# 20 meter bandstop filter notes

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## 1 Introduction

A shorted half-wavelength stub cut for 20 meters acts as a bandstop filter for 10 and 20 meters, while passing 40 and 15 meters. The measured performance of such a stub, made from RG-213 coax, is shown in fig. 1.

The main advantage of the stub is the ease of construction. The insertion loss on 40 meters is about 0.08dB, which corresponds to a loss of 27 watts with a 1500 watt output. The stub shown was tuned to give maximum insertion loss (about 29dB) in the 20 meter cw band; a stub tuned to the center of 20 meters will give a matched insertion loss between 27 dB and 29 dB across the band.

In a nonmatched condition for the 20 meter harmonic, typical of a monoband 40m antenna, the position of the stub on the transmission line can greatly affect its attenuation of the harmonic. The dependence of the stub position on a nonmatched line is well known, see, for example, ref. [1], and it has been recently stressed by W2VJN[2] and later K9YC[3] that the line is often unmatched at harmonic frequencies. Since the stub attenuates by shorting the transmission line with a low impedance at the attenuation frequency, if the mismatched line has a voltage minimum where the stub is placed, the stub effectiveness can be greatly reduced. As K9YC describes, a double stub separated by a quarter wavelength at the suppression frequency will ameliorate this problem since the first stub establishes a low impedance at its position, and this is transformed to a high impedance which is then shorted by the second stub. This transformation occurs even for a badly placed first stub. The total matched insertion loss for a double stub is twice the insertion loss in dB (i.e. 50 db to 60 dB) for harmonics.

A lumped element filter is an alternative to stubs. We know that the effective Q for stubs is rather poor at HF. For example an eighth wavelength shorted length of RG-213 at 14 MHz has a reactance of about 50 ohms and a resistance of about 0.93 ohms, corresponding to a Q of 54. Matters get worse at lower frequencies and better at higher frequencies, so at 3.5 MHz a shorted eighth wavelength of RG-213 has a Q of about 30 while for 144MHz the Q is about 160. Capacitors at HF often have a Q greater than 1000, while toroids have typical Qs of 100 to 200, and carefully constructed air-core coils can have substantially higher Qs.

The attempt here will be to design and build a bandstop filter for 20 meters that meets or exceeds the specifications for a double stub filter. The advantages of a successful

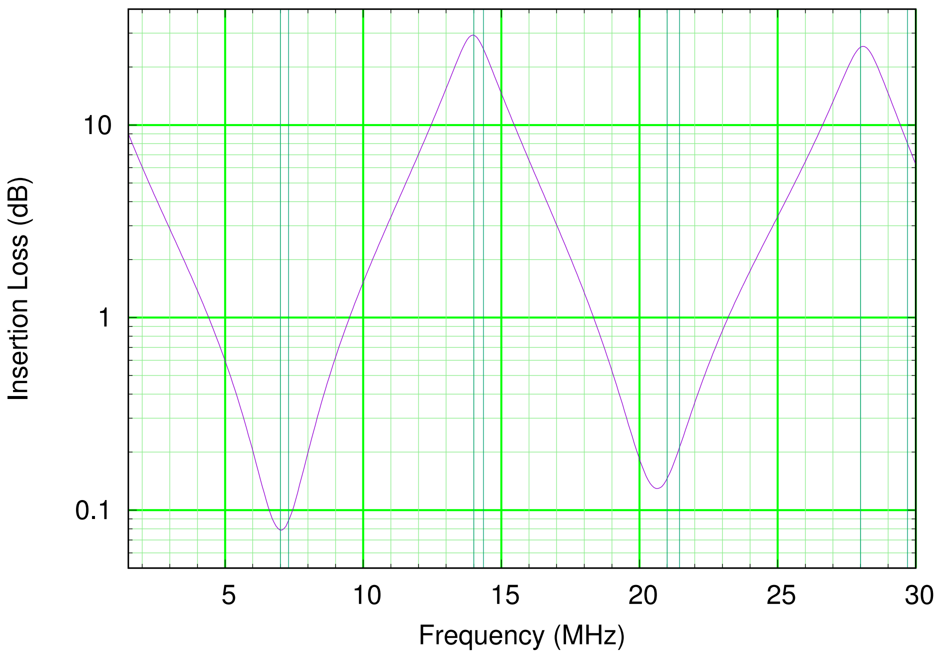


Figure 1: The measured insertion loss, i.e. the 50 ohm $-S_{21}$, of a tee-connected, shorted, half-wavelength 20 meter stub made from RG-213.

design are smaller size and less insertion loss at 40 meters. Unlike stubs, the suppression will only be at one frequency. The trade off is that the filter will be useable on the other HF contest bands 160, 80, 40, 15, 10.

# 2 Bandstop filter design

I used the ELSIE software package[4] to design the filter. On the new design page, I chose

- shunt-input bandstop filter
- Chebychev
- 12 MHz ripple bandwidth
- 14.175 MHz center frequency
- Third order
- 50 Ohm input termination

The filter topology is shown in fig. 2, with the component values $C_1 = C_3 = 51.8277\,\text{pF}$, $L_1 = L_3 = 2.43239\,\mu\text{H}$, $C_2 = 518.277\,\text{pF}$, $L_2 = 243.239\,\text{nH}$, As expected, the three tuned circuits are all resonant at 14.175 MHz.

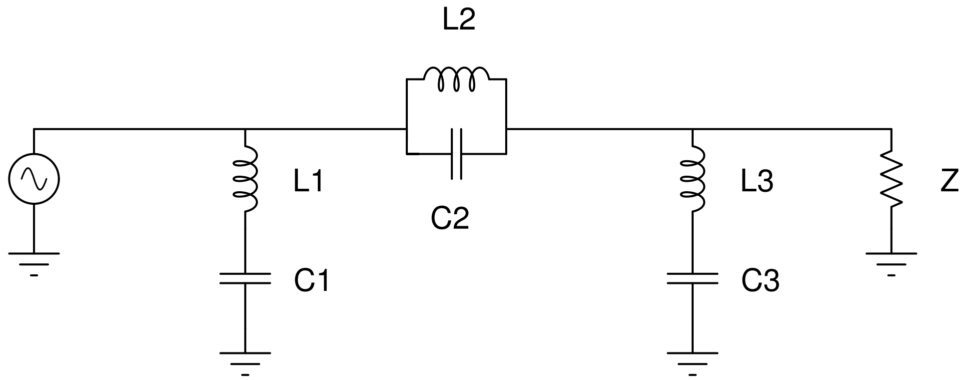


Figure 2: The topology used for the 20 meter bandstop filter.

| Band V/I | $C_1$ | $C_2$ | $C_3$ | $L_1$ | $L_2$ | $L_3$ |
|---|---|---|---|---|---|---|
| 160 Voltage | 395 | 25 | 395 | 8 | 25 | 8 |
| 160 Current | 0.25 | 0.16 | 0.25 | 0.25 | 7.9 | 0.25 |
| 80 Voltage | 421 | 54 | 421 | 34 | 54 | 34 |
| 80 Current | 0.53 | 0.67 | 0.53 | 0.53 | 8.44 | 0.53 |
| 40 Voltage | 528 | 124 | 528 | 141 | 124 | 140 |
| 40 Current | 1.21 | 2.84 | 1.21 | 1.21 | 10.7 | 1.21 |
| 15 Voltage | 326 | 222 | 322 | 717 | 222 | 709 |
| 15 Current | 2.15 | 14.68 | 2.12 | 2.15 | 6.67 | 2.12 |
| 10 Voltage | 133 | 119 | 133 | 521 | 119 | 520 |
| 10 Current | 1.17 | 10.51 | 1.17 | 1.17 | 2.68 | 1.17 |

Table 1: The calculated maximum RMS volages and currents for a 2:1 SWR.

Since these values are all realizable, no transformation[5] of the filter topology is necessary.

Quick estimates of the capacitor currents and voltages indicated that modern 1000V rf rated mica capacitors[6] would work well. These are available in standard values of 500pF and 50pF both 5 percent. Substituting these value, and reresonating the inductors, the inductors become 252 nH and 2.52 $\mu$H.

Table 1, shows the calculated maximum RMS voltages and currents across and through the components at an SWR of 2:1. These were done with a straightforward series parallel numerical analysis. The magnitude of the reflection coefficient was set from the SWR, and its phase was brute force varied from 0 to $2\pi$, and the maximum values recorded.

Typical values of the maximum allowed current through the capacitors is given in the datasheet[6]. The allowed current rises linearly with frequency for all values given until it plateaus at 10 to 20 Amperes.

A quick calculation shows that the linear region is simply given by the voltage rating of the capcitor, that is

$$I_{\mathrm{RMS}} = 2\pi f C V_{\mathrm{RMS}} = 2\pi f C \frac{V_{\mathrm{peak}}}{\sqrt{2}}\,. \tag{1}$$

In this region, if the peak voltage is less than the rated DC voltage, the current is safe. All of the plateaus are above 10 A, so only the 15 meter value of 14.68 A for the 500 pF capacitor is problematic. This is to be expected since 15 meters has the smallest ratio of frequencies and is therefore the closest band to 20 meters, and the filter is most stressed there. The 470 pF 1000V curve appears to have a value of around 15A at the plateau, but it could easily be as small as 12A. Reducing the SWR to 1.4:1 brings the maximum current in $C_2$ down to 12.3 A. Alternatively, dropping the allowed power to 1000 W at 2:1 SWR brings the current down to a safe level.

The other possible problem is the current through $L_1$ and $L_3$ at 15 meters and possibly 10 meters. Since I wanted to use toroids, a quick calculation shows they might have heating problems. I will look at this below, but since it is unlikely that a 20 meter bandstop filter will be used on a 10 or 15 meter amplifier, it does not seem worthwhile to redesign the filter to handle these bands. The filter is therefore designed to be used only on 160, 80, and 40 meters, 1500 watts with a 2:1 SWR.

Note, SO-239 connectors are only rated at 500 ac volts peak. For an SWR of $s$, that corresponds a peak voltage of $V_p = \sqrt{s \cdot 100 \cdot 1500}$. For an SWR of 2:1, this is 548 volts. So a 2:1 SWR at 1500 watts is a little beyond the SO-239 rating. To stay within the rating, the SWR would need to be below $\frac{500^2}{150000} = \frac{5}{3} = 1.67$.

Just as for the stubs, the performance of the filter can be degraded under mismatched conditions as would normally occur at 20 meters for a 40 meter monoband antenna. However, the filter as designed has the same transformation properties as the double stub described by K9YC. That is, if the filter is connected at a point where the 20 meter impedance is already low, the series resonant circuit $L_3$ $C_3$ across that point will be relatively inaffective. However, the parallel resonant circuit then gives a high series impedance which forms a voltage divider with the low impedance output. This voltage divider is connected across the low-impedance series resonant circuit $L_1$ $C_1$. The net result is good suppression in all cases just as for the double stub filter.

# 3 Construction

## 3.1 Low power filter

To validate the design, I first built a 100 W version. I built this low power filter in a 2 inch by 3 inch by 5 inch aluminum box (similar to a Hammond Manufacturing 1411N). I mounted SO-239 connectors to the box ends.

Since $\sqrt{\frac{1500}{100}} \sim 4$, all of the maximum voltage and currents will be lower by that factor. In this case, low loss ceramic capacitors will work. I used TDK type CC45 ceramic capacitors rated at 3KV. I used a parallel combination of 2 220 pF capacitors and one 68 pF capacitor for the 500 pF capacitor, and 47 pF capacitors for the 50 pF. The

40m current through the $L_1$ and $L_3$ inductors is low, so they can be toroidal inductors. I used Micrometals T130-17 cores wound initially with 24 turns of 16 gauge enameled copper wire. These were connected in series with the 47pF capacitors from the SO-239 connectors to ground, and then paralleled by $\frac{1}{4}$ watt, 49.9 Ohm resistors. The return loss was then swept and the toroids adjusted until the return loss peaked at 14.175 MHz.

The Wheeler formula for solenoid inductors was then used to find a reasonable sized inductor. I wound it with 14 gauge copper wire to make it easily adjustable. However, I designed it to have a wire spacing such that it would be approximately space wound with 6 gauge wire since I have about 10 feet of that on hand. The plan was to wind a coil of approximately the same geometry with 6 gauge wire for the 1500 watt filter.

I then removed the toroids and their capacitors and series connected the two SO-239 connectors with the parallel combination of the 500 pF capacitor and the coil. I measured the frequency of the peak in insertion loss, and adjusted the coil for resonance at 14.175 MHz.

Once this was accomplished, I reconnected the toroid circuits to complete the filter. The final air coil is 1 inch in diameter, 1.25 inches long, made with 3.5 turns of 14 gauge wire.

The initial response was quite good. Slight adjustments to the toroids gave an insertion loss close to the theoretical curve. The return loss at high frequencies is not as good as predicted. Presumably my junk box capacitors values are to blame for that.

The quickly thrown together test filter is shown in fig. 3

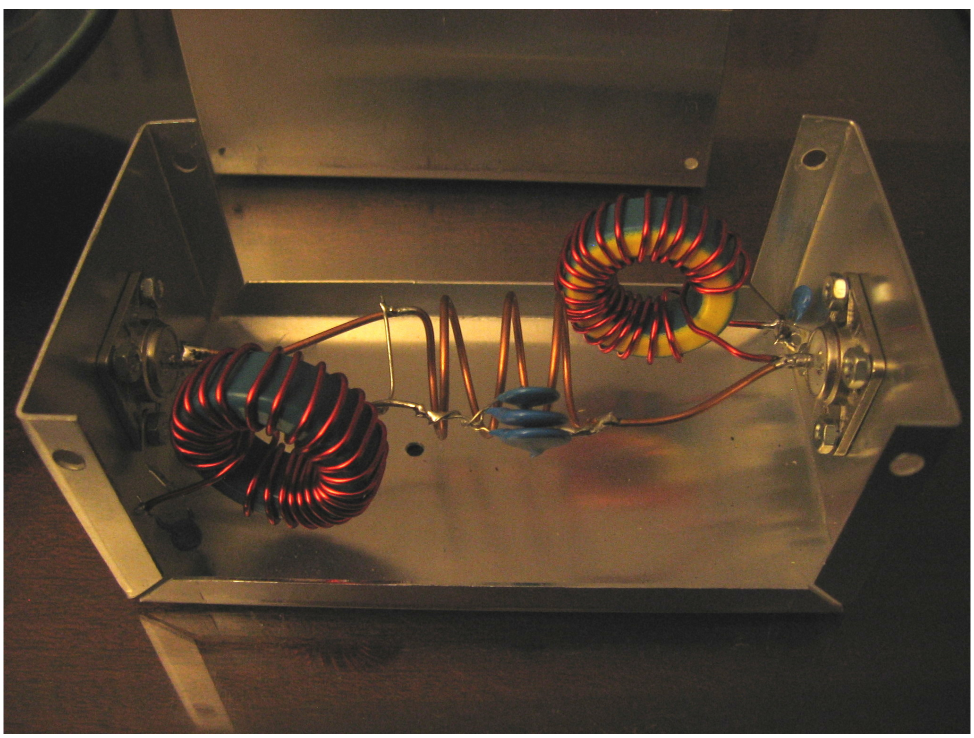

Figure 3: The low-power test filter as constructed.

The measured insertion and return loss of the filter are shown in fig. 4.

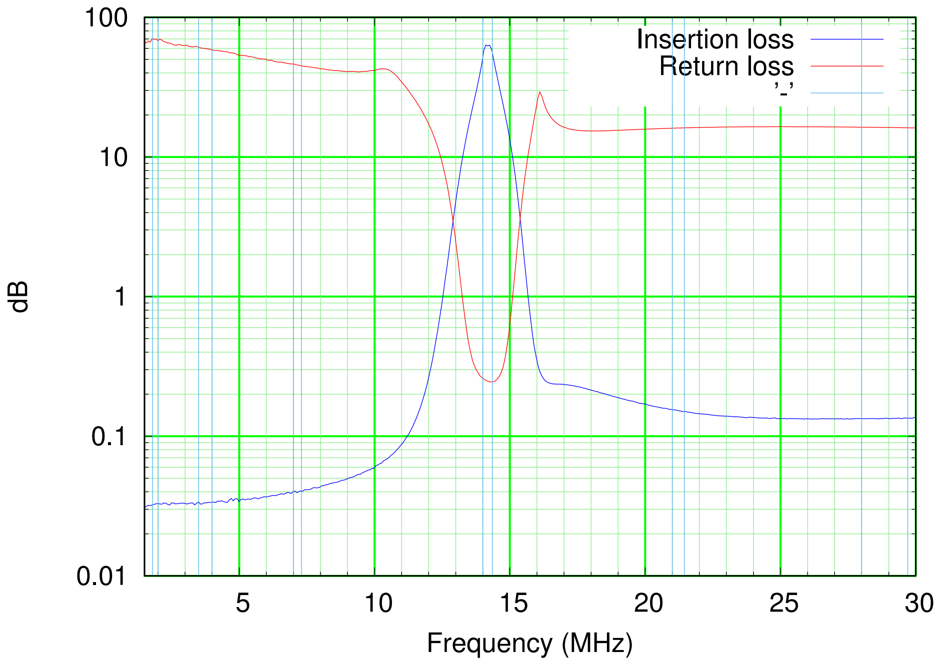


Figure 4: The measured insertion and return loss for the low-power filter.

### 3.2 High power filter

It is clear that on 40 meters, the quality of $L_2$ determines the filter 40 meter insertion loss. Using toroids for $L_1$ and $L_3$ does not substantially degrade performance and means that the filter does not require internal shielding – simplifying construction.

I wound $L_2$ from 6 gauge bare copper wire, and to improve its $Q$, I built the filter in somewhat larger Hammond Manufacturing 1411PU 6 inch by 5 inch by 4 inch aluminum box. Cornell Dubilier MC22FF501J-F, and MC12FF500J-F capacitors (500 pF and 50pF, 1000V, rf rated, mica capacitors, respectively) were used. To resonate with this capacitor in the bigger box, $L_2$ was reduced to 3 turns with a slightly larger diameter. The filter as built is shown in fig. 5.

The return and insertion loss are shown in fig. 6.

## Appendix – Toroid losses

Micrometals defines the peak flux density by calculating the voltage from Faraday's law (in SI units)

$$\boldsymbol{\nabla} \times \boldsymbol{E} = -\partial_t \boldsymbol{B} \,. \tag{2}$$

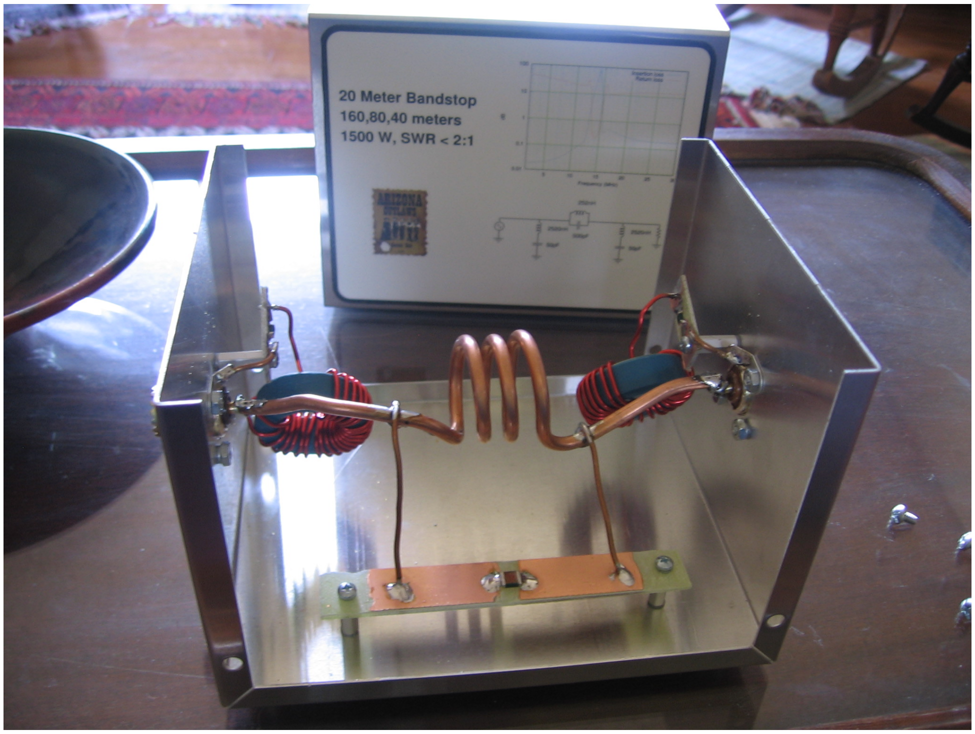


Figure 5: The high-power test filter as constructed.

which on integrating over an area and applying Stokes theorem becomes

$$\oint d\boldsymbol{l} \cdot \boldsymbol{E} = -\partial_t \Phi \tag{3}$$

where $\Phi$ is the total flux. Integrating around the wire, assuming a perfect conductor, gives zero, and the integral across the ends of the wire is the voltage, so the result is

$$V_{\text{peak}} = 2\pi f \Phi_{\text{peak}} \,. \tag{4}$$

Micrometals apparently defines $B_{\text{peak}}$ to be $\Phi_{\text{peak}}$ divided by the area, which they approximate by $NA$ where $A$ is the cross sectional area of the core and $N$ the number of turns. Converting to rms voltage, $V_{\text{peak}} = \sqrt{2} V_{\text{rms}}$, the result is

$$B_{\text{peak}} = \frac{V_{\text{rms}}}{NA\sqrt{2}\pi f} \,. \tag{5}$$

The SI flux density unit is a Tesla. Micrometals uses the Gaussian unit Gauss for flux density, and the Gaussian unit of length (cm), but does not use the Gaussian unit of voltage, instead using volts. One Tesla is $10^4$ Gauss, 1 cm = 0.01 meter, so

$$B_{\text{peak}}(Gauss) = \left[ 10^4 \frac{\text{Gauss}}{\text{Tesla}} 10^4 \frac{\text{cm}^2}{\text{m}^2} \right] \frac{V_{\text{rms}}(Volts)}{\sqrt{2}\pi f(Hz) NA(cm^2)} \,. \tag{6}$$

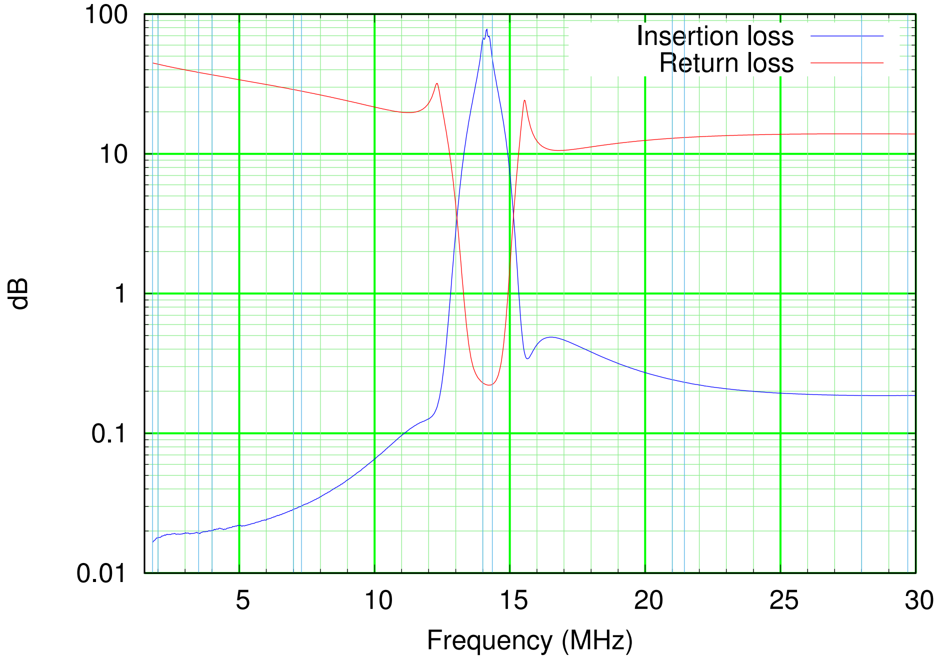


Figure 6: The measured insertion and return loss for the high-power filter.

Substituting $4.44 \simeq \sqrt{2}\pi$ gives the equation in the Micrometals application note [7].

We see from table 1, the flux density for $L_1$ and $L_3$ will be highest on 15 meters, with 40 meters the second highest. This is expected since these are the closest bands to 20 meters.

A Micrometals T130-17 core has outer diameter, $d_o = 33.0$mm, inner diameter $d_i = 19.8$mm, and height $h = 11.1$mm with cross sectional area $A = 0.74\text{cm}^2$ and volume 6.1 $\text{cm}^3$. For 23 turns, and the voltages from table 1, we have peak fluxes of about 26 Gauss on 40 meters, and 46 Gauss on 15 meters.

I could not find any core loss data published by Micrometals for type 17 material. Han et al. [8] published measurements for core loss of typye 17 material at 30 MHz and higher. They were unable to measure core loss at 20 MHz. Their result at 30 MHz was fit to what they call the Steinmetz equation

$$P_V\left(\frac{\text{mW}}{\text{cm}^3}\right) = KB^{\beta}_{\text{peak}}(\text{Gauss}) \tag{7}$$

with numerical values of $K = 0.03621$, $\beta = 2.76$.

Plugging in values for 15 meters, the toroids may be problematic there.

I also measured the rf resistance at 21 MHz for the T130-17 toroids to be 1.2 Ohms. This would correspond to a total wire plus core loss of about 6 watts, which according to Micrometals will cause substantial heating of the core.

The bottom line is do not use this filter on 15 meters.

# References

[1] R. W. P. King, H. R. Mimno, and A. H. Wing, *Transmission Lines Antennas and Wave Guides*, 1st ed. New York: MacGraw-Hill, 1945.

[2] G. Cutsogeorge, "Optimizing the performance of harmonic attenuation stubs," *National Contest Journal*, pp. 3–4, January/February 2015. [Online]. Available: http://ncjweb.com/bonus-content/Cutsogeorge-Stubs2.pdf

[3] J. Brown. (2015) Optimizing the location of stubs for harmonic suppression. [Online]. Available: http://audiosystemsgroup.com/LocatingStubs.pdf

[4] Elsie software. [Online]. Available: http://www.tonnesoftware.com/elsie.html

[5] S. B. Cohn, "Direct-coupled-resonator filters," *Proceedings of the IRE*, vol. 45, pp. 187–196, 1957.

[6] Types mc and mcn multilayer rf capacitors. [Online]. Available: http://www.cde.com/resources/catalogs/MC.pdf

[7] J. Cox, "Iron powder core selection for rf power applications." [Online]. Available: www.micrometals.com/appnotes/appnotedownloads/ipcs4rfp.pdf

[8] Y. Han, A. L. G. Cheung, C. R. Sullivan, and D. J. Perreault, "Evaluation of magnetic materials for very high frequency power applications," *IEEE Transactions on Power Electronics*, vol. 27, pp. 425–435, 2012.