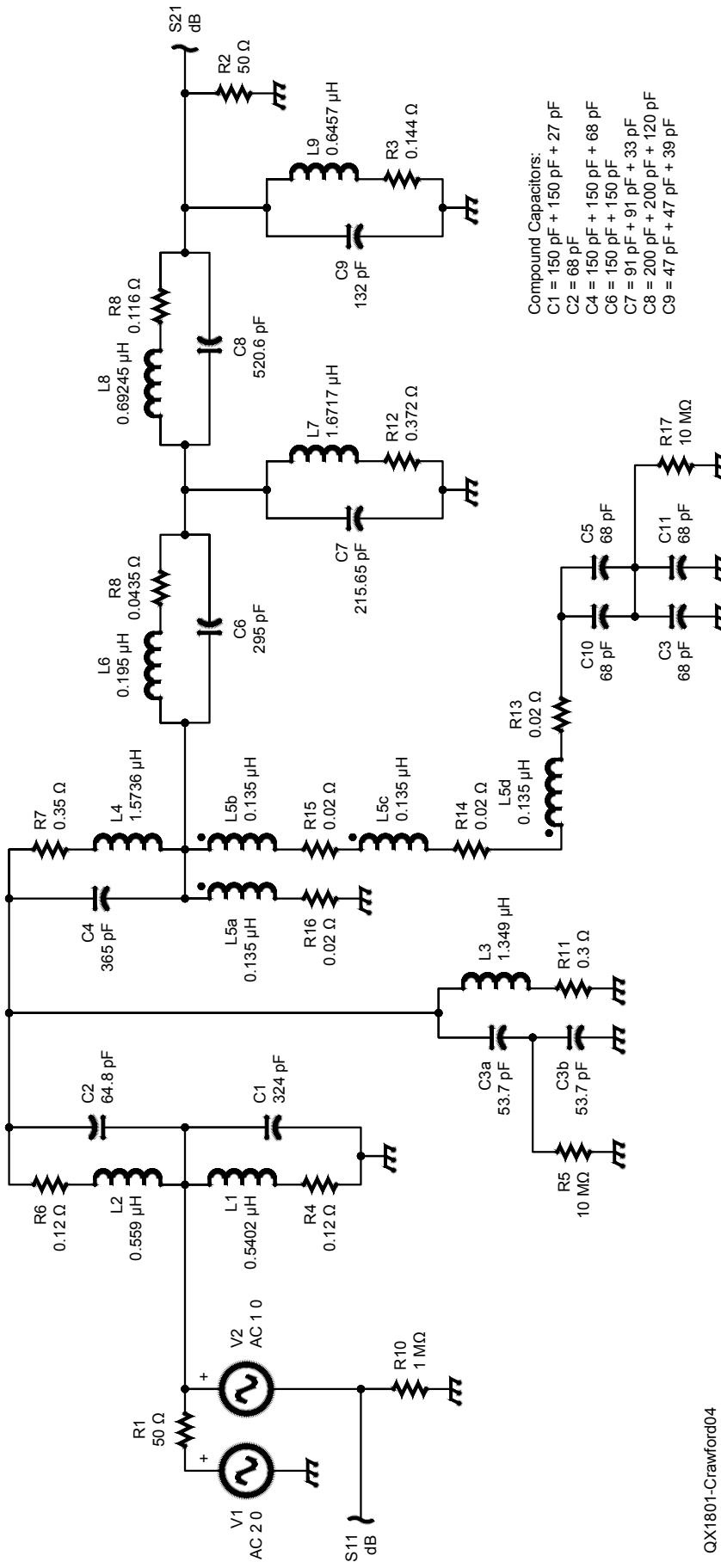


Figure 4 — Completed schematic for 20 m band pass filter.

Improving Filter Realizability and Performance

Although Figure 4 is the schematic of the filter as constructed, it began with the *Elsie* based schematic shown in Figure 5. There is quite a difference. A *SIMetrix* evaluation of Figure 5 revealed excessively high currents — more than 25 A — in resonator 4, the 178.7 nH inductor and 807 pF capacitor. There are some techniques that can be used to attack this problem. The first has already been employed by widening the passband to 5 MHz, and thus reducing Q_{BP} . This filter was designed around an impedance of 50 Ω. What if we designed it at 100 Ω and used impedance transformers at the input and output? This is an available option, but not selected for the following reasons.

Resonator 3 becomes large, elevating concerns about self-resonance in important parts of the stop band. Capacitor C2 takes on decreasing values making the idea of paralleling multiple capacitors troublesome. Simple *L*-networks at the input and output do not have sufficient bandwidth to comfortably handle 5 MHz. Other techniques could ameliorate this issue, but were not elected here.

We identified the use of Norton Transforms, and their use is now briefly outlined. Several references¹³ go into greater detail for the interested reader. Norton transformations appear in several forms and are shown in the Table of Figure A. These transformations allow different capacitor and inductor arrangements to be replaced equivalently with a different capacitor and inductor arrangement accompanied by an ideal transformer. Through use of a $1:n$ Norton transformation near the input of the filter and a $n:1$ complimentary Norton transformation near the filter output, an impedance transformation can be inserted almost anywhere within the filter. The wide bandwidth characteristics of the ideal transformers are retained, as is not the case for an *L*-network matching implementation where filter impedance scaling is used.

The Table of Figure A addresses both capacitors and inductors. In the case of a series capacitor, row one of the Table, the equivalency using the pi-pad connected capacitors and ideal transformer is used. Had the series inductor been used, the pi-pad of inductors and ideal transformer would be used. These transforms are rather simply derived from cascaded *ABCD* matrices for the two circuits we wish to equate, and the relationships derived.

It bears repeating that this equivalency of the Norton transform technique is superior over bandwidth to a narrow-band solution of an input and output *L*-network. As the filter's center frequency increases, *L*-networks can

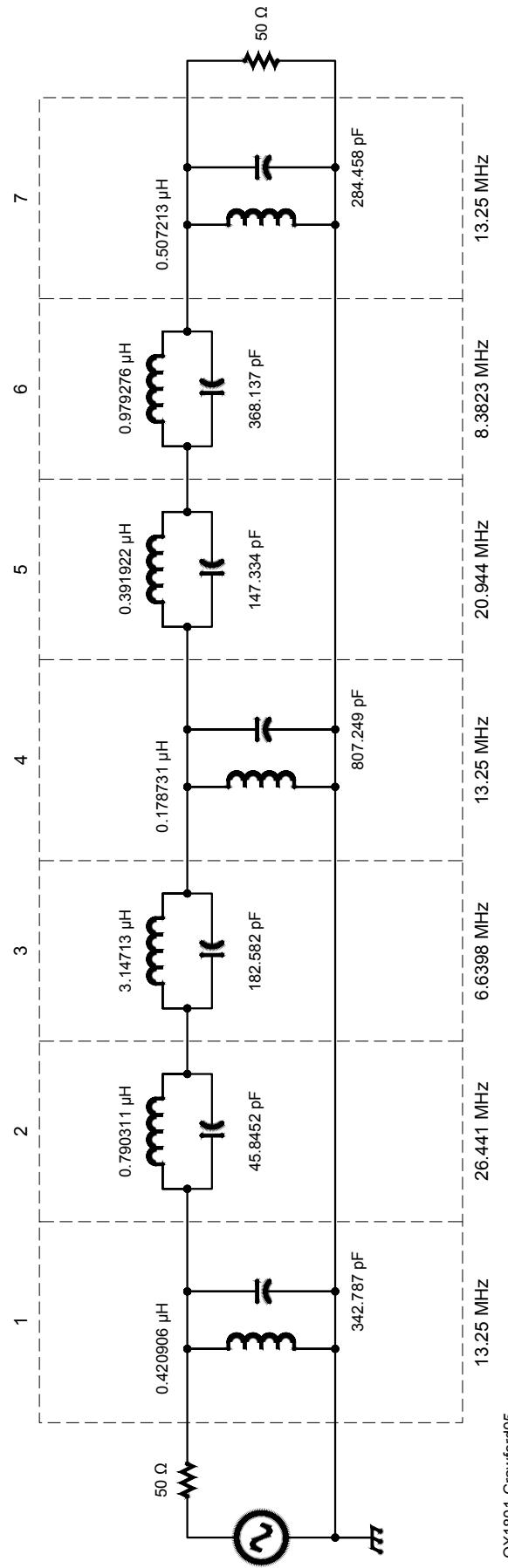
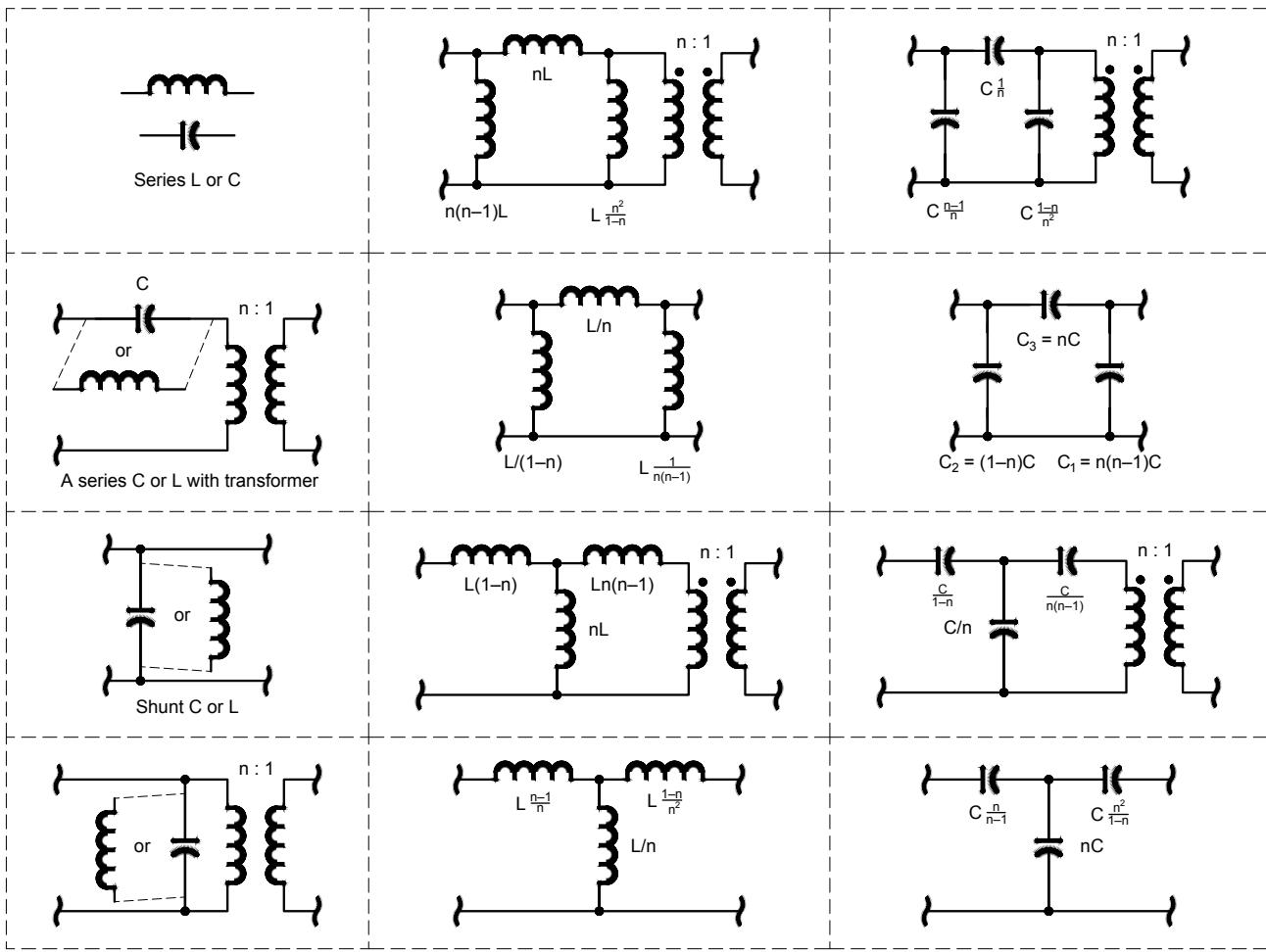


Figure 5 — A 5-th order elliptic filter designed using *Elsie*.



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Figure A — Table of multiple Norton Transforms.

be utilized with more success, not being relegated to only capacitor networks — which for 1500 W run up costs — and the Norton Transform.

An additional Norton Transform¹⁴, and the one used in this design, is that shown in Figure 6. This transform allows for an impedance step-up or step-down for a parallel LC network with an ideal capacitor.

Capacitors C_1 , C_2 , and C_3 in Figure 6 are derived from the simple algebraic expressions in the row-2 column-3 entry of the Figure A Table. The “ n ” is the transformer turns ratio.

There is a catch to the Norton approach, however. Upon study of the relationships in the Table of Figure A, one finds that there are always some resulting negative valued components. Consequently, when using the Norton technique, other components must be present to absorb these negative valued L s and C s. One of the intermediate steps in the 20 m filter design is shown in Figure 7 where, indeed, there are negative component values. Figures 8 and 9 show that these

negative values are handled in the same manner traditional inductors and capacitors are combined.

The route to the final 20 m band pass filter design employed the Norton technique two times at two different locations within the filter. The first was required to alleviate

the negative components that would result from the second transformation. The extra L s and C s in Figure 4 arise from the Norton Transform application. The two additional LC resonators (a) eliminate concerns about the otherwise floating node at this point, and (b) help equalize component values while

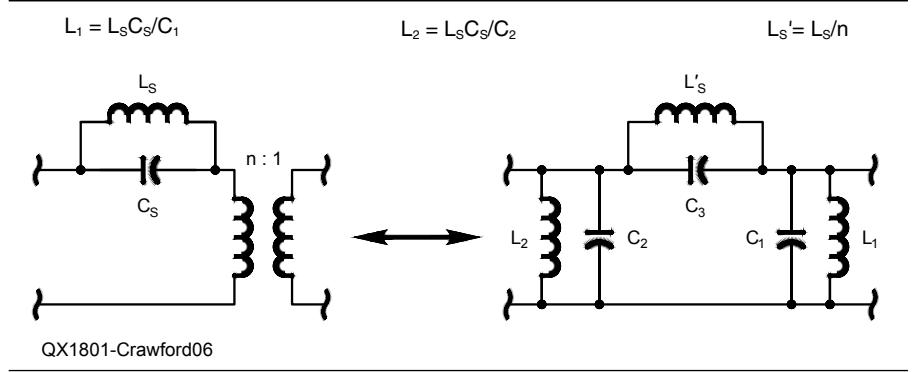


Figure 6 — An additional Norton Transform used in this design.

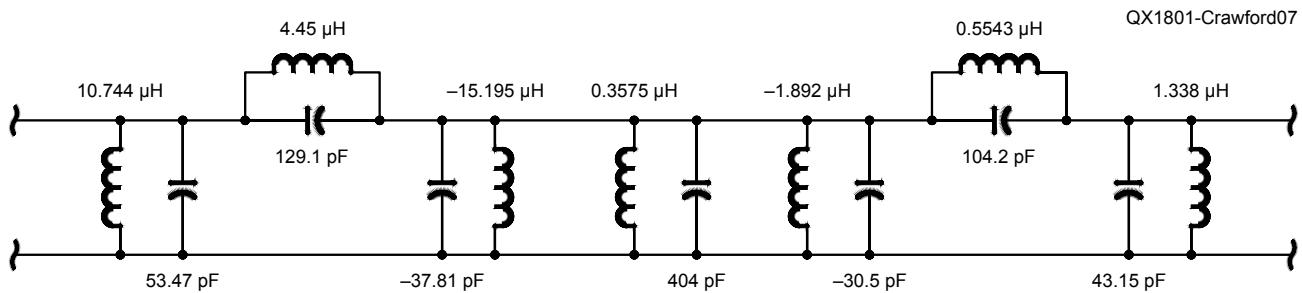


Figure 7 — One of the intermediate steps in the 20 m filter design is shown. Figures 8 and 9 show that these negative values are handled in the same manner traditional inductors and capacitors are combined.

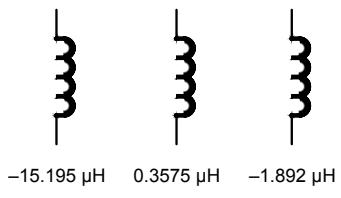


Figure 8 — Combining inductors that have negative values.

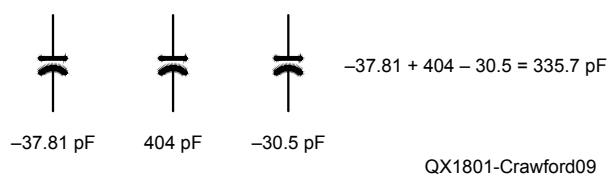


Figure 9 — Combining capacitors that have negative values.

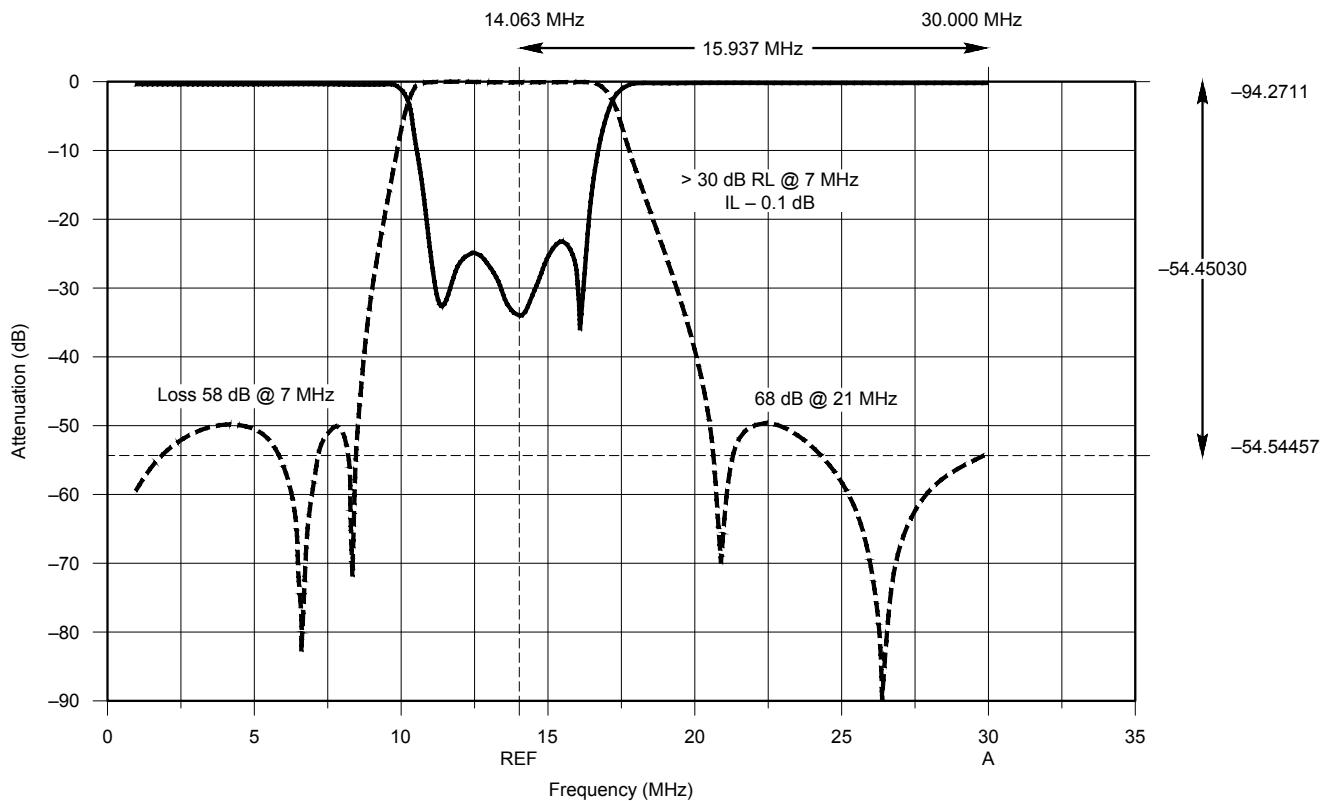


Figure 10 — Filter performance of the circuit in Figure 4 simulated with *SiMetrix*.

also diminishing currents and voltages in many cases. An in-depth discussion of the steps taken in the design is available at www.k0zr.com.

A Toroidal Transformer Solution

Techniques of optimizing inductor Q_s , diminishing Q_{BP} , impedance scaling the filter directly, or manipulation through the use of Norton Transforms have been offered. There is yet another valuable technique presented to finalize the design. This technique is used throughout the low-power W3NQN filter designs¹⁵.

While many good characteristics resulted from the Norton Transformation application, currents in the center resonator were considered too high. To reduce the current to more acceptable levels, the W3NQN technique of a multi-filar toroidal transformer is used. The prevailing voltages were sufficiently low such that core saturation was of no concern.

Further study of Figure 4 shows that a four-winding toroid ($L5a, b, c, d$) is used. The transformer is a quadrifilar transformer, four turns, making an impedance transformation of 16 times. The toroids are two stacked Amidon T-130-17cores. If one pulls apart this assembly of four coupled inductors, as shown in Figure 4, you see that indeed this is an autotransformer composed of the four different windings. Because there are four windings and the composite resonator is tapped just above the first one, an impedance change of 4^2 results, thus changing the inductor-capacitor currents from about 30 A to about 6.5 A. To maintain the same original LC resonant frequency, the factor of 16 multiplies the effective inductance so the accompanying capacitor must be reduced by a factor of 16. One additional consideration is that if the tap-point voltage were 270 V, the voltage on the resonating capacitor will be four times this value, or nearly 1.1 kV.

Figure 10 shows the filter performance, simulated by *SIMetrix*, of the circuit in Figure 4, and shown assembled in Figure 11. The insertion loss is theoretically about 0.1 dB, and the passband return loss for the 20 m band is better than 30 dB. Table 3 show a summary of the insertion loss and return loss of the filter. Figure 12 shows the S21 performance and Figure 13 shows the measured return loss.

The Filter Assembly

Figure 11 shows an image of the filter assembly. The board dimension are approximately 5.5 inches by 11 inches. The larger current-carrying inductors are wound with use #12 AWG thermaleze-coated wire. The several smaller inductors, including the toroid, use #14 AWG wire. All capacitors are CDV16s available through Mouser. These capacitors should safely handle 5 to 6 A each in the HF range. When needed, multiple values are placed in parallel for current sharing. Figure 14 is an image of the completed filter.

Table 3
Insertion loss and return loss.

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

Filter Tuning

This filter topology lends itself nicely to final tuning. Each parallel LC resonator frequency can be easily found using,

$$f_0 = \frac{1}{2\pi\sqrt{LC}}.$$

The resonators with resonant frequencies

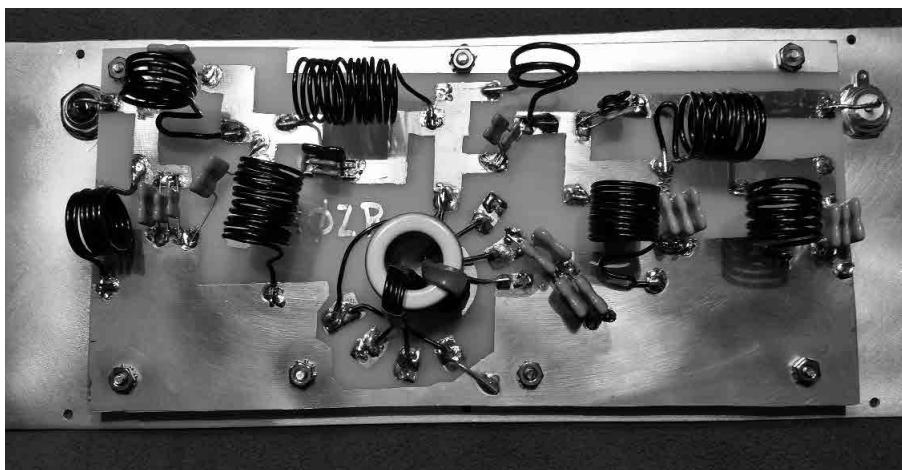


Figure 11 — The filter assembly. [Jeff Crawford, K0ZR, photo]

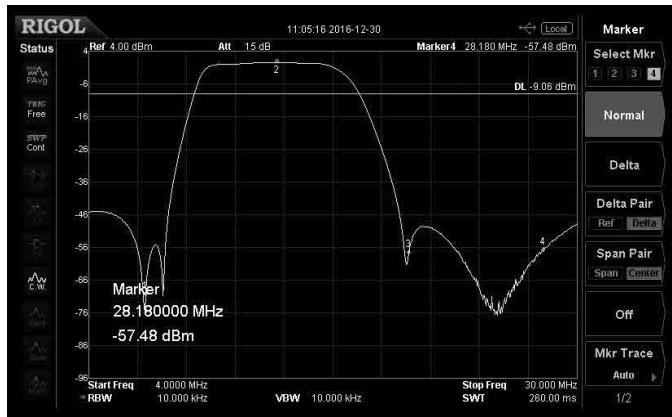


Figure 12 — The S22 performance of the filter.



Figure 13 — Measured return loss of the filter.



Figure 14 — The completed 20 m band pass filter. [Jeff Crawford, K0ZR, photo]

outside the passband form the deep notches in the stop bands. Those resonators are adjusted first, then the remaining resonators are adjusted to optimize passband return loss. It bears emphasizing the importance of tuning the filter pass band by optimizing return loss, *not* insertion loss.

Prior to assembly, each inductor was paralleled with a known capacitance and adjusted to what should be the resonant frequency for the “design-to” inductor value and known capacitor. This will save you many headaches in your assembly and tuning process.

Summary

The filter was designed to achieve a minimum of 50 dB stop band attenuation, while also taking advantage of the transmission zeros at 7 and 21 MHz. In operation at the 1500 W level, only inductors L_1 and L_2 were elevated in temperature, and only slightly, after ten minutes of constantly calling CQ. The cores were absolutely cold. The insertion loss is difficult to measure with the Rigol spectrum analyzer and tracking generator. The insertion loss appears to be about 0.1 dB. I will place a fan on the backside of the filter so as to lessen concerns about component heating. Component cost for this 20 m filter is approximately \$100.

Jeff Crawford, K0ZR, was licensed in 1969 at age 15 with the call sign WA0ZRT. He upgraded to the Amateur Extra class in 1976, and adopted call sign K0ZR. He earned a B.S. in Zoology from the University of Nebraska, in 1975, a BSEE from the University of Nebraska in 1983, and an MSEE from the University of Southern California in 1988. Jeff is a member of ARRL, Loudoun Amateur Radio Group, Potomac Valley Radio Club, and CWOPs, His

first welding project was a 63 foot free-standing tower, still standing after almost 40 years. Jeff has designed and built assemblies to tip-over his crank-up tower, a base for a quarter-wave 80 m vertical, and a traveling hoist system in an out building. He enjoys design and analysis of RF and microwave systems. Jeff is an active contestor in the larger world-wide contests. He has 304 DXCC entities confirmed on LOTW. Professionally, he is employed by a government think tank, specializing in RF and microwave hardware and systems.

Notes

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