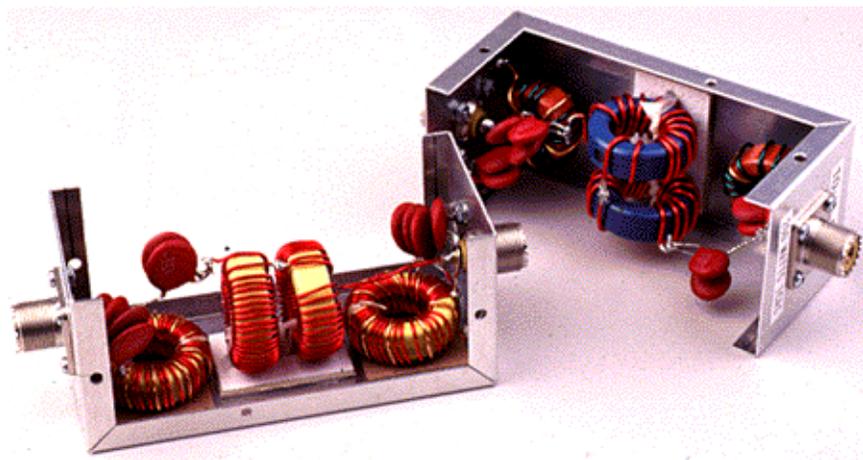


Clean Up Your Signals with Band-Pass Filters

Part 1—These inexpensive, easily built filters can be the buffer you need between the signals you want to hear and send and those you don't!

By Ed Wetherhold, W3NQN



An inside view of 160 meter and 10 meter BPFs housed in 2¹/₄×2¹/₄×5-inch (HWD) boxes equipped with SO-239 connectors at each end. With the exception of the 10 meter BPF, the filters are now housed in slightly larger boxes—2¹/₈×3×5¹/₄-inch (HWD)—to lessen component crowding and minimize response degeneration when the cover is installed.

In January 1996, Tim Duffy, K3LR, chief operator of one of the top-three-ranked multi-multi contest stations, contacted me. Tim asked me to design a series of band-pass filters (BPFs) that he could use between his 150 W transceivers and 1500 W amplifiers. He wanted to reduce the transfer of harmonics and broadband phase noise to the amplifiers. The filters would also make the transceivers, when receiving, less susceptible to overload from the adjacent-band transmitters. This project is the result of my efforts in designing, constructing and testing BPFs for six amateur bands—160, 80, 40, 20, 15 and 10 meters.

These filters provide a level of selectivity and reliability unachieved in currently available filters—including commercial units. This performance is obtained by employing a three-resonator filter configuration using inductors wound on powdered-iron or phenolic toroidal cores, and series-parallel connected, high-voltage, low-loss NP0 ceramic capacitors. If you're a contester (especially one who's involved in multi-multi contesting and ARRL Field-Day activity), these filters are *what you need!* [1] (*Editor's note:* You don't have to operate multi-multi to reap the benefits of these filters. Do remember that these are *single-band* filters and must be switched appropriately when changing bands or you'll let the smoke out.)

Background

Multiple-transmitter interference was discussed recently by Thomas Moliere, DL7AV, in his *CQ Contest* article. [2] Tom's article is an excellent list of the many interference sources common to this type of operation. Multiple-transmitter operation at a single site creates an unusual RF environment that results in many unexpected interference problems not normally encountered with single-station setups. For example, in addition to transmitter harmonics, there can be receiver blocking and damage, and the generation of harmonics and intermodulation at the nonlinear receiver input. Tom's article shows how the addition of 1500 W, high-power, low-pass filters at the transmitter's outputs proved helpful. Here, I'll concentrate on filters designed to be placed *between a transceiver and the input to a high-power amplifier*. The advantage of this approach is that the filters can be constructed using standard, readily available components for an ICAS power level of 200 W. (This is not the case for filter components required for use at a power level of 1500 W.)

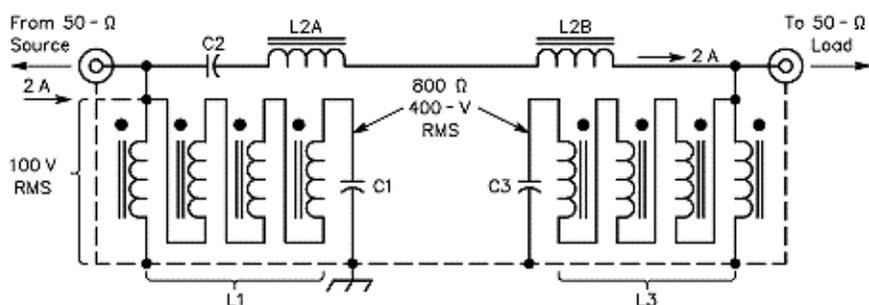
A recent *QST* harmonic-attenuation BPF article describes a three-resonator design. [3] Although the BPF performance is

satisfactory, it is limited to a power level of 100 W, and the L1 and L3 reactances are too low for good Q. Also, the design information is limited, so readers can't confirm the author's designs and try others. Another QST project uses capacitively coupled input and output resonators. [4] The omission of a series inductor in the coupling circuit, however, results in poor stopband attenuation above the upper cut-off frequency. A commercially available BPF used by many multi-multi contesters also exhibits similar poor, high-frequency attenuation.

The new BPFs provide for higher-power handling, increased reliability and greater attenuation. To help you evaluate many different designs and select the most-promising ones for trial, I've included a 62-line BASIC program that allows you to change component values and explore the resultant effect on the BPF bandwidth, return loss and stopband attenuation. [5]

Band-Pass Filter Design

I selected a three-resonator Chebyshev BPF for use on the six bands (see Figure 1). This filter is a satisfactory compromise between adequate selectivity and acceptable complexity. The filter consists of input and output parallel-tuned shunt resonators with their tops initially coupled by a series-tuned resonator. The resonators, capacitors and inductors are numbered from left to right as 1, 2 and 3.



Notes:

1. The listed RMS voltages and current in the diagram are for an input power of 200 W.
2. L1 and L3 are quadrifilar wound with 50-Ω taps.
3. L2A and L2B are wound on separate toroidal cores.
4. ● = Phasing

Figure 1—Schematic of a three-resonator band-pass filter (BPF). L1 and L3 are *quadrifilar* wound and equipped with a 50-Ω tap point. L2A and L2B are wound on separate toroidal cores. The RMS voltages and currents shown represent a filter input power of 200 W.

Component Values

Resonators 1 and 3 have identical component values. To minimize the component-value spread to less than 3.6 to 1, the filter impedance is made 450 or 800 Ω, and the input and output shunt inductors tapped appropriately to obtain 50 Ω. This allows L1 and L3 to have a reactance of more than 100 Ω, assuring a reasonable Q. By comparison, the three-resonator BPF designs in the September 1988 QST article (see Note 3) have L1 and L3 reactances of only 25 Ω.

Winding Type and Inductance Value

Figure 1 is the basic circuit for all the filters, but shows the *quadrifilar-wound* inductors used in the 160, 40, 15 and 10 meter BPFs. Although a quadrifilar winding is generally preferred for all designs to minimize component value spread, inductance-value limitations with the quadrifilar winding prevent using that configuration for the 80 meter filter. On that band, *trifilar windings* are used for L1 and L3. *Trifilar windings* were also used in the 20 meter BPF because they give an attenuation maximum at the second harmonic frequency of the 14-MHz signal.

Using a trifilar or quadrifilar winding means that a particular inductance value cannot be obtained by simply adding or removing one or two turns. Instead, any turns added or removed must be done *as multiples or submultiples* of three or four turns to maintain the trifilar or quadrifilar configuration. This winding limitation makes it more difficult to obtain a particular inductance value. In spite of this, the trifilar and quadrifilar windings are preferred because much better interwinding coupling results over the entire inductor. This provides a correspondingly greater stopband attenuation than is possible with the more common progressive winding.

Achieving a 50-W Tap

The taps on L1 and L3 are used as the input and output connections to the filter and should be terminated in a 50 Ω impedance. (These taps also serve as the final connection points for the series-resonant circuit of C2 and L2.) Because the impedance of an inductor varies as the square of turns, the input/output tap is placed at a point having $1/3$ or $1/4$ of the total number of turns so that the tap impedance is $1/9$ or $1/16$ of the total impedance. For example, on a trifilar-wound inductor of a 450 Ω resonator, the impedance at the $1/3$ tap is $450/3^2 = 50 \Omega$. The impedance at the $1/4$ tap of a quad-rifilar-wound inductor of an 800 Ω resonator is $800/4^2 = 50 \Omega$.

Calculating BPF Component Values

I'll use the calculation of the 160 meter filter components as an example of the procedure followed to design all the filters. First, a preliminary design is made to find trial component values, then the BASIC program (see Note 5) is used to evaluate a number of designs using a range of C1 and C2 values near the trial C1 and C2 values. The most-promising design is assembled and tested. If its performance is satisfactory, the design is accepted.

To calculate the preliminary 160 meter BPF component values, I use the procedure described in *The ARRL Handbook*. [6] Because the 160 meter bandwidth is 0.20 MHz, I first design a 50 Ω Chebyshev low-pass filter having a ripple cut-off frequency of 0.28 MHz, slightly greater than the actual bandwidth. (A Chebyshev filter has many possible designs and is preferred to a Butterworth, which has only one possible design.) To minimize reflective losses, I use a return loss greater than 22 dB.

Referring to *Handbook* Table 16.2, the normalized component values associated with $N = 3$ and $RL = 26$ dB, where $C1, C3 = 0.6292$ F and $L2 = 0.9703$ H, are used to calculate a C-in/out low-pass filter having a cutoff frequency of 0.28 MHz. The capacitive and inductive *scaling factors* based on a 50 Ω filter are 11368×10^{-12} and 28.421×10^{-6} , respectively. Multiplying the normalized C and L values by the respective scaling factors gives C1, C3 and L2 values of 7153 pF and 27.58 μ H, respectively.

The 50 Ω low-pass filter is transformed into a band-pass filter by resonating C1, C3 and L2 at 1.87 MHz, the geometric center frequency of the band-pass filter. For example, $L1$ and $L3 = 25330/((F^2) \times C1)$, and $C2 = 25330/((F^2) \times L2)$, where F, C and L are in megahertz, picofarads and micro-henries, respectively. The three-resonator BPF has shunt input and output parallel-tuned resonators with C1 and C3 = 7153 pF and L1 and L3 = 1.013 μ H; the series-tuned resonator has C2 and L2 values of 262.6 pF and 27.58 μ H. Although this BPF has the desired theoretical response, the design is *not optimum* because of the wide spread in the ratio of the C1 and C2 values. For example, the ratio of $C1/C2 = 27$. At the center frequency, the reactances of C1 and C3 and L1 and L3 are only 11.9 Ω ; this is too low a reactance to obtain a reasonable Q. In comparison, the C2 and L2 reactances are satisfactory at 324 Ω , where a satisfactory Q is much easier to achieve.

To improve the Qs of the C1 and L1 and C3 and L3 shunt resonators, and to reduce the component-value spread, the impedance of the shunt resonators can be raised by a factor of 4, 9 or 16, to 200, 450 or 800 Ω , respectively. The 50 Ω series C2/L2 resonator then connects to input/output taps at $1/2$, $1/3$ or $1/4$ of the turns above the ground connection. For the 160 meter BPF used in this example, L1 and L3 are quadrifilar wound with the 50 Ω taps connected to the top of the first quarter winding above ground. The L1 and L3 and C1 and C3 values then become 16 and $1/16$ times the 50 Ω values, or 16.2 μ H and 447 pF, respectively. By using a quadrifilar tap, the component spread is reduced to about 1.7, and the L1 and L3 reactance at the center frequency is 16 times greater than before, or 190 Ω . This level of reactance is much more suitable for achieving a satisfactory inductor Q.

BASIC Program Calculates Many BPF Designs

Having decided on a winding style for L1 and L3, and finding trial values for C1 and C2, I then employ the BASIC program to evaluate many different designs (see Note 5). Although the program designs 160 meter BPFs, it is easily modified for other bands by changing the center-frequency variable (FC) from 1.87 to a new center frequency, and changing the C1 and C2 values to those appropriate for the new band.

To use the program, enter values for the variables TR, FC, C1 and C2. TR is the tap ratio, with the digits 3 and 4 indicating trifilar and quadrifilar windings, respectively. FC is the center frequency in megahertz (MHz), and C1 and C2 values are in picofarads (pF). The trial values of C1 and C2 were previously hand-calculated to be 447 pF and 263 pF, respectively, and some preliminary computer trials confirmed that the acceptable range of C1 is between 440 to 450 pF; consequently, this range is used for C1. C2's range was specified to be between 250 to 300 pF. When the program is run with these variables, the result lists six designs each for C1 = 440 and 450 pF, and for C2 ranging from 250 to 300 pF. If necessary, smaller capacitance increments can be used. For this demonstration, a step of 10 for C1 and C2 is adequate.

See **Table 1** for the tabulation of 160-meter BPF parameters for TR = 4 with an L1 and L3 impedance of 800 Ω . Of all the

possibilities, I considered only the design for C1 and C2 = 440 and 250 because I had capacitors on hand to realize those values. The 16.46 μH inductance value is obtained by using a quadrifilar winding on a Micrometals T130-6 core. Examination of the ripple cutoff frequencies of 1.75 and 1.99 MHz, the frequencies of 1.16 and 3.02 MHz at the 35-dB level and the return loss of 30 dB show that all are satisfactory. This design is acceptable, and its design, performance and construction parameters are listed in **Table 2A** for reference.

Table 1—Tabulation of 160 Meter Band-Pass Filter Parameters for TR=4, Z(W)=800

F-C	-FAp (MHz)	+FAp	BW 3 dB (%)	F-	35 dB (MHz)	F+	RL (dB)	RC (%)	C1, C3 (pF)	C2 (pF)	L1, L3 (μH)	L2 (μH)	XL1 (Ω)
1.87	1.75	1.99	27.4	1.16		3.02	30.0	3.15	440	250	16.46	28.97	193
1.87	1.74	2.01	28.1	1.15		3.03	27.7	4.12	440	260	16.46	27.86	193
1.87	1.73	2.02	28.8	1.15		3.05	25.8	5.11	440	270	16.46	26.83	193
1.87	1.72	2.03	29.4	1.14		3.07	24.3	6.11	440	280	16.46	25.87	193
1.87	1.71	2.04	30.0	1.13		3.09	22.9	7.13	440	290	16.46	24.98	193
1.87	1.70	2.05	30.7	1.13		3.11	21.8	8.15	440	300	16.46	24.15	193
1.87	1.75	2.00	27.2	1.17		3.00	28.6	3.70	450	250	16.10	28.97	189
1.87	1.74	2.01	27.9	1.16		3.01	26.6	4.70	450	260	16.10	27.86	189
1.87	1.73	2.03	28.5	1.15		3.03	24.8	5.72	450	270	16.10	26.83	189
1.87	1.72	2.04	29.1	1.15		3.05	23.4	6.76	450	280	16.10	25.87	189
1.87	1.71	2.05	29.8	1.14		3.07	22.2	7.80	450	290	16.10	24.98	189
1.87	1.70	2.06	30.4	1.13		3.08	21.1	8.85	450	300	16.10	24.15	189

Table 2A—Parameters for 160, 80, and 40 Meter Band-Pass Filters

Parameters/Band (MHz)	160 Meters (1.8 - 1.94)	80 Meters (3.5 - 3.91)	40 Meters (7 - 7.3)
Fc, BW (MHz), %BW	1.870, 0.120, 6.42%	3.700, 0.410, 11.1%	7.150, 0.300, 4.20%
F-Ap, +BW Ap (MHz)	1.753, 1.995, 0.242	3.366, 4.067, 0.701	6.740, 7.584, 0.844
RL (dB), RC%	30.0, 3.15%	22.8, 7.22%	30.2, 3.08%
3-dB BW (MHz) and %BW	0.5128, 27.4%	1.174, 31.7%	1.801, 25.2%
Calc freqs (MHz) @ 35 dB	1.16, 3.02	2.18, 6.27	4.60, 11.1
L1,L3; Qu & XL @ Fc	16.46 μH , 195, 193	4.93 μH , 170, 115	3.96 μH , 150, 178
Core & AL (nH/N2)	T130-6, 9.6	T130-17, 4.0	T130-17, 4.0
Total wire turns and gauge	40: 10 #16, 30 #18	33: 11 #16, 22 #18	28: 7 #16, 21 #18
No. and type of turns	10 quadrifilar	11 trifilar	7 quadrifilar
Wire lengths (inches) and gauge	17.5 #16; 45.7 #18	19 #16; 34 #18	13 #16; 33 #18
L2(μH), Q & XL @ Fc	28.97, 320, 340	11.94, 250, 287	8.26, 220, 371
L2a and L2b (μH)	14.8, 14.2	5.97, 5.97	4.13, 4.13
No. turns (a, b) and core type	39, 38 on T130-6	37, 37 on T130-17	30, 30 on T130-17
Wire lengths (inches) and gauge	60 and 59, #18	58 and 57, #18	47 and 48, #18
C1, C3 (pF)	440=110 p 110 p 110 p 110	375=110 p 110 p 110 p 30 p 15	125=(220 s 220) p 15
C2 (pF)	250=110 p 110 p 30	155=31 p 31 p 31 p 31 p 31	60=30 p 30
C1/C2 Ratio	1.76	2.42	2.08

Notes

In the C1 and C3 capacitance declarations, a *p* or an *s* indicates that two or more capacitors are connected in *parallel* or *series*, respectively, to obtain the design capacitance value (ie, 30 p 30 means two 30 pF capacitors are connected in parallel (60 pF total); 15 s 15 means two 15 pF capacitors are connected in series (7.5 pF total).

The 110, 30 and 15 pF capacitors are Tusonix NP0 10% disc ceramics with voltage ratings of 2, 3 and 4 kV, respectively. The 220 pF disc ceramic capacitors are Ceramite NP0 5% 1 kV, type 10TCCT22.

Toroidal cores are Micrometals T130-6 (yellow) or T130-17 (blue/yellow).

L1 and L3 are tuned by injecting a signal at the center frequency into the 50 Ω tap through a 2.2 kΩ resistor. A 50 Ω detector is coupled to the inductor with a one-turn loop and the inductor turns are adjusted for a maximum signal level indication on the detector output meter.

For tuning L2, the assembled filter is terminated at one end with a 50 Ω load while the other end of the filter is connected to a return loss bridge. While monitoring the detected output of the return-loss bridge with an oscilloscope, adjust the L2 windings for an optimum return-loss response over the filter passband.

Table 2B—Parameters for 20, 15 and 10 Meter Band-Pass Filters

Parameters/Band (MHz)	20 Meters (14 - 14.4)	15 Meters (21 - 21.45)	10 Meters (28 - 29.7)
Fc, BW (MHz), %BW	14.88, 2.38, 16%	21.22, 0.450, 2.12%	28.84, 1.70, 5.89%
F-Ap-,+ BW Ap (MHz)	13.74, 16.12, 2.38	20.43, 22.04, 1.605	27.18, 30.60, 3.424
RL (dB), RC%	27.3 4.34	35.1, 1.75	24.4, 6.01
3 dB BW (MHz) and %BW	4.59, 30.9%	4.057, 19.1%	6.035, 20.9%
Calc freqs (MHz) @ 35 dB	8.79, 25.2	15.0, 30.0	20.1, 41.3
L1, L3; Qu, and XL @ Fc	1.27 μH, 140, 119	1.053 μH, 120, 140	0.761 μH, 100, 138
Core and AL (nH/N2)	T130-17, 4.0	T130-0, 1.50	T106-0, 1.90
Total turns and gauge	15: 5 t #15, 10 t #15	20: 5 t #16, 15 t #16	16: 4 t #16, 12 t #16
No. and type of turns	5 trifilar	5 quadrifilar	4 quadrifilar
Wire lengths (inches) and gauge	10.2 and 17.8 15	10 #16; 25 #16	8 #16; 19.6 #16
L2 (μH), Qu and XL @ Fc	3.18, 220, 297	3.75, 180, 500	2.36, 130, 428
L2a and L2b (μH)	1.70, 1.48	1.87, 1.88	1.10 and 1.26
No. turns (a, b) and core type	18, 17 on T130-17	19, 19 on T130-17	14, 15 on T130-17
Wire lengths (inches) and gauge	30, 28.5 #16	31.0, 31.0, #16 green	23.5, 25.1, #16
C1, C3 (pF)	90(30 p 30 p 30)	53.4=(100 s 100) p (6.8 s 6.8)	40.0=(15 p 15) p (20 s 20)
C2 (pF)	36(30 p 30 p 30)s (30 p 30)	15=(15 p 15) s (15 p 15)	12.9=[(15 s 15) p 15] s (15 p 15)
C1/C2 ratio	2.5	3.56	3.1

Notes

In the C1 and C3 capacitance declarations, a *p* or an *s* indicates that two or more capacitors are connected in *parallel* or *series*, respectively, to obtain the design capacitance value (ie, 30 p 30 means two 30 pF capacitors are connected in parallel (60 pF total); 15 s 15 means two 15 pF capacitors are connected in series (7.5 pF total).

The 15 and 30 pF capacitors are Tusonix, NP0, 10% disc ceramics with a 3 and 4-kV voltage rating, respectively. Other capacitors are CeraMite, NP0, 5%, disc ceramics rated at 1 kVDC @ 300 VAC RMS.

All cores are Micrometals T130-17, T130-0 or T106-0.

See notes of **Table 2A** for the suggested tuning procedures.

ELSIE Confirms the Design

The correctness of the design selected from the 160 meter BPF tabulation was further confirmed by evaluating it with a filter analysis program named *ELSIE*, available from Trinity Software. [7] *ELSIE*-calculated plots of return loss and insertion loss are presented in the Appendix (next month) along with the component values used in the computer simulation. The design selected from the BASIC program 160 meter tabulation is confirmed by noting that the *ELSIE*-calculated return loss of 30 dB and the 35 dB frequencies of the plots are *identical* with the computer-calculated values obtained with the BASIC program. In the *ELSIE* insertion-loss plot, the measured insertion loss above 2.2 MHz is greater than the *ELSIE*-calculated curve. I'll explain the reason for this later.

If the first attempts to obtain a satisfactory L1 and L3 value are unsuccessful, try other quadrifilar windings with four more or four fewer turns, and find the exact C1 capacitance needed to resonate the inductor at the center frequency. Then, use the program with C1's value fixed and let C2 vary over a limited range to find an acceptable design. No doubt many possible designs will be available, but only those with a 3 dB percentage bandwidth of less than 32% are narrow enough to be useful.

Toroidal-Core Selection

All inductors used in the BPFs are wound on powdered-iron or phenolic toroidal cores. Although this approach is more expensive than using simpler solenoidal-wound air-core coils, the toroidal type is preferred because of its self-shielding characteristic that allows the inductors to be physically close with little interaction, making for a more compact filter.

For power levels of less than 1 W, core sizes of 0.44 inches outer diameter or less are commonly used. At power levels of 150 to 200 W, however, the core size must be much larger to dissipate the heat resulting from core and winding losses without excessive temperature rise. Many years ago, it was a common misconception that core saturation was the primary limiting factor in high-power RF applications. However, Micrometals' core-loss measurements of their iron-powder cores at high frequencies show that, with sine-wave signals, *excessive temperature rise* resulting from the losses in the winding and core material is the limiting factor. [8, 9] Micrometals specifies the maximum permissible core temperature at 100°C, but any long-term temperature above 90°C accelerates the deterioration of the core's binding material. Consequently, a temperature rise of less than 40°C is preferred so that in an ambient of 90°F (typical temperature for a hot day), the core temperature will be not more than $32 + 40^{\circ}\text{C} = 72^{\circ}\text{C}$, or well below 90°C.

After each BPF design was complete, the filter was assembled and tested under a 200 W continuous load to confirm that the temperature rises of all inductors and capacitors is acceptable. Tim Duffy, K3LR, did most of the early power testing of the BPFs under load. John Brosnahan, W0UN, provided additional power testing later during the development of the BPFs. Both Tim and John provided network analyzer plots of insertion loss and return loss using Hewlett-Packard equipment usually not available to the average amateur experimenter. Their valuable assistance was crucial in confirming both the power capabilities and response performance of the BPFs, permitting the BPF development to progress with the assurance the designs were completely satisfactory. Several times different design variations were tried until Tim was satisfied that the BPF performance would be acceptable for the multi-multi applications.

For the 160 meter BPF, Micrometals T130-6 (yellow) cores are used for L1, L2 and L3. To minimize the temperature rise of L2, it is necessary to use two separate, series-connected inductors. For L1 and L3, a 10 turn quadrifilar winding (40 turns total) of #16 and #18 magnet wire delivered the design inductance of 16.46 μH with a measured Q of 195 at the center frequency. L2 is made of two T130-6 cores with 39 and 38 turns of #18 wire. Its Q measures 320. Details associated with the 160 meter BPF inductor assemblies are listed in the second column in **Table 2A**.

Inductors for the higher bands were designed similarly; their assembly details are listed in Tables **2A** and **2B**. For the 80, 40 and 20 meter BPFs, the L1, L3 and L2 cores are also T130s, but the material is -17 (blue/yellow) with a lower μ than the -6 material. L1 and L3 for the 15 and 10 meter BPFs are wound on T130 and T106 phenolic (tan) cores, while L2 remains two separate series-connected inductors, each wound on a T130-17 core.

Capacitor Selection

The voltage across C1 and C3 is 300 and 400 V RMS, respectively, for a 200 W input to a 50 Ω terminated BPF having a trifilar or quadrifilar-wound L1 and L3. Consequently, the RMS voltage rating of C1 and C3 should be greater than 400 V. Based on a 2 A current flowing through C2, the RMS voltage across C2 can range from a low of 574 V to as much as 1 kV. The C2 voltage depends on the product of two times the reactance of C2 at the center frequency. The C2 voltage rating should be greater than 680 and 580 V for the 160 and 80 meter BPFs, respectively, and even higher for the 40 through 10 meter BPFs because of their higher C2

reactances.

In addition to a safe voltage rating, the capacitors must be capable of handling the current associated with a 200 W RF signal passing through the BPF. For C2, the RMS current level is 2 A. Finally, so that filter construction is economically practical, the capacitors must be readily available and reasonably priced.

I selected Tusonix NP0 10% ceramic capacitors, having an outer diameter of about 16 mm. [10] Although 5% tolerance units cost only a few cents more than the 10% tolerance capacitors, the 10% tolerance is specified so that the trimming operation associated only with the 5% tolerance can be omitted. [11] Harry Roseberry, W1HRZ, of the Tusonix Customer Engineering Dept, advised me that the nontrimmed 10% capacitors are better-suited for RF applications than the trimmed capacitors. Although this capacitor type is not formally rated by Tusonix for RF service, it is nevertheless widely used in nonstringent commercial and Amateur Radio RF applications. To minimize the chance of failure caused by greater-than-anticipated voltages or currents, and to derate the standard dc voltage rating by 50% for RF applications, capacitors with dc ratings of 2, 3 and 4 kV are used. To minimize the temperature rise caused by the 2 A current passing through C2, two or more capacitors in parallel are used at C2. The capacitors and their connection configurations used in each BPF are listed in Tables 2A and 2B.

The values of capacitance and voltage found to be most convenient are 110 pF/2 kV, 30 pF/3 kV and 15 pF/4 kV. Various series and parallel combinations of these three values make it possible to match all the design values. In some cases—because of tolerance variations in the powdered-iron cores used for L1 and L3—a few additional picofarads are needed across C1 and C3 to precisely tune both resonators to the center frequency. To achieve this, two small-value Ceramite 1000 VDC/300 VAC capacitors (type 10TCCQ) are wired in series and placed in parallel across the Tusonix capacitors. The Ceramite capacitors are available in small quantities from Newark Electronics. [12]

Tune In Next Month...

...when I'll tackle filter assembly and tuning, and put the wraps on this project.

Ed Wetherhold, W3NQN, received a degree in Radio Engineering from Tri-State University, Angola, Indiana, in 1956. From 1962 to 1992, he was employed at the Annapolis Signal Analysis Center of Alliant Techsystems, Inc (Alliant Techsystems was formerly the Defense Division of Honeywell, Inc), as a communications systems test engineer and as a certified TEMPEST Professional Level II.

Ed obtained his Amateur Radio license in 1947, while serving in the Air Force as a radio mechanic instructor at Scott AFB, in Illinois. For the past 15 years, he has been a technical advisor to the ARRL on passive LC filters.

Ed's many articles on simplified filter design have been published in the electronics trade and Amateur Radio journals, such as *Interference Technology Engineers' Master (ITEM)*, *QST*, *QEX*, *CQ* and *Practical Wireless*, and in professional EMC journals. *The 1998 ARRL Handbook* contains Ed's SVC filter design tables and an explanation of how to design passive LC filters.

While not working on filters, Ed is active as a tournament tennis player and is ranked Number 1 in the Men's 70 singles and doubles in the USTA's Middle Atlantic section. You can contact Ed at 1426 Catlyn PI, Annapolis, MD 21401, or by telephone at 410-268-0916.

Clean Up Your Signals with Band-Pass Filters

Part 2—You're now much closer to saying goodbye to your unwanted signal problems!

Last month,¹³ I described the design of band-pass filters that provide a level of selectivity and reliability heretofore unavailable. In this installment, I'll delve into filter assembly, tuning and performance.

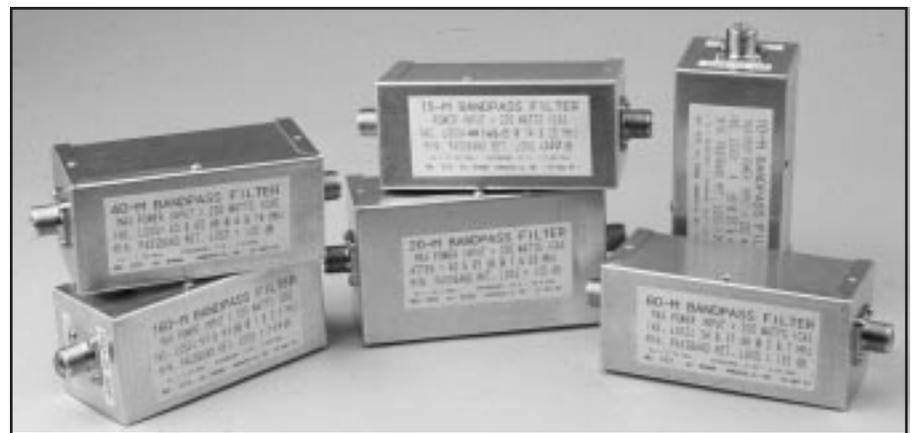
Filter Assembly and Tuning

For the prototypes shown here, 2¹/₄×5-inch (HWD) aluminum boxes are used. While tuning the 20 meter filters, I discovered that the passband return loss was adversely affected when the enclosure cover was installed. That effect was traced to the close proximity between the inductors and the sides of the smaller box. Using a box with a 3-inch width eliminates the problem. Now, except for the 10 meter filter, all the other filters are housed in 2¹/₈×3×5¹/₄-inch (HWD) boxes such as the LMB 880.

The detector I used to tune the resonators has a sensitivity of -12 dBm, and is described in Wes (W7ZOI) Hayward's article, "Beyond the Dipper."¹⁴ Instead of the two hot-carrier diodes specified by Hayward, I use 1N4148 diodes as they are more conveniently available and are adequate for use up to 10 meters.

Before installing C1 and C3, use a digital capacitance meter to ensure they're the same value, preferably within a picofarad. As shown in Figure 1 (see part 1), the 50 Ω taps on the L1 and L3 inductors are connected to the SO-239 UHF connectors; capacitors C1 and C3 are connected between ground lugs and the ends of the inductor windings. The components for the other BPFs are installed similarly. The L2 inductors are secured by tying them to a strip of cardboard that is attached to the box bottom with RTV silicone sealant.

Using the tuning procedure described in the notes of Table 1A, tune resonators 1 and 3 to the design center frequency. When tuning these two circuits, use a digital frequency counter to ensure they resonate as closely as possible to the same frequency. If these circuits are not tuned to the same frequency, it may be impossible to obtain an acceptable



return-loss response. Micrometals specifies a 5% core-permeability tolerance. The resulting L1 and L3 inductance variation is compensated for by squeezing or spreading the turns on the cores until both resonators tune to the same frequency.

To resonate L1 and L3 to within 0.2% of the *design center frequency*, a small amount of capacitance (3 to 10 pF) must usually be added in parallel with C1 and C3. The amount of additional capacitance required is determined by noting the exact values of C1 and C3, and measuring the initial resonant frequency. The required additional capacitance is equal to

$$\{[(F1/Fc)^2] \times C1\} - C1 \quad (\text{Eq 1})$$

For example, assume C1 and C3 of the 40-meter BPF design both measure 123 pF, and C1 and L1 and C3 and L3 each resonate initially at 7.30 and 7.35 MHz. Squeeze and spread the turns on L1 and L3 until the circuits resonate at the same frequency, say, 7.33 MHz. Because C1 and C3 were previously matched to exactly the same value, L1 and L3 now must also be of equal value because the 1 and 3 resonators are both tuned to 7.33 MHz. Eq 1 is used to find the required additional capacitance to add to C1 and C3:

$$Ca = (7.33/7.15)^2 \times 123 - 123 = 6.3 \text{ pF} \quad (\text{Eq 2})$$

A close approximation of this value can be obtained by connecting in series two 12 pF, 1 kVDC Ceramite capacitors. Any final resonator adjustments are made by squeezing and spreading the L1 and L3 inductor turns until both resonators are tuned to the same frequency with 0.2% of the design value.

After resonators 1 and 3 are tuned, note the resonant frequency and preliminarily tune resonator 2 to this frequency to find the proper number of turns on L2A and L2B. Do this by grounding the center pins of both SO-239 connectors to short resonators 1 and 3 to ground. Then, couple a test signal and detector to L2A and L2B using single-turn loops. Add or remove turns on L2 until peak output is obtained at the design center frequency. Final tuning of L2 is done by squeezing and spreading turns on L2A and L2B while using an oscilloscope to observe the passband return-loss response. (See the Appendix for details of the return-loss measurement procedure.)

Tuning is complete when a three-peak passband return-loss response—typical of a third-order band-pass filter—is obtained.

¹³Notes appear on page 41.

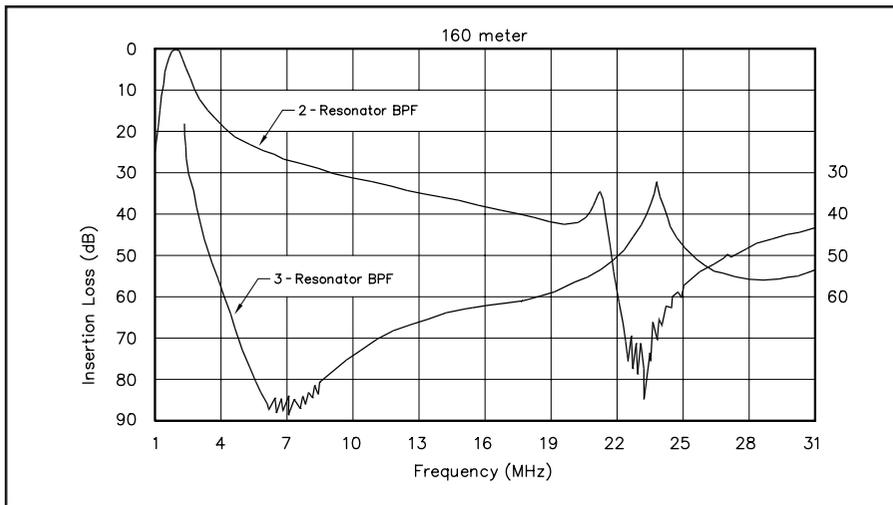


Figure 2— Insertion-loss responses of a commercial 160 meter, two-resonator BPF, and the 160 meter, three-resonator BPF described here. Note how the three-resonator BPF exhibits substantially more loss than the two-resonator BPF between 4 and 10 MHz.

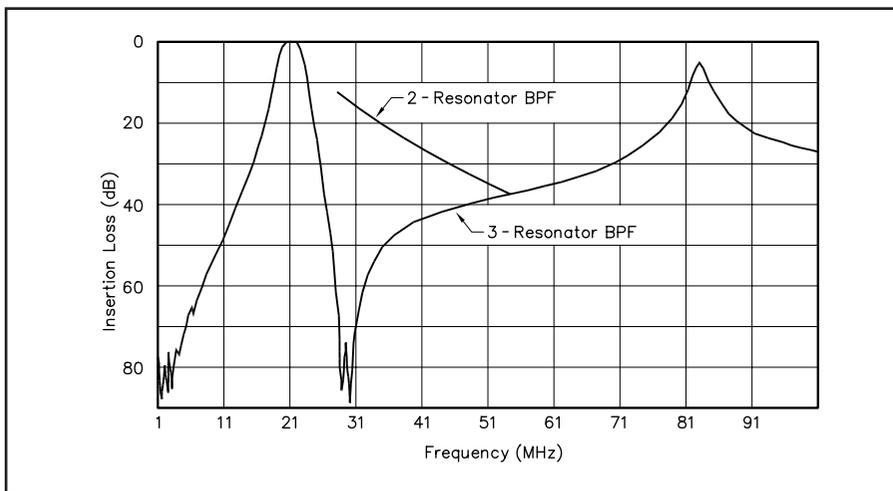


Figure 3— Insertion-loss responses of a commercial 15 meter, two-resonator BPF and the three-resonator BPF. The three-resonator BPF has more than 60 dB loss in the 10 meter band compared to less than 15 dB for the two-resonator BPF, and the three-resonator BPF loss is greater than the two-resonator BPF up to about 52 MHz.

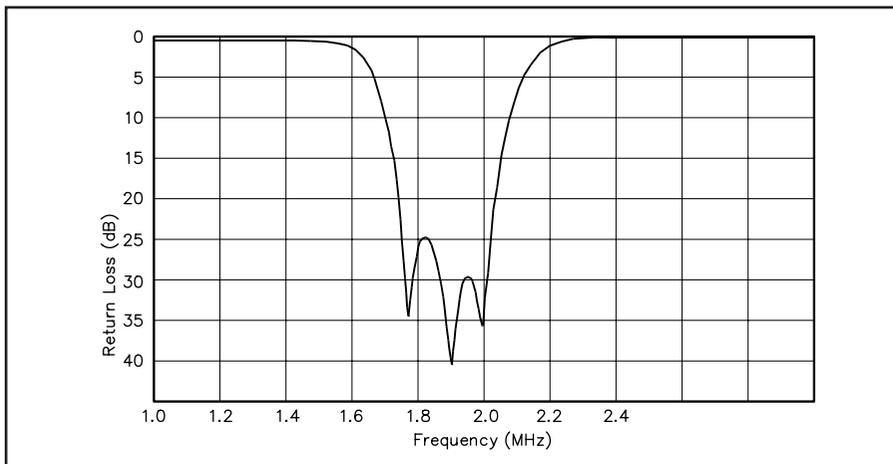


Figure 4— Passband return-loss response of the 160-meter, three-resonator BPF. The three sharp peaks in the passband are typical of the three-resonator BPF. Return-loss minimums of a perfectly tuned BPF are at the same level; however, the minimum return loss of 25 dB indicates that the tuning is quite satisfactory.

The two valleys of minimum return loss should theoretically be equal to the calculated return-loss value listed in Table 1; however, any passband minimum return loss greater than 22 dB is quite acceptable. For the 160 meter BPF, the calculated minimum return loss is 30 dB.

BPF Passband and Stopband Performance

The stopband insertion-loss responses of the 160 and 15 meter three-resonator BPFs are shown in Figures 2 and 3. The stopband insertion-loss response of the commercial two-resonator BPF is included for comparison to illustrate the superior performance of the three-resonator BPF. The superior performance of the three-resonator BPF is mainly due to the series-coupling circuit between resonators 1 and 3 having an inductor in addition to a capacitor, whereas the two-resonator BPF has only a capacitor. Consequently, at frequencies above the upper passband cut-off frequency, the two-resonator BPF looks like a high-pass filter and provides little attenuation to the higher frequencies, whereas the series inductor in the three-resonator BPF provides increasingly greater attenuation as the frequency increases. The stopband responses of the other three-resonator BPFs are similar to the 160 meter and 15 meter BPFs.

Comparing the 160 meter BPF measured insertion-loss response with that of the computer-calculated response in Figure A1 (see the Appendix), you can see that the increase in the measured response is more abrupt than the calculated response. This abrupt attenuation rise is typical of all six BPFs and is attributed to the effect of imperfect coupling between the L1 and L3 50 Ω tap and the rest of the windings.¹⁵ As the tap approaches the top of the coil, the effect of imperfect coupling becomes less and less, so when the tap reaches the top of the coil, the calculated and measured responses are virtually identical. Of course, the component values and impedance level will then be impractical.

In Figure 3, the undesired abrupt drop in stopband attenuation of the 15 meter BPF above 30 MHz is also attributed to the less-than-perfect coupling between the tapped windings of L1 and L3 (see Note 15). However, the use of the quadrifilar winding in L1 and L3 minimizes the effect of the imperfect interwinding coupling and produces an acceptable loss response of more than 40 dB between 35 and 50 MHz. Compare this 40 dB level of attenuation with the substantially lower 14 to 16 dB attenuation of the 15 meter BPF shown in Figure 5, page 36 of the June 1994 *QST* article (see Note 4).

As the frequency decreases below the lower BPF cut-off frequency, the increasing attenuation is caused primarily by the increasing reactance of the series capacitor, and the series inductor of the three-resonator BPF becomes increasingly ineffective. Consequently, the low-frequency attenuation responses of the two and three-resonator BPFs are similar.

Depending on the band, the measured

passband insertion loss of the three-resonator BPF ranges from 0.25 to 0.50 dB. The corresponding power dissipated in the BPF, relative to a 200 W input, is about 11 and 22 W, respectively. The major portion of the power loss occurs in the inductor cores, but there is sufficient surface area in the 1.30-inch powdered-iron cores to dissipate this amount of power without an excessive rise in temperature.

Figure 4 shows a network analyzer plot, provided by WØUN, of the passband return-loss response of the 160 meter, three-resonator BPF. Because being able to observe the return-loss response is crucial in making the final adjustment to L2, it soon became obvious that this particular test had to be performed at the time of assembly. Consequently, I developed a return-loss test procedure using circuits similar to those described by Randy Henderson, WI5W,¹⁶ for testing all the BPFs. Details of the return-loss test procedure and the associated circuits are described in the Appendix. An example of the performance of these filters used during the 1997 ARRL International CW DX Contest is indicated by the comment of N6RO: "The combination of double stubs (tnx K2KW) and W3NQN band-pass filters (tnx K3LR) provided interference-free op-

eration for the first time ever."¹⁷

Summary

The new BPF design, construction and testing techniques discussed in this article advance the current state of the Amateur Radio art. These new techniques should serve as a useful guide until they are superseded by future improvements.

Assembled and tested band-pass filters are available from me. Send me a business-size, self-addressed, stamped envelope for details.

Acknowledgments

A project of this size and length could not have been completed without the assistance of many others. I am grateful to Tim Duffy, K3LR, for proposing this project and providing recommendations based on his unique experience in running one of the top multi-multi contest operations. His network analyzer plots and heating tests assured that the final BPF designs would meet the special requirements of the multi-multi contest. Tim's contacts with the multi-multi contest-operator fraternity was crucial in distributing a number of BPF sets to those needing a more reliable BPF than currently available. During the later development and testing

phase, John Brosnahan, WØUN, provided additional test data with his network analyzer plots and heat run tests.

My thanks to Frank Glatz, Regional Sales Manager of Tusonix, for providing samples of Tusonix capacitors. The samples permitted testing of prototype designs to confirm that the capacitor types and values selected were satisfactory before quantity orders were placed. Harry Roseberry, W1HRZ, of Tusonix Customer Engineering assisted in selecting the proper capacitor types.

Dale Nicol of Micrometals, Inc. discussed with me the temperature limitations of powdered-iron cores, and Micrometals provided the many samples I needed to find the most suitable cores for each BPF design.

Notes

¹³Ed Wetherhold, W3NQN, "Clean Up Your Signals with Band-Pass Filters," QST, May 1998, pp 44-48.

¹⁴Wes Hayward, W7ZOI, "Beyond the Dipper," QST, May 1986, p 17, Figure 6.

¹⁵Philip Geffe, *Simplified Modern Filter Design*, p 32, Figure 4-4, "Tapped Coil Hazards"; John F. Rider Publisher, Inc, New York, 1963.

¹⁶Randy Henderson, WI5W, "A Swept-Frequency Generator for Crystal-Filter Evaluation," QEX, Mar 1994, pp 3-8. See page 6, Figure 6, for the deflection and sweep circuit.

¹⁷1997 ARRL International DX Contest CW Results, QST, Sep 1997, p 112, third column (under photo).

Appendix

Return-Loss Measurement Equipment and Procedures

Figures A2 and A3 show the deflection and sweep circuit and a block diagram of the equipment used in the return-loss testing of the BPFs. For return-loss testing of the 160, 80, 40, 20 and 15-meter BPFs, I used a

B&K Precision 4040 sweep generator. Because that sweep generator has an upper-frequency limit of 26 MHz, I used a Mini-Circuits POS-50 VCO driven by the deflection and sweep circuit shown in Figure A2

when testing the 10 meter BPFs.

Except for revising the deflection and sweep output circuit with the addition of an **AUTO/MAN** switch and using smaller capacitors to increase the sweep rate and ramp

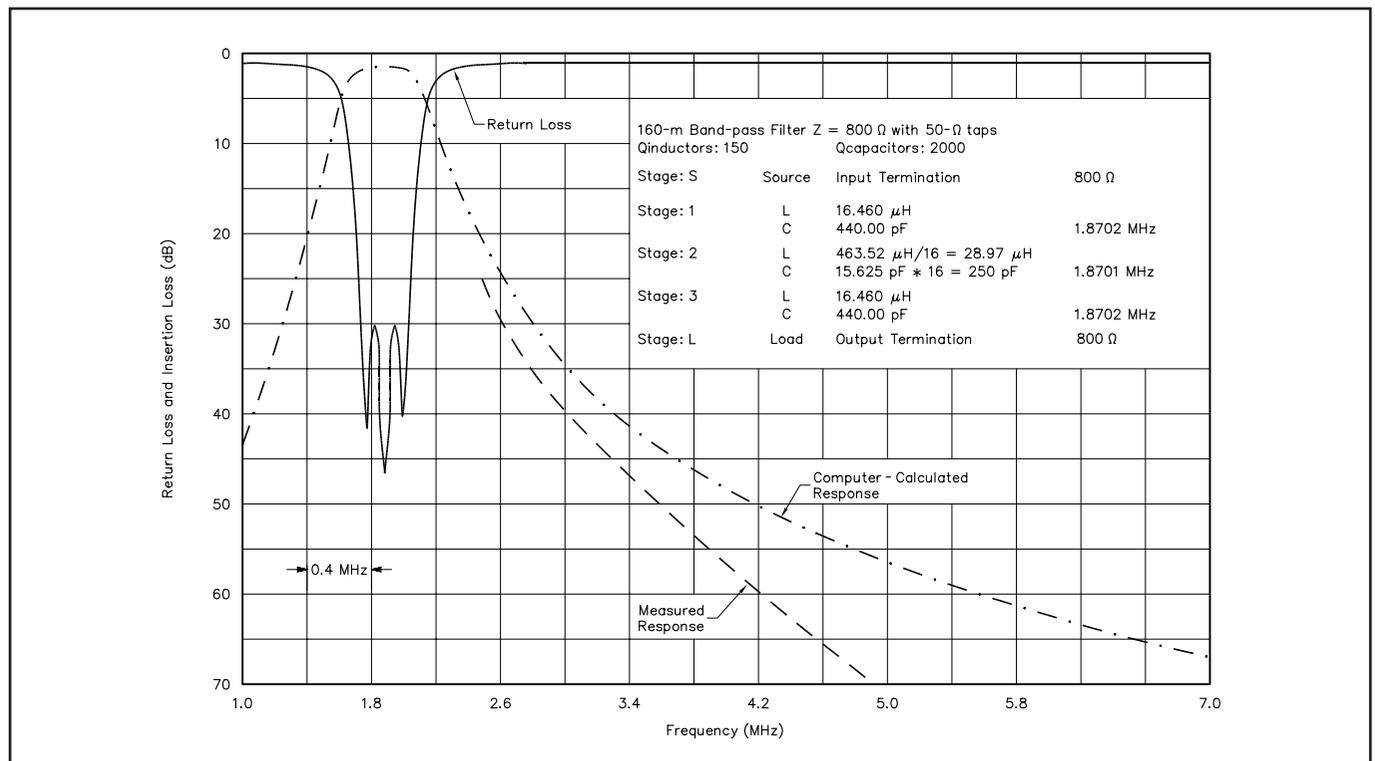


Figure A1—Computer-calculated and plotted return loss and insertion loss responses of a 160 meter band-pass filter selected for assembly.

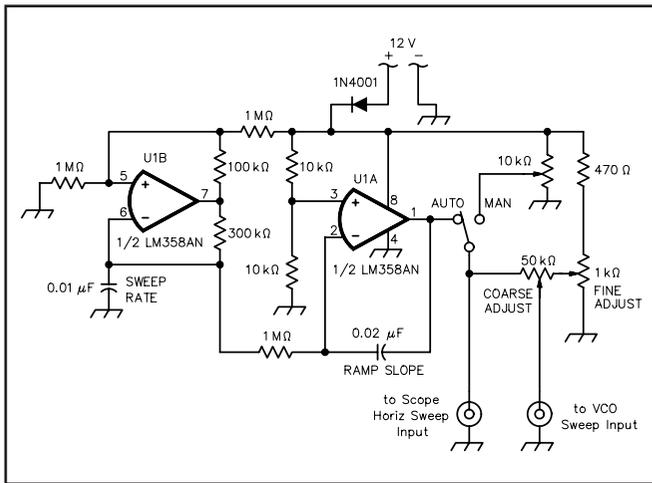


Figure A2—Deflection and sweep circuit used with the Mini-Circuits POS-50 VCO for return-loss testing of the 10 meter BPF.

slope, it's the same as that used by Randy Henderson, W15W, in his March 1994 *QEX* article.* It was from Randy's article that I got the idea of using a VCO with a deflection and sweep circuit, a return-loss bridge, a log RF detector and an oscilloscope to observe the return-loss responses of the BPFs. I used commercially available swept oscillators and Zack Lau's RF log detector circuit† instead of the four-stage amplifier and detector shown in Henderson's article. A similar article in *Popular Electronics*,‡ features a sweep oscillator and an RF detector for filter testing.

For the 10 meter BPF tests, I connected the +7 dBm output of the VCO to the input of an RF return-loss bridge (RLB) described on pages 26.41 and 26.42 of *The 1997 ARRL Handbook*. The UNKNOWN Z port of the RLB attaches to the input of a 50-Ω-terminated BPF and the RF OUTPUT port con-

nects to the log RF detector.

After some trial and error, a combination of sweep rates and 'scope adjustments produced a swept-frequency return-loss response that related to adjustments made to the windings on the center resonator inductor. Using this equipment and return-loss test procedure, it was possible to adjust all the BPFs to obtain either a three-peak passband return-loss response, or a minimum passband return loss greater than 23 dB indicating that optimum BPF tuning was achieved.

By replacing the BPF with a 60 Ω resistor in the return-loss test setup, I obtained a 20 dB return-loss reference curve against which the BPFs' return loss could be compared. Whenever the minimum return-loss response of a BPF is below the 20-dB refer-

ence curve, it indicates that the BPF is not properly tuned and further adjustment of the L2A and L2B windings is necessary.

*Randy Henderson, W15W, "A Swept-Frequency Generator for Crystal-Filter Evaluation," *QEX*, Mar 1994, pp 3-8. See page 6, Figure 6, for the deflection and sweep circuit.

†Zack Lau, W1VT, "A Logarithmic RF Detector for Filter Tuning," *QEX*, Oct 1988, pp 10-11. (Note: In Figure 1, p 10, the ground connection of pin 1 is missing.)

‡John J. Yacono's column *Think Tank* in *Popular Electronics*, Feb 1997, pp 77-78; letter from Douglas Ripka, Rebersburg, Pennsylvania: "RF Signal-Strength Circuit." Figure 4 shows a circuit featuring the Motorola MC3356P RSSI IC (received signal-strength indicator) used as an RF log detector to drive the vertical input of an oscilloscope. **Q57-**

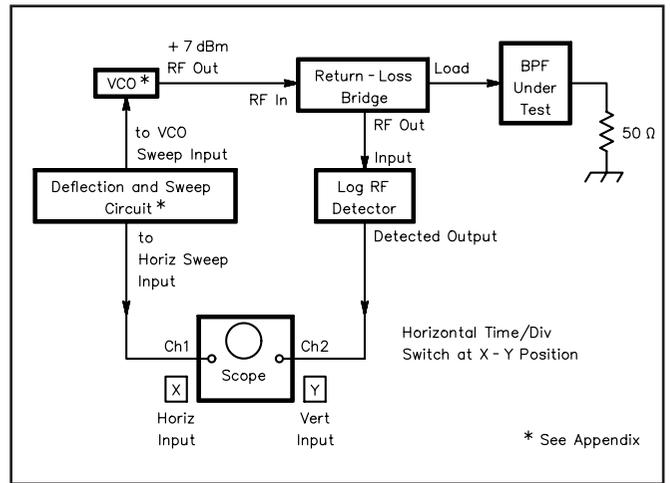


Figure A3—Block diagram showing the equipment arrangement used for measuring the return-loss testing of the BPFs. A B&K Precision Model 4040 sweep generator was used as the VCO on the 160, 80, 40, 20 and 15-meter bands. On 10 meters, a Mini-Circuits POS-50 replaced the B&K Precision unit. The oscilloscope is a TENMA Model 72-3055, with a 20 MHz bandwidth.

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