Chapter 12

RF and AF Filters

This chapter contains design information and examples of the most common filters used by radio amateurs. The initial sections describing basic concepts, lumped element filters and some design examples were prepared by Jim Tonne, WB6BLD. The following theory sections describing crystal, SAW, transmission-line and active filters were prepared by Reed Fisher, W1CQH, and others. Rounding out the chapter are a number of useful filter designs for immediate construction. References are given for further reading and design information.

Basic Concepts

Electrical filters are used to pass signals of certain frequencies and reject others. The electronics industry has advanced to its current level in large part because of the successful use of filters. Filters are used in receivers so that the listener can hear only the desired signal; other signals are rejected. Filters are used in transmitters to pass only one signal and reject those that might interfere with other spectrum users. The simplified SSB receiver shown in **Fig 12.1** illustrates the use of several common types of filters.

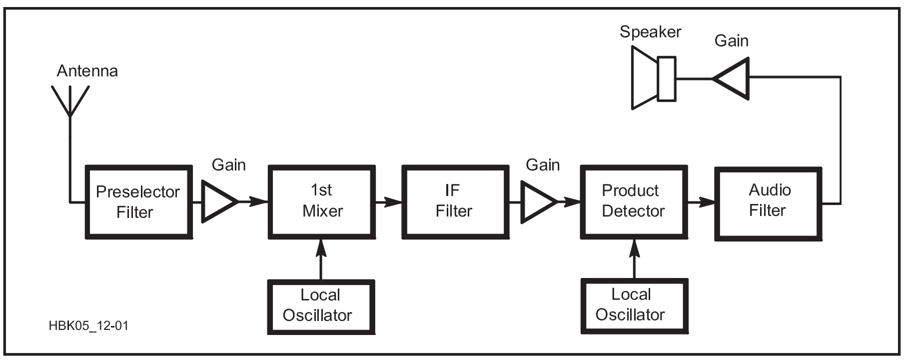


Fig 12.1 — One-band SSB receiver. At least three filters are used between the antenna and speaker.

A filter is placed between the antenna and the first mixer. This filter is usually called a "preselector" filter. It passes all frequencies within a given amateur band with low loss. Out-of-band signals are rejected to prevent them from overloading the first mixer (a common problem with both medium wave and shortwave broadcast stations). The preselector filter is almost always built with lumped-element or "LC" technology.

An intermediate frequency (IF) filter is placed between the first and second mixers. It is a band-pass filter that passes only the desired SSB signal. The age of the receiver probably determines which of several filter technologies is used. As an example, 50-kHz LC filters and 455-kHz crystal or mechanical filters were used through the 1960s. Later model receivers usually use quartz crystal

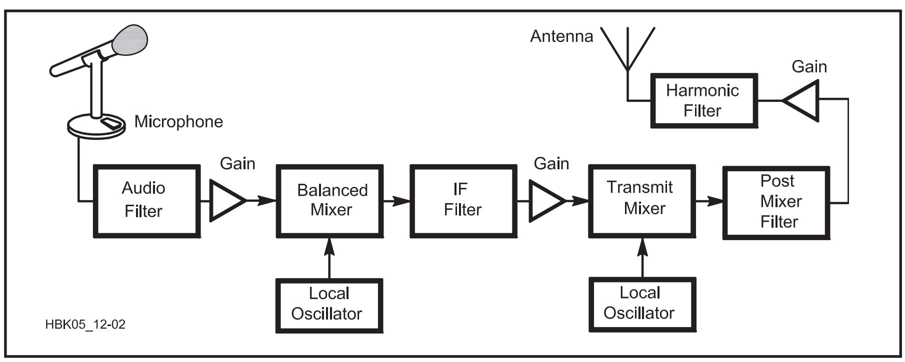


Fig 12.2 — One-band SSB transmitter. At least three filters are needed to ensure a clean transmitted signal.

Table 12.1
Typical Filter Bandwidths for Typical Signals

Source	*Required Bandwidth*
High-fidelity speech and music	20 Hz to 15 kHz
Telephone-quality speech	200 Hz to 3 kHz
Radiotelegraphy (Morse code, CW)	200 Hz
HF RTTY	1000 Hz (varies with frequency shift)
NTSC television	60 Hz to 4.5 MHz
SSTV	200 Hz to 3 kHz
1200 bit/s packet	200 Hz to 3 kHz

filters with center frequencies typically between 3 and 11 MHz. In all cases, the filter bandwidth must be less than 3 kHz to effectively reject adjacent SSB stations.

Finally, a 300-Hz to 3-kHz audio band-pass filter is placed somewhere between the detector and the speaker. It rejects unwanted products of detection, power supply hum and noise. The audio filter is often implemented with active filter technology.

The complementary SSB transmitter block diagram is shown in **Fig 12.2**. A similar array of filters appears in reverse order. First is a 300-Hz to 3-kHz audio filter, which rejects out-of-band audio signals such as 60-Hz power supply hum. It is placed between the microphone and the balanced mixer.

The IF filter is next. Since the balanced mixer generates both lower and upper sidebands, the IF filter is placed at the mixer output to pass only the desired lower (or upper) sideband. In commercial SSB transceivers this filter is usually the same as the IF filter used in the receive mode.

Finally, a band-pass filter covering the amateur band in use is placed at the output of the transmit mixer to reject unwanted frequencies generated by the mixer and prevent them from being amplified and transmitted. Low-pass filters are usually placed at the transmitter output to pass the transmitted signal and attenuate its harmonics.

Table 12.1 shows the usual signal bandwidths for several signal types.

FILTER MAGNITUDE RESPONSES

A filter passes signals in a range of frequencies — the *passband* — while rejecting signals outside that range — the *stop band*. To pass signals from dc up to some cutoff frequency we would use a *low-pass* filter. To pass signals above a cutoff we would use a *high-pass* filter. Similarly, to pass signals within a range of frequencies we would use a *band-pass* filter. To pass signals at all frequencies except those within a specified range dictates the use of a *band-stop* filter. The region between the passband and the stop band is logically called the *transition region*.

Fig 12.3 illustrates the magnitude response of a low-pass filter. Signals lower than the cutoff frequency (3 MHz in this case) are passed with some small amount of attenuation while signals higher than that frequency are attenuated. The degree of attenuation is dependent on several variables, the filter complexity being a major factor.

Fig 12.4 illustrates the magnitude response of a high-pass filter. Signals above the cutoff frequency are passed with minimum attenuation while signals below that frequency are attenuated. Again, the degree of attenuation is dependent on several variables.

Fig 12.5 illustrates the magnitude response of a band-pass filter. Signals within the band-pass range (between a lower band edge and an upper band edge) are passed with minimum attenuation while signals outside that range are attenuated. In this example the filter was designed to be 3 dB down at 2 MHz and at 4 MHz.

Fig 12.6 illustrates the magnitude response of a band-stop filter. Signals within the band-stop range are attenuated while all other signals are passed with minimum attenuation. A notch filter is an example of a band-stop design.

FILTER ORDER

The steepness of the descent from the passband to the area of attenuation — the stop band — is dependent on the complexity of the filter, properly called the *order*.

Fig 12.7 shows the magnitude response of a set of low-pass filters with orders varying from very simple ("N=2") to the more complex ("N=10"). In each case on this plot the 3-dB point remained the same, at 3 Hz.

As we go from the passband into the stop band, the attenuation increases. The rate of attenuation at frequencies well outside the cutoff, expressed in terms of dB per octave, will be 6.02 times the order. As the order goes up the rate also goes up.

FILTER FAMILIES

There is no single "best" way to develop the parts values for a filter. Instead we have to decide on some traits and then choose the most appropriate *family* for our design. Different families have different traits, and filter families are commonly named after the mathematician or other developer responsible for them.

A major trait is the sharpness of the corner (cutoff), or performance in terms of how close to an ideal magnitude response we need. Ideally a filter would pass all frequencies up to some point with no attenuation and totally

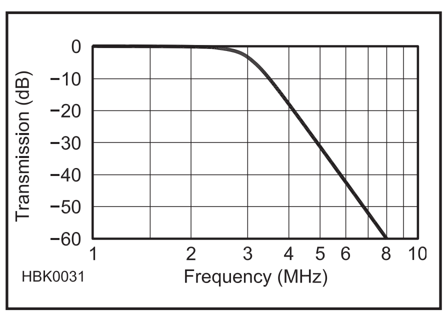


Fig 12.3 — Example of a low-pass response plot.

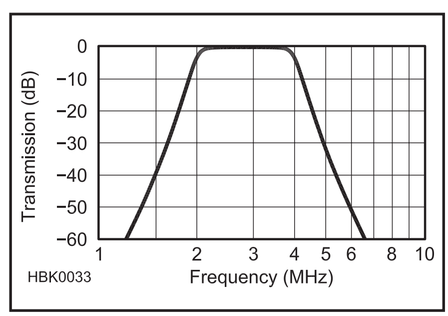


Fig 12.5 — Example of a band-pass response plot.

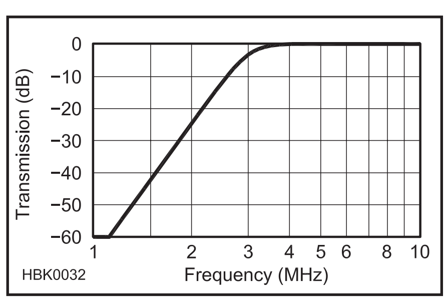


Fig 12.4 — Example of a high-pass response plot.

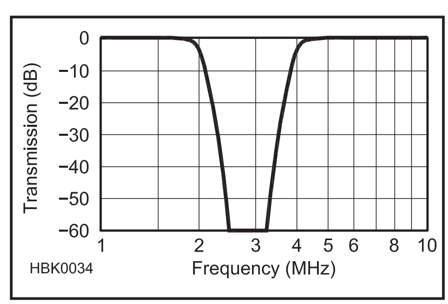


Fig 12.6 — Example of a band-stop response plot.

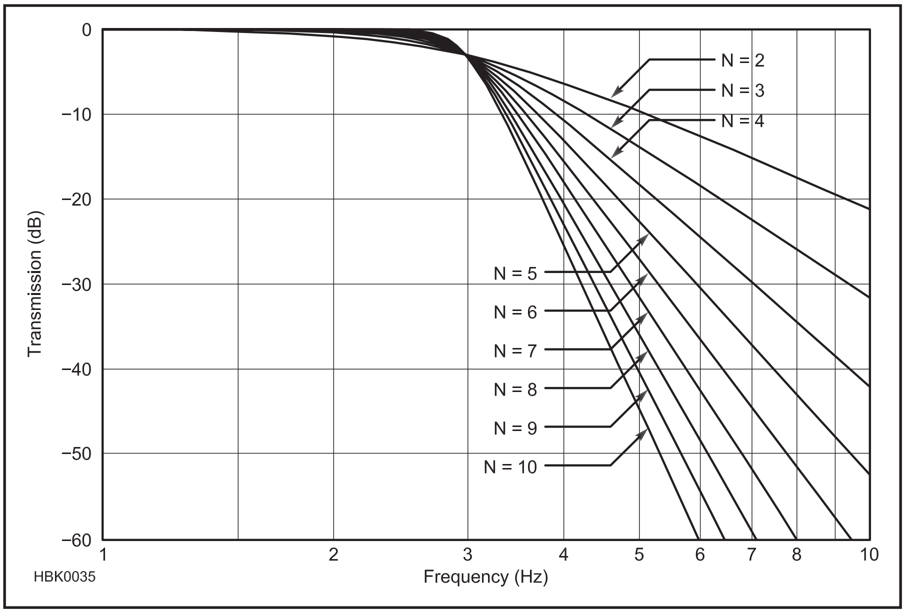


Fig 12.7 — Magnitude responses for low-pass filters with orders ranging from very simple (N=2) to very complex (N=10).

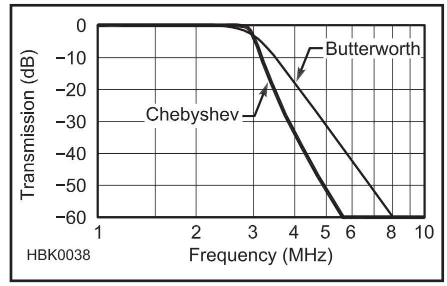


Fig 12.10 — A Chebyshev filter (0.2 dB passband ripple) allows a sharper cutoff than a Butterworth design with no passband ripple.

reject everything beyond that point. With a flat magnitude response (and so no ripples in the passband) we have what is known as a *Butterworth* family design. This family also goes by the name of Maximally Flat Magnitude or Maximally Flat Gain. An example of the magnitude response of a filter from the Butterworth family is shown in **Fig 12.8**.

By allowing magnitude ripples in the passband, we can get a somewhat steeper descent from the passband into the stop band particularly just outside the cutoff frequency. A family that does this is the *Chebyshev* family. **Fig 12.9** illustrates how allowing 1 dB of magnitude ripple in the passband provides a sharper filter cutoff. That plot compares the 1-dB Chebyshev with the no-ripple Butterworth filter, down to 12 dB of attenuation.

Fig 12.9 also illustrates the usual definition of the bandwidth of a Butterworth filter (the 3-dB point) and the bandwidth of a Chebyshev filter (the ripple bandwidth). The Butterworth filter is defined by specifying the order and the 3-dB (half-power) bandwidth. The Chebyshev filter is commonly defined by specifying the order, the *ripple* bandwidth and also the amount of passband ripple. In Fig 12.9, the Chebyshev filter has 1 dB of ripple; its ripple bandwidth is 1000 Hz. The Butterworth filter has a 3-dB bandwidth also of 1000 Hz. Some filter textbooks use the 3-dB point to define Chebyshev filters; most use the ripple bandwidth as illustrated here. The schematics (if you ignore parts values) of those two families are identical.

Even small amounts of ripple can be beneficial in terms of cutoff sharpness. **Fig 12.10** compares a Butterworth filter (with the narrow line plot, no ripple in the passband) with a Chebyshev filter (wide line plot, 0.2 dB of ripple in the passband) down to 60 dB of attenuation. Even that small amount of ripple in the Chebyshev filter passband allows a

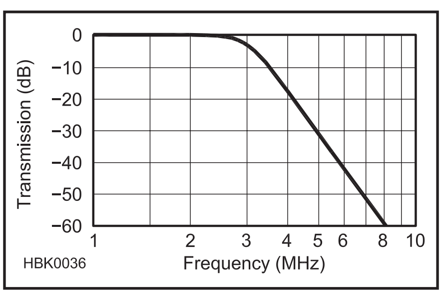


Fig 12.8 — Filters from the Butterworth family exhibit flat magnitude response in the passband.

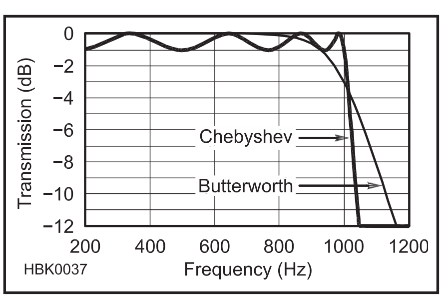


Fig 12.9 — This plot compares the response of a Chebyshev filter with a 1-dB ripple bandwidth of 1000 Hz and a Butterworth filter 3-dB bandwidth of 1000 Hz.

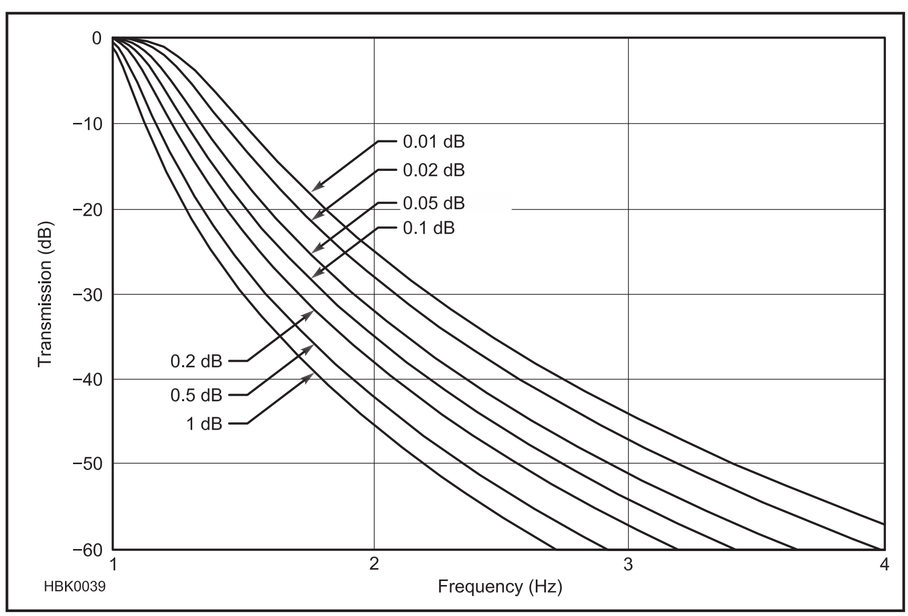


Fig 12.11 — These response plots illustrate the Chebyshev family with various values of passband ripple. These plots are for a seventh-order low-pass design with ripple values from 0.01 to 1 dB. Ignoring the effects in the passband of high ripple values, increasing the ripple will allow somewhat steeper descent into the stop band area, and better ultimate attenuation in the stop band.

noticeably improved rate of descent from the passband into the stop band. This descent is steeper immediately above the band edge when compared with the Butterworth design. As the passband ripple specification is increased, the sharpness of the corner improves. The ultimate rate of descent is the same as for the Butterworth family. For this comparison the frequencies with 3 dB of attenuation are the same for each filter.

For Chebyshev filters, when the value of the passband ripple is changed, the magnitude response in the stop band region also changes. **Fig 12.11** compares the response of Chebyshev filters with passband ripple ranging from 0.01 to 1 dB.

It is possible to get an even steeper descent into the stop band by adding to the design "traps" whose frequencies are carefully calculated. The resonant frequencies of those traps are in the stop band region and are set to yield best performance. When this is properly done, we have a *Cauer* family design (also called the elliptic-function design). The Cauer filter has a descent rate from the passband into the stop band that is the steepest of all analog filter types provided that the behavior in the passband is uniform (either no ripple or uniform ripple). **Fig 12.12** shows the response of the Chebyshev and Cauer designs for comparison.

The Cauer filter is defined by specifying the order, the ripple bandwidth, and the passband ripple, just as for the Chebyshev. And again an alternative bandwidth definition is to use the 3-dB point instead of the ripple bandwidth. The Cauer family requires one more specification: the *stop band frequency* and/or the *stop band depth*. In Fig 12.12, the stop band frequency is about 3.47 MHz and the stop band depth is 50 dB. For this comparison both designs have the same passband ripple value of 0.2 dB.

A downside to the Cauer filter is that the attenuation in the stop band is some chosen minimum rather than ever increasing, as is the case with the other families. In the far stop band region the Cauer ultimately rolls off at about 6 or 12 dB per octave, depending on the order. Odd-ordered Cauer filters have an ultimate rolloff rate of about 6 dB per octave while the even-ordered versions have an ultimate rolloff rate of about 12 dB per octave.

TIME DELAY

Another trait that sometimes influences the choice of which filter family to use is the way the signal delay in the passband of the filter varies with frequency. The wide line (lower plot) in **Fig 12.13** illustrates the time delay characteristics of a Chebyshev low-pass filter, while the upper plot (narrow line) shows the

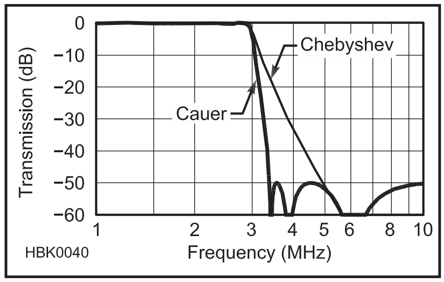


Fig 12.12 — The Cauer family has an even steeper descent rate from the passband into the stop band than the Chebyshev family. Note that the ultimate attenuation in the stop band is a design parameter rather than ever-increasing as is the case with other families.

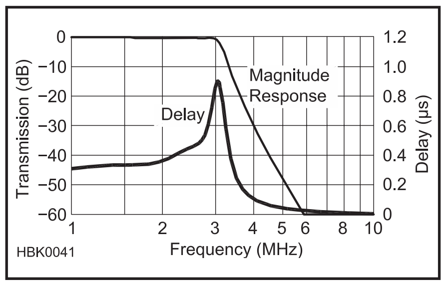


Fig 12.13 — Magnitude response and time delay of a Chebyshev low-pass filter.

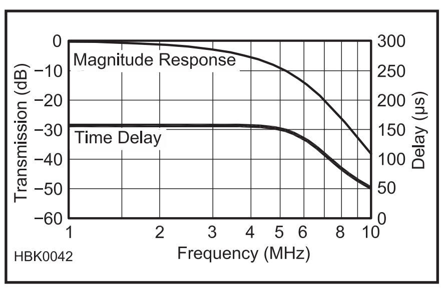


Fig 12.14 — Magnitude response and time delay of a Bessel low-pass filter.

magnitude response. As shown in Fig 12.13, the time delay of components near the cutoff frequency becomes quite large when compared to the delay of components at lower test frequencies. This is a result of the phase shift of the filter's transmission being nonlinear with frequency; it is usually greater near the cutoff or band edge.

If a uniform time delay for signals throughout the passband is needed, then the Bessel filter family should be selected. The Bessel filter can be used as a delay-line or time-delay element although the gentle magnitude rolloff in the passband may need to be taken into account. The magnitude and delay charac-

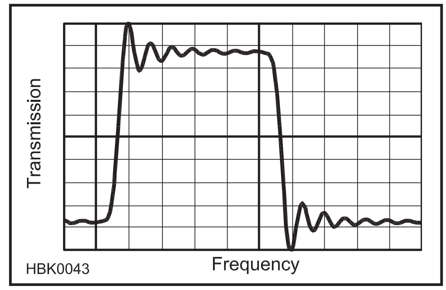


Fig 12.15 — Transient response of a Chebyshev low-pass filter.

teristics for the Bessel family are shown in **Fig 12.14**.

The constant-delay characteristic of the Bessel (the bottom plot in Fig 12.14) extends into the region where attenuation is occurring (into the stop band). A downside of the Bessel family is that the sharpness of the corner of that filter is quite poor; it is not very good as a magnitude-response-shaping filter. The Bessel family is characterized largely by its constant time delay in the passband (for the low-pass topology). The Bessel filter bandwidth is commonly defined by its 3-dB point (as with the Butterworth) or sometimes by the one-radian lag frequency.

TRANSIENT RESPONSE

Some applications require that a square wave signal applied to the input of a low-pass design have a minimum overshoot as seen at the output. If a sharp-cutoff filter is used in such an application, an overshoot will be added to the square wave. If we have a sharp cutoff low-pass filter of a given bandwidth and apply a square wave of one tenth that frequency to its input, the appearance of the signal as it exits from the filter may be as shown in **Fig 12.15**. The scales for both the X- and the Y-axes would depend on the frequency and magnitude of the waveform.

A square wave (used in this discussion as a test waveform) is composed of a fundamental and an endless series of odd harmonics. If harmonics only up to a certain number are used to create the square wave — that is, if the square waveform is passed through a sharp-cutoff low-pass filter — then that waveform as it exits from the filter will have the overshoot or ringing as shown.

All of the traits mentioned so far in this chapter apply to a filter regardless of how it is implemented, whether it is fabricated using inductors and capacitors ("*lumped element*") or op amps with resistors and capacitors ("*active*"). In general they are also true if the filter is implemented using other technologies.

Lumped-Element Filters

LOW-PASS FILTERS

This part of the chapter deals with filters fabricated using discrete inductors and capacitors, (which gives rise to their name, *lumped element*). The basic idea behind filters built using inductors and capacitors is shown in **Fig 12.16**.

In the first case (Fig 12.16A), less power is delivered to the load at higher frequencies because the reactance of the inductor in series with the load increases as the test frequency increases. The voltage appearing at the load goes down as the frequency increases. This configuration would pass direct current (dc) and reject higher frequencies, and so it would be a *low-pass* filter.

With the second case (Fig 12.16B), less power is delivered to the load at higher frequencies because the reactance of the capacitor in parallel with the load decreases as the test frequency increases. Again, the voltage appearing at the load goes down as the frequency increases and so this, too, would be called a low-pass filter. In the real world, combinations of both series and parallel components are used to form a low-pass filter.

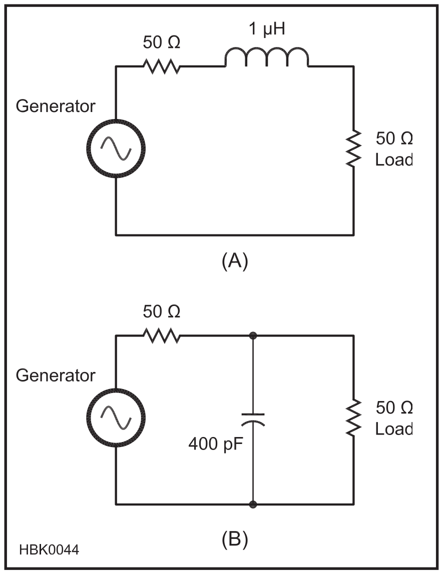


Fig 12.16 — An elementary low-pass filter can be formed using a series inductor (A) or a shunt capacitor (B).

A *high-pass* filter can be made using the opposite configuration — series capacitors and shunt inductors. And a *band-pass* (or *band-stop*) filter can be made using series and parallel tuned circuit pairs. When multiple components are used they must be chosen carefully and that will now be discussed.

We spoke of filter order, or complexity, earlier in this chapter. **Fig 12.17** illustrates *capacitor-input* low-pass filters with orders of 3, 4 and 5. For example, the third-order filter in Fig 12.17A has three elements (two capacitors and one inductor), while the fifth-order design in Fig 12.17C has five elements total (three capacitors, two inductors). For comparison, a third-order *inductor-input* filter is illustrated in **Fig 12.18**.

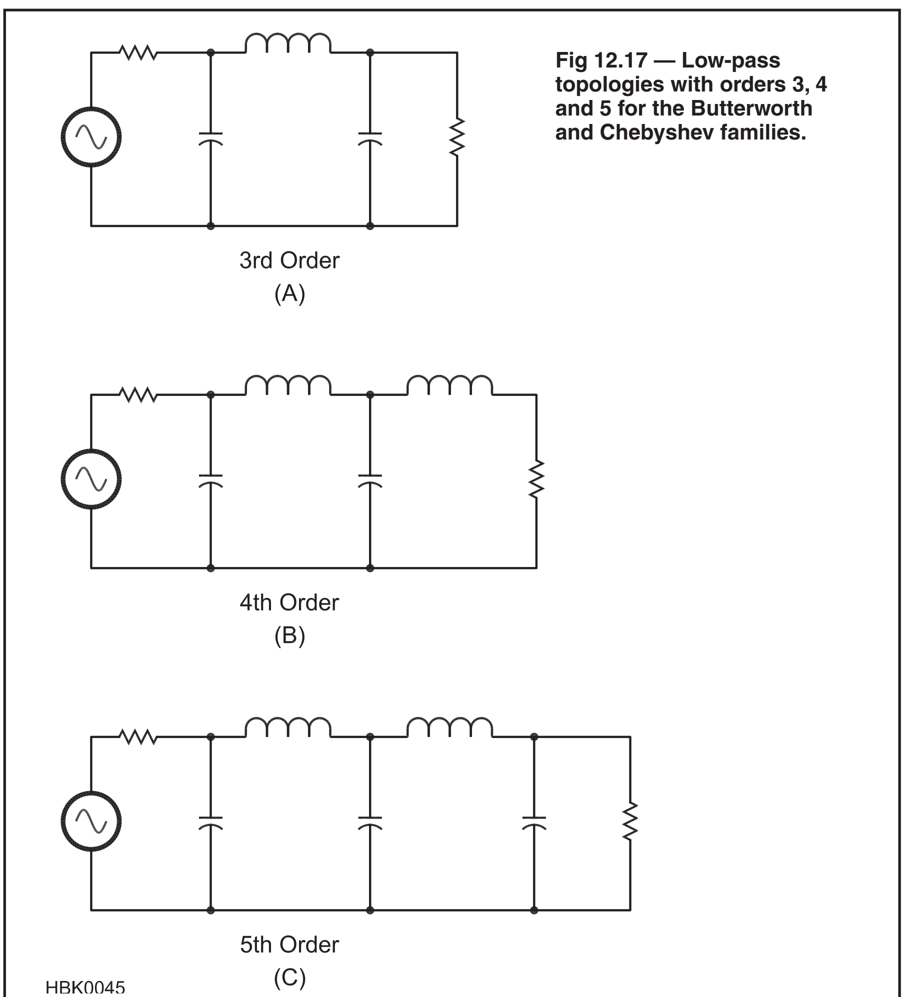


Fig 12.17 — Low-pass topologies with orders 3, 4 and 5 for the Butterworth and Chebyshev families.

As mentioned previously, the Cauer family has parts carefully added to produce dips or notches (properly called "zeros") in the stop band. Schematics for the Cauer versions of capacitor-input low-pass filters with orders of 3, 4 and 5 are shown in **Fig 12.19**. The capacitors in parallel with the series inductors create the notches at calculated frequencies to allow the Cauer filter to be implemented.

The capacitor-input and the inductor-input versions of a given low-pass design have identical characteristics for their magnitude, phase and time responses, but they differ in the impedance seen looking into the filter. The capacitor-input filter has a low impedance in the stop band while the inductor-input filter has a high impedance in the stop band.

HIGH-PASS FILTERS

A high-pass filter passes signals above its cutoff frequency and attenuates those below. The basic idea behind the high-pass filter is shown in **Fig 12.20.**

The reactance of the series capacitor in Fig 12.20A increases as the test frequency is lowered and so at lower frequencies there will be less power delivered to the load. Similarly, the reactance of the shunt inductor in Fig 12.20B decreases at lower frequency, with

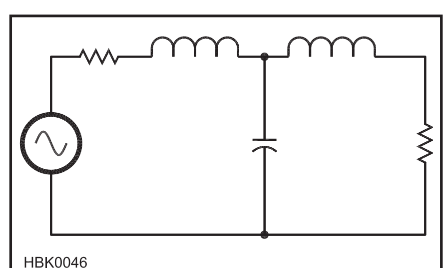


Fig 12.18 — A third-order inductor-input low-pass filter.

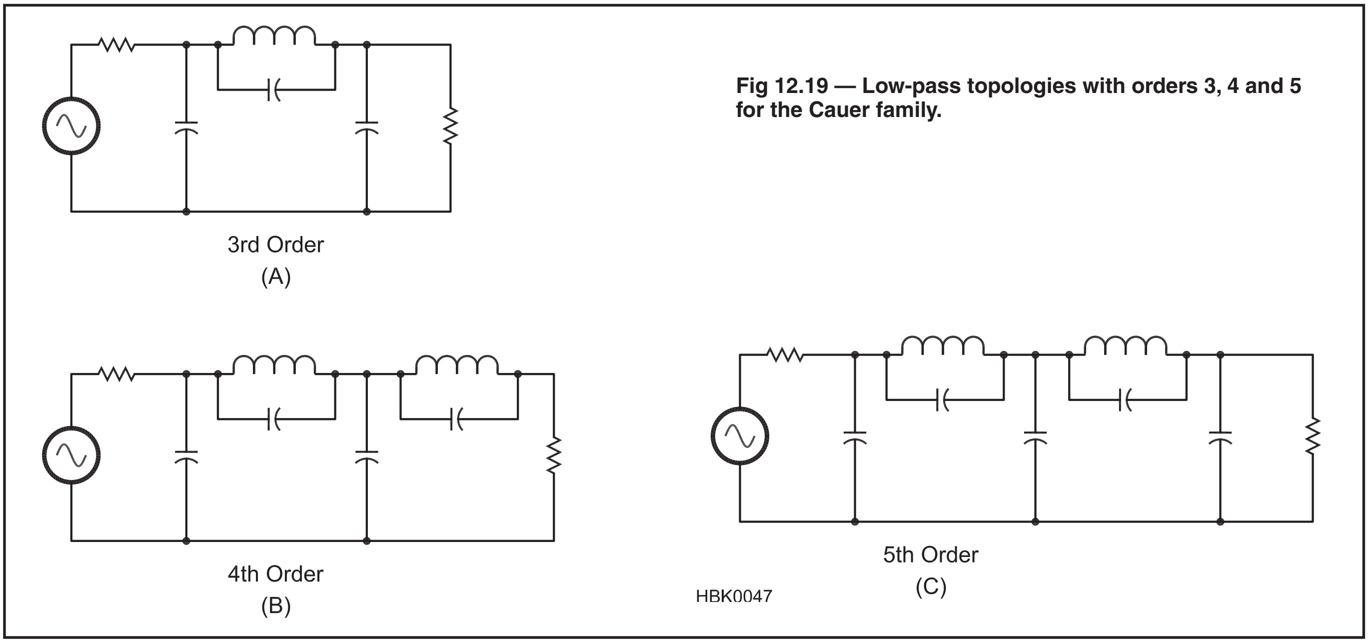


Fig 12.19 — Low-pass topologies with orders 3, 4 and 5 for the Cauer family.

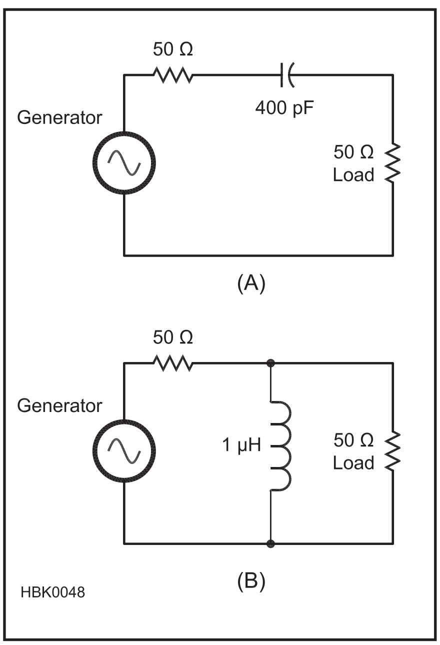


Fig 12.20 — An elementary high-pass filter can be formed using a series capacitor (A) or a shunt inductor (B).

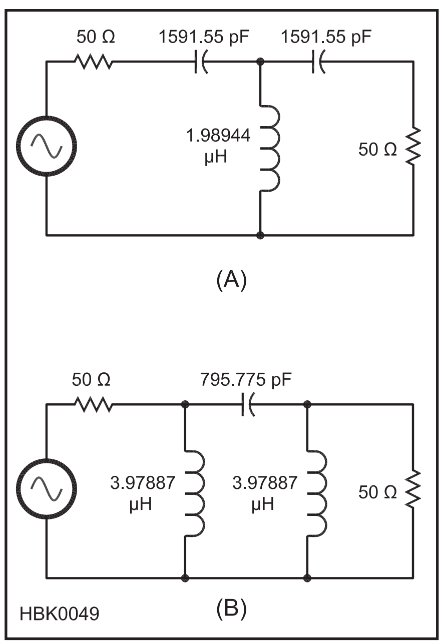


Fig 12.21 — Capacitor-input (A) and inductor-input (B) high-pass filters. Both designs have a 2 MHz cutoff.

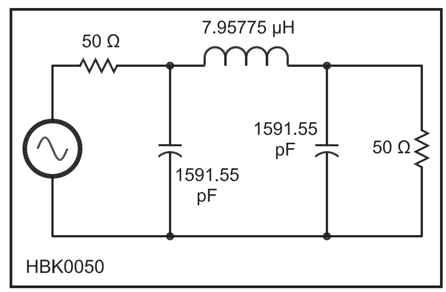


Fig 12.22 — Third-order low-pass filter with a bandwidth of 2 MHz in a 50 Ω system.

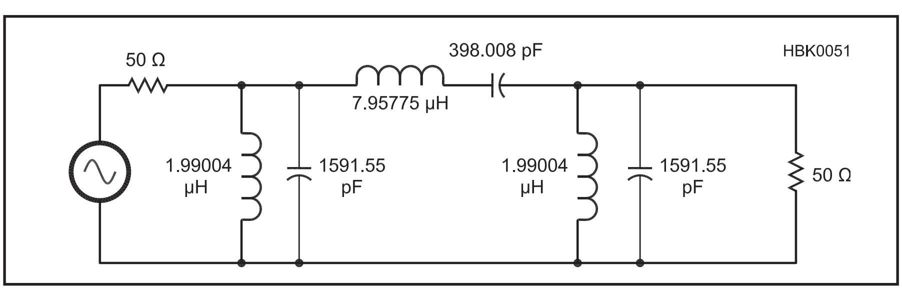


Fig 12.23 — The low-pass filter of Fig 12.22 can be transformed to a bandpass filter by resonating the shunt capacitors with a parallel inductor and resonating the series inductor with a series capacitor. Width is 2 MHz and the center is at 2.828 MHz.

the same effect. And as with the other filter types, a high-pass filter in a real world design would typically use both series and shunt components.

An example of a high-pass filter application would be one designed to pass amateur-band signals in the range of 3.5 MHz and above while rejecting broadcast signals in the range of 1.7 MHz and below. A high-pass filter with a design cutoff of 2 MHz is illustrated in **Fig 12.21**. It can be implemented as a capacitor-input (Fig 12.21A) or inductor-input (Fig 12.21B) design. In each case, signals above the cutoff are passed with minimum attenuation while signals below the cutoff are attenuated, in a manner similar to the action of a low-pass filter.

LOW-PASS TO BAND-PASS TRANSFORMATION

A band-pass design can be defined in part by specifying a *bandwidth* and a *center frequency*. (An alternative method is to specify a lower band edge and an upper band edge.) A low-pass design such as the one shown in

Fig 12.17A can be converted to a band-pass filter by resonating each of the elements to the center frequency. **Fig 12.22** shows a third-order low-pass filter with a design bandwidth of 2 MHz for use in a 50-Ω system. It should be mentioned that the next several designs and the various manipulations were done using a computer; other design methods will be shown later in the chapter.

If the shunt elements are now resonated with a parallel component, and if the series elements are resonated with a series component, the result is a band-pass filter as shown in **Fig 12.23**. The series inductor value and the shunt capacitor values are the same as those for the original low-pass design. Those components have been resonated at the center frequency for the filter (2.828 MHz in this case).

Now we can compare the magnitude response of the original low-pass filter with the band-pass design; the two responses are shown in **Fig 12.24**. The magnitude response of the low-pass version of this filter is down 3 dB at 2 MHz. The band-pass response is down 3 dB at 2 MHz and 4 MHz (the difference from the high side to the low side is 2 MHz). The response of the low-pass design is down 33 dB at 7 MHz while the response of the band-pass version is down 33 dB at 1 and 8 MHz (the difference from high side to low side is 7 MHz).

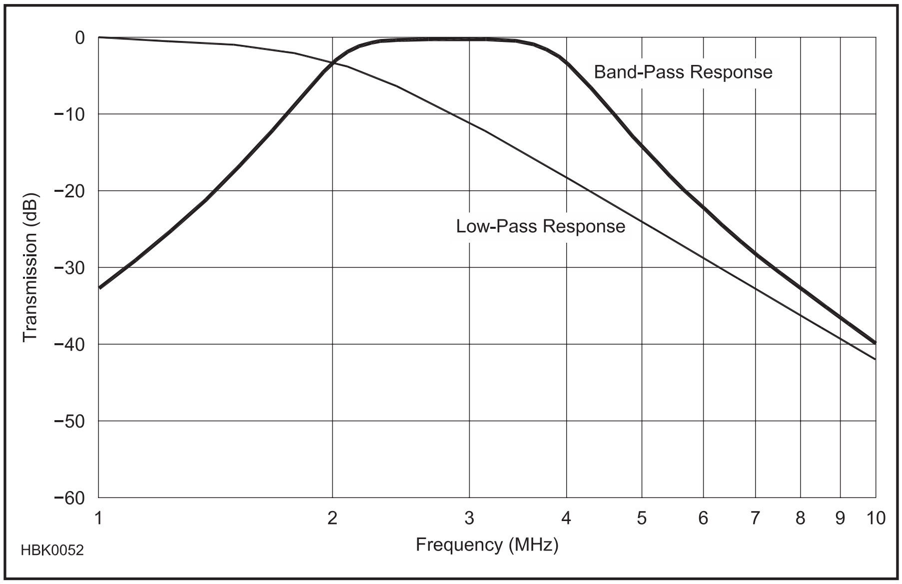


Fig 12.24 — Comparison of the response of the low-pass filter of Fig 12.22 with the band-pass design of Fig 12.23.

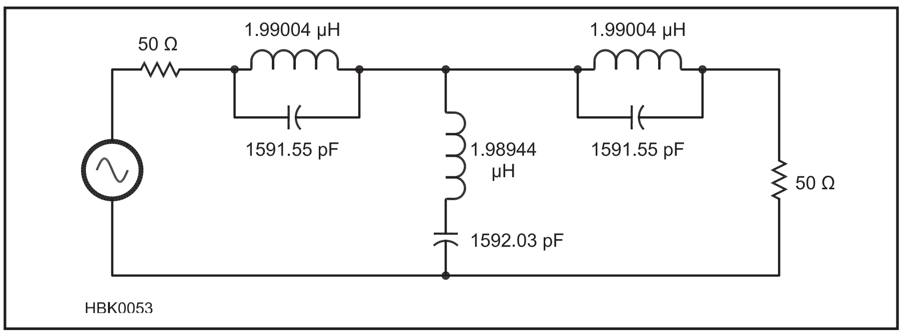


Fig 12.25 — The high-pass filter of Fig 12.21A can be converted to a band stop filter with a width of 2 MHz, centered at 2.828 MHz.

HIGH-PASS TO BAND-STOP TRANSFORMATION

Just as a low-pass filter can be transformed to a band-pass type, a high-pass filter can be transformed into a *band-stop* (also called a *band-reject*) filter. The procedures for doing this are similar in nature to those of the transformation from low-pass to band-pass. As with the band-pass filter example, to transform a high-pass to a band-stop we need to specify a center frequency.

Fig 12.25 shows how to convert the 2 MHz capacitor-input high-pass filter of Fig 12.21A to a band-stop filter centered at 2.828 MHz. The original high-pass components are resonated at the chosen center frequency to form a band-stop filter. Either a capacitor-input high-pass or an inductor-input high-pass may be transformed in this manner. In this case the series capacitor values and the shunt inductor values for the band-stop are the same as those for the high-pass. The series elements are resonated with an element in parallel with them. Similarly, the shunt elements are also resonated with an element in series with them. In each case the pair resonate at the center frequency of the band-stop. **Fig 12.26** shows the responses of both the original high-pass and the resulting band-stop filter.

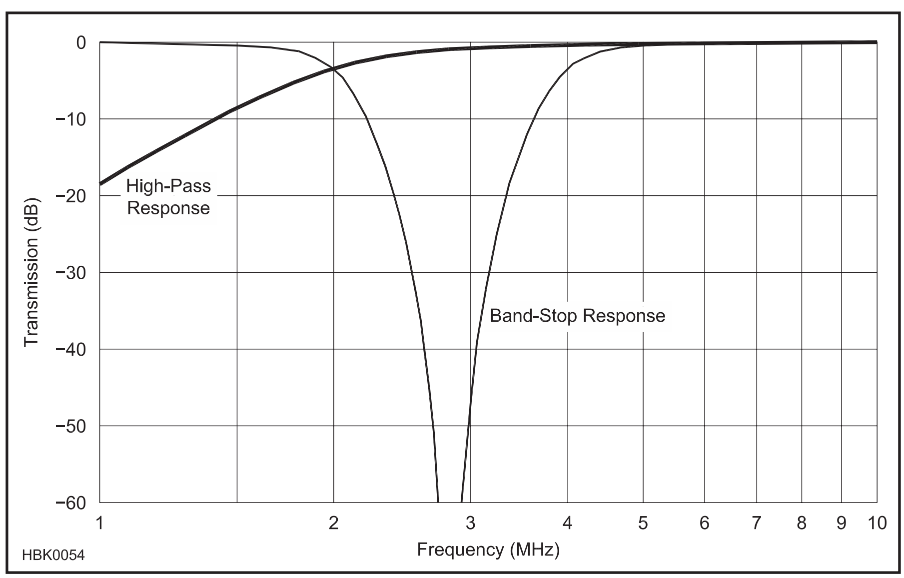


Fig 12.26 — Comparison of the response of the high-pass filter of Fig 12.21A with the band-stop design of Fig 12.25.

REFINEMENTS IN BAND-PASS DESIGN

The method of transforming a low-pass

filter to a band-pass filter described previously is simple, but it has a major drawback. The resulting component values are often awkward for building a narrowband band-pass filter. This situation can be avoided by using various manipulations as shown next. They are tedious and are best done using a computer. Some manual design methods are touched on later in this chapter.

Fig 12.27 shows a band-pass filter centered at 3 MHz with a width of 100 kHz that was designed using the method previously described. This filter is impractical for several reasons. The shunt capacitors (0.0318 µF) will probably have poor characteristics at the center frequency of 3 MHz because of their series inductance. Similarly, the series inductor (159 µH) will certainly have a parallel resonance because of distributed capacitance, and that resonance is quite likely to alter the filter's response. For a narrowband band-pass filter, this method of transformation from low-pass to band-pass is not practical.

A much better approach to the narrowband band-pass filter design task is to change the topology to a *nodal-capacitor-coupled* design. **Fig 12.28** shows how the filter in Fig 12.27 can be redesigned using such an approach.

Some component values are improved, but the shunt capacitor values are still quite large for the center frequency of 3 MHz. We can scale the impedance of the filter upward to reduce the values of these capacitors and so minimize this problem. If we make the filter for a 500 Ω system instead of a 50 Ω system, the reactances of all parts will be multiplied by a factor of 10. The 0.03 µF capacitors will decrease in value to only 3000 pF, a much better value for a center frequency of 3 MHz. The shunt inductor values and the nodal coupling capacitor values (about 75 pF in this case) are also realistic. The 500 Ω version of the filter in Fig 12.28 is shown in **Fig 12.29**.

From a component-value viewpoint this is a better design, but it must be terminated in 500 Ω at each end. To use this filter in a 50 Ω system, impedance matching must be added as shown in **Fig 12.30**. This is the kind of topology used in a radio receiver in its preselector in the front end, or in the IF system, or anywhere else that a narrow — percentage-wise — filter is needed.

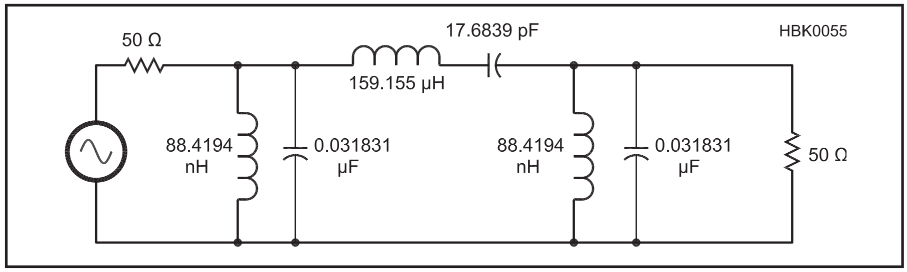


Fig 12.27 — This band-pass filter, centered at 3 MHz with a width of 100 kHz, was designed using a simple transformation. The resulting component values make the design impractical, as discussed in the text.

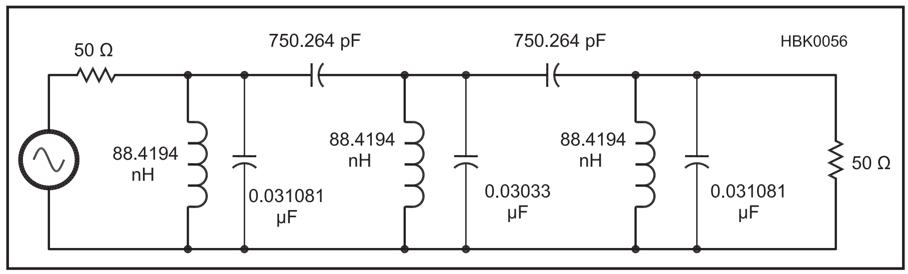


Fig 12.28 — A narrowband band-pass filter using a nodal-capacitor-coupled design improves component values somewhat.

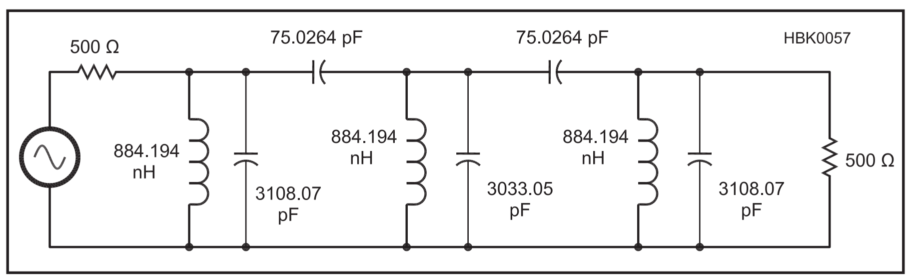


Fig 12.29 — A 500 Ω version of the filter in Fig 12.28.

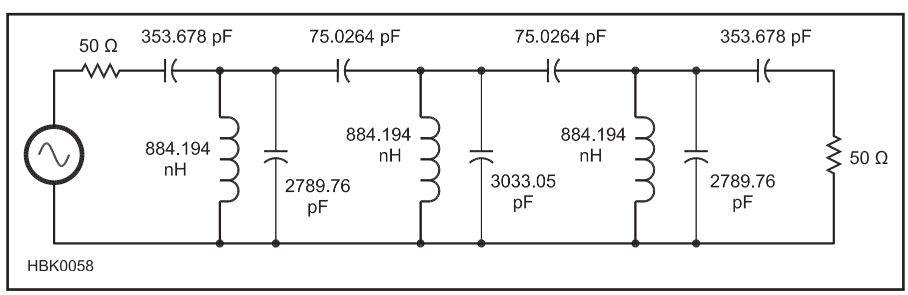


Fig 12.30 — Final design of the narrowband band-pass filter with 50 Ω terminations.

When the nodal capacitor-coupled band-pass topology is used for a filter whose bandwidth is wide, the attenuation on the high-frequency side of the response curve will be less than on the low-frequency side. This characteristic should be taken into account when attenuation of harmonics of signals in the passband is of concern. The filter shown in **Fig 12.31A** is designed to pass the amateur 75-meter band and suffers from this defect. By going to the nodal inductor-coupled topology we can correct this problem. The resulting design is shown in Fig 12.31B. Another way to accomplish the task is by going to a mesh capacitor-coupled design as shown in Fig 12.31C. The last two designs have identical magnitude responses.

The design in Fig 12.31B has only three capacitors (a minimum-capacitor design) while the design in Fig 12.31C has only three inductors (a minimum-inductor design).

The magnitude response of the capacitor-coupled design in Fig 12.31A is shown in **Fig 12.32** as the solid line. The designs in Fig 12.31B and C have magnitude responses as shown in Fig 12.32 as a dashed line. Note that the descent into the stop band on the high-frequency side is at a greater rate for the designs in Fig 12.31B and C. This will be the case for those filters with relatively wide (percentage-wise) designs. Such a design would be useful where harmonics of a signal are to be especially attenuated. It might

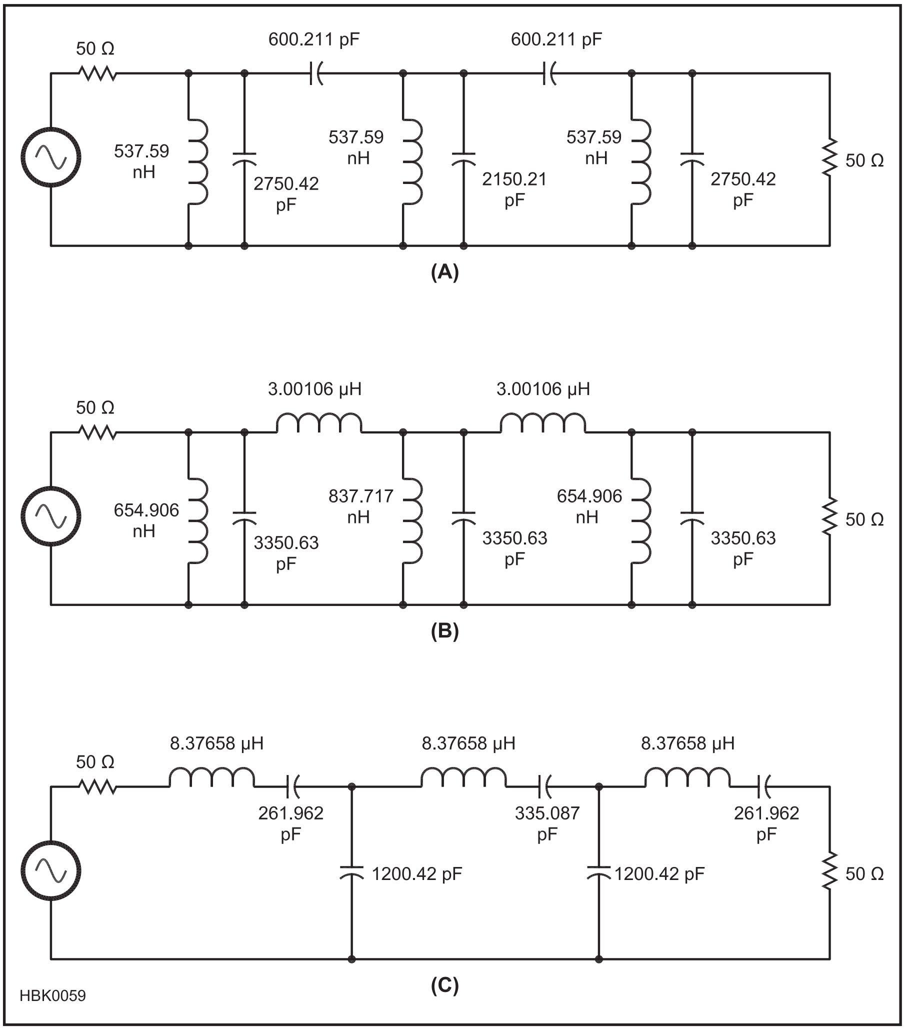


Fig 12.31 — Three band-pass filters designed to pass the 75-meter amateur band. The nodal-capacitor-coupled topology is shown at A, nodal-inductor-coupled at B, and mesh capacitor-coupled at C.

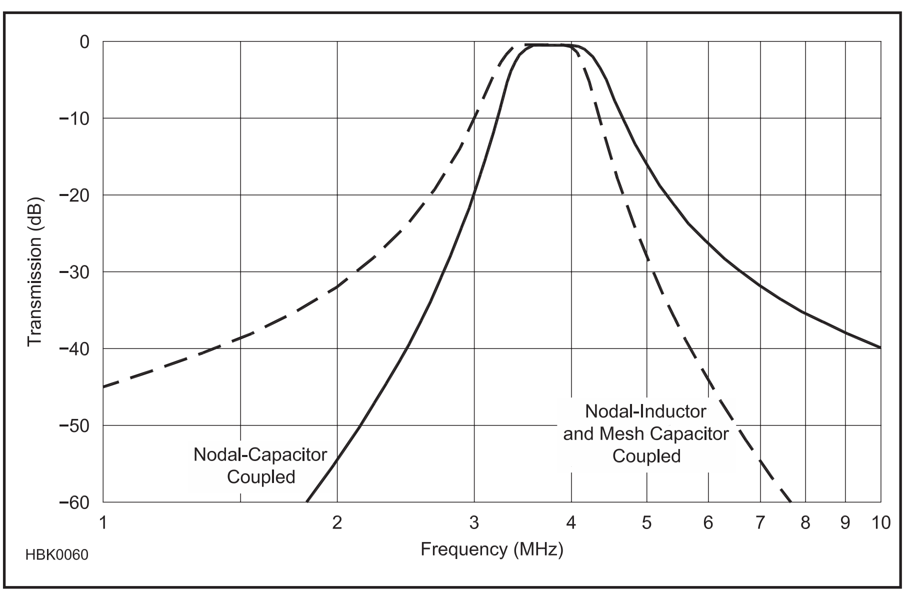


Fig 12.32 — Responses of the filters in Fig 12.31.

also improve image rejection in a receiver RF stage where high-side injection of the local oscillator is used.

When a filter is designed "by the book" as shown here, the component values will invariably be uncommon. The nearest 5% values can sometimes be substituted, but it is always wise to analyze the resulting response using a computer. The better design programs offer this as an option. The program *SVC Filter Designer* (SVC is *standard value component*) on the CD included with this book expedites the design of low-pass (and high-pass) filters and automatically selects the nearest 5% capacitor values. It selects the nearest 5% inductor values as an option. It also shows the resulting response degradations (which may be minor).

Many of the filter design and analysis programs available show the responses of the network as designed and also allow the selection of nearest-5% values, tuning and other useful features. Those free or low-cost filter design programs including both design and analysis capabilities are listed in the references at the end of this chapter.

EFFECT OF COMPONENT Q

When components with less-than-ideal characteristics are used to fabricate a filter, the performance will also be less than ideal. One such item to be concerned about is component "Q." This is a measure of the loss of a part, as determined by its resistive component. Q is the ratio of reactance to the resistance internal to a component at a given test frequency. The resistance referred to here includes not only the value as measured by an ohmmeter but would also include an equivalent resistance attributable to losses in any core material used.

The Q values for capacitors are usually greater than 500 and may reach a few thousand. Q values for inductors seldom reach 500 and may be as low as 20 or even worse for miniaturized parts. A good toroidal inductor can have a Q value in the vicinity of 250 to 400.

Q values can affect both the loss of the filter when it is inserted into the circuit and the sharpness of the cutoff. Band-pass filters (and especially narrowband band-pass filters) are more vulnerable to this problem than low-pass and high-pass filters. **Fig 12.33** shows the effect of finite values of inductor Q values on the response of a low-pass filter. The Q values for each plot are as shown.

Inadequate component Q values introduce loss and more importantly they introduce rounded corners at cutoff, especially problematic in the case of narrowband band-pass filters. **Fig 12.34** illustrates the effect of finite inductor Q values on a narrowband band-pass filter. In the case of a band-pass filter, the Q values required to support a given shape

Try *Elsie* for "LC"!

The *ARRL Handbook CD* at the back of this book includes *Elsie*, a *Windows* program for design and analysis of lumped element filters. In addition to providing parts values for filters with various topologies from various families, tools are included to assist with practical construction. *Elsie* is provided courtesy of Jim Tonne, WB6BLD, who wrote the introductory section of this chapter. See the "About the Included CD-ROM" section for more information.

Table 12.2
Passband Ripple, VSWR and Return Loss

Passband Ripple (dB)	*VSWR*	*Return Loss (dB)*
0.0005	1.022	39.38
0.001	1.031	36.37
0.002	1.044	33.36
0.005	1.07	29.39
0.01	1.101	26.38
0.02	1.145	23.37
0.05	1.24	19.41
0.1	1.355	16.42
0.2	1.539	13.46
0.5	1.984	9.636
1	2.66	6.868

Note: As the passband ripple specification is changed so do the other items. Conversely, to get a particular value of VSWR or return loss, use this table to see the passband ripple that should be used to design the filter.

are much higher than those required for the low-pass (by the ratio of center to width). Capacitor Q values are generally much high than inductor Q values and so contribute far less to this effect.

In general, component-value adjustment will not be able to fully compensate for inadequate component Q values. However, if the filter is deliberately mismatched (by changing the input and/or output terminations) then a limited amount of response-*shape* correction can sometimes be achieved by network component value optimization ("tweaking"). The loss caused by Q problems (at dc in the case of a low-pass or at the center frequency in the case of a band-pass) may increase if such correction is attempted.

SIDE EFFECTS OF PASSBAND RIPPLE

Especially in RF applications it is desirable to design a filter such that the impedance seen looking into the input side remains fairly constant over the passband. Increasing the value of passband ripple increases the rate of descent from the passband into the stop band, giving a sharper cutoff. But it also degrades the uniformity of the impedance across the passband as seen looking into the input of the filter. This may be shown in terms of VSWR or return loss; those are simply different ways of stating the same effect.

Designs using a low value of passband ripple are preferred for RF work. Audio-frequency applications are generally not as critical, and so higher ripple values (up to about 0.2 dB) may be used in audio work.

Table 12.2 shows the values of VSWR and return loss for various values of passband ripple. Note that lower values for passband ripple yield better values for VSWR and return loss. Specifying a filter to have a passband ripple value of 0.01 dB will result in that filter's VSWR figure to be about 1.1:1. Or restated, the return loss will be about 26 dB. These values will be a function of frequency and at some test frequencies may be much better.

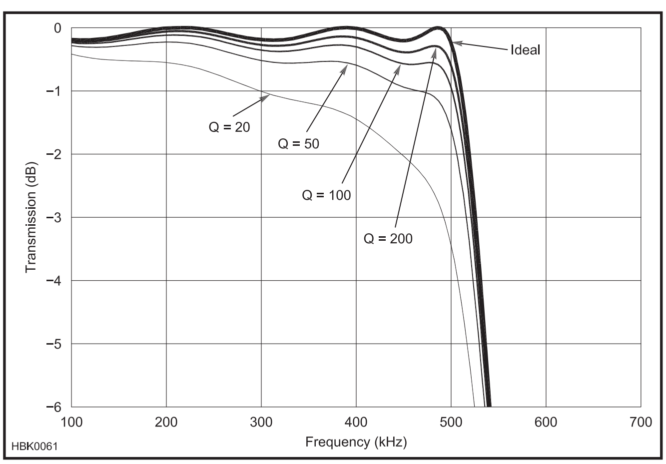


Fig 12.33 — Effect of inductor Q values on a low-pass filter.

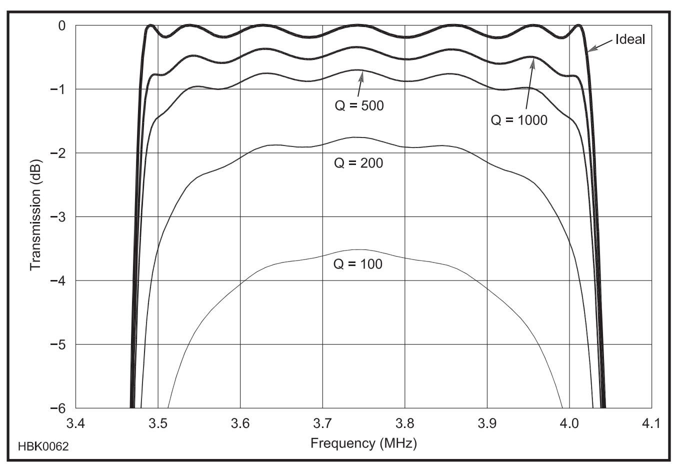


Fig 12.34 — Effect of inductor Q values on a narrowband band-pass filter.

Filter Design Examples

Although desktop computers have greatly simplified the design of lumped-element filters, going through the design process may give some degree of insight into the filter design. Working through the equations will assuredly give the designer an appreciation of what a computer with filter design and analysis software can do.

NORMALIZED VALUES

The following design examples will use *tables of normalized values.* These tables are in fact listings of previously computed component values for a low-pass configuration, designed for a bandwidth of 1 radian per second and with 1-Ω input termination. For an actual filter design they are *denormalized* to the desired bandwidth and impedance values.

The tables presented here are for various families (Butterworth, Bessel, Chebyshev and Cauer), with various values of passband ripple specification and so on. These low-pass designs may be converted to other types — such as a band-pass — with a few arithmetic operations. The numbers shown in these normalized-value tables are the capacitor values in farads and the inductor values in Henries.

The Butterworth family (**Table 12.3**) is used when there should be no ripple at all in the passband, when the response(s) are to be as flat as possible. This trait is particularly apparent near 0 Hz for the low-pass type, at band center for the band-pass type, and at infinity for the high-pass type. The resulting magnitude response will also have a relatively gentle corner from passband into the stop band.

When a low-pass filter with constant time delay throughout the passband is desired then the Bessel family, whose normalized values are shown in **Table 12.4**, should be used.

The Chebyshev family is used when a sharper cutoff is desired for a given number of components and where at least a small amount of ripple is allowable in the passband. The Chebyshev filter tables are broken into groups according to logically chosen values of passband ripple. The tables presented here are for passband ripple values of 0.01 dB (**Table 12.5**, for critical RF work), 0.044 dB (**Table 12.6**, an intermediate value offering 20 dB of return loss or 1.2:1 VSWR) and 0.2 dB (**Table 12.7**, for audio and less-critical work where a steeper descent into the stop band is of primary concern). Some published tables show figures for passband ripple values less than 0.01 dB. These are difficult to implement because of the tight component tolerances required to achieve the expected responses.

When steepness of descent from passband into stop band is the item of greatest importance, then the Cauer filter family is used. Cauer filters involve a more complicated set of choices. In addition to selecting a passband ripple, the designer must also assign a stop band depth (or stop band frequency). Some of the items interact**;** they can't all be selected arbitrarily.

The most likely combinations of items to be chosen for Cauer filters are presented in

Table 12.3
Butterworth Normalized Values

Order	*G(1)*	*G(2)*	*G(3)*	*G(4)*	*G(5)*	*G(6)*	*G(7)*	*G(8)*	*G(9)*	*G(10)*	*G(11)*	*R_{load}*
3	1	2	1									1
4	0.7654	1.848	1.848	0.7654								1
5	0.618	1.618	2	1.618	0.618							1
6	0.5176	1.414	1.932	1.932	1.414	0.5176						1
7	0.445	1.247	1.802	2	1.802	1.247	0.445					1
8	0.3902	1.111	1.663	1.962	1.962	1.663	1.111	0.3902				1
9	0.3473	1	1.532	1.879	2	1.879	1.532	1	0.3473			1
10	0.3129	0.908	1.414	1.782	1.975	1.975	1.782	1.414	0.908	0.3129		1
11	0.2846	0.8308	1.31	1.683	1.919	2	1.919	1.683	1.31	0.8308	0.2846	1

Table 12.4
Bessel Normalized Values

Order	*G(1)*	*G(2)*	*G(3)*	*G(4)*	*G(5)*	*G(6)*	*G(7)*	*G(8)*	*G(9)*	*G(10)*	*R_{load}*
3	2.203	0.9705	0.3374								1
4	2.24	1.082	0.6725	0.2334							1
5	2.258	1.111	0.804	0.5072	0.1743						1
6	2.265	1.113	0.8538	0.6392	0.4002	0.1365					1
7	2.266	1.105	0.869	0.702	0.5249	0.3259	0.1106				1
8	2.266	1.096	0.8695	0.7303	0.5936	0.4409	0.2719	0.0919			1
9	2.265	1.086	0.8639	0.7407	0.6306	0.5108	0.377	0.2313	0.078		1
10	2.264	1.078	0.8561	0.742	0.6493	0.5528	0.4454	0.327	0.1998	0.0672	1

Table 12.5
Chebyshev Normalized Values
Passband ripple 0.01

Order	*G(1)*	*G(2)*	*G(3)*	*G(4)*	*G(5)*	*G(6)*	*G(7)*	*G(8)*	*G(9)*	*G(10)*	*G(11)*	*R_{load}*
3	0.6292	0.9703	0.6292									1
4	0.7129	1.2	1.321	0.6476								0.9085
5	0.7563	1.305	1.577	1.305	0.7563							1
6	0.7814	1.36	1.69	1.535	1.497	0.7098						0.9085
7	0.7969	1.392	1.748	1.633	1.748	1.392	0.7969					1
8	0.8073	1.413	1.782	1.683	1.853	1.619	1.555	0.7334				0.9085
9	0.8145	1.427	1.804	1.713	1.906	1.713	1.804	1.427	0.8145			1
10	0.8196	1.437	1.819	1.731	1.936	1.759	1.906	1.653	1.582	0.7446		0.9085
11	0.8235	1.444	1.83	1.744	1.955	1.786	1.955	1.744	1.83	1.444	0.8235	1

Table 12.6
Chebyshev Normalized Values
Passband ripple 0.044

Order	*G(1)*	*G(2)*	*G(3)*	*G(4)*	*G(5)*	*G(6)*	*G(7)*	*G(8)*	*G(9)*	*G(10)*	*G(11)*	R_{load}
3	0.855	1.104	0.855									1
4	0.9347	1.293	1.581	0.7641								0.8175
5	0.9747	1.372	1.805	1.372	0.9747							1
6	0.9972	1.413	1.896	1.55	1.729	0.8153						0.8175
7	1.011	1.437	1.943	1.621	1.943	1.437	1.011					1
8	1.02	1.452	1.969	1.657	2.026	1.61	1.776	0.8341				0.8175
9	1.027	1.462	1.986	1.677	2.067	1.677	1.986	1.462	1.027			1
10	1.031	1.469	1.998	1.69	2.091	1.709	2.067	1.633	1.797	0.8431		0.8175
11	1.035	1.474	2.006	1.698	2.105	1.727	2.105	1.698	2.006	1.474	1.035	1

Table 12.7
Chebyshev Normalized Values
Passband ripple 0.2

Order	*G(1)*	*G(2)*	*G(3)*	*G(4)*	*G(5)*	*G(6)*	*G(7)*	*G(8)*	*G(9)*	*G(10)*	*G(11)*	R_{load}
3	1.228	1.153	1.228									1
4	1.303	1.284	1.976	0.8468								0.65
5	1.339	1.337	2.166	1.337	1.339							1
6	1.36	1.363	2.239	1.456	2.097	0.8838						0.65
7	1.372	1.378	2.276	1.5	2.276	1.378	1.372					1
8	1.38	1.388	2.296	1.522	2.341	1.493	2.135	0.8972				0.65
9	1.386	1.394	2.309	1.534	2.373	1.534	2.309	1.394	1.386			1
10	1.39	1.398	2.318	1.542	2.39	1.554	2.372	1.507	2.151	0.9035		0.65
11	1.393	1.402	2.324	1.547	2.401	1.565	2.401	1.547	2.324	1.402	1.393	1

Tables 12.8 through **12.19**. These tables are for passband ripple values of 0.01, 0.044 and 0.2 dB (the same values that were chosen for the Chebyshev family) and stop band depths of 30, 40, 50 and 60 dB. The frequency at which the attenuation first reaches the design stop band depth value is shown in the final column, labeled F_{stop}. As with the Chebyshev filters, some published tables show ripple values of less than 0.01 dB but such designs are difficult to implement in practice because of the tight tolerances required on all of the components.

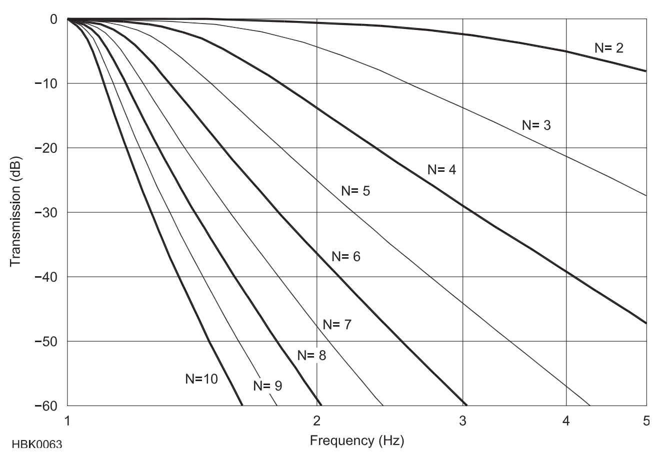


Fig 12.35 — This graph shows transmission for filters with orders from 2 through 10 for Chebyshev low-pass filter with 0.01 dB passband ripple. For a given order, higher passband ripple values will have higher attenuation (less transmission) at a given frequency while Butterworth filters will have less attenuation at a given frequency.

DESIGN EXAMPLE — LOW-PASS

Here is an example of using normalized-value tables to design a low-pass filter. **Fig 12.35** illustrates the attenuation to be expected for a Chebyshev low-pass filter with 0.01 dB of passband ripple and various orders as shown. For example, a fourth-order low-pass would have an attenuation of about 14 dB one octave above cutoff. If the intended application requires a response at least 25 dB down at one octave above cutoff, then a fifth-order filter is needed.

Fig 12.36 shows the layout of a fifth-order low-pass filter that will meet the requirements. The next step is to define the input termination (50 Ω in this case) and the ripple bandwidth (4.2 MHz for this example).

Next, refer to Table 12.5 to obtain the normalized component values for the 1-Ω and 1 radian/second low-pass. Choosing the line of component values for fifth order, G(1) is 0.7563, G(2) is 1.305, G(3) is 1.577, G(4) is 1.305 and G(5) is 0.7563. The last value, R_{load}, is used to calculate the output termination.

The calculations for each filter component are shown in Fig 12.36A, and the resulting component values are shown in Fig 12.36B. In the equations for calculating parts values, $\Omega_c = 2\pi F_{width}$. In this case, $\Omega_c = 2 \times 3.1416 \times 4.2 = 26.389$.

R is the filter's input termination, in this case 50 Ω. The termination on the output side of the filter will be the input termination value multiplied by R_{load}. In this case, $R_{load} = 1$, so the input and output terminations are equal. (Chebyshev *even-ordered* low-pass and high-pass filters have an output termination different from the input, as shown in the last column under R_{load}.)

DESIGN EXAMPLE — LOW-PASS (CAUER)

The design of a Cauer low-pass filter is very similar to the design of a Chebyshev, but it has added capacitors to form the traps. Graphs showing the stop band performance or loss of the various Cauer filters in the tables are impractical because of the large number of options to be chosen.

This example illustrates the design of a fifth-order Cauer capacitor-input low-pass filter with 0.044 dB passband ripple and 40-dB stop band depth. The input termination will be 50 Ω and the ripple bandwidth will again be 4.2 MHz.

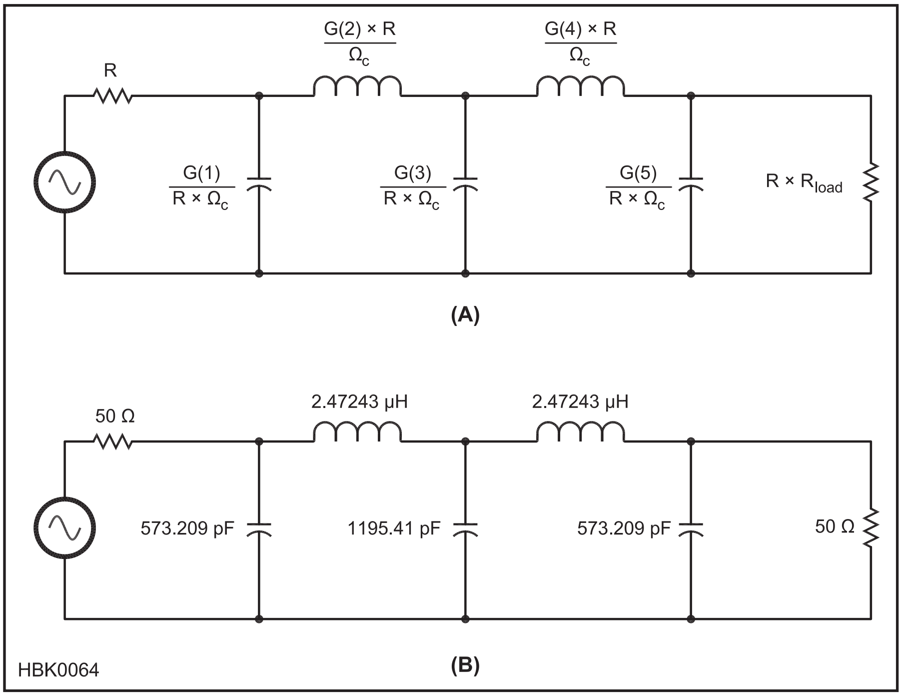


Fig 12.36 — Design example for a Chebyshev low-pass filter using the normalized filter tables. The topology and design equations are shown at A, with the resulting calculated parts values at B.

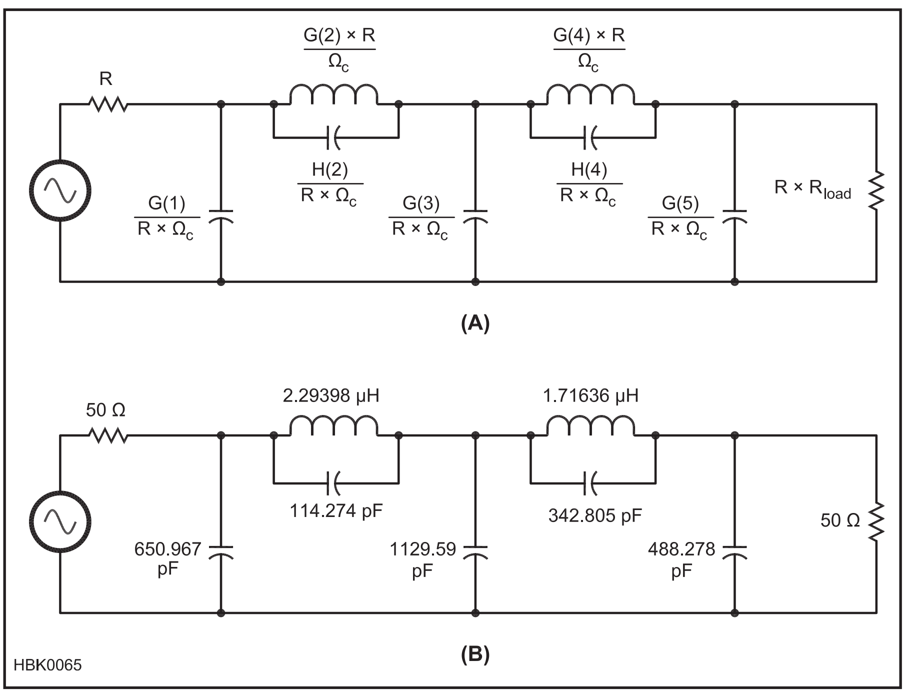


Fig 12.37 — Design example for a Cauer low-pass filter using the normalized filter tables. The topology and design equations are shown at A, with the resulting calculated parts values at B.

Fig 12.37 shows the filter layout. The first step is to obtain from Table 12.13 the normalized 1-Ω and 1 radian/second component values. Choosing the line of component values for fifth order, G(1) is 0.8597, G(2) is 1.211, G(3) is 1.491, G(4) is 0.9058 and G(5) is 0.6448. In addition, the "trap" capacitor values must be retrieved. The trap capacitors are shown as "H" values. H(2) is 0.1509 and H(4) is 0.4527. The output termination is 1 and so is same as the input termination.

The filter with the design equations appears in Fig 12.37A, with the resulting component values in Fig 12.37B.

DESIGN EXAMPLE — HIGH-PASS

The design of a high-pass filter will now be illustrated. Fig 12.35 can be used to estimate the response of a high-pass design just as it was for a low-pass. The only manipulation that needs to be done is to use the *reciprocal of frequency* when estimating the magnitude response. The response will be down a given number of dB at half the cutoff instead of twice the cutoff, as an example.

For this example, we will use a fourth-order capacitor-input Butterworth high-pass type with the 3-dB point at 250 Hz and system impedance of 600 Ω. The schematic is shown in **Fig 12.38**.

The normalized values for the low-pass are taken from Table 12.3, and the design equations are shown in Fig 12.38A. From the line for Order = 4, the value for G(1) is 0.7654, G(2) is 1.848, G(3) is 1.848 and G(4) is 0.7654.

For the high-pass configuration, the output termination is based on the reciprocal of R_{load}. In this case, the table shows that the output termination is 1, so it is the same as the input. (Unlike the Chebyshev family, the output termination for the Butterworth family is always the same as the input termination.)

Fig 12.38B shows the component values for this filter.

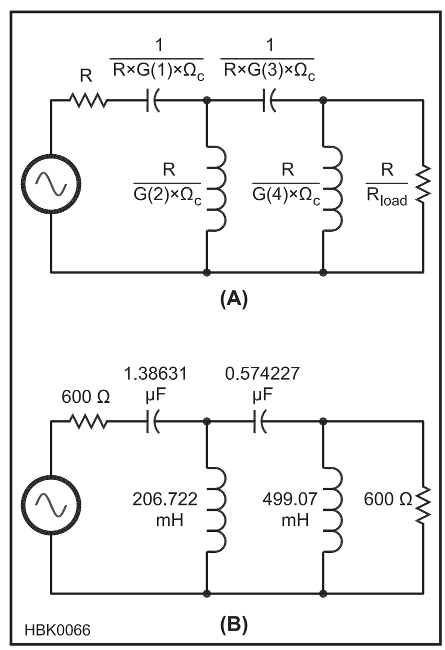


Fig 12.38 — Design example for a Butterworth high-pass filter using the normalized filter tables. The topology and design equations are shown at A, with the resulting calculated parts values at B. For a Cauer family high-pass, trap capacitors are inserted in series with the shunt inductors. Their value is calculated by the same expression as the series capacitors, but using the H( ) values from the Cauer tables.

Table 12.8
Cauer Normalized Values
Passband ripple 0.01, Stop band depth 30

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	0.5835	0.8848	0.06895	0.5835							1	3.524
4	0.4844	0.954	0.1799	1.109	0.6392	0					1	2.215
5	0.6173	1.106	0.1913	1.264	0.7084	0.6533	0.3403				1	1.418
6	0.4158	0.936	0.3958	1.074	0.7642	0.7979	0.9396	0.7292	0		1	1.252
7	0.6095	1.117	0.2497	1.027	0.5503	1.459	0.8809	0.5513	1.175	0.1594	1	1.105

Table 12.9
Cauer Normalized Values
Passband ripple 0.01, Stop band depth 40

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	0.6074	0.9299	0.03092	0.6074							1	5.12
4	0.5463	1.057	0.09503	1.144	0.634	0					1	2.885
5	0.6678	1.179	0.1179	1.359	0.9065	0.3574	0.4904				1	1.687
6	0.5089	1.063	0.2549	1.208	0.992	0.4761	1.066	0.7275	0		1	1.416
7	0.6636	1.197	0.173	1.186	0.7766	0.8845	1.04	0.7415	0.7093	0.3321	1	1.191

Table 12.10
Cauer Normalized Values
Passband ripple 0.01, Stop band depth 50

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	0.6194	0.9518	0.01415	0.6194							1	7.47
4	0.5813	1.116	0.05176	1.165	0.6309	0					1	3.797
5	0.7007	1.225	0.07357	1.432	1.045	0.2096	0.5872				1	2.045
6	0.5739	1.156	0.168	1.317	1.167	0.3023	1.157	0.7255	0		1	1.634
7	0.7019	1.252	0.122	1.319	0.9711	0.5823	1.192	0.899	0.4614	0.4575	1	1.309

DESIGN EXAMPLE — WIDE BAND-PASS

The design of a band-pass filter of appreciable percentage bandwidth will be looked at next. For the purposes of this discussion, “appreciable” means a bandwidth of about 20% of the center frequency. Again, Fig 12.35 can be used to approximate the required order, bearing in mind that it is precise for passband ripple of 0.01 dB.

Fig 12.39 shows this example, a third-order shunt-input Chebyshev type with 0.2 dB of passband ripple with an input termination of 75 Ω. The center frequency is to be 4 MHz and the ripple bandwidth is to be 1 MHz.

The normalized values for the low-pass are taken from Table 12.7. Choosing the line of component values for third order, the value for G(1) is 1.228, G(2) is 1.153 and G(3) is 1.228.

In the equations used to calculate the actual parts values, Q_L is the loaded Q. This is the ratio of F_{center} to F_{width}. Note that in this topology both Ω_c and Ω_o are used; as

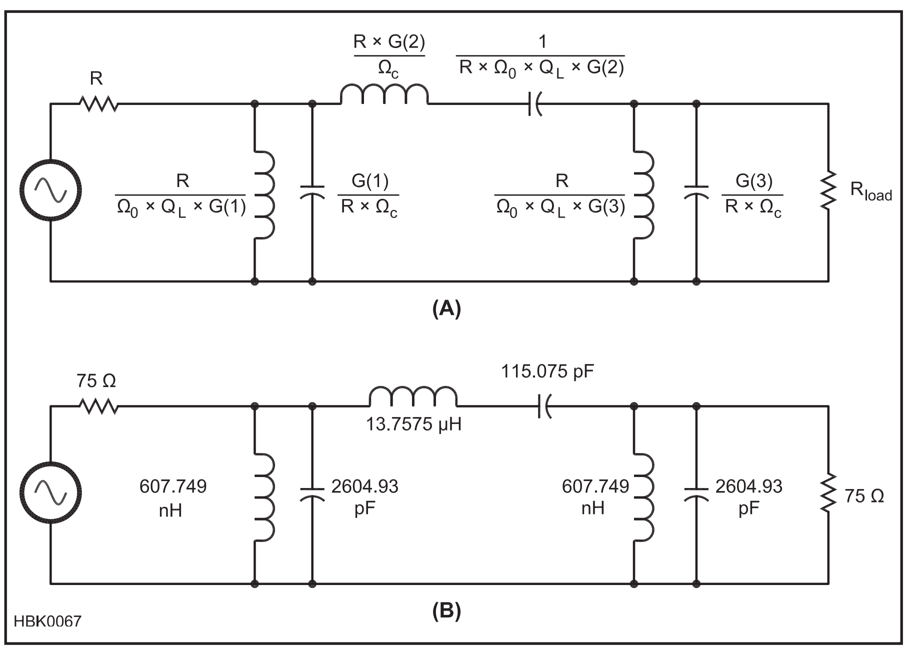


Fig 12.39 — Design example for a Chebyshev wide band-pass filter using the normalized filter tables. The topology and design equations are shown at A, with the resulting calculated parts values at B.

Table 12.11
Cauer Normalized Values
Passband ripple 0.01, Stop band depth 60

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	0.6246	0.9617	0.00652	0.6246							1	10.94
4	0.6011	1.15	0.02862	1.178	0.629	0					1	5.025
5	0.7209	1.254	0.04608	1.482	1.137	0.1269	0.6486				1	2.514
6	0.6191	1.222	0.1121	1.399	1.296	0.1978	1.223	0.7238	0		1	1.913
7	0.7283	1.291	0.08673	1.425	1.131	0.3982	1.322	1.024	0.312	0.5489	1	1.463

Table 12.12
Cauer Normalized Values
Passband ripple 0.044, Stop band depth 30

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	0.7873	0.9891	0.09924	0.7873							1	2.788
4	0.6172	1.046	0.2284	1.239	0.8101	0					1	1.883
5	0.794	1.119	0.2452	1.353	0.6862	0.8268	0.4694				1	1.286
6	0.5391	0.9646	0.4632	1.064	0.7337	0.9564	0.9908	0.8958	0		1	1.171
7	0.7808	1.11	0.3034	1.058	0.4585	1.892	0.8492	0.5271	1.379	0.2962	1	1.065

Table 12.13
Cauer Normalized Values
Passband ripple 0.044, Stop band depth 40

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	0.8229	1.05	0.04455	0.8229							1	4.02
4	0.6958	1.176	0.1209	1.287	0.8064	0					1	2.429
5	0.8597	1.211	0.1509	1.491	0.9058	0.4527	0.6448				1	1.504
6	0.6471	1.112	0.299	1.227	0.9893	0.5657	1.126	0.8992	0		1	1.304
7	0.8479	1.204	0.2085	1.242	0.6828	1.118	1.045	0.7251	0.833	0.4797	1	1.132

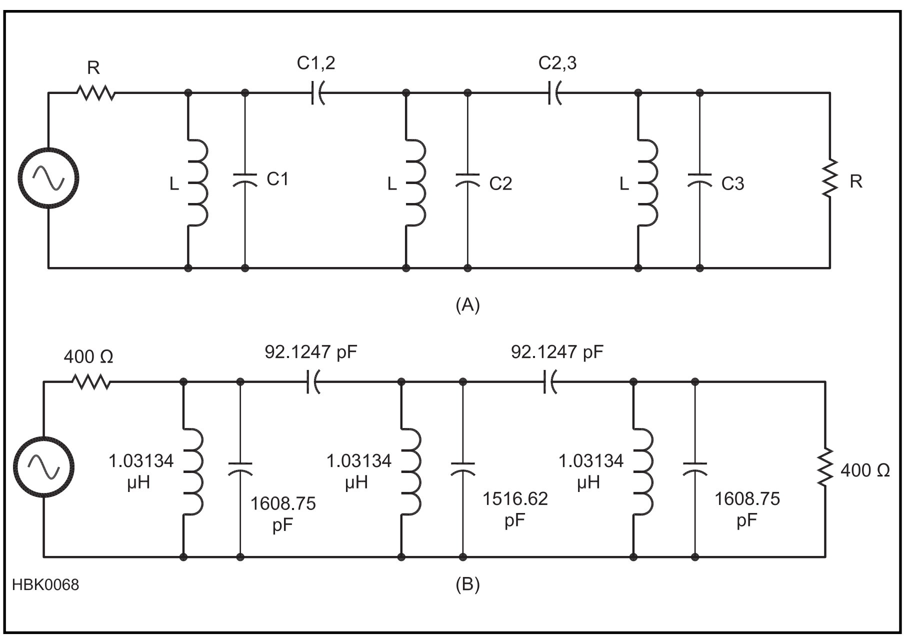


Fig 12.40 — Design example for a Chebyshev narrow band-pass filter using the normalized filter tables. The topology is shown at A and the and design equations are in the text, with the resulting calculated parts values at B.

in the previous examples, $\Omega_c = 2\ \pi\ F_{width}$, while $\Omega_o = 2\ \pi\ F_{center}$.

The filter in schematic form with the equations appears in Fig 12.39A, and calculated values are shown in Fig 12.39B.

DESIGN EXAMPLE — NARROW BAND-PASS

Narrow band-pass filters involve more calculations than wide band-pass types. The topology used in this discussion is be a set of shunt-to-ground parallel-tuned resonators coupled by relatively small value coupling capacitors. This topology is shown in **Fig 12.40**.

This example will be for a band-pass filter with a bandwidth of 200 kHz, which is centered at 3.8 MHz, using the Chebyshev family with a passband ripple of 0.044 dB. The order will be three, and the normalized values are found in Table 12.6.

As in the wide band-pass example, the equations used to calculate the actual parts values use Q_L (the loaded Q), which is the ratio of F_{center} to F_{width}. Once again both Ω_c

Table 12.14
Cauer Normalized Values
Passband ripple 0.044, Stop band depth 50

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	0.8401	1.079	0.02038	0.8401							1	5.85
4	0.7411	1.253	0.06596	1.316	0.8039	0					1	3.178
5	0.9012	1.269	0.094	1.594	1.064	0.2658	0.761				1	1.803
6	0.724	1.22	0.1977	1.362	1.192	0.3584	1.227	0.9004	0		1	1.487
7	0.8928	1.27	0.1462	1.401	0.8857	0.7262	1.232	0.8927	0.5426	0.6158	1	1.229

Table 12.15
Cauer Normalized Values
Passband ripple 0.044, Stop band depth 60

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	0.8481	1.093	0.0094	0.8481							1	8.553
4	0.767	1.297	0.0365	1.333	0.8025	0					1	4.192
5	0.9275	1.307	0.05888	1.667	1.172	0.1613	0.8371				1	2.199
6	0.7782	1.298	0.1323	1.465	1.346	0.2346	1.301	0.9008	0		1	1.725
7	0.9275	1.317	0.104	1.531	1.056	0.4957	1.395	1.027	0.3689	0.7207	1	1.359

Table 12.16
Cauer Normalized Values
Passband ripple 0.2, Stop band depth 30

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	1.116	0.9996	0.1584	1.116							1	2.207
4	0.8176	1.087	0.3037	1.335	1.065	0					1	1.61
5	1.084	1.031	0.3443	1.482	0.5907	1.153	0.6819				1	1.18
6	0.7319	0.9415	0.5675	1.024	0.6677	1.186	1.002	1.149	0		1	1.106
7	1.065	1.01	0.4035	1.124	0.3345	2.756	0.8246	0.451	1.785	0.5138	1	1.036

Table 12.17
Cauer Normalized Values
Passband ripple 0.2, Stop band depth 40

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	1.174	1.08	0.07104	1.174							1	3.147
4	0.9224	1.254	0.1607	1.399	1.067	0					1	2.047
5	1.175	1.14	0.2106	1.686	0.8175	0.6264	0.8974				1	1.351
6	0.8638	1.107	0.3667	1.221	0.9522	0.6869	1.139	1.164	0		1	1.21
7	1.156	1.116	0.2739	1.35	0.5366	1.558	1.073	0.6422	1.069	0.7205	1	1.084

Table 12.18
Cauer Normalized Values
Passband ripple 0.2, Stop band depth 50

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	1.203	1.118	0.03253	1.203							1	4.554
4	0.9842	1.355	0.08773	1.438	1.067	0					1	2.654
5	1.233	1.211	0.1309	1.842	0.9885	0.3676	1.047				1	1.595
6	0.9599	1.23	0.2433	1.387	1.19	0.4317	1.246	1.173	0		1	1.36
7	1.213	1.192	0.1904	1.548	0.7322	0.9884	1.313	0.8105	0.694	0.8771	1	1.161

Table 12.19
Cauer Normalized Values
Passband ripple 0.2, Stop band depth 60

Order	*G(1)*	*G(2)*	*H(2)*	*G(3)*	*G(4)*	*H(4)*	*G(5)*	*G(6)*	*H(6)*	*G(7)*	R_{load}	F_{stop}
3	1.216	1.137	0.01501	1.216							1	6.64
4	1.02	1.413	0.04859	1.46	1.067	0					1	3.484
5	1.272	1.256	0.08207	1.953	1.108	0.2237	1.149				1	1.924
6	1.029	1.32	0.1634	1.517	1.375	0.2818	1.328	1.178	0		1	1.56
7	1.257	1.244	0.1348	1.716	0.9025	0.6672	1.527	0.9477	0.4718	1.001	1	1.269

and Ω_o are used: $\Omega_c = 2\pi F_{width}$, while $\Omega_o = 2\pi F_{center}$.

The same inductor, L, is used for each resonator. The value of that inductor is given by:

$$L = \frac{R}{\Omega_o \times Q_L \times G(1)} \quad (1)$$

Each resonator has a basic tuning capacitor whose value is given by:

$$C_{basic} = \frac{G(1)}{R \times \Omega_c} \quad (2)$$

Next calculate the inter-resonator coupling capacitors:

$$C1,2 = \frac{G(1)}{R \times \Omega_o}\sqrt{\frac{1}{G(1) \times G(2)}} \quad (3)$$

$$C2,3 = \frac{G(1)}{R \times \Omega_o}\sqrt{\frac{1}{G(2) \times G(3)}} \quad (4)$$

The actual shunt tuning capacitors for each resonator are then the basic tuning capacitor minus the coupling capacitors on each side, as shown here:

$C1 = C_{basic} - C1,2$
$C2 = C_{basic} - C1,2 - C2,3$
$C3 = C_{basic} - C2,3$

The completed design is shown in Fig 12.40B. It should be evident how to extend this procedure to higher-order filters.

Quartz Crystal Filters

Practical inductor Q values effectively set the minimum achievable bandwidth limits for LC band-pass filters. Higher-Q circuit elements must be employed to extend these limits. These high-Q resonators include PZT ceramic, mechanical and coaxial devices. However, the quartz crystal provides the highest Q and best stability with temperature and time of all available resonators. Quartz crystals suitable for filter use are fabricated over a frequency range from audio to VHF.

The quartz resonator has the equivalent circuit shown in **Fig 12.41**. L_s, C_s and R_s represent the *motional* reactances and loss resistance. C_p is the parallel plate capacitance formed by the two metal electrodes separated by the quartz dielectric. Quartz has a dielectric constant of 3.78. **Table 12.20** shows parameter values for typical moderate-cost quartz resonators. Q_U is the resonator unloaded Q.

$$Q_U = 2\pi f_s r_s \quad (5)$$

Q_U is very high, usually exceeding 25,000. Thus the quartz resonator is an ideal component for the synthesis of a high-Q band-pass filter.

A quartz resonator connected between generator and load, as shown in **Fig 12.42A**,

Table 12.20
Typical Parameters for AT-Cut Quartz Resonators

Freq (MHz)	*Mode n*	*rs (Ω)*	*Cp (pF)*	*Cs (pF)*	*L (mH)*	*Q_U*
1.0	1	260	3.4	0.0085	2900	72,000
5.0	1	40	3.8	0.011	100	72,000
10.0	1	8	3.5	0.018	14	109,000
20	1	15	4.5	0.020	3.1	26,000
30	3	30	4.0	0.002	14	87,000
75	3	25	4.0	0.002	2.3	43,000
110	5	60	2.7	0.0004	5.0	57,000
150	5	65	3.5	0.0006	1.9	27,000
200	7	100	3.5	0.0004	2.1	26,000

Courtesy of Piezo Crystal Co, Carlisle, PA

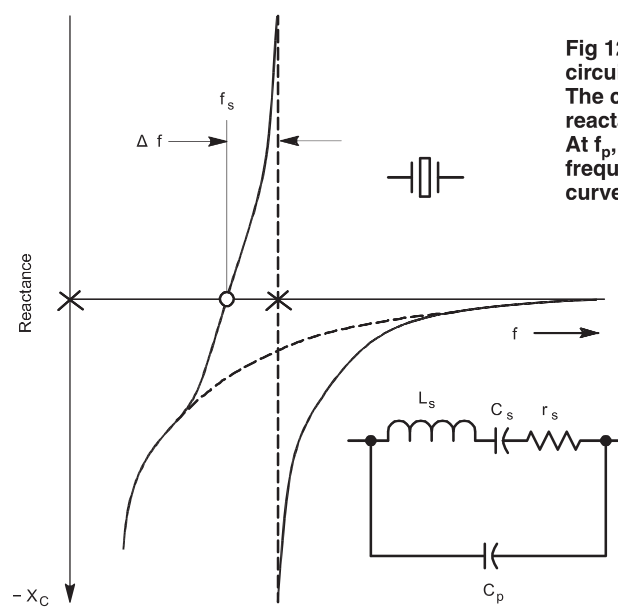


Fig 12.41 — Equivalent circuit of a quartz crystal. The curve plots the crystal reactance against frequency. At f_p, the resonance frequency, the reactance curve goes to infinity.

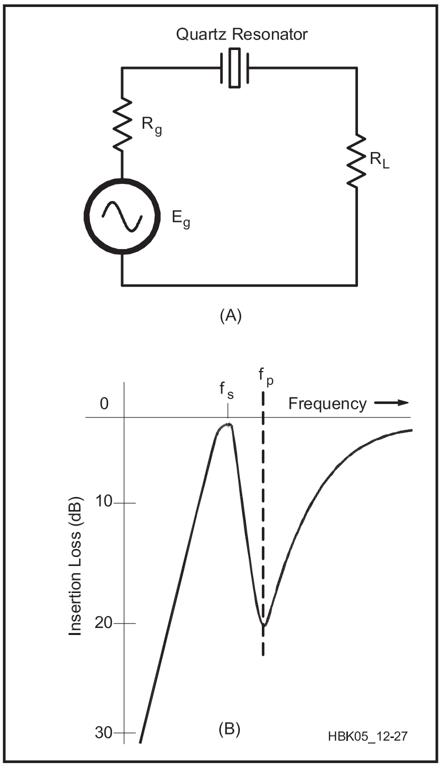


Fig 12.42 — A: Series test circuit for a crystal. In the test circuit the output of a variable frequency generator, e_g, is used as the test signal. The frequency response in B shows the highest attenuation at resonance (f_p). See text.

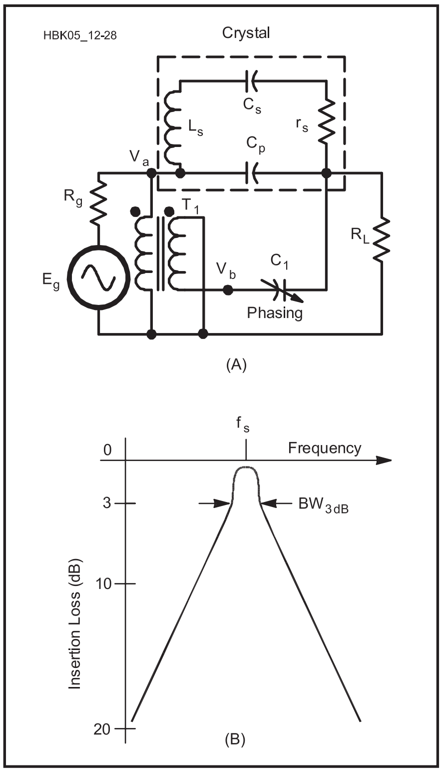


Fig 12.43 — The practical one-stage crystal filter in A has the response shown in B. The phasing capacitor is adjusted for best response (see text).

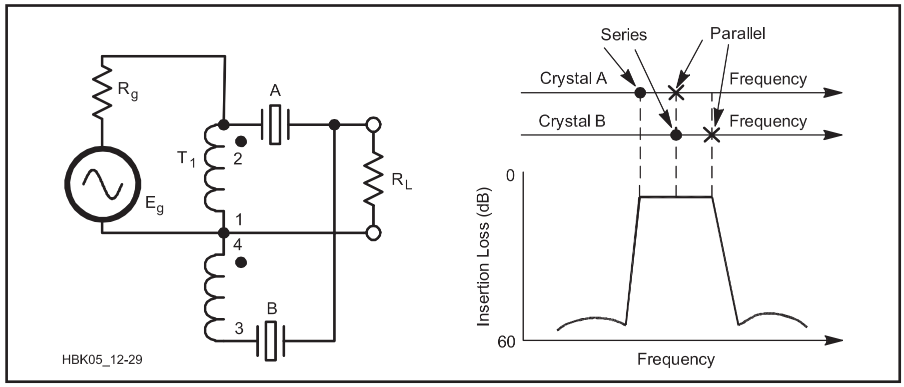


Fig 12.44 — A half-lattice crystal filter. No phasing capacitor is needed in this circuit.

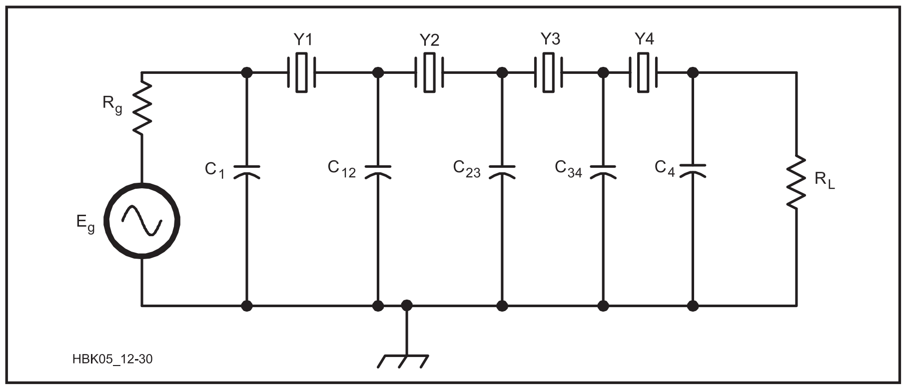


Fig 12.45 — A four-stage crystal ladder filter. The crystals must be chosen properly for best response.

produces the frequency response of Fig 12.42B. There is a relatively low loss at the series resonant frequency f_s and high loss at the parallel resonant frequency f_p. The test circuit of Fig 12.42A is useful for determining the parameters of a quartz resonator, but yields a poor filter.

A crystal filter developed in the 1930s is shown in **Fig 12.43A**. The disturbing effect of C_p (which produces f_p) is canceled by the *phasing capacitor,* C1. The voltage-reversing transformer T1 usually consists of a bifilar winding on a ferrite core. Voltages V_a and V_b have equal magnitude but 180° phase difference. When C1 = C_p, the effect of C_p will disappear and a well-behaved single resonance will occur as shown in Fig 12.43B. The bandpass filter will exhibit a loaded Q given by:

$$Q_L = \frac{2 \pi f_S L_S}{R_L} \qquad (6)$$

This single-stage "crystal filter," operating at 455 kHz, was present in almost all high-quality amateur communications receivers up through the 1960s. When the filter was switched into the receiver IF amplifier the bandwidth was reduced to a few hundred Hz for Morse code reception.

The half-lattice filter shown in **Fig 12.44** is an improvement in crystal filter design. The quartz resonator parallel-plate capacitors, C_p, cancel each other. Remaining series resonant circuits, if properly offset in frequency, will produce an approximate 2-pole Butterworth or Chebyshev response. Crystals A and B are usually chosen so that the parallel resonant frequency (f_p) of one is the same as the series resonant frequency (f_s) of the other.

Half-lattice filter sections can be cascaded to produce a composite filter with many poles. Until recently, most vendor-supplied commercial filters were lattice types. Ref 11 discusses the computer design of half-lattice filters.

Many quartz crystal filters produced today use the ladder network design shown in **Fig 12.45**. In this configuration, all resonators have the same series resonant frequency f_s. Inter-resonator coupling is provided by shunt capacitors such as C12 and C23. Refs 12 and 13 provide good ladder filter design information.

CRYSTAL-FILTER EVALUATION

Crystal filters are often constructed of surplus crystals or crystals whose characteristics are not exactly known. Randy Henderson, WI5W, developed a swept frequency generator for testing these filters. It was first described in March 1994 *QEX*. This test instrument adds to the ease and success in quickly building filters from inexpensive microprocessor crystals.

A template, containing additional information, is available on the CD included with this book.

An Overview

The basic setup is shown in **Fig 12.46**. The VCO is primarily a conventional LC-tuned Hartley oscillator with its frequency tuned over a small range by a varactor diode (MV2104). Other varactors may be used as long as the capacitance specifications aren't too different. Change the 5-pF coupling capacitor to expand the sweep width if desired.

The VCO signal goes through a buffer

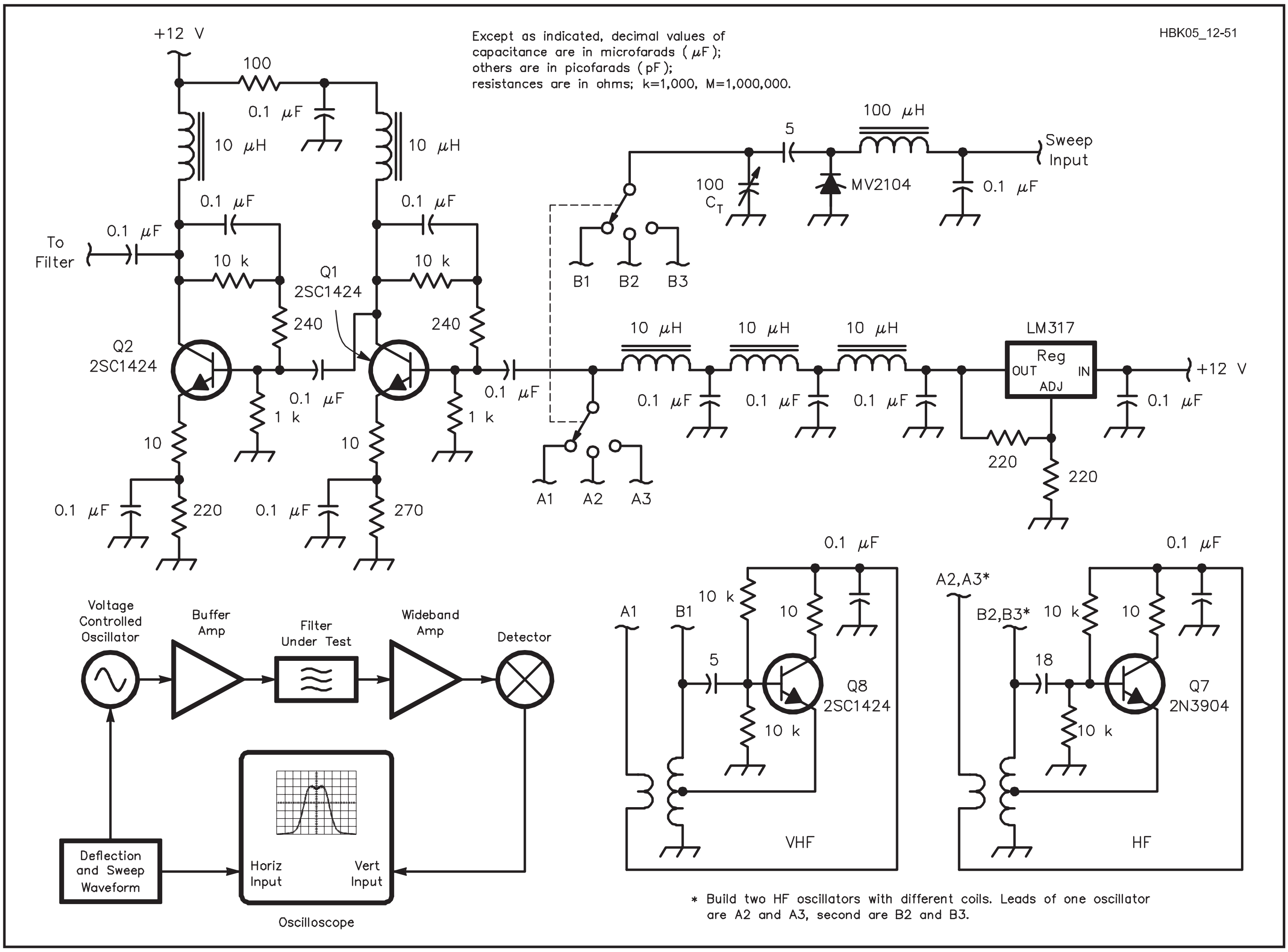


Fig 12.46 — The test set block diagram, lower left, starts with a swept frequency oscillator, shown in the schematic. If a commercial swept-frequency oscillator is available, it can be substituted for the circuit shown.

amplifier to the filter under test. The filter is followed by a wide-bandwidth amplifier and then a detector. The output of the detector is a rectified and filtered signal. This varying dc voltage drives the vertical input of an oscilloscope. At any particular time, the deflection and sweep circuitry commands the VCO to "run at this frequency." The same deflection voltage causes the oscilloscope beam to deflect left or right to a position corresponding to the frequency.

Any or all of these circuits may be eliminated by the use of appropriate commercial test equipment. For example, a commercial sweep generator would eliminate the need for everything but the wide-band amplifier and detector. Mini-Circuits Labs and many others sell devices suitable for the wide-band amplifiers and detector.

The generator/detector system covers approximately 6 to 74 MHz in three ranges. Each tuning range uses a separate RF oscillator module selected by switch S1. The VCO output and power-supply input are multiplexed on the "A" lead to each oscillator. The tuning capacitance for each VCO is switched into the appropriate circuit by a second set of contacts on S1. C_T is the coarse tuning adjustment for each oscillator module.

Two oscillator coils are wound on PVC plastic pipe. The third, for the highest frequency range, is self-supporting #14 copper wire. Although PVC forms with Super Glue dope may not be "state of the art" technology, frequency stability is completely adequate for this instrument. See **Table 12.21** for details.

The oscillator and buffer stage operate at low power levels to minimize frequency drift caused by component heating. Crystal filters cause large load changes as the frequency is swept in and out of the passband. These large changes in impedance tend to "pull" the oscil-

Table 12.21
VCO Coils

Coil	*Inside Diameter (inches)*	*Length (inches)*	*Turns, Wire*	*Inductance (μH)*
large	0.85	1.1	18 t, #28	5.32
medium	0.85	0.55	7 t, #22	1.35
small	0.5	0.75	5 t, #14	0.27

The two larger coils are wound on ¾-inch PVC pipe and the smaller one on a ½-inch drill bit. Tuning coverage for each oscillator is obtained by squeezing or spreading the turns before gluing them in place. The output windings connected to A1, A2 and A3 are each single turns of #14 wire spaced off the end of the tapped coils. The taps are approximate and 25 to 30% of the full winding turns—up from the cold (ground) end of the coils.

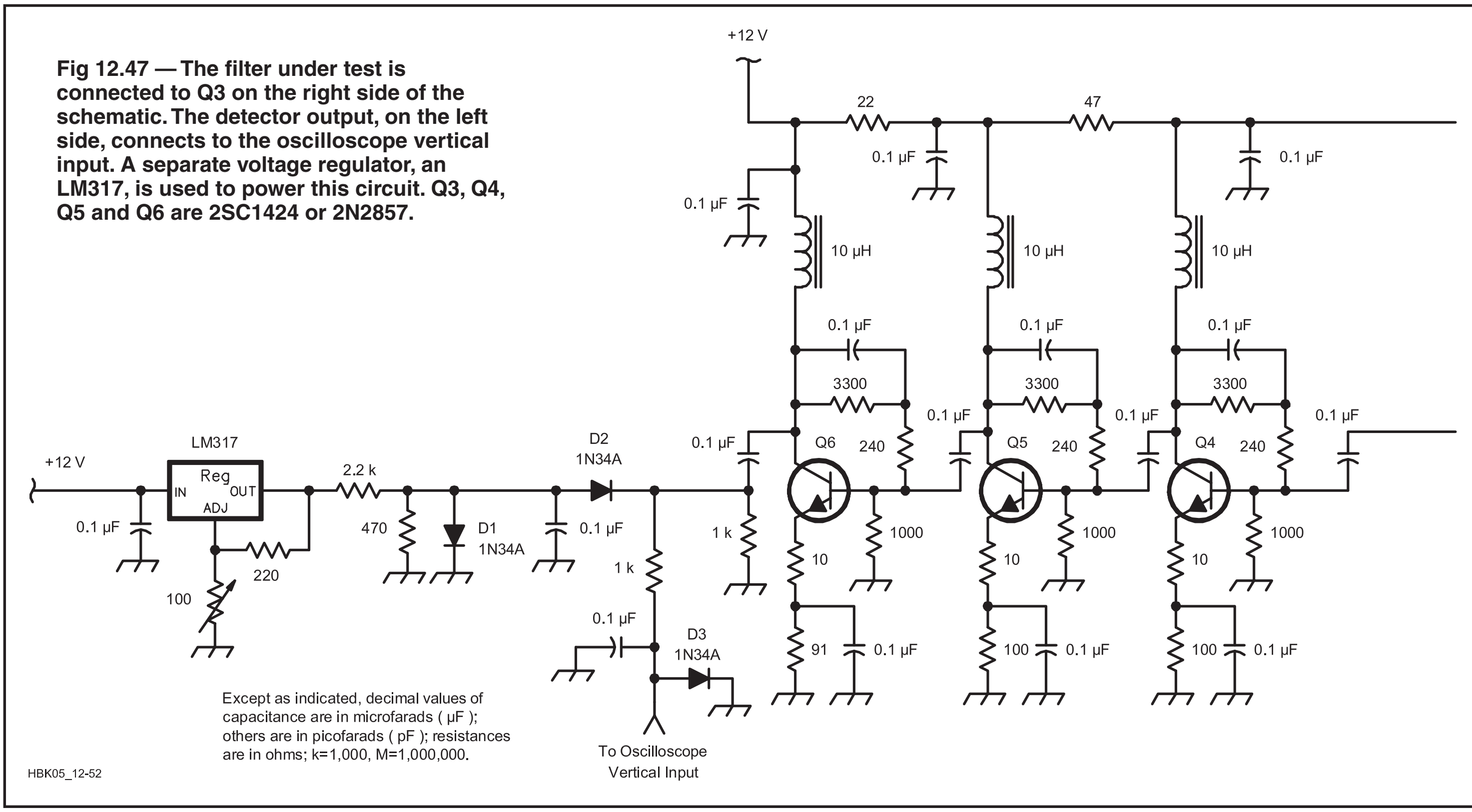


Fig 12.47 — The filter under test is connected to Q3 on the right side of the schematic. The detector output, on the left side, connects to the oscilloscope vertical input. A separate voltage regulator, an LM317, is used to power this circuit. Q3, Q4, Q5 and Q6 are 2SC1424 or 2N2857.

lator frequency and cause inaccuracies in the passband shape depicted by the oscilloscope. Therefore a buffer amplifier is a necessity. The wide-band amplifier in **Fig 12.47** is derived from one in ARRL's *Solid State Design for the Radio Amateur* (out of print).

S2 selects a 50-Ω, 10-dB attenuator in the input line. When the attenuator is in the line, it provides a better output match for the filter under test. The detector uses some forward bias for D2. A simple unbiased diode detector would offer about 50 dB of dynamic range. Some dc bias increases the dynamic range to almost 70 dB. D3, across the detector output (the scope input), increases the vertical-amplifier sensitivity while compressing or limiting the response to high-level signals. With this arrangement, high levels of attenuation (low-level signals) are easier to observe and low attenuation levels are still visible on the CRT. The diode only kicks in to provide limiting at higher signal levels.

The horizontal-deflection sweep circuit uses a dual op-amp IC (see **Fig 12.48**). One section is an oscillator; the other is an integrator. The integrator output changes linearly with time, giving a uniform brightness level as the trace is moved from side to side. Increasing C1 decreases the sweep rate. Increasing C2 decreases the slope of the output waveform ramp.

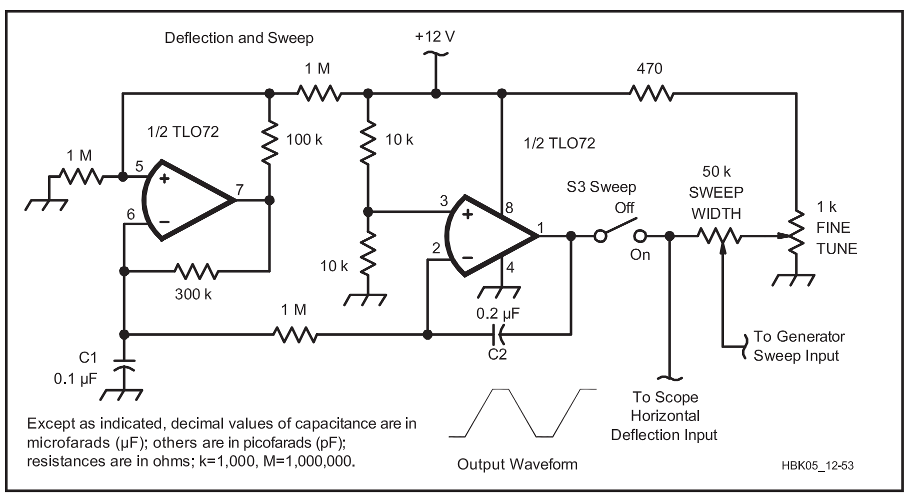


Fig 12.48 — The sweep generator provides both an up and down sweep voltage (see text) for the swept frequency generator and the scope horizontal channel.

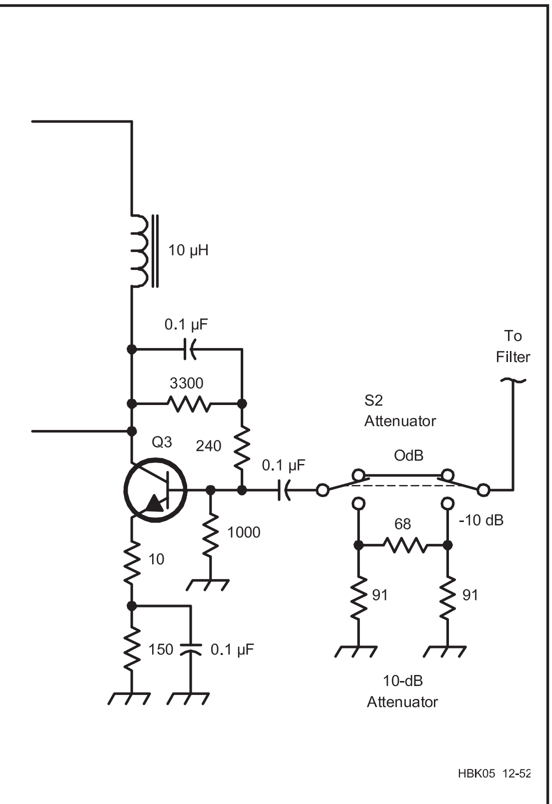
Operation

The CRT is swept in both directions, left to right and right to left. The displayed curve is a result of changes in frequency, not time. Therefore it is unnecessary to incorporate the usual right-to-left, snap-back and retrace blanking used in oscilloscopes.

S3 in Fig 12.48 disables the automatic sweep function when opened. This permits manual operation. Use a frequency counter to measure the VCO output, from which bandwidth can be calculated. Turn the fine-tune control to position the CRT beam at selected points of the passband curve. The difference in frequency readings is the bandwidth at that particular point or level of attenuation.

Substitution of a calibrated attenuator for the filter under test can provide reference readings. These reference readings may be used to calibrate an otherwise uncalibrated scope vertical display in dB.

The buffer amplifier shown here is set up to drive a 50-Ω load, and the wide-bandwidth amplifier input impedance is about 50 Ω. If the filter is not a 50-Ω



unit, however, various methods can be used to accommodate the difference. For example, a transformer may be used for widely differing impedance levels, whereas a minimum-loss resistive pad may be preferable where impedance levels differ by a factor of approximately 1.5 or less—presuming some loss is acceptable.

References

A. Ward, "Monolithic Microwave Integrated Circuits," Feb 1987 *QST*, pp 23-29.

Z. Lau, "A Logarithmic RF Detector for Filter Tuning," Oct 1988 *QEX*, pp 10-11.

Monolithic Crystal Filters

A monolithic (Greek: one-stone) crystal filter has two sets of electrodes deposited on the same quartz plate, as shown in **Fig 12.49**. This forms two resonators with acoustic (mechanical) coupling between them. If the acoustic coupling is correct, a 2-pole Butterworth or Chebyshev response will be achieved. More than two resonators can be fabricated on the same plate yielding a multipole response. Monolithic crystal filter technology is popular because it produces a low parts count, single-unit filter at lower cost than a lumped-element equivalent. Monolithic crystal filters are typically manufactured in the range from 5 to 30 MHz for the fundamental mode and up to 90 MHz for the third-overtone mode. Q_L ranges from 200 to 10,000.

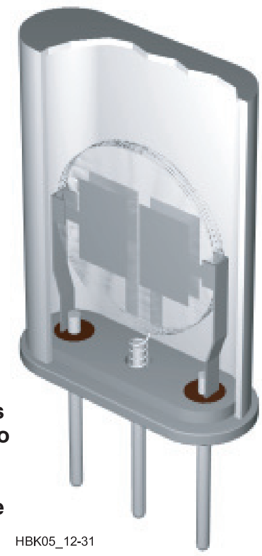


Fig 12.49 — Typical two-pole monolithic crystal filter. This single small (½ to ¾-inch) unit can replace 6 to 12, or more, discrete components.

SAW Filters

The resonators in a monolithic crystal filter are coupled together by bulk acoustic waves. These acoustic waves are generated and propagated in the interior of a quartz plate. It is also possible to launch, by an appropriate transducer, acoustic waves that propagate only along the surface of the quartz plate. These are called "surface-acoustic-waves" because they do not appreciably penetrate the interior of the plate.

A surface-acoustic-wave (SAW) filter consists of thin aluminum electrodes, or fingers, deposited on the surface of a piezoelectric substrate as shown in **Fig 12.50**. Lithium Niobate ($LiNbO_3$) is usually favored over quartz because it yields less insertion loss. The electrodes make up the filter's transducers. RF voltage is applied to the input transducer and generates electric fields between the fingers. The piezoelectric material vibrates, launching an acoustic wave along the surface. When the wave reaches the output transducer it produces an electric field between the fingers. This field generates a voltage across the load resistor.

Since both input and output transducers are not entirely unidirectional, some acous-

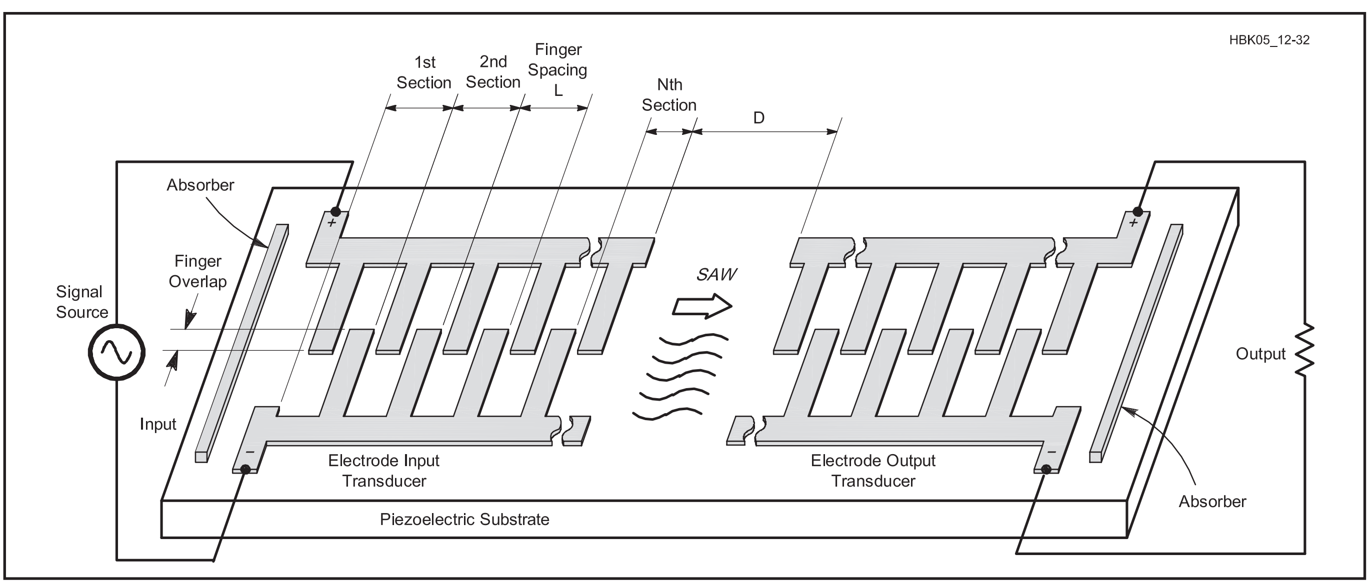


Fig 12.50 — The *interdigitated* transducer, on the left, launches SAW energy to a similar transducer on the right (see text).

tic power is lost in the acoustic absorbers located behind each transducer. This lost acoustic power produces a mid-band electrical insertion loss typically greater than 10 dB. The SAW filter frequency response is determined by the choice of substrate material and finger pattern. The finger spacing, (usually one-quarter wavelength) determines the filter center frequency. Center frequencies are available from 20 to 1000 MHz. The number and length of fingers determines the filter loaded Q and shape factor.

Loaded Qs are available from 2 to 100, with a shape factor of 1.5 (equivalent to a dozen poles). Thus the SAW filter can be made broadband much like the LC filters that it replaces. The advantage is substantially reduced volume and possibly lower cost. SAW filter research was driven by military needs for exotic amplitude-response and time-delay requirements. Low-cost SAW filters are presently found in television IF amplifiers where high mid-band loss can be tolerated.

Transmission-Line Filters

LC filter calculations are based on the assumption that the reactances are *lumped*—the physical dimensions of the components are considerably less than the operating wavelength. Therefore the unavoidable interturn capacitance associated with inductors and the unavoidable series inductance associated with capacitors are neglected as secondary effects. If careful attention is paid to circuit layout and miniature components are used, lumped LC filter technology can be used up to perhaps 1 GHz.

Transmission-line filters predominate from 500 MHz to 10 GHz. In addition they are often used down to 50 MHz when narrowband (Q_L > 10) band-pass filtering is required. In this application they exhibit considerably lower loss than their LC counterparts.

Replacing lumped reactances with selected short sections of TEM transmission lines results in transmission-line filters. In TEM, or *Transverse Electromagnetic Mode,* the electric and magnetic fields associated with a transmission line are at right angles (transverse) to the direction of wave propagation. Coaxial cable, stripline and microstrip are examples of TEM components. Waveguides and waveguide resonators are not TEM components.

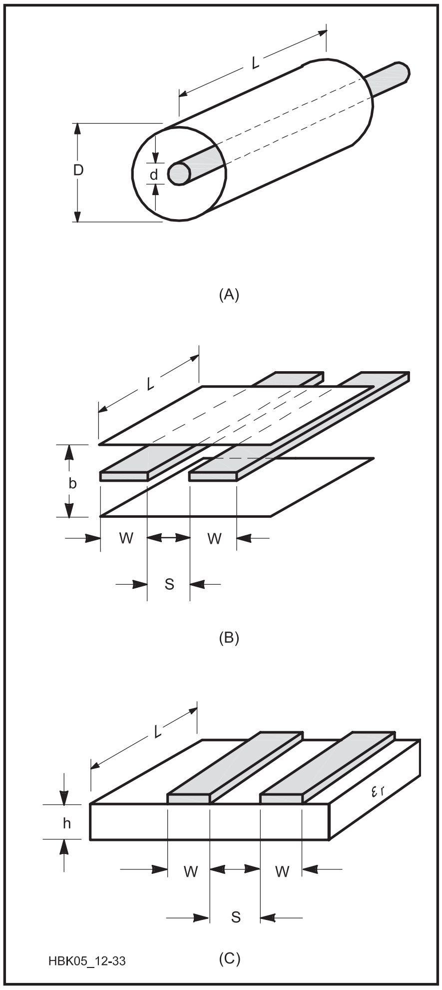


Fig 12.51 — Transmission lines. A: Coaxial line. B: Coupled stripline, which has two ground planes. C: Microstripline, which has only one ground plane.

TRANSMISSION LINES FOR FILTERS

Fig 12.51 shows three popular transmission lines used in transmission-line filters. The circular coaxial transmission line (coax) shown in Fig 12.51A consists of two concentric metal cylinders separated by dielectric (insulating) material. Coaxial transmission line possesses a characteristic impedance given by:

$$Z_0 = \frac{138}{\sqrt{\varepsilon}} \log\left(\frac{D}{d}\right) \tag{7}$$

A plot of Z_0 vs D/d is shown in **Fig 12.52**. At RF, Z_0 is an almost pure resistance. If the distant end of a section of coax is terminated in Z_0, then the impedance seen looking into the input end is also Z_0 at all frequencies. A terminated section of coax is shown in **Fig 12.53A**. If the distant end is not terminated

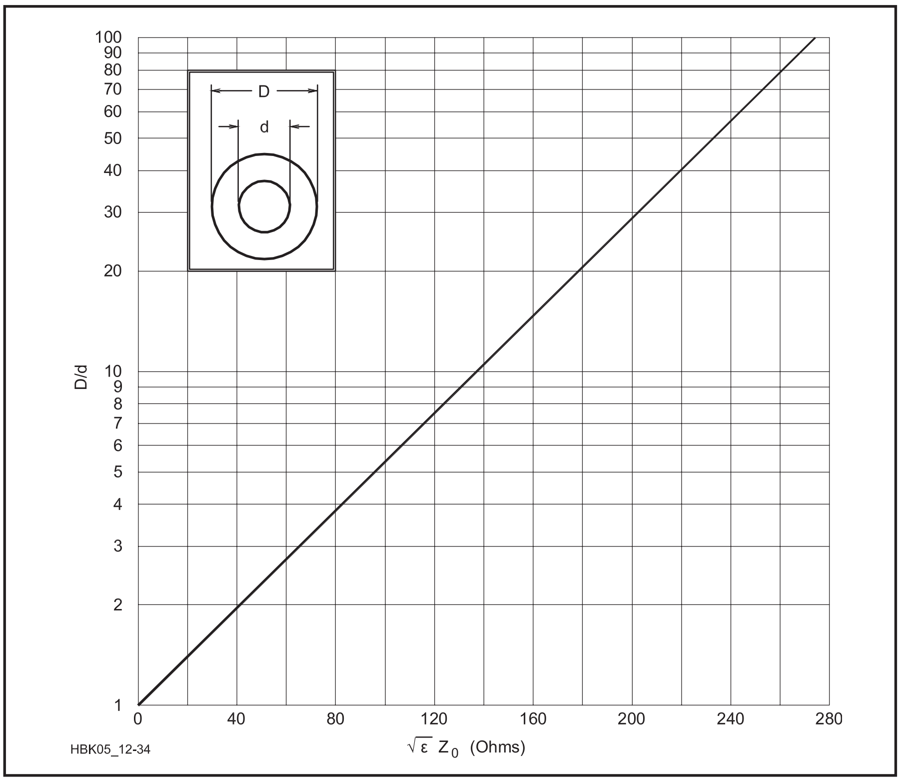


Fig 12.52 — Coaxial-line impedance varies with the ratio of the inner- and outer-conductor diameters. The dielectric constant, ε, is 1.0 for air and 2.32 for polyethylene.

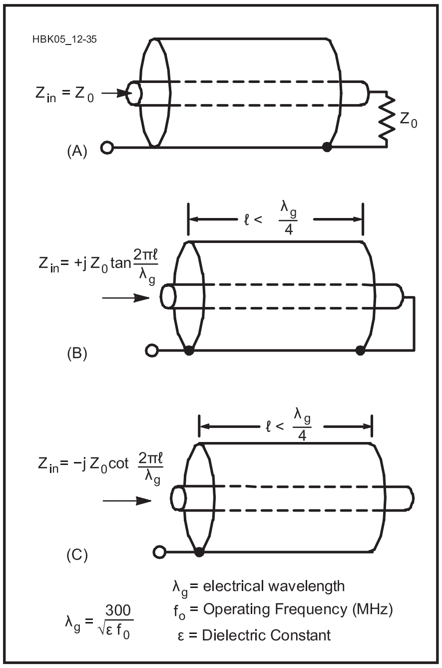


Fig 12.53 — Transmission line stubs. A: A line terminated in its characteristic impedance. B: A shorted line less than ¼-λ long is an *inductive* stub. C: An open line less than ¼-λ long is a *capacitive* stub.

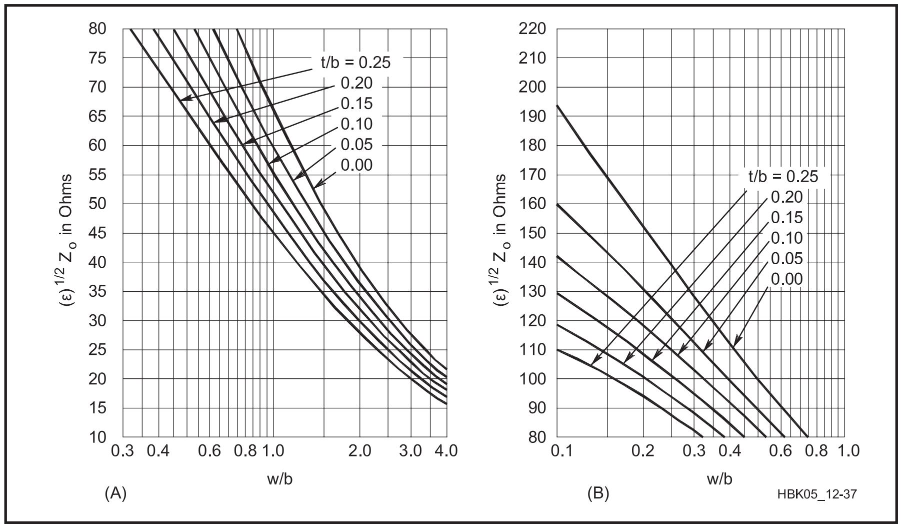


Fig 12.55 — The Z_0 of stripline varies with *w*, *b* and *t* (conductor thickness). See Fig 12.51B. The conductor thickness is *t* and the plots are normalized in terms of *t/b*.

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$\frac{\ell}{\lambda_g}$	Shorted Stub X_L Ω	Open Stub X_C Ω
0	0	∞
0.05	16.2	154
0.10	36.3	68.8
0.125	50	50
0.15	68.8	36.3
0.20	154	16.2
0.25	∞	0

Fig 12.54 — Stub reactance for various lengths of transmission line. Values are for Z_0 = 50 Ω. For Z_0 = 100 Ω, double the tabulated values.

Z_0 Ω	ε =1 (AIR) W/h	ε =2.3 (RT/Duroid) W/h	ε =2.3 (RT/Duroid) $\sqrt{\varepsilon_e}$	ε =4.5 (G−10) W/h	ε =4.5 (G−10) $\sqrt{\varepsilon_e}$
25	12.5	7.6	1.4	4.9	2.0
50	5.0	3.1	1.36	1.8	1.85
75	2.7	1.6	1.35	0.78	1.8
100	1.7	0.84	1.35	0.39	1.75
	$\sqrt{\varepsilon}$ =1				

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Fig 12.56 — Microstrip parameters (after H. Wheeler, *IEEE Transactions on MTT*, March 1965, p 132). ε_e is the effective ε.

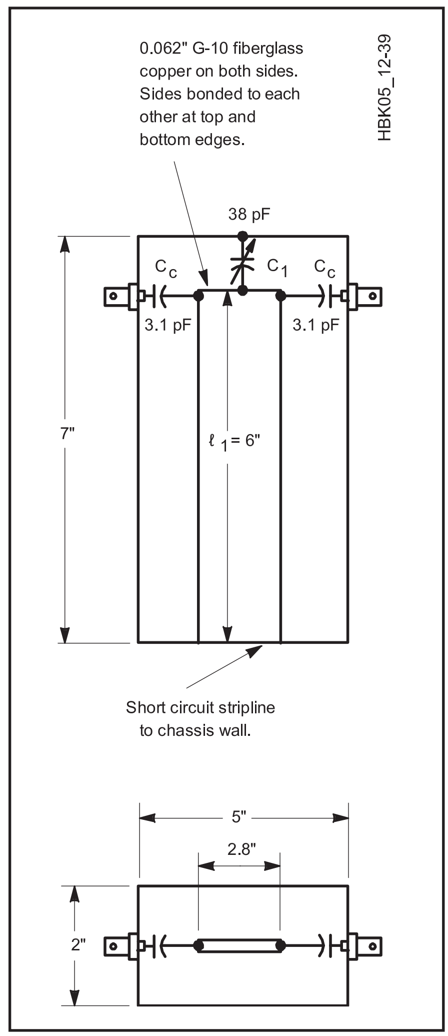


Fig 12.57 — This 146-MHz stripline band-pass filter has been measured to have a Q_L of 63 and a loss of approximately 1 dB.

in Z_0, the input impedance will be some other value. In Fig 12.53B the distant end is short-circuited and the length is less than ¼ λ. The input impedance is an inductive reactance as seen by the notation +*j* in the equation in part B of the figure.

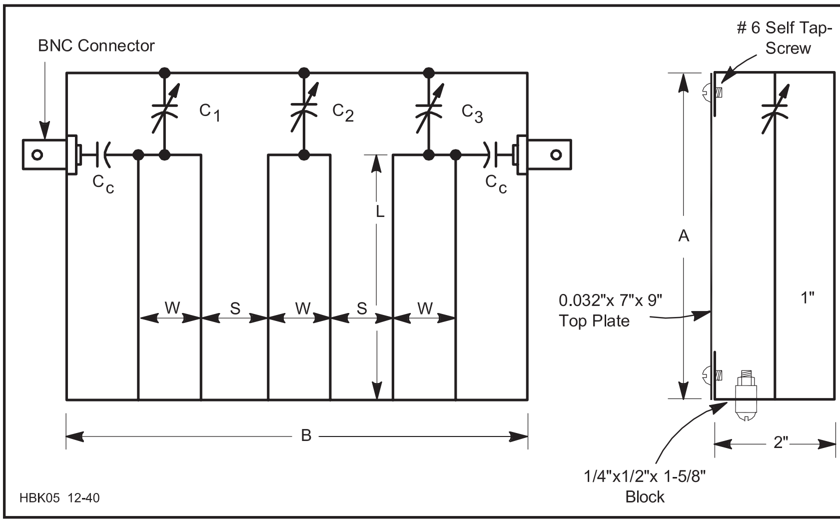
The input impedance for the case of the open-circuit distant end, is shown in Fig 12.53C. This case results in a capacitive reactance (–*j*). Thus, short sections of coaxial line (stubs) can replace the inductors and capacitors in an LC filter. Coax line inductive stubs usually have lower loss than their lumped counterparts.

X_L vs ℓ for shorted and open stubs is shown in **Fig 12.54**. There is an optimum value of Z_0 that yields lowest loss, given by

$$Z_0 = \frac{75}{\sqrt{\varepsilon}} \quad (8)$$

If the dielectric is air, Z_0 = 75 Ω. If the dielectric is polyethylene (ε = 2.32) Z_0 = 50 Ω. This is the reason why polyethylene dielectric flexible coaxial cable is usually manufactured with a 50-Ω characteristic impedance.

The first transmission-line filters were built from sections of coaxial line. Their mechanical fabrication is expensive and it is difficult to provide electrical coupling between line sections. Fabrication difficulties are reduced by the use of shielded strip transmission line



Dimension	*52 MHz*	*146 MHz*	*222 MHz*
A	**9"**	**7"**	**7"**
B	**7"**	**9"**	**9"**
L	**7⅜"**	**6"**	**6"**
S	**1"**	**1¹⁄₁₆"**	**1⅜"**
W	**1"**	**1⅝"**	**1⅝"**
Capacitance (pF)			
C1	**110**	**22**	**12**
C2	**135**	**30**	**15**
C3	**110**	**22**	**12**
C_c	**35**	**6.5**	**2.8**
Q_L	**10**	**29**	**36**
Performance			
BW3 (MHz)	**5.0**	**5.0**	**6.0**
Loss (dB)	**0.6**	**0.7**	**—**

Fig 12.58 — This Butterworth filter is constructed in combline. It was originally discussed by R. Fisher in December 1968 *QST*.

(stripline) shown in Fig 12.51B. The outer conductor of stripline consists of two flat parallel metal plates (ground planes) and the inner conductor is a thin metal strip. Sometimes the inner conductor is a round metal rod. The dielectric between ground planes and strip can be air or a low-loss plastic such as polyethylene. The outer conductors (ground planes or shields) are separated from each other by distance *b*.

Striplines can be easily coupled together by locating the strips near each other as shown in Fig 12.51B. Stripline Z_0 vs width (w) is plotted in **Fig 12.55**. Air-dielectric stripline technology is best for low bandwidth (Q_L > 20) band-pass filters.

The most popular transmission line is microstrip (unshielded stripline), shown in Fig 12.51C. It can be fabricated with standard printed-circuit processes and is the least expensive configuration. Unfortunately, microstrip is the lossiest of the three lines; therefore it is not suitable for narrow band-pass filters. In microstrip the outer conductor is a single flat metal ground-plane. The inner conductor is a thin metal strip separated from the ground-plane by a solid dielectric substrate. Typical substrates are 0.062-inch G-10 fiberglass ($\varepsilon = 4.5$) for the 50- MHz to 1-GHz frequency range and 0.031-inch Teflon ($\varepsilon = 2.3$) for frequencies above 1 GHz.

Conductor separation must be minimized or free-space radiation and unwanted coupling to adjacent circuits may become problems. Microstrip characteristic impedance and the effective dielectric constant (ε) are shown in **Fig 12.56**. Unlike coax and stripline, the effective dielectric constant is less than that of the substrate since a portion of the electromagnetic wave propagating along the microstrip "sees" the air above the substrate.

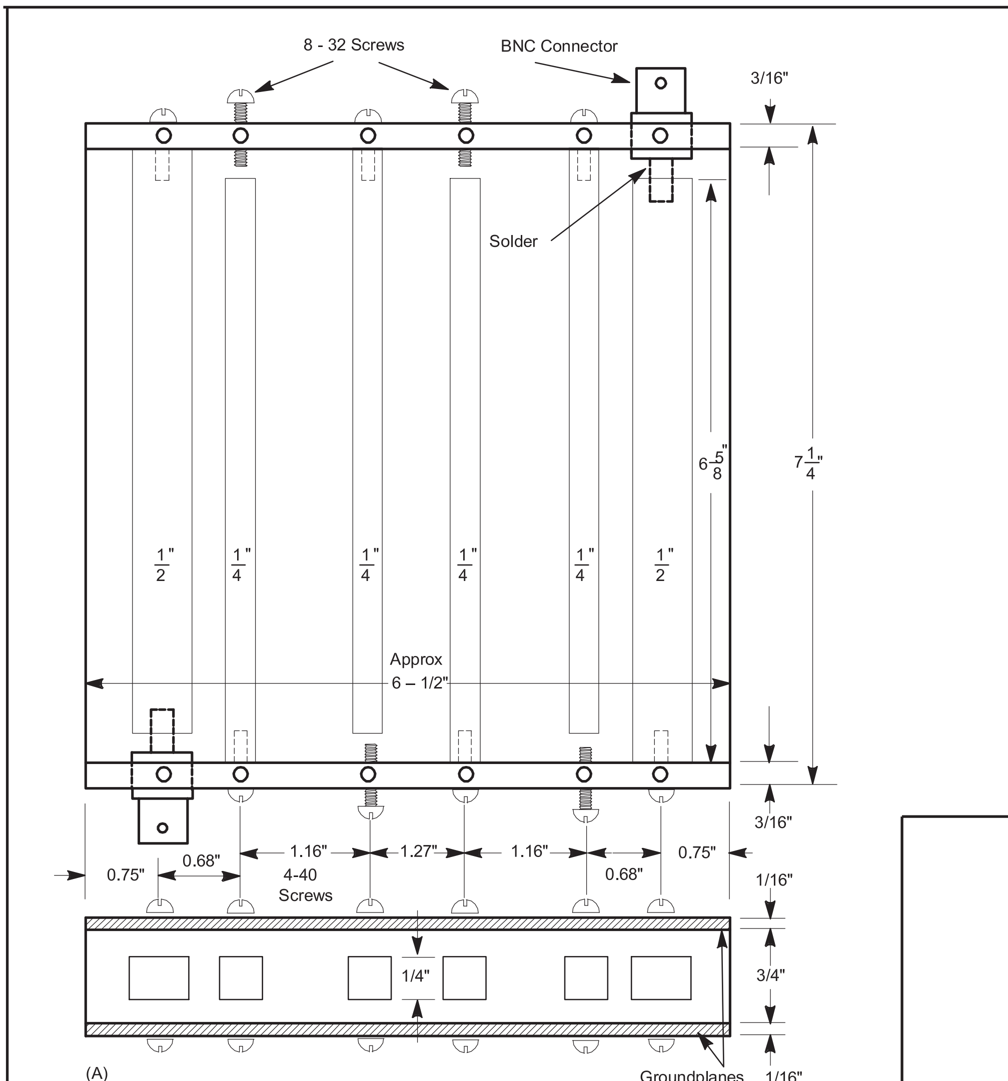


Fig 12.59 — These 3-pole Butterworth filters (upper: 432 MHz, 8.6 MHz bandwidth, 1.4 dB pass-band loss; lower: 1296 MHz, 110 MHz bandwidth, 0.4 dB pass-band loss) are constructed as interdigitated filters. The material is from R. E. Fisher, March 1968 *QST*.

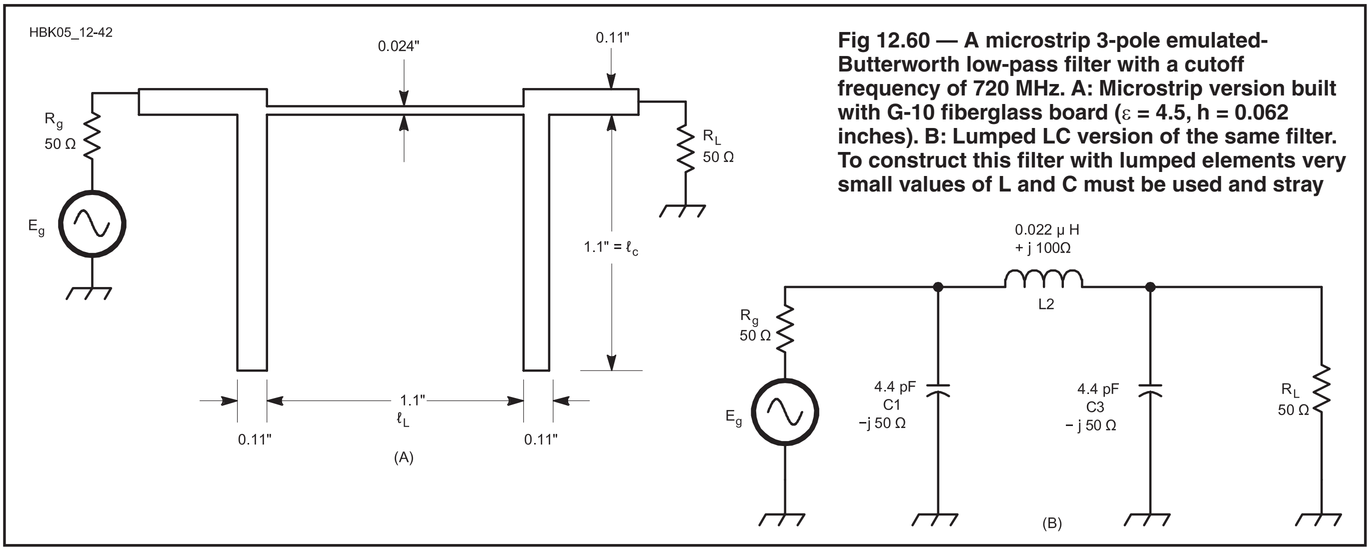
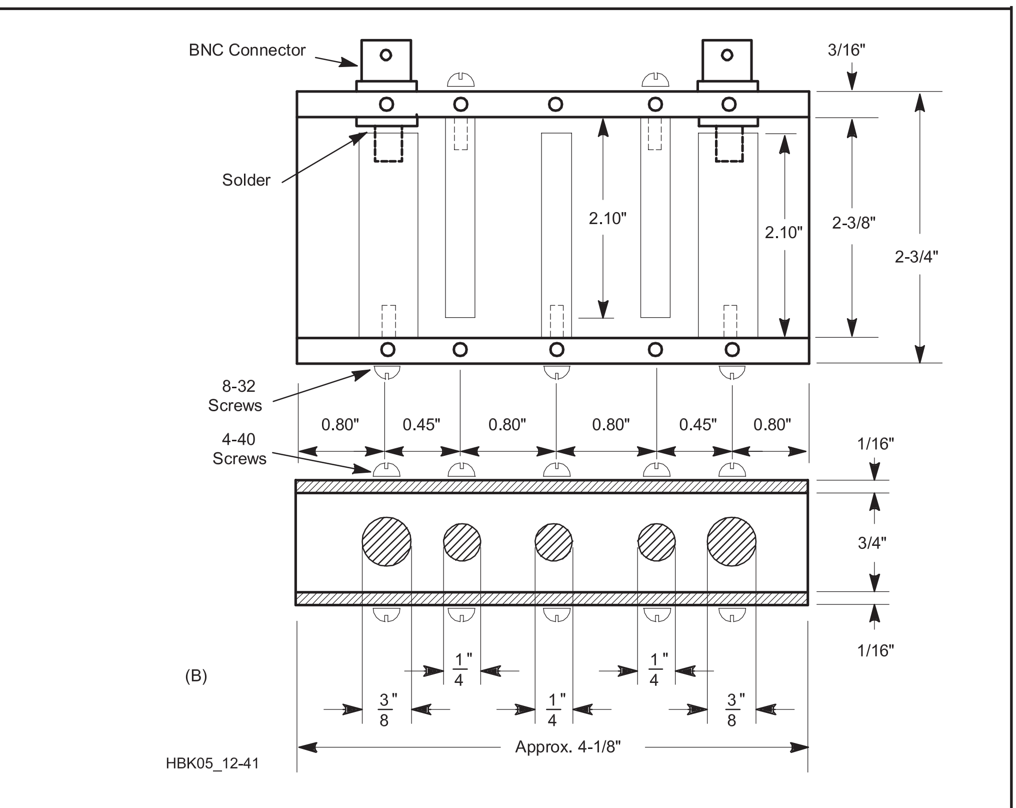


Fig 12.60 — A microstrip 3-pole emulated-Butterworth low-pass filter with a cutoff frequency of 720 MHz. A: Microstrip version built with G-10 fiberglass board (ε = 4.5, h = 0.062 inches). B: Lumped LC version of the same filter. To construct this filter with lumped elements very small values of L and C must be used and stray



The least-loss characteristic impedance for stripline and microstrip-lines is not 75 Ω as it is for coax. Loss decreases as line width increases, which leads to clumsy, large structures. Therefore, to conserve space, filter sections are often constructed from 50-Ω stripline or microstrip stubs.

TRANSMISSION-LINE BAND-PASS FILTERS

Band-pass filters can also be constructed from transmission-line stubs. At VHF the stubs can be considerably shorter than a quarter wavelength yielding a compact filter structure with less mid-band loss than its LC counterpart. The single-stage 146-MHz stripline band-pass filter shown in **Fig 12.57** is an example. This filter consists of a single inductive 50-Ω strip-line stub mounted into a 2 × 5 × 7-inch aluminum box. The stub is resonated at 146 MHz with the "APC" variable capacitor, C1. Coupling to the 50-Ω generator and load is provided by the coupling capacitors C_c. The measured performance of this filter is: f_o = 146 MHz, BW = 2.3 MHz (Q_L = 63) and mid-band loss = 1 dB.

Single-stage stripline filters can be coupled together to yield multistage filters. One method uses the capacitor coupled band-pass filter synthesis technique to design a 3-pole filter. Another method allows closely spaced inductive stubs to magnetically couple to each other. When the coupled stubs are grounded on the same side of the filter housing, the structure is called a "combline filter." Three examples of combline band-pass filters are shown in **Fig 12.58**. These filters are constructed in 2 × 7 × 9-inch chassis boxes.

QUARTER-WAVE TRANSMISSION-LINE FILTERS

Fig 12.56 shows that when $\ell = 0.25\ \lambda_g$, the shorted-stub reactance becomes infinite. Thus, a ¼-λ shorted stub behaves like a parallel-resonant LC circuit. Proper input and output coupling to a ¼-λ resonator yields a practical band-pass filter. Closely spaced ¼-λ resonators will couple together to form a multistage band-pass filter. When the resonators are grounded on opposite walls of the filter housing, the structure is called an "interdigital filter" because the resonators look like interlaced fingers. Two examples of 3-pole UHF interdigital filters are shown in **Fig 12.59**. Design graphs for round-rod interdigital filters are given in Ref 16. The ¼-λ resonators may be tuned by physically changing their lengths or by tuning the screw opposite each rod.

If the short-circuited ends of two ¼-λ resonators are connected to each other, the resulting ½-λ stub will remain in resonance, even when the connection to ground-plane is removed. Such a floating ½-λ microstrip line, when bent into a U-shape, is called a “hairpin” resonator. Closely coupled hairpin resonators can be arranged to form multistage band-pass filters. Microstrip hairpin band-pass filters are popular above 1 GHz because they can be easily fabricated using photo-etching techniques. No connection to the ground-plane is required.

TRANSMISSION-LINE FILTERS EMULATING LC FILTERS

Low-pass and high-pass transmission-line filters are usually built from short sections of transmission lines (stubs) that emulate lumped LC reactances. Sometimes low-loss lumped capacitors are mixed with transmission-line inductors to form a hybrid component filter. For example, consider the 720-MHz, 3-pole microstrip low-pass filter shown in **Fig 12.60A** that emulates the LC filter shown in Fig 12.60B. C1 and C3 are replaced with 50-Ω open-circuit shunt stubs ℓ_C long. L2 is replaced with a short section of 100-Ω line ℓ_L long. The LC filter, Fig 12.60B, was designed for f_c = 720 MHz. Such a filter could be connected between a 432-MHz transmitter and antenna to reduce harmonic and spurious emissions. A reactance chart shows that X_C is 50 Ω, and the inductor reactance is 100 Ω at f_c. The microstrip version is constructed on G-10 fiberglass 0.062-inch thick, with ε = 4.5. Then, from Fig 12.56, *w* is 0.11 inch and $\ell_C = 0.125\ \lambda_g$ for the 50-Ω capacitive stubs. Also, from Fig 12.56, *w* is 0.024 inch and ℓ_L is 0.125 ℓ_g for the 100-Ω inductive line. The inductive line length is approximate because the far end is not a short circuit. ℓ_g is 300/(720 × 1.75) = 0.238 m, or 9.37 inches. Thus ℓ_C is 1.1 inch and ℓ_L is 1.1 inch.

This microstrip filter exhibits about 20 dB of attenuation at 1296 MHz. Its response rises again, however, around 3 GHz. This is because the fixed-length transmission-line stubs change in terms of wavelength as the frequency rises. This particular filter was designed to eliminate third-harmonic energy near 1296 MHz from a 432-MHz transmitter and does a better job in this application than the Butterworth filter in Fig 12.59 which has spurious responses in the 1296-MHz band.

Helical Resonators

Ever-increasing occupancy of the radio spectrum brings with it a parade of receiver overload and spurious responses. Overload problems can be minimized by using high-dynamic-range receiving techniques, but spurious responses (such as the image frequency) must be filtered out before mixing occurs. Conventional tuned circuits cannot provide the selectivity necessary to eliminate the plethora of signals found in most urban and many suburban neighborhoods. Other filtering techniques must be used.

Helical resonators are usually a better choice than ¼-λ cavities on 50, 144 and 222 MHz to eliminate these unwanted inputs. They are smaller and easier to build. In the frequency range from 30 to 100 MHz it is difficult to build high-Q inductors, and coaxial cavities are very large. In this frequency range the helical resonator is an excellent choice. At 50 MHz for example, a capacitively tuned,¼-λ coaxial cavity with an unloaded Q of 3000 would be about 4 inches in diameter and nearly 5 ft long. On the other hand, a helical resonator with the same unloaded Q is about 8.5 inches in diameter and 11.3 inches long. Even at 432 MHz, where coaxial cavities are common, the use of helical resonators results in substantial size reductions.

> **Helical Filter Design Software**
>
> The *ARRL Handbook CD* at the back of this book includes *Helical*, a *Windows* program for design and analysis of helical-resonator filters by Jim Tonne, WB6BLD. The software is interactive and plots performance changes as various parameters are altered. See the “About the Included CD-ROM” section for more information.

The helical resonator was described by the late Jim Fisk, W1HR, in a *QST* article as a coil surrounded by a shield, but it is actually a shielded, resonant section of helically wound transmission line with relatively high characteristic impedance and low axial propagation velocity. The electrical length is about 94% of an axial ¼-λ or 84.6°. One lead of the helical winding is connected directly to the shield and the other end is open circuited as shown in **Fig 12.61**. Although the shield may be any shape, only round and square shields will be considered here.

DESIGN

The unloaded Q of a helical resonator is determined primarily by the size of the shield. For a round resonator with a copper coil on a low-loss form, mounted in a copper shield, the unloaded Q is given by

$$Q_U = 50\,D\sqrt{f_0} \qquad (9)$$

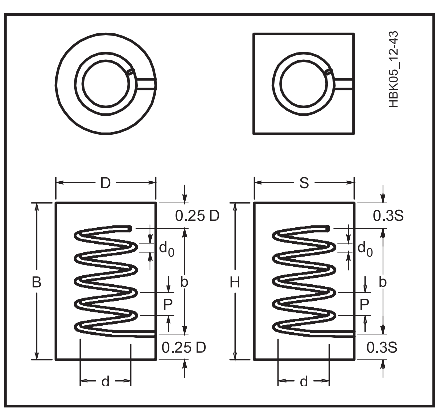


Fig 12.61 — Dimensions of round and square helical resonators. The diameter, D (or side, S) is determined by the desired unloaded Q. Other dimensions are expressed in terms of D or S (see text).

where

D = inside diameter of the shield, in inches

f_o = frequency, in MHz.

D is assumed to be 1.2 times the width of one side for square shield cans. This formula includes the effects of losses and imperfections in practical materials. It yields values of unloaded Q that are easily attained in practice. Silver plating the shield and coil increases the unloaded Q by about 3% over that predicted by the equation. At VHF and UHF, however, it is more practical to increase the shield size slightly (that is, increase the selected Q_U by about 3% before making the calculation). The fringing capacitance at the open-circuit end of the helix is about 0.15 D pF (that is, approximately 0.3 pF for a shield 2 inches in diameter). Once the required shield size has been determined, the total number of turns, N, winding pitch, P and characteristic impedance, Z_0, for round and square helical resonators with air dielectric between the helix and shield, are given by:

$$N = \frac{1908}{f_0 D} \qquad (10A)$$

$$P = \frac{f_0 D^2}{2312} \qquad (10B)$$

$$Z_0 = \frac{99{,}000}{f_0 D} \qquad (10C)$$

$$N = \frac{1590}{f_0 S} \qquad (10D)$$

$$P = \frac{f_0 S^2}{1606} \qquad (10E)$$

$$Z_0 = \frac{82{,}500}{f_0 S} \qquad (10F)$$

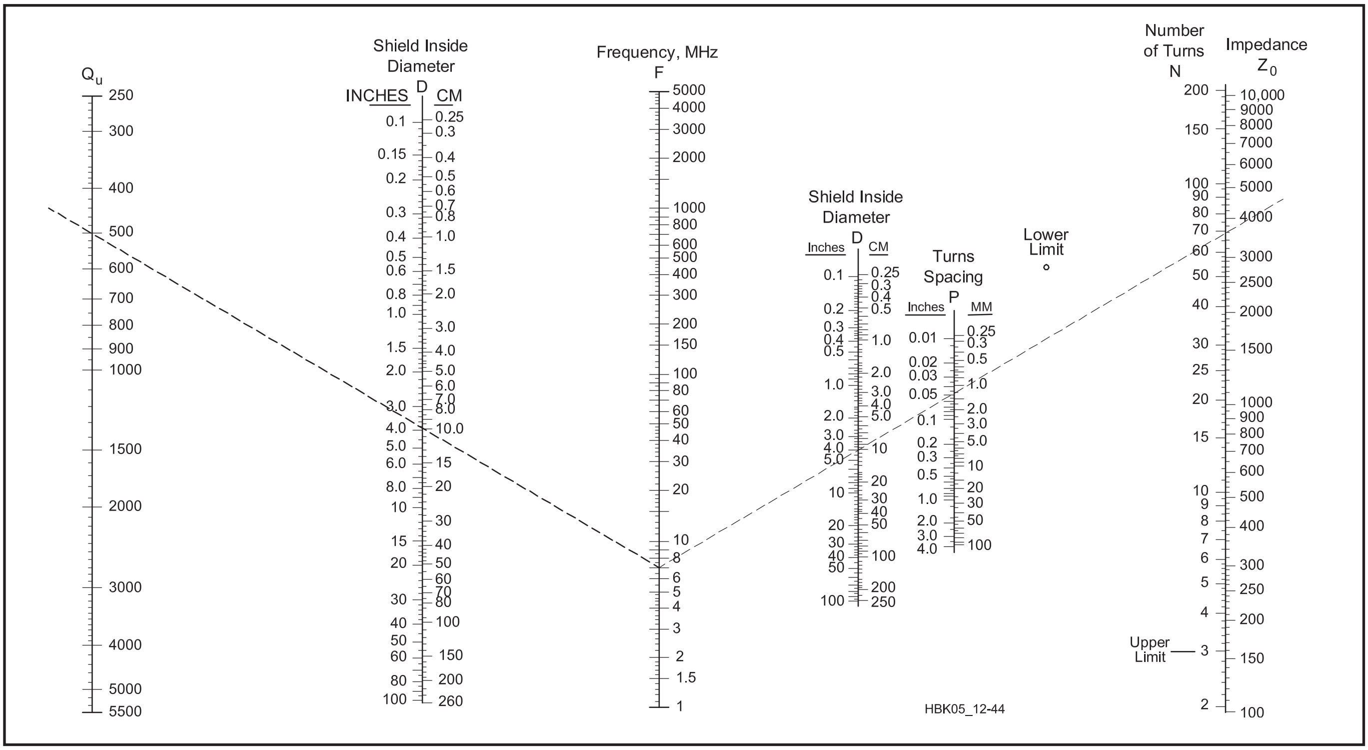


Fig 12.62 — The design nomograph for round helical resonators starts by selecting Q_U and the required shield diameter. A line is drawn connecting these two values and extended to the frequency scale (example here is for a shield of about 3.8 inches and Q_U of 500 at 7 MHz). Finally the number of turns, N, winding pitch, P, and characteristic impedance, Z_0, are determined by drawing a line from the frequency scale through selected shield diameter (but this time to the scale on the right-hand side. For the example shown, the dashed line shows P ≈ 0.047 inch, N = 70 turns, and Z_n = 3600 Ω).

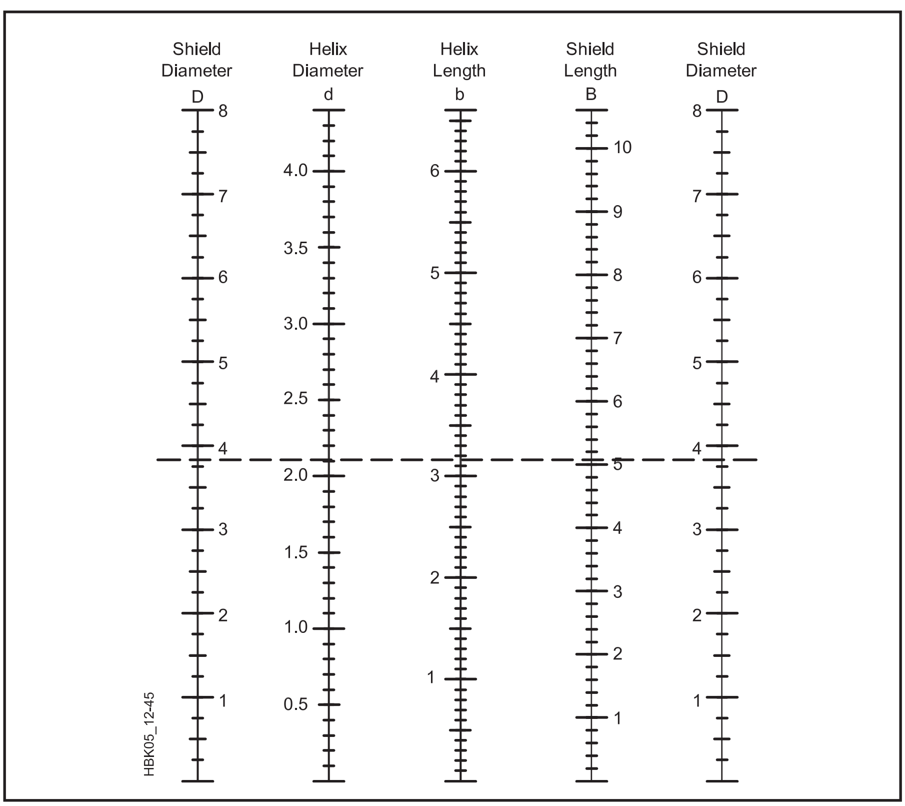


Fig 12.63 — The helical resonator is scaled from this design nomograph. Starting with the shield diameter, the helix diameter, d, helix length, b, and shield length, B, can be determined with this graph. The example shown has a shield diameter of 3.8 inches. This requires a helix mean diameter of 2.1 inches, helix length of 3.1 inches, and shield length of 5 inches.

In these equations, dimensions D and S are in inches and f_0 is in megahertz. The design nomograph for round helical resonators in **Fig 12.62** is based on these formulas.

Although there are many variables to consider when designing helical resonators, certain ratios of shield size to length and coil diameter to length, provide optimum results. For helix diameter, d = 0.55 D or d = 0.66 S. For helix length, b = 0.825D or b = 0.99S. For shield length, B = 1.325 D and H = 1.60 S.

Fig 12.63 simplifies calculation of these dimensions. Note that these ratios result in a helix with a length 1.5 times its diameter, the condition for maximum Q. The shield is about 60% longer than the helix—although it can be made longer—to completely contain the electric field at the top of the helix and the magnetic field at the bottom.

The winding pitch, P, is used primarily to determine the required conductor size. Adjust the length of the coil to that given by the equations during construction. Conductor size ranges from 0.4 P to 0.6 P for both round and square resonators and are plotted graphically in **Fig 12.64**.

Obviously, an area exists (in terms of frequency and unloaded Q) where the designer must make a choice between a conventional cavity (or lumped LC circuit) and a helical resonator. The choice is affected by physical shape at higher frequencies. Cavities are

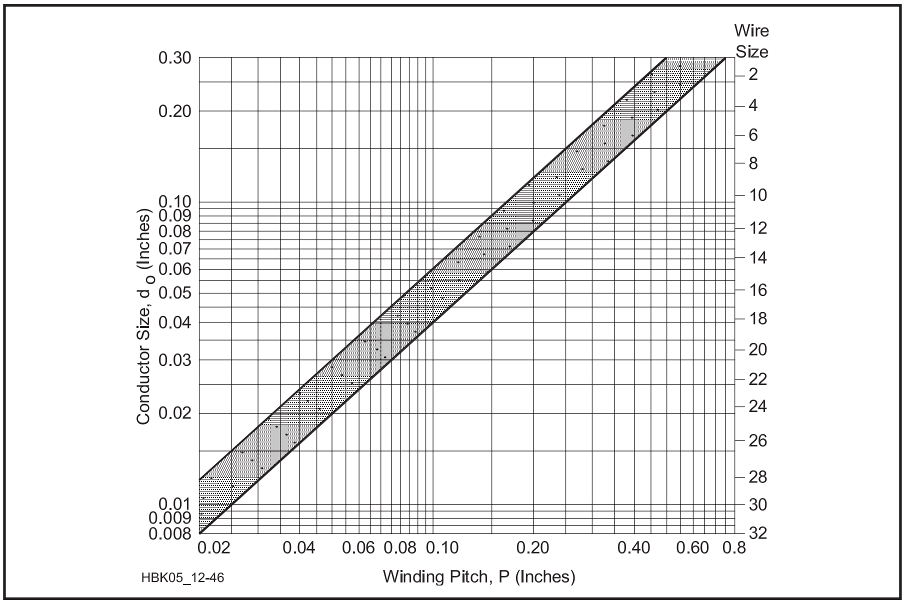


Fig 12.64 — This chart provides the design information of helix conductor size vs winding pitch, P. For example, a winding pitch of 0.047 inch results in a conductor diameter between 0.019 and 0.028 inch (#22 or #24 AWG).

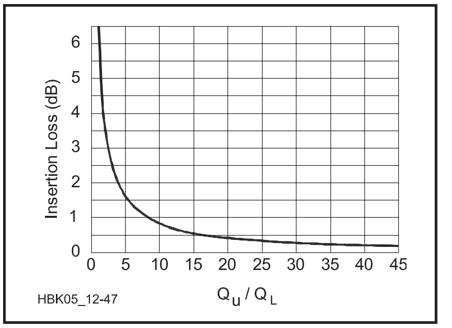


Fig 12.65 — The ratio of loaded (Q_L) to unloaded (Q_U) Q determines the insertion loss of a tuned resonant circuit.

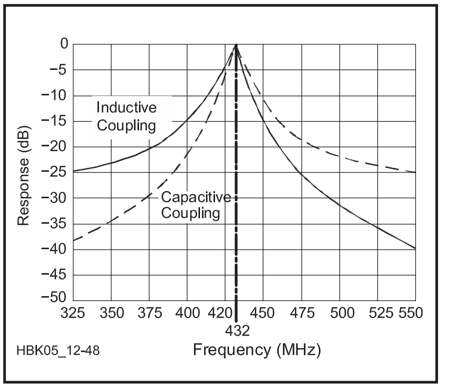


Fig 12.66 — This response curve for a single-resonator 432-MHz filter shows the effects of capacitive and inductive input/output coupling. The response curve can be made symmetrical on each side of resonance by combining the two methods (inductive input and capacitive output, or vice versa).

long and relatively small in diameter, while the length of a helical resonator is not much greater than its diameter. A second consideration is that point where the winding pitch, P, is less than the radius of the helix (otherwise the structure tends to be nonhelical). This condition occurs when the helix has fewer than three turns (the "upper limit" on the design nomograph of Fig 12.62).

CONSTRUCTION

The shield should not have any seams parallel to the helix axis to obtain as high an unloaded Q as possible. This is usually not a problem with round resonators because large-diameter copper tubing is used for the shield, but square resonators require at least one seam and usually more. The effect on unloaded Q is minimum if the seam is silver soldered carefully from one end to the other.

Results are best when little or no dielectric is used inside the shield. This is usually no problem at VHF and UHF because the conductors are large enough that a supporting coil form is not required. The lower end of the helix should be soldered to the nearest point on the inside of the shield.

Although the external field is minimized by the use of top and bottom shield covers, the top and bottom of the shield may be left open with negligible effect on frequency or unloaded Q. Covers, if provided, should make electrical contact with the shield. In those resonators where the helix is connected to the bottom cover, that cover must be soldered solidly to the shield to minimize losses.

TUNING

A carefully built helical resonator designed from the nomograph of Fig 12.62 will resonate very close to the design frequency. Slightly compress or expand the helix to adjust resonance over a small range. If the helix is made slightly longer than that called for in Fig 12.63, the resonator can be tuned by pruning the open end of the coil. However, neither of these methods is recommended for wide frequency excursions because any major deviation in helix length will degrade the unloaded Q of the resonator.

Most helical resonators are tuned by means of a brass tuning screw or high-quality air-variable capacitor across the open end of the helix. Piston capacitors also work well, but the Q of the tuning capacitor should ideally be several times the unloaded Q of the resonator. Varactor diodes have sometimes been used where remote tuning is required, but varactors can generate unwanted harmonics and other spurious signals if they are excited by strong, nearby signals.

When a helical resonator is to be tuned by a variable capacitor, the shield size is based on the chosen unloaded Q at the operating frequency. Then the number of turns, N *and* the winding pitch, P, are based on resonance at 1.5 f_0. Tune the resonator to the desired operating frequency, f_0.

INSERTION LOSS

The insertion loss (dissipation loss), I_L, in decibels, of all single-resonator circuits is given by

$$I_L = 20 \log_{10} \left(\frac{1}{1 - \frac{Q_L}{Q_U}} \right) \quad (11)$$

where
Q_L = loaded Q
Q_U = unloaded Q

This is plotted in **Fig 12.65**. For the most practical cases ($Q_L > 5$), this can be closely approximated by $I_L \approx 9.0\ (Q_L/Q_U)$ dB. The selection of Q_L for a tuned circuit is dictated primarily by the required selectivity of the circuit. However, to keep dissipation loss to 0.5 dB or less (as is the case for low-noise VHF receivers), the unloaded Q must be at least 18 times the Q_L.

COUPLING

Signals are coupled into and out of helical resonators with inductive loops at the bottom of the helix, direct taps on the coil or a combination of both. Although the correct tap point can be calculated easily, coupling by loops and probes must be determined experimentally.

The input and output coupling is often provided by probes when only one resonator is used. The probes are positioned on opposite sides of the resonator for maximum isolation. When coupling loops are used, the plane of the loop should be perpendicular to the axis of the helix and separated a small distance from the bottom of the coil. For resonators with only a few turns, the plane of the loop can be tilted slightly so it is parallel with the slope of the adjacent conductor.

Helical resonators with inductive coupling (loops) exhibit more attenuation to signals above the resonant frequency (as compared to attenuation below resonance), whereas resonators with capacitive coupling (probes) exhibit more attenuation below the passband, as shown for a typical 432-MHz resonator in **Fig 12.66**. Consider this characteristic when choosing a coupling method. The passband can be made more symmetrical by using a combination of coupling methods (inductive input and capacitive output, for example).

If more than one helical resonator is required to obtain a desired band-pass characteristic, adjacent resonators may be coupled through apertures in the shield wall between the two resonators. Unfortunately, the size and location of the aperture must be found empirically, so this method of coupling is not very practical unless you're building a large number of identical units.

Since the loaded Q of a resonator is determined by the external loading, this must be considered when selecting a tap (or position of a loop or probe). The ratio of this external loading, R_b, to the characteristic impedance, Z_0, for a ¼-λ resonator is calculated from:

$$K = \frac{R_b}{Z_0} = 0.785\left(\frac{1}{Q_L} - \frac{1}{Q_U}\right) \quad (12)$$

Even when filters are designed and built properly, they may be rendered totally ineffective if not installed properly. Leakage around a filter can be quite high at VHF and UHF, where wavelengths are short. Proper attention to shielding and good grounding is mandatory for minimum leakage. Poor coaxial cable shield connection into and out of the filter is one of the greatest offenders with regard to filter leakage. Proper dc-lead bypassing throughout the receiving system is good practice, especially at VHF and above. Ferrite beads placed over the dc leads may help to reduce leakage. Proper filter termination is required to minimize loss.

Most VHF RF amplifiers optimized for noise figure do not have a 50-Ω input impedance. As a result, any filter attached to the input of an RF amplifier optimized for noise figure will not be properly terminated and filter loss may rise substantially. As this loss is directly added to the RF amplifier noise figure, carefully choose and place filters in the receiver.

Active Filters

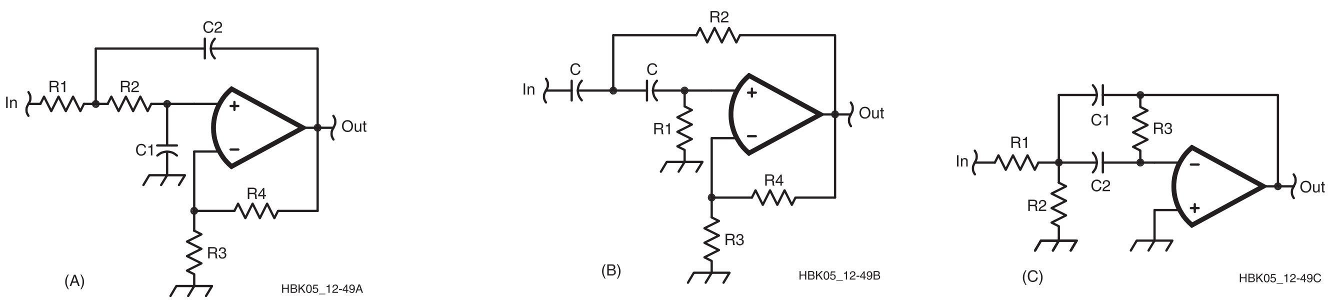
Passive HF filters are made from combinations of inductors and capacitors. These may be used at low frequencies, but the inductors often become a limiting factor because of their size, weight, cost and losses. The active filter is a compact, low-cost alternative made with op amps, resistors and capacitors. They often occupy a fraction of the space required

Table 12.22
Factor "a" for Low- and High-Pass Filters in Fig 12.67

No. of Stages	*Stage 1*	*Stage 2*	*Stage 3*	*Stage 4*
1	1.414	—	—	—
2	0.765	1.848	—	—
3	0.518	1.414	1.932	—
4	0.390	1.111	1.663	1.962

These values are truncated from those of Appendix C of Ref 21, for even-order Butterworth filters.

Unless otherwise specified, values of R are in ohms, C is in farads, F in hertz and ω in radians per second. Calculations shown here were performed on a scientific calculator.



Low-Pass Filter

$$C_1 \le \frac{\left[a^2 + 4(K-1)\right]C_2}{4}$$

$$R_1 = \frac{2}{\left[aC_2 + \sqrt{\left[a^2 + 4(K-1)\right]C_2^{\,2} - 4C_1C_2}\,\right]\omega_C}$$

$$R_2 = \frac{1}{C_1C_2R_1\omega_C^{\,2}}$$

$$R_3 = \frac{K(R_1+R_2)}{K-1} \quad (K>1)$$

$$R_4 = K(R_1+R_2)$$

where
K = gain
f_c = –3 dB cutoff frequency
$\omega_c = 2\pi f_c$
C_2 = a standard value near $10/f_c$ (in µF)

Note: For unity gain, short R4 and omit R3.

Example:
a = 1.414 (see table, one stage)
K = 2
f = 2700 Hz
ω_c = 16,964.6 rads/sec
C_2 = 0.0033 µF
C1 ≤ 0.00495 µF (use 0.0050 µF)
R1 ≤ 25,265.2 Ω (use 24 kΩ)
R2 = 8,420.1 Ω (use 8.2 kΩ)
R3 = 67,370.6 Ω (use 68 kΩ)
R4 = 67,370.6 Ω (use 68 kΩ)

High-Pass Filter

$$R_1 = \frac{4}{\left[a + \sqrt{a^2 + 8(K-1)}\,\right]\omega_C\, C}$$

$$R_2 = \frac{1}{\omega_C^{\,2}C^2R_1}$$

$$R_3 = \frac{KR_1}{K-1} \quad (K>1)$$

$$R_4 = KR_1$$

where
K = gain
f_c = –3 dB cutoff frequency
$\omega_c = 2\pi f_c$
C = a standard value near $10/f_c$ (in µF)

Note: For unity gain, short R4 and omit R3.

Example:
a = 0.765 (see table, first of two stages)
K = 4
f = 250 Hz
ω_c = 1570.8
C = 0.04 µF (use 0.039 µF)
R1 = 11,123.2 Ω (use 11 kΩ)
R2 = 22,722 Ω (use 22 kΩ)
R3 = 14,830.9 Ω (use 15 kΩ)
R4 = 44,492.8 Ω (use 47 kΩ)

Band-Pass Filter

Pick K, Q, $\omega_o = 2\pi f_c$
where f_c = center freq.
Choose C
Then

$$R1 = \frac{Q}{K_0\omega_0 C}$$

$$R2 = \frac{Q}{\left(2Q^2 - K_0\right)\omega_0 C}$$

$$R3 = \frac{2Q}{\omega_0 C}$$

Example:
K = 2, f_o = 800 Hz, Q = 5 and C = 0.022 µF
R1 = 22.6 kΩ (use 22 kΩ)
R2 = 942 Ω (use 910 Ω)
R3 = 90.4 kΩ (use 91 kΩ)

Fig 12.67 — Equations for designing a low-pass RC active audio filter are given at A. B, C and D show design information for high-pass, band-pass and band-reject filters, respectively. All of these filters will exhibit a Butterworth response. Values of K and Q should be less than 10. See Table 12.22 for values of "a".

by an LC filter. While active filters have been traditionally used at low and audio frequencies, modern op amps with small-signal bandwidths that exceed 1 GHz have extended their range into MF and HF.

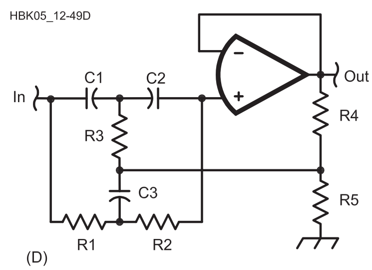
Active filters can perform any common filter function: low pass, high pass, band-pass, band reject and all pass (used for phase or time delay). Responses such as Butterworth, Chebyshev, Bessel and elliptic can be realized. Active filters can be designed for gain, and they offer excellent stage-to-stage isolation.

Despite the advantages, there are also some limitations. They require power, and performance may be limited by the op amp's finite input and output levels, gain and bandwidth. While LC filters can be designed for high-power applications, active filters usually are not.

The design equations for various filters are shown in **Fig 12.67**. **Fig 12.68** shows a typical application of a two-stage, band-pass filter. A two-stage filter is considered the minimum acceptable for CW, while three or four stages will prove more effective under some conditions of noise and interference.

SWITCHED CAPACITOR FILTERS

The *switched capacitor filter,* or SCF, uses an IC to synthesize a high-pass, low-pass, band-pass or notch filter. The performance of multiple-pole filters is available, with Q and bandwidth set by external resistors. An external clock frequency sets the filter center frequency, so this frequency may be easily changed or digitally controlled. Dynamic range of 80 dB, Q of 50, 5-pole equivalent design and maximum usable frequency of 250 kHz are available for such uses as audio CW and RTTY filters. In addition, all kinds of digital tone signaling such as DTMF and modem encoding and decoding are being designed with these circuits.



Band-Reject Filter

$$F_0 = \frac{1}{2\pi R1C1}$$

$$K = 1 - \frac{1}{4Q}$$

$$R >> (1-K)R1$$

where

$$C1 = C2 = \frac{C3}{2} = \frac{10\,\mu F}{f_0}$$

R1 = R2 = 2R3
R4 = (1 – K)R
R5 = K × R

Example:
f_0 = 500 Hz, Q=10
K = 0.975
C1 = C2 = 0.02 µF (or use 0.022 µF)
C3 = 0.04 µF (or use 0.044 µF)
R1 = R2 = 15.92 kΩ (use 15 kΩ)
R3 = 7.96 kΩ (use 8.2 kΩ)
R >> 398 Ω (390 Ω)
R4 = 9.95 Ω (use 10 Ω)
R5 = 388.1 Ω (use 390 Ω)

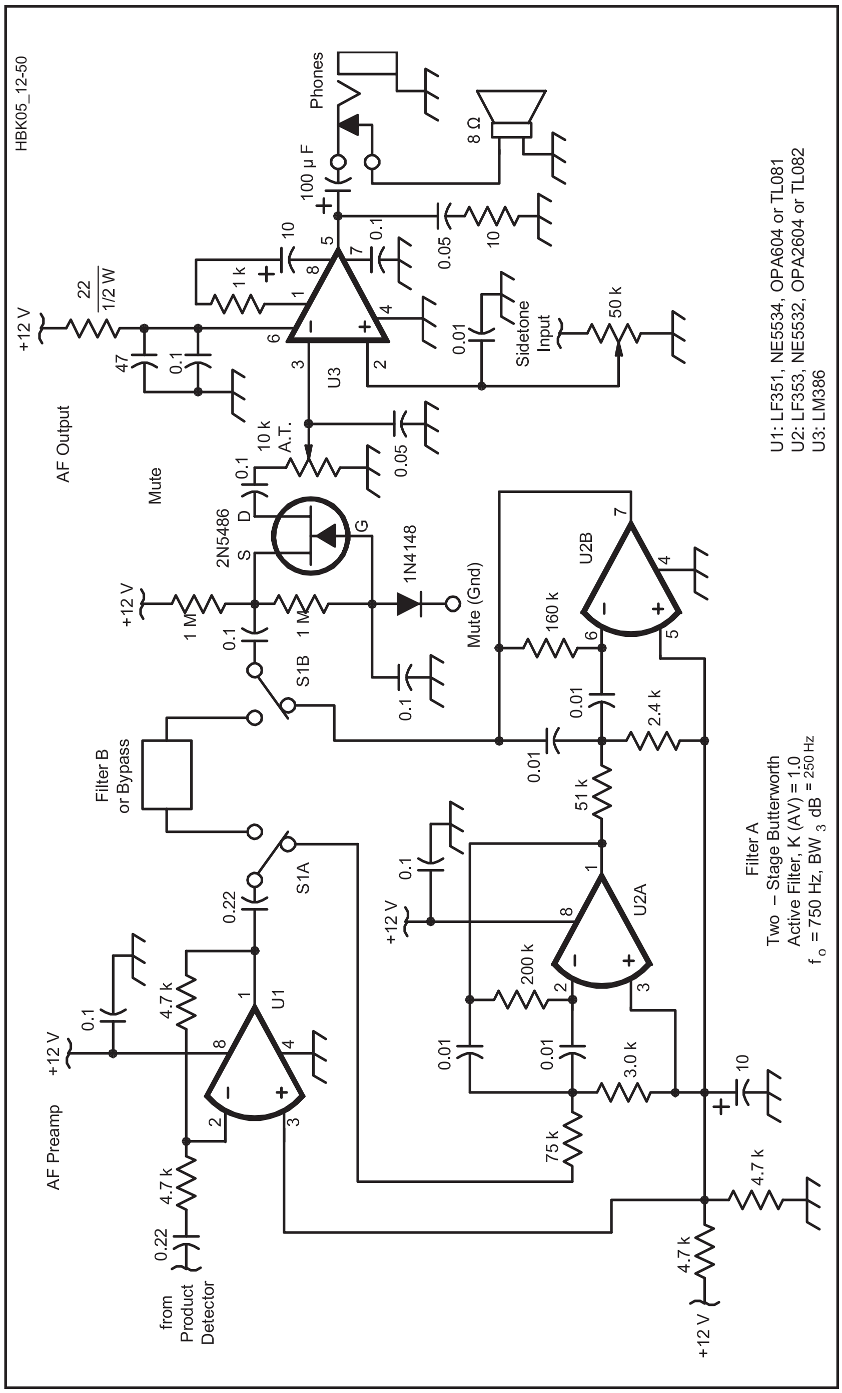


Fig 12.68 — Typical application of a two-stage active filter in the audio chain of a QRP CW transceiver. The filter can be bypassed, or another filter can be switched in by S1.

A BROADCAST-BAND REJECTION FILTER

Inadequate front-end selectivity or poorly performing RF amplifier and mixer stages often result in unwanted cross-talk and overloading from adjacent commercial or amateur stations. The filter shown is inserted between the antenna and receiver. It attenuates the out-of-band signals from broadcast stations but passes signals of interest (1.8 to 30 MHz) with little or no attenuation.

The high signal strength of local broadcast stations requires that the stop-band attenuation of the high-pass filter also be high. This filter provides about 60 dB of stop-band attenuation with less than 1 dB of attenuation above 1.8 MHz. The filter input and output ports match 50 Ω with a maximum SWR of 1.353:1 (reflection coefficient = 0.15). A 10-element filter yields adequate stop-band attenuation and a reasonable rate of attenuation rise. The design uses only standard-value capacitors.

BUILDING THE FILTER

The filter parts layout, schematic diagram, response curve and component values are shown in **Fig 12.69**. The standard capacitor values listed are within 2.8% of the design values. If the attenuation peaks (f2, f4 and f6) do not fall at 0.677, 1.293 and 1.111 MHz, tune the series-resonant circuits by slightly squeezing or separating the inductor windings.

Construction of the filter is shown in **Fig 12.70**. Use Panasonic P-series polypropylene (type ECQ-P) capacitors. These capacitors are available through Digi-Key and other suppliers. The powdered-iron T50-2 toroidal cores are available through Amidon, Palomar Engineers and others.

For a 3.4-MHz cutoff frequency, divide the L and C values by 2. (This effectively doubles the frequency-label values in Fig 12.69.) For the 80-m version, L2 through L6 should be 20 to 25 turns each, wound on T50-6 cores. The actual turns required may vary one or two from the calculated values. Parallel-connect capacitors as needed to achieve the nonstandard capacitor values required for this filter.

FILTER PERFORMANCE

The measured filter performance is shown in Fig 12.69. The stop-band attenuation is more than 58 dB. The measured cutoff frequency (less than 1 dB attenuation) is under 1.8 MHz. The measured passband loss is less than 0.8 dB from 1.8 to 10 MHz. Between 10 and 100 MHz, the insertion loss of the filter gradually increases to 2 dB. Input impedance was measured between 1.7 and 4.2 MHz. Over the range tested, the input impedance of the filter remained within the 37 to 67.7-Ω input-impedance window (equivalent to a maximum SWR of 1.353:1).

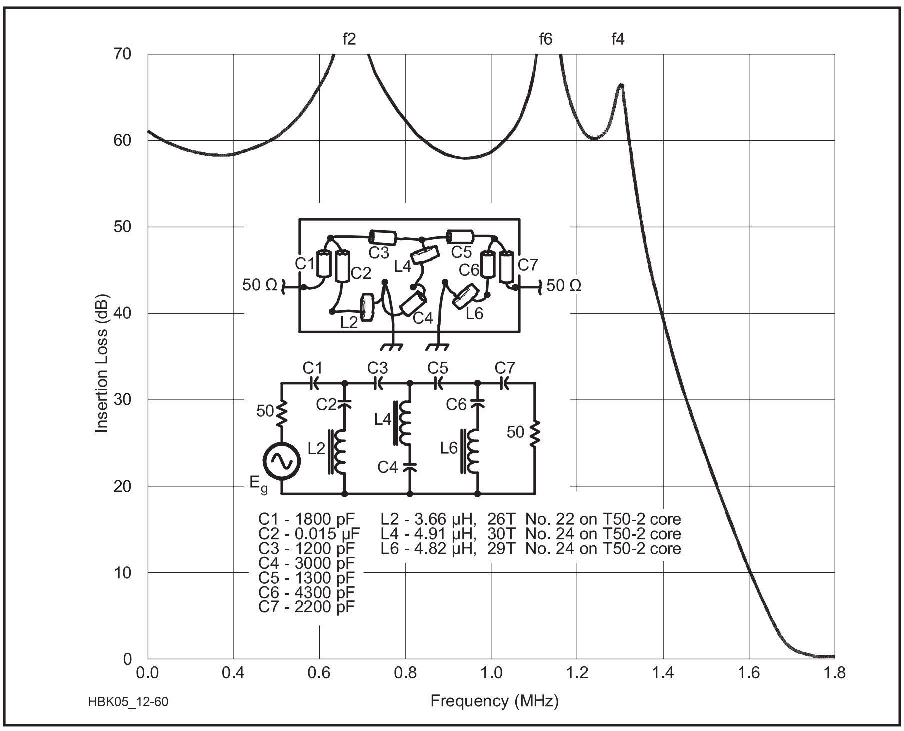


Fig 12.69 — Schematic, layout and response curve of the broadcast band rejection filter.

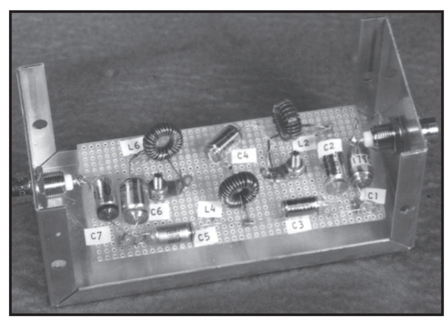


Fig 12.70 — The filter fits easily in a 2 × 2 × 5-inch enclosure. The version in the photo was built on a piece of perfboard.

A WAVE TRAP FOR BROADCAST STATIONS

Nearby medium-wave broadcast stations can sometimes cause interference to HF receivers over a broad range of frequencies. A wave trap can catch the unwanted frequencies and keep them out of your receiver.

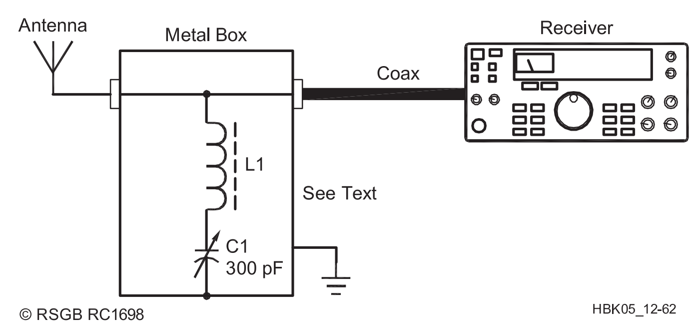


Fig 12.71 — The wave trap consists of a series tuned circuit, which 'shunts' signals on an unwanted frequency to ground.

CIRCUIT DESCRIPTION

The way the circuit works is quite simple. Referring to **Fig 12.71**, you can see that it consists essentially of only two components, a coil L1 and a variable capacitor C1. This series-tuned circuit is connected in parallel with the antenna circuit of the receiver. The characteristic of a series-tuned circuit is that the coil and capacitor have a very low impedance (resistance) to frequencies very close to the frequency to which the circuit is tuned. All other frequencies are almost unaffected. If the circuit is tuned to 1530 kHz, for example, the signals from a broadcast station on that frequency will flow through the filter to ground, rather than go on into the receiver. All other frequencies will pass straight into the receiver. In this way, any interference caused in the receiver by the station on 1530 kHz is significantly reduced.

CONSTRUCTION

This is a series-tuned circuit that is adjustable from about 540 kHz to 1600 kHz. It is built into a metal box, **Fig 12.72**, to shield it from other unwanted signals and is connected as shown in Fig 12.71. To make the inductor, first make a *former* by winding two layers of paper on the ferrite rod. Fix this in place with black electrical tape. Next, lay one end of the wire for the coil on top of the former, leaving about an inch of wire protruding beyond the end of the ferrite rod. Use several turns of electrical tape to secure the wire to the former. Now, wind the coil along the former, making sure the turns are in a single layer and close together. Leave an inch or so of wire free at the end of the coil. Once again, use a couple of turns of electrical tape to secure the wire to the former. Finally, remove half an inch of enamel from each end of the wire.

Alternatively, if you have an old AM transistor radio, a suitable coil can usually be recovered already wound on a ferrite rod. Ignore any small coupling coils. Drill the box to take the components, then fit them in and solder together as shown in **Fig 12.73**. Make sure the lid of the box is fixed securely in place, or the wave trap's performance will be adversely affected by pick-up on the components.

CONNECTION AND ADJUSTMENT

Connect the wave trap between the antenna and the receiver, then tune C1 until the interference from the offending broadcast station is a minimum. You may not be able to eliminate interference completely, but this handy little device should reduce it enough to listen to the amateur bands. Lets say you live near an AM transmitter on 1530 kHz, and the signals break through on your 1.8-MHz receiver. By tuning the trap to 1530 kHz, the problem is greatly reduced. If you have problems from more than one broadcast station, the problem needs a more complex solution.

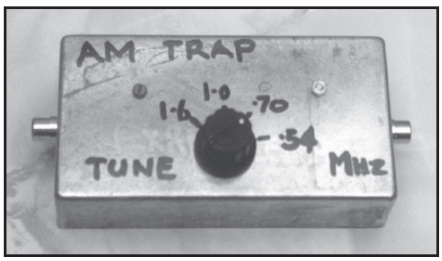


Fig 12.72 — The wave trap can be roughly calibrated to indicate the frequency to which it is tuned.

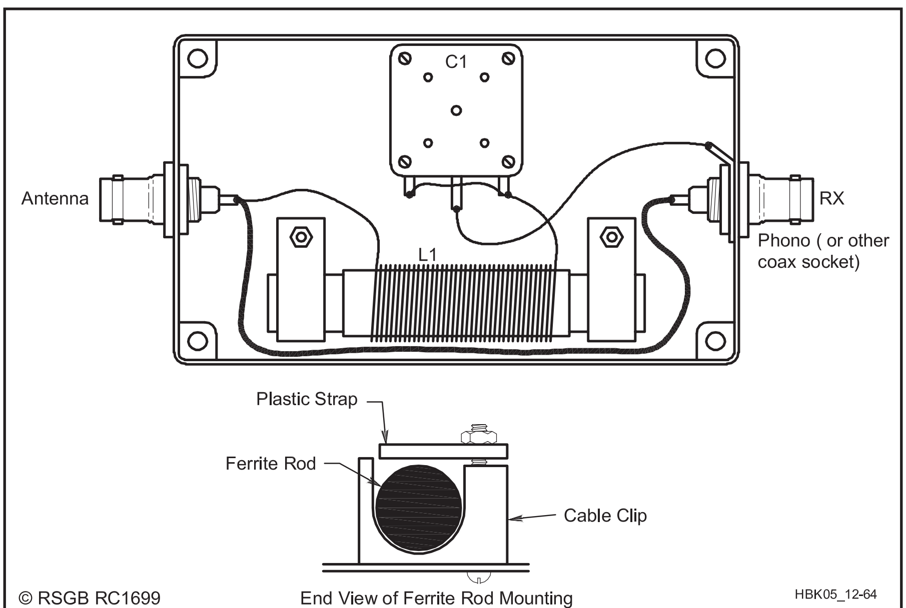


Fig 12.73 — Wiring of the wave trap. The ferrite rod is held in place with cable clips.
C1 — 300 pF polyvaricon variable.
L1 — 80 turns of 30 SWG enameled wire, wound on a ferrite rod.
Associated items: Case (die-cast box), knobs to suit, connectors to suit, nuts and bolts, plastic cable clips.

OPTIMIZED HARMONIC FILTERS

Low-pass filters should be placed at the output of transmitters to ensure that they meet the various regulatory agency requirements for harmonic suppression. These are commonly designed to pass a single amateur band and provide attenuation at harmonics of that band sufficient to meet the requirements. The material presented here by Jim Tonne, WB6BLD, is based on material originally published in the September/October 1998 issue of *QEX*. The basic approach is to use a computer to optimize the performance in the passband (a single amateur band) while simultaneously maximizing the attenuation at the second and third harmonic of that same band. When this is done, the higher harmonics will also be well within spec.

The schematic of this filter along with parts values for the 3.5 to 4.0 MHz amateur band is shown in **Fig 12.74**. The responses of that filter are shown in **Fig 12.75**.

Component values for the 160-meter through the 6-meter amateur bands are shown in **Table 12.23**. The capacitors are shown in pF and the inductors in µH. The capacitors are the nearest 5% values; both the nearest 5% and the exact inductor values are shown.

Using the nearest-5% inductor values will result in satisfactory operation. If the construction method is such that exact-value (adjustable) inductors can be used then the "Exact" values are preferred. These values were obtained from the program *SVC Filter Designer* which is on the *ARRL Handbook CD* included with this book.

SVC Filter Design Software

The *ARRL Handbook CD* at the back of this book includes *SVC Filter Designer*, a *Windows* program for design lumped element high-pass and low-pass filters by Jim Tonne, WB6BLD. The software shows ideal values and also the nearest 5% values for capacitors and inductors. It also analyzes those filters and shows the deviation of key responses from ideal when those 5% values are used. See the "About the Included CD-ROM" section for more information.

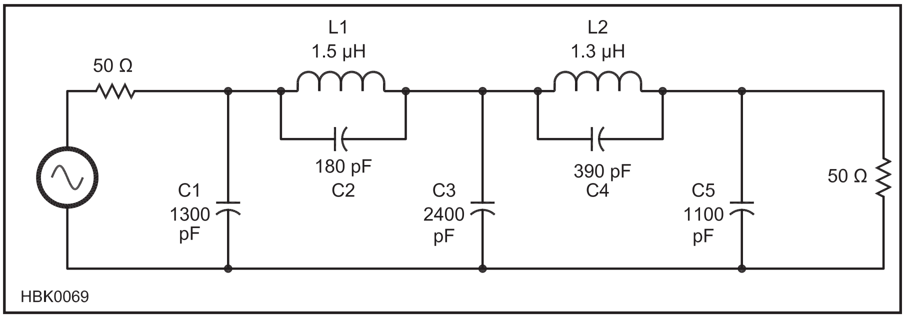


Fig 12.74 — Optimized low-pass filter. This design is for the 80-meter amateur band. It is similar to a Cauer design but the parts values have been optimized as described in the text and in the referenced issue of *QEX*.

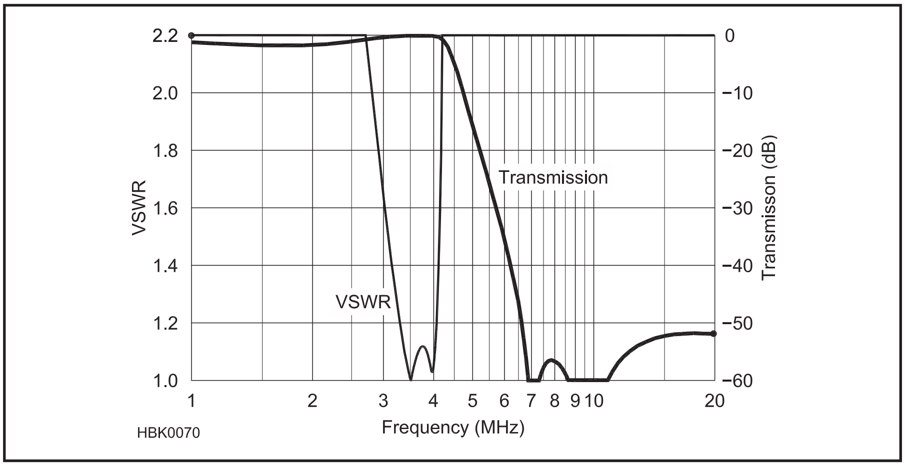


Fig 12.75 — Responses of the filter shown in Fig 12.74. Note the good (low) values for VSWR from 3.5 MHz to 4 MHz. At the same time the harmonics are attenuated to meet regulations. Responses for the other amateur bands are very similar except for the frequency scaling.

Table 12.23
Values for the Optimized Harmonic Filters

Band (meters)	*C1 (pF)*	*L2, 5% (µH)*	*L2, Exact (µH)*	*C2 (pF)*	*C3 (pF)*	*L4, 5% (µH)*	*L4, Exact (µH)*	*C4 (pF)*	*C5 (pF)*
160	2400	3.0	2.88	360	4700	2.4	2.46	820	2200
80	1300	1.5	1.437	180	2400	1.3	1.29	390	1100
60	910	1.0	1.029	120	1600	0.91	0.8897	270	750
40	680	0.75	0.7834	91	1300	0.62	0.6305	220	560
30	470	0.56	0.5626	68	910	0.47	0.4652	160	430
20	330	0.39	0.3805	47	620	0.33	0.3163	110	300
17	270	0.30	0.3063	36	510	0.27	0.2617	82	240
15	220	0.27	0.2615	30	430	0.22	0.2245	68	200
12	200	0.24	0.241	27	390	0.20	0.2042	62	180
10	180	0.20	0.2063	24	330	0.18	0.1721	56	150
6	91	0.11	0.108	13	180	0.091	0.0911	30	82

THE DIPLEXER FILTER

This section, covering diplexer filters, was written by William E. Sabin, WØIYH. The diplexer is helpful in certain applications, and Chapter 11 shows them used as frequency mixer terminations.

Diplexers have a constant filter-input resistance that extends to the stop band as well as the passband. Ordinary filters that become highly reactive or have an open or short-circuit input impedance outside the passband may degrade performance.

Fig 12.76 shows a *normalized* prototype 5-element, 0.1-dB Chebyshev low-pass/high-pass (LP/HP) filter. This idealized filter is driven by a voltage generator with zero internal resistance, has load resistors of 1.0 Ω and a cutoff frequency of 1.0 radian per second (0.1592 Hz). The LP prototype values are taken from standard filter tables.[1] The first element is a series inductor. The HP prototype is found by:

a) replacing the series L (LP) with a series C (HP) whose value is 1/L, and

b) replacing the shunt C (LP) with a shunt L (HP) whose value is 1/C.

For the Chebyshev filter, the return loss is improved several dB by multiplying the prototype LP values by an experimentally derived number, K, and dividing the HP values by the same K. You can calculate the LP values in henrys and farads for a 50-Ω RF application with the following formulas:

$$L_{LP} = \frac{K L_{P(LP)}\ R}{2\pi f_{CO}}\ ;\ C_{LP} = \frac{K C_{P(LP)}}{2\pi f_{CO}\ R}$$

where

$L_{P(LP)}$ and $C_{P(LP)}$ are LP prototype values
$K = 1.005$ (in this specific example)
$R = 50\ \Omega$
f_{CO} = the cutoff (–3-dB response) frequency in Hz.

For the HP segment:

$$L_{HP} = \frac{L_{P(HP)}\ R}{2\pi f_{CO}\ K}\ ;\ C_{HP} = \frac{C_{P(HP)}}{2\pi f_{CO}\ KR}$$

where $L_{P(HP)}$ and $C_{P(HP)}$ are HP prototype values.

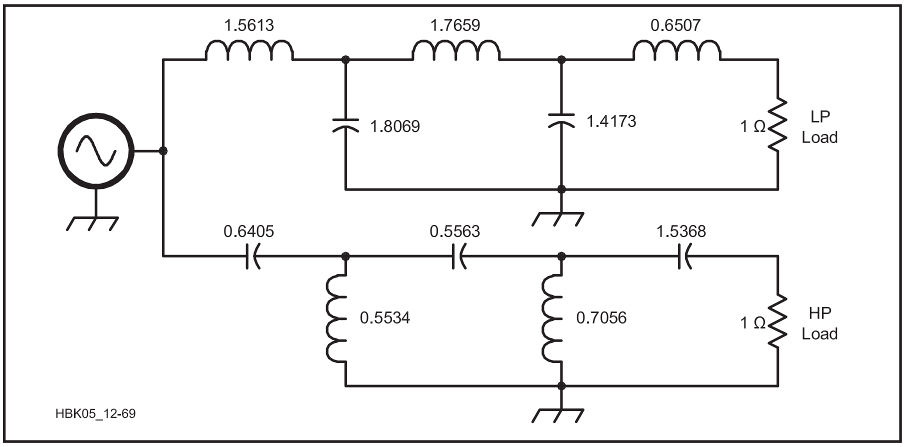


Fig 12.76 — Low-pass and high-pass prototype diplexer filter design. The low-pass portion is at the top, and the high-pass at the bottom of the drawing. See text.

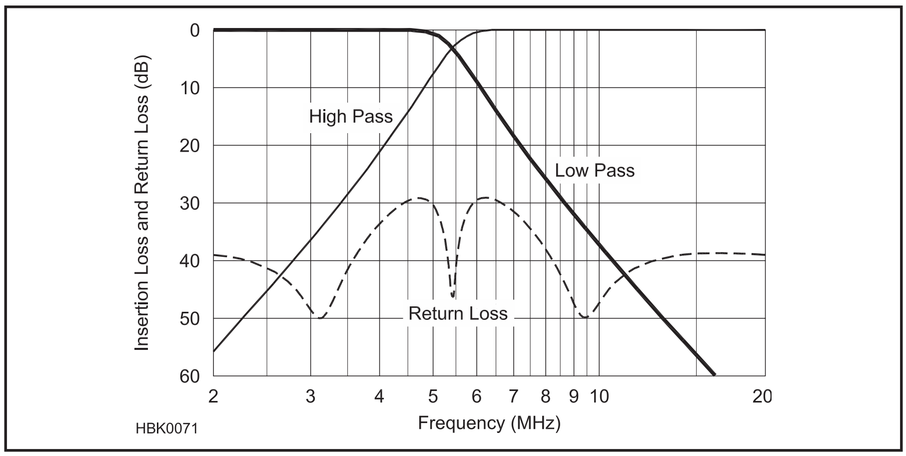


Fig 12.77 — Response for the low-pass and high-pass portions of the 80-meter diplexer filter. Also shown is the return loss of the filter.

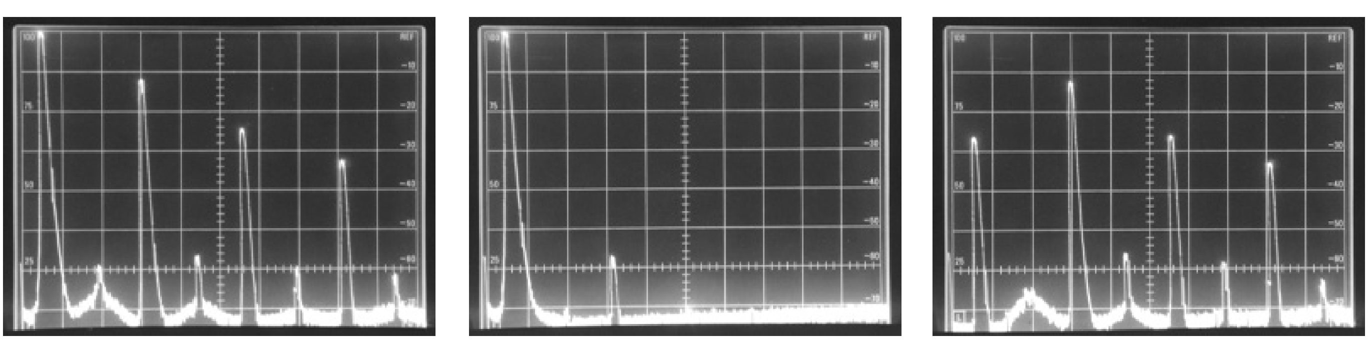
Fig 12.77 shows the LP and HP responses of a diplexer filter for the 80-meter band. The following items are to be noted:

• The 3 dB responses of the LP and HP meet at 5.45 MHz.

• The input impedance is close to 50 Ω at all frequencies, as indicated by the high value of return loss (SWR <1.07:1).

• At and near 5.45 MHz, the LP input reactance and the HP input reactance are conjugates; therefore, they cancel and produce an almost perfect 50-Ω input resistance in that region.

• Because of the way the diplexer filter



(A) (B) (C)

Fig 12.78 — At A, the output spectrum of a push-pull 80-meter amplifier. At B, the spectrum after passing through the low-pass filter. At C, the spectrum after passing through the high-pass filter.

is derived from synthesis procedures, the transfer characteristic of the filter is mostly independent of the actual value of the amplifier dynamic output impedance.[2] This is a useful feature, since the RF power amplifier output impedance is usually not known or specified.

• The 80-meter band is well within the LP response.

• The HP response is down more than 20 dB at 4 MHz.

• The second harmonic of 3.5 MHz is down only 18 dB at 7.0 MHz. Because the second harmonic attenuation of the LP is not great, it is necessary that the amplifier itself be a well-balanced push-pull design that greatly rejects the second harmonic. In practice this is not a difficult task.

• The third harmonic of 3.5 MHz is down almost 40 dB at 10.5 MHz.

Fig 12.78A shows the unfiltered of a solid-state push-pull power amplifier for the 80-meter band. In the figure you can see that:

• The second harmonic has been suppressed by a proper push-pull design.

• The third harmonic is typically only 15 dB or less below the fundamental.

The amplifier output goes through our diplexer filter. The desired output comes from the LP side, and is shown in Fig 12.78B. In it we see that:

• The fundamental is attenuated only about 0.2 dB.

• The LP has some harmonic content; however, the attenuation exceeds FCC requirements for a 100-W amplifier.

Fig 12.78C shows the HP output of the diplexer that terminates in the HP load or *dump* resistor. A small amount of the fundamental frequency (about 1%) is also lost in this resistor. Within the 3.5 to 4.0 MHz band, the filter input resistance is almost exactly the correct 50-Ω load resistance value. This is because power that would otherwise be *reflected* back to the amplifier is absorbed in the dump resistor.

Solid state power amplifiers tend to have stability problems that can be difficult to debug.[3] These problems may be evidenced by level changes in: load impedance, drive, gate or base bias, B+, etc. Problems may arise from:

• The reactance of the low-pass filter outside the desired passband. This is especially true for transistors that are designed for high-frequency operation.

• Self resonance of a series inductor at some high frequency.

• A stop band impedance that causes voltage, current and impedance reflections back to the amplifier, creating instabilities within the output transistors.

Intermodulation performance can also be degraded by these reflections. The strong third harmonic is especially bothersome for these problems.

The diplexer filter is an approach that can greatly simplify the design process, especially for the amateur with limited PA-design experience and with limited home-lab facilities. For these reasons, the amateur homebrew enthusiast may want to consider this solution, despite its slightly greater parts count and expense.

The diplexer is a good technique for narrowband applications such as the HF amateur bands.[4] From Fig 12.77, we see that if the signal frequency is moved beyond 4.0 MHz the amount of desired signal lost in the dump resistor becomes large. For signal frequencies below 3.5 MHz the harmonic reduction may be inadequate. A single filter will not suffice for all the HF amateur bands.

This treatment provides you with the information to calculate your own filters. A *QEX* article has detailed instructions for building and testing a set of six filters for a 120-W amplifier. These filters cover all nine of the MF/HF amateur bands.[5]

You can use this technique for other filters such as Bessel, Butterworth, linear phase, Chebyshev 0.5, 1.0, etc.[6] However, the diplexer idea does *not* apply to the elliptic function types.

The diplexer approach is a resource that can be used in any application where a constant value of filter input resistance over a wide range of passband and stop band frequencies is desirable for some reason. Computer modeling is an ideal way to finalize the design before the actual construction. The coil dimensions and the dump resistor wattage need to be determined from a consideration of the power levels involved, as illustrated in Fig 12.78.

Another significant application of the diplexer is for elimination of EMI, RFI and TVI energy. Instead of being reflected and very possibly escaping by some other route, the unwanted energy is dissipated in the dump resistor.[7]

Diplexer Filter Design Software

The *ARRL Handbook CD* at the back of this book includes *DiplexerDesigner*, a *Windows* program for design and analysis of diplexer filters by Jim Tonne, WB6BLD. The software is interactive and plots performance changes as various parameters are altered. See the “About the Included CD-ROM” section for more information.

Notes

[1]Williams, A. and Taylor, F., *Electronic Filter Design Handbook*, any edition, McGraw-Hill.

[2]Storer, J.E., *Passive Network Synthesis*, McGraw-Hill 1957, pp 168-170. This book shows that the input resistance is ideally constant in the passband and the stop band and that the filter transfer characteristic is ideally independent of the generator impedance.

[3]Sabin, W. and Schoenike, E., *HF Radio Systems and Circuits*, Chapter 12, Noble Publishing, 1998. Also the previous edition of this book, *Single-Sideband Systems and Circuits*, McGraw-Hill, 1987 or 1995.

[4]Dye, N. and Granberg, H., *Radio Frequency Transistors, Principles and Applications*, Butterworth-Heinemann, 1993, p 151.

[5]Sabin, W.E. WØIYH, “Diplexer Filters for the HF MOSFET Power Amplifier,” *QEX*, Jul/Aug, 1999. Also check ARRLWeb at **www.arrl.org/qex/**.

[6]See note 1. *Electronic Filter Design Handbook* has LP prototype values for various filter types, and for complexities from 2 to 10 components.

[7]Weinrich, R. and Carroll, R.W., “Absorptive Filters for TV Harmonics,” *QST*, Nov 1968, pp 10-25.

BAND-PASS FILTERS FOR 144 OR 222 MHZ

Spectral purity is necessary during transmitting. Tight filtering in a receiving system ensures the rejection of out-of-band signals. Unwanted signals that lead to receiver overload and increased intermodulation-distortion (IMD) products result in annoying in-band "birdies." One solution is the double-tuned band-pass filters shown in **Fig 12.79.** They were designed by Paul Drexler, W2PED. Each includes a resonant trap coupled between the resonators to provide increased rejection of undesired frequencies.

Many popular VHF conversion schemes use a 28-MHz intermediate frequency (IF), yet proper filtering of the image frequency is often overlooked in amateur designs. The low-side injection frequency used in 144-MHz mixing schemes is 116 MHz and the image frequency, 88 MHz, falls in TV channel 6. Inadequate rejection of a broadcast carrier at this frequency results in a strong, wideband signal at the low end of the 2-m band. A similar problem on the transmit side can cause TVI. These band-pass filters have effectively suppressed undesired mixing products. See **Fig 12.80** and **12.81**.

The circuit is constructed on a double-sided copper-clad circuit board. Minimize component lead lengths to eliminate resistive losses and unwanted stray coupling. Mount the piston trimmers through the board with the coils soldered to the opposite end, parallel to the board. The shield between L1 and L3 decreases mutual coupling and improves the frequency response. Peak C1 and C3 for optimum response.

L1, C1, L3 and C3 form the tank circuits that resonate at the desired frequency. C2 and L2 reject the undesired energy while allowing the desired signal to pass. The tap points on L1 and L3 provide 50-Ω matching; they may be adjusted for optimum energy transfer. Several filters have been constructed using a miniature variable capacitor in place of C2 so that the notch frequency could be varied.

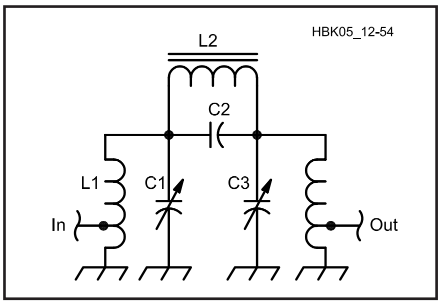


Fig 12.79 — Schematic of the VHF band-pass filter for receiving systems. Capacitor values are the same for either band.
C1, C3— 1-7 pF piston trimmer.
C2 — 1 pF.
L1, L3 — 144 MHz, 7t no. 18, 0.25-in ID, tap 1.5t; 222 MHz, 4t no. 18, 0.25-in ID, tap 1.5t.
L2 — 144 MHz, 27t no. 26 enam on T37-10; 222 MHz, 15t no. 24 enam on T44-10.

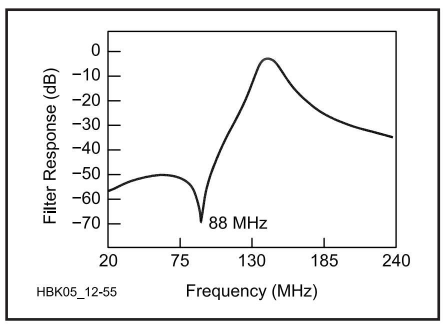


Fig 12.80 — Filter response plot of the 144-MHz band-pass filter, with an image-reject notch for a 28 MHz IF.

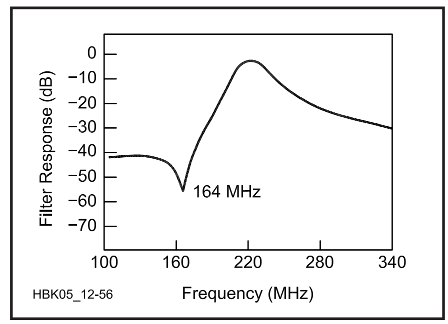


Fig 12.81 — Filter response plot of the 222-MHz band-pass filter, with an image-reject notch for a 28 MHz IF.

A HIGH PERFORMANCE, LOW COST 1.8 TO 54 MHZ LOW PASS FILTER

The low-pass filter shown in **Fig 12.82** offers low insertion loss, mechanical simplicity, easy construction and operation on all amateur bands from 160 through 6 meters. Originally built as an accessory filter for a 1500 W 6-meter amplifier, the filter easily handles legal limit power. No complicated test equipment is necessary for alignment. It was originally described by Bill Jones, K8CU, in November 2002 *QST*.

Although primarily intended for coverage of the 6-meter band, this filter has low insertion loss and presents excellent SWR characteristics for all HF bands. Although harmonic attenuation at low VHF frequencies near TV channels 2, 3 and 4 does not compare to filters designed only for HF operation, the use of this filter on HF is a bonus to 6-meter operators who also use the HF bands. Six-meter operators may easily tune this filter for low insertion loss and SWR in any favorite band segment, including the higher frequency FM portion of the band.

ELECTRICAL DESIGN

The software tool used to design this low pass filter is *Elsie* by Jim Tonne, WB6BLD, which may be found on the CD included with this book. The *Elsie* format data file for this filter, DC54.lct, may be downloaded for your own evaluation from the author's Web site at **www.realhamradio.com**.

Fig 12.83 is a schematic diagram of the filter. The use of low self-inductance capacitors with Teflon dielectric easily allows legal limit high power operation and aids in the ultimate stop band attenuation of this filter. Capacitors with essentially zero lead length will not introduce significant series inductance that upsets filter operation. This filter also uses a trap that greatly attenuates second harmonic frequencies of the 6-meter band.

MECHANICAL DETAILS

A detailed mechanical drawing is shown in **Fig 12.84**. One design goal of this filter was easy tuning with modest home test equipment. To realize this, build the coils carefully. The homemade coils solder directly to the top surface of the brass capacitor plates. Capacitors C1 and C2 are made using a brass to Teflon

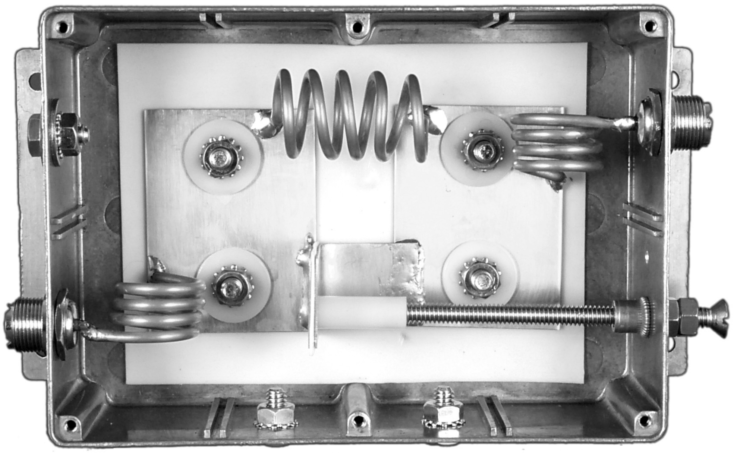


Fig 12.82 — The 1.8 to 54 MHz low-pass filter is housed in a diecast box.

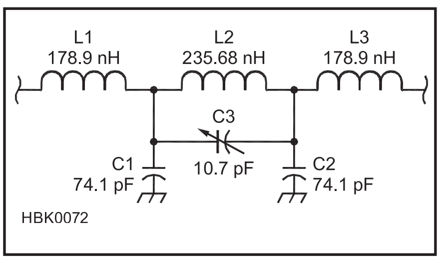


Fig 12.83 — The low pass filter schematic.

C1, C2 — 74.1 pF. 2 × 2.65 inch brass plate sandwiched with 0.03125 inch thick Teflon sheet. The metal enclosure is the remaining grounded terminal of this capacitor.

C3 — See text and Fig 12.84.

L1, L3 — 178.9 nH. Wind with ⅛ inch OD soft copper tubing, 3.5 turns, 0.75 inch diameter form, 0.625 inch long, ¼ inch lead length for soldering to brass plate. The length of the other lead to RF connector as required.

L2 — 235.68 nH. Wind with ⅛ inch OD soft copper tubing, 5 turns, 0.75 inch diameter form, 1.75 inches long. Leave ¼ inch lead length for soldering.

to aluminum case sandwich. C3 is made from two pieces of 0.032-inch thick brass plate and a Teflon insulator. The filter inductors are mounted at right angles to each other to help maintain good stop band attenuation.

One *Elsie* software tool will calculate the details of each inductor. Inductors L1 and L3 are designed with a half turn winding. This allows short connections to the brass capacitor plate and the RF connectors mounted on the enclosure walls. The coils are physically spaced with ¼-inch lead lengths, and then soldered to the brass plates.

Many of the parts required to make this filter are available at hardware stores. In particular, the 1 and 2 inch wide brass strips (sold as Hobby or Miniature brass), ⅛ inch diameter soft copper tubing, nuts, bolts, and nylon spacers and washers are commonly available at low cost. It is important that the specified 0.03125-inch thickness of Teflon be used since another size will result in a different capacitor value. If you have another Teflon thickness available, you will need to calculate the specific capacitor values depending upon the new thickness and brass plate sizes.

The opaque white Teflon used here has a dielectric constant of 2.1. The clear varieties typically have lower values and will result in different capacitance for the same size brass plates. The capacitance will decrease if the assembly bolts are loose, so be sure to have the bolts tightened. Also, use the 0.064-inch thick brass plate for the bolted down capacitors. When under compression, the thinner brass size used for the variable capacitor tends to flex more and doesn't fit as flat to the Teflon.

A separate Teflon sheet, also used in the variable capacitor, is glued to the stationary brass plate. This insulator is used to prevent a short circuit in case the tuning screw is tightened too much. Teflon is extremely slick, and doesn't glue well unless chemically prepared. One way to get acceptable glue joint performance between the brass support plate and the insulator is to scuff the Teflon and brass surfaces with sandpaper. The intention is to increase the available surface area as much as possible, and provide more places for the glue to fasten to. Glue the Teflon in place with a bead of RTV or epoxy. After drying, the Teflon sheet can be intentionally peeled from the brass plate, but it appears to hold reasonably well. Special Teflon that has been treated to allow good adhesion is available, but the expense isn't justified for this simple application. This Teflon variable capacitor insulator sheet measures 1.5 inches wide by 1.75 inches tall and is larger than the two brass plates. This gives an outside edge insulation some safety margin.

Calculating Capacitance

The 0.064-inch thick brass capacitor plates have two 0.5-inch holes in them for the mounting bolts and washers. The surface area of each hole is πR^2, so the two holes combined have a total surface area of 0.3925 square inch. The brass plate size is 2 inches by 2.65 inches. This equals 5.3 square inches of surface area. Subtracting the area of the two holes gives a total surface area of 4.9 square inches. The formula for capacitance is

$$C = 0.2248\ (kA/d)$$

where

C = capacitance in pF
K = dielectric constant of Teflon
A = surface area of one plate in square inches
d = thickness of insulator

The dielectric constant of the Teflon used here is 2.1, and the thickness used is 0.03125 inch. The calculated capacitance of each plate equals 74.1 pF. Measured values agree closely with this number. When built as described, the capacitor plates measured between 2% and 2.5% of the calculated value. This is acceptable for a practical filter.

The brass sheet material acts like a large heat sink, so an adequate soldering iron is required. A large chisel point 125-W iron will work well. The soldering heat does not affect the Teflon material. However, beware of the temptation to use a small propane torch. Two bolts in each capacitor hold the Teflon sheet and brass plates firmly together. The bolts are insulated from the brass plates by nylon spacers the same thickness as the brass. The nylon plunger for the tuning capacitor needs to be drilled and tapped to accept the ¼ × 20 thread of the adjustment bolt. A threaded insert or PEM nut in the enclosure provides support for the tuning bolt.

ADJUSTMENT

C3 is shunted across coil L2. This coil and capacitor combination acts like a tunable trap for second harmonic frequencies when operating in the 6-meter band. After soldering into place, the flexible tuning plate of this capacitor is simply bent toward the adjustment screw. Brass of this thickness has a definite spring effect. Just bend the plate well toward the tuning screw, and then tighten the tuning bolt inward. This will result in a stable variable capacitor.

Six-Meter Alignment Procedure

If you wish to use this filter from 1.8 to 30 MHz only (no 6 meter operation), C3 adjustment is not critical and does not affect HF SWR performance. Simply set the C3 plates 0.1 inch apart and disregard the following steps. However, don't eliminate the capacitor entirely. The software predicts degraded VHF response with it missing. For use on the HF bands only, the tuning screw and associated nylon plunger may be omitted.

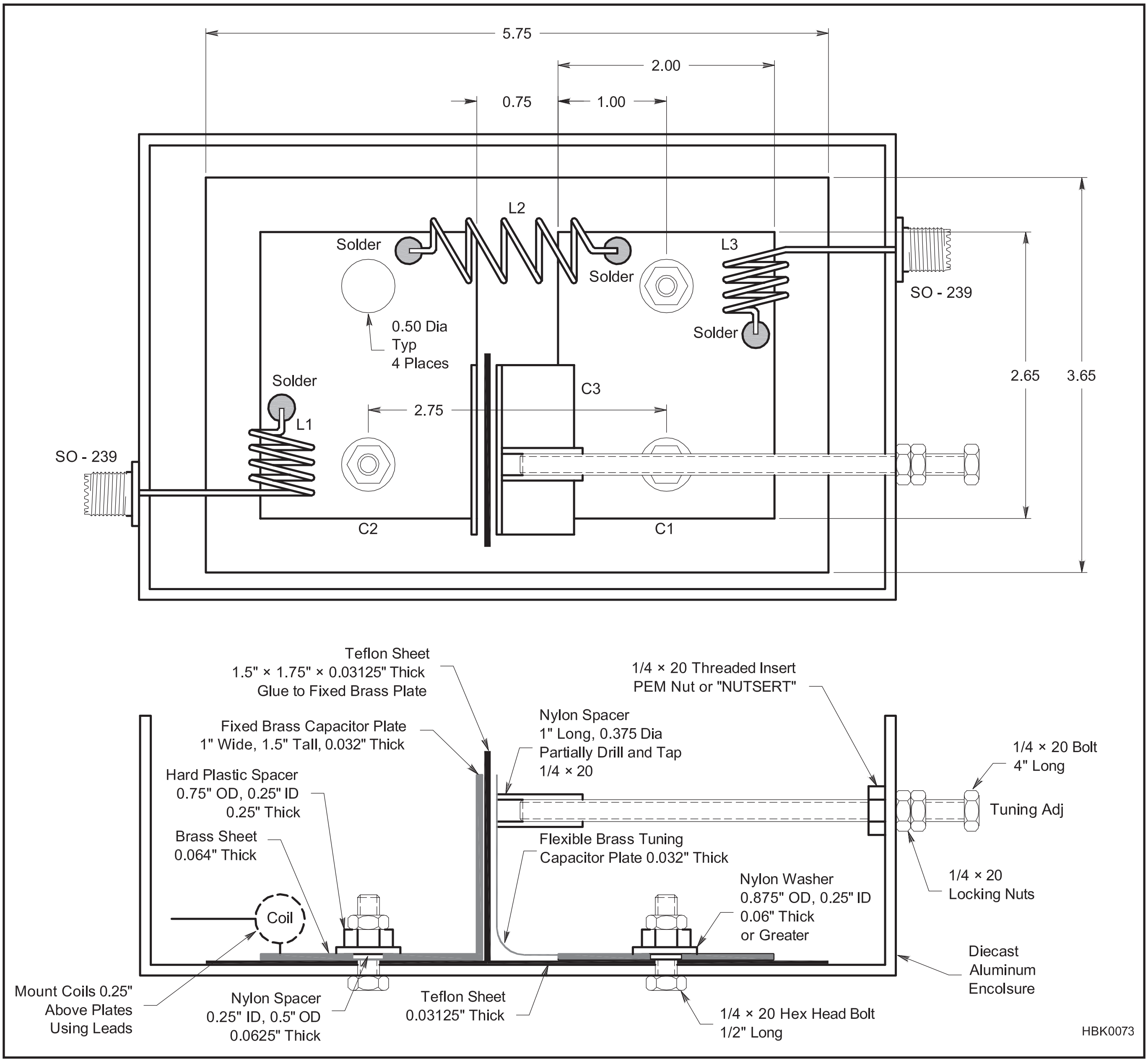


Fig 12.84 — Assembly drawing for the low pass filter. All dimensions are in inches. See Table 12.24 for a complete parts list.

Table 12.24
Low-Pass Filter Parts List

Qty	*Description*
1	Miniature brass strip, 1 × 12 in., 0.032 in. thick (C3)
1	Miniature brass strip, 2 × 12 in., 0.064 in. thick (C1, C2)
5 ft	⅛ inch diameter soft copper tubing
4	¼ × 20 × ½ in. long hex head bolt
4	Plastic spacer or washer, 0.5 in. OD, 0.25 in. ID, 0.0625 in. thick
6	¼ × 20 hex nut with integral tooth lock washer
1	¼ × 20 × 4 in. long bolt
1	¼ × 20 threaded nut insert, PEM nut, or “Nutsert”
1	1 × 0.375 in. diameter nylon spacer. ID smaller than 0.25 in.(used for C3 plunger).
4	Nylon spacer, 0.875 in. OD, 0.25 to 0.34 in. ID, approx. 0.065 in. or greater thickness (used to attach brass capacitor plates).

Aluminum diecast enclosure is available from Jameco Electronics (**www.jameco.com**) part no. 11973. The box dimensions are 7.5 × 4.3 × 2.4 in. The 0.03125 in. thick Teflon sheet is available from McMaster-Carr Supply Co (**www.mcmaster.com**), item #8545K21 is available as a 12 × 12 in. sheet.

Normally, tuning this filter would be a challenge, since three variables (with two interacting) are involved (L1, L2 and C3). I realized that the *Elsie* software “Tune” mode held the answer. After studying what the software predicted, I generated this tuning procedure. My very first attempt to exactly tune this filter was successful, and was completed in just a few minutes. This method was predicted by software and then confirmed in practice. A common variable SWR analyzer is required. These steps may seem complicated, but are actually pretty straightforward once you get a feel for it. Read first before you start adjusting.

Step One

Adjust C3 until the top plate spacing is about 0.1 inch apart. Using an SWR analyzer, search for a very low SWR null anywhere in the vicinity of about 45 to 60 MHz. If a low SWR value (near 1:1) can be found, even though the frequency of the low SWR isn't where you want it, proceed to step two. Otherwise, adjust the input coil L1 by expanding or compressing the turns until a low SWR can be obtained anywhere in 45 to 60 MHz range. If you have a way to measure the notch response at 100.2 MHz, proceed to step two. Otherwise, proceed to step three.

Step Two

Now apply a 100.2 MHz signal to the filter input. Adjust C3 until the 6 meter second harmonic at 100.2 MHz is nulled on the filter output. Hook up the SWR analyzer again, and sweep the 6-meter band with the SWR analyzer. If the low SWR frequency is too low, adjust L2 for less inductance (expand turns farther apart), and then readjust the variable capacitor to bring the notch back on frequency. Continue these iterations until the SWR null is where you want, and the notch frequency is correctly set.

If the desired SWR low spot is too high in frequency, adjust L2 for more inductance (compress the coil turns), and then readjust the variable capacitor for the second harmonic notch. Continue this until both the low SWR frequency location and the notch null are set where you want.

You may need to unsolder one end of coil L2 to allow the adjustment for a longer or shorter coil length as you expand or compress turns. Just solder the end again after you make your length correction. Note that you will probably need to install the enclosure lid during the very final tuning steps.

I was able to reduce the second harmonic into the noise floor of an IFR-1200S spectrum display, but the lid needed to be installed. The lid also interacts with the variable capacitor. Once the SWR and the notch frequency are set, the tuning process is complete and the filter is optimally adjusted. Do not perform step three below.

Step Three

This step is only performed if you don't have a way to generate the 100.2 MHz input signal, and then detect a null on the filter's output terminal. C3 will become your SWR adjustment to move the SWR null spot to the portion of the 6-meter band you desire. If you run out of adjustment range (C3 turned all the way in), just compress the L2 turns together and try again. If C3 is backed completely off, just expand the L2 turns and try again.

After your SWR is set, you are finished. Although the second harmonic notch probably isn't exactly on frequency, you will still have good (but not optimum) suppression since the notch is very deep

PERFORMANCE DISCUSSION

Assuming the 6-meter SWR is set to a low value for a favorite part of the band, the worst case calculated forward filter loss is about 0.18 dB. The forward loss is better in the HF bands, with a calculated loss of only 0.05 dB from 1.8 through 30 MHz. The filter cutoff frequency is about 56 MHz, and the filter response drops sharply above this. There are parasitic capacitors on coils L1 and L3. These are also included in this filter analysis. The calculated self-capacitance of each coil is almost 1 pF. These small capacitors are included on the schematic and are also included in the software for the model. These capacitors occur naturally, so do not solder a

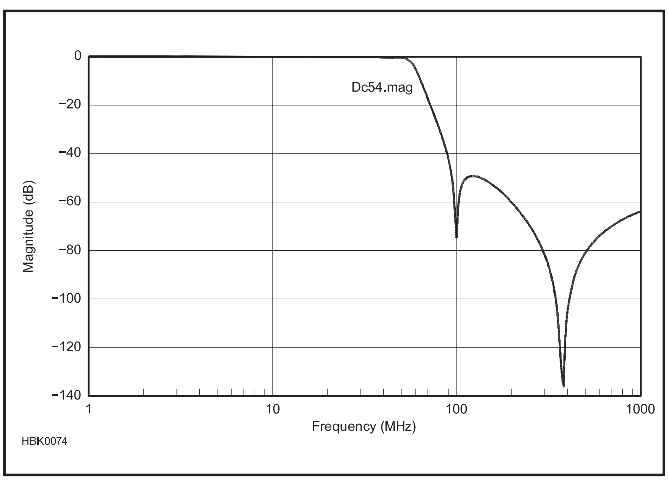


Fig 12.85 — Modeled filter response from 1 to 1000 MHz.

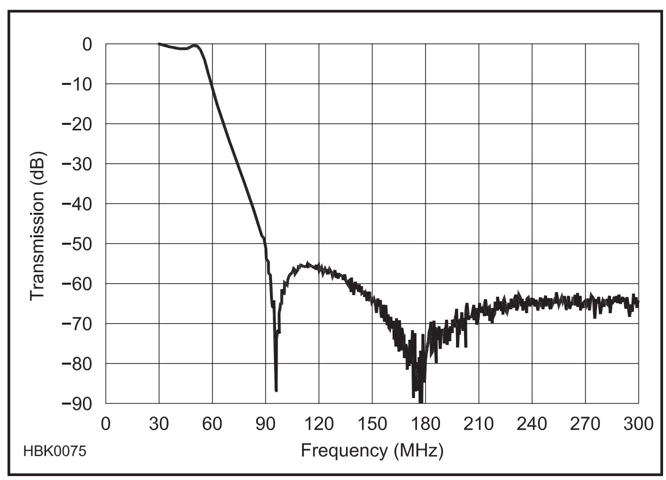


Fig 12.86 — Measured filter response from 300 kHz to 300 MHz.

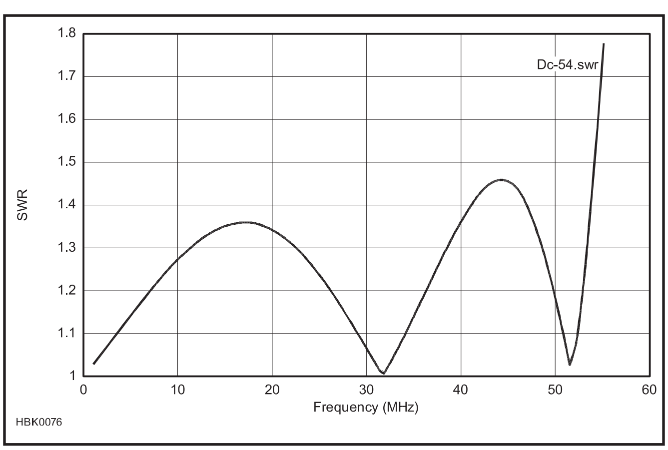


Fig 12.87 — The filter SWR from 1 to 55 MHz.

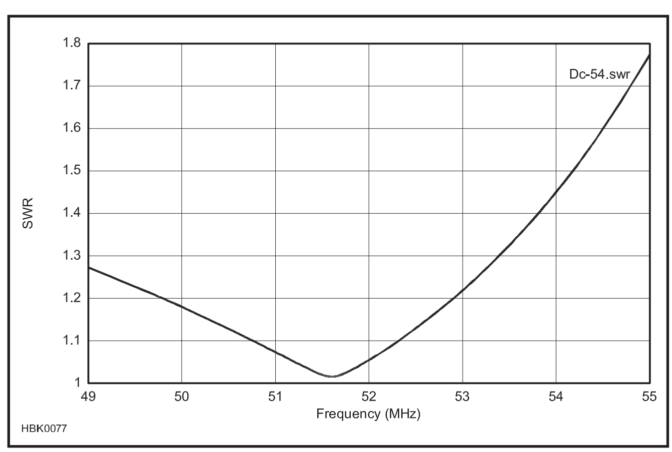


Fig 12.88 — Six-meter filter SWR.

1-pF capacitor across each of the end coils in this filter. The capacitors have the effect of placing additional notches somewhere in the UHF region. The calculated self-resonant frequency of L1 and L3 is about 365 MHz.

Fig 12.85 shows the calculated filter response from 1 to 1000 MHz. The impressive notch near 365 MHz is because of these inherent stray capacitances across each of the coils. Slight variations in each coil will make slightly different tuned traps. This will introduce a stagger-tuned effect that results in a broader notch.

These exact capacitance values are hard to predict because of variations in home made coil dimensions and exact placement of each coil inside the enclosure. The best way to determine their effect is to physically measure the UHF response of this filter. Using low self-inductance capacitors in a VHF filter helps to take advantage of predicted filter attenuation at extended stop-band frequencies.

The measured response of the completed filter is shown in **Fig 12.86**. It corresponds closely with the modeled response. The SWR across the HF bands and 6 meters is shown in **Fig 12.87**. **Fig 12.88** shows only the 6-meter band SWR.

AN EASY-TO-BUILD, HIGH-PERFORMANCE PASSIVE CW FILTER

Modern commercial receivers for amateur radio applications have featured CW filters with digital signal processing (DSP) circuits. These DSP filters provide exceptional audio selectivity with the added advantages of letting the user change the filter's center frequency and bandwidth. Yet in spite of these improvements, many hams are dissatisfied with DSP filters due to increased distortion of the CW signal and the presence of a constant low-level, wide-band noise at the audio output. One way to avoid this distortion and noise is to switch to a selective passive filter that generates no noise! Although the center frequency and bandwidth of the passive filter is fixed and cannot be changed, this is not a serious problem once a center frequency preferred by the user is chosen. The bandwidth can be made narrow enough for good selectivity with no ringing that frequently occurs when the bandwidth is too narrow. This passive CW filter project was designed, built and refined over many years by ARRL Technical Advisor Edward E. Wetherhold, W3NQN.

The effectiveness of an easy-to-build, high-performance passive CW filter in providing distortion-free and noise-free CW reception—when compared with several commercial amateur receivers using DSP filtering—was experienced by Steve Root, KØSR. He reported that when he replaced his DSP filter with the passive CW filter that he assembled, he had the impression that the signals in the filter passband were amplified. In reality, the noise floor appeared to drop one or two dB. When attempting to hear low-level DX CW signals, Steve now prefers the passive CW filter over DSP filters.[1] The CW filter assembled and used by KØSR is the passive five-resonator CW filter that has been widely published in many Handbooks and magazines since 1980 (see references 2-11 at the end of this text).

If you want to build the high-performance passive five-resonator CW filter and experience no-distortion and no-noise CW reception, this article will show you how.

This inductor-capacitor CW filter uses one stack of 85-mH inductors and two modified separate inductors in a five-resonator circuit that is easy to assemble, gives high performance and is low cost. Although these inductors have been referred to as "88 mH" over the past 25 years, their actual value is closer to 85 mH, and for that reason the designs presented in this article are based on an inductor value of 85 mH.

Five band-pass filter designs for center frequencies between 546 Hz and 800 Hz are listed in **Table 12.25**. Select the center frequency that matches your transceiver sidetone frequency. If you are using a direct conversion receiver or an old receiver with a BFO, you may select any of the designs having a center frequency that you find easy on your ears. The author can provide a kit of parts with detailed instructions for assembling this filter at a nominal cost. For contact information, see the end of this text.

The actual 3-dB bandwidth of the filters is between 250 and 270 Hz depending on the center frequency. This bandwidth is narrow enough to give good selectivity, and yet broad enough for easy tuning with no ringing. Five high-Q resonators provide good skirt selectivity that is adequate for interference-free CW reception. Simple construction, low cost and good performance make this filter an ideal first project for anyone interested in putting together a useful station accessory, provided you operate CW mode of course!

DESIGNS AND INTERFACING

Fig 12.89 shows the filter schematic diagram. Component values are given in Table 12.25 for five center-frequency designs. All designs are to be terminated in an impedance between 200 and 230 Ω and standard commercial 8 Ω to 200 Ω audio transformers are used to match the filter input and output to the 8 Ω audio output jack on your receiver — and to an 8 Ω headset. Details are discussed a bit later in this text to interface using headphones with other than 8 Ω impedances that are now quite common.

CONSTRUCTION

The encircled numbers in Fig 12.89 indicate the filter circuit nodes for reference. **Fig 12.90A** shows the L2 and L4 inductor lead connections for the 546-Hz design where no turns need to be removed; the two inductors are used in their original condition. For all other designs, turns need to be removed from each of the windings. The number of turns requiring removal from the L2 and L4 windings is listed in Table 12.25.

Fig 12.90B shows a pictorial of the filter assembly and the connections between the capacitors and the 85-mH stack terminals. Inductors L1, L3 and L5 are contained within the inductor stack and are interconnected using the terminal lugs on the stack as shown in

Table 12.25

CW Filter Using One 85-mH Inductor Stack and Two Modified 85-mH Inductors

Center Freq. (Hz)	*546*	*600*	*700*	*750*	*800*
C1, C5 (nF)	1000	828	608	530	466
C2, C4 (µF)	1.0	1.0	1.0	1.0	1.0
C3 (nF)	333	276	202.7	176.5	155
L2, L4 (mH)	85	70.36	51.69	45.0	39.6
Remove Turns*	None	66	160	200	232

*The total number of turns removed, split equally from each of the two windings of L2. Do the same also for L4. (For example, for a 700-Hz center frequency, remove 80 turns from each of the two windings of L2, for a total of 160 turns removed from L2. Repeat exactly for L4.)

For all designs: L1, L5 = 85 mH; L3 = 255 mH (three 85 mH inductors). Although the surplus inductors are commonly considered to be 88 mH, the actual value is closer to 85 mH. For this reason, all designs are based on the 85-mH value. L2 and L4 have white cores, Magnetic Part No. 55347, OD Max = 24.3mm, ID Min = 13.77mm, HT = 9.70mm; µ = 200, AL = 169 mH/1000T ±8%. The calculated 3-dB BW is 285 Hz and is the same for all designs; however, the actual bandwidth is 5 to 10-percent narrower depending on the inductor Q at the edges of the filter passband.

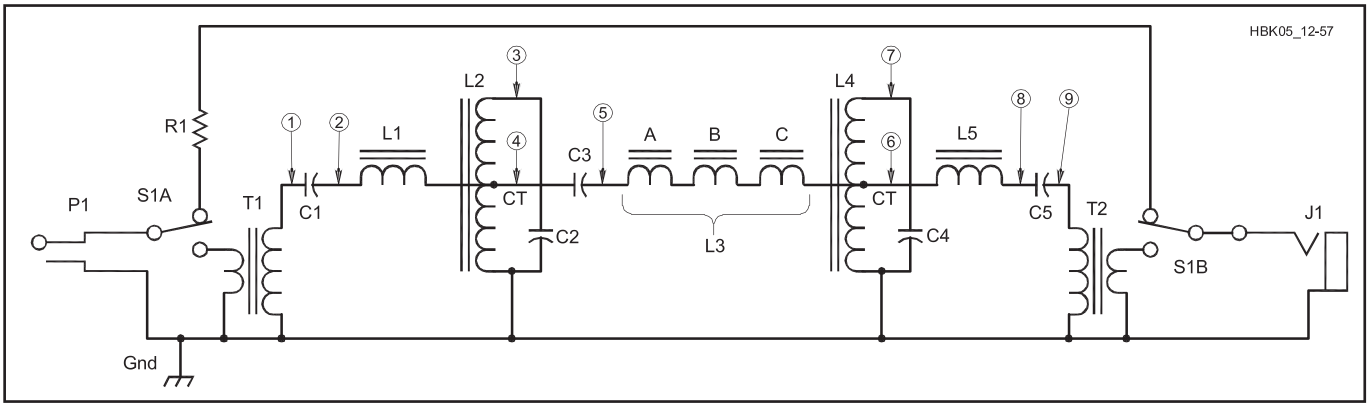


Fig 12.89 — Schematic diagram of the five-resonator CW filter. See Table 12.25 for capacitor and inductor values to build a filter with a center frequency of 546, 600, 700, 750 or 800 Hz.

P1 — Phone plug to match your receiver audio output jack.

J1 — Phone jack to match your headphone.

R1 — 6.8 to 50 Ω, ¼-W, 10% resistor (see text).

S1 — DPDT switch.

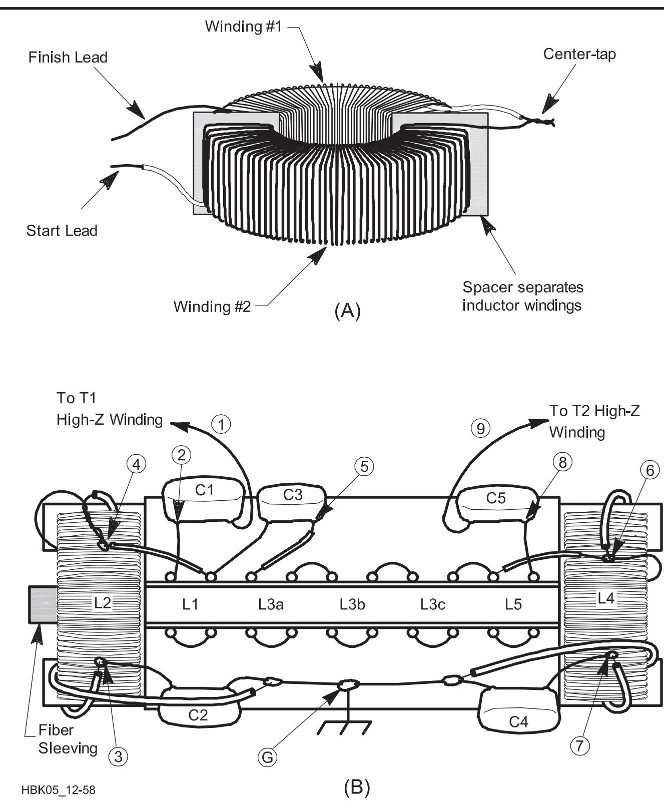
T1, T2 — 200 to 8-Ω impedance-matching transformers, 0.4-W, Miniature Core Type EI-24, Mouser No. 42TU200.

Note: The circled numbers identify the circuit nodes corresponding to the same nodes labeled in the pictorial diagram in Fig 12.90.

the pictorial diagram. The encircled numbers show the circuit nodes corresponding to those in Fig 12.89.

After the correct number of turns are removed from L2 and L4, the leads are gently scraped until you see copper and then the start lead (with sleeving) of one winding is connected to the finish lead of the other winding to make the center tap. The center tap lead and the other start and finish leads of L2 and L4 are connected as indicated in Fig 12.90B. L2 and L4 are fastened to opposite ends of the stack with clear silicone sealant that is available in a small tube at low cost from your local hardware store. Use the silicone sealant to fasten C2 and C4 to the side of the stack. The capacitor leads of C1, C3 and C5 are adequate to support the capacitors when their leads are soldered to the stack terminals. Fig 12.90C is a photo of the assembled filter installed in a Jameco plastic box. Transformers T1 and T2 are secured to the bottom of the plastic box with more silicone sealant and are placed on opposite sides of the DPDT switch. See the photograph for the placement of the phone jack and plug.

After the stack and capacitor wiring is completed, the correctness of the wiring is checked before installing the stack in the box. To do this, check the measured node-to-node resistances of the filter with the values listed in **Table 12.26**.



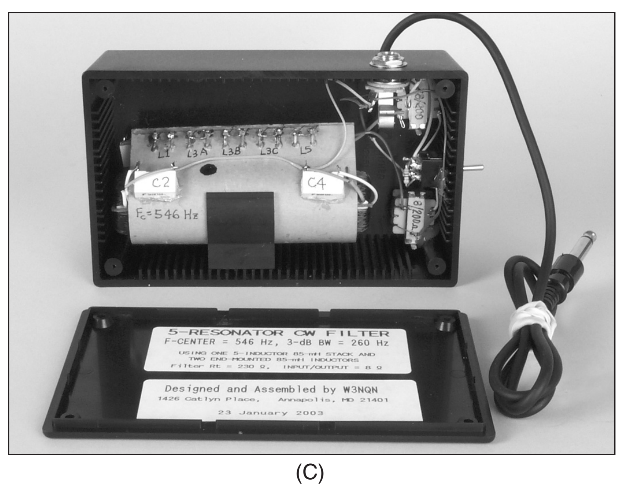


Fig 12.90 — Part A shows a pictorial diagram of the lead-connection details for L2 and L4. Part B shows the filter wiring diagram, including the inductor stack wiring of L1, L3 and L5. Part C is a photo of the assembled filter installed in a Jameco H2581 plastic box. The bypass switch (S1) and input/output transformers (T1, T2) are on the right side of the box.

Table 12.26

Node-to-Node Resistances for the 546-Hz CW Filter

From Node	*To Node*	*Component Designation*	*Resistance (ohms ±20%)*
1	GND	T1 hi-Z winding	12
2	GND	L1 + 1/2(L2)	12
3	GND	L2	8
4	GND	1/2(L2)	4
5	GND	L3 + 1/2(L4)	28
6	GND	1/2(L4)	4
7	GND	L4	8
8	GND	L5 + 1/2(L4)	12
9	GND	T2 hi-Z winding	12
2	4	L1	8
5	6	L3	24
6	8	L5	8
2	3	L1 + 1/2(L2)	12
8	7	L5 + 1/2(L4)	12

Notes

1. See Figs 12.89 and 12.90 for the filter node locations.
2. Check your wiring using the resistance values in this table. If there is a significant difference between your measured values and the table values, you have a wiring error that must be corrected!
3. The resistances of L2 and L4 in the four other filters will be somewhat less than the 546-Hz values.

For accurate measurements, use a high-quality digital ohmmeter.

INTERFACING TO SOURCE AND LOAD

The T1 and T2 transformers match the filter to the receiver low-impedance audio output and to an 8 Ω headset or speaker. If your headset impedance is greater than 200 Ω, omit T2 and connect a ½-watt resistor from node 9 (C5 output lead) to ground. Choose the resistor so the parallel combination of the headset impedance and the resistor gives the correct filter termination impedance (within about 10% of 230 Ω).

PERFORMANCE

The measured 30-dB and 3-dB bandwidths of the 750-Hz filter are about 567 and 271 Hz, respectively. The 30/3-dB shape factor is 2.09. Use this factor to compare the selectivity performance of this filter with others. **Fig 12.91** shows the measured relative attenuation responses of the 546-Hz and 750-Hz filters. These responses were measured in a 200-Ω system without the transformers. All attenuation levels were measured relative to a 0 dB attenuation level at the filter center frequency.

The measured insertion loss of these passive filters with transformers is slightly less than 3 dB and this is typical of filters of this type. This small loss is compensated by slightly increasing the receiver audio gain.

R1 is selected to maintain a relatively constant audio level when the filter is switched in or out of the circuit. The correct value of R1 for your audio system should be determined by experiment and probably will be between 6.8 and 50 Ω. Start with a short circuit across the S1A and B terminals and gradually increase the resistance until the audio level appears to be the same with the filter in or out of the circuit.

Thousands of hams have constructed this five-resonator filter, and many have commented on its ease of assembly, excellent performance and lack of hiss and ringing!

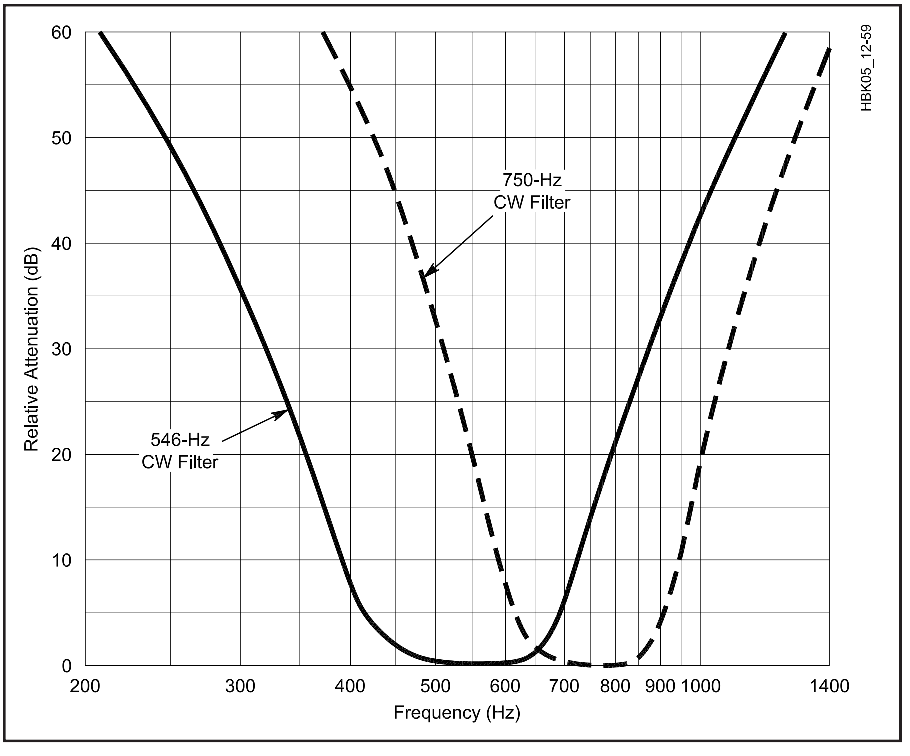


Fig 12.91 — Measured attenuation responses of the 546- and 750-Hz filters. The responses are plotted relative to the zero dB attenuation levels at the center frequencies of the filters. The other filter response curves are similar, but centered at their design frequency.

ORDERING PARTS/CONTACTING THE AUTHOR

The author can provide a kit of parts with detailed instructions for assembling this filter at a nominal cost. The kit includes an inductor stack and two inductors, a pre-punched plastic box with a plastic mounting clip for the inductor stack, five matched capacitors, two transformers, a phone plug and jack and a miniature DPDT switch. Write to Ed Wetherhold, W3NQN, 1426 Catlyn Place, Annapolis, MD 21401-4208 for details about parts and prices. Be sure to include a self-addressed, stamped 9.5 × 4-inch envelope with your request.

Notes

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OTHER FILTER PROJECTS

Filters for specific applications may be found in other chapters of this *Handbook*. Receiver input filters, transmitter filters, interstage filters and others can be separated from the various projects and built for other applications. Since filters are a first line of defense against *electromagnetic interference* (EMI) problems, additional filter projects appear in the **EMI/Direction Finding** chapter

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