Narrow Band-Pass Filters for HF

Band-pass filters can be critical components in competitive stations. This setup may help put your station on the map.

By William E. Sabin, W0IYH

There are nine relatively narrow HF Amateur Radio frequency bands. In homebrew equipment designed for these bands, a narrow band-pass filter (NBPF) that attenuates frequencies above and below a particular band can be very useful. Harmonic, sub-harmonic, image, intermodulation, overload and mixer spurious products (harmonic intermodulation) are problems that these filters can greatly alleviate in receivers and transmitters. This article describes simple filters at medium cost and performance levels that are suitable for many of the kinds of homebrew projects Amateurs build. They can be cascaded in filter-amplifier-filter arrangements for highly advanced performance.

Construction details, including simulated frequency responses of the filters, can be downloaded from the *QEX/Communications Quarterly* Web page.[1] The plots can be studied to see if they are adequate for the task at hand. A 32 MHz low-pass filter is included that provides additional attenuation beyond the HF region. My actual filters agree quite closely with simulations down to the –60 dB level, except for small differences in the passbands.

[1]Notes appear on page 17.

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The filters use two resonators. In the interest of simplicity, low parts count, low dc power consumption (0.9 W for any number of filters, one at a time) and low internally generated intermodulation distortion (IMD), a pair of inexpensive, miniature RadioShack SPDT relays (275-241) is used in each filter. A PIN-diode switching approach for lower-level applications will be discussed later. Fig 1 shows simulated, idealized responses of three types of two-resonator filters. One has greater selectivity on the "high" side and one is better on the "low" side. These types are quite useful in various applications. The symmetrical response is not as easy to implement in practice in a NBPF, but very easy on a computer. Because the frequency scale is logarithmic, the shape of these plots is constant as they "slide" horizontally.

The filters will deliver 10 W continuous output with negligible warming. On each band, and at S = +37 dBm (5 W) input for each of two in-band tones, a third-order input intercept point (I_3 in dBm) was determined. A 40 dB each-tone-to-intermod ratio (IMR_3) computes to an I_3 of about +57 dBm, using Eq 1. A value of IMR_3 for other values of input per tone can be estimated also from Eq 1. Do not "hot-switch" the relays. I also suggest using type-2 (μ=10) or type-6 (μ=8) powdered-iron cores.

$$I_3(\mathrm{dBm}) = S(\mathrm{dBm}) + 0.5 \bullet IMR_3(\mathrm{dB})$$
$$IMR_3(\mathrm{dB}) = 2.0 \bullet \left[I_3(\mathrm{dBm}) - S(\mathrm{dBm})\right] \quad \text{(Eq 1)}$$

NBPF Circuits

Fig 2A is a typical high-side filter. The shunt C_S couples the two resonators. Since its reactance decreases as frequency increases, the resonators become more isolated from each other at higher frequencies. Fig 2B uses a top-coupling capacitor C_T and the low side is improved because the reactance of C_T increases at low frequencies. Capacitive dividers C1 and C2 provide a desirable broadband interface with adjacent circuits.

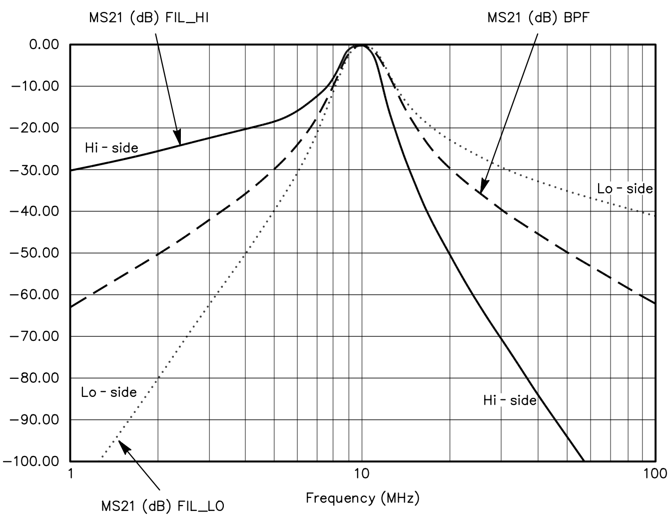


Fig 1—*ARRL Radio Designer* predicted response of symmetrical, low-side and high-side band-pass filters.

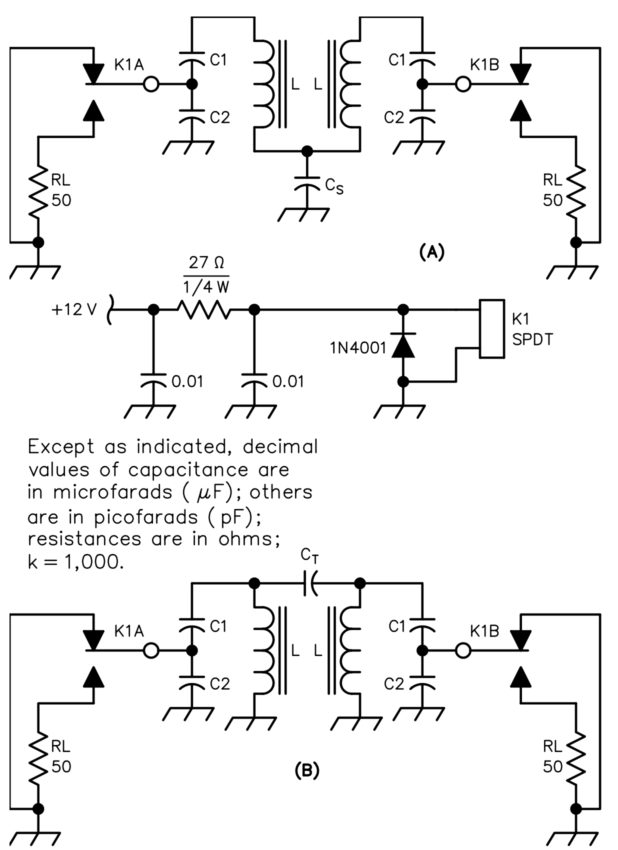


Fig 2—Two-resonator NBPF circuits. (A) is a bottom-coupled "high side" filter; (B) is a top-coupled "low-side" filter.

The filters are designed to operate between two 50-Ω resistances, but we begin the design with R much greater than 50 Ω. The filters are based on the Butterworth approach. There are certain approximations involved in the design of narrow-band coupled resonators[2,3,4] that are related to the ways that coupling reactances and impedance-transforming networks vary with frequency. The method used here gets very close to the final filter using simple design equations and a program like *Mathcad,* then tweaks the design with *ARRL Radio Designer*. A simple test setup is used to make final adjustments to the hardware.

Low-pass and NBPF Prototypes

Comparisons between Chebyshev and Butterworth filters

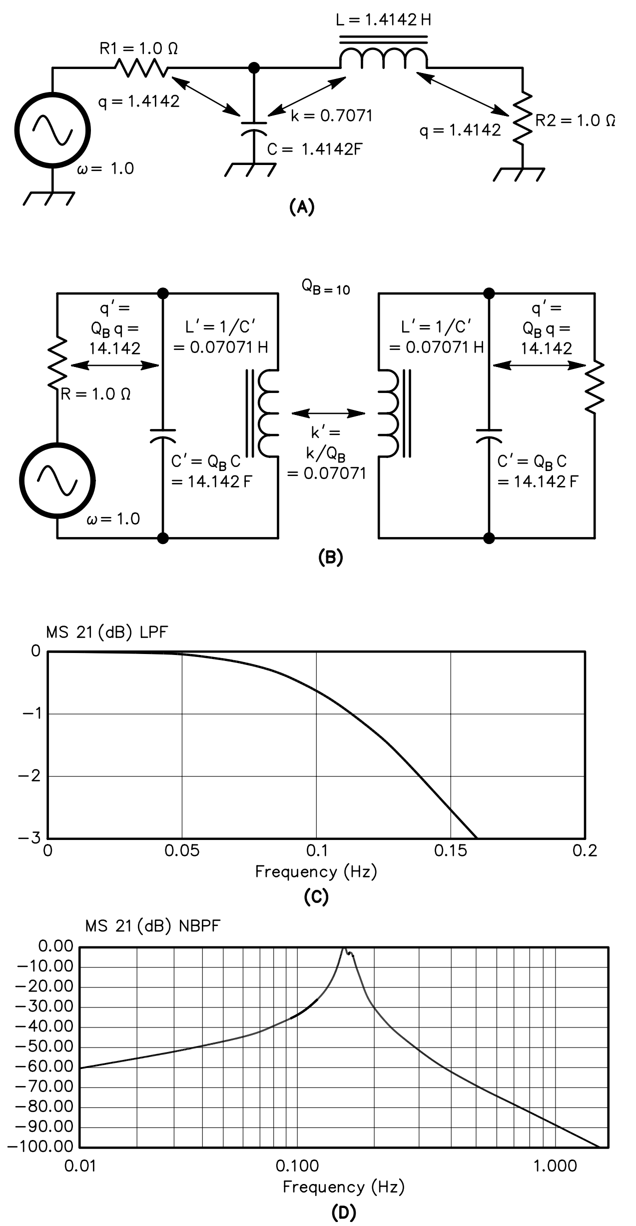


Fig 3—Prototype filters: (A) low-pass filter; (B) narrow band-pass filter; (C) a low-pass response; (D) a narrow band-pass response.

led me to the Butterworth as a better choice for the NBPF. We use two numbers to get started: q and k. For the Butterworth, q = 1.4142 and k = 0.7071. The significance of these numbers is seen in Fig 3A, the two-element Butterworth prototype LPF from which the NBPF is derived. Variable q is the "Q" of C in parallel with R1 at ω = 1.0:

$$q = R\frac{1}{X_C} = R \bullet \omega \bullet C = (1.0)(1.0) \bullet C = 1.4142;\ C = 1.4142\ F \quad \text{(Eq 2)}$$

k is the "coefficient of coupling" (explained later) from C to L; L is given by:

$$L = \frac{1}{\omega^2 k^2 C} = \frac{1}{(1.0)^2(0.7071)^2(1.4142)} = 1.4142\ H \quad \text{(Eq 3)}$$

$R2$ is found by noting that q is also the Q of L in series with $R2$ at ω =1.0:

$$R2 = \frac{\omega L}{q} = \frac{(1.0)(1.4142)}{1.4142} = 1.0\ \Omega \quad \text{(Eq 4)}$$

The response of Fig 3A is down 3 dB at ω =1.0 rad/s (0.1592 Hz) as shown in Fig 3C.

The numbers q and k also apply to the NBPF. If the coefficient of coupling (k in Fig 3A) could be achieved without a direct connection between C and L, we could remove this connection. First, we must do the following: Multiply q by a large number, which we call Q_B, for example 10, and divide k by Q_B. To keep the discussion brief, Fig 3B shows the method and the component values. The response (Fig 3D) of Fig 3B is centered at ω = 1.0 rad/s (0.1592 Hz). Close to this frequency, above and below, the response is very nearly the Butterworth. Far away from the center frequency the similarity changes; that is, the NBPF is a *narrow-band approximation* to Butterworth *near* ω = 1.0. At low frequencies, the response is –20 dB per decade and at high frequencies the response is –40 dB per decade. The 3-dB bandwidth of Fig 3D is nearly:

$$\frac{1}{Q_B} = 0.1 \text{ rads/sec (0.0159 Hz)} \quad \text{(Eq 5)}$$

The next task is to scale this NBPF prototype to its final HF values. For small HF filters that use low-cost inductors and capacitors, values of Q_B from 4 to 30 are practical.

Designing the Two-Resonator NBPF

Refer now to Fig 4. We need the value of Q_B for the final filter

$$Q_B = \frac{F_0}{BW_{3dB}} = \frac{\sqrt{F_{HI} \bullet F_{LO}}}{F_{HI} - F_{LO}} \quad \text{(Eq 6)}$$

where F_{LO} and F_{HI} are the 3-dB edges of the filter's passband. F_{LO} and F_{HI} are positioned so that the frequency response over the amateur band varies no more than a few tenths of a decibel. They should also be selected initially so that the filter is centered near the geometric center of the amateur-band limits. For example, F_0 (in MHz) for the 80-meter band is $\sqrt{3.5 \bullet 4.0} \approx 3.74$. We will need to fine-tune these numbers later. We now need Q_N, the Q of each resonator when loaded by R, and K, the coefficient of coupling between the resonators.

$$Q_N = q \bullet Q_B;\ K = \frac{k}{Q_B} \quad \text{(Eq 7)}$$

In Fig 4, there are three values to be determined: C, L and R, which are related as shown in Eq 8:

$$C = \frac{1}{\omega_0^2 \bullet L} = \frac{Q_N}{\omega_0 \bullet R};\ \omega_0 = 2\pi\sqrt{F_{HI} \bullet F_{LO}} \quad \text{(Eq 8)}$$

For a selected value of Q_N, it is clear from this equation that after C is chosen, L and R are both determined. The goal is to make all three "reasonable" values that are inexpensive and appropriate for the particular frequency band. For example, we would not choose C = 1000 pF for the 10-meter band because that would make L unreasonably small and difficult. We would not choose C = 10 pF for the 160-meter band. Experience and "feel" are valuable tools for this. Having made an educated choice for C, L can then be achieved by winding the right number of turns on the right toroid core. R is then constrained to the value found in Eq 8 and should be between 500 Ω and 2500 Ω.

In Fig 4, the next step is to couple the two resonators so that the desired bandwidth and passband response are achieved. For the top-coupled filter (as in Fig 2B), C_T is used and C is reduced to the value C':

$$C_T = \frac{C \bullet k}{Q_B} = C \bullet K;\ C' = C \bullet (1 - K) \quad \text{(Eq 9)}$$

For the shunt-coupled filter (as in Fig 2A), C_S is used and L is increased to L':

$$C_S = \frac{1}{K \bullet \omega_0^2 \bullet L};\ L' = L \bullet (1 + K) \quad \text{(Eq 10)}$$

Refer to Fig 5. In Eq 8 we found that, having chosen a reasonable C, R is determined. So the final step is to use capacitive dividers to transform R to R_L = 50 Ω; but first, R must be broken up into two parts. One is the resistance of the coil and the other is R_S, the external loading resistance as shown in Fig 5.

$$R_S = \frac{1}{\left(\frac{1}{R} - \frac{1}{\omega_0 \bullet L \bullet Q_L}\right)};\ \omega_0 \bullet L \bullet Q_L >> R \quad \text{(Eq 11)}$$

where Q_L is assumed to be known by measurement at ω_0. Note that the coil resistance must be much greater than R, which implies that L and Q_L cannot be too small. R_S is then to be transformed to R_L. For the capacitor dividers, I use exact equations rather than the approximate ones that are found in many references. The values of C2 and C1 (C_F is defined below) are given by:

$$C2 = \frac{\sqrt{\frac{R_L}{R_S} \bullet \left[1 + (\omega_0 \bullet C_F \bullet R_S)^2\right] - 1}}{\omega_0 \bullet R_L};\ C1 = \frac{1 + (\omega_0 \bullet C2 \bullet R_L)^2}{\omega_0^2 \bullet R_L \bullet (C_F \bullet R_S - C2 \bullet R_L)} \quad \text{(Eq 12)}$$

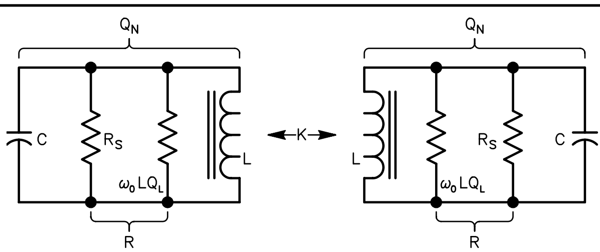


Fig 4—Coupled, loaded resonators.

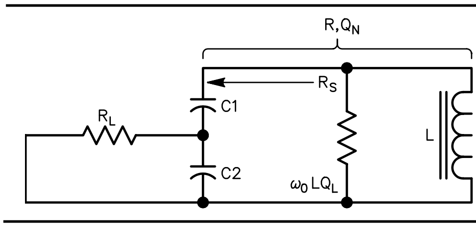


Fig 5—A capacitor-divider impedance transformer.

If different values of R_L at either end of the filter are desired, Eq 12 can be used to find new values for C1 and C2, with no changes elsewhere. Two conditions must be satisfied in Eq 12:

$$\frac{R_L}{R_S} \bullet \left[1+(\omega_0 \bullet C_F \bullet R_S)^2\right] > 1 \text{ and } \frac{R_S}{R_L} > 1 \qquad \text{(Eq 13)}$$

In Eqs 12 and 13, C' from Eq 9 is the correct value of C_F to use if top coupling is used. C from Eq 8 is the correct value if C_S is used. C_S replaces a coupling inductor L_M as shown in Note 3, Fig 6.15. This substitution causes very little error within the narrow passband but greatly increases high-side attenuation, as Fig 1 shows.

Final Design

The filter design is now almost complete and looks like Fig 2A or 2B, and the losses due to the coils (measured Q_L from 160 to 220) have been adequately accounted for. The coil losses affect the attenuation in the passband, which is to be 2 dB or less. I adjusted Q_B (Eq 6) and C (Eq 8) and used *ARRL Radio Designer* and my lab equipment to get the desired passband response and to get standard C values, if possible. I then slightly adjusted L to get reasonably close to the desired response. There is also a small mismatch loss because the input and output impedances are not exactly 50 Ω. *Mathcad* quickly recalculates the component value improvements using the equations in this article. I run *Mathcad* and *Radio Designer* simultaneously and click back and forth. One problem to avoid is making the passband too narrow, in which case the passband attenuation increases more than we might want. In general, it is much better to let the powerful software that is available take care of the design tweaking than to get involved in a purely experimental approach. The *Mathcad* (a spreadsheet program is also good) and *Radio Designer* worksheets that I used are included in the data package (see Note 1) and are very convenient for those who want to design or modify filters. For more in-depth material on the NBPF, look at the references in Notes 2, 3 and 4.

Core Flux

For a power input of P_{IN} watts, the capacitive divider increases the input voltage V_{IN} to V_{TOP} at the top of the coils according to Fig 6:

$$V_{TOP} = \sqrt{P_{IN} \bullet R} \qquad \text{(Eq 14)}$$

and the capacitors are appropriately rated. The question occurs whether the powdered-iron-core inductors will have too much flux at the higher impedance. The answer is no; for a *specific* core, the flux is nearly constant. Suppose one circuit has resistance value R_A and another has R_B. The voltage ratio is:

$$\frac{V_B}{V_A} = \sqrt{\frac{R_B}{R_A}} \qquad \text{(Eq 15)}$$

and the inductance ratio is:

$$\frac{L_B}{L_A} = \frac{R_B}{R_A} = \left(\frac{N_B}{N_A}\right)^2 \qquad \text{(Eq 16)}$$

The flux, ϕ, in a particular core is equal to a constant, K_ϕ, times the volts per turn, V/N, of the winding. Combining Eqs 15 and 16 into this relationship, we get the resulting flux ratio:

$$\frac{\phi_B}{\phi_A} = \frac{V_B}{V_A} \bullet \frac{N_A}{N_B} = \sqrt{\frac{R_B}{R_A}} \bullet \sqrt{\frac{R_A}{R_B}} = 1 \qquad \text{(Eq 17)}$$

which is approximately correct for the filters described in this article. This question often occurs because of the influence of core flux levels on nonlinearities.

Construction

Fig 7A is a close-up of two of the high-side (shunt-coupled) filters and Fig 7B shows two of the low-side (top-coupled) filters. Each filter is 1×4 inches, and a standard 4×6-inch PC board provides five individual filter boards. These individual boards are available in any quantity from FAR Circuits.[5] Fig 8 shows the construction of my filter assembly on a 5×7×2-inch chassis. The band switch, the low-pass filter and the method of mounting the filters, five to each side, are shown. The idea was to minimize the chassis footprint of the filter assembly by using the vertical style of construction.

The filters should be connected by short lengths of miniature 50-Ω coax. This method is somewhat tedious to implement, but helps to preserve the 50-Ω interface and

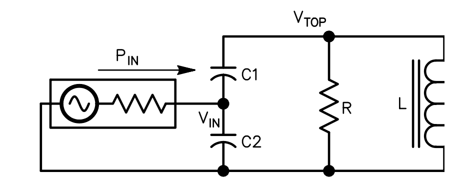
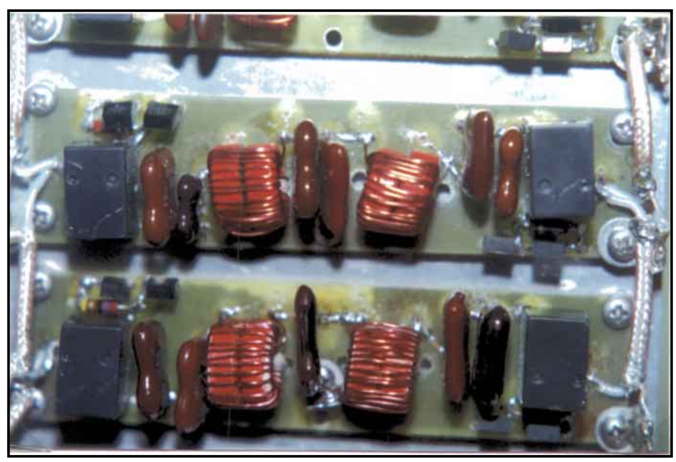
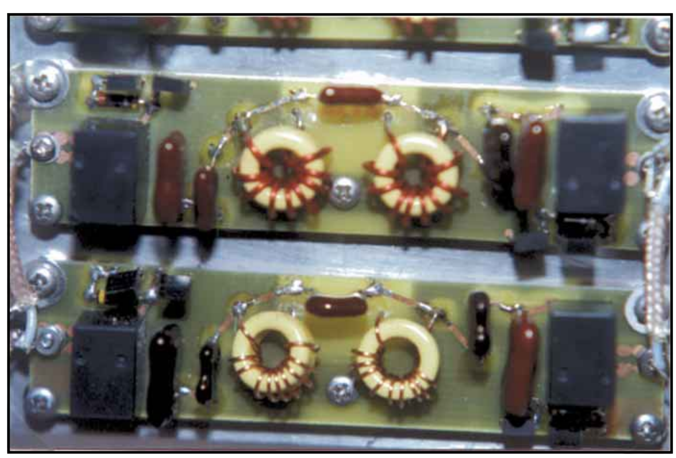


Fig 6—Voltage and flux transformation in a capacitive divider.



(A)



(B)

Fig 7—(A) Shunt-coupled NBPFs. (B) Top-coupled NBPFs.

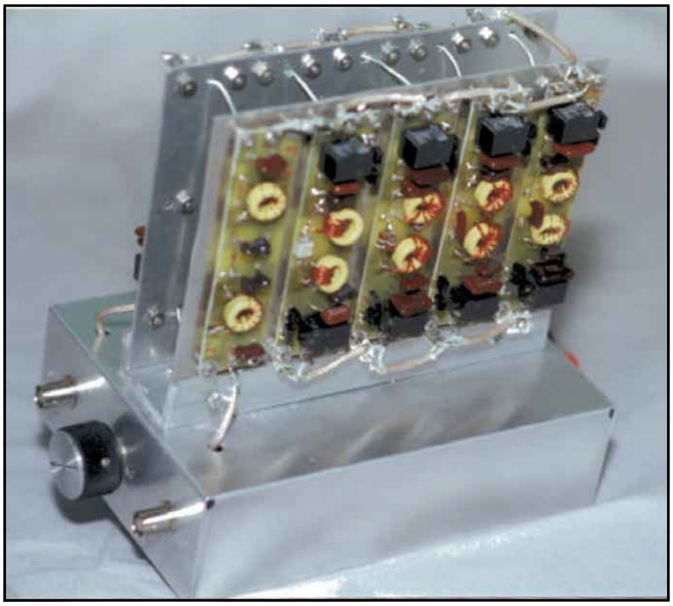


Fig 8—A complete NBPF assembly. The 32-MHz LPF is at the left.

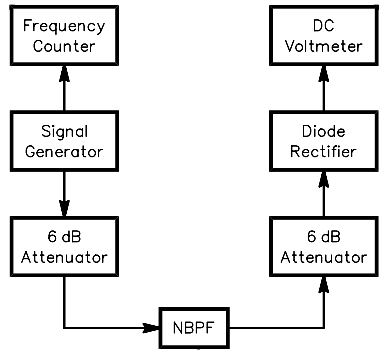


Fig 9—Block diagram of a test setup to adjust the NBPF passband.

effectively reduces stopband leakage. The **IN** connector coax goes to the 160-meter input first, then to the other inputs. The filter outputs all go to the LPF input and the LPF output goes to the **OUT** connector (of course, **IN** and **OUT** may be reversed). Notice the single-point grounding of the coax braids at the input and output of each filter (verified effective). The relay-coil switching can be done electronically, under software control, in an actual application. The low-pass-filter board is an NBPF board, slightly modified (see Fig 8); a separate board design is not necessary.

Fig 9 is a simple test setup that can be used to finalize the passband response and is highly recommended if the use of a spectrum analyzer and tracking generator is not feasible. The capacitors should ideally be within 2% of the values that are suggested in the datasheets. A digital or analog capacitance meter that has an accuracy of better than 1% is a valuable asset for filter construction. Because of the tolerances of capacitors, this selection process is a source of some difficulty that requires patience and an assortment of parts from which to choose. If necessary, use two capacitors in parallel: a "main" low-side value and a small "tweak" value. You can modify slightly the Cs values in Eqs 6, 7, 8, 9, 10, 11, 12 and 13 to values that match what you have on hand or can easily get. Use *Radio Designer* to fine-tune the inductance, using the capacitance values you have chosen.

Filter tests within the passband, using the setup of Fig 9, will require some minor adjustments of the inductors by spreading or compressing turns. I found this procedure to be less effective on the 160 and 80 filters than on those for the higher frequency bands. A turn more or less on the toroid cores may be indicated (start with an extra turn and remove it if necessary). Experimental adjustments of C_S (for shunt-coupled) or C_T (for top-coupled) can be made to fine-tune the shape of the passband, if necessary. The accumulation of small uncertainties ("fuzziness" is the operative word these days) in the actual filter module often makes this process desirable and quite permissible.

Here are some suggestions that will improve the broadband attenuation of the filter boards. For the high-side filter, Fig 7A, remove the printed-circuit traces that go the location where a top-coupling capacitor C_T would be located. Be sure to use C_S capacitors that have low self-inductance.

For the low-side filter (Fig 7B) remove the two strips that go to the shunt capacitor (C_S) location. Connect each coil ground lead directly to ground by running the coil wires through the pads and soldering them to the ground plane. Be sure to use the grounding screw from the center of the PC board to the metal mounting plate. I use #4-40 hex nuts and #4 flat washers as spacers in five locations. In addition, it is important to avoid stray coupling between the filters and adjacent metal surfaces and circuits that might degrade the stopband (verified).

For the simple style of construction shown in Figs 7 and 8, an ultimate attenuation of 70 dB from 1.8 through 30 MHz, is a reasonable expectation that is good enough for many applications. For more-stringent needs, a filter-amplifier-filter arrangement will provide enough ultimate attenuation for just about any application. I prefer this approach to more elaborate individual filters because the actual hardware's ultimate broadband attenuation is much better. The amplifier can be a low-gain (4-6 dB), unilateral, grounded-gate amplifier that has a 50-Ω dynamic (loss-less) input resistance, a physical 50-Ω output resistance and dynamic range suitable to the application. Calculate the cascaded noise figure and intercepts.

Another option is to sharpen the selectivity of the filter. This can be done by narrowing the passband (Eq 6) and calculating new component values (Eqs 7, 8, 9, 10, 11, 12 and 13). *Radio Designer* will show that the passband attenuation may in-crease by a decibel.

For low-level applications, such as medium-performance receivers, the filters can be switched with PIN diodes, although I much prefer the relays. The download file (see Note 1) shows an approach that I have used; it works quite well in the HF bands. The references in Notes 6 and 7 should be consulted for further information on this approach.

Notes

[1]You can download this package from the ARRL Web **http://www.arrl.org/files/qex/**. Look for NPBF.ZIP.

[2]H. J. Blinchikoff and A. I. Zverev, *Filtering in the Time an Frequency Domain* (Wiley and Sons, 1976) Chapter 4.

[3]A. I. Zverev, *Handbook of Filter Synthesis* (Wiley and Sons, 1967), pp 300-306.

[4]W. E. Sabin, W0IYH, "Designing Narrow Band-Pass Filters with a BASIC Program," *QST*, May 1983.

[5]FAR Circuits, 18N640 Field Ct, Dundee, IL 60118; tel 847-836-9148 (Voice mail), fax 847-836-9148 (same as voice mail); e-mail **farcir@ais.net**; URL **http://www.cl.ais.net/farcir/**.

[6]*The ARRL Handbook*, 1995-2000 editions, p 17.31. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Check out the full ARRL publications line at **http://www.arrl.org/catalog**.

[7]W. E. Sabin, W0IYH, "Mechanical Filters in HF Receiver Design," *QEX*, Mar 1996.