Contents

Chapter 11

RF and AF Filters

This chapter discusses the most common types of filters used by radio amateurs. The sections describing basic concepts, lumped element filters and some design examples were initially prepared by Jim Tonne, W4ENE, and updated by Ward Silver, NØAX. The CD-ROM's design example for active filters was provided by Dan Tayloe, N7VE, and the section on crystal filters was developed by Dave Gordon-Smith, G3UUR.

Chapter 11 — CD-ROM Content

Supplemental Articles

- "Using Active Filter Design Tools" by Dan Tayloe, N7VE
- "Crystal Parameter Measurements Simplified", by Chuck Adams, K7QO

Projects

- "An Easy-to-Build, High-Performance Passive CW Filter" by Ed Wetherhold, W3NQN
- "A High Performance, Low Cost 1.8 to 54 MHz Low Pass Filter" by Bill Jones, K8CU

Chapter 17 Software

The following *Windows* software by Jim Tonne, W4ENE, is available from **www.arrl.org/arrl-handbook-reference**.

- *DiplexerDesigner* for design and analysis of diplexer filters
- *Elsie* for design and analysis of lumped-element LC filters
- *Helical* for design and analysis of helical-resonator bandpass filters
- *SVC Filter Designer* for design and analysis of lumped-element high-pass and low-pass filters
- *QuadNet* for design and analysis of active all-pass networks for SSB operation

11.1 Introduction

Electrical filters are circuits used to process signals based on their frequency. For example, most 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. **Table 11.1** shows the usual signal bandwidths for several signal types.

The simplified SSB receiver shown in **Fig 11.1** illustrates the use of several common types of filters. A *preselector* filter is placed between the antenna and the first mixer to pass all frequencies within a given amateur band with low loss. Strong out-of-band signals from broadcast, commercial or military stations are rejected to prevent them from overloading the first mixer. 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 to pass only the desired signal. Finally, an audio filter is placed somewhere between the detector and the speaker rejects unwanted products of detection, power supply hum and noise. The audio

Table 11.1
Typical Filter Bandwidths for Typical Signals

Source	*Required Bandwidth*
Fast-scan analog television (ATV)	4.5 MHz
Broadcast-quality speech and music	15 kHz (from 20 Hz to 15 kHz)
Communications-quality speech	3 kHz (from 300 Hz to 3 kHz)
Slow-scan television (SSTV)	3 kHz (from 300 Hz to 3 kHz)
HF RTTY	250 to 1000 Hz (depends on frequency shift)
Radiotelegraphy (Morse code, CW)	200 to 500 Hz
PSK31 digital modulation	100 Hz

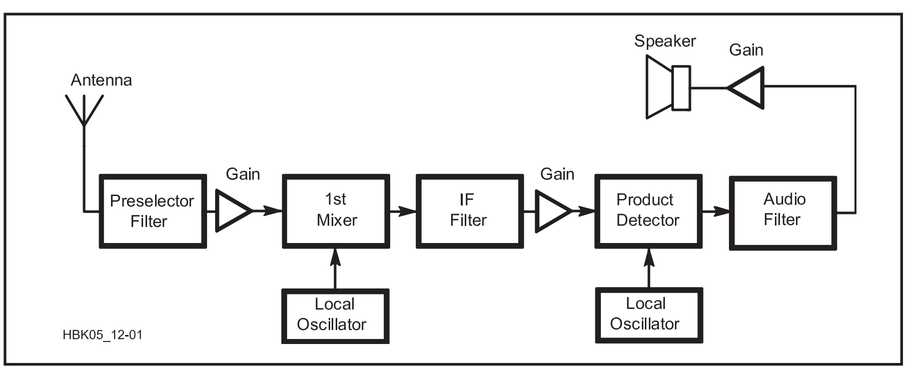


Fig 11.1 — Single-band SSB receiver. At least three filters are used between the antenna and speaker.

filter is often implemented with *active filter* technology.

The complementary SSB transmitter block diagram is shown in **Fig 11.2**. A similar array of filters appears in reverse order. First, an audio filter between the microphone and the balanced mixer rejects out-of-band audio signals such as 60 Hz hum. Since the balanced mixer generates both lower and upper sidebands, an IF filter is placed at the mixer output to pass only the desired lower (or upper) sideband. Finally, a filter at the output of the transmit mixer passes only signals within the amateur band in use rejects unwanted frequencies generated by the mixer to prevent them from being amplified and transmitted. Filters at the transmitter output attenuate the harmonics of the transmitted signals.

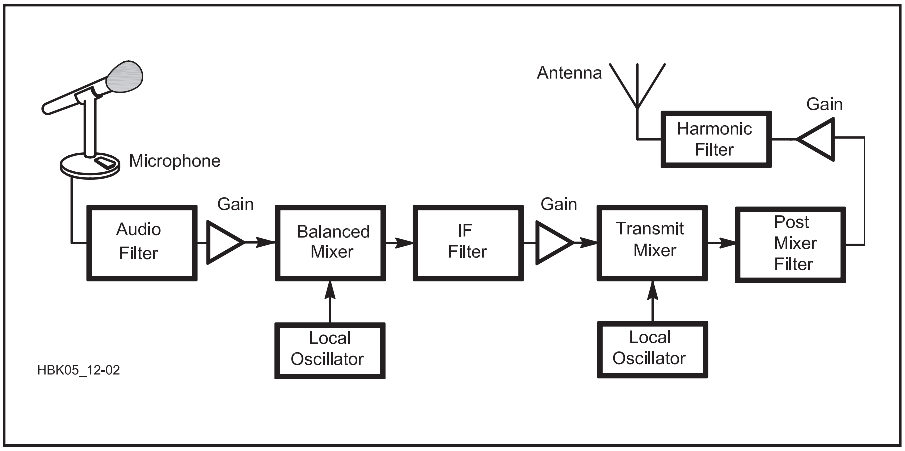


Fig 11.2 — Single-band SSB transmitter. At least three filters are needed to ensure a clean transmitted signal.

11.2 Filter Basics

A common type of filter, the *band-pass*, 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* at which output power is halved (reduced by 3 dB) we would use a *low-pass* filter. To pass signals above a cutoff frequency (also called the "3-dB frequency") 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 requires a *band-stop* or *notch filter*. The region between the passband and the stop band is logically called the *transition region*. The attenuation for signal frequencies in the stop band far from the cutoff frequency is called the *ultimate attenuation*.

Many of the basic terms used to describe filters are defined in the chapters on **Electrical Fundamentals** and **Analog Basics**. In addition, a glossary is provided at the end of this chapter.

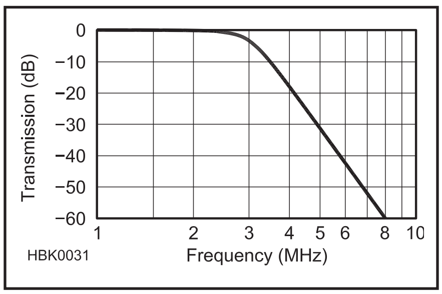


Fig 11.3 — Example of a low-pass magnitude response.

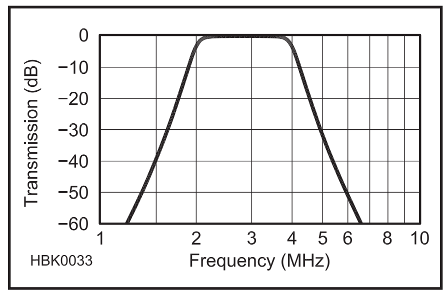


Fig 11.5 — Example of a band-pass magnitude response.

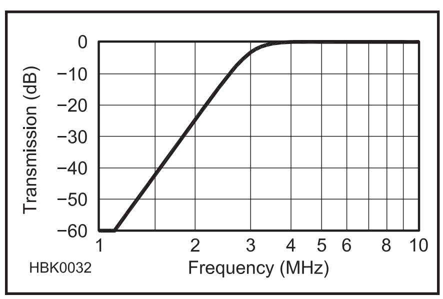


Fig 11.4 — Example of a high-pass magnitude response.

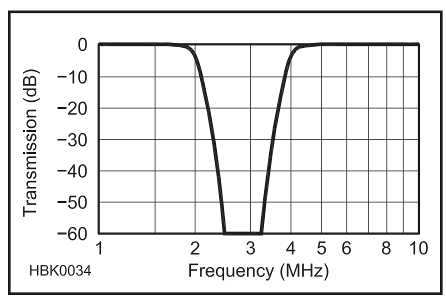


Fig 11.6 — Example of a band-stop magnitude response.

11.2.1 Filter Magnitude Responses

Fig 11.3 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, filter complexity being a major factor.

Of the graphs in this chapter that show a filter's magnitude response, the vertical axes are labeled "Transmission (dB)" with 0 dB at the top of the axis and negative values increasing towards the x-axis. Increasingly negative values of transmission in dB are the same as increasingly positive values of attenuation in dB. For example, –40 dB transmission is the same as 40 dB of attenuation.

Fig 11.4 illustrates the magnitude response of a high-pass filter. Signals above the 3 MHz 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 11.5 illustrates the magnitude response of a band-pass filter. Signals within the band-pass range (between the lower and upper cutoff frequencies) are passed with minimum attenuation while signals outside that range are attenuated. In this example the filter was designed with cutoff frequencies of 2 MHz and 4 MHz, for a passband width of 2 MHz.

Fig 11.6 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 a type of band-stop filter with a narrow stop-band in which the attenuation is a maximum at a single frequency.

An *ideal filter* — a low-pass filter, for example — would pass all frequencies up to some point with no attenuation at all and

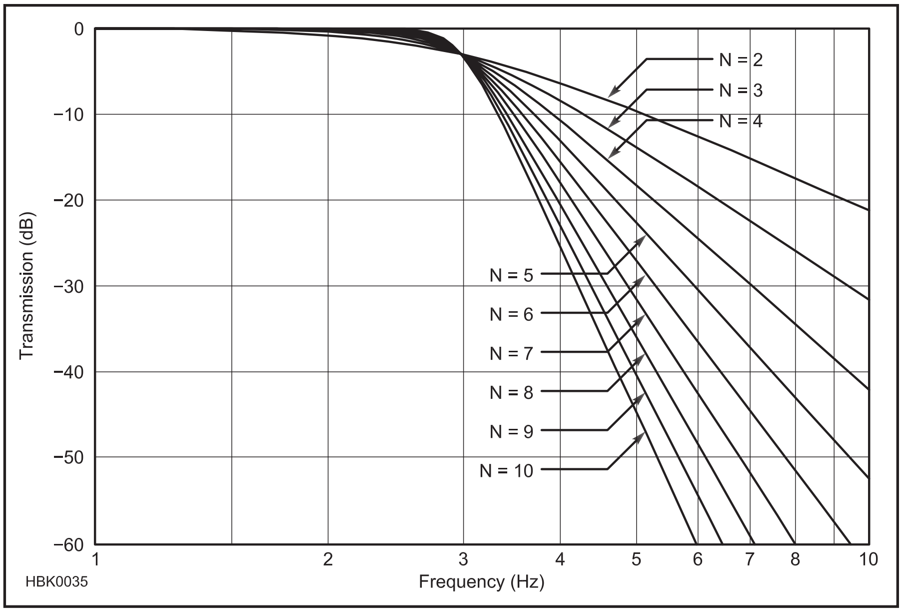


Fig 11.7 — Magnitude responses for low-pass filters with orders ranging from very simple (N=2) to very complex (N=10).

totally reject everything beyond that point. This is known as a *brick wall* response because the filter's passband and stop band are *flat*, meaning no attenuation, and the rolloff in the transition region is infinitely steep. The magnitude response of such a filter would be drawn as a flat line representing 0 dB attenuation up to the cutoff frequency that then abruptly changes at the cutoff frequency to a flat line at infinite attenuation throughout the stop band to infinite frequency.

11.2.2 Filter Order

The steepness of the descent from the passband to the region of attenuation — the stop band — is dependent on the complexity of the filter, properly called the *order.* In a lumped element filter made from inductors and capacitors, the order is determined by the number of separate energy-storing elements (either L or C) in the filter. For example, an RC filter with one resistor and one capacitor has an order of 1. A notch filter made of a single series-LC circuit has an order of 2. The definition of order for active and digital signal processing filters can be more complex, but the general understanding remains valid that the order of a filter determines how rapidly its response can change with frequency.

For example, **Fig 11.7** shows the magnitude response of one type of low-pass filter with orders varying from very simple ("N=2") to the more complex ("N=10"). In each case on this plot the 3-dB point (the filter's cutoff or corner frequency) remained the same, at 3 Hz.

As frequency increases through the transition region from the passband into the stop band, the attenuation increases. *Rolloff*, the rate of change of the attenuation for frequencies above cutoff, is expressed in terms of dB of change per octave of frequency. Rolloff for this type of filter design is always 6.02 times the order, regardless of the circuit used to fabricate the circuit. As the order goes up, the rate of change also goes up. A filter's response with large values of rolloff is called "sharp" or "steep" and "soft" or "shallow" if rolloff is low.

11.2.3 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 engineer responsible for defining their behavior mathematically.

Real filters respond more gradually and with greater variation than ideal filters. Two primary traits of the most importance to amateurs are used to describe the behavior of filter families: *ripple* (variations in the magnitude response within the passband and stop band) and rolloff. Different families of filters have different degrees of ripple and rolloff (and other characteristics, such as phase response).

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 both **Fig 11.7** and **Fig 11.8**. (Most filter discussions are based on low-pass filters because the same concepts are easily extended to other types of filters and much of the mathematics behind filters is equivalent for the various types of frequency responses.)

By allowing magnitude response ripples in the passband, we can get a somewhat steeper rolloff from the passband into the stop band, particularly just beyond the cutoff frequency. A family that does this is the *Chebyshev* family. **Fig 11.9** illustrates how allowing magnitude ripple in the passband provides a sharper filter rolloff. This plot compares the 1-dB ripple Chebyshev with the no-ripple Butterworth filter down to 12 dB of attenuation.

Fig 11.9 also illustrates the usual definition of bandwidth for a low-pass Butterworth filter — the filter's 3-dB or cutoff frequency. For a low-pass Chebyshev filter, *ripple bandwidth* is used — the frequency range over which the filter's passband ripple is no greater than the specified limit. For example, the ripple bandwidth of a low-pass Chebyshev filter designed to have 1 dB of passband ripple is the highest frequency at which attenuation is 1 dB or less. The 3-dB bandwidth of the frequency of the filter will be somewhat greater.

A Butterworth filter is defined by specifying the order and bandwidth. The Chebyshev filter is defined by specifying the order, the ripple bandwidth, and the amount of passband ripple. In Fig 11.9, the Chebyshev filter has 1 dB of ripple; its ripple bandwidth is 1000 Hz.

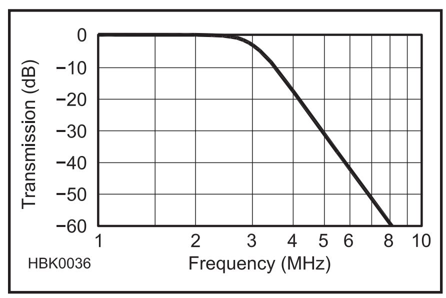


Fig 11.8 — Filters from the Butterworth family exhibit flat magnitude response in the passband.

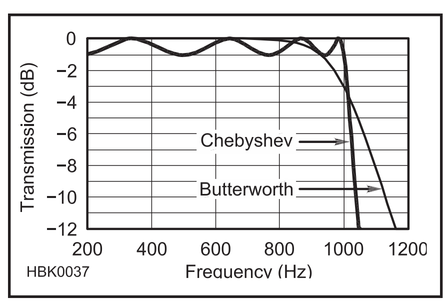


Fig 11.9 — This plot compares the response of a Chebyshev filter with a 1-dB ripple bandwidth of 1000 Hz and a Butterworth filter 3-dB bandwidth of 1000 Hz.

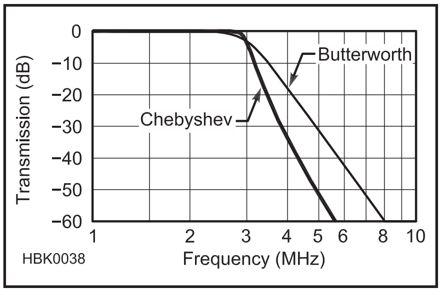


Fig 11.10 — A Chebyshev filter (0.2 dB passband ripple) allows a sharper cutoff than a Butterworth design with no passband ripple.

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 increasing a filter's rolloff. **Fig 11.10** 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. (For this comparison the cutoff frequencies at 3 dB of attenuation are the same for each filter.) Even that small amount of ripple in the Chebyshev filter passband allows a noticeably steeper rolloff between the passband and the stop band. As the passband ripple specification is increased, the steepness of the transition from passband to stop band increases, compared to the Butterworth family although the rolloff of the two families eventually becomes equal.

For Chebyshev filters, when the value of the passband ripple is changed, the magnitude response in the stop band region also changes. **Fig 11.11** compares the stop band response of Chebyshev filters with passband ripple ranging from 0.01 to 1 dB.

It is possible to obtain even steeper rolloff into the stop band by adding "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 done, we have a *Cauer* family design (also called the *elliptic-function* design). The rolloff of the Cauer filter from the passband into the stop band is the steepest of all analog filter types provided that the behavior in the passband is uniform (either no ripple or a uniform amount of ripple). **Fig 11.12** 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. Again, an alternative bandwidth definition is to use the 3-dB frequency instead of the ripple bandwidth. The Cauer family requires one more specification: the *stop band frequency* and/or the *stop band depth*. The stop band frequency is the lowest frequency of a null or notch in the stop band. The stop band depth is the minimum amount of attenuation allowed in the stop band. In Fig 11.12, the stop band frequency is about 3.5 MHz and the stop band depth is 50 dB. For this comparison both designs have the same passband ripple value of 0.2 dB.

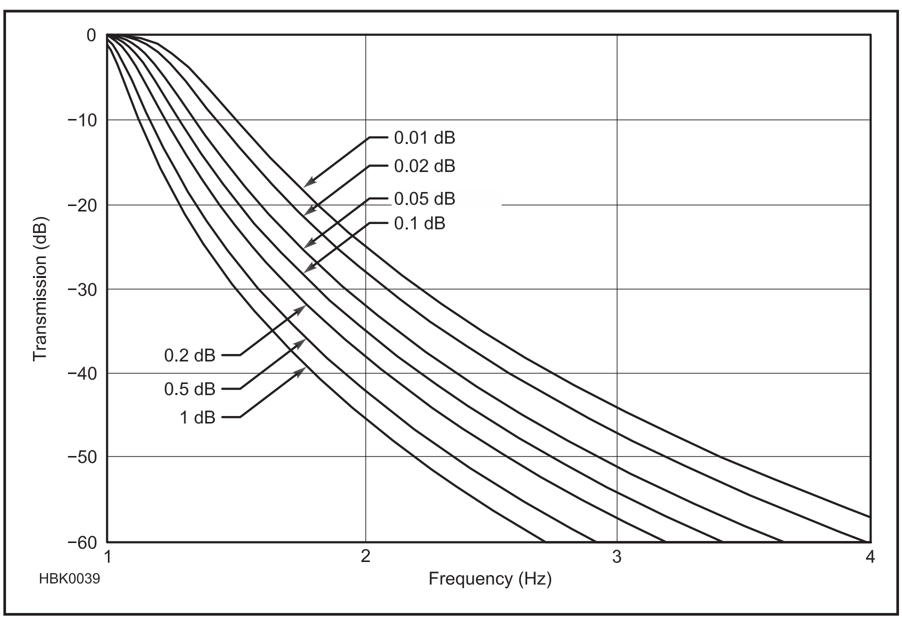


Fig 11.11 — These stop band response plots illustrate the Chebyshev family with various values of passband ripple. These plots are for a seventh-order low-pass design with ripple values from 0.01 to 1 dB. Ignoring the effects in the passband of high ripple values, increasing the ripple will allow somewhat steeper rolloff into the stop band area, and better ultimate attenuation in the stop band.

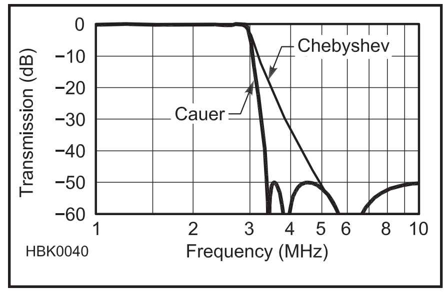


Fig 11.12 — The Cauer family has an even steeper rolloff from the passband into the stop band than the Chebyshev family. Note that the ultimate attenuation in the stop band is a design parameter rather than ever-increasing as is the case with other families.

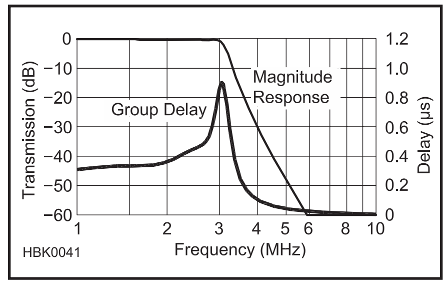


Fig 11.13 — Magnitude response and group delay of a Chebyshev low-pass filter.

A downside to the Cauer filter is that the ultimate attenuation is some chosen value rather than ever increasing, as is the case with the other families. In the far stop band region (where frequency is much greater than the cutoff frequency) the rolloff of Cauer filters ultimately reaches 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.

11.2.4 Group Delay

Another trait that can influence the choice of which filter family to use is the way the transit time of signals through the filter varies with frequency. This is known as *group delay*. ("Group" refers to a group of waves of similar frequency and phase moving through a media, in this case, the filter.) The wide line (lower plot) in **Fig 11.13** illustrates the group delay characteristics of a Chebyshev low-pass filter, while the upper plot (narrow line) shows the magnitude response. As shown in Fig 11.13, the group delay of components near the cutoff frequency becomes quite large when compared to that of components at lower and higher frequencies. This is a result of the phase shift of the filter's transmission being nonlinear with frequency; it is usually greater

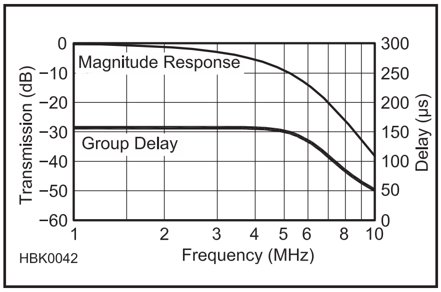


Fig 11.14 — Magnitude response and group delay of a Bessel low-pass filter.

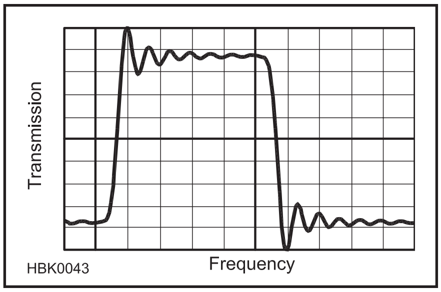


Fig 11.15 — Transient response of a Chebyshev low-pass filter.

near the cutoff frequency. By delaying signals at different frequencies different amounts, the signal components are "smeared" in time. This causes distortion of the signal and can seriously disrupt high-speed data signals.

If a uniform group 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 rolloff of the magnitude response may need to be taken into account. The magnitude and delay characteristics for the Bessel family are shown in **Fig 11.14**. Comparing the Bessel filter's response to that of the Butterworth in Fig 11.8 shows the difference in roll off.

A downside of Bessel family filters is that the rolloff characteristic is quite poor; they are not very good as a magnitude-response-shaping filter. The Bessel family is characterized largely by its constant group delay in the passband (for a low- or high-pass filter) shown as the bottom plot in Fig 11.14. The *constant-delay* characteristic of the Bessel extends into the stop band. The Bessel filter bandwidth is commonly defined by its 3-dB point (as with the Butterworth).

Because the Bessel filter is used when phase response is important, it is often characterized by the frequency at which a specific amount of phase shift occurs, usually one radian. (One radian is equal to $360 / 2\pi = 57.3°$.) Since the delay causes the output signal to lag behind the input signal, the filter is specified by its *one-radian lag frequency*.

11.2.5 Transient Response

Some applications require that a signal with sharp rising and falling edges (such as a digital data waveform) applied to the input of a low-pass design have a minimum *overshoot* or *ringing* as seen at the filter's output. Overshoot (and undershoot) occurs when a signal exceeds (falls below) the final amplitude temporarily before settling at its final value. Ringing is a repeated sequence of overshoot and undershoot. If a sharp-cutoff filter is used in such an application, overshoot or ringing will occur. The appearance of a signal such as a square wave with sharp rising and falling edges as it exits from the filter may be as shown in **Fig 11.15**. The scales for both the X- and the Y-axes would depend on the frequency and magnitude of the waveform.

The 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 order are used to create the square wave — that is, if the square wave is passed through a sharp-cutoff low-pass filter that attenuates higher frequencies — then that waveform will have the overshoot or ringing as shown as it exits from the filter.

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 *passive* lumped-element inductors and capacitors or using op amps with resistors and capacitors (an *active* filter). The traits are general descriptions of filter behavior and can be applied to any type of filter technology.

11.2.6 Filter Family Summary

Selecting a filter family is one of the first steps in filter design. To make that choice easier, the following list of filter family attributes is provided:

• Butterworth — No ripple in passband, smooth transition region, shallow rolloff for a given filter order, high ultimate attenuation, smooth group delay change across transition region

• Chebyshev — Some passband ripple, abrupt transition region, steep rolloff for a given filter order, peak in group delay near cutoff frequency, high ultimate attenuation

• Cauer (or Elliptical-function) — Some passband ripple, abrupt transition region, steepest roll off, ripple in stop band due to traps, ultimate attenuation smaller than Butterworth and Chebyshev, group delay peaks near cutoff frequency and in stop band

• Bessel — No ripple in passband, smooth transition region, constant group delay in passband, shallow rolloff for a given filter order, smooth changes in group delay, high ultimate attenuation

All characterizations such as "steep" and "abrupt" are relative with respect to filter designs from other families with similar orders. Other factors, such as number of components, sensitivity to component value and so on may need to be considered when selecting a filter family for a specific application.

11.3 Lumped-Element Filters

This part of the chapter deals with passive LC filters fabricated using discrete inductors and capacitors (which gives rise to their name, *lumped-element*). We will begin with a discussion of basic low-pass filters and then generalize to other types of filters.

11.3.1 Low-Pass Filters

A very basic LC filter built using inductors and capacitors is shown in **Fig 11.16**. In the first case (Fig 11.16A), 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 11.16B), 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.

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 pairs of series and parallel tuned circuits. These filters, made from alternating LC elements or LC tuned circuits, are called *ladder filters*.

We spoke of filter order, or complexity, earlier in this chapter. **Fig 11.17** illustrates *capacitor-input* low-pass filters with orders of 3, 4 and 5. Remember that the order corresponds to the number of energy-storing elements. For example, the third-order filter

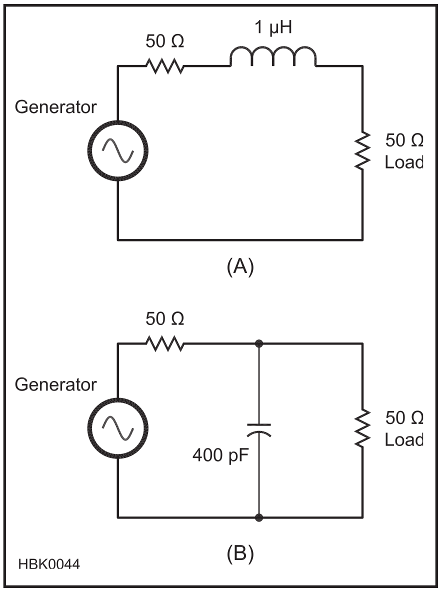


Fig 11.16 — A basic low-pass filter can be formed using a series inductor (A) or a shunt capacitor (B).

in Fig 11.17A has three energy-storing elements (two capacitors and one inductor), while the fifth-order design in Fig 11.17C has five elements total (three capacitors, two inductors). For comparison, a third-order filter is illustrated in **Fig 11.18**. The filters in Fig 11.17 are *capacitor-input* filters because a capacitor is connected directly across the input source. The filter in Fig 11.18 is an *inductor-input* filter.

As mentioned previously, the Cauer family has traps (series or parallel tuned circuits) 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 11.19**. 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 low impedance in the stop band while the inductor-input filter has high impedance in the stop band.

11.3.2 High-Pass Filters

A high-pass filter passes signals above its cutoff frequency and attenuates those below. Simple high-pass equivalents of the filters in Fig 11.16 are shown in **Fig 11.20.**

The reactance of the series capacitor in Fig 11.20A 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 11.20B decreases at lower frequency, with the same effect. Similarly to the low-pass filter designs presented in Fig 11.17, a high-pass filter in a real world design would typically use both series and shunt components but with the positions of inductors and capacitors exchanged.

An example of a high-pass filter application would be a broadcast-reject filter designed to pass amateur-band signals in the range of 3.5 MHz and above while rejecting broadcast signals at 1.7 MHz and below. A high-pass filter with a design cutoff of 2 MHz is illustrated in **Fig 11.21**. It can be implemented as a capacitor-input (Fig 11.21A) or inductor-input (Fig 11.21B) 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.

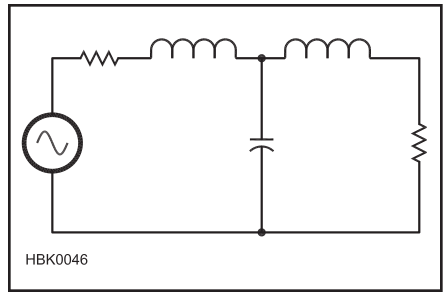


Fig 11.18 — A third-order inductor-input low-pass filter.

11.3.3 Low-Pass to Band-Pass Transformation

A band-pass filter is defined in part by a *bandwidth* and a *center frequency*. (An alternative method is to specify a lower and an upper cutoff frequency.) A low-pass design such as the one shown in Fig 11.17A can be converted to a band-pass filter by resonating each of the elements at the center frequency. **Fig 11.22** 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 11.23**. 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 11.24**. 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

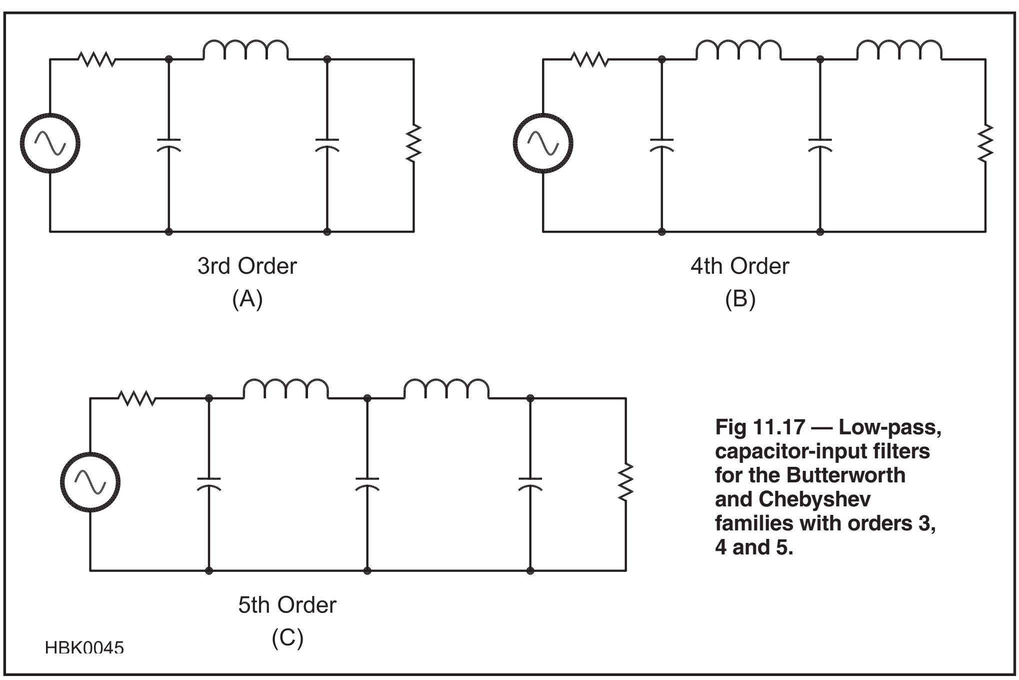


Fig 11.17 — Low-pass, capacitor-input filters for the Butterworth and Chebyshev families with orders 3, 4 and 5.

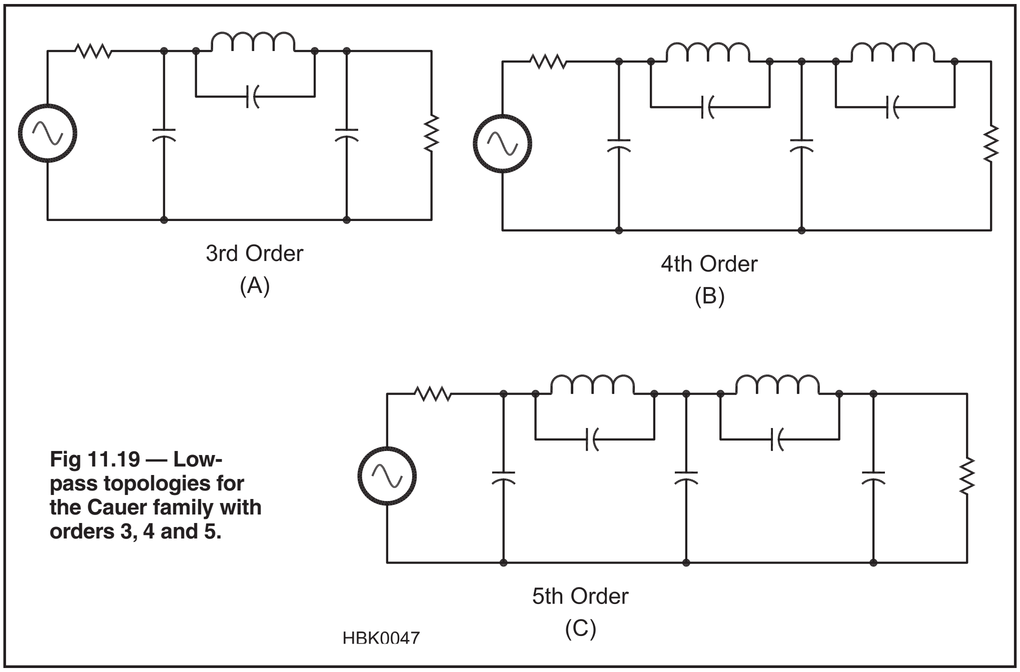


Fig 11.19 — Low-pass topologies for the Cauer family with orders 3, 4 and 5.

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).

11.3.4 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. The bandwidth of a band-stop filter is measured between the frequencies at which the magnitude response drops 3 dB in the transition region into the stop band.

Fig 11.25 shows how to convert the 2 MHz capacitor-input high-pass filter of Fig 11.21A 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 11.26** shows the responses of both the original high-pass and the resulting band-stop filter.

11.3.5 Refinements in Band-Pass Design

The method of transforming a low-pass filter to a band-pass filter described previously is

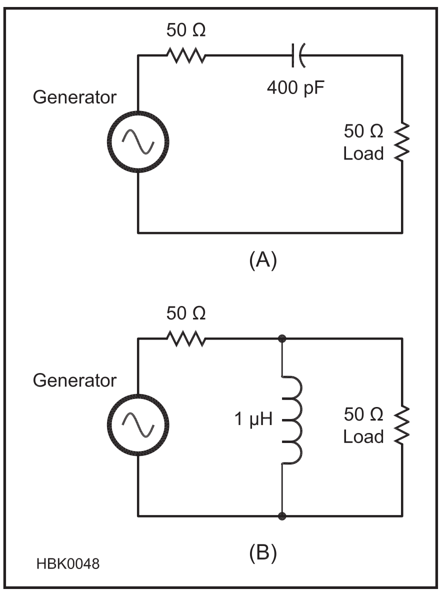


Fig 11.20 — A basic high-pass filter can be formed using a series capacitor (A) or a shunt inductor (B).

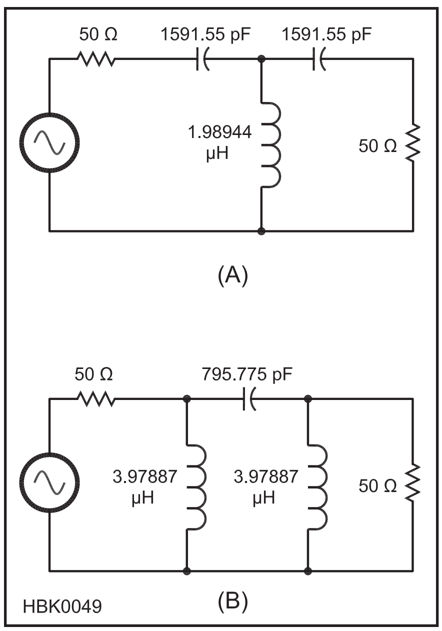


Fig 11.21 — Capacitor-input (A) and inductor-input (B) high-pass filters. Both designs have a 2 MHz cutoff.

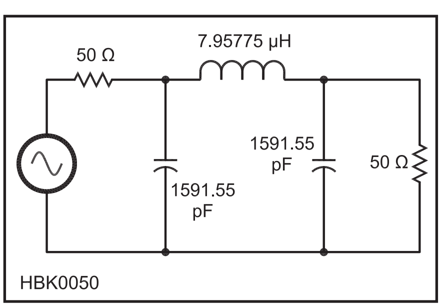


Fig 11.22 — Third-order low-pass filter with a bandwidth of 2 MHz in a 50 Ω system.

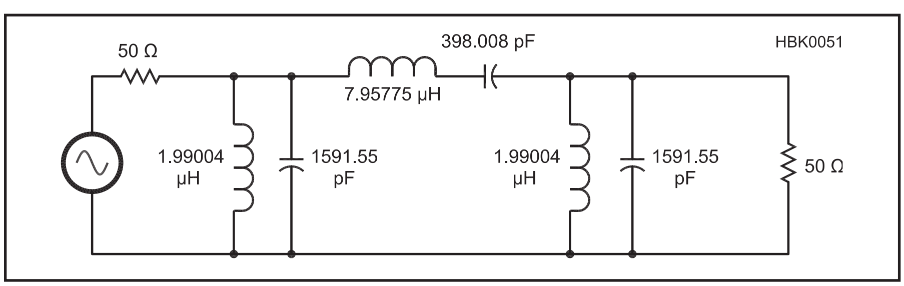


Fig 11.23 — The low-pass filter of Fig 11.22 can be transformed to a band-pass filter by resonating the shunt capacitors with a parallel inductor and resonating the series inductor with a series capacitor. Bandwidth is 2 MHz and the center frequency is 2.828 MHz.

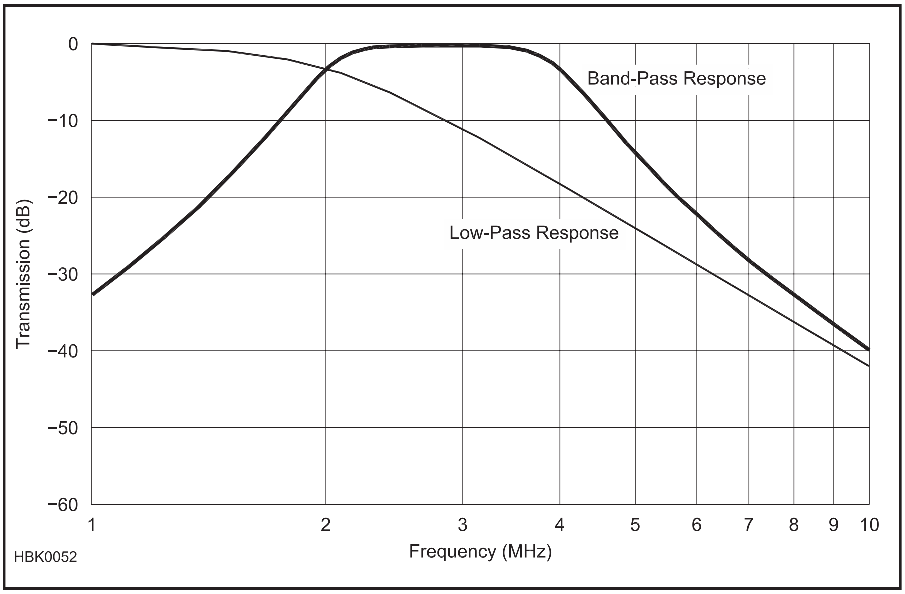


Fig 11.24 — Comparison of the response of the low-pass filter of Fig 11.22 with the band-pass design of Fig 11.23.

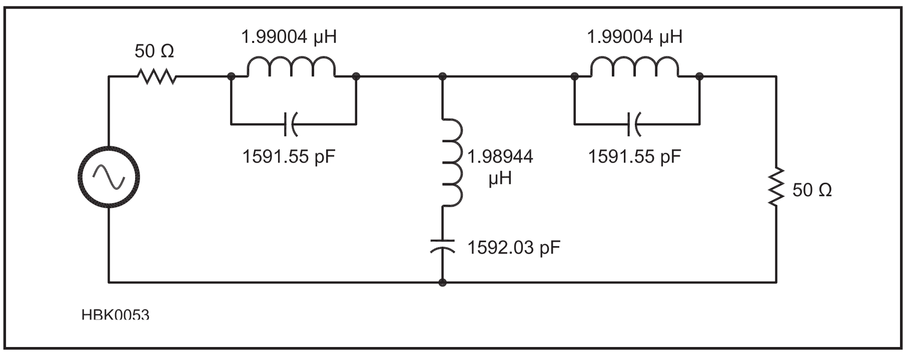


Fig 11.25 — The high-pass filter of Fig 11.21A can be converted to a band-stop filter with a bandwidth of 2 MHz, centered at 2.828 MHz.

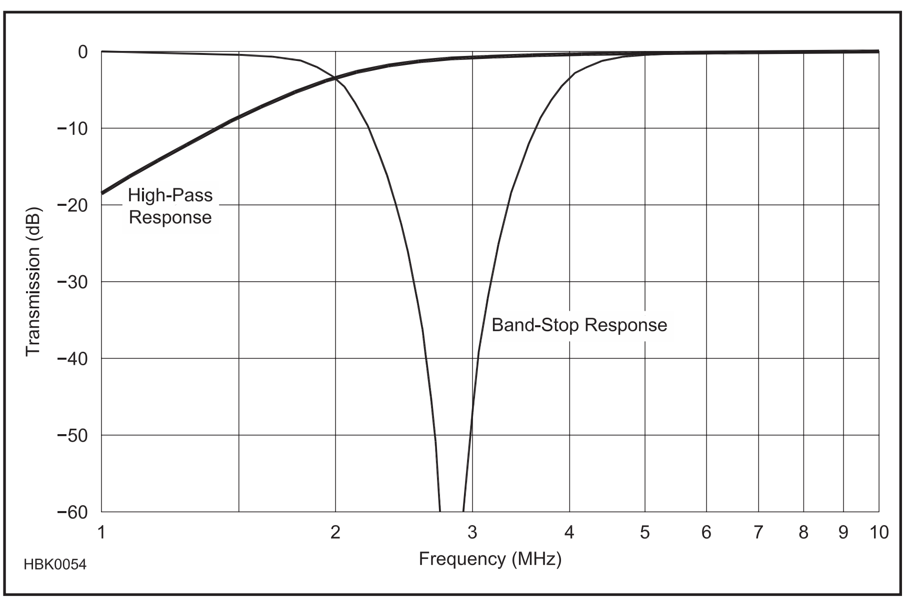


Fig 11.26 — Comparison of the response of the high-pass filter of Fig 11.21A with the band-stop design of Fig 11.25.

simple, but it has a major drawback. The resulting component values are often awkward for building a narrowband band-pass filter or inappropriate for the frequencies involved. This can be addressed through additional transformation techniques or by changing the filter to a different *topology* (the general organization of the filter circuit elements and how they are connected).

The following discussion shows an example of band-pass filter design, starting with a low-pass filter and then applying various techniques to create a filter with appropriate component values and improved response characteristics. These transformations are tedious and are best done using filter design software. Some manual design methods are presented later in this chapter.

Fig 11.27 shows a band-pass filter centered at 3 MHz with a width of 100 kHz that was designed by resonating the elements of a 3-MHz low-pass ladder-type filter as 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 the series inductance likely to be present in a practical capacitor of that value. Similarly, because of parasitic capacitance, the series inductor (159 µH) will certainly have a parallel self-resonance that 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. (See the **RF Techniques** chapter for a discussion of parasitic effects.)

A better approach to narrowband band-pass filter design may be found by changing the filter topology. There are many different filter topologies, such as the LC "ladder" filters discussed in the preceding sections. One such topology is the *nodal-capacitor-coupled* design and **Fig 11.28** shows how the filter in Fig 11.27 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. In this case, the technique of *impedance scaling* offers some improvement. The filter was originally designed for a 50-Ω system. By scaling the filter impedance upwards from 50 to 500 Ω, the reactances of all elements are multiplied by a factor of 500 / 50 = 10. (Capacitors will get smaller and inductors larger.) 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 11.28 is shown in **Fig 11.29**.

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 components

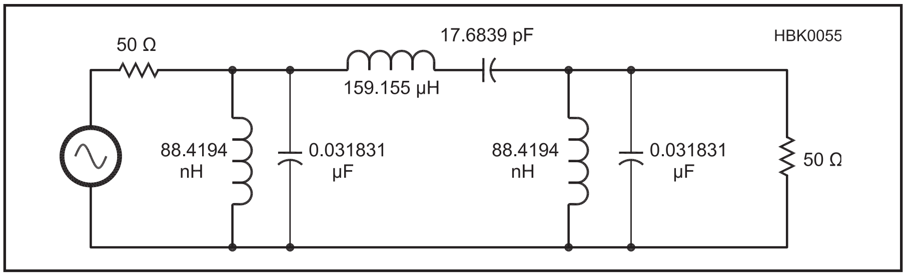


Fig 11.27 — This band-pass filter, centered at 3 MHz with a bandwidth of 100 kHz, was designed using a simple transformation. The resulting component values make the design impractical, as discussed in the text.

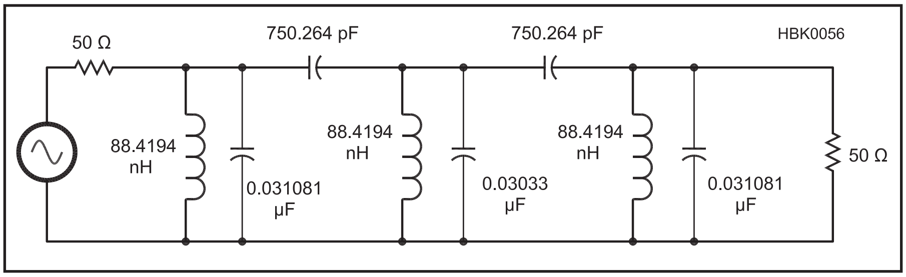


Fig 11.28 — A narrowband band-pass filter using a nodal-capacitor-coupled design improves component values somewhat.

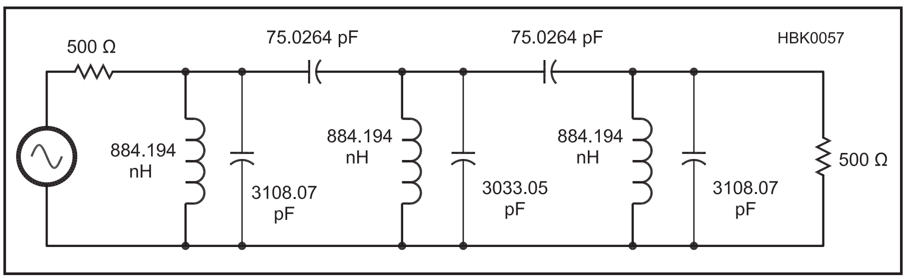


Fig 11.29 — The filter in Fig 11.28 scaled to an impedance of 500 Ω.

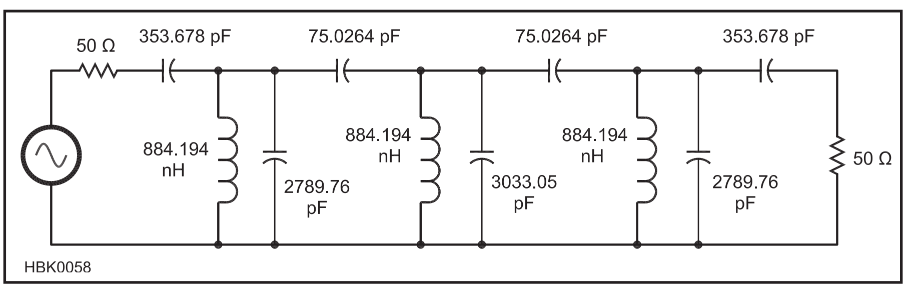


Fig 11.30 — Final design of the narrowband band-pass filter with 50 Ω terminations.

must be added as shown in **Fig 11.30**. This topology is typical of a radio receiver front-end preselector or anywhere else that a narrow — in terms of percentage of the center frequency — filter is needed.

When the nodal capacitor-coupled band-pass topology is used for a filter whose bandwidth is wide (generally, a bandwidth 20% or more of the center frequency, f_0, resulting in a filter Q = f_0 / BW less than 5), the attenuation at frequencies above the center frequency will be less than below the center frequency. This characteristic should be taken into account when attenuation of harmonics of signals in the passband is of concern. The filter shown in **Fig 11.31A** 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 11.31B. Another way to accomplish the task is by going to a *mesh capacitor-coupled* design as shown in Fig 11.31C. The last two designs have identical magnitude responses.

Two other topologies are shown for comparison. The design in Fig 11.31B has only three capacitors (a minimum-capacitor design) while the design in Fig 11.31C has only three inductors (a minimum-inductor design).

The magnitude response of the capacitor-coupled design in Fig 11.31A is shown in **Fig 11.32** as the solid line. The designs in Fig 11.31B and C have magnitude responses as shown in Fig 11.32 as a dashed line. Note that the rolloff on the high-frequency side is steeper for the designs in Fig 11.31B 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 also improve image rejection in a receiver RF stage where high-side injection of the local oscillator is used.

Try *Elsie* for Filter Design!

Elsie, a *Windows* program for design and analysis of lumped-element LC filters, is available from **www.arrl.org/arrl-handbook-reference**. 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, W4ENE, who wrote the introductory section of this chapter.

11.3.6 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." As described in the **Electrical Fundamentals** chapter, Q is a measure of the loss in an inductor or capacitor, as determined by its resistive component. Q is the ratio of component's reactance to the loss resistance and is specified at a given test frequency. The loss resistance referred to here includes not only the value as measured by an ohmmeter but includes all sources of loss, such as skin effect, dielectric heating and so on.

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 *insertion loss* of signals passing through the filter and

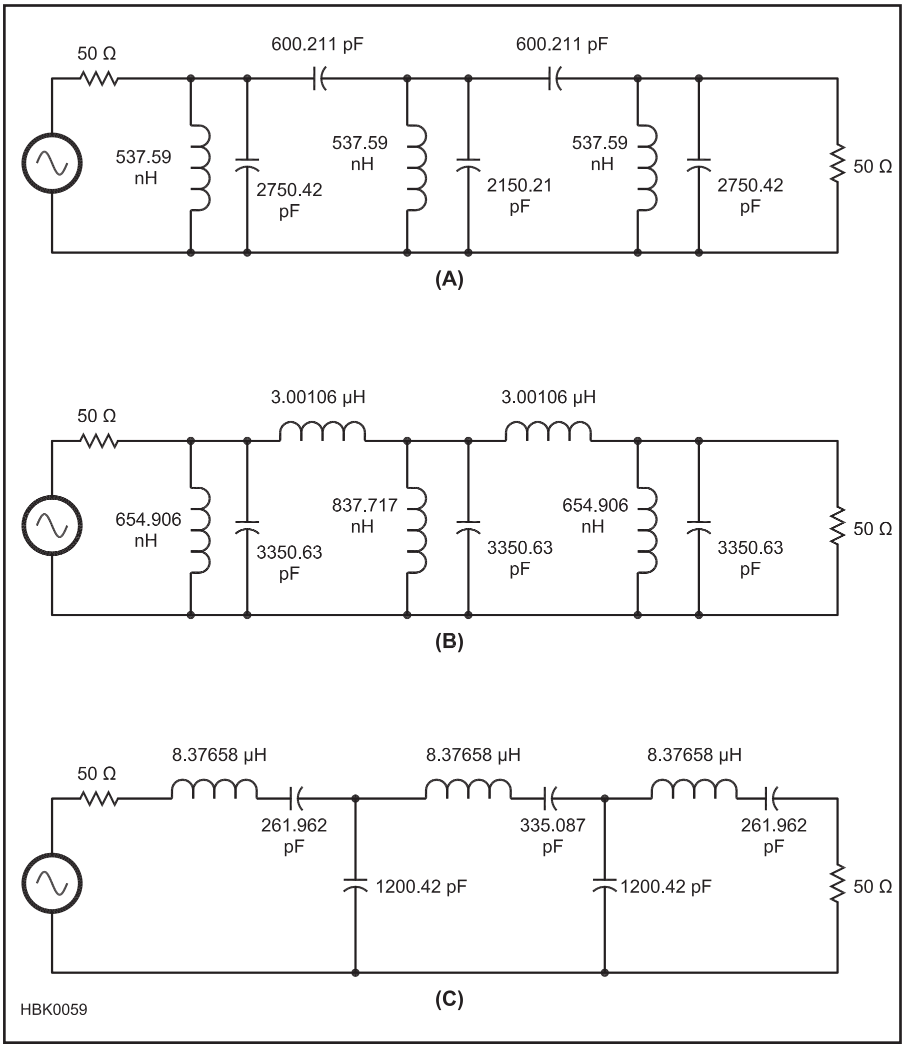


Fig 11.31 — Three band-pass filters designed to pass the 75 meter amateur band. The nodal-capacitor-coupled topology is shown at A, nodal-inductor-coupled at B, and mesh capacitor-coupled at C.

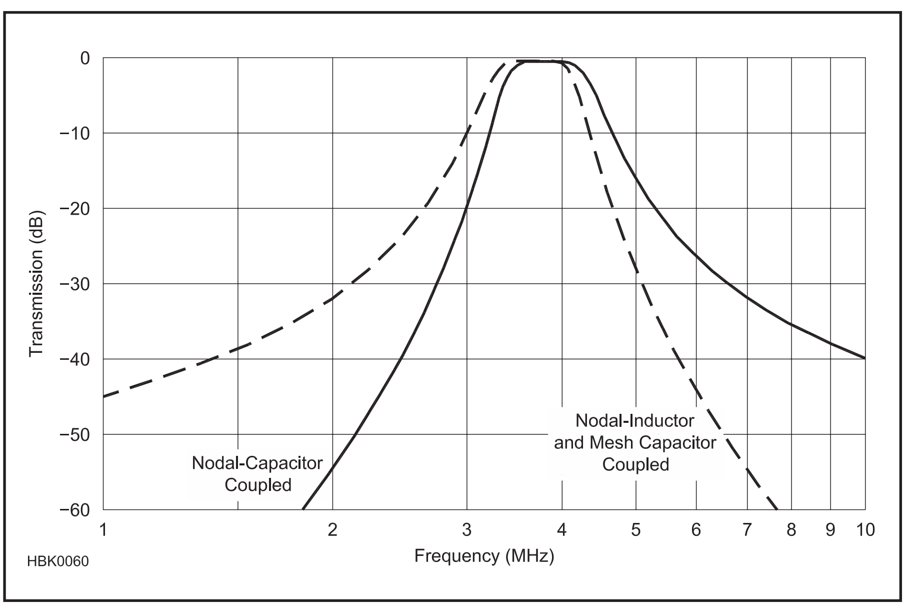


Fig 11.32 — Responses of the filters in Fig 11.31.

the steepness of the filter's rolloff. Band-pass filters (and especially narrowband band-pass filters) are more vulnerable to this problem than low-pass and high-pass filters. **Fig 11.33** 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 compromise the filter's response at cutoff, especially problematic in the case of narrowband band-pass filters. **Fig 11.34** 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 response shape are much higher than those required for the low-pass or high-pass filter (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.

11.3.7 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

Table 11.2
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 find the passband ripple that should be used to design the filter.

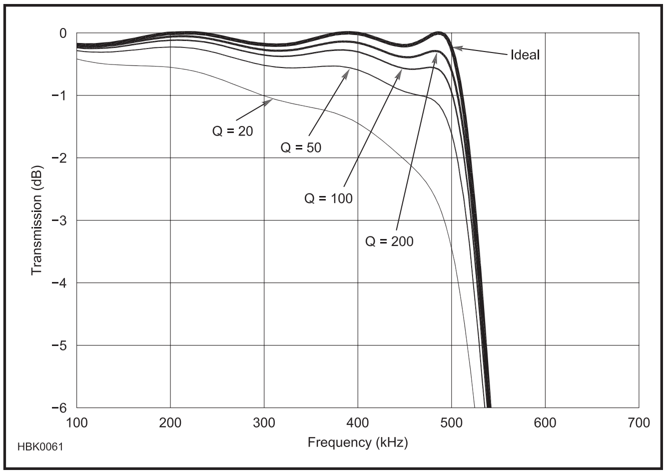


Fig 11.33 — Effect of inductor Q values on a low-pass filter.

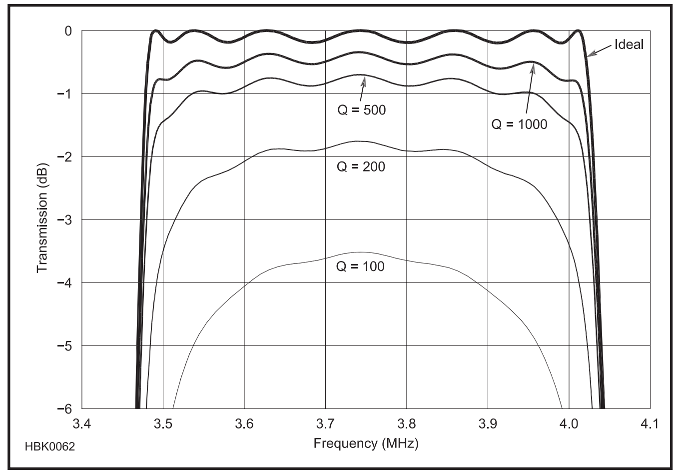


Fig 11.34 — Effect of inductor Q values on a narrowband band-pass filter.

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. (*Return loss* is explained in the **RF Techniques** chapter.)

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 11.2 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.

11.4 Filter Design Examples

Each topology of filter has its own mathematical techniques to determine the filter components, given the frequencies and attenuations that define the filter behavior. These techniques are very precise, involving the use of tables and scaling techniques as discussed in the following sections. With the availability of filter design software, such as *Elsie* and *SVC Filter Designer* (both available from **www.arrl.org/arrl-handbook-reference**) filters using different topologies can be designed easily, however. (Additional programs and design resources are listed in the References section at the end of this chapter.)

Whether designing a filter manually or through the use of software, the design process begins with developing a set of filter *performance requirements* for:

• Type of response — low-pass, high-pass, band-pass, band-stop, or all-pass.

• Passband ripple — none or some maximum amount.

• Cutoff frequency (or frequencies) or bandwidth.

• Steepness of rolloff in transition region.

• Minimum stop-band attenuation and stop-band depth (optional).

• Group delay (optional).

The requirements shown as optional may not be important for simple filters. The next step is to identify the filter families that can meet those requirements. For example, if no ripple is allowed in the passband, the Butterworth or Bessel filter families would be suitable. If a very steep rolloff is required, try the Chebyshev or Cauer family.

After entering the basic requirements into the filter design software and observing the results of the calculation, the design is likely to require some adjustment. Component values can be unrealistic or the required order may be too high for practical implementation. At this point, your experience with filter design helps guide the choices of what changes to make — a different filter family, raising or lowering some of your requirements and so forth. The key is to experiment and observe to build understanding of filters.

Filter design and analysis programs show the responses of the resulting network in graphic form. This makes it easy to compare your filter to your requirements, making design adjustments quickly. Many design packages also allow the selection of nearest-5% values, tuning and other useful features. For example, the program *SVC Filter Designer* (SVC is *standard value component*) 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).

Although filter-design software has greatly simplified the design of lumped-element filters, going through the manual design process, which is based on the use of tables of component values, will give some insight into how to make better use the software to develop and refine a filter.

SVC Filter Design Software

SVC Filter Designer, a *Windows* program for the design of lumped-element high-pass and low-pass filters by Jim Tonne, W4ENE, is available from **www.arrl.org/arrl-handbook-reference**. 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.

11.4.1 Normalized Values

The equations used to design filters are quite complex, so to ease the design process, tables of component values have been developed for different families of filters. Developing a set of

Table 11.3
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 11.4
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 11.5
Chebyshev Normalized Values
Passband ripple 0.01 dB

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 11.6
Chebyshev Normalized Values
Passband ripple 0.044 dB

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 11.7
Chebyshev Normalized Values
Passband ripple 0.2 dB

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

Table 11.8
Cauer Normalized Values
Passband ripple 0.01 dB, Stop band depth 30 dB

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 11.9
Cauer Normalized Values
Passband ripple 0.01 dB, Stop band depth 40 dB

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 11.10
Cauer Normalized Values
Passband ripple 0.01 dB, Stop band depth 50 dB

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

Table 11.11
Cauer Normalized Values
Passband ripple 0.01 dB, Stop band depth 60 dB

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 11.12
Cauer Normalized Values
Passband ripple 0.044 dB, Stop band depth 30 dB

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 11.13
Cauer Normalized Values
Passband ripple 0.044 dB, Stop band depth 40 dB

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

Table 11.14
Cauer Normalized Values
Passband ripple 0.044 dB, Stop band depth 50 dB

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 11.15
Cauer Normalized Values
Passband ripple 0.044 dB, Stop band depth 60 dB

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 11.16
Cauer Normalized Values
Passband ripple 0.2 dB, Stop band depth 30 dB

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 11.17
Cauer Normalized Values
Passband ripple 0.2 dB, Stop band depth 40 dB

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 11.18
Cauer Normalized Values
Passband ripple 0.2 dB, Stop band depth 50 dB

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 11.19
Cauer Normalized Values
Passband ripple 0.2 dB, Stop band depth 60 dB

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

tables for all possible filter requirements would be impractical, so *tables of normalized values* are used based on a low-pass filter with a bandwidth of 1 radian per second and with 1-Ω input termination impedance. Performance requirements for the desired filter are *normalized* to 1 rad/sec bandwidth and 1-Ω terminations and the tables used to determine the various component values for those frequencies and impedance. That filter is then transformed to the desired response if necessary — high-pass, band-pass, etc. Then the filter is converted to the desired frequency and impedance by *denormalizing* or *scaling* the component values with another set of transformations. In this way, a (relatively) small set of tables can be used to design filters of any type as you'll see in the following examples.

Summarizing, the table-based design process consists of:

- Determining performance requirements.
- Normalizing the requirements to a 1-Hz, 1-Ω, low-pass response.
- Selecting a family and order.
- Obtaining normalized component values from the tables.
- Transforming the filter to the desired response.
- Denormalizing the component values for frequency and impedance.

11.4.2 Filter Family Selection

The tables presented here are for low-pass filters from different families (Butterworth, Bessel, Chebyshev and Cauer), with various specifications for passband ripple and so on. In the normalized tables, capacitor values are in farads (F) and inductor values are in henries (H). These are converted to actual component values through the process of denormalization described below.

The Butterworth family (described by normalized component values in **Table 11.3**) is used when there should be no ripple at all in the passband, so that the response is to be as flat as possible. This trait is particularly apparent near dc for the low-pass response, at the center frequency for a band-pass response, and at infinity for the high-pass response. The resulting magnitude response will also have a relatively gentle transition from the passband into the stop band.

When a low-pass filter with constant group delay throughout the passband is desired then the Bessel family, whose normalized component values are shown in **Table 11.4**, 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 11.5**, for critical RF work), 0.044 dB (**Table 11.6**, an intermediate value offering 20 dB of return loss or 1.2:1 VSWR) and 0.2 dB (**Table 11.7**, for audio and less-critical work where a steeper rolloff 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 rolloff 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 **Tables 11.8** to **11.19**. 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.

DESIGN EXAMPLE — CHEBYSHEV LOW-PASS

Here is an example of using normalized-value tables to design a low-pass Chebyshev filter with 0.01 dB of passband ripple, a bandwidth of 4.2 MHz, a rolloff requirement of 25 dB of attenuation one octave above cutoff, and an input termination of 50 Ω. **Fig 11.35** illustrates the attenuation to be expected for various orders (N) of a Chebyshev low-pass filter with 0.01 dB of passband ripple. The next step is to determine the lowest order that can meet the requirements for roll off. Based on the rolloff requirement, a fifth-order filter is the lowest order filter for which the attenuation curve is below 25 dB at twice the normalized cutoff frequency of 1. (If 14 dB of attenuation one octave above the cutoff frequency had been required, a fourth-order filter would suffice.)

Fig 11.36 shows the schematic of a fifth-order low-pass filter that will meet the requirements. (If a higher- or lower-order filter is required, add or subtract elements, beginning with G(1) at the filter input, with capacitors as the parallel or shunt elements and inductors as the series elements.)

Next, refer to Table 11.5 to obtain the normalized component values for the 1-Ω and 1 radian/second low-pass filter. Choosing the table row with component values for Order = 5, 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 equations to calculate the actual component values are shown in Fig 11.36A. Ω_c is the denormalizing factor used to scale the normalized filter component values to the desired frequency

$$\Omega_c = 2\pi\ F_{width} \qquad (1)$$

In this case, $\Omega_c = 2 \times 3.1416 \times 4.2 \times 10^6 = 26.389 \times 10^6$. 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}.) After denormalizing, the resulting component values are shown in Fig 11.36B. At this point in a table-based design process, the closest available standard value or adjustable components would be used to fabricate the actual circuit.

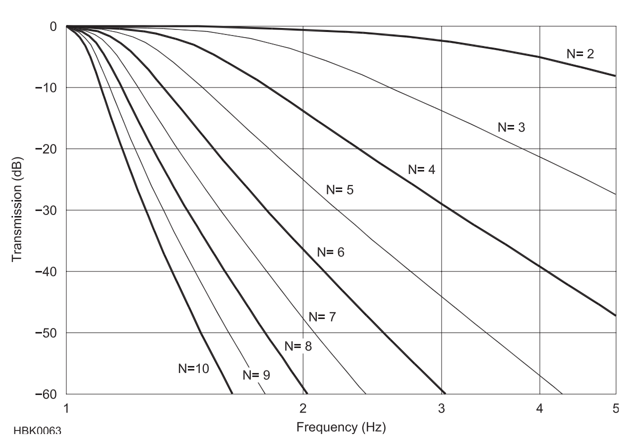


Fig 11.35 — This graph shows transmission for Chebyshev low-pass filters with 0.01 dB passband ripple filters with orders from 2 through 10. For a given order, higher passband ripple values will have higher attenuation (less transmission) at a given frequency while Butterworth filters will have less attenuation at a given frequency.

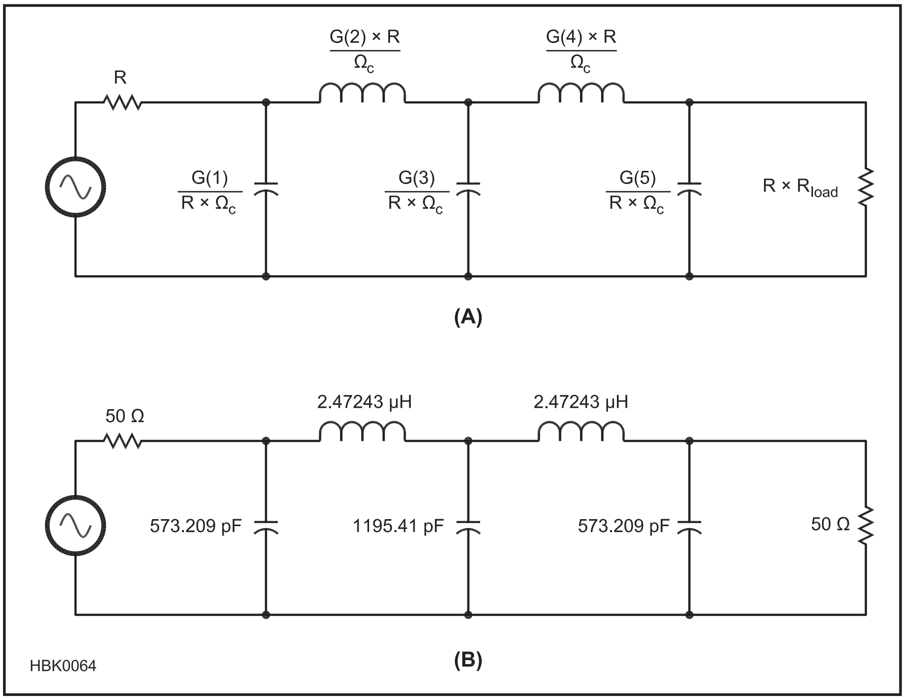


Fig 11.36 — Design example for a Chebyshev low-pass filter using the normalized filter tables. The topology and design equations are shown at A, with the resulting calculated parts values at B.

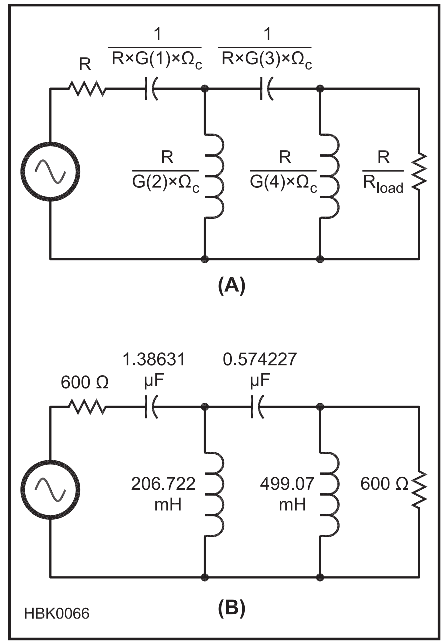


Fig 11.38 — Design example for a Butterworth high-pass filter using the normalized filter tables. The topology and design equations are shown at A, with the resulting calculated parts values at B. For a Cauer family high-pass, trap capacitors would be inserted in series with the shunt inductors. Their value is calculated by the same expression as the series capacitors, but using the *H(n)* values from the Cauer tables.

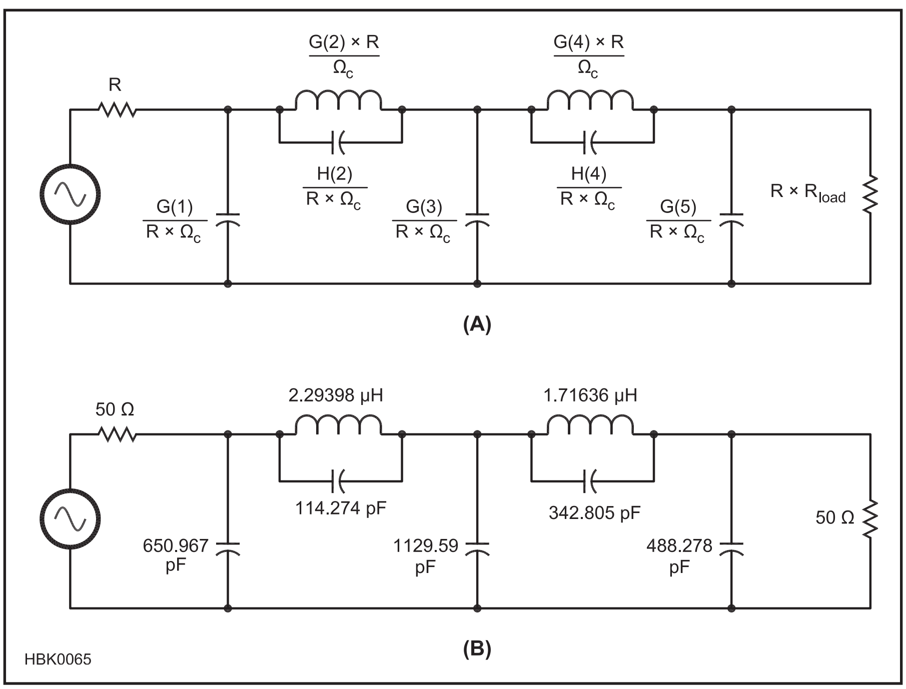


Fig 11.37 — Design example for a Cauer low-pass filter using the normalized filter tables. The topology and design equations are shown at A, with the resulting calculated parts values at B.

DESIGN EXAMPLE — CAUER LOW-PASS

The design of a Cauer low-pass filter is very similar to the design of a Chebyshev, but it has added capacitors across the series inductors to form the traps that create the stop band notches. 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 11.37 shows the filter schematic. The first step is to obtain from Table 11.13 the normalized 1-Ω and 1 radian/second component values. Choosing the table row of component values for Order = 5, 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 the same as the input termination.

The filter schematic with the denormalization equations appears in Fig 11.37A. Ω_c and R have the same definitions as in the previous example. The resulting actual component values for the required frequency and impedance are shown in Fig 11.37B.

DESIGN EXAMPLE — HIGH-PASS BUTTERWORTH

The design of a high-pass filter will now be illustrated. Fig 11.35 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. For example, if the attenuation requirement for the high-pass filter is for 10 dB of attenuation at 0.2 times the cutoff frequen-cy, then when using the low-pass filter response graphs, look up the attenuation at 1/0.2 = 5 times the cutoff frequency, instead.

For this example, we will design a fourth-order capacitor-input Butterworth high-pass response with the 3-dB cutoff frequency at 250 Hz and system impedance of 600 Ω. The filter schematic is shown in **Fig 11.38**. (Remember that even-order ladder filters can have either a series or parallel element at the input.)

The normalized values for the low-pass response are taken from Table 11.3, and the design equations are shown in Fig 11.38A. From the table row of component values 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.

Ω_c is computed the same way as in the previous examples. 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 11.38B shows the resulting denormalized component values for the actual filter.

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 20% of the center frequency or higher (a filter Q of 5 or less).

Fig 11.39 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. Again, Fig 11.35 can be used to determine the required filter order, bearing in mind that it is precise for passband ripple of 0.01 dB.

The normalized values for the low-pass are taken from Table 11.7. Choosing the table row of component values for Order = 3, 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. Ω_c is defined as in the previous examples, while

$$\Omega_o = 2\pi \ F_{center} \quad (2)$$

The filter schematic with the denormalizing equations appears in Fig 11.39A, and calculated values for the actual filter are shown in Fig 11.39B.

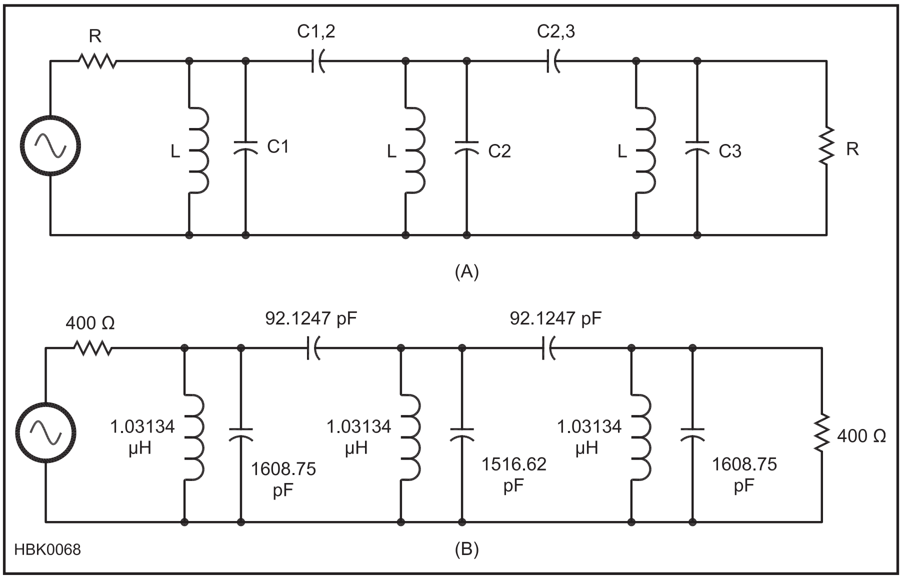


Fig 11.40 — Design example for a Chebyshev narrow band-pass filter using the normalized filter tables. The topology is shown at A and the design equations are given in the text, with the resulting calculated parts values at B.

DESIGN EXAMPLE — NARROW BAND-PASS

Narrow band-pass filters involve more calculations than wide band-pass types. The filter topology used in this discussion is shown in **Fig 11.40** — a set of shunt-connected parallel-tuned resonators coupled by relatively small-value coupling capacitors.

This example will be for a band-pass filter with a bandwidth of 200 kHz, 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 11.6.

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 and Ω_o are used as before The same inductor, L, is used for each resonator. The value of that inductor is given by:

$$L = \frac{R}{\Omega_o \times Q_L \times G(1)} \quad (3)$$

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

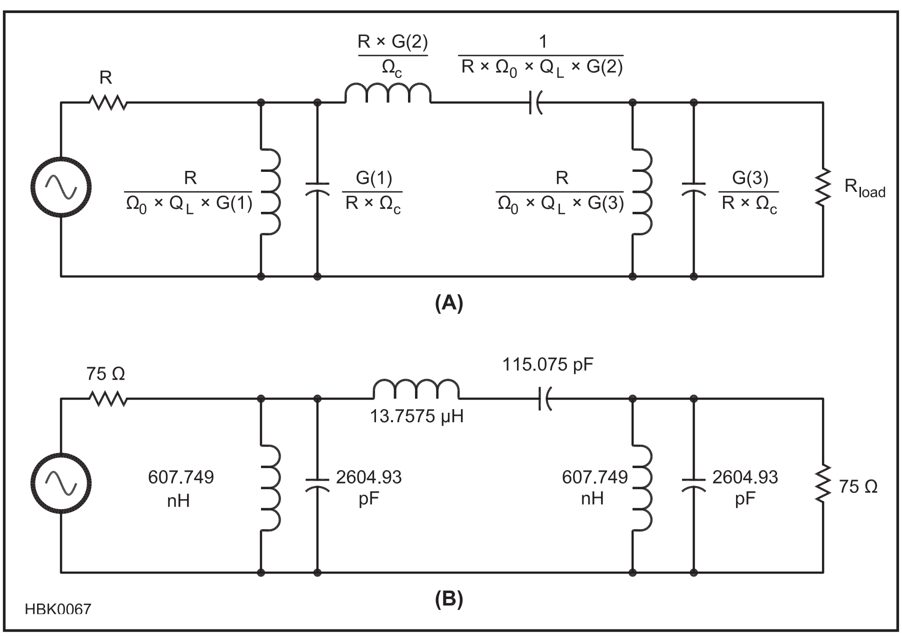


Fig 11.39 — Design example for a Chebyshev wide band-pass filter using the normalized filter tables. The topology and design equations are shown at A, with the resulting calculated parts values at B.

$$C_{basic} = \frac{G(1)}{R \times \Omega_c} \quad (4)$$

Next calculate the inter-resonator coupling capacitors:

$$C1,2 = \frac{G(1)}{R \times \Omega_o} \sqrt{\frac{1}{G(1) \times G(2)}} \quad (5)$$

$$C2,3 = \frac{G(1)}{R \times \Omega_o} \sqrt{\frac{1}{G(2) \times G(3)}} \quad (6)$$

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 11.40B. It should be evident how to extend this procedure to higher-order filters.

11.5 Active Audio Filters

Below RF, in what is broadly referred to as the "audio" range between a few Hz and a few hundred kHz, designers have several choices of filter technology.

- Passive LC
- Digital Signal Processing (DSP)
- Switched-Capacitor Audio Filter (SCAF)
- Active RC

LC audio filters are not used much in current designs except in high-power audio applications, such as speaker crossover networks, and are not covered here. LC filters were once popular as external audio filters for CW reception. These designs tended to be large and bulky, often using large surplus 44 or 88 mH core inductors. (A classic *ARRL Handbook* project to construct a passive LC CW filter can be found on the CD-ROM included with this book.) LC filters used in very low-level receiver applications can be very compact, but these tend to suffer from relatively high insertion loss, which reduces the receiver sensitivity (if no preamp is used ahead of the filter) or its large-signal dynamic range (if a loss-compensating preamp is used ahead of the filter). In addition, LC filters used in low-level receiver applications tend to pick up 60/120 Hz hum from stray magnetic fields from ac power transformers. Even "self shielded" inductors can produce noticeable ac hum pick up when followed by 90 to 120 dB of audio gain!

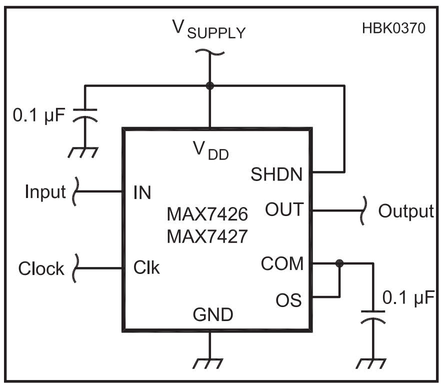


Fig 11.41 — Simple SCAF low pass filter, taken from the MAX7426 data sheet.

DSP filtering can yield filters that are superior to analog filters. (See the **DSP and Software Radio Design** chapter.) High signal-level DSP-based external audio filters such as the popular Timewave DSP-9 (**www.timewave.com**) and MFJ Enterprises MFJ-784B (**www.mfjenterprises.com**) have been available for some time. In addition, if the DSP is coupled to a very high-performance audio A/D converter such as an Asahi-Kasei AK5394A, DSP filtering can provide excellent filtering for even very low-level audio signals.

The main drawback to DSP processing is that of expense. A low-end DSP external audio filter costs at least $100, while a DSP with a high-end A/D converter (such as an audio sound card front-end to a PC) can run in the $150 to $400+ price range, which does not include the cost of the host PC.

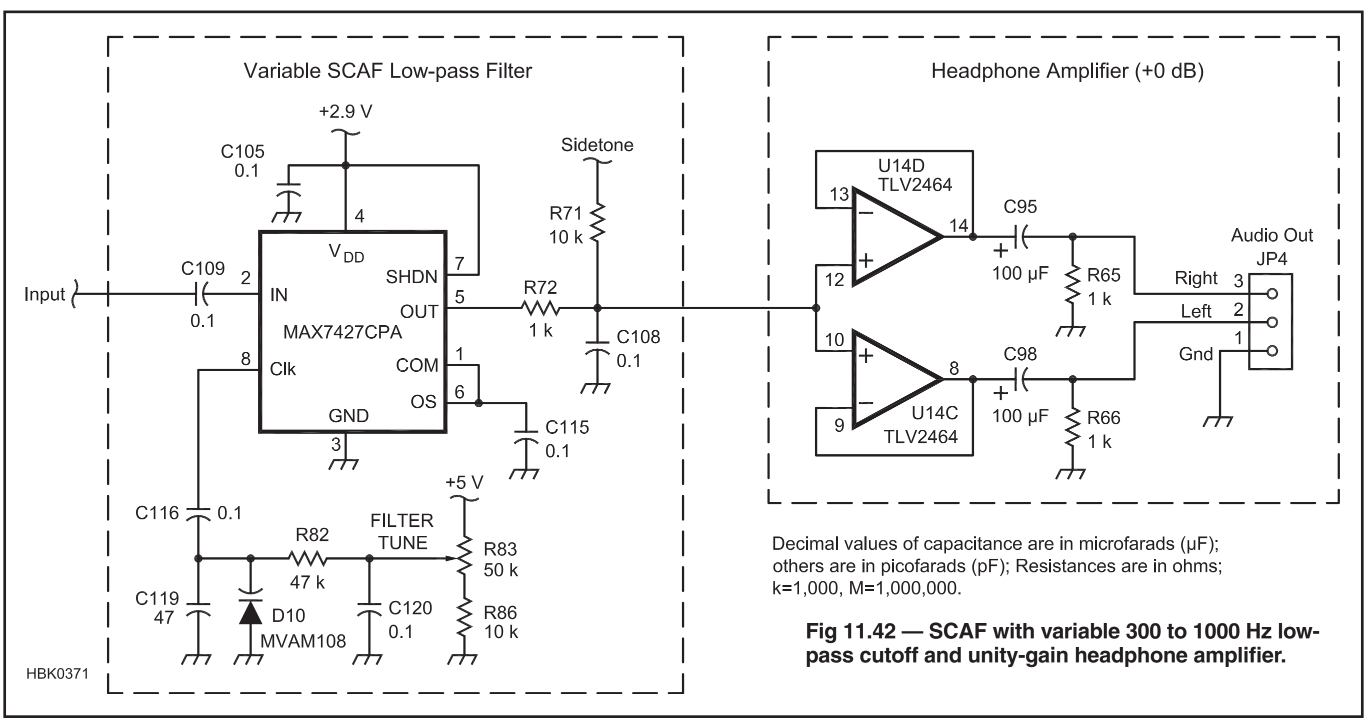


Fig 11.42 — SCAF with variable 300 to 1000 Hz low-pass cutoff and unity-gain headphone amplifier.

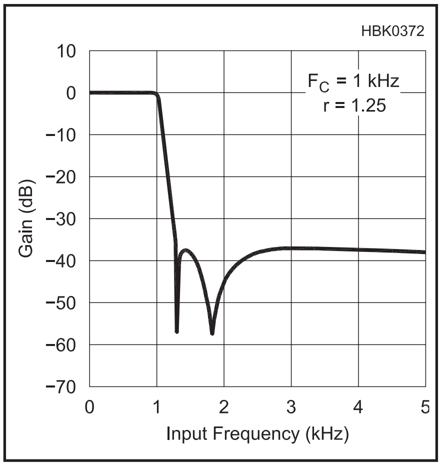


Fig 11.43 — Frequency response of a low-pass filter using the MAX7426 set to a 1 kHz cutoff frequency.

11.5.1 SCAF Filters

Simple SCAF filter designs can produce extremely effective low-power audio filters. The implementation of these filters in practical IC form usually involves small-value capacitors and high-value resistors. The use of high-value resistors tends to generate enough noise to make these filters unsuitable for very low-level audio processing such as the front end of a receiver audio chain. Thus, SCAF filters tend to be limited to filtering at the output end of the audio chain — headphone or speaker level audio applications, an area in which they very much excel. The cutoff frequency of a SCAF filter is set by an external clock signal. Thus, this class of filter naturally lends itself to use in variable frequency filters. For simplicity, excellent frequency selectivity and relatively low cost, SCAF filters are highly recommended when additional audio filtering is desired on the output of an existing receiver.

An example of a very simple, but highly effective SCAF low-pass filter IC is the Maxim MAX7426 (5 V supply) or MAX7427 (3 V supply) in **Fig 11.41.** The filter's cutoff frequency can be set by placing an appropriately sized capacitor across the clock oscillator inputs because the part can generate its own internal clock signal. For example, connecting a 180 pF capacitor across the MAX7426's clock inputs will produce a 1 kHz low-pass filter. The MAX7427 (3-V version) was used in the NC2030 QRP transceiver along with a MVAM108 varactor diode to create a low-pass filter that was tunable from 300 Hz to 1 kHz.

A portion of that schematic is shown in **Fig 11.42**. The SCAF low-pass filter is followed by a 3-V unity-gain headphone amplifier (U14C and U14D) as the filter IC itself cannot directly drive headphones. The low-pass cutoff frequency is tuned by using R83 to vary the voltage applied across the MVAM108 varactor diode D10, thus changing the capacitance across the clock input (pin 8) of the chip. C116 was used to isolate the dc voltage across the varactor diode from the bias voltage on the clock input line. The frequency response of such a filter is shown in **Fig 11.43**.

As can be seen, this is an extremely sharp filter, producing almost 40 dB of attenuation very close to the cut off frequency — not too bad for an inexpensive part. The slightly more expensive MAX7403 gives an even steeper 80 dB cutoff. Both parts are specified for an 80 dB signal-to-noise ratio based on an assumed 4-V_{P-P} signal. Sensitive modern in-ear type headphones require only 20 mV_{P-P} to produce a fairly loud signal. A 20 mV signal is 46 dB below 4 V_{P-P}, so at such headphone levels, the noise floor is actually only 34 dB below the 4 V_{P-P} signal — fairly quiet, but is a lot less noise margin than one would tend to think given the 80 dB specification.

Again, with a noise floor only 34 dB below typical headphone levels, SCAF filters such as these are fine at the end of the receiver chain, but are not useful as an audio filter at very low signal levels early on in a receiver.

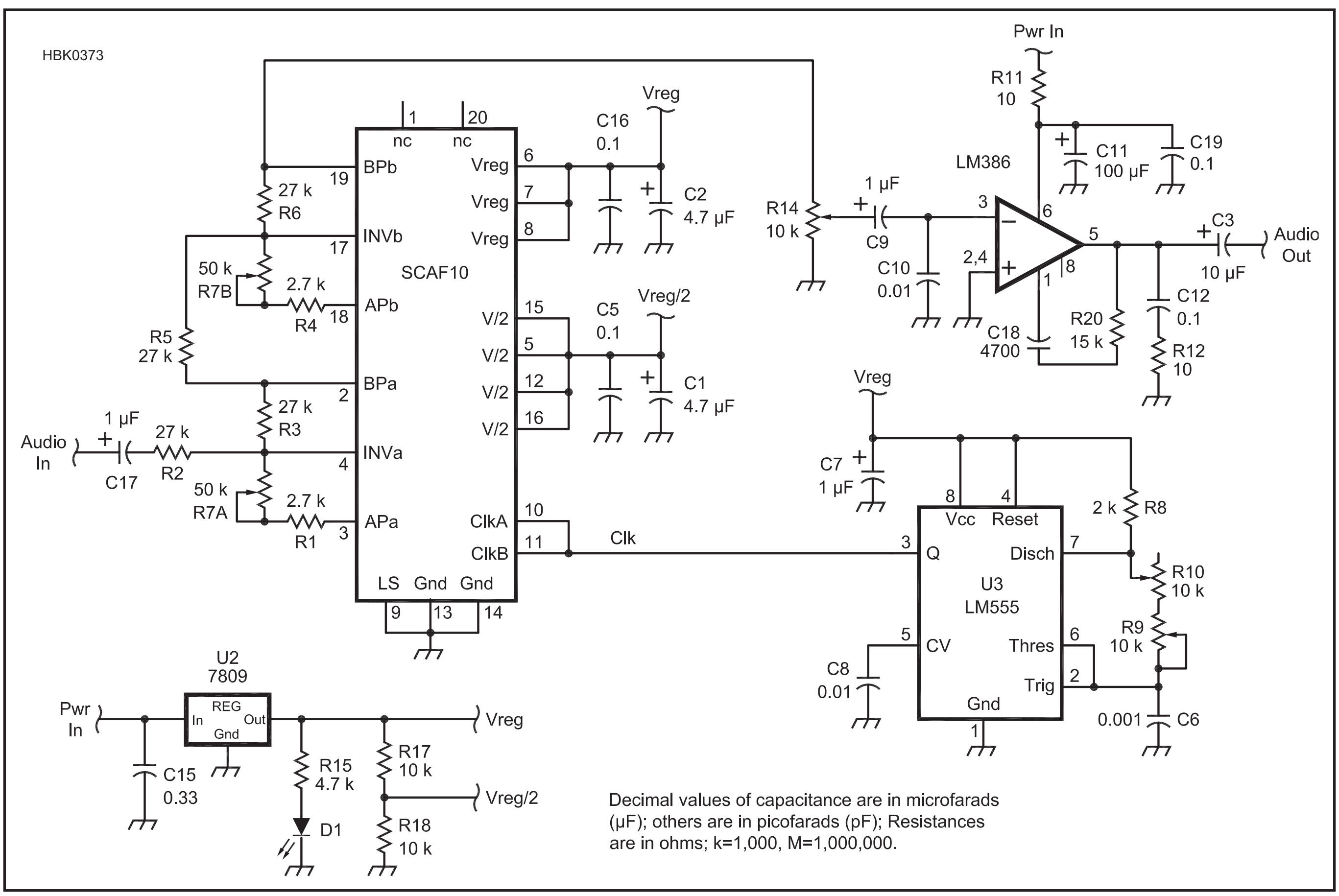


Fig 11.44 — Example of a simple SCAF band-pass filter from a New England QRP Club kit.

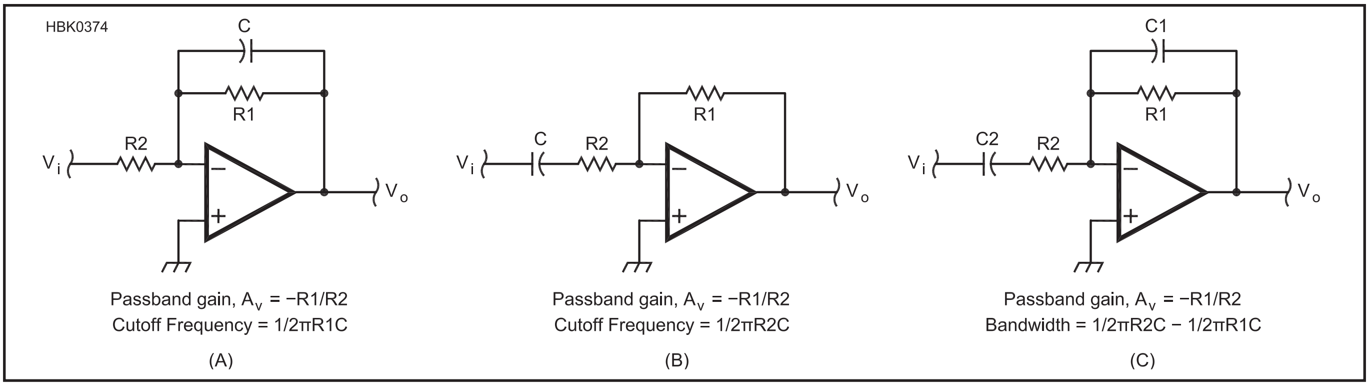
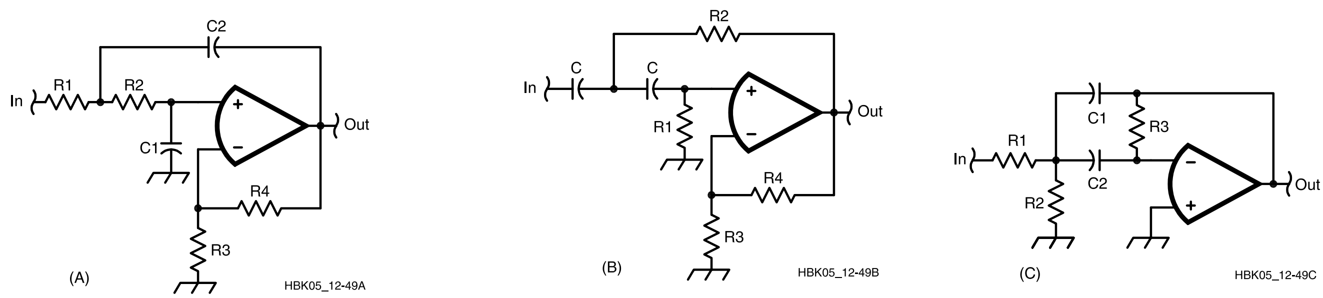


Fig 11.45 — Simple active filters. A low-pass filter is shown at A; B is a high-pass filter. The circuit at C combines the low- and high-pass filters into a wide band-pass filter.

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 rad/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 rad/sec
- 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

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

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

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

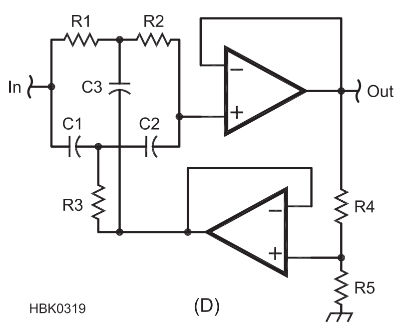
Example:
K = 2, f_o = 800 Hz, Q = 5 and C = 0.022 μF
R1 = 22.6 kΩ (use 22 kΩ)
R2 = 942 Ω (use 910 Ω)
R3 = 90.4 kΩ (use 91 kΩ)

Fig 11.46 — 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 11.20 for values of "a".

Fig 11.44 is an example of a slightly more complex SCAF band-pass filter. This filter features both a variable center frequency (450 Hz to 1000 Hz) and a variable bandwidth (90 Hz to 1500 Hz). This very popular filter was designed by the New England QRP Club (**www.newenglandqrp.org**) and is currently offered in kit form.

This filter features both sections of the SCAF10 IC configured as identical band-pass filters. The bandwidth (Q) is adjustable via R7A and R7B, while the center frequency is adjustable by R10 and the trimmer R9 as these resistors set the frequency of the LM555 clock generator.

As in the previous SCAF filter example, this filter also includes an audio amplifier (LM386) for driving either an external speaker or headphones as none of the SCAF filter ICs are capable of driving headphones directly.



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 >> 1 kΩ
R4 = 25 Ω (use 24 Ω)
R5 = 975 Ω (use 910 Ω)

11.5.2 Active RC Filters

Active RC filters based on op amp circuits can be used in either high-level audio output or very low-level direct-conversion receiver front-end filtering applications and thus are extremely flexible. Unlike LC filters, active RC filters can provide both filtering and gain at the same time, eliminating the relatively high insertion loss of a physically small, sharp LC filter. In addition, an active RC filter is not susceptible to the same ac hum pickup in low signal level applications as LC filters.

Active filters can be designed for gain and they offer excellent stage-to-stage isolation. The circuits require only resistors and capacitors, avoiding the limitations associated with inductors. By using gain and feedback, filter Q is controllable to a degree unavailable to passive LC filters. 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. Active filters that drive speakers or other heavy loads usually employ an audio output amplifier circuit to boost the output power after a low-power filter stage.

A particular advantage of active RC filters for use in receivers is that they can be capable of extremely low noise operation, allowing the filtering of extremely small signals. At the same time they are capable of handling extremely large signals. Op amps are often capable of using ±18-V dual supply voltages or 36-V single supply voltages allowing the construction of a very high performance audio filter/amplifier chain that can handle signals up to 33 V_{P-P}. The ability to provide gain while also providing filtering provides a lot of flexibility in managing the sensitivity of a receiver audio chain.

The main disadvantage of active RC filters is their relatively high parts count compared to other filter types. In addition, active RC filters tend to be fixed-frequency designs unlike SCAF filters whose frequency can be moved simply by changing the clock frequency that drives the SCAF IC.

11.5.3 Active Filter Responses

Active filters can implement any of the passive LC filter responses described in the preceding section: low-pass, high-pass, band-pass, band-stop and all-pass. Filter family responses such as Butterworth, Chebyshev, Bessel and Cauer (elliptic) can be realized. All of the same family characteristics apply equally to passive and active filters and will not be repeated here. (Op amps are discussed in the **Analog Basics** chapter.)

Fig 11.45 presents circuits for first-order low-pass (Fig 11.45A) and high-pass (Fig 11.45B) active filters. The frequency response of these two circuits is the same as a parallel and series RC circuit, respectively, except that these two circuits can have a passband gain greater than unity. Rolloff is shallow at 6 dB/octave. The two responses can be combined to form a simple band-pass filter (Fig 11.45C). This combination cannot produce sharp band-pass filters because of the shallow rolloff.

To achieve high-order responses with steeper rolloff and narrower bandwidths, more complex circuits are required in which combinations of capacitors and resistors create *poles* and *zeroes* in the frequency response. (Poles and zeros are described in the **Analog Basics** chapter.) The various filter response families are created by different combinations of additional poles and zeroes. There are a variety of circuits that can be configured to implement the equations that describe the various families of filter responses. The most common circuits are *Sallen-Key* and *multiple-feedback*, but there are numerous other choices.

There are many types of active filters — this section presents some commonly used circuits as examples. A book on filter design (see the References section) will present more choices and how to develop designs based on the different circuit and filter family types. In addition, op amp manufacturer's publish numerous application notes and tutorials on active filter design. Several are listed in the References section of this chapter.

The set of tutorials published by Analog Devices is particularly good.

SECOND-ORDER ACTIVE FILTERS

Fig 11.46 shows circuits for four second-order filters: low-pass (Fig 11.46A), high-pass (Fig 11.46B), band-pass (Fig 11.46C) and band-reject or notch (Fig 11.46D). Sequences of these filters are used to create higher-order circuits by connecting them in series. Two second-order filter stages create a fourth-order filter, and so forth.

The low-pass and high-pass filters use the Sallen-Key circuit. Note that the high-pass circuit is just the low-pass circuit with the positions of R1-R2 and C1-C2 exchanged. R3 and R4 are used to control gain in the low- and high-pass configurations. The band-pass filter is a multiple-feedback design. The notch filter is based on the twin-T circuit. All of the filter design equations and tables will result in a Butterworth family response. (For the Chebyshev and other filter responses, consult the references listed at the end of the chapter.)

The circuits in Fig 11.46 assume the op amp is operating from a balanced, bipolar power supply, such as ±12 V. If a single supply is used (such as +12 V and ground), the circuit

Table 11.20
Factor "a" for Low- and High-Pass Filters in Fig 11.46

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 Williams, *Electronic Filter Design Handbook*, for even-order Butterworth filters

Active Quadrature Network Design Software

QuadNet, a *Windows* program for design and analysis of active quadrature ("90-degree") networks for use in SSB transmitters and receivers, is available from **www.arrl.org/arrl-handbook-reference**. Written by Jim Tonne, W4ENE, *QuadNet* handles orders 2 to 10, odd and even, with tuning modes and analysis.

must have a dc offset added and blocking capacitors between filter sections to prevent the dc offset from causing the op amp to saturate. The references listed at the end of the chapter provide more detailed information on single-supply circuit design.

Avoid electrolytic or tantalum capacitors as frequency-determining components in active filter design. These capacitors are best used for bypassing and power filtering as their tolerance is generally quite low, they have significant parasitic effects, and are usually polarized. Very small values of capacitance (less than 100 pF) can be affected by stray capacitance to other circuit components and wiring. High-order and high-Q filters require close attention to component tolerance and temperature coefficients, as well.

SECOND-ORDER ACTIVE FILTER DESIGN PROCEDURES

The following simple procedures are used to design filters based on the schematics in Fig 11.46. Equations and a design example are provided in the figure.

Low- and High-Pass Filter Design

To design a low- or high-pass filter using the circuits in Fig 11.46, start by determining your performance requirements for filter order (2, 4, 6, or 8), gain (K) and cutoff frequency (f_C). Calculate $\omega_C = 2\pi f_C$ and C_2 or C as required. **Table 11.20** provides design coefficients to create the Butterworth response from successive Sallen-Key low- and high-pass stages. A different coefficient is used for each stage. Obtain design coefficient "a" from Table 11.20. Calculate the remaining component values from the equations provided.

Band-Pass Filter Design

To design a band-pass filter as shown in Fig 11.46C, begin by determining the filter's required Q, gain, and center frequency, f_0. Choose a value for C and solve for the resistor values. Very high or very low values of resistance (above 1 MΩ or lower than 10 Ω) should be avoided. Change the value of C until suitable values are obtained. High gain and Q may be difficult to obtain in the same stage with reasonable component values. Consider a separate stage for additional gain or to narrow the filter bandwidth.

Band-Reject (Notch) Filter Design

Band-reject filter design begins with the selection of center frequency, f_0, and Q. Calculate the value of K. Choose a value for C1 that is approximately 10 µF / f_0. This determines the values of C2 and C3 as shown. Calculate the value for R1 that results in the desired value for f_0. This determines the values of R2 and R3 as shown. Select a convenient value for R4 + R5 that does not load the output amplifier. Calculate R4 and R5 from the value of K.

The depth of the notch depends on how closely the values of the components match the design values. The use of 1%-tolerance resistors is recommended and, if possible, matched values of capacitance. If all identical components are used, two capacitors can be paralleled to create C3 and two resistors in parallel create R3. This helps to minimize thermal drift.

11.5.4 Active Filter Design Tools

While the simple active filter examples presented above can be designed manually, more sophisticated circuits are more easily designed using filter-design software. Follow the same general approach to determining the filter's performance requirements and then the filter family as was presented in the section on LC filters. You can then enter the values or make the necessary selections for the design software. Once a basic design has been calculated, you can then "tweak" the design performance, use standard value components and make other adjustments. The design example presented below shows how a real analog design is assembled by understanding the performance requirements and then using design software to experiment for a "best" configuration.

Op amp manufacturers such as Texas Instruments and National Semiconductor (originally separate companies but now merged) have made available sophisticated "freeware" filter design software. These packages are extremely useful in designing active RC filters. They begin by collecting specifications from the user and then creating a basic circuit. Once the initial circuit has been designed, the user can adjust specifications, component values, and op amp types until satisfied with the final design.

Free filter design software can be obtained from the following sources:

• Analog Devices — *Analog Filter Wizard 2.0* (**www.analog.com/designtools/en/filterwizard**)

• National Semiconductor — *Webench* (**http://webench.national.com**)

• Texas Instruments — *FilterPro* (**www.ti.com**, search for "FilterPro")

Even if a manual design process is followed, using a software tool to double-check the results is a good way to verify the design before building the circuit.

An extended design example using the Texas Instruments *FilterPro* software is included on this book's CD-ROM. In the example, Dan Tayloe, N7VE explains the process of designing a high-performance 750 Hz low-pass filter, illustrating the power of using sophisticated interactive tools that enable design changes on-the-fly. The reader is encouraged to follow along and experi-ment with *FilterPro* as a means of becoming familiar with the software so that it can be used for other filter design tasks. Similar processes apply to other filter design software tools.

11.6 Quartz Crystal Filters

Inductor Q values effectively limit the minimum bandwidth that can be achieved with LC band-pass filters. Higher-Q circuit elements, such as quartz crystal, PLZT ceramic and constant-modulus metal alloy resonators, are required to extend these limits. Quartz crystals offer the highest Q and best stability with time and temperature of all available resonators. They are manufactured for a wide range of frequencies from audio to VHF, using cuts (crystal orientations) that suit the frequency and application of the resonator. The AT cut is favored for HF fundamental and VHF overtone use, whereas other cuts (DT, SL and E) are more convenient for use at lower frequencies.

Each crystal plate has several modes of mechanical vibration. These can be excited electrically thru the piezoelectric effect, but generally resonators are designed so as to maximize their response on a particular operating frequency using a crystal cut that provides low loss and a favorable temperature coefficient. Consequently, for filter design, quartz crystal resonators are modeled using the simplified equivalent circuit shown in **Fig 11.47**. Here L_m, C_m and r_m represent the *motional* parameters of the resonator at the main operating frequency — r_m being the loss resistance, which is also known as the *equivalent series resistance*, or ESR. C_o is a combination of the static capacitance formed between the two metal electrodes with the quartz as dielectric (ε_r = 4.54 for AT-cut crystals) and some additional capacitance introduced by the metal case, base and mount. There is a physical relationship between C_m and the static capacitance formed by the resonator electrodes, but, unfortunately, the added holder capacitance masks this direct relationship causing C_o/C_m to vary from 200 to over 500. However, for modern fundamental AT-cut crystals between 1 and 30 MHz, their values of C_m are typically between 0.003 and 0.03 pF. Theoretically, the motional inductance of a quartz crystal should the same whether it is operated on the fundamental or one of its overtones, making the motional capacitance at the nth overtone $1/n^2$ of the value at the fundamental. However, this relationship is modified by the effect of the metal electrodes deposited on either side of the crystal plate and in practice the motional inductance increases with the frequency of the overtone. This makes the motional capacitance at the overtone substantially less than C_m/n^2.

An important parameter for crystal filter design is the unloaded Q at f_s, the resonant frequency of the series arm. This is usually denoted by Q_U.

$$Q_u = 2 \pi f_s L_m/r_m \quad (7)$$

Q_U is very high, often exceeding 100,000 in the lower HF region. Even VHF overtone crystals can have Q_U values over 20,000, making it possible to design quartz crystal band-pass filters with a tremendously wide range of bandwidths and center frequencies.

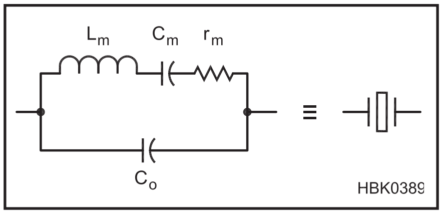


Fig 11.47 — Equivalent circuit of a quartz crystal and its circuit symbol.

The basic filtering action of a crystal can be seen from **Fig 11.48**, which shows a plot of attenuation vs frequency for the test circuit shown in the inset**.** The series arm of the crystal equivalent circuit forms a series-tuned circuit, which passes signals with little attenuation at its resonant frequency, f_s, but appears inductive above this frequency and parallel resonates with C_o at f_p to produce a deep notch in transmission. The difference between f_s and f_p is known as the pole-zero separation, or PZ spacing, and is dependent on C_m/C_o as well as f_s. Further information on quartz crystal theory and operation can be found in the **Oscillators and Synthesizers** chapter and Ref 1.

A simple crystal filter developed in the 1930s is shown in **Fig 11.49**. The voltage-reversing transformer T1 was usually an IF transformer, but nowadays could be a bifilar winding on a ferrite core. Voltages V_a and V_b have equal magnitude but 180° phase difference. When $C_1 = C_o$, the effect of C_o will disappear and a well-behaved single resonance will occur as indicated by the solid line in Fig 11.49B. However, if C_1 is adjusted to unbalance the circuit, a transmission zero (notch) is produced well away from the pass band and by increasing the amount of imbalance this can be brought back towards the edge of the pass band to attenuate close-by interfering CW signals. If C_1 is reduced in value from the balanced setting, the notch comes in from the high side and if C_1 is increased, it comes in from the low side. The dotted curve in Fig 11.49B illustrates how the notch can be set with C_1 less than C_o to suppress adjacent signals just above the pass band. In practice a notch depth of up to 60 dB can be achieved. This form of "crystal gate" filter, operating at 455 kHz, was present in many high-quality amateur communications receivers from the 1930s through the 1960s. When the filter was switched into the receiver IF amplifier the bandwidth was reduced to a few hundred Hz for CW reception. The close-in

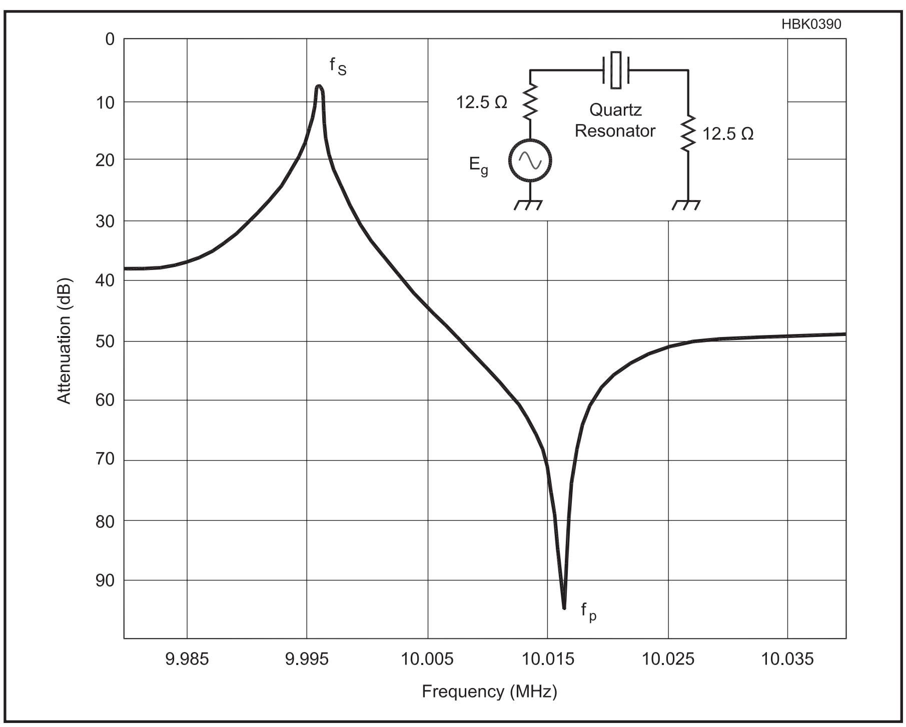


Fig 11.48 — Response of 10 MHz crystal (Cm = 0.0134 pF, Lm = 18.92 mH, ESR = 34 Ω) in a series test circuit (see inset) showing peak of transmission (lowest attenuation) at the series resonant frequency, f_s, and a null (maximum attenuation) caused by the parallel resonance due to C_o at f_p.

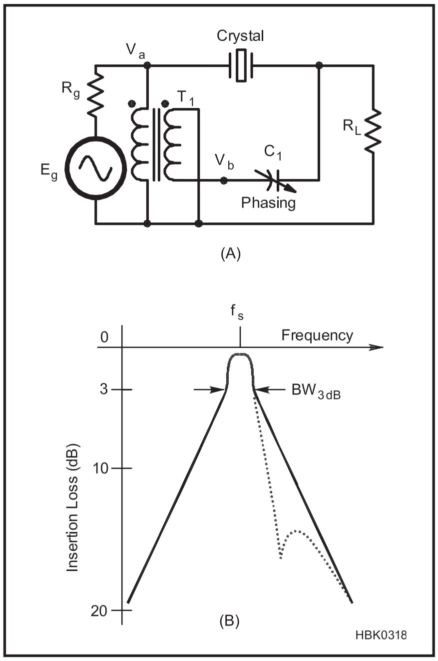


Fig 11.49 — Classic single-crystal filter in A has the response shown in B. The phasing capacitor can be adjusted to balance out C_o (solid line), or set to a lesser or greater value to create a movable null to one side, or other, of the passband (dotted line).

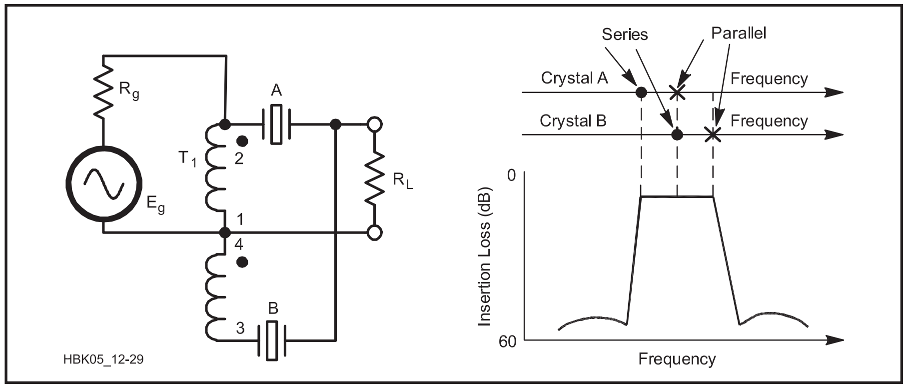


Fig 11.50 — A half-lattice crystal filter. The C_o of one crystal can be made to balance the C_o of the other, or C_o across the higher crystal can be deliberately increased to create nulls on either side of the passband.

range of the notch was sometimes improved by making C_1 part of a differential capacitor that could add extra capacitance to either the C_1 or C_o side of the IF transformer. This design could also be used to good effect at frequencies up to 1.7 MHz with an increased minimum bandwidth. However, any crystal gate requires considerable additional IF filtering to achieve a reasonable ultimate attenuation figure, so it should not be the only form of selectivity used in an IF amplifier.

The half-lattice filter shown in **Fig 11.50** offers an improvement in performance over a single-crystal filter. The quartz crystal static capacitors, C_o, cancel each other. The remaining series-resonant arms, if offset in frequency and terminated properly, will produce an approximate 2-pole Butterworth or Chebyshev response. The crystal spacing for simple Chebyshev designs is usually around two-thirds of the bandwidth. Half-lattice filter sections can be cascaded to produce composite filters with multiple poles. Many of the older commercial filters are coupled half-lattice types using 4, 6 or 8 crystals, and this is still the favored technology for some crystal filters at lower frequencies. Very often extra capacitance was added across one crystal in each half lattice to unbalance C_o and provide deep transmission zeroes on either side of the pass band to sharpen up the close-in response at the expense of the attenuation further out. Ref 11 discusses the computer design of half-lattice filters.

Most commercial HF and VHF crystal filters produced today use dual monolithic structures as their resonant elements. These are a single quartz plate onto which two sets of metal electrodes have been deposited, physically separated to control the acoustic (mechanical) coupling between them. An example of a dual monolithic filter (2-pole) is shown in **Fig 11.51.** These are available with center frequencies from 9 MHz to well over 120 MHz. In effect, the dual monolithic structure behaves just like a pair of coupled crystal resonators. There is a subtle difference, however, because the static capacitance in duals appears across the input and output terminations and doesn't produce a null above the pass band, as it would for two electrically coupled crystal resonators. Multi-pole filters can be built by coupling duals together using external capacitors, the C_o from each dual input and output being absorbed in the coupling capacitors or terminations, or by including more acoustically coupled resonators in the electrode structure on a single quartz plate. Though not common, 3-pole and 4-pole monolithic filters housed in standard, single crystal holders are available from some manufacturers.

Fig 11.51 — A dual monolithic crystal filter has two sets of electrodes acoustically coupled to provide a two-pole response in a single crystal unit. The center lead is connected to the case and grounded in normal operation.

11.6.1 Filter Parameters

An ideal band-pass filter would pass the desired signal with no loss and completely attenuate everything else. Practically, it is not possible to achieve such a response with a finite number of resonators, and approximations to this ideal have to be accepted. Greater stop-band attenuation and steeper sides can be achieved if more and more crystals are used, and the response gets nearer to the ideal "brick-wall" one. This feature of a filter is expressed as a *shape factor*, which specifies the ratio of the bandwidth at an attenuation of 60 dB to the bandwidth at 6 dB — both these levels being taken relative to the actual pass band peak to eliminate insertion loss from the calculation. An ideal brickwall filter would have a shape factor of 1, and practical filters have shape factors that depend on the number of crystals used in the design and the type of response chosen for the pass band. A 1 dB Chebyshev design, for example, can typically produce shape factors that vary from about 4.1 for a 4-pole to 1.5 for a 10-pole, but the actual figures obtained in practice are very much dependent on Q_u and how much greater it is than the filter Q, defined by f_o/BW, where f_o is the center frequency and BW the 3 dB bandwidth. The ratio of Q_u to f_o/BW is often quoted as Q_o, or q_o, and this, along with the order and type of response, determines the insertion loss of a crystal filter. Q_o also determines how closely the pass band follows the design response, and how much the passband ripple is smoothed out and the edges of the response rounded off by crystal loss.

Commercial filter manufacturers usually choose the Chebyshev equiripple design for SSB, AM and FM bandwidths because it gives the best compromise between passband response and the steepness of the sides, and 1 dB-ripple Chebyshev designs are pretty standard for speech bandwidths. Tolerances in component values and crystal frequencies can cause the ripple in the pass band to exceed 1 dB, so often the maximum ripple is specified as 2 dB even though the target ripple is lower. Insertion loss, the signal loss going thru the filter, also varies with the type of response and increases as the order of the filter increases. The insertion loss for a given order and bandwidth is higher for high-ripple Chebyshev designs than it is for low-ripple ones, and Butterworth designs have lower insertion loss than any Chebyshev type for a given bandwidth. Pass-band amplitude response and shape factor are important parameters for assessing the performance of filters used for speech communications.

However, group delay is also important for data and narrow bandwidth CW reception. Differential group delay can cause signal distortion on data signals if the variations are greater than the automatic equalizer can handle. Ringing can be an annoying problem when using very narrow CW filters, and the group delay differential across the pass band must be minimized to reduce this effect. When narrow bandwidth filters are being considered, shape factor has to be sacrificed to reduce differential group delay and its associated ringing problems. It's no good having a narrow Chebyshev design with a shape factor of 2 if the filter produces unacceptable ringing and is intolerable to use in practice.

Both Bessel and linear phase (equiripple 0.05º) responses have practically constant, low group delay across the entire pass band and well beyond on either side, making either a good choice for narrow CW or specialized data use. They also have the great advantage of offering the lowest possible insertion loss of all the types of response currently in use, which is important when Q_o is low, as it often is for very narrow bandwidth filters. The insertion loss of the Bessel design is marginally lower than that of the linear phase, but the latter has a superior shape factor giving it the best balance of low group delay and good selectivity. A 6-pole linear phase (equiripple 0.05°) design has a shape factor of 3.39, whereas a 6-pole Bessel has 3.96. The Gaussian-to-12 dB response has a better shape factor than that of either the linear phase or Bessel designs but the group delay across the pass band is not as flat and pronounced peaks (ears) are beginning to appear at the band edges with a 6-pole design. The Gaussian-to-6 dB group delay is reasonably flat for 3- and 4-pole filters, but significant ears appear towards the passband edges in designs with 5 poles or more.

11.6.2 Crystal Filter Design

A wide variety of crystals are produced for use with microprocessors and other digital integrated circuits. They are offered in several case styles, but the most common are HC-49/U and HC-49/US. Crystal resonators in the larger HC-49/U style case are fabricated on 8 mm diameter quartz discs, whereas those in the squat HC-49/US cases are fabricated on 8 mm by 2 mm strips of quartz. At any given frequency, C_m will be lower for HC-49/US crystals because the active area is smaller than it is in the larger HC-49/U crystals. Both types are cheap and have relatively small frequency spreads, making them ideal for use in the LSB ladder configuration suggested by Dishal (Ref 2) — see **Fig 11.52**. This arrangement requires the motional inductances of all the crystals to be identical, and each loop in isolation (crystal and coupling capacitors either side) to be resonant at the same frequency. Series capacitors to trim individual crystals are needed to achieve this in some of the more advanced designs, where production frequency spreads are not sufficient to satisfy the latter requirement. Refs 3, 4, 5, 6, 7, 8 and 11 contain design information on Dishal LSB crystal ladder filters. Design software associated with Ref 11 (**11x09_Steder-Hardcastle.zip**) can also be downloaded from **www.arrl.org/qexfiles**.

The *min-loss* form of Cohn ladder filter, where $C_{12} = C_{23} = C_{34}$, has become very popular in recent years because it's so simple to design and build. However, it suffers from the drawback that the ripple in its passband response increases dramatically with increasing order, and ringing can be a real problem at bandwidths below 500Hz for Cohn *min-loss* filters of 6th-order, or more. The ripple may not be a problem in most narrow filters because it's smoothed out almost completely by loss, but the ringing can be tiring. For wider bandwidths, where the ratio of Q_u to f_o/BW is much greater, the ripple is not smoothed out, and is very evident. One way round this problem, without sacrificing simplicity, is to use the arrangement shown in **Fig 11.53**, which was originally designed for variable bandwidth applications. It was devised to make the mesh frequencies track together as variable coupling changed the bandwidth. However, it has the great advantage that it can be optimized for almost equal ripple at maximum bandwidth, making it an ideal al-

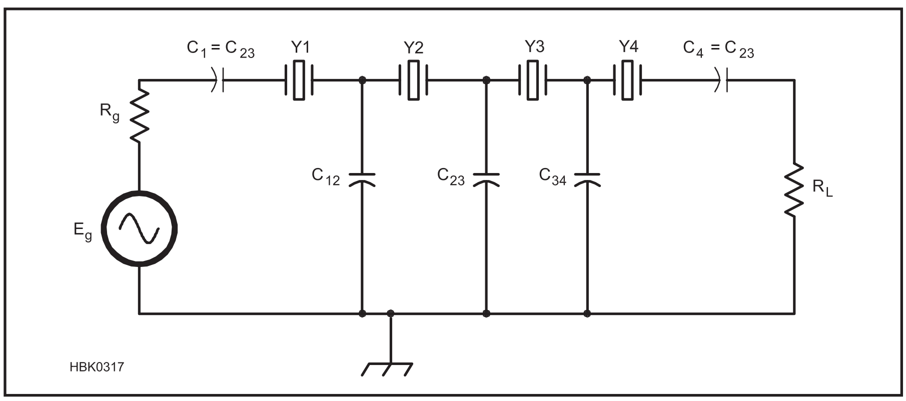


Fig 11.52 — Dishal LSB crystal-ladder filter configuration. Crystals must have identical motional inductances, and the coupling capacitors and termination resistors are selected according to the bandwidth and type of passband response required.

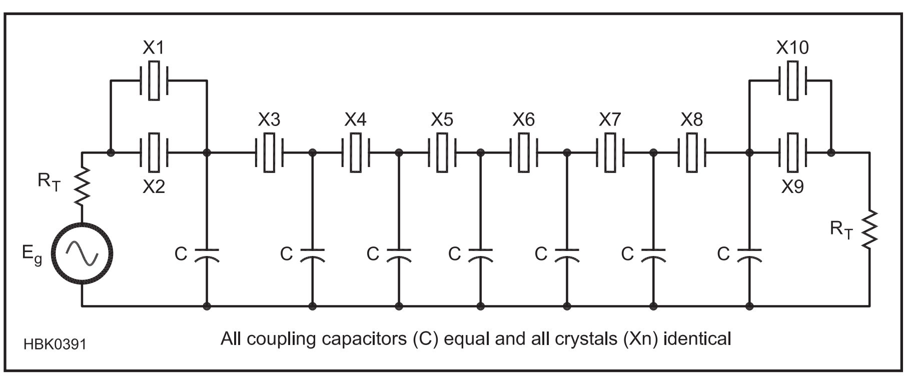


Fig 11.53 — Configuration of improved crystal ladder filter using identical crystals and equal coupling capacitor values. The parallel resonator end-sections (PRES) can provide excellent passband responses, giving either quasi-equiripple (QER) or minimum-loss (PRESML) responses with just a change in termination resistance.

Table 11.21
3 dB-down k & q Values for Quasi-Equiripple (QER) Ladder Filters

Order	*q*	k_{12}	k_{23}	*Shape Factor*	*Max Ripple (dB)*
4	0.9942	0.7660	0.5417	4.56	0.002
5	1.0316	0.7625	0.5391	3.02	0.018
6	1.0808	0.7560	0.5346	2.31	0.09
7	1.1876	0.7459	0.5275	1.90	0.16
8	1.2532	0.7394	0.5228	1.66	0.31
9	1.3439	0.7335	0.5187	1.50	0.42
10	1.4115	0.7294	0.5158	1.40	0.60
11	1.4955	0.7261	0.5134	1.33	0.72
12	1.5506	0.7235	0.5116	1.28	0.90

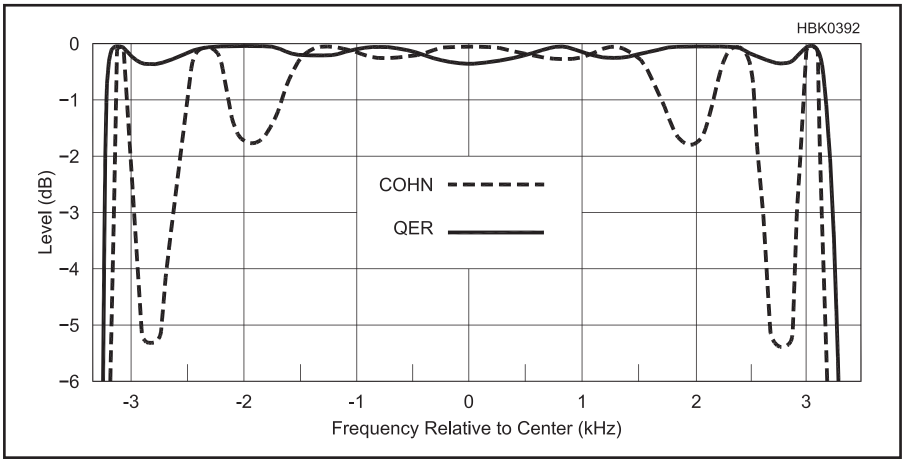


Fig 11.54 — Comparison of 8-pole Cohn min-loss passband response with that of the quasi-equiripple (QER) type. Note the almost equal ripple in the passband of the QER response.

ternative for fixed bandwidth speech applications where the Cohn *min-loss* pass band is poor. Two crystals are used in parallel to halve the motional inductance and double the motional capacitance of the resonators in the end sections, and although the two additional crystals do not increase the order of the filter by two, they do reduce the passband ripple substantially while maintaining the simplicity of design and construction offered by the Cohn *min-loss* filter. In addition, the group delay of the parallel-resonator-end-section (PRES) configuration is less than that of the Cohn *min-loss*. All the coupling capacitors are equal and the filter can be terminated to achieve a quasi-equiripple response (QER), so that its pass band resembles that of a Chebyshev design, or minimum loss (PRESML) with a response like that of the Cohn *min-loss*. **Fig 11.54** shows the Cohn *min-loss* and QER passband responses with infinite crystal Q_u for comparison. Values of k and q for QER filters from 4 to 12 poles are given in **Table 11.21,** along with the maximum ripple and shape factor for each order. The coupling capacitor value for any bandwidth can be determined from k_{23} using Eq 8. The end-section resonators formed by the two parallel crystals have twice the effective motional capacitance of the inner resonators, and since k_{12} is always 1.414 times k_{23} for the QER design the value of C_{12} will be the same as that calculated for C_{23} and all the other coupling capacitors, C, in the design.

$$C = f_o C_m/(BW\ k_{23}) \quad (8)$$

The termination resistance, R_T, must be calculated using half the motional inductance of a single crystal, as illustrated in Eq 9, where L_m is the motional inductance of one of the parallel crystals used in the end-sections.

$$R_T = 2\pi\ BW\ L_m/2q = \pi\ BW\ L_m/q \quad (9)$$

An alternative way of producing a QER filter, which has the added benefit of improving the symmetry of the ladder filter response, is to add transformers with bifilar windings to each end of the filter so that the end sections are effectively like crystal gate filters with C_1 approximately equal to $2C_o$. Normally, Dishal LSB ladder filters have an asymmetrical response because each crystal produces a transmission zero at f_p, on the high side of the filter pass band This asymmetry can be counteracted to a certain extent if the value of C_1 in each end section is adjusted to over-compensate for C_o and produce nulls on the low side of the pass band to even up the overall response. Whereas the presence of C_o causes the motional inductance of crystals to increase and their motional capacitance to decrease throughout the pass band, over compensating the end crystals to produce nulls on the low side has the opposite effect and the end-section crystals appear as if their motional capacitance is higher and their motional inductance lower, like two crystals in parallel. Therefore, they can be made to produce the QER type of response with suitable coupling and terminations. Ref 12 provides more information on this topic.

11.6.3 Crystal Characterization

Simple crystal filters can be constructed using cut-and-try methods, but sometimes the results are very disappointing. The only sure way to guarantee good results is to fully characterize the crystals beforehand, so that only the most suitable ones are used in a design appropriate for the crystal motional parameters and the application. When crystal parameters are known, computer modeling can be used to assess the effect of Q_u and C_o on bandwidth, before proceeding to the construction phase. In addition to the *Elsie* design program available from **www.arrl.org/arrl-handbook-reference**, AADE Filter Design and Analysis (**www.aade.com**) provides a free filter modeling program.

Crystal characterization can be done with very limited or very advanced test equipment, the main difference being the accuracy of the results. The *phase-zero* method for measuring C_m used in industry can be implemented by amateurs if a dual-beam oscilloscope is available to substitute as a phase detector — Ref 7 gives details of this method. However, many successful crystal filter constructors have achieved excellent results with a very limited amount of home-built test equipment. Ref 6 describes a simple switched-capacitor test oscillator for measuring C_m that was developed by G3UUR. His technique requires a frequency counter, a small 12 V power supply and very little construction effort. Using care and a more exact expression for C_m than the one given in Ref 6, this oscillator method can achieve results that are comparable with professional techniques. The circuit can also be modified to include relative Q or ESR measurement if a multimeter is available.

Values of Q_u for crystals can vary considerably, even within the same batch, and the ratio of the best to the worst, excluding dead ones, can be as high as 6 for cheap mass-produced crystals. This ratio can still be more than 2 for batches of high quality crystals. The relative activity of each crystal needs to be established to weed out the poor ones, and an estimate

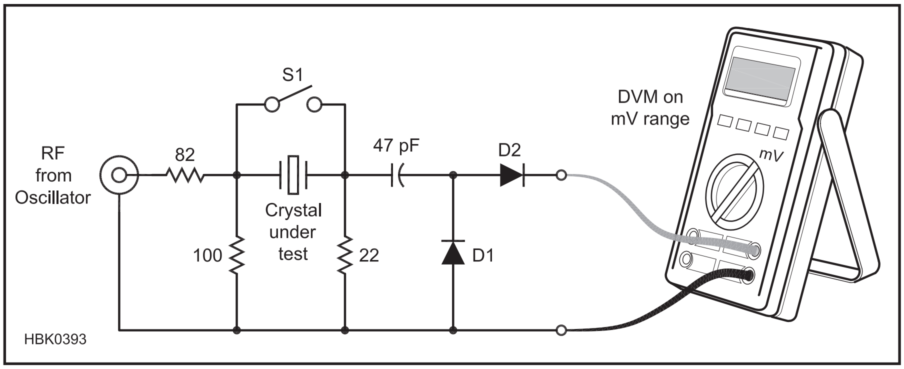


Fig 11.55 — A simple jig for measuring crystal ESR. The jig may be driven by a signal generator, or a modified crystal test oscillator, and the signal through the crystal detected on a DMM using the mV dc range. Resistors are ⅛ or ¼ W, 5%. D1 and D2 are small signal germanium or Schottky barrier diodes. S1 is a miniature pushbutton switch.

for Q_u is required to more accurately model the filter performance prior to construction. A modified version of the switched-capacitor crystal test oscillator is shown in **Fig 11.55**. The RF detector circuit at the output of the oscillator provides a means of assessing the relative activity of each crystal being tested. A DMM on a suitable dc voltage range attached across points M and G (ground) will give a digital readout roughly equal to the peak-to-peak RF voltage produced by each crystal. Crystals with higher Q values will produce higher output voltages, so each crystal can be ranked according to its output reading relative to others in the batch. For convenience, a socket should be used for the crystal under test. If crystals with wire leads are to be measured, this can be fashioned from a dual-in-line IC or transistor socket. There are also small PCB connectors that might make suitable sockets.

The formula for C_m presented in Ref 6 is much simplified and less accurate than the exact derivation for C_m. Better accuracy can be achieved with Eq 10, where F_1 is the frequency obtained with S1 closed and F_2 is the one with S1 open.

$$C_m = 2\,(F_2 - F_1)\,[C_S + C_O + C_R]/F_1 \qquad (10)$$

All capacitances in this equation are in pF and the frequencies in Hz. $C_R = 4C_S\,C_o/C_F$ and C_o is the total parallel capacitance of the crystal, including the contribution from the metal case — it's assumed that the metal case is floating during these measurements and is not grounded, or being held. The feedback capacitors, C_F, are 390 pF in **Fig 11.56** and the series capacitor, C_S, is 68 pF. C_o typically varies from 2.5 to 5.5 pF for HC49/U crystals in the 4 to 12 MHz range and about 1.5 to 3.5 pF for the smaller HC49/US crystals. LC meters that can measure down to 0.01 pF and 1 nH can be constructed using PIC technology. Commercial LC meters with amazingly good specifications are also available at moderate prices if a PIC LC meter seems too ambitious a project for home construction at this stage. Obviously, great care must be exercised to avoid stray capacitance and errors in setting zero when measuring such small values of capacitance.

When the value of C_o cannot be determined easily, or the utmost accuracy is not required, a simplified version of Eq 10 may be used. An average value for the type of crystals being tested can be assigned to C_o at the expense of a few percentage points loss in accuracy. For HC49/U crystals, a reasonable average for C_o is 3.75 pF and using the values for C_F and C_S shown in Fig 11.56**,** the more exact expression simplifies to Eq 11.

$$C_m = 148\,(F_2 - F_1)/F_1 \qquad (11)$$

Again, F_1 is the frequency registered on the counter when S1 is closed and F_2 when it's open. Both frequencies are in Hz and C_m comes out in pF. If more accuracy is required, the value of C_m obtained with Eq 11 can be used to estimate C_o from Eq 12 and that value used in Eq 10 to achieve a closer estimate for C_m.

$$C_o = 175\,C_m + 0.95 \qquad (12)$$

Also, the series-resonant frequency f_S for each crystal can be estimated by calculating the amount by which F_1 is higher than f_s and then subtracting that from F_1. This frequency difference, ΔF_1, is given by Eq 13 for the value of C_F used in Fig 11.56.

$$\Delta F_1 = C_m\,F_1/400 \qquad (13)$$

Ranking crystals according to their oscillator output is sufficient to be able to select the best (highest Q) crystals for a filter, but if you want to use computer modeling to correct for the influence of loss on bandwidth, then an estimate of ESR or Q_u is necessary. This can be done quite simply with the test oscillator shown in Fig 11.56, but requires just a little more time and effort than just ranking relative Q by RF output. There will be a spread of output levels corresponding to the range of Q values. Pair up two crystals from the low end of the spread with similar output figures (within 5%) and reasonably closely matched frequencies (within 50 Hz). Pair up another couple of crystals from the high end. Connect each pair of similar crystals in parallel and place them in the test oscillator a pair at a time with S_1 and S_2 closed. The output level should be much higher using a parallel pair than for each crystal alone. Now open S_2 and adjust the variable resistor VR1 until an output level is reached that is equal to the average of the levels previously obtained with the two crystals individually. After removing each pair check the resistance of VR1 using your DMM. Test points TP1 and TP2 are

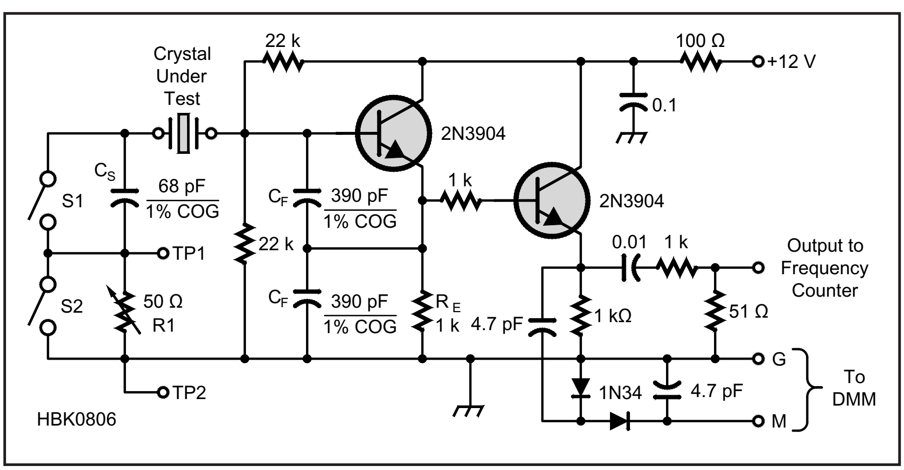


Fig 11.56 — Circuit of a modified switched-capacitor test oscillator which can be used to measure crystal motional parameters C_m and r_m.

Measuring Crystal Parameters

The following sidebar is an overview of the Jan/Feb 2016 *QEX* article "Crystal Parameter Measurements Simplified" by Chuck Adams, K7QO. The complete article is available as a PDF file on this book's CD-ROM.

This section describes a simple workbench technique developed by K7QO to characterize the electrical parameters of individual crystals. The parameters are discussed elsewhere in this section. (See also the section "Crystal Oscillators" in the Oscillators and Synthesizers chapter.) In addition to the test fixture, the procedure requires a digital RF signal generator, a frequency counter, an accurate L/C meter, and an RF voltmeter or RF probe used with a voltmeter.

The electrical equivalent circuit for a quartz crystal is shown in **Fig 11.A**. A test capacitor with a known value, C_X, is added in series with the crystal to shift the crystal's resonant frequency. By measuring the crystal with and without the effects of C_X, values for L_m, R_m, C_m, and C_0 can be determined.

In order to make the measurements reliably and to minimize the effects of stray capacitance and inductance a simple test fixture is used. The schematic of the test fixture is shown in **Fig 11.B**. The input and output impedance of the fixture is close to 50 Ω, but is not critical. R2 and R3 are small to reduce the loaded Q of the crystal. The resonant frequency of the crystal is not affected by the resistor values. Small values for R2 and R3 narrow the resonant peak and minimize the effects of stray capacitance. **Fig 11.C** is a photo of the test fixture being used.

The crystal is installed in the test fixture and measurements (described completely in the PDF article) are made with C_X both in-circuit and shorted by the jumper, JMP. From the shift in resonant frequency and amplitude described in the full article, all four primary crystal parameters can be obtained. In the PDF article, K7QO also describes a simple method of using a Colpitts crystal oscillator to find closely matched crystals in a batch.

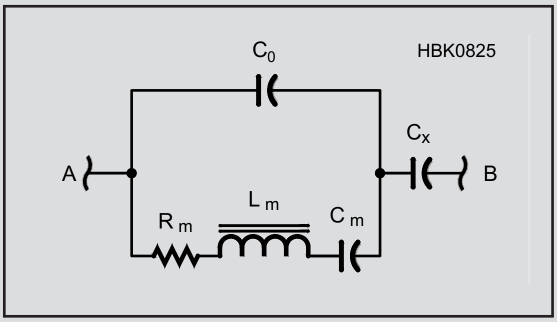


Fig 11.A — This schematic shows the equivalent circuit for a crystal in series with the test capacitor,

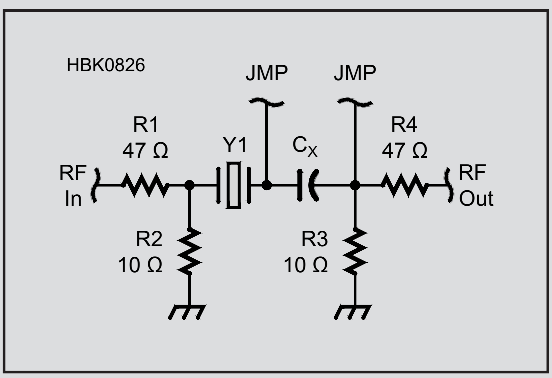


Fig 11.B — The schematic diagram for the K7QO test fixture. Note that R1 = R4 and R2 = R3. Y1 is the crystal being tested and C_X is the test capacitance added to shift the crystal's resonant frequency. K7QO uses a 47 pF capacitor for C_X. JMP represents a jumper to short-circuit C_X.

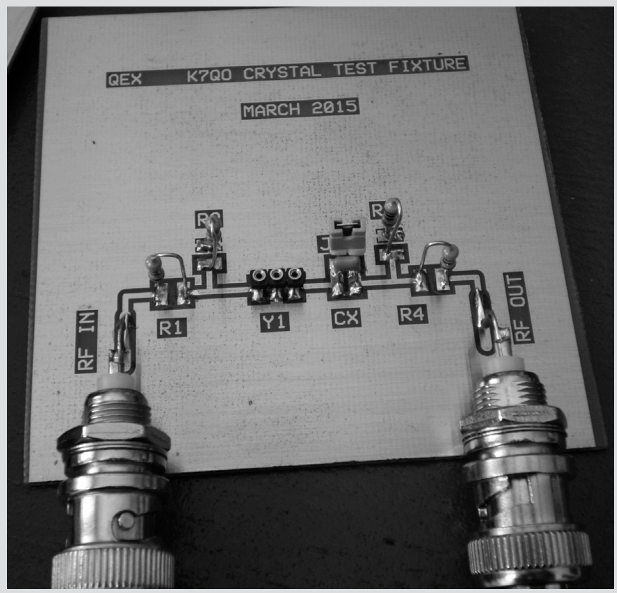


Fig 11.C — The finished crystal test fixture. The center pin of the crystal socket is grounded to reduce the effects of the parasitic capacitance of the socket which is in parallel with the crystal's C_0.

provided for such measurements and should be feed-through types mounted on the front panel or side of the oscillator box. The ESR value of the two similar crystals in the pair is approximately twice the value of the resistance measured across the test points. This method may be very crude, but it will give you an ESR value that is certainly better than 20%, and probably within 10% of the value measured by more accurate means. Once ESR values for the crystals at the extremes of the spread are established, one can be done in the middle of the range for good measure and values roughly assigned to those in between. Then, an average of the motional parameters and Q values of the set of crystals chosen for a particular filter can be used for modeling purposes.

Should a more accurate means of measuring crystal ESR be required, the phase-zero method or a VNA should be considered. Whatever the means of crystals characterization, a spreadsheet to record the data should be prepared beforehand and some means of marking the crystals with a number or letter organized. Sticky white dots are probably the most convenient way to identify each crystal. They adhere well to metal surfaces and can be written on with a ball-point pen. Alternatively, a permanent marker pen directly on the metal case could be tried, but is sometimes not all that permanent.

11.6.4 Crystal Filter Evaluation

The simplest means of assessing the performance of a crystal filter is to temporarily install it in a finished transceiver, or receiver, and use a strong on-air signal, or locally generated carrier, to run thru the filter pass band and down either side to see if there are any anomalies. Provided that the filter crystals have been carefully characterized in the first instance, and computer modeling has shown that the design is close to what's required, this may be all that's required to confirm a successful project. However, more elaborate checks on both the pass band and stop band can be made if a DDS signal generator, or vector network analyzer (VNA), is available. A test set-up for evaluating the response of a filter using a DDS generator requires an oscilloscope to display the response, whereas a VNA controlled via the USB port of a PC can display the response on the PC screen with suitable software — see the N2PK website **www.n2pk.com**. In addition, it can measure the phase and work out the group delay of the filter. Construction of a VNA is a very worthwhile project, and in addition to its many other applications it can be used to characterize the crystals prior to making the filter as well as evaluating filter performance after completion.

11.7 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 11.57**. 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 in response, 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 acoustic 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.

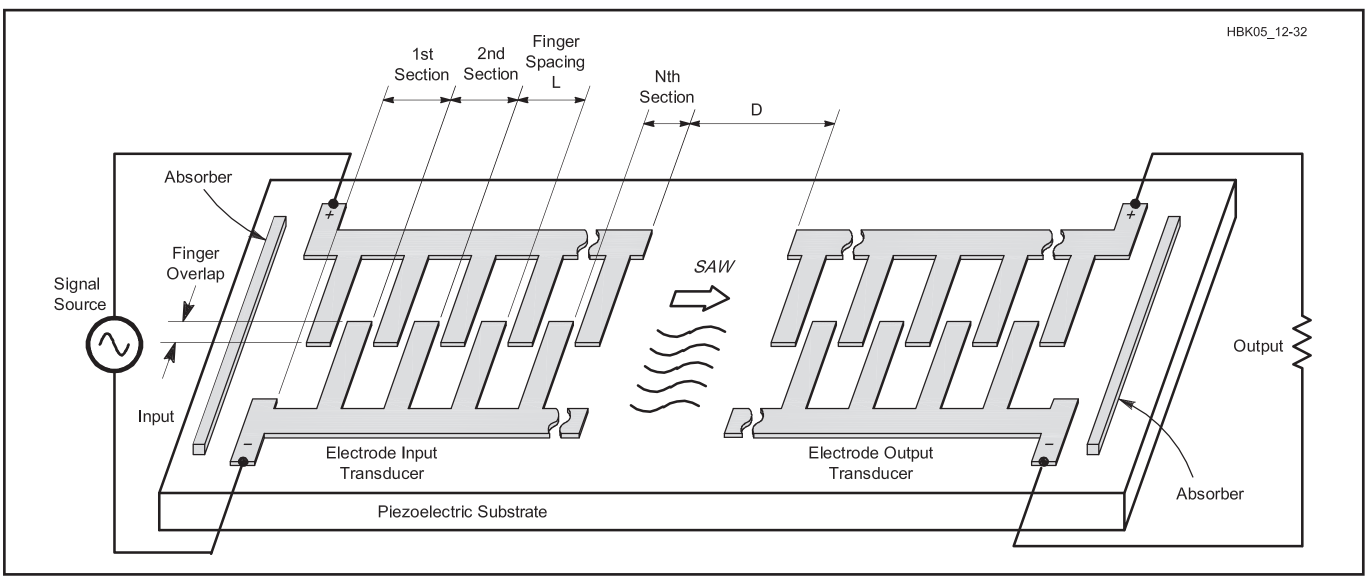


Fig 11.57 — The *interdigitated* transducer, on the left, launches SAW energy to a similar transducer on the right (see text).

11.8 Transmission Line Filters

LC filter calculations are based on the assumption that the reactances are *lumped*—that the physical dimensions of the components are considerably less than the operating wavelength. In such cases the unavoidable inter-turn parasitic capacitance associated with inductors and the unavoidable series parasitic inductance associated with capacitors are neglected as being 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.

Replacing lumped reactances with selected short sections of *Transverse Electromagnetic Mode* (TEM) transmission lines results in transmission line filters. (In TEM 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.

Coaxial cable transmission line filters are often used at HF and VHF frequencies. Stripline and microstrip 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 and are useful at frequencies where coaxial transmission lines are too lossy. A detailed treatment of the use of coaxial cable to form transmission line filters is presented in the **Transmission Lines** chapter. This section focuses on stripline and microstrip filters used at VHF and above.

11.8.1 Stripline and Microstrip Filters

Fig 11.58 shows three popular transmission lines used in transmission line filters. The circular coaxial transmission line (*coax*) shown in Fig 11.58A consists of two concentric metal cylinders separated by dielectric (insulating) material. 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 (*stripline*) shown in Fig 11.58B. 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 11.58B. Stripline Z_0 vs width (*w*) is plotted in **Fig 11.59**. Air-dielectric stripline technology is best for low bandwidth ($Q_L > 20$) band-pass filters.

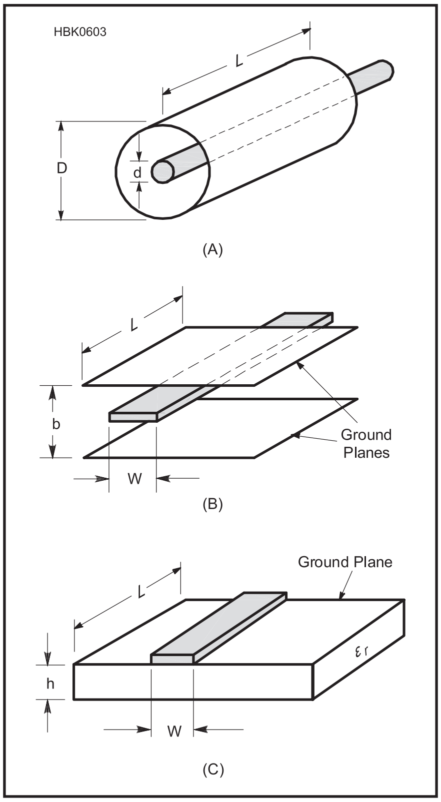


Fig 11.58 — Transmission lines. A: Coaxial line. B: Coupled stripline, which has two ground planes. C: Microstrip has only one ground plane.

The most popular transmission line at UHF and microwave is *microstrip* (unshielded stripline), shown in Fig 11.58C. It can be fabricated with standard printed-circuit processes and is the least expensive configuration. 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 (ε = 4.5) for the 50 MHz to 1 GHz frequency range and 0.031 inch Teflon (ε = 2.3) for frequencies above 1 GHz. Unfortunately, microstrip has the most loss of the three types of transmission line; therefore it is not suitable for narrow, high-Q, band-pass filters.

Conductor separation must be minimized or radiation from the line and unwanted coupling to adjacent circuits may become problems. Microstrip characteristic impedance and the effective dielectric constant (ε) are shown in **Fig 11.60**. 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 travels in the air above the substrate.

The characteristic impedance for stripline and microstrip-lines that results in the lowest loss 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 even though the loss at that characteristic impedance is not a minimum for that type of transmission line.

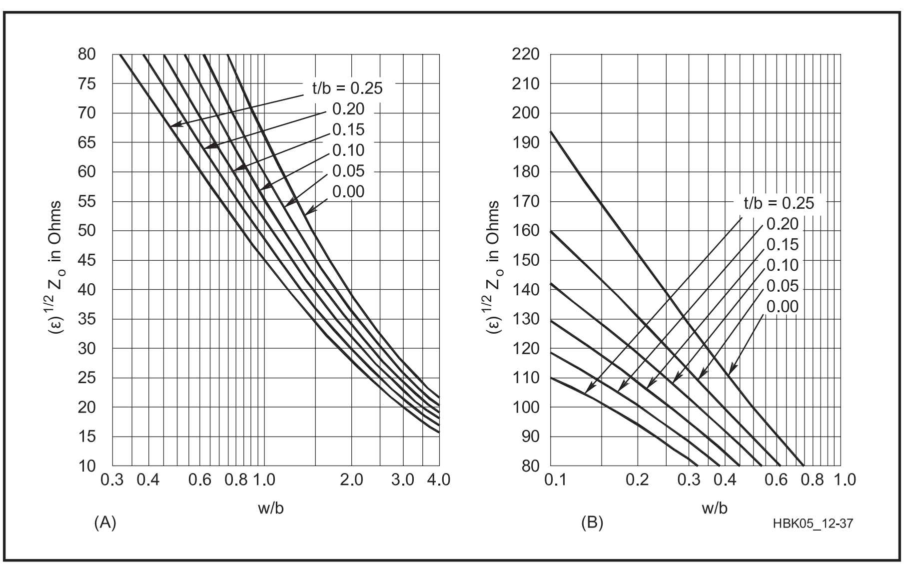


Fig 11.59 — The Z_0 of stripline varies with *w*, *b* and *t* (conductor thickness). See Fig 11.58B. The conductor thickness is *t* and the plots are normalized in terms of *t/b*.

Z_0 Ω	ε =1 (AIR) W/h	ε =2.3 (RT/Duroid) W/h	$\sqrt{\varepsilon_e}$	ε =4.5 (G−10) W/h	$\sqrt{\varepsilon_e}$
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				

HBK05_12-38

Fig 11.60 — Microstrip parameters (after H. Wheeler, *IEEE Transactions on MTT*, March 1965, p 132). ε_e is the effective ε.

11.8.2 Transmission Line Band-Pass Filters

Band-pass filters can also be constructed from transmission line stubs. (See the **Transmission Lines** chapter for information on stub behavior and their use as filters at HF and VHF.) At VHF the stubs can be considerably shorter than a quarter-wavelength (¼ λ), 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 11.61** 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.

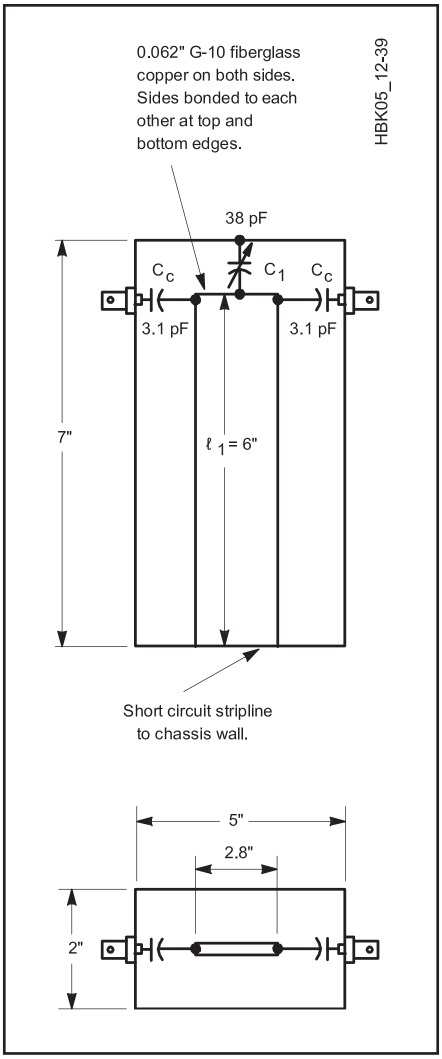


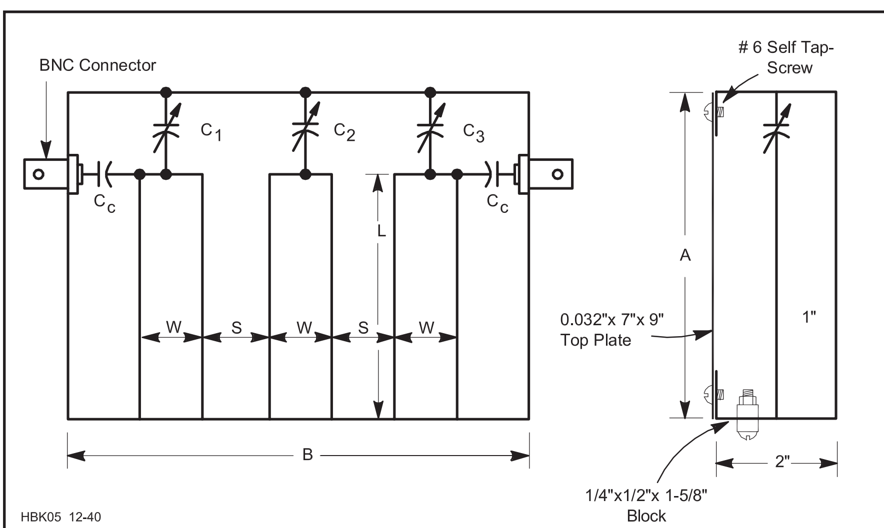
Fig 11.61 — This 146 MHz stripline band-pass filter has been measured to have a Q_L of 63 and a loss of approximately 1 dB.

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 11.62**. These filters are constructed in 2 × 7 × 9 inch aluminum boxes.

11.8.3 Quarter-Wave Transmission Line Filters

The reactance of a ¼ λ shorted-stub is infinite, as discussed in the **Transmission Lines** chapter. 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 11.63**. Design graphs for round-rod interdigital filters are given in Ref 9. 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 the 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



Dimension (inches)	*52 MHz*	*146 MHz*	*222 MHz*
A	9	7	7
B	7	9	9
L	7⅜	6	6
S	1	1 1/16	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 11.62 — This Butterworth filter is constructed in combline. It was originally discussed by R. Fisher in December 1968 *QST*.

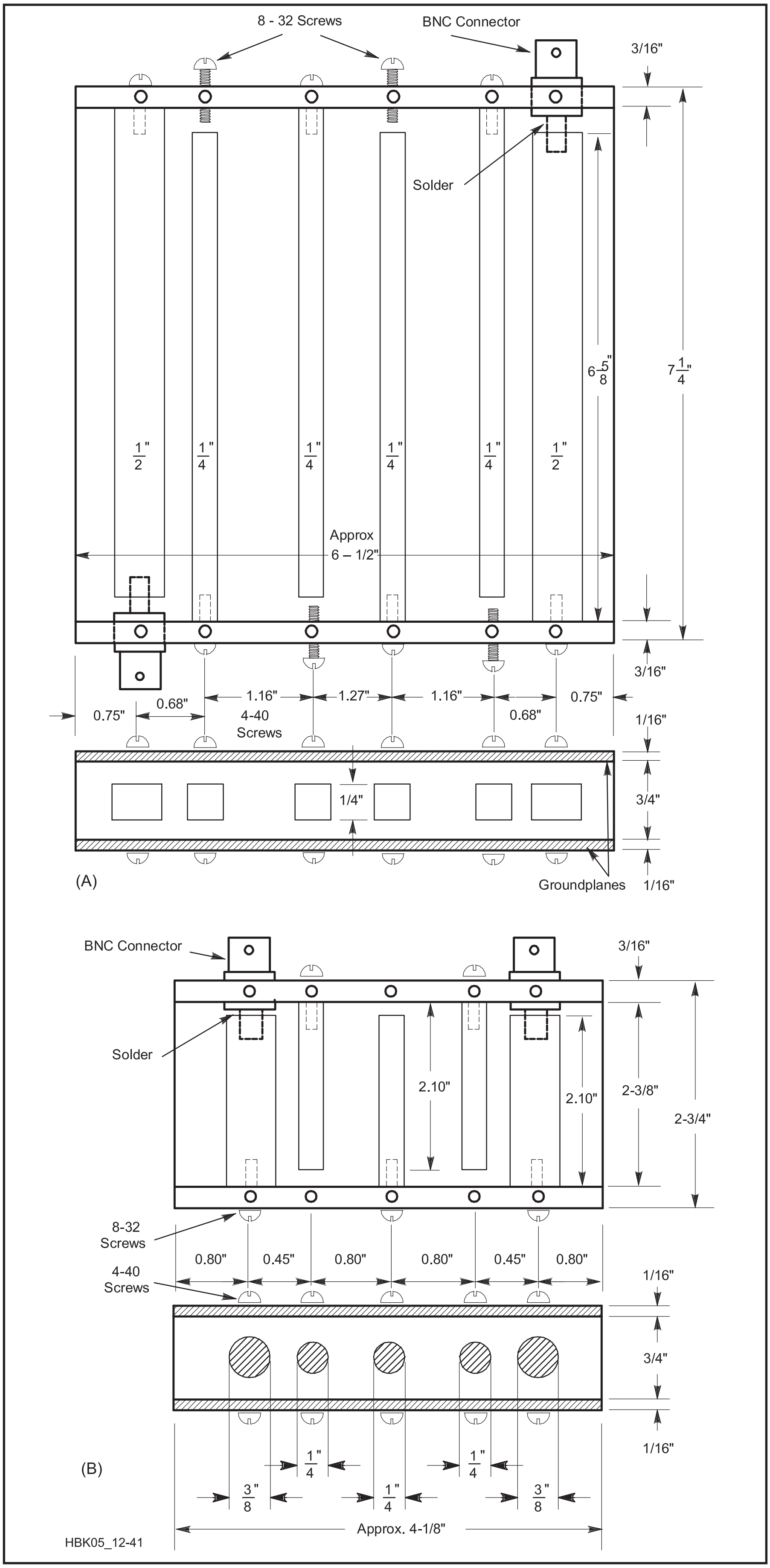


Fig 11.63 — These 3-pole Butterworth filters for 432 MHz (shown at A, 8.6 MHz bandwidth, 1.4 dB passband loss) and 1296 MHz (shown at B, 110 MHz bandwidth, 0.4 dB passband loss) are constructed as interdigitated filters. The material is from R. E. Fisher, March 1968 *QST.*

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.

11.8.4 Emulating LC Filters with Transmission Line 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 11.64A** (on page 32) that emulates the LC filter shown in Fig 11.64B. 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 11.64B, 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 $\varepsilon = 4.5$. Then, from Fig 11.60, *w* is 0.11 inch and $\ell_C = 0.125\ \lambda_g$ for the 50 Ω capacitive stubs. Also, from Fig 11.60, *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 inches.

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 han the Butterworth filter in Fig 11.63 which has spurious responses in the 1296 MHz band.

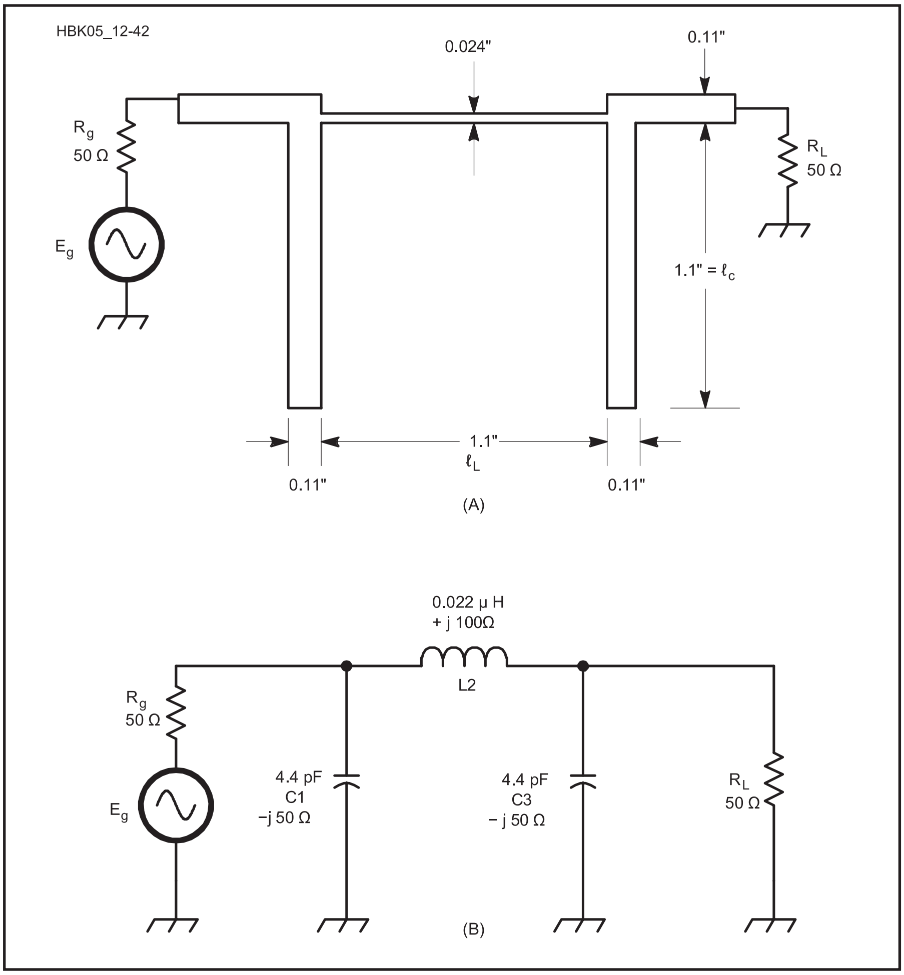


Fig 11.64 — A microstrip 3-pole emulated-Butterworth low-pass filter with a cutoff frequency of 720 MHz. A: Microstrip version built with G-10 fiberglass board (ε = 4.5, h = 0.062 inches). B: Lumped LC version of the same filter. To construct this filter with lumped elements very small values of L and C must be used and stray capacitance and inductance would have to be reduced to a tiny fraction of the component values.

11.9 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 environments. 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 because they are smaller and easier to build than coaxial cavity resonators, although their Q is not as high as that of cavities. 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.

The helical resonator was described by the late Jim Fisk, W1HR, in a June 1976 *QST* article. [Ref 10] The resonator is described 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 propagation velocity along the axis of the helix. 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 11.65**. Although the shield may be any shape, only round and square shields will be considered here.

11.9.1 Helical Resonator 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

$$Q_U = 50\ D \sqrt{f_0} \qquad (14)$$

where

D = inside diameter of the shield, in inches

f_0 = 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 imperfec-

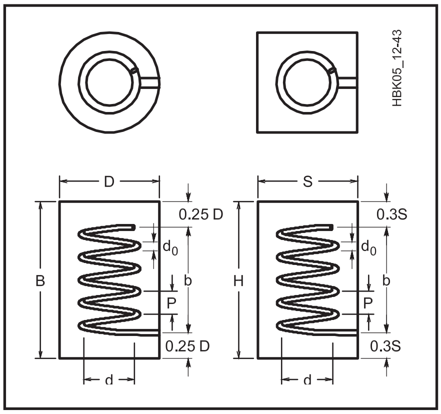


Fig 11.65 — Dimensions of round and square helical resonators. The diameter, D (or side, S) is determined by the desired unloaded Q. Other dimensions are expressed in terms of D or S (see text).

tions 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 QU by about 3% before making the calculation). The fringing capacitance at the open-circuit end of the helix is about 0.15D 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:

$$N = \frac{1908}{f_0 D} \quad (15A)$$

$$P = \frac{f_0 D^2}{2312} \quad (15B)$$

$$Z_0 = \frac{99{,}000}{f_0 D} \quad (15C)$$

$$N = \frac{1590}{f_0 S} \quad (15D)$$

$$P = \frac{f_0 S^2}{1606} \quad (15E)$$

$$Z_0 = \frac{82{,}500}{f_0 S} \quad (15F)$$

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 11.66** 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.

Design of filter dimensions can be done using the nomographs in this section or with computer software. The program *Helical* for designing and analyzing helical filters is available from **www.arrl.org/arrl-handbook-reference**. Use of the nomographs is described in the following paragraphs.

Fig 11.67 simplifies calculation of these dimensions. Note that these ratios result in a helix with a length 1.5 times its diameter,

Helical Filter Design Software

Helical, a *Windows* program by Jim Tonne, W4ENE, for design and analysis of helical-resonator bandpass filters usually used in the VHF and UHF frequency ranges, is available from **www.arrl.org/arrl-handbook-reference**.

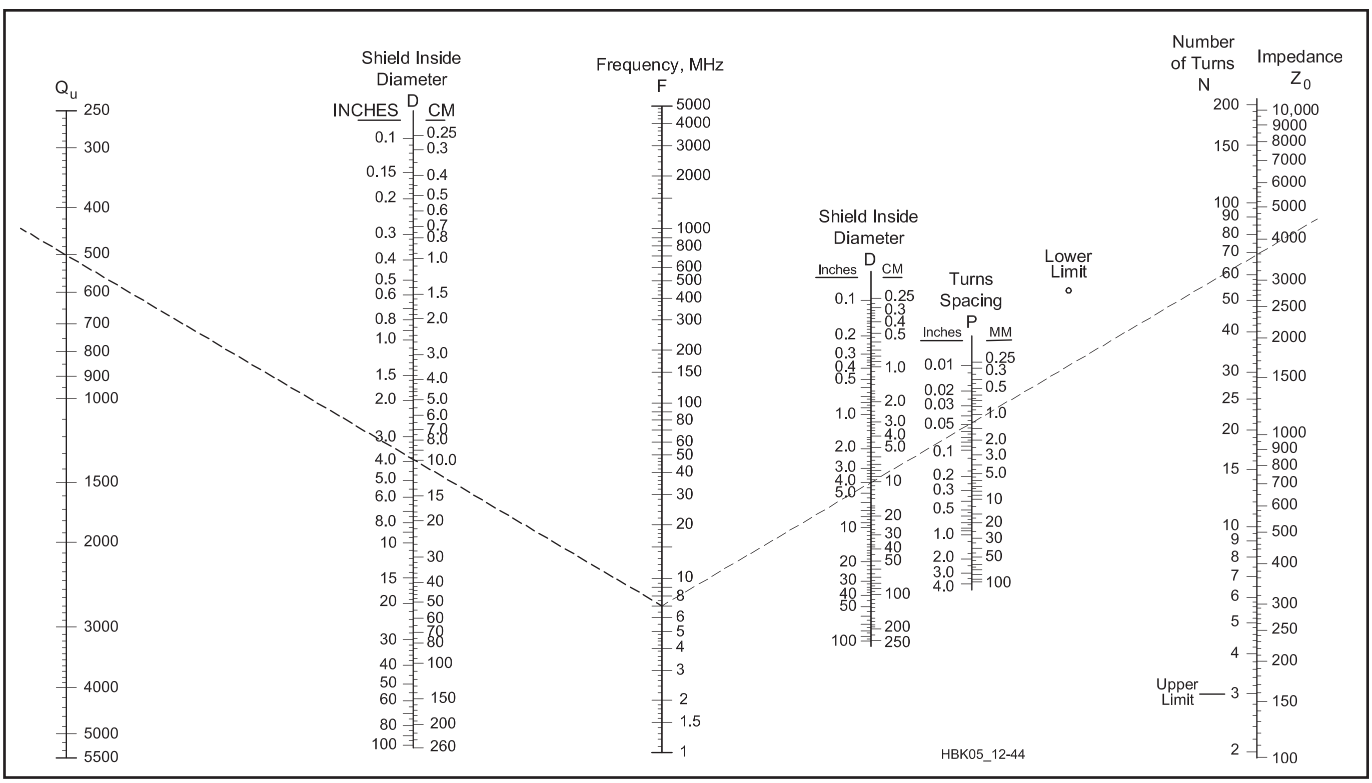


Fig 11.66 — The design nomograph for round helical resonators starts by selecting Q_U and the required shield diameter. A line is drawn connecting these two values and extended to the frequency scale (example here is for a shield of about 3.8 inches and Q_U of 500 at 7 MHz). Finally the number of turns, 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).

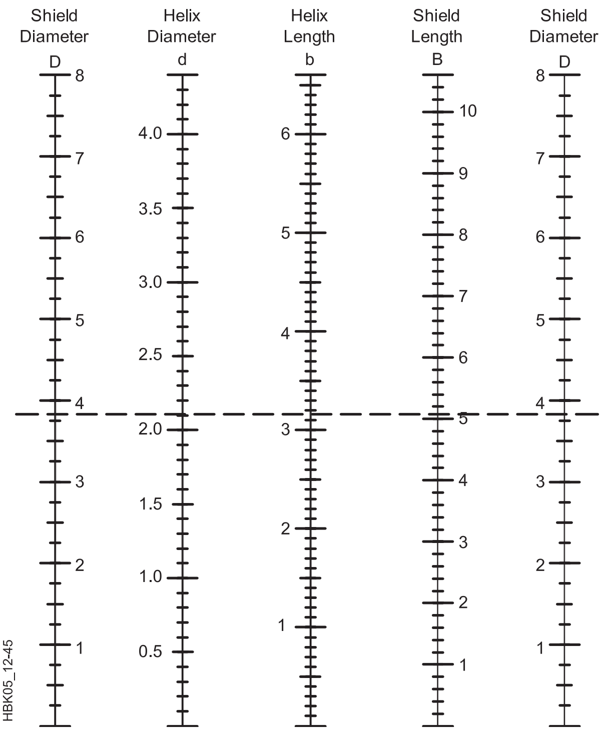


Fig 11.67 — The helical resonator is scaled from this design nomograph. Starting with the shield diameter, the helix diameter, d, helix length, b, and shield length, B, can be determined with this graph. The example shown has a shield diameter of 3.8 inches. This requires a helix mean diameter of 2.1 inches, helix length of 3.1 inches, and shield length of 5 inches.

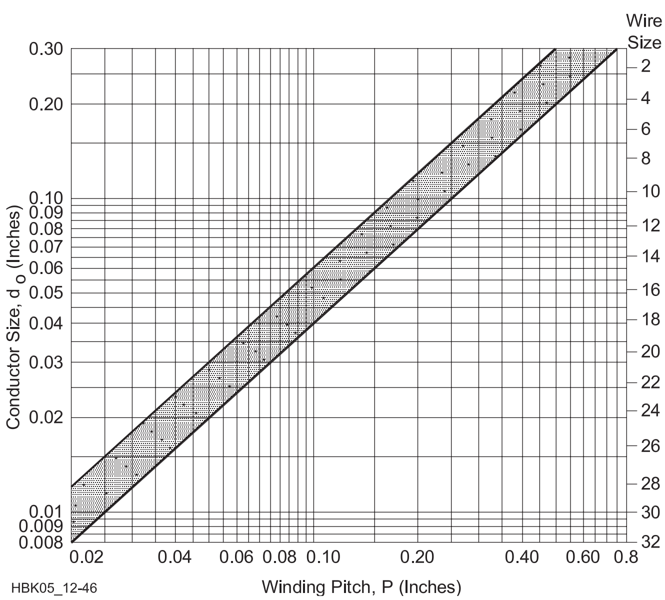


Fig 11.68 — This chart provides the design information of helix conductor size vs winding pitch, P. For example, a winding pitch of 0.047 inch results in a conductor diameter between 0.019 and 0.028 inch (#22 or #24 AWG).

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 11.68**.

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 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 non-helical). This condition occurs when the helix has fewer than three turns (the "upper limit" on the design nomograph of Fig 11.66).

11.9.2 Helical Filter 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 minimized 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.

11.9.3 Helical Resonator Tuning

A carefully built helical resonator designed from the nomograph of Fig 11.66 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 11.67, 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.

11.9.4 Helical Resonator Insertion Loss

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

$$I_L = 20 \log_{10} \left(\frac{1}{1 - \frac{Q_L}{Q_U}} \right) \qquad (16)$$

where
Q_L = loaded Q
Q_U = unloaded Q

This is plotted in **Fig 11.69**. 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.

11.9.5 Coupling Helical Resonators

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 11.70**. 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).

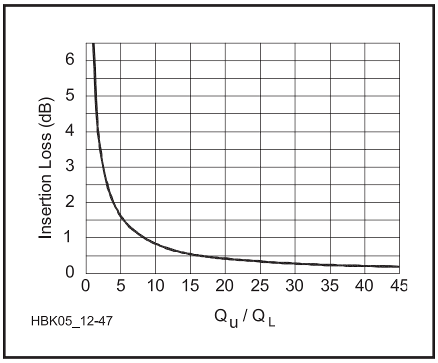


Fig 11.69 — The ratio of loaded (Q_L) to unloaded (Q_U) Q determines the insertion loss of a tuned resonant circuit.

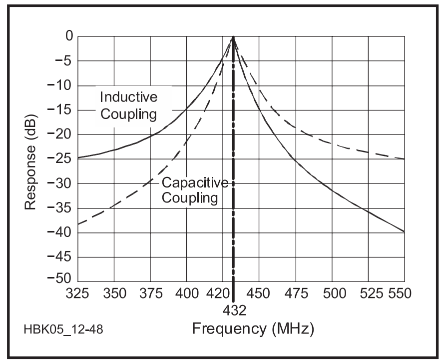


Fig 11.70 — This response curve for a single-resonator 432 MHz filter shows the effects of capacitive and inductive input/output coupling. The response curve can be made symmetrical on each side of resonance by combining the two methods (inductive input and capacitive output, or vice versa).

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:

$$K = \frac{R_b}{Z_0} = 0.785 \left(\frac{1}{Q_L} - \frac{1}{Q_U} \right) \qquad (17)$$

11.10 Use of Filters at VHF and UHF

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.

11.11 Filter Projects

The filter projects to follow are by no means the only filter projects in this book. Filters for specific applications may be found in other chapters of this *Handbook*. Receiver input filters, transmitter filters, inter-stage filters and others can be extracted 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 **RF Interference** chapter.

11.11.1 Crystal Ladder Filter for SSB

One of the great advantages of building your own crystal filter is the wide choice of crystal frequencies currently available. This allows the filter's center frequency to be chosen to fit more conveniently between adjacent amateur bands than those commercially available on 9 and 10.7 MHz do.

The 8.5 MHz series crystals used in this project were chosen in preference to ones on 9 MHz to balance the post-mixer filtering requirements on the 30 and 40 meter bands, and reduce the spurious emission caused by the second harmonic of the IF when operating on the 17 meter band. The crystals have the equivalent circuit shown in **Fig 11.71** and were obtained from Digi-Key (part number X418-ND).

The crystal-to-case capacitance becomes part of the coupling capacitance when the case is grounded, as it should be for best ultimate attenuation. The motional capacitance and inductance can vary by as much as ±5% from crystal to crystal, although the resonant frequency only varies by ±30 ppm. The ESR (equivalent series resistance) exhibits the most variation with mass-produced crystals, and in the case of these 8.5 MHz crystals can be anything from under 25 Ω to over 125 Ω.

Since the frequency variation has a spread of more than twice what can be tolerated in a 2.4 kHz SSB filter design, and the ESR is so variable, individual crystals need to be checked and selected for frequency matching and ESR. In order to allow plenty for selection, 30 crystals were purchased for the prototype. A frequency counter capable of making measurements with 10 Hz resolution, a DMM with a dc mV range and two simple self-constructed test circuits are required for crystal selection.

The project filter shown in **Fig 11.72** is based on a 7-pole QER (quasi-equiripple) design that has a shape factor of 1.96 and offers simplicity, flexibility, and a great passband. The 39 pF coupling capacitors can be silver mica or low-k disc ceramic types with a tolerance of ±2%. The termination resistance shown in the diagram has been reduced to allow for the loss in the end crystals (roughly 36 Ω for each parallel pair). The total termination resistance should be 335 Ω, theoretically, and the actual value used should be adjusted to reflect the effective ESR of your end pairs.

If more than nine suitable crystals are available from the batch tested, the order of the filter can be increased without changing the value of the coupling capacitors. The bandwidth will change very little — by less than +1% per unit increment in order. The termination resistance will need to be reduced as the order is increased, however, by the ratio of the q1 values given in Table 11.21. Increasing the order will improve the shape factor and further reduce adjacent channel interference, but also increase the passband ripple.

Effectively, each crystal in the middle section of the filter has a load of around 20.25 pF because of the 39 pF capacitors and 1.5 pF crystal-to-case capacitance on either side. The end crystal pairs are also shifted up in frequency as if they have the same load. Therefore, for best matching, all crystals should be checked in an oscillator with this load capacitance. The oscillator for this test is shown in **Fig 11.73**, and it will be seen that it shares many common parts with the VXO circuit used for ESR measurements shown in Fig 11.56. The 22 pF input capacitor and two 470 pF feedback capacitors in series present about 20 pF to the crystal under test. A transistor socket can be used as a quick means of connecting the crystals in circuit, rather than soldering and de-soldering each one in turn. If test crystals are soldered, adequate time should be allowed for them to cool so that their frequency stabilizes before

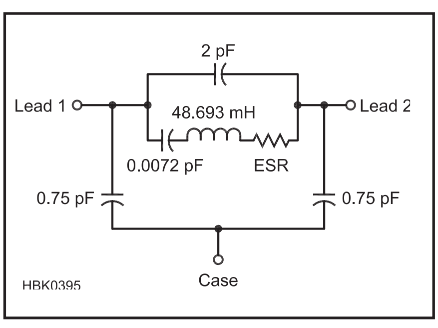


Fig 11.71 — Real electrical equivalent circuit of an ECS 8.5 MHz crystal with its metal case grounded.

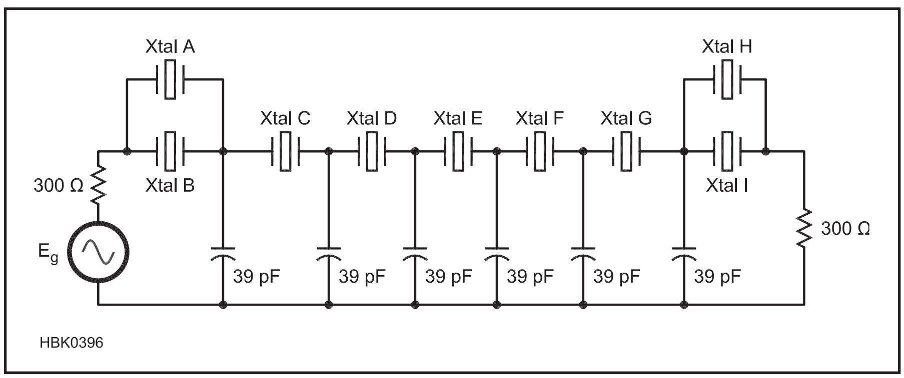


Fig 11.72 — 7-Pole 2.4 kHz QER ladder filter using ECS 8.5 MHz crystals (ECS-85-S-4) from Digi-Key (X418-ND).

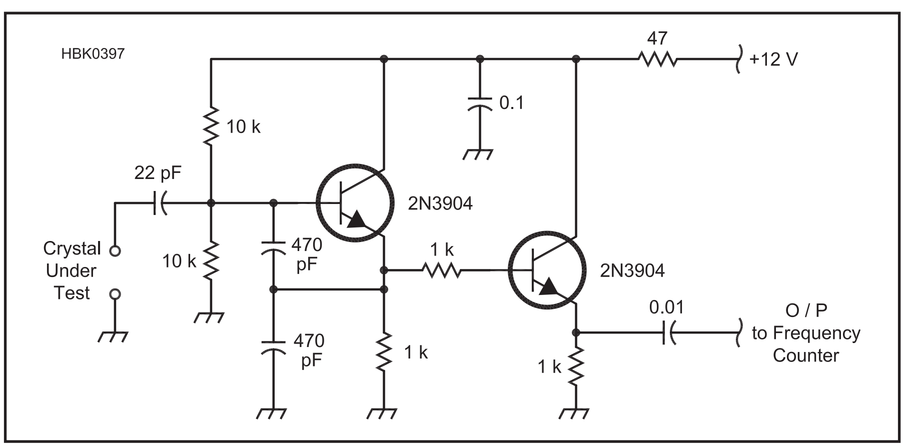


Fig 11.73 — Test oscillator with 20 pF capacitive load for matching crystal frequencies.

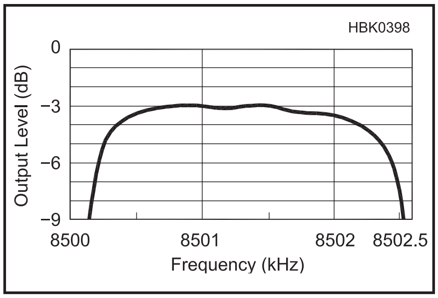


Fig 11.74 — Passband response of prototype 8.5 MHz QER crystal ladder filter with 3 dB insertion loss.

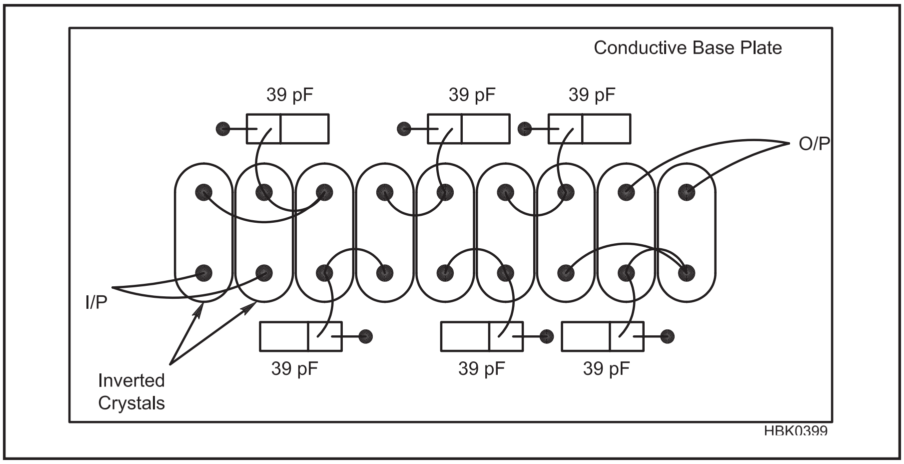


Fig 11.75 — Suggested construction arrangement using inverted crystals with their cases soldered to a conductive base plate and direct point-to-point wiring using only component leads.

Table 11.22
Measured Parameters for the 9 Crystals Selected for Use in the Prototype 7-pole 2.4 kHz SSB Filter.

Xtal	*Freq (20pF)*	*ESR (Ω)*	*Q*
A	8501.342 kHz	87	30k
B	8501.441 kHz	60	43k
C	8501.482 kHz	33	78k
D	8501.557 kHz	32	81k
E	8501.530 kHz	27	97k
F	8501.372 kHz	28	93k
G	8501.411 kHz	52	50k
H	8501.477 kHz	108	24k
I	8501.400 kHz	54	48k

a reading is taken.

The crystals need to be numbered in sequence with a permanent marker to identify them, and their 20 pF load frequencies recorded in a table or spreadsheet as they are measured. When this is complete, the oscillator can then be converted to a VXO, as shown in Fig 11.56, and using this and the jig in Fig 11.55 the ESR of each crystal can be assessed by comparison with metal or carbon film resistors that give the same output readings on the DMM.

The variable capacitor used in the VXO to check the crystals for the prototype filter was a polyvaricon type with maximum capacitance of 262 pF and the inductor was a 10 µH miniature RF choke. If smaller values of variable capacitor are used the inductor value will need to be increased in order to ensure the frequency swing is great enough to tune thru the peak of the lowest frequency crystal in the batch.

Once all the crystals have been checked and their 20 pF frequencies and ESR values recorded, the best selection strategy is to look over the figures for a group of nine crystals that have a spread of frequencies of less than 10% of the 2.4 kHz bandwidth. They can then be considered for position in the filter on the basis of their ESR values. The ones with the lowest values should be selected for the middle positions, with the lowest of all as the central crystal.

The crystals with the highest ESR values should be paired up for use as the end parallel crystals because their loss can be absorbed in the terminations. Try to use a pair of crystals with similar values of overall ESR if you want the two terminating resistors to have the same value after subtracting the effective loss resistance of each pair of end crystals from the required theoretical termination resistance of 335 Ω. **Table 11.22** shows how the nine crystals chosen from the batch of 30 obtained for the prototype were selected for position to obtain the passband curve shown in **Fig 11.74**. It can be seen that the spread of frequencies for a load of 20 pF in this case was 215 Hz, and the values of ESR varied from 27 Ω (best) to 108 Ω (worst). The prototype bandwidth (–6 dB) came out at 2.373 kHz with 39 pF coupling capacitors that were all approximately 1% high of their nominal value. A random selection of ±2% capacitors should produce a bandwidth of between 2.35 and 2.45 kHz. If a wider bandwidth is required, 33 pF coupling can be used instead of 39 pF, with the theoretical termination resistance increased to 393 Ω. This should provide a bandwidth of around 2.8 kHz.

The motional capacitance of the ECS 9 MHz series crystals available from Digi-Key (part number X419-ND) should be only slightly higher than that of the 8.5 MHz crystals, so they could be used in this design with a corresponding increase in bandwidth — probably around 300 Hz, making the overall filter bandwidth at 9 MHz about 2.7 kHz (±60 Hz with 2% tolerance, 39 pF capacitors).

Construction can be a matter of availability and ingenuity. Reclaimed filter cans from unwanted wide-bandwidth commercial crystal filters could be utilized if they can be picked up cheaply enough. Otherwise, inverted crystals can be soldered side by side, and in line, to a base plate made of a piece of PCB material, copper, or brass sheet as shown in **Fig 11.75**. The coupling capacitors can then be soldered between the crystal leads and the base plate, or crystal cases, depending on which is more convenient. Excellent ultimate attenuation (>120 dB) can be achieved using direct wiring and good grounding like this, particularly if lead lengths are kept as short as possible and if additional shielding is added at each end of the filter to prevent active circuitry at either end of the filter from coupling to and leaking signal around it. A case made from pieces of the same material can be soldered around the base plate to form a fully shielded filter unit with feed-thru insulators for the input and output connections.

11.11.2 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 11.76**. The standard capa-

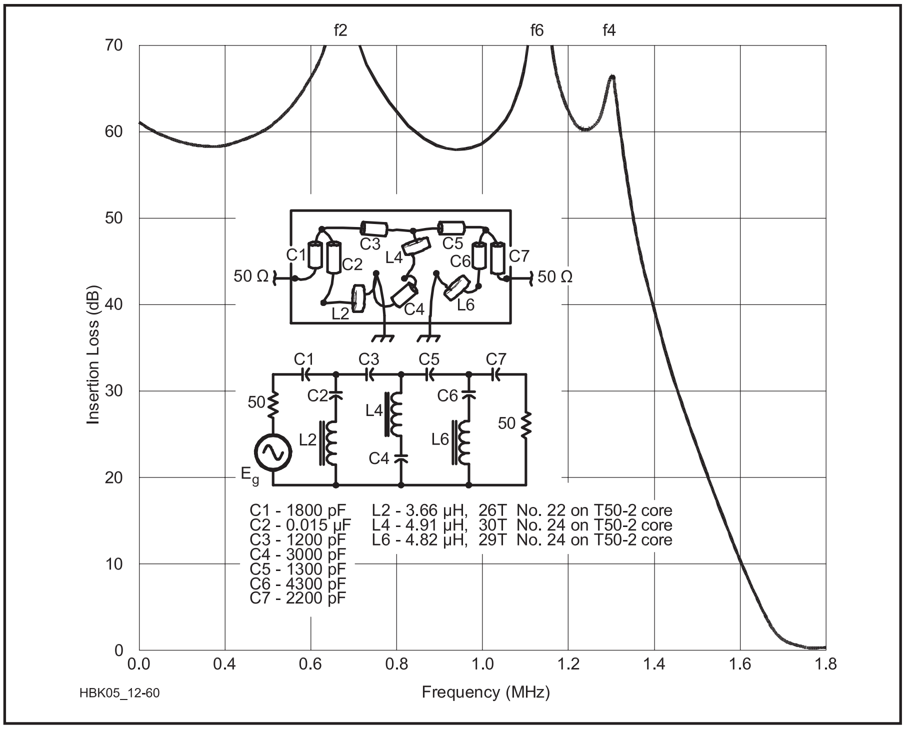


Fig 11.76 — Schematic, layout and response curve of the broadcast band rejection filter.

ci-tor 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 11.77**. Use polypropylene film-type 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 11.76.) For the 80 meter 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 11.76. 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).

11.11.3 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.

CIRCUIT DESCRIPTION

The way the circuit works is quite simple. Referring to **Fig 11.78**, 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 11.79**, to shield it from other unwanted signals and is connected

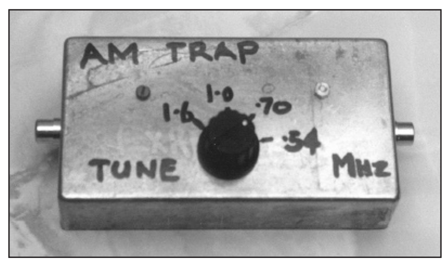


Fig 11.79 — The wave trap can be roughly calibrated to indicate the frequency to which it is tuned.

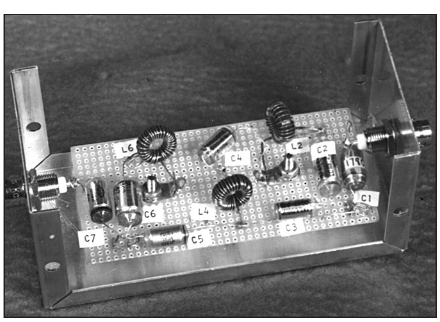


Fig 11.77 — The filter fits easily in a 2 × 2 × 5 inch enclosure. The version in the photo was built on a piece of perfboard.

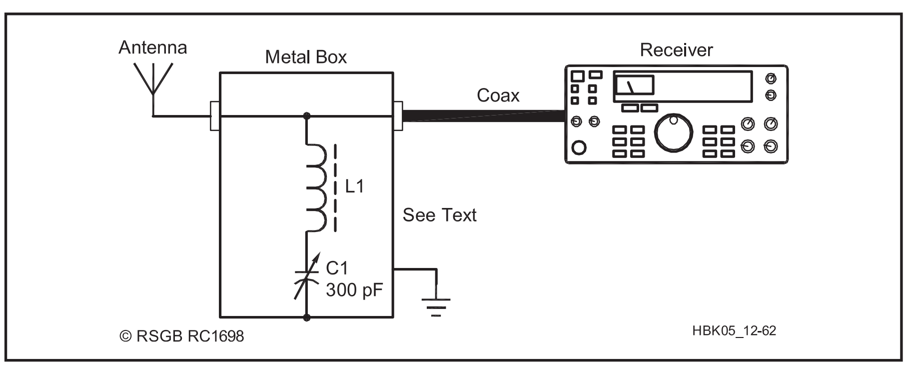


Fig 11.78 — The wave trap consists of a series tuned circuit, which 'shunts' signals on an unwanted frequency to ground.

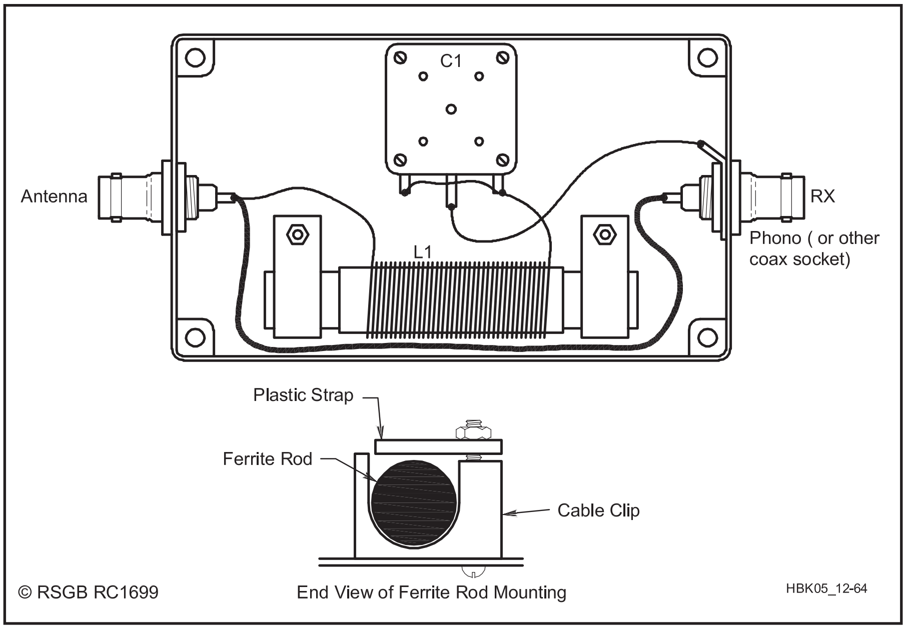


Fig 11.80 — Wiring of the wave trap. The ferrite rod is held in place with cable clips.

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.

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.

11.11.4 Optimized Harmonic Transmitting 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, W4ENE, 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

as shown in Fig 11.78. 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 11.80**. 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. Let's say you live near an AM transmitter on 1530 kHz, and the signals break through on

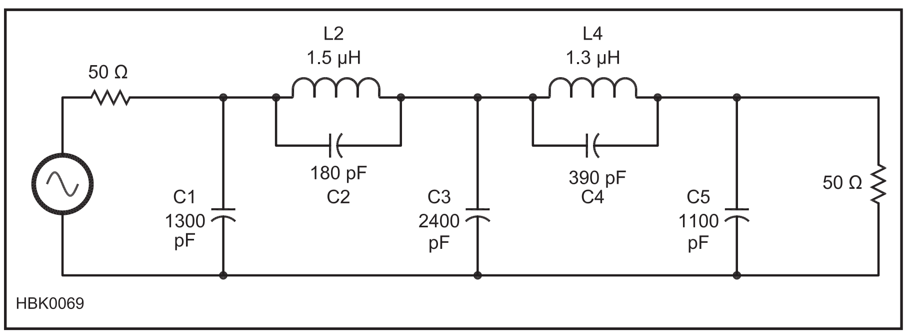


Fig 11.81 — Optimized low-pass filter. This design is for the 80 meter amateur band. It is similar to a Cauer design but the parts values have been optimized as described in the text and in the Sep/Oct 1998 issue of *QEX*.

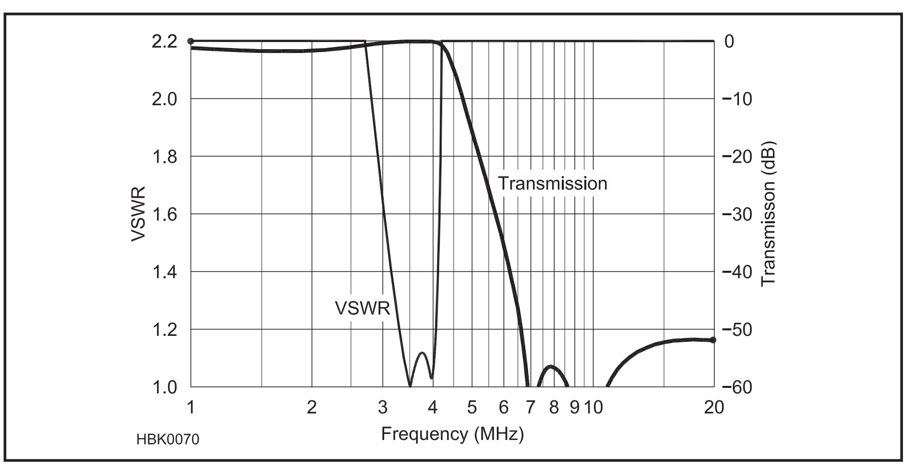


Fig 11.82 — Responses of the filter shown in Fig 11.81. Note the low values of SWR from 3.5 to 4 MHz. At the same time the harmonics are attenuated to meet regulations. Responses for the other amateur bands are very similar except for the frequency scaling.

values for the 3.5 to 4.0 MHz amateur band is shown in **Fig 11.81**. The responses of that filter are shown in **Fig 11.82**.

Component values for the 160 meter through the 6 meter amateur bands are shown in **Table 11.23**. 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 available from **www.arrl.org/arrl-handbook-reference**.

Table 11.23
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

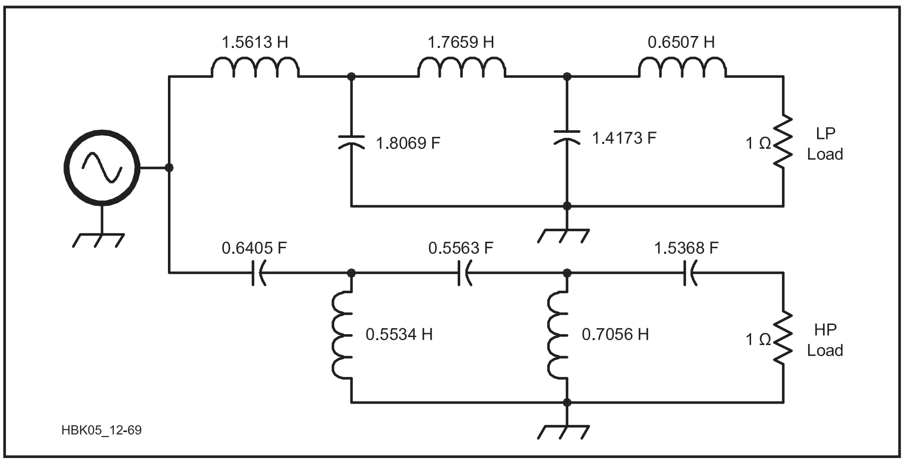


Fig 11.83 — Low-pass and high-pass prototype diplexer filter design. The low-pass portion is at the top, and the high-pass at the bottom of the drawing. See text.

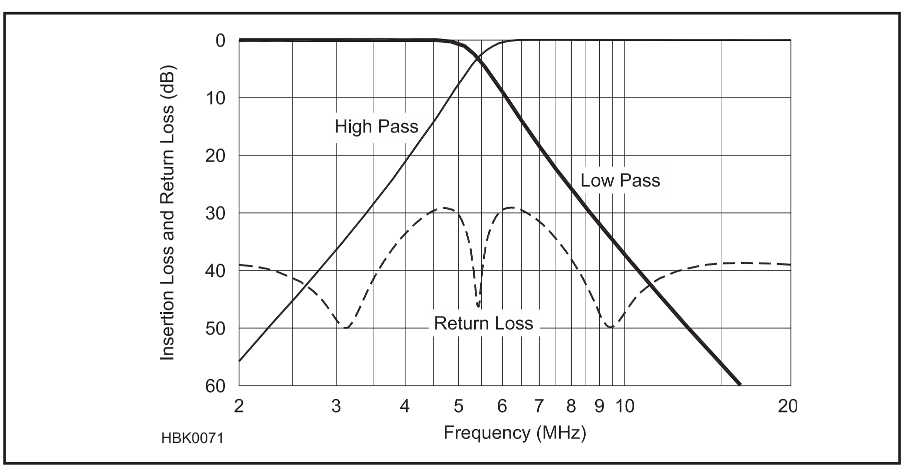


Fig 11.84 — Response for the low-pass and high-pass portions of the 80 meter diplexer filter. Also shown is the return loss of the filter.

11.11.5 Diplexer Filter

This section, covering diplexer filters, was written by William E. Sabin, WØIYH. The diplexer is helpful in certain applications, such as frequency mixer terminations.

(The terms "diplexer" and "duplexer" are often confused. A duplexer is a device such as a circulator that allows a transmitter and a receiver to use the same antenna *without* the use of filters. Diplexers use filters so that the signal frequencies must be far apart, such as on different bands. 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 of the devices to which they are attached. (For example, impedances far from 50 Ω outside the operating frequency range may cause an amplifier to develop parasitic oscillations.)

Fig 11.83 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

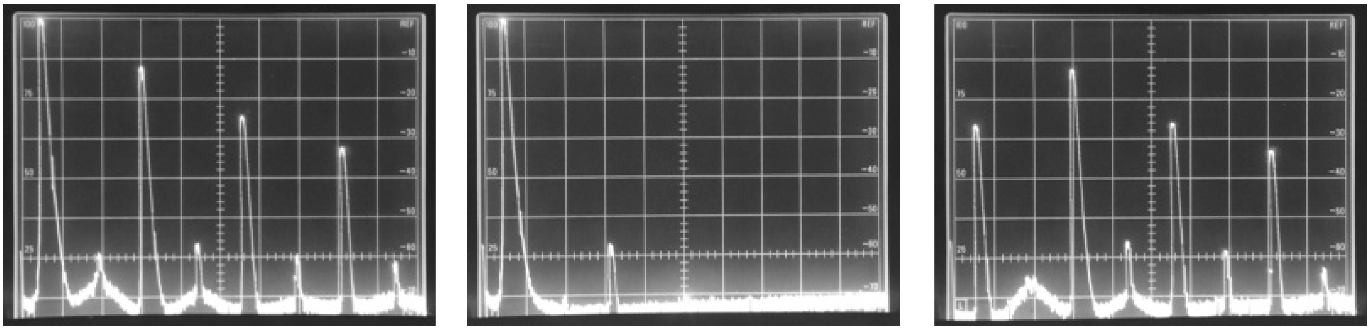
$L_{P(LP)}$ and $C_{P(LP)}$ are LP prototype values
K = 1.005 (in this specific example)
R = 50 Ω
f_{CO} = the cutoff (–3 dB response) frequency in Hz.

For the HP segment:

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

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

Fig 11.84 shows the LP and HP responses of a diplexer filter for the 80 meter band. The



(A) (B) (C)

Fig 11.85 — 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.

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 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 11.85A shows the unfiltered output 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 11.85B. 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 11.85C 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, supply voltage, 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 11.84, 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.

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

DiplexerDesigner, a *Windows* program for design and analysis of diplexer filters by Jim Tonne, W4ENE, is available from **www.arrl.org/arrl-handbook-reference**.

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 the ARRL website 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.

11.11.6 High-Performance, Low-Cost 1.8 to 54 MHz Low-Pass Filter

The low-pass filter shown in **Fig 11.86** 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*. The complete original *QST* article for this project is included on the CD-ROM included with this book. The article supplies complete assembly and alignment drawings.

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, W4ENE, which is available from **www.arrl.org/arrl-handbook-reference**. The *Elsie* format data file for this filter, DC54.lct, may be downloaded for your own evaluation from the author's website at **www.realhamradio.com**.

Fig 11.87 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. The parts list for the filter is given in **Table 11.24.**

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.

Fig 11.88 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.

Table 11.24
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.

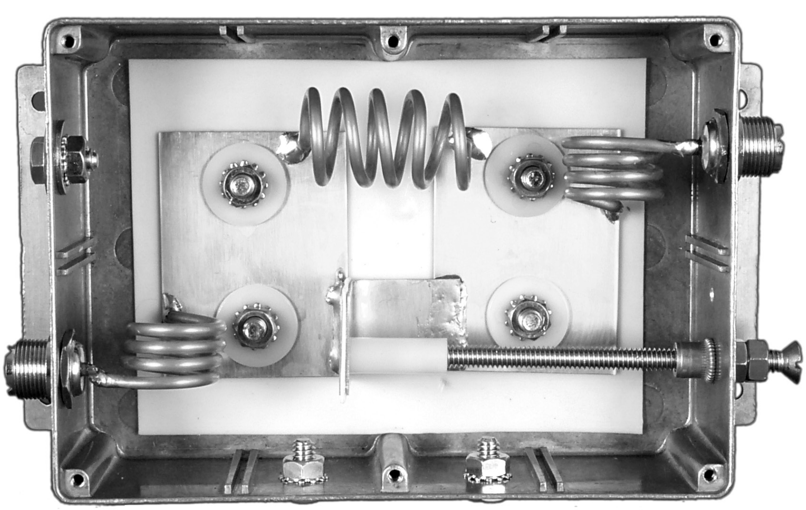


Fig 11.86 — The 1.8 to 54 MHz low-pass filter is housed in a die-cast box.

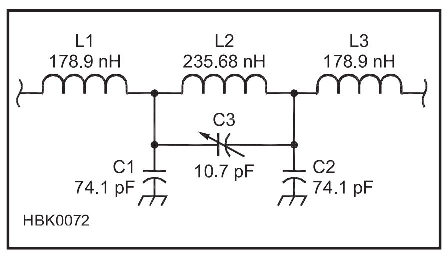


Fig 11.87 — The low-pass filter schematic. See construction details in the article on the CD-ROM.
C1, C2 — 74.1 pF. 2 × 2.65 inch brass plate sandwiched with 0.03125 inch thick Teflon sheet. The metal enclosure is the remaining grounded terminal of this capacitor.
C3 — Homemade variable using brass and Teflon.
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.

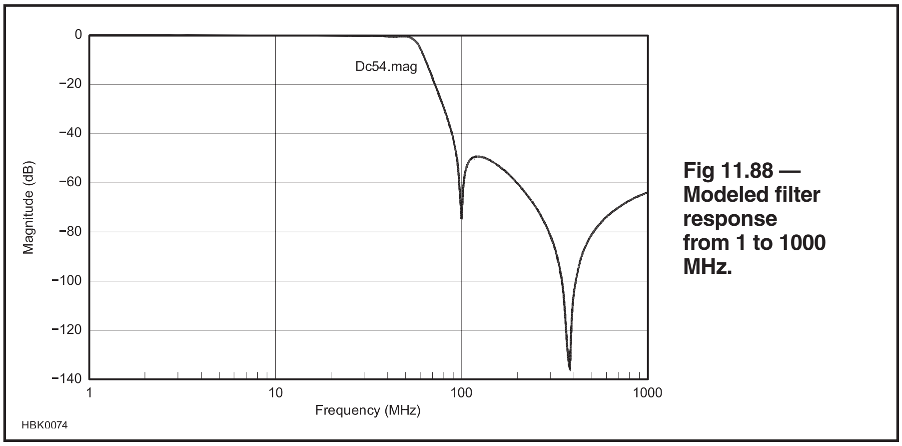


Fig 11.88 — Modeled filter response from 1 to 1000 MHz.

11.11.7 Band-Pass Filter for 145 MHz

The following project is based on a design from the RSGB *Radio Communication Handbook*, *11th edition.* This filter is intended to reduce harmonics and other out-of-band spurious emissions when transmitting and suppress strong out-of-band incoming signals which could overload the receiver. The filter's schematic and response curve are shown in **Fig 11.89.** This filter design is not suitable for use as a repeater duplexer for in-band signals sharing a common antenna — a high-Q cavity resonator is required.

Direct inductive coupling from the proximity of the coils is used between the first two and the last two resonant circuits. C_C performs "top coupling" between the center two sections where a shield prevents stray coupling. The input and output connections are tapped down on their respective coils to transform the 50 Ω source and load into the proper impedances for terminating the filter

At VHF, self-supporting coils and mica, ceramic or air-dielectric trimmer capacitors give adequate results for most applications. Piston ceramic trimmers can be used for receiving or for low-power signals of a few watts. At higher power, use an air-variable capacitor. The filter is assembled in a 4 × 2½ × 1½ inch die-cast aluminum box so there is room for either piston or air-variable capacitors.

The band-pass filter is made from four parallel resonant circuits formed by the inductance of the coils and their associated parasitic inter-turn capacitance described in the **RF Techniques** chapter. A 1-6 pF series capacitor (C in Fig 11.89) connected to ground adjusts the resonant frequency of each coil. A ceramic capacitor should be used for C_C which has a value of 0.5 pF.

The dimensions of the coils are important to control the self-resonant frequency. Each coil is constructed from 6½ turns of #17 AWG solid, bare wire (1.15 mm dia), ⅜ inch (9.5 mm) in diameter. Each turn is spaced 1 wire diameter apart. The original design used British #18 SWG wire which is slightly thicker (1.22 mm) so coils made with #17 AWG wire will have a slightly higher resonant frequency. The taps for the coils are made 1 turn from the grounded end of the coil as shown in the figure.

Adjustment will be required after assembly to tune out the stray capacitances and inductances. A sweep generator and oscilloscope (see the **Test Equipment and Measurements** chapter) provide the most practical adjustment method. A variable oscillator with frequency counter and a voltmeter with RF probe, plus a good deal of patience, can also do the job.

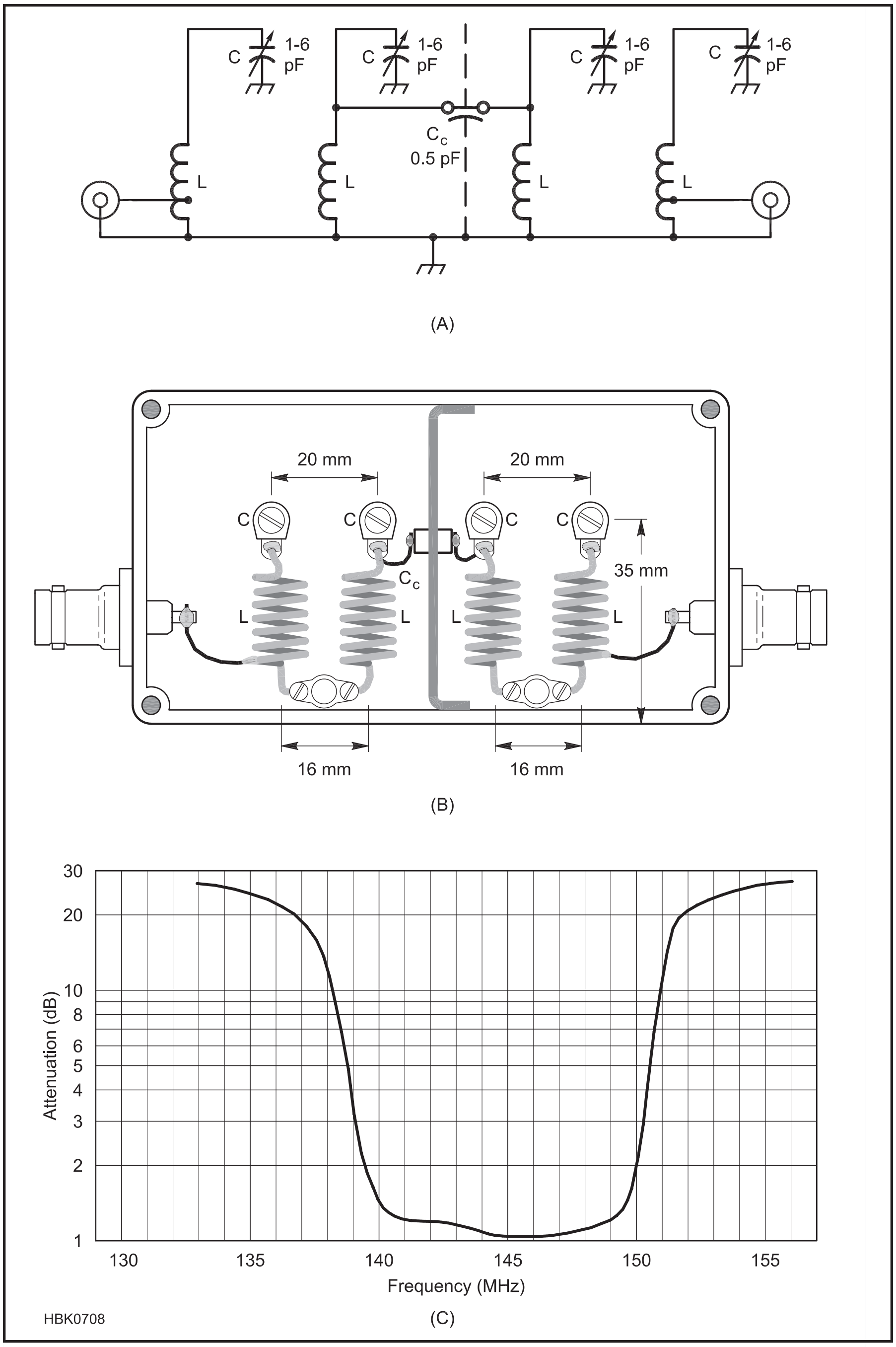


Fig 11.89 — A four-section band-pass filter for 145 MHz for attenuating strong out-of-band signals and reducing harmonics or other out-of-band spurious emissions.

11.12 Filter Glossary

Active filter — A filter that uses active (powered) devices to implement its function.
All-pass — Filter response in which the magnitude response does not change with frequency, but the phase response does change with frequency.
Amplitude response — see **magnitude response**
Band-pass — Filter response in which signals are passed in a range of frequencies and rejected outside that range.
Band-stop — Filter response in which signals are rejected in a range of frequencies and passed outside that range (also called *band-reject* or *notch* filter).
Bandwidth — Range of frequencies over which signals are passed (low-pass, high-pass, band-pass) or rejected (band-stop).
Brick wall response — An ideal filter response in which signals are either passed with no attenuation or attenuated completely.
Cutoff frequency — Frequency at which a filter's output is 3 dB below its passband output (also called *corner frequency* or *3 dB frequency*).
Decade — A ratio of 10 in frequency.
Equiripple — Equalized ripple in a filter's magnitude response across the passband, stop band, or both.
Flat — Refers to a filter's magnitude response that is constant across a range of frequencies.
Group delay — The transit time of signals through a filter.
High-pass — Filter response in which signals above the cutoff frequency are passed and rejected at lower frequencies.
Ideal filter — Filter that passes signals without loss or attenuates them completely. An ideal filter has no transition regions. (See also **brick wall response**).
Insertion loss — The loss incurred by signals in a filter's passband.
Low-pass — Filter response in which signals below the cutoff frequency are passed and rejected at higher frequencies.
Lumped elements — Discrete inductors and capacitors; a lumped-element filter made from discrete inductors and capacitors.
Magnitude response — Graph of a filter's output amplitude versus frequency.
Microstrip — A type of transmission line made from a strip of metal separated from a ground plane by a layer of insulating material, such as on a printed-circuit board..
Normalize — The technique of converting numeric values to their ratio with respect to some reference value. (To denormalize is to reverse the normalization, converting the ratios back to the original values.)
Notch filter — see **band-stop filter**.
Octave — A ratio of two in frequency (see also **decade**).
Overshoot — The condition in which the output of a circuit, in responding to a change in its input, temporarily exceeds the steady-state value that the input should cause.
Overtone — Vibration mode at a higher frequency than the fundamental mode, usually harmonically related.
Passband — The range of frequencies passed by a filter.
Passive filter — A filter that does not require power to perform its function (see also **lumped element**).
Phase response — Graph of the difference in angular units (degrees or radians) between a filter's input and output versus frequency.
Radian — Unit of angular measurement equal to $1/2\pi$ of a circle, equal to $360 / 2\pi$ degrees
Ringing — The condition in which the output of a circuit, in responding to a change in its input, exhibits a damped alternating sequence of exceeding and falling below the steady-state value that the input should cause before settling at the steady-state value.
Ripple — A regular variation with frequency in a filter's magnitude response.
Rolloff — The rate of change in a filter's magnitude response in the transition region and stop band.
Scaling — Changing a filter's impedance or frequency characteristics through multiplication or division by a constant.
Shape factor — The ratio of a filter's bandwidth between the points at which its magnitude response is 6 dB and 60 dB below the response in the filter's passband.
Stop band — The range of frequencies that are rejected by a filter.
Stripline — A transmission line consisting of a metal strip suspended between two ground planes.
TEM — Transverse electromagnetic mode in which the electric and magnetic fields of electromagnetic energy are aligned perpendicularly to the direction of motion.
Topology — The arrangement of connections of components in the filter. For example, "capacitor-input" and "inductor-input" are two different topologies.
Transition region — Range of frequencies between a filter's passband and stop band.

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