Filters and Projects

16

This chapter contains basic design information and examples of the most common filters used by radio amateurs. It was prepared by Reed Fisher, W2CQH, and includes a number of design approaches, tables and filters by Ed Wetherhold, W3NQN, and others. The chapter is divided into two major sections. The first section contains a discussion of filter theory with some design examples. It includes the tools needed to predict the performance of a candidate filter before a design is started or a commercial unit purchased. Extensive references are given for further reading and design information. The second section contains a number of selected practical filter designs for immediate construction.

Basic Concepts

A filter is a network that passes signals of certain frequencies and rejects or attenuates those of other frequencies. The radio art owes its success to effective filtering. Filters allow the radio receiver to provide the listener with only the desired signal and reject all others. Conversely, filters allow the radio transmitter to generate only one signal and attenuate others that might interfere with other spectrum users.

The simplified SSB receiver shown in **Fig 16.1** illustrates the use of several common filters. Three of them are located between the antenna and the speaker. They provide the essential receiver filter functions. A preselector filter is placed between the antenna and the first mixer. It passes all frequencies between 3.8 and 4.0 MHz with low loss. Other frequencies, such as out-of-band signals, are rejected to prevent them from overloading the first mixer (a common problem with short-

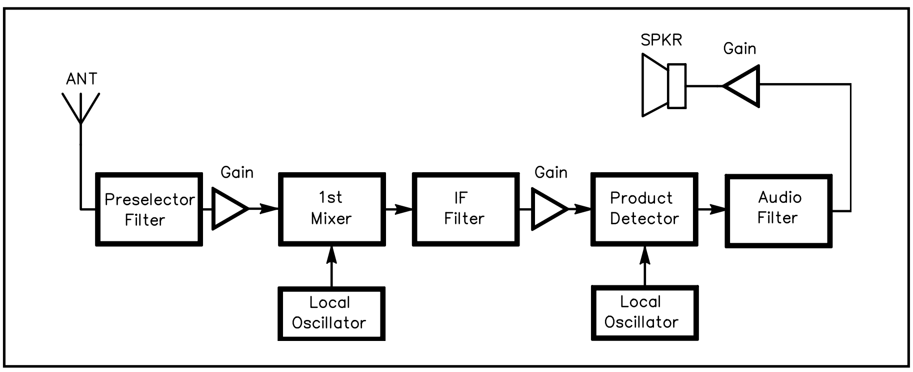


Fig 16.1—One-band SSB receiver. At least three filters are used between the antenna and speaker.

wave broadcast stations). The preselector filter is almost always built with LC filter technology.

An intermediate frequency (IF) filter is placed between the first and second mixers. It is a band-pass filter that passes the desired SSB signal but rejects all others. The age of the receiver probably determines which of several filter technologies is used. As an example, 50 kHz or 455 kHz LC filters and 455 kHz mechanical filters were used through the 1960s. Later model receivers usually use quartz crystal filters with center frequencies between 3 and 9 MHz. In all cases, the filter bandwidth must be less than 3 kHz to effectively reject adjacent SSB stations.

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

The complementary SSB transmitter block diagram is shown in **Fig 16.2**. The same array of filters appear in reverse order.

First is a 300-Hz to 3-kHz audio filter, which rejects out-of-band audio signals such as 60-Hz power supply hum. It is placed between the microphone and the balanced mixer.

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

Finally, a 3.8 to 4.0-MHz band-pass filter is placed between transmit mixer and antenna to reject unwanted frequencies generated by the mixer and prevent them from being amplified and transmitted.

This chapter will discuss the four most common types of filters: low-pass, high-pass, band-pass and band-stop. The idealized characteristics of these filters are shown in their most basic form in **Fig 16.3**.

A low-pass filter permits all frequencies below a specified cutoff frequency to be transmitted with small loss, but will attenuate all frequencies above the cutoff frequency. The "cutoff frequency" is usually specified to be that frequency where the filter loss is 3 dB.

A high-pass filter has a cutoff frequency above which there is small transmission loss, but below which there is considerable attenuation. Its behavior is opposite to that of the low-pass filter.

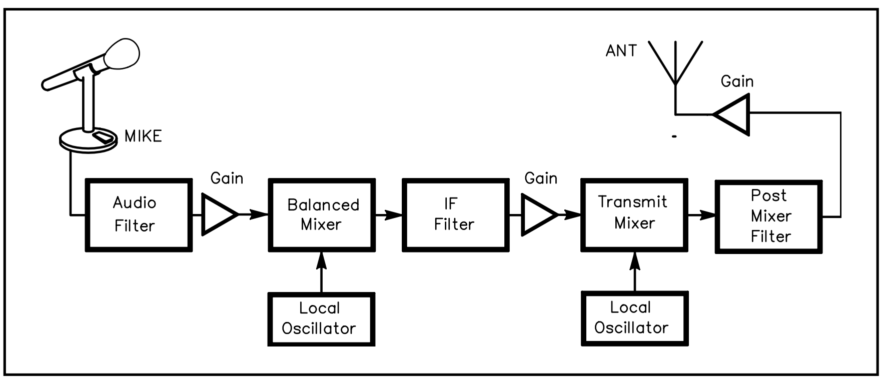


Fig 16.2—One-band transmitter. At least three filters are needed to ensure a clean transmitted signal.

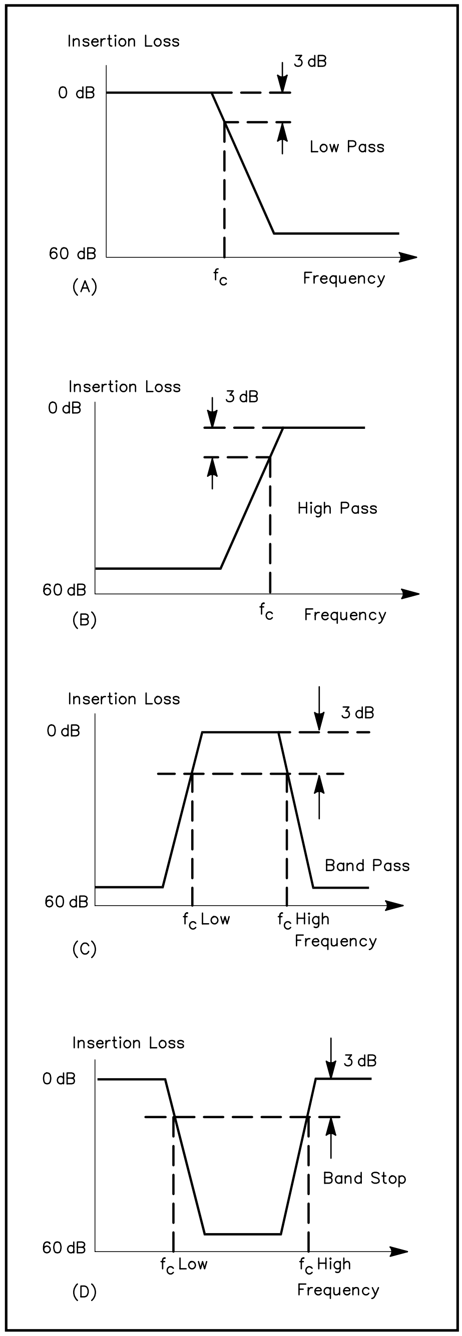


Fig 16.3—Idealized filter responses. Note the definition of f_c is 3 dB down from the break points of the curves.

A band-pass filter passes a selected band of frequencies with low loss, but attenuates frequencies higher and lower than the desired passband. The passband of a filter is the frequency spectrum that is conveyed with small loss. The transfer characteristic is not necessarily perfectly uniform in the passband, but the variations usually are small.

A band-stop filter rejects a selected band of frequencies, but transmits with low loss frequencies higher and lower than the desired stop band. Its behavior is opposite to that of the band-pass filter. The stop band is the frequency spectrum in which attenuation is desired. The attenuation varies in the stop band rising to high values at frequencies far removed from the cutoff frequency.

FILTER FREQUENCY RESPONSE

The purpose of a filter is to pass a desired frequency (or frequency band) and reject all other undesired frequencies. A simple single-stage low-pass filter is shown in **Fig 16.4**. The filter consists of an inductor, L. It is placed between the voltage source e_g and load resistance R_L. Most generators have an associated "internal" resistance, which is labeled R_g.

When the generator is switched on, power will flow from the generator to the load resistance R_L. The purpose of this low-pass filter is to allow maximum power flow at low frequencies (below the cutoff frequency) and minimum power flow at high frequencies. Intuitively, frequency filtering is accomplished because the inductor has reactance that vanishes at dc but becomes large at high frequency. Thus, the current, I, flowing through the load resistance, R_L, will be maximum at dc and less at higher frequencies.

The mathematical analysis of Fig 16.4 is as follows: For simplicity, let $R_g = R_L = R$.

$$i = \frac{e_g}{2R + jX_L} \qquad (1)$$

where

$X_L = 2\pi f L$

f = generator frequency.

Power in the load, P_L, is:

$$P_L = \frac{{e_g}^2 R_L}{4R^2 + {X_L}^2} \qquad (2)$$

Available (maximum) power will be delivered from the generator when:

$X_L = 0$ and $R_g = R_L$

$$P_O = \frac{{E_g}^2}{4R_g} \qquad (3)$$

The filter response is:

$$\frac{P_L}{P_O} = \frac{\text{power in the load}}{\text{available generator power}} \qquad (4)$$

The filter cutoff frequency, called f_c, is the generator frequency where

$$2R = X_L \text{ or } f_c = \frac{R}{\pi L} \qquad (5)$$

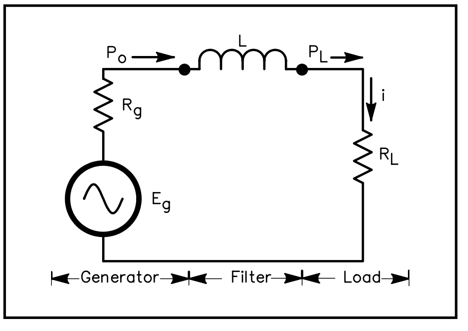


Fig 16.4—A single-stage low-pass filter consists of a series inductor. DC is passed to the load resistor unattenuated. Attenuation increases (and current in the load decreases) as the frequency increases.

As an example, suppose $R_g = R_L = 50\ \Omega$ and the desired cutoff frequency is 4 MHz. Equation 4 states that the cutoff frequency is where the inductive reactance $X_L = 100\ \Omega$. At 4 MHz, using the relationship $X_L = 2\pi f L$, $L = 4\ \mu H$. If this filter is constructed, its response should follow the curve in **Fig 16.5**. Note that the gentle rolloff in response indicates a poor filter. To obtain steeper rolloff a more sophisticated filter, containing more reactances, is necessary. Filters are designed for a specific value of purely resistive load impedance called the *terminating resistance*. When such a resistance is connected to the output terminals of a filter, the impedance looking into the input terminals will equal the load resistance throughout most of the passband. The degree of mismatch across the passband is shown by the SWR scale at the left-hand side of Fig 16.5. If maximum power is to be extracted from the generator driving the filter, the generator resistance must equal the load resistance. This condition is called a "doubly terminated" filter. Most passive filters, including the LC filters described in this chapter, are designed for double termination. If a filter is not properly terminated, its passband response changes.

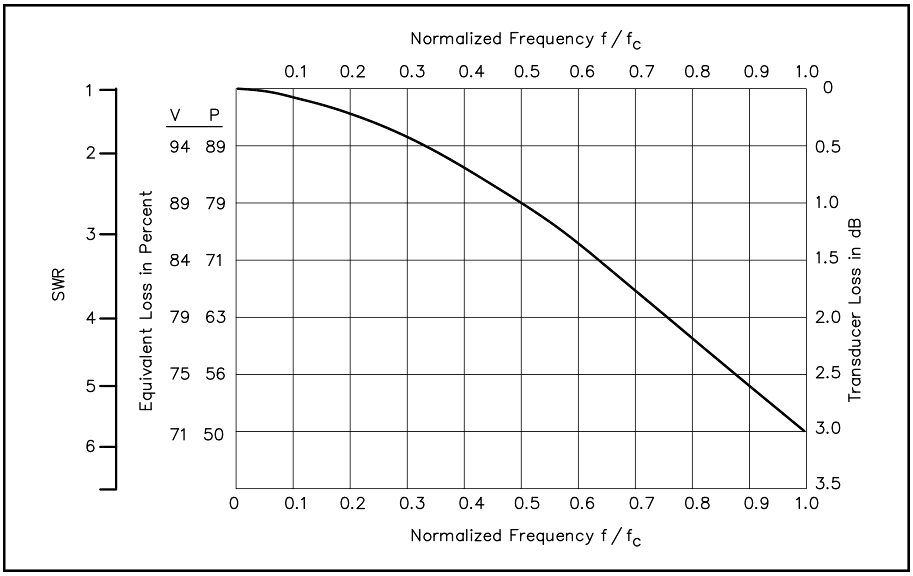


Fig 16.5—Transmission loss of a simple filter plotted against normalized frequency. Note the relationship between loss and SWR.

Certain classes of filters, called "transformer filters" or "matching networks" are specifically designed to work between unequal generator and load resistances. Band-pass filters, described later, are easily designed to work between unequal terminations.

All passive filters exhibit an undesired nonzero loss in the passband due to unavoidable resistances associated with the reactances in the ladder network. All filters exhibit undesired transmission in the stop band due to leakage around the filter network. This phenomenon is called the "ultimate rejection" of the filter. A typical high-quality filter may exhibit an ultimate rejection of 60 dB.

Band-pass filters perform most of the important filtering in a radio receiver and transmitter. There are several measures of their effectiveness or *selectivity*. Selectivity is a qualitative term that arose in the 1930s. It expresses the ability of a filter (or the entire receiver) to reject unwanted adjacent signals. There is no mathematical measure of selectivity.

The term *Q* is quantitative. A band-pass filter's *quality factor* or Q is expressed as Q = (filter center frequency)/(3-dB bandwidth). *Shape factor* is another way some filter vendors specify band-pass filters. The shape factor is a ratio of two filter bandwidths. Generally, it is the ratio (60-dB bandwidth) / (6-dB bandwidth), but some manufacturers use other bandwidths. An ideal or *brick-wall* filter would have a shape factor of 1, but this would require an infinite number of filter elements. The IF filter in a high-quality receiver may have a shape factor of 2.

POLES AND ZEROS

In equation 1 there is a frequency called the "pole" frequency that is given by $f_p = 0$.

In equation 1 there also exists a frequency where the current i becomes zero. This frequency is called the *zero frequency* and is given by: f_0 = infinity. Poles and zeros are intrinsic properties of all networks. The poles and zeros of a network are related to the values of inductances and capacitances in the network.

Poles and zero locations are of interest to the filter theorist because they allow him to predict the frequency response of a proposed filter. For low-pass and high-pass filters the number of poles equals the number of reactances in the filter network. For band-pass and band-stop filters the number of poles specified by the filter vendors is usually taken to be half the number of reactances.

LC FILTERS

Perhaps the most common filter found in the Amateur Radio station is the inductor-capacitor (LC) filter. Historically, the LC filter was the first to be used and the first to be analyzed. Many filter synthesis techniques use the LC filter as the mathematical model.

LC filters are usable from dc to approximately 1 GHz. Parasitic capacitance associated with the inductors and parasitic inductance associated with the capacitors make applications at higher frequencies impractical because the filter performance will change with the physical construction and therefore is not totally predictable from the design equations. Below 50 or 60 Hz, inductance and capacitance values of LC filters become impractically large.

Mathematically, an LC filter is a linear, lumped-element, passive, reciprocal network. Linear means that the ratio of output to input is the same for a 1-V input as for a 10-V input. Thus, the filter can accept an input of many simultaneous sine waves without intermodulation (mixing) between them.

Lumped-element means that the inductors and capacitors are physically much smaller than an operating wavelength. In this case, conductor lengths do not contribute significant inductance or capacitance, and the time that it takes for signals to pass through the filter is insignificant. (Although the different times that it takes for different frequencies to pass through the filter—known as group delay— is still significant for some applications.)

The term *passive* means that the filter does not need any internal power sources. There may be amplifiers before and/or after the filter, but no power is necessary for the filter's equations to hold. The filter alone always exhibits a finite (nonzero) insertion loss due to the unavoidable resistances associated with inductors and (to a lesser extent) capacitors. Active filters, as the name implies, contain internal power sources.

Reciprocal means that the filter can pass power in either direction. Either end of the filter can be used for input or output.

TIME DOMAIN VS FREQUENCY DOMAIN

Humans think in the time domain. Life experiences are measured and recorded in the stream of time. In contrast, Amateur Radio systems and their associated filters are often better understood when viewed in the frequency domain, where frequency is the relevant system parameter. *Frequency* may refer to a sine-wave voltage, current or electromagnetic field. The sine-wave voltage, shown in **Fig 16.6**, is a waveform plotted against time with equation V = A sin(2π f t). The sine wave has a peak amplitude A (measured in volts) and frequency, f (measured in cycles/second or Hertz). A graph showing frequency on the horizontal axis is called a spectrum. A filter response curve is plotted on a spectrum graph.

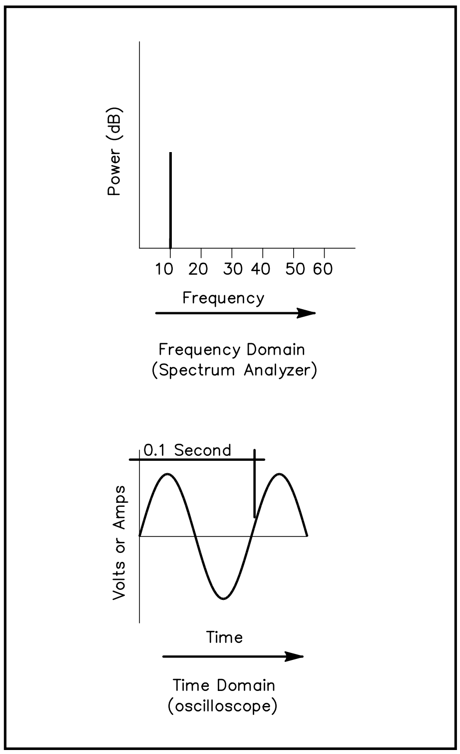


Fig 16.6—Ideal sine-wave voltage. Only one frequency is present.

Historically, radio systems were best analyzed in the frequency domain. The radio transmitters of Hertz (1865) *and* Marconi (1895) consisted of LC resonant circuits excited by high-voltage spark gaps. The transmitters emitted packets of damped sine waves. The low-frequency (200-kHz) antennas used by Marconi were found to possess very narrow bandwidths, and it seemed natural to analyze antenna performance using sine-wave excitation. In addition, the growing use of 50 and 60-Hz alternating current (ac) electric power systems in the 1890s demanded the use of sine-wave mathematics to analyze these systems. Thus engineers trained in ac power theory were available to design and build the early radio systems.

In the frequency domain, the radio world is imagined to be composed of many sine waves of different frequencies flowing endlessly in time. It can be shown by the Fourier transform (Ref 7) that all periodic waveforms can be represented by summing sine waves of different frequencies. For example, the square-wave voltage shown in **Fig 16.7** can be represented by a "fundamental" sine wave of frequency f = 1/t and all its odd harmonics: 3f, 5f, 7f and so on. Thus, in the frequency domain a sine wave is a *narrowband* signal (zero bandwidth) and a square wave is a "wideband" signal.

If the square-wave voltage of Fig 16.7 is passed through a low-pass filter, which removes some of its high-frequency components, the waveform of **Fig 16.8** results. The filtered square wave now has a rise time, which is the time required to rise from 10% to 90% of its peak value (A). The rise time is approximately:

$$\tau_R = \frac{0.35}{f_c} \qquad (6)$$

where f_c is the cutoff frequency of the low-pass filter.

Thus a filter distorts a time-domain signal by removing some of its high-frequency components. Note that a filter cannot distort a sine wave. A filter can only change the amplitude and phase of sine waves. A linear filter will pass multiple sine waves without producing any intermodulation or "beats" between frequencies—this is the definition of *linear*.

The purpose of a radio system is to convey a time-domain signal originating at a source to some distant point with minimum distortion. Filters within the radio system transmitter and receiver may intentionally or unintentionally distort the source signal. A knowledge of the source signal's frequency-domain bandwidth is required so that an appropriate radio system may be designed.

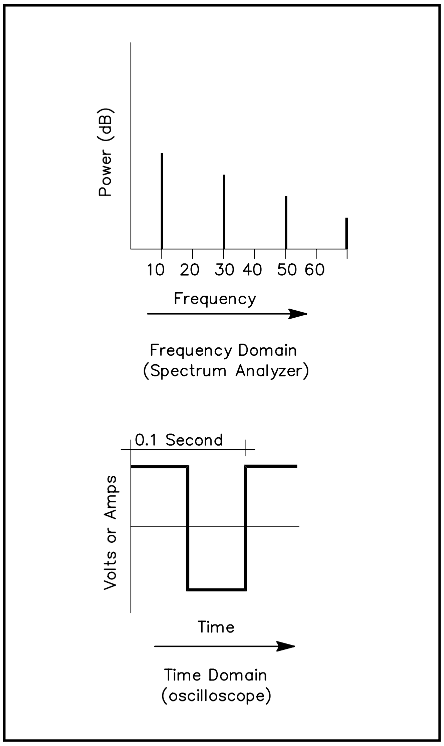


Fig 16.7—Square-wave voltage. Many frequencies are present, including f = 1/t and odd harmonics 3f, 5f, 7f with decreasing amplitudes.

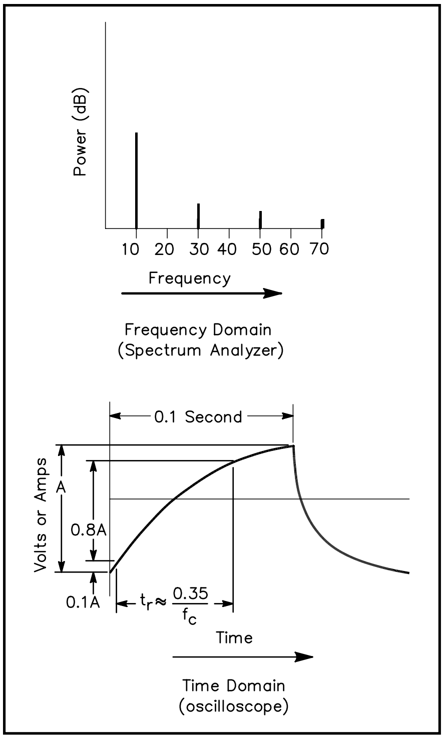


Fig 16.8—Square-wave voltage filtered by a low-pass filter. By passing the square wave through a filter, the higher frequencies are attenuated. The rectangular shape (fast rise and fall items) are *rounded* because the amplitude of the higher harmonics are decreased.

Table 16.1 shows the minimum necessary bandwidth of several common source signals. Note that high-fidelity speech and music requires a bandwidth of 20 Hz to 15 kHz, which is that transmitted by high-quality FM broadcast stations. However, telephone-quality speech requires a bandwidth of only 200 Hz to 3 kHz. Thus, to minimize transmit spectrum, as required by the FCC, filters within amateur transmitters are required to reduce the speech source bandwidth to 200 Hz to 3 kHz at the expense of some speech distortion. After modulation the transmitted RF bandwidth will exceed the filtered source bandwidth if inefficient (AM or FM) modulation methods are employed. Thus the post-modulation *emission bandwidth* may be several times the original filtered source bandwidth. At the receiving end of the radio link, band-pass filters are required to accept only the desired signal and sharply reject noise and adjacent channel interference.

Table 16.1
Typical Filter Bandwidths for Typical Signals.

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

As human beings we are accustomed to operation in the time domain. Just about all of our analog radio connected design occurs in the frequency domain. This is particularly true when it comes to filters. Although the two domains are convertible, one to the other, most filter design is performed in the frequency domain.

Filter Synthesis

The image-parameter method of filter design was initiated by O. Zobel (Ref 1) of Bell Labs in 1923. Image-parameter filters are easy to design and design techniques are found in earlier editions of the ARRL *Handbook*. Unfortunately, image parameter theory demands that the filter terminating impedances vary with frequency in an unusual manner. The later addition of "m-derived matching half sections" at each end of the filter made it possible to use these filters in many applications. In the intervening decades, however, many new methods of filter design have brought both better performance and practical component values for construction.

MODERN FILTER THEORY

The start of modern filter theory is usually credited to S. Butterworth and S. Darlington (Refs 3 and 4). It is based on this approach: Given a desired frequency response, find a circuit that will yield this response.

Filter theorists were aware that certain known mathematical polynomials had "filter like" properties when plotted on a frequency graph. The challenge was to match the filter components (L, C and R) to the known polynomial poles and zeros. This pole/zero matching was a difficult task before the availability of the digital computer. Weinberg (Ref 5) was the first to publish computer generated tables of normalized low-pass filter component values. ("Normalized" means 1-Ω resistor terminations and cutoff frequency $\omega_c = 2\pi f_c = 1$ radian/s.)

An ideal low-pass filter response shows no loss from zero frequency to the cutoff frequency, but infinite loss above the cutoff frequency. Practical filters may approximate this ideal response in several different ways.

Fig 16.9 shows the Butterworth or "maximally-flat" type of approximation. The Butterworth response formula is:

$$\frac{P_L}{P_O} = \frac{1}{1+\left(\frac{\omega}{\omega_c}\right)^{2n}} \qquad (7)$$

where

ω = frequency of interest,
ω_c = cutoff frequency
n = number of poles (reactances)
P_L = power in the load resistor
P_O = available generator power.

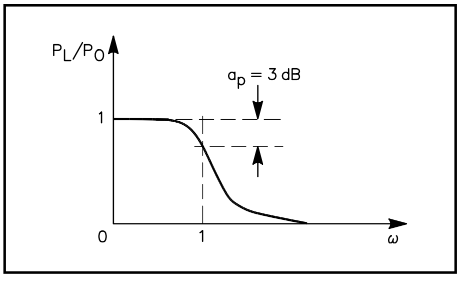


Fig 16.9—Butterworth approximation of an ideal low-pass filter response. The 3-dB attenuation frequency (f_c) is normalized to 1 radian/s.

The passband is exceedingly flat near zero frequency *and* very high attenuation is experienced at high frequencies, but the approximation for both pass and stop bands is relatively poor in the vicinity of cutoff.

Fig 16.10 shows the Chebyshev approximation. Details of the Chebyshev response formula can be found in Refs 5 and 17. Use of this reference as well as similar references for Chebyshev filters requires detailed familiarity with Chebyshev polynomials.

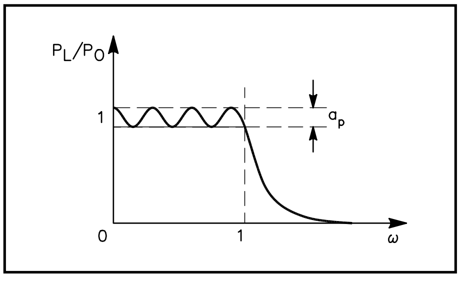


Fig 16.10—Chebyshev approximation of an ideal low-pass filter. Notice the ripple in the passband.

IMPEDANCE AND FREQUENCY SCALING

Fig 16.11A shows normalized component values for Butterworth filters up to ten poles. Fig 16.11B shows the schematic diagrams of the Butterworth low-pass filter. Note that the first

Prototype Butterworth Low-Pass Filters

	C1	*L2*	*C3*	*L4*	*C5*	*L6*	*C7*	*L8*	*C9*	*L10*
	L1	*C2*	*L3*	*C4*	*L5*	*C6*	*L7*	*C8*	*L9*	*C10*
n										
1	2.0000									
2	1.4142	1.4142								
3	1.0000	2.0000	1.0000							
4	0.7654	1.8478	1.8478	0.7654						
5	0.6180	1.6180	2.0000	1.6180	0.6180					
6	0.5176	1.4142	1.9319	1.9319	1.4142	0.5176				
7	0.4450	1.2470	1.8019	2.0000	1.8019	1.2470	0.4450			
8	0.3902	1.1111	1.6629	1.9616	1.9616	1.6629	1.1111	0.3902		
9	0.3473	1.0000	1.5321	1.8794	2.0000	1.8794	1.5321	1.0000	0.3473	
10	0.3129	0.9080	1.4142	1.7820	1.9754	1.9754	1.7820	1.4142	0.9080	0.3129

(A)

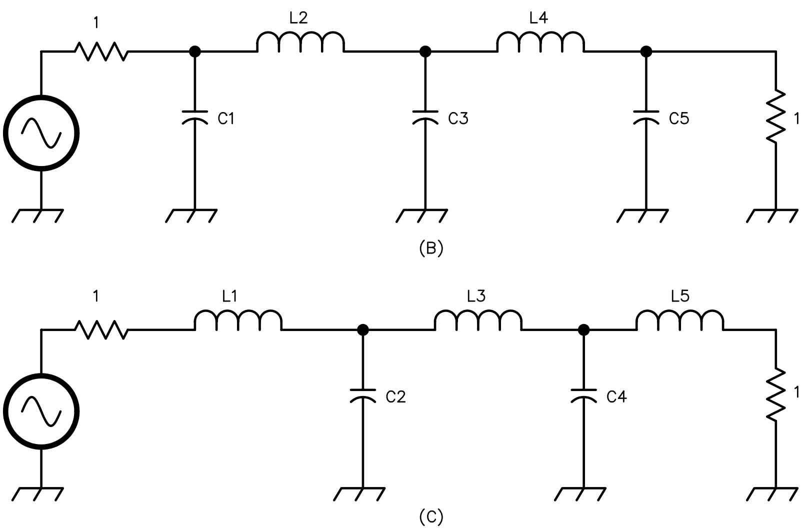


Fig 16.11—Component values for Butterworth low-pass filters. Greater values of n require more stages.

reactance in Fig 16.11B is a shunt capacitor C1, whereas in Fig 16.11C the first reactance is a series inductor L1. Either configuration can be used, but a design using fewer inductors is usually chosen.

In filter design, the use of *normalized* values is common. Normalized generally means a design based on 1-Ω terminations and a cutoff frequency (passband edge) of 1 radian/second. A filter is *denormalized* by applying the following two equations:

$$L' = \left(\frac{R'}{R}\right)\left(\frac{\omega}{\omega'}\right)L \qquad (8)$$

$$C' = \left(\frac{R}{R'}\right)\left(\frac{\omega}{\omega'}\right)C \qquad (9)$$

where

L', C', ω' and R' are the new (desired) values

L and C are the values found in the filter tables

R = 1 Ω

ω = 1 radian/s.

For example, consider the design of a 3-pole Butterworth low-pass filter for a transmitter speech amplifier. Let the desired cutoff frequency be 3000 Hz and the desired termination resistances be 1000 Ω. The normalized prototype, taken from Fig 16.11B is shown in **Fig 16.12**. The new (desired) inductor value is:

$$L' = \left(\frac{1000\ \Omega}{1\Omega}\right)\left(\frac{1\ \text{radian/second}}{2\pi(3000)\ \text{Hz}}\right) 2\ \text{H}$$

or L' = 0.106 H.

The new (desired) capacitor value is:

$$C' = \left(\frac{1\Omega}{1000\ \Omega}\right)\left(\frac{1\ \text{radian/second}}{2\pi(3000)\ \text{Hz}}\right) 1\ \text{F}$$

or C' = 0.053 µF.

The final denormalized filter is shown in **Fig 16.13**. The filter response, in the passband, should obey curve n = 3 in **Fig 16.14**. To use the normalized frequency response curves in Fig 16.14 calculate the frequency ratio f/f_c where f is the desired frequency and f_c is the cutoff frequency. For the filter just designed, the loss at 2000 Hz can be found as follows: When f is 2000 Hz, the frequency ratio is: f/f_c = 2000/3000 = 0.67. Therefore the predicted loss (from the n = 3 curve) is about 0.37 dB.

When f is 4000 Hz, the filter is operating in the stop-band (Fig 16.17). The resulting frequency ratio is: f/f_c = 4000/3000 = 1.3. Therefore the expected loss is about 8 dB. Note that as the number of reactances (poles) increases the filter response approaches the low-pass response of Fig 16.3A.

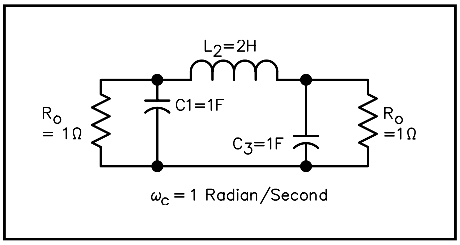


Fig 16.12—A 3-pole Butterworth filter designed for a normalized frequency of 1 radian/s.

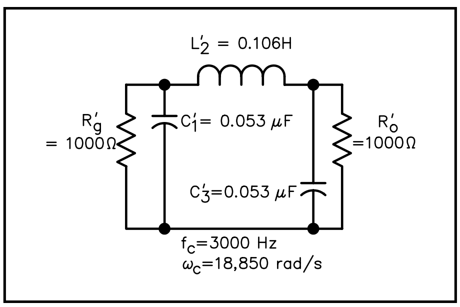


Fig 16.13—A 3-pole Butterworth filter scaled to 3000 Hz.

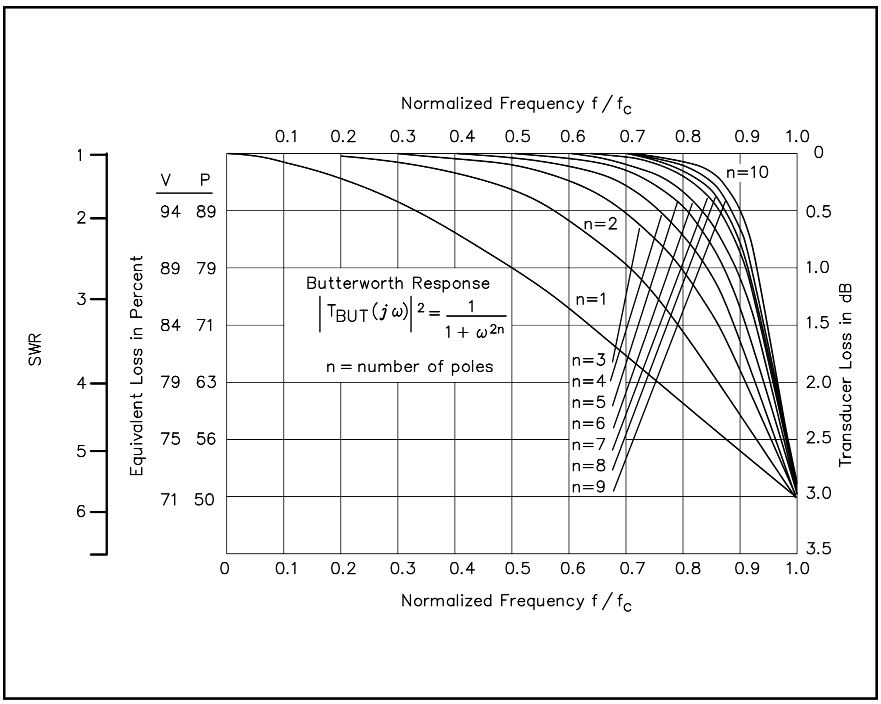


Fig 16.14—Passband loss of Butterworth low-pass filters. The horizontal axis is normalized frequency (see text).

BAND-PASS FILTERS—SIMPLIFIED DESIGN

The design of band-pass filters may be directly obtained from the low-pass prototype by a frequency translation. The low-pass filter has a “center frequency” (in the parlance of band-pass filters) of 0 Hz. *The frequency translation from 0 Hz to the band-pass filter center frequency, f, is obtained by replacing in the low-pass prototype all shunt capacitors with parallel tuned circuits and all series inductors with series tuned circuits.*

As an example, suppose a band-pass filter is required at the front end of a home-brew 40-m QRP receiver to suppress powerful adjacent broadcast stations. The proposed filter has these characteristics:

- Center frequency, f_c = 7.15 MHz
- 3-dB bandwidth = 360 kHz
- terminating resistors = 50 Ω
- 3-pole Butterworth characteristic.

Start the design for the normalized 3-pole Butterworth low-pass filter (shown in Fig 16.11). First determine the center frequency from the band-pass limits. This frequency, f_O, is found by determining the geometric mean of the band limits. In this case the band limits are 7.15 + 0.360/2 = 7.33 MHz and 7.15 –0.360/2 = 6.97 MHz; then

$$f_O = \sqrt{f_{lo} \times f_{hi}} = \sqrt{6.97 \times 7.33} = 7.14\,\text{MHz} \qquad (10)$$

where

f_{lo} = low frequency end of the band-pass (or band-stop)

f_{hi} = high frequency end of the band-pass (or band-stop)

[Note that in this case there is little difference between 7.15 (bandwidth center) and 7.147 (band-edge geometric mean) because the bandwidth is small. For wide-band filters, however, there can be a significant difference.]

Next, denormalize to a new interim low-pass filter having R' = 50 Ω and f' = 0.36 MHz.

$$L' = \left(\frac{50}{1}\right)\left(\frac{1}{2 \times \pi \times 0.36 \times 10^6}\right) 2\,\text{H} = 44.2\,\mu\text{H}$$

$$C' = \left(\frac{1}{50}\right)\left(\frac{1}{2 \times \pi \times 0.36 \times 10^6}\right) 1\,\text{F} = 0.0088\,\mu\text{F}$$

This interim low-pass filter, shown in **Fig 16.15**, has a cutoff frequency f_c = 0.36 MHz and is terminated with 50-Ω resistors. The desired 7.147-MHz band-pass filter is achieved by parallel resonating the shunt capacitors with inductors and series resonating the series inductor with a series capacitor. All resonators must be tuned to the center frequency. Therefore, variable capacitors or inductors are required for the resonant circuits. Based on the L' and C' just calculated the parallel-resonating inductor values are:

$$L1 = L3 = \frac{1}{C'\left(2 \times \pi f_o\right)^2} = 0.056\,\mu\text{H}$$

The series-resonating capacitor value is:

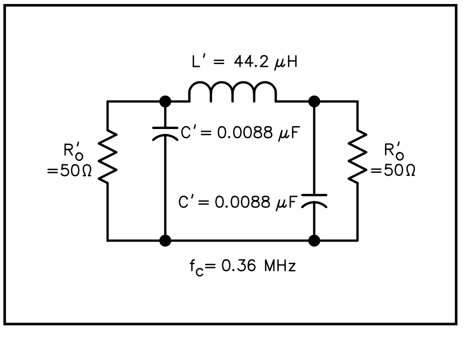


Fig 16.15—Interim 3-pole Butterworth low-pass filter designed for cutoff at 0.36 MHz.

$$C2 = \frac{1}{L'\left(2 \times \pi \times f_O\right)^2} = 11\,pF$$

The final band-pass filter is shown in **Fig 16.16**. The filter should have a 3-dB bandwidth of 0.36 MHz. That is, the 3-dB loss frequencies are 6.97 MHz and 7.33 MHz. The filter's loaded Q is: Q = 7.147/0.36 or approximately 20.

The filter response, in the passband, falls on the "n = 3" curve in **Fig 16.17**. To use the normalized frequency response curves, calculate the frequency ratio f/f_C. For this band-pass case, f is the difference between the desired attenuation frequency and the center frequency, while f_c is the upper 3-dB frequency minus the center frequency. As an example the filter loss at

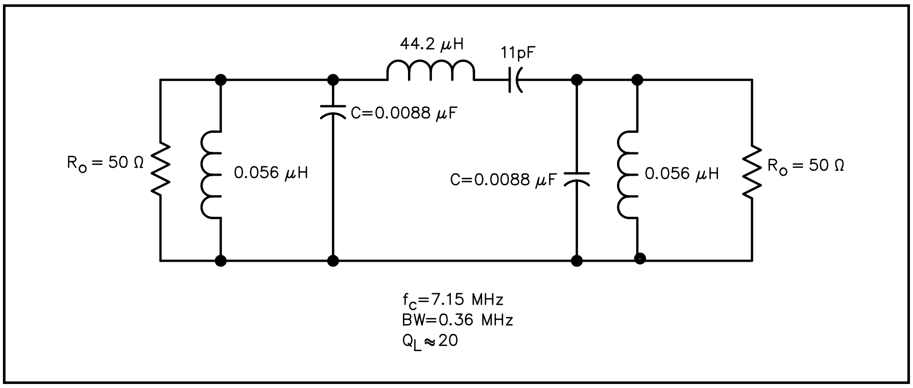


Fig 16.16—Final filter design consists of the low-pass filter scaled to a center frequency of 7.15 MHz.

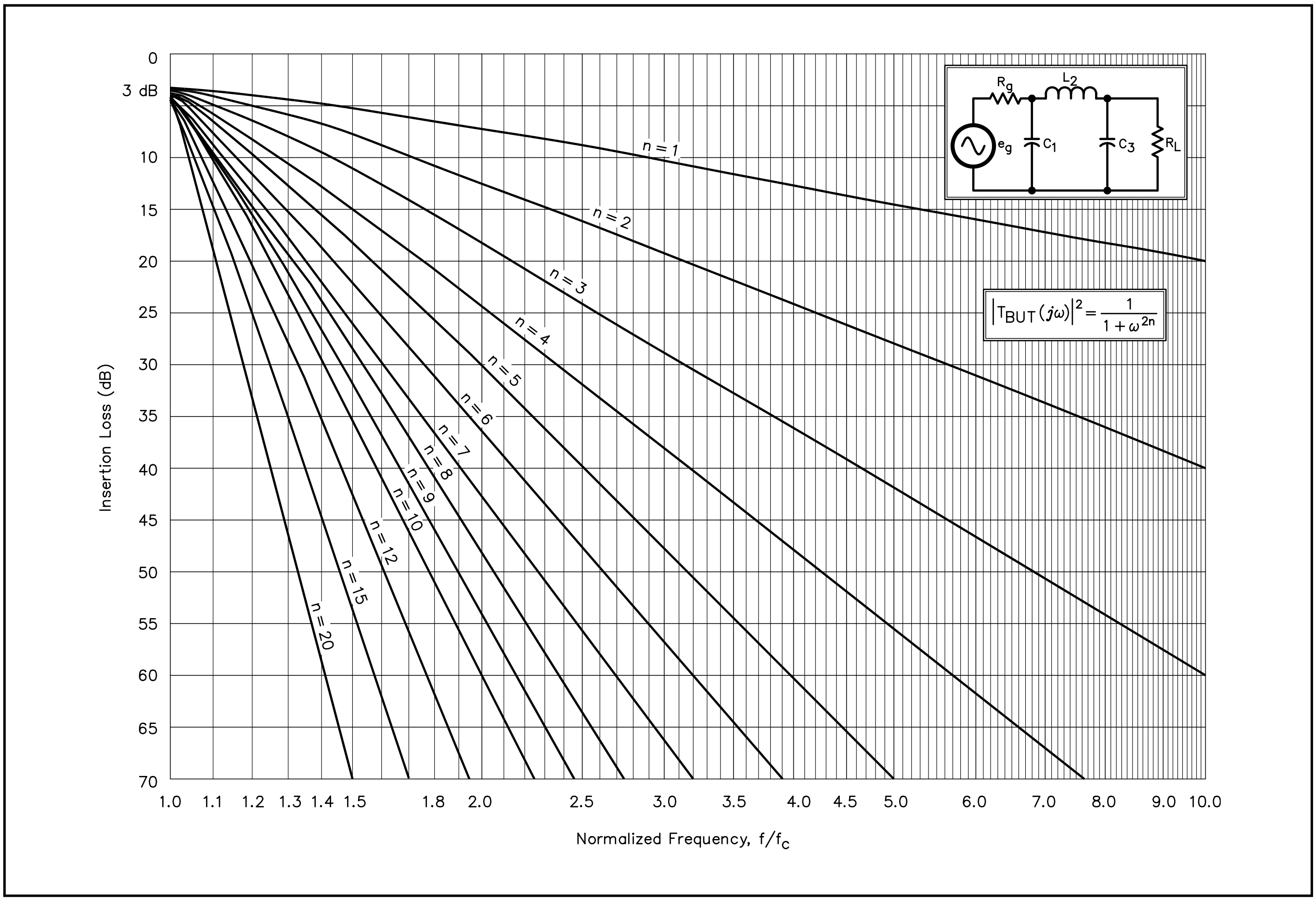


Fig 16.17—Stop-band loss of Butterworth low-pass filters. The almost vertical angle of the lines representing filters with high values of n (10, 12, 15, 20) show the slope of the filter will be very high (sharp cutoff).

7.5 MHz is found by using the normalized frequency ratio given by:

$$\frac{f}{f_c} = \frac{7.5 - 7.147}{7.33 - 7.147} = 1.928$$

Therefore, from Fig 16.17 the expected loss is about 17 dB.

At 6 MHz the loss may be found by:

$$\frac{f}{f_c} = \frac{7.147 - 6}{7.33 - 7.147} = 6.26$$

The expected loss is approximately 47 dB. Unfortunately, awkward component values occur in this type of band-pass filter. The series resonant circuit has a very large LC ratio and the parallel resonant circuits have very small LC ratios. The situation worsens as the filter loaded Q_L ($Q_L = f_0/BW$) increases. Thus, this type of band-pass filter is generally used with a loaded Q less than 10.

Good examples of low-Q band-pass filters of this type are demonstrated by W3NQN's High Performance CW Filter and Passive Audio Filter for SSB in the 1995 and earlier editions of this handbook.

[Note: This analysis used the geometric f_c with the assumption that the filter response is symmetrical about f_c, which it is not. A more rigorous analysis yields 16.9 dB at 7.5 MHz and 50.7 dB at 6 MHz. —Ed.]

Q Restrictions—Band-pass Filters

Most filter component value tables assume lossless reactances. In practice, there are always resistance losses associated with capacitors and inductors (especially inductors). Lossy reactances in low-pass filters modify the response curve. There is finite loss at zero frequency and the cutoff "knee" at f_c will not be as sharp as predicted by theoretical response curves.

The situation worsens with band-pass filters. As loaded Q is increased the midband insertion loss may become intolerable. Therefore, before a band-pass filter design is started, estimate the expected loss.

An approximate estimate of band-pass filter midband response is given by:

$$\frac{P_L}{P_O} = \left(1 - \frac{Q_L}{Q_U}\right)^{2N} \quad (11A)$$

where:

P_L = power delivered to load resistor R_L

P_O = power available from generator:

$$P_O = \frac{{e_g}^2}{4R_L} \quad (11B)$$

Q_U = unloaded Q of inductor:

$$Q_U = \frac{2\pi \times f_0 \times L}{R} \quad (11C)$$

R = inductor series resistance

L = inductance

Q_L = filter loaded Q

$$Q_L = \frac{f_0}{BW_3} \quad (11D)$$

BW_3 = 3 dB bandwidth

N = number of filter stages.

This equation assumes that all losses are in the inductors. For example, the expected loss of the 7.15-MHz filter shown in Fig 16.16 is found by assuming $Q_U = 150$. Q_L is found by equation 11D to be = 7.147/0.36 = 19.8 or approximately 20. Since N = 3 then:

$$\frac{P_L}{P_O} = \left(1 - \frac{20}{150}\right)^6$$

from equation (11A), which equals 0.423. Expressed as dB this is equal to 10 log (0.423) = – 3.73 dB.

Therefore this filter may not be suitable for some applications. If the insertion loss is to be kept small there are severe restrictions on Q_L/Q_U. With typical lumped inductors Q_U seldom exceeds 200. Therefore, LC band-pass filters are usually designed with Q_L not exceeding 20 as shown in **Fig 16.18**.

This loss vs bandwidth trade-off is usually why the final intermediate frequency (IF) in older radio receivers was very low. These units used the equivalent of LC filters in their IF coupling. Generally, for SSB reception the desired receiver bandwidth is about 2.5 kHz. Then 50 kHz was often chosen as the final IF since this implies a loaded Q_L of 20. AM broadcast receivers require a 10-kHz bandwidth and use a 455-kHz IF, which results in $Q_L = 45$. FM broadcast receivers require a 200-kHz bandwidth and use a 10.7-MHz IF and $Q_L = 22$.

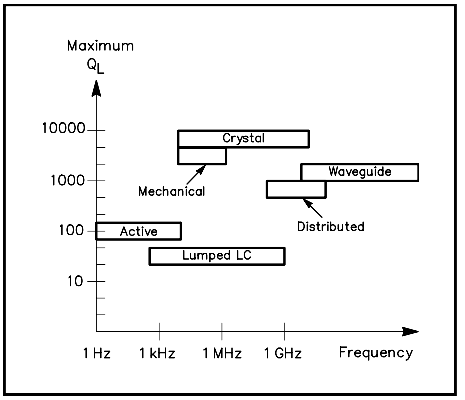


Fig 16.18—Frequency range and maximum loaded Q of band-pass filters. Crystal filters are shown with the highest Q_L and LC filters the lowest.

Filter Design Using Standard Capacitor Values

Practical filters must be designed using commercially available components. Therefore a set of tables, based upon *standard value capacitors* (SVC), has been generated to facilitate this real design process. The procedure presented here uses eight computer-calculated tables of performance parameters and component values for 5- and 7-branch Chebyshev and 5-branch elliptic 50-Ω filters. The tables permit the quick and easy selection of an equally terminated passive LC filter for applications where the attenuation response is of primary interest. All of the capacitors in the Chebyshev designs and the three nonresonating capacitors in the elliptic designs have standard, off-the-shelf values to simplify construction. Although the tables cover only the 1 to 10-MHz frequency range, a simple scaling procedure gives standard-value capacitor (SVC) designs for any impedance level and virtually any cutoff frequency.

The full tables are printed in the **References** chapter of this *Handbook*. Extracts from the tables are reprinted in this section to illustrate the design procedure.

The following text by Ed Wetherhold, W3NQN, is adapted from his paper entitled *Simplified Passive LC Filter Design for the EMC Engineer*. It was presented at an IEEE International Symposium on Electromagnetic Compatibility in 1985.

The approach is based upon the fact that for most nonstringent filtering applications, it is not necessary that the actual cutoff frequency exactly match the desired cutoff frequency. A deviation of 5% or so between the actual and desired cutoff frequencies is acceptable. This permits the use of design tables based on standard capacitor values instead of passband ripple attenuation or reflection coefficient.

STANDARD VALUES IN FILTER DESIGN CALCULATIONS

Capacitors are commercially available in special series of preferred values having designations of E12 (10% tolerance) and E24 (5% tolerance; Ref 22) The reciprocal of the E-number is the power to which 10 is raised to give the step multiplier for that particular series.

First the normalized Chebyshev and elliptic component values are calculated based on many ratios of standard capacitor values. Next, using a 50-Ω impedance level, the parameters of the designs are calculated and tabulated to span the 1-10 MHz decade. Because of the large number of standard-value capacitor (SVC) designs in this decade, the increment in cutoff frequency from one design to the next is sufficiently small so that virtually any cutoff frequency requirement can be satisfied. Using such a table, the selection of an appropriate design consists of merely scanning the cutoff frequency column to find a design having a cutoff frequency that most closely matches the desired cutoff frequency.

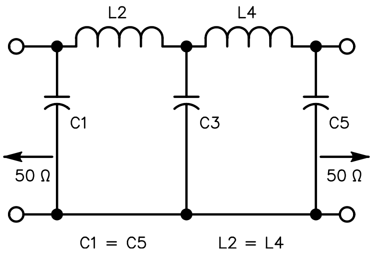
CHEBYSHEV AND ELLIPTIC FILTERS

Low-pass and high-pass 5- and 7-element Chebyshev and 5-branch elliptic designs were selected for tabulation because they are easy to construct and will satisfy the majority of nonstringent filtering requirements where the amplitude response is of primary interest. The precalculated 50-Ω designs are presented in eight tables of five low-pass and three high-pass designs with cutoff frequencies covering the 1-10 MHz decade. The applicable filter configuration and attenuation response curve accompany each table. In addition to the component values, attenuation vs frequency data and SWR are also included in the table. The passband attenuation ripples are so low in amplitude that they are swamped by the filter losses and are not measurable. For this reason, they are not shown in the response curves.

LOW-PASS TABLES

Fig 16.19 is an extract from the tables for the low-pass 5- and 7-element Chebyshev capacitor input/output configuration in the **References** chapter of this *Handbook*. This filter configuration is generally preferred to the alternate inductor input/output configuration because it requires fewer inductors. Generally, decreasing input impedance with increasing frequency in the stop band presents no problems. **Fig 16.20** shows the

The schematic for a 5-element capacitor input/output Chebyshev low-pass filter.

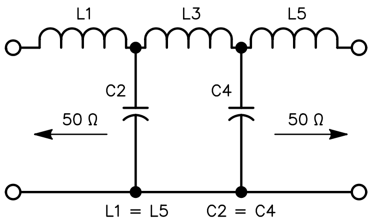


5-Element Chebyshev 50-Ω Lowpass SVC-Filter Designs C-In/Out, E24 Capacitor Values

	— FREQUENCY (MHz) —				MAX	C1,5	L2,4	C3
No.	F_{co}	3 dB	20 dB	40 dB	SWR	(pF)	(µH)	(pF)
1	1.01	1.15	1.53	2.25	1.355	3600	10.8	6200
2	1.02	1.21	1.65	2.45	1.212	3000	10.7	5600
3	1.15	1.29	1.71	2.51	1.391	3300	9.49	5600
4	1.10	1.32	1.81	2.69	1.196	2700	9.88	5100
5	1.25	1.41	1.88	2.75	1.386	3000	8.67	5100
6	1.04	1.37	1.94	2.94	1.085	2200	9.82	4700
7	1.15	1.41	1.95	2.92	1.155	2400	9.37	4700

Fig 16.19—A portion of a 5-element Chebyshev low-pass filter design table for 50-Ω impedance, C-in/out and standard E24 capacitor values. The full table is printed in the References chapter.

Schematic for a 5-element inductor-input/output Chebyshev low-pass filter. See the corresponding table in References chapter for the attenuation response curve.



5-Element Chebyshev 50-Ω Low-pass Filters with L-In/Out and Standard-value L and C

	— FREQUENCY (MHz) —				MAX	L1,5	C2,4	L3
No.	F_{co}	3 dB	20 dB	40 dB	SWR	(µH)	(pF)	(µH)
1	0.744	1.15	1.69	2.60	1.027	5.60	4700	13.7
2	0.901	1.26	1.81	2.76	1.055	5.60	4300	12.7
3	1.06	1.38	1.94	2.93	1.096	5.60	3900	11.8
4	1.19	1.47	2.05	3.07	1.138	5.60	3600	11.2
5	1.32	1.58	2.17	3.23	1.192	5.60	3300	10.6
6	0.911	1.39	2.03	3.12	1.030	4.70	3900	11.4
7	1.08	1.50	2.16	3.29	1.056	4.70	3600	10.6
8	1.25	1.63	2.30	3.48	1.092	4.70	3300	9.92
9	1.42	1.77	2.46	3.68	1.142	4.70	3000	9.32
10	1.61	1.92	2.63	3.90	1.209	4.70	2700	8.79
11	1.05	1.64	2.41	3.72	1.025	3.90	3300	9.63

Fig 16.20—A portion of a 5-element Chebyshev low-pass filter design table for 50-Ω impedance, L-in/out and standard-value L and C. The full table is printed in the References chapter.

corresponding information for low-pass applications, but with an inductor input/output configuration. This configuration is useful when the filter input impedance in the stop band must rise with increasing frequency. For example, some RF transistor amplifiers may become unstable when terminated in a low-pass filter having a stop-band response with a decreasing input impedance. In this case, the inductor-input configuration may eliminate the instability. (Ref 23) Because only one capacitor value is required in the designs of Fig 16.20, it was feasible to have the inductor value of L1 and L5 also be a standard value. **Fig 16.21** is extracted from the table for the low-pass 5-branch elliptic filter with the capacitor input/output configuration, in which the nonresonating capacitors (C1, C3 and C5) are standard values. The alternate inductor input/output elliptic configuration is seldom used and therefore it is not included.

HIGH-PASS TABLES

A high-pass 5-element Chebyshev capacitor input/output configuration is shown in the table extract of **Fig 16.22**. Because the inductor input/output configuration is seldom used, it was not included. High-pass tables for elliptical filters appear in the **References** chapter.

	(MHz)					(pF)					(µH)		(MHz)	
No.	F_{co}	F-3dB	F-A_s	A_s (dB)	MAX SWR	C1	C3	C5	C2	C4	L2	L4	F2	F4
1	0.795	0.989	1.57	47.4	1.092	2700	5600	2200	324	937	12.1	10.1	2.54	1.64
2	1.06	1.20	1.77	46.2	1.234	2700	4700	2200	341	982	9.36	7.56	2.82	1.85
3	1.47	1.57	2.15	45.4	1.586	2700	3900	2200	364	1045	6.32	4.88	3.32	2.23
4	0.929	1.18	1.91	48.0	1.077	2200	4700	1800	257	743	10.2	8.59	3.11	1.99
5	1.27	1.45	2.17	46.7	1.215	2200	3900	1800	271	779	7.85	6.39	3.45	2.26
6	1.69	1.82	2.54	45.9	1.489	2200	3300	1800	287	821	5.64	4.42	3.96	2.64
7	1.12	1.44	2.41	49.8	1.071	1800	3900	1500	192	549	8.45	7.25	3.95	2.52
8	1.49	1.73	2.70	48.8	1.183	1800	3300	1500	200	570	6.75	5.62	4.33	2.81
9	2.11	2.27	3.27	47.8	1.506	1800	2700	1500	213	604	4.55	3.64	5.12	3.40
10	1.28	1.66	2.63	46.3	1.064	1500	3300	1200	192	561	7.20	6.00	4.28	2.74
11	1.79	2.06	2.99	44.8	1.195	1500	2700	1200	204	592	5.52	4.42	4.75	3.11
12	2.52	2.70	3.63	43.8	1.525	1500	2200	1200	220	636	3.71	2.82	5.58	3.76

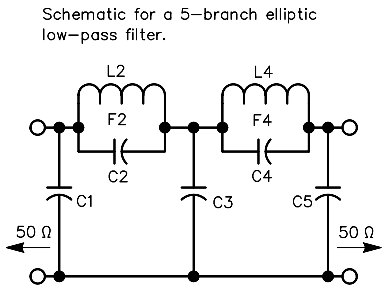
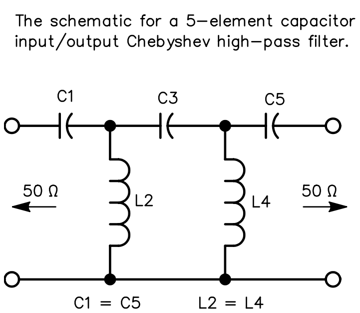


Fig 16.21—A portion of a 5-element elliptic low-pass filter design table for 50-Ω impedance, Standard E12 capacitor values for C1, C3 and C5. The full table is printed in the References chapter.



	FREQUENCY (MHz)							
No.	F_{co}	3 dB	20 dB	40 dB	MAX SWR	C1,5 (pF)	L2,4 (µH)	C3 (pF)
1	1.04	0.726	0.501	0.328	1.044	5100	6.45	2200
2	1.04	0.788	0.554	0.366	1.081	4300	5.97	2000
3	1.17	0.800	0.550	0.359	1.039	4700	5.85	2000
4	1.07	0.857	0.615	0.410	1.135	3600	5.56	1800
5	1.17	0.877	0.616	0.406	1.076	3900	5.36	1800
6	1.33	0.890	0.609	0.397	1.034	4300	5.26	1800
7	1.12	0.938	0.686	0.461	1.206	3000	5.20	1600
8	1.25	0.974	0.693	0.461	1.109	3300	4.86	1600
9	1.38	0.994	0.691	0.454	1.057	3600	4.71	1600
10	1.54	1.00	0.683	0.444	1.028	3900	4.67	1600

Fig 16.22—A portion of a 5-element Chebyshev high-pass filter design table for 50-Ω impedance, C-in/out and standard E24 capacitor values. The full table is printed in the References chapter.

SCALING TO OTHER FREQUENCIES AND IMPEDANCES

The tables shown are for the 1-10 MHz decade and for a 50-Ω equally terminated impedance. The designs are easily scaled to other frequency decades and to other equally terminated impedance levels, however, making the tables a universal design aid for these specific filter types.

Frequency Scaling

To scale the frequency and the component values to the 10-100 or 100- 1000 MHz decades, multiply all tabulated frequencies by 10 or 100, respectively. Then divide all C and L values by the same number. The A_S and SWR data remain unchanged. To scale the filter tables to the 0.1-1 kHz, 1-10 kHz or the

10-100 kHz decades, divide the tabulated frequencies by 1000, 100 or 10, respectively. Next multiply the component values by the same number. By changing the "MHz" frequency headings to "kHz" and the "pF" and "µH" headings to "nF" and "mH," the tables are easily changed from the 1-10 MHz decade to the 1-10 kHz decade and the table values read directly. Because the impedance level is still at 50 Ω, the component values may be awkward, but this can be corrected by increasing the impedance level by ten times using the impedance scaling procedure described below.

Impedance Scaling

All the tabulated designs are easily scaled to impedance levels other than 50 Ω, while keeping the convenience of standard-value capacitors and the "scan mode" of design selection. If the desired new impedance level differs from 50 Ω by a factor of 0.1, 10 or 100, the 50-Ω designs are scaled by shifting the decimal points of the component values. The other data remain unchanged. For example, if the impedance level is increased by ten or one hundred times (to 500 or 5000 Ω), the decimal point of the capacitor is shifted to the left one or two places and the decimal point of the inductor is shifted to the right one or two places. With increasing impedance the capacitor values become smaller and the inductor values become larger. The opposite is true if the impedance decreases.

When the desired impedance level differs from the standard 50-Ω value by a factor such as 1.2, 1.5 or 1.86, the following scaling procedure is used:

1. Calculate the impedance scaling ratio:

$$R = \frac{Z_x}{50} \tag{12}$$

where Z_x is the desired new impedance level, in ohms.

2. Calculate the cutoff frequency (f_{50co}) of a "trial" 50-Ω filter,

$$f_{50co} = R \times f_{xco} \tag{13}$$

where R is the impedance scaling ratio and f_{xco} is the desired cutoff frequency of the filter at the new impedance level.

3. From the appropriate SVC table select a design having its cutoff frequency closest to the calculated f_{50co} value. The tabulated capacitor values of this design are taken directly, but the frequency and inductor values must be scaled to the new impedance level.

4. Calculate the exact f_{xco} values, where

$$f_{xco} = \frac{f'_{50co}}{R} \tag{14}$$

and f'_{50co} is the tabulated cutoff frequency of the selected design. Calculate the other frequencies of the design in the same way.

5. Calculate the inductor values for the new filter by multiplying the tabulated inductor values of the selected design by the square of the scaling ratio, R.

For example, assume a 600-Ω elliptic low-pass filter is desired with a cutoff frequency of 1.0 kHz. The elliptic low-pass table is frequency scaled to the 1-10 kHz decade by changing the table headings to kHz, nF and mH. A suitable design is then selected for scaling to 60 Ω. The 60-Ω design is then scaled to 600 Ω by shifting the decimal point to complete the scaling procedure. The calculations for this example follow, using the five steps outlined above and using the table extract in Fig 16.21:

1. $R = \frac{Z_x}{50} = \frac{60}{50} = 1.2$

2. $f_{50co} = 1.2 \times 1.0$ kHz = 1.2 kHz

3. From the elliptic low-pass table (Fig 16.21), designs 5 and 10 have cutoff frequencies closest to the F_{50co} of 1.2 kHz. Either design is suitable and design 5 is chosen because of its better selectivity. The tabulated capacitor values of 2200 nF, 3900 nF, 1800 nF, 271 nF and 779 nF are copied directly.

4. All frequencies of the final design are calculated by dividing the tabulated frequencies (in kHz) of design 5 by the impedance scaling ratio, 1.2:

$$f_{co} = \frac{1.27}{1.2} = 1.06$$

$$f_{3dB} = \frac{1.45}{1.2} = 1.21$$

$$f_{AS} = \frac{2.17}{1.2} = 1.88$$

Note that a cutoff frequency of 1.0 kHz was desired, but a 1.06-kHz cutoff frequency will be accepted in exchange for the convenience of using an SVC design.

5. The L2 and L4 inductor values of design 5 are scaled to 60 Ω by multiplying them by the square of the impedance ratio, where R = 1.2 and R^2 = 1.44:

$$L2 = 1.44 \times 7.85 \text{ mH} = 11.3 \text{ mH}$$

$$L4 = 1.44 \times 6.39 \text{ mH} = 9.20 \text{ mH}$$

The 60-Ω design is now impedance scaled to 600 Ω by shifting the decimal points of the capacitor values to the left and the decimal points of the inductor values to the right. The final scaled component values for the 600-Ω filter are:

C1 = 0.22 µF
C3 = 0.39 µF
C5 = 0.18 µF
C2 = 27.1 nF
C4 = 77.9 nF
L2 = 113 mH
L4 = 92.0 mH

How to Use the Filter Tables

1. 50-Ω impedance level: Before selecting a filter design, the important parameters of the filter must be known, such as type (low-pass or high-pass), cutoff frequency, impedance level, preferred input element (for low-pass only) *and* an approximation of the required stop band attenuation. It is obvious which tables to use for low-pass or high-pass applications, but it is not so obvious which one design of the many possible choices is optimum for the intended application.

Generally, the Chebyshev will be preferred over the elliptic because the Chebyshev does not require tuning of the inductors. If the relatively gradual attenuation rise of the Chebyshev is not satisfactory, however, then the elliptic should be considered. For audio filtering, the elliptic designs with high values of SWR are preferred because these designs have a much more abrupt attenuation rise than the Chebyshev. For RF applications, SWR values less than 1.2 are recommended to minimize undesired reflections. Low SWR is also important when cascading high-pass and low-pass filters to achieve a band-pass response more than two octaves wide. Each filter will operate as expected if it is correctly terminated, but this will occur only if both designs have the relatively constant terminal impedance that is associated with low SWR.

Once you know the filter type and response needed, select the table of designs most appropriate for the application on a trial basis. From the chosen table, scale the 1-10 MHz data to the desired frequency decade and search the cutoff frequency column for a value nearest the desired cutoff frequency. After finding a possible design, check the stop-band attenuation levels to see if they are satisfactory. Then check the SWR to see if it is appropriate for the application. Finally, check the component values to see if they are convenient. For example, in the audio-frequency range, the capacitor values probably will be in the microfarad range and capacitors in this size are available only in the E12 series of standard values. Then connect the components in accordance with the diagram shown in the table from which the design was selected.

2. Impedance levels other than 50 Ω: First calculate a “trial” filter design using the impedance scaling procedure previously explained. Then search the appropriate table for the best match to the trial filter and scale the selected design to the desired impedance level. In this way, the convenient scan mode of filter selection is used regardless of the desired impedance level.

Chebyshev Filter Design (Normalized Tables)

The figures and tables in this section provide the tools needed to design Chebyshev filters including those filters for which the previously published *standard value capacitor* (SVC) designs might not be suitable. **Table 16.2** lists normalized low-pass designs that, in addition to low-pass filters, can also be used to calculate high-pass, band-pass and band-stop filters in either the inductor or capacitor input/output configurations for equal impedance terminations. **Table 16.3** provides the attenuation for the resultant filter.

This material was prepared by Ed Wetherhold, W3NQN, who has been the author of a number of

Table 16.2

Element values of Chebyshev low-pass filters normalized for a ripple cutoff frequency (Fa_p) of one radian/sec ($^1/_{2\pi}$ Hz) and 1-Ω terminations.

Use the top column headings for the low-pass C-in/out configuration and the bottom column headings for the low-pass L-in/out configuration. Fig 16.23 shows the filter schematics.

N	*RC (%)*	*Ret Loss (dB)*	*F3/F_{ap} Ratio*	*C1 (F)*	*L2 (H)*	*C3 (F)*	*L4 (H)*	*C5 (F)*	*L6 (H)*	*C7 (F)*	*L8 (H)*	*C9 (F)*
3	1.000	40.00	3.0094	0.3524	0.6447	0.3524						
3	1.517	36.38	2.6429	0.4088	0.7265	0.4088						
3	4.796	26.38	1.8772	0.6292	0.9703	0.6292						
3	10.000	20.00	1.5385	0.8535	1.104	0.8535						
3	15.087	16.43	1.3890	1.032	1.147	1.032						
5	0.044	67.11	2.7859	0.2377	0.5920	0.7131	0.5920	0.2377				
5	0.498	46.06	1.8093	0.4099	0.9315	1.093	0.9315	0.4099				
5	1.000	40.00	1.6160	0.4869	1.050	1.226	1.050	0.4869				
5	1.517	36.38	1.5156	0.5427	1.122	1.310	1.122	0.5427				
5	2.768	31.16	1.3892	0.6408	1.223	1.442	1.223	0.6408				
5	4.796	26.38	1.2912	0.7563	1.305	1.577	1.305	0.7563				
5	6.302	24.01	1.2483	0.8266	1.337	1.653	1.337	0.8266				
5	10.000	20.00	1.1840	0.9732	1.372	1.803	1.372	0.9732				
5	15.087	16.43	1.1347	1.147	1.371	1.975	1.371	1.147				
7	1.000	40.00	1.3004	0.5355	1.179	1.464	1.500	1.464	1.179	0.5355		
7	1.427	36.91	1.2598	0.5808	1.232	1.522	1.540	1.522	1.232	0.5808		
7	1.517	36.38	1.2532	0.5893	1.241	1.532	1.547	1.532	1.241	0.5893		
7	3.122	30.11	1.1818	0.7066	1.343	1.660	1.611	1.660	1.343	0.7066		
7	4.712	26.54	1.1467	0.7928	1.391	1.744	1.633	1.744	1.391	0.7928		
7	4.796	26.38	1.1453	0.7970	1.392	1.748	1.633	1.748	1.392	0.7970		
7	8.101	21.83	1.1064	0.9390	1.431	1.878	1.633	1.878	1.431	0.9390		
7	10.000	20.00	1.0925	1.010	1.437	1.941	1.622	1.941	1.437	1.010		
7	10.650	19.45	1.0885	1.033	1.437	1.962	1.617	1.962	1.437	1.033		
7	15.087	16.43	1.0680	1.181	1.423	2.097	1.573	2.097	1.423	1.181		
9	1.000	40.00	1.1783	0.5573	1.233	1.550	1.632	1.696	1.632	1.550	1.233	0.5573
9	1.517	36.38	1.1507	0.6100	1.291	1.610	1.665	1.745	1.665	1.610	1.291	0.6100
9	2.241	32.99	1.1271	0.6679	1.342	1.670	1.690	1.793	1.690	1.670	1.342	0.6679
9	2.512	32.00	1.1206	0.6867	1.357	1.688	1.696	1.808	1.696	1.688	1.357	0.6867
9	4.378	27.17	1.0915	0.7939	1.419	1.786	1.712	1.890	1.712	1.786	1.419	0.7939
9	4.796	26.38	1.0871	0.8145	1.427	1.804	1.713	1.906	1.713	1.804	1.427	0.8145
9	4.994	26.03	1.0852	0.8239	1.431	1.813	1.712	1.913	1.712	1.813	1.431	0.8239
9	8.445	21.47	1.0623	0.9682	1.460	1.936	1.692	2.022	1.692	1.936	1.460	0.9682
9	10.000	20.00	1.0556	1.025	1.462	1.985	1.677	2.066	1.677	1.985	1.462	1.025
9	15.087	16.43	1.0410	1.196	1.443	2.135	1.617	2.205	1.617	2.135	1.443	1.196
N	*RC (%)*	*Ret Loss (dB)*	*F3/F_{ap} Ratio*	*L1 (H)*	*C2 (F)*	*L3 (H)*	*C4 (F)*	*L5 (H)*	*C6 (F)*	*L7 (H)*	*C8 (F)*	*L9 (H)*

Table 16.3

Normalized Frequencies at Listed Attenuation Levels for Chebyshev Low-Pass Filters with N = 3, 5, 7 and 9.

		Attenuation Levels (dB)										
N	*RC(%)*	*1.0*	*3.01*	*6.0*	*10*	*20*	*30*	*40*	*50*	*60*	*70*	*80*
3	1.000	2.44	3.01	3.58	4.28	6.33	9.27	13.59	19.93	29.25	42.92	63.00
3	1.517	2.15	2.64	3.13	3.74	5.52	8.08	11.83	17.35	25.46	37.36	54.83
3	4.796	1.56	1.88	2.20	2.60	3.79	5.53	8.08	11.83	17.35	25.45	37.35
3	10.000	1.31	1.54	1.78	2.08	3.00	4.34	6.33	9.26	13.57	19.90	29.20
3	15.087	1.20	1.39	1.59	1.85	2.63	3.79	5.52	8.06	11.81	17.32	25.41
5	1.000	1.46	1.62	1.76	1.94	2.39	2.97	3.69	4.62	5.79	7.27	9.13
5	1.517	1.38	1.52	1.65	1.80	2.22	2.74	3.41	4.26	5.33	6.69	8.40
5	4.796	1.19	1.29	1.39	1.50	1.82	2.22	2.74	3.41	4.26	5.33	6.69
5	6.302	1.16	1.25	1.34	1.44	1.74	2.12	2.61	3.24	4.04	5.05	6.34
5	10.000	1.11	1.18	1.26	1.35	1.61	1.95	2.39	2.96	3.69	4.61	5.78
5	15.087	1.07	1.13	1.20	1.28	1.51	1.82	2.22	2.74	3.41	4.25	5.33
7	1.000	1.23	1.30	1.37	1.45	1.65	1.89	2.18	2.53	2.95	3.44	4.04
7	1.517	1.19	1.25	1.32	1.39	1.57	1.80	2.07	2.39	2.79	3.25	3.81
7	4.796	1.10	1.15	1.19	1.25	1.39	1.57	1.80	2.07	2.39	2.79	3.25
7	8.101	1.07	1.11	1.15	1.19	1.32	1.49	1.69	1.94	2.24	2.60	3.03
7	10.000	1.05	1.09	1.13	1.18	1.30	1.45	1.65	1.89	2.18	2.53	2.94
7	15.087	1.04	1.07	1.10	1.14	1.25	1.39	1.57	1.80	2.07	2.39	2.78
9	1.000	1.13	1.18	1.22	1.26	1.38	1.51	1.67	1.85	2.07	2.32	2.61
9	1.517	1.11	1.15	1.19	1.23	1.34	1.46	1.61	1.78	1.99	2.22	2.50
9	4.796	1.06	1.09	1.11	1.15	1.23	1.34	1.46	1.61	1.78	1.99	2.22
9	8.445	1.04	1.06	1.09	1.11	1.19	1.28	1.40	1.53	1.69	1.88	2.10
9	10.000	1.03	1.06	1.08	1.10	1.18	1.27	1.38	1.51	1.67	1.85	2.07
9	15.087	1.02	1.04	1.06	1.08	1.15	1.23	1.34	1.46	1.61	1.78	1.99

articles and papers on the design of LC filters. It is a complete revision of his previously published filter design material and provides both insight to the design and actual designs in just a few minutes.

For a given number of elements (N), increasing the filter reflection coefficient (RC or ρ) causes the attenuation slope to increase with a corresponding increase in both the passband ripple amplitude (a_p) and SWR and with a decrease in the filter return loss. All of these parameters are mathematically related to each other. If one is known, the others may be calculated. Filter designs having a low RC are preferred because they are less sensitive to component and termination impedance variations than are designs having a higher RC. The RC percentage is used as the independent variable in Table 16.2 because it is used as the defining parameter in the more frequently used tables, such as those by Zverev and Saal (see Refs 17 and 18).

The return loss is tabulated instead of passband ripple amplitude (a_p) because it is easy to measure using a return loss bridge. In comparison, ripple amplitudes less than 0.1 dB are difficult to measure accurately. The resulting values of attenuation are contained in Table 16.3 and corresponding values of a_p and SWR may be found by referring to the Equivalent Values of Reflection Coefficient, Attenuation, SWR and Return Loss table in the **References** chapter. The filter used (low pass, high-pass, band-pass and so on) will depend on the application and the stop-band attenuation needed.

The filter schematic diagrams shown in **Fig 16.23** are for low-pass and high-pass versions of the Chebyshev designs listed in Table 16.2. Both low-pass and high-pass equally terminated configurations and component values of the C-in/out or L-in/out filters can be derived from this single table. By using a simple procedure, the low-pass and high-pass designs can be transformed into corresponding band-pass

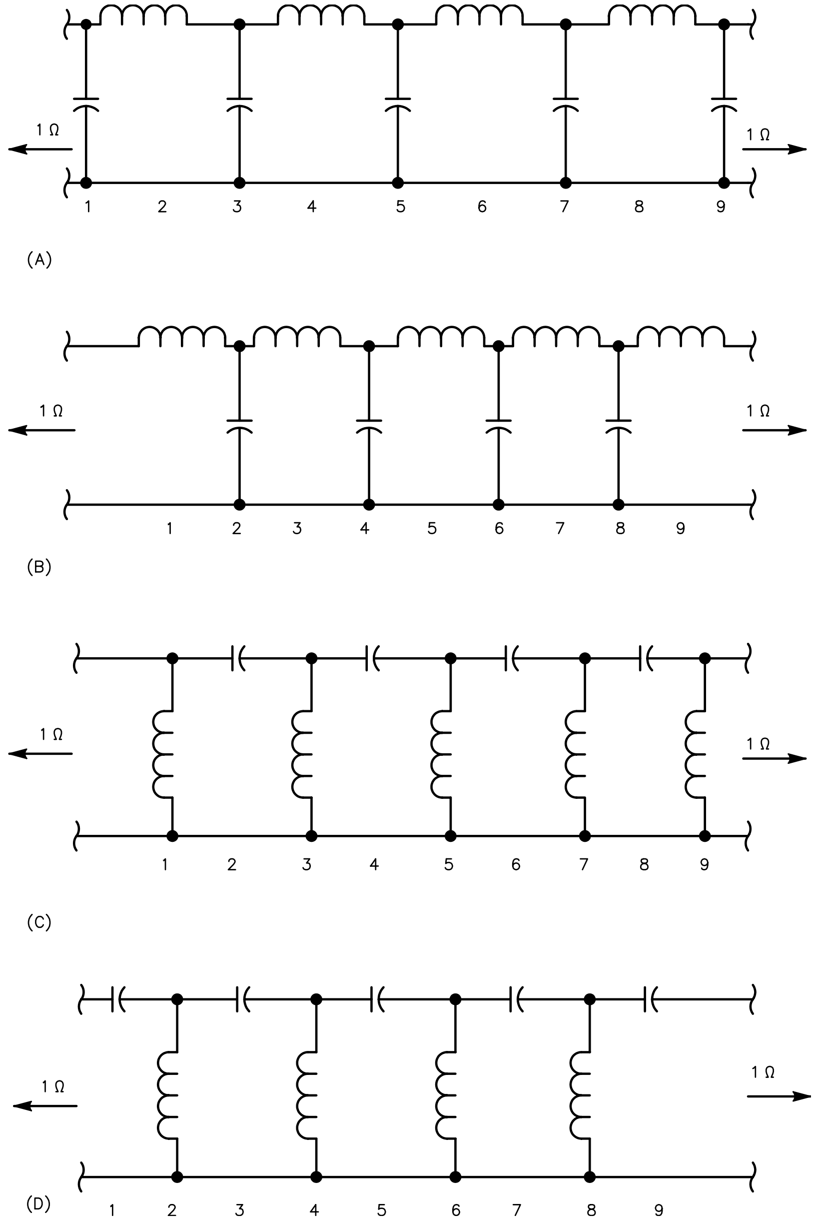


Fig 16.23—The schematic diagrams shown are low-pass and high-pass Chebyshev filters with the C-in/out and L-in/out configurations. For all normalized values see Table 16.2.

A: C-in/out low-pass configuration. Use the C and L values associated with the top column headings of the Table.

B: L-in/out low-pass configuration. For normalized values, use the L and C values associated with the bottom column headings of the Table.

C: L-in/out high-pass configuration is derived by transforming the C-in/out low-pass filter in A into an L-in/out high-pass by replacing all Cs with Ls and all Ls with Cs. The reciprocals of the lowpass component values become the highpass component values. For example, when n = 3, RC = 1.00% and C1 = 0.3524 F, L1 and L3 in C become 2.838 H.

D: The C-in/out high-pass configuration is derived by transforming the L-in/out low-pass in B into a C-in/out high-pass by replacing all Ls with Cs and all Cs with Ls. The reciprocals of the low-pass component values become the high-pass component values. For example, when n = 3, RC = 1.00% and L1 = 0.3524 H, C1 and C3 in D become 2.838 F.

and band-stop filters. The normalized element values of the low-pass C-in/out and L-in/out designs, Fig 16.23A and B, are read directly from the table using the values associated with either the top or bottom column headings, respectively.

The first four columns of Table 16.2 list *N* (the number of filter elements), *RC* (reflection coefficient percentage), *return loss* and the ratio of the 3-dB-to-F_{ap} frequencies. The passband maximum ripple amplitude (a_p) is not listed because it is difficult to measure. If necessary it can be calculated from the reflection coefficient. The $F3/F_{ap}$ ratio varies with N and RC; if both of these parameters are known, the $F3/F_{ap}$ ratio may be calculated. The remaining columns list the normalized Chebyshev element values for equally terminated filters for Ns from 3 to 9 in increments of 2.

The Chebyshev passband ends when the passband attenuation first exceeds the maximum ripple amplitude, a_p. This frequency is called the "ripple cutoff frequency, F_{ap}" and it has a normalized value of unity. All Chebyshev designs in Table 16.2 are based on the ripple cutoff frequency instead of the more familiar 3-dB frequency of the Butterworth response. However, the 3-dB frequency of a Chebyshev design may be obtained by multiplying the ripple cutoff frequency by the $F3/F_{ap}$ ratio listed in the fourth column.

The element values are normalized to a ripple cutoff frequency of 0.15915 Hz (one radian/sec) and 1-Ω terminations, so that the low-pass values can be transformed directly into high-pass values. This is done by replacing all Cs and Ls in the low-pass configuration with Ls and Cs and by replacing all the low-pass element values with their reciprocals. The normalized values are then multiplied by the appropriate C and L scaling factors to obtain the final values based on the desired ripple cutoff frequency and impedance level. The listed C and L element values are in farads and henries and become more reasonable after the values are scaled to the desired cutoff frequency and impedance level.

The normalized designs presented are a mixture: Some have integral values of *reflection coefficient (RC)* (1% and 10%) while others have "integral" values of *passband ripple amplitude* (0.001, 0.01 and 0.1 dB). These ripple amplitudes correspond to reflection coefficients of 1.517, 4.796 and 15.087%, respectively. By having tabulated designs based on integral values of both reflection coefficient and passband ripple amplitude, the correctness of the normalized component values may be checked against those same values published in filter handbooks whichever parameter, RC or a_p, is used.

In addition to the customary normalized design listings based on integral values of reflection coefficient or ripple amplitude, Table 16.2 also includes unique designs having special element ratios that make them more useful than previously published tables. For example, for N = 5 and RC = 6.302, the ratio of C3/C1 is 2.000. This ratio allows 5-element low-pass filters to be realized with only one capacitor value because C3 may be obtained by using parallel-connected capacitors each having the same value as C1 and C5.

In a similar way, for N = 7 and RC = 8.101, C3/C1 and C5/C1 are also 2.000. Another useful N = 7 design is that for RC = 1.427%. Here the L4/L2 ratio is 1.25, which is identical to 110/88. This means a seventh-order C-in/out low-pass audio filter can be realized with four surplus 88-mH inductors. Both L2 and L6 can be 88 mH while L4 is made up of a series connection of 22 mH and 88 mH. The 22-mH value is obtained by connecting the two windings of one of the four surplus inductors in parallel. Other useful ratios also appear in the N = 9 listing for both C3/C1 and L4/L2.

Except for the first two N = 5 designs, all designs were calculated for a reflection coefficient range from 1% to about 15%. The first two N = 5 designs were included because of their useful L3/L1 ratios. Designs with an RC of less than 1% are not normally used because of their poor selectivity. Designs with RC greater than 15% yield increasingly high SWR values with correspondingly increased objectionable reflective losses and sensitivity to termination impedance and component value variations.

Low-pass and high-pass filters may be realized in either a C-input/output or an L-input/output configuration. The C-input/output configuration is usually preferred because fewer inductors are required, compared to the L-input/output configuration. Inductors are usually more lossy, bulky and expensive than capacitors. The selection of the filter order or number of filter elements, N, is determined by the desired stop-band attenuation rate of increase and the tolerable reflection coefficient or SWR. A steeper attenuation slope requires either a design having a higher reflection coefficient or more circuit elements. Consequently, to select an optimum design, the builder must determine the amount of attenuation required in the stop band and the permissible maximum amount of reflection coefficient or SWR.

Table 16.3 shows the theoretical normalized frequencies (relative to the ripple cutoff frequency) for the listed attenuation levels and reflection coefficient percentages for Chebyshev low-pass filters of 3, 5, 7 and 9 elements. For example, for N = 5 and RC = 15.087%, an attenuation of 40 dB is reached at 2.22 times the ripple cutoff frequency (slightly more than one octave). The tabulated data are also applicable to high-pass filters by simply taking the reciprocal of the listed frequency. For example, for the same previous N and RC values, a high-pass filter attenuation will reach 40 dB at 1/2.22 = 0.450 times the ripple cutoff frequency.

The attenuation levels are theoretical and assume perfect components, no coupling between filter sections and no signal leakage around the filter. A working model should follow these values to the 60 or 70-dB level. Beyond this point, the actual response will likely degrade somewhat from the theoretical.

Fig 16.24 shows four plotted attenuation vs normalized frequency curves for N = 5 corresponding to the normalized frequencies in Table 16.3 At two octaves above the ripple cutoff frequency, f_c, the attenuation slope gradually becomes 6 dB per octave per filter element.

LOW-PASS AND HIGH-PASS FILTERS

Low-Pass Filter

Let's look at the procedure used to calculate the capacitor and inductor values of low-pass and high-pass filters by using two examples. Assume a 50-Ω low-pass filter is needed to give more than 40 dB of attenuation at $2f_c$ or one octave above the ripple-cutoff frequency of 4.0 MHz. Referring to Table 16.3, we see from the 40-dB column that a filter with 7 elements (N = 7) and a RC of 4.796% will reach 40 dB at 1.80 times the cutoff frequency or 1.8 × 4 = 7.2 MHz. Since this design has a reasonably low reflection coefficient and will satisfy the attenuation requirement, it is a good choice. Note that no 5-element filters are suitable for this application because 40 dB of attenuation is not achieved one octave above the cutoff frequency.

From Table 16.2, the nor-

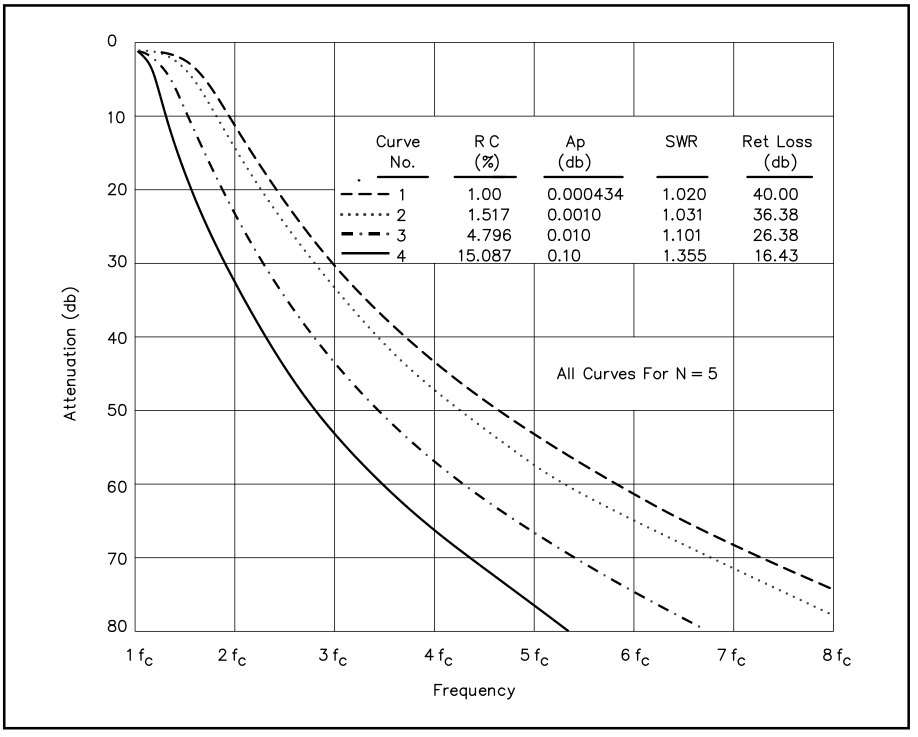


Fig 16.24—The graph shows attenuation vs frequency for four 5-element low-pass filters designed with the information obtained from Table 16.2. This graph demonstrates how reflection coefficient percentage (RC), maximum passband ripple amplitude (a_p), SWR, return loss and attenuation rolloff are all related. The exact frequency at a specified attenuation level can be obtained from Table 16.3.

malized component values corresponding to N = 7 and RC = 4.796% for the C-in/out configuration are: C1, C7 = 0.7970 F, L2, L6 = 1.392 H, C3, C5 = 1.748 F and L4 = 1.633 H. See Fig 16.23A for the corresponding configuration. The C and L normalized values will be scaled from a ripple cutoff frequency of one radian/sec and an impedance level of 1 Ω to a cutoff frequency of 4.0 MHz and an impedance level of 50 Ω. The C_s and L_s scaling factors are calculated:

$$C_s = \frac{1}{2\pi Rf} \tag{15}$$

$$L_s = \frac{R}{2\pi f} \tag{16}$$

where:
R = impedance level
f = cutoff frequency.

In this example:

$$C_S = \frac{1}{2\pi Rf} = \frac{1}{2\pi \times 50 \times 4 \times 10^6} = 795.8 \times 10^{-12}$$

$$L_S = \frac{R}{2\pi f} = \frac{50}{2\pi \times 4 \times 10^6} = 1.989 \times 10^{-6}$$

Using these scaling factors, the capacitor and inductor normalized values are scaled to the desired cutoff frequency and impedance level:

C1, C7 = 0.797 × 795.8 pF = 634 pF
C3, C5 = 1.748 × 795.8 pF = 1391 pF
L2, L6 = 1.392 × 1.989 µH = 2.77 µH
L4 = 1.633 × 1.989 µH = 3.25 µH

High-Pass Filter

The procedure for calculating a high-pass filter is similar to that for a low-pass filter, except a low-pass-to-high-pass transformation must first be performed. Assume a 50-Ω high-pass filter is needed to give more than 40 dB of attenuation one octave below (f_c/2) a ripple cutoff frequency of 4.0 MHz. Referring to Table 16.3, we see from the 40-dB column that a 7-element low-pass filter with RC of 4.796% will give 40 dB of attenuation at 1.8f_c. If this filter is transformed into a high-pass filter, the 40-dB level is reached at f_c/1.80 or at 0.556f_c = 2.22 MHz. Since the 40-dB level is reached before one octave from the 4-MHz cutoff frequency, this design will be satisfactory.

From Fig 16.23, we choose the low-pass L-in/out configuration in B and transform it into a high-pass filter by replacing all inductors with capacitors and all capacitors with inductors. Fig 16.23D is the filter configuration after the transformation. The reciprocals of the low-pass values become the high-pass values to complete the transformation. The high-pass values of the filter shown in Fig 16.23D are:

$$C1, C7 = \frac{1}{0.7970} = 1.255\,F$$

$$L2, L6 = \frac{1}{1.392} = 0.7184\,H$$

$$C3, C5 = \frac{1}{1.748} = 0.5721\,F$$

and

$$L4 = \frac{1}{1.633} = 0.6124\,H$$

Using the previously calculated C and L scaling factors, the high-pass component values are calculated the same way as before:

C1, C7 = 1.255 × 795.8 pF = 999 pF
C3, C5 = 0.5721× 795.8 pF = 455 pF
L2, L6 = 0.7184 × 1.989 µH = 1.43 µH
L4 = 0.6124 × 1.989 µH = 1.22 µH

BAND-PASS FILTERS

Band-pass filters may be classified as either narrowband or broadband. If the ratio of the upper ripple cutoff frequency to the lower cutoff frequency is greater than two, we have a wideband filter. For wideband filters, the band-pass filter (BPF) requirement may be realized by simply cascading separate high-pass and low-pass filters having the same design impedance. (The assumption is that the filters maintain their individual responses even though they are cascaded.) *For this to be true, it is important that both filters have a relatively low reflection coefficient percentage (less than 5%) so the SWR variations in the passband will be small.*

For narrowband BPFs, where the separation between the upper and lower cutoff frequencies is less than two, it is necessary to transform an appropriate low-pass filter into a BPF. That is, we use the low-pass normalized tables to design narrowband BPFs.

We do this by first calculating a low-pass filter (LPF) with a cutoff frequency equal to the desired bandwidth of the BPF. The LPF is then transformed into the desired BPF by resonating the low-pass components at the geometric center frequency of the BPF.

For example, assume we want a 50-Ω BPF to pass the 75/80-m band and attenuate all signals outside the band. Based on the passband ripple cutoff frequencies of 3.5 and 4.0 MHz, the geometric center frequency = $(3.5 \times 4.0)^{0.5} = (14)^{0.5}$ = 3.741657 or 3.7417 MHz. Let's slightly extend the lower and upper ripple cutoff frequencies to 3.45 and 4.058 MHz to account for possible component tolerance variations and to maintain the same center frequency. We'll evaluate a low-pass 3-element prototype with a cutoff frequency equal to the BPF passband of (4.058–3.45)MHz = 0.608 MHz as a possible choice for transformation.

Further, assume it is desired to attenuate the second harmonic of 3.5 MHz by at least 40 dB. The following calculations show how to design an N = 3 filter to provide the desired 40-dB attenuation at 7 MHz and above.

The bandwidth (BW) between 7 MHz on the upper attenuation slope (call it "f+") of the BPF and the corresponding frequency at the same attenuation level on the lower slope (call it "f–") can be calculated based on $(f+)(f-) = (f_c)^2$ or

$$f- = \frac{14}{7} = 2\text{ MHz}$$

Therefore, the bandwidth at this unknown attenuation level for 2 and 7 MHz is 5 MHz. This 5-MHz BW is normalized to the ripple cutoff BW by dividing 5.0 MHz by 0.608 MHz:

$$\frac{5.0}{0.608} = 8.22$$

We now can go to Table 16.3 and search for the corresponding normalized frequency that is closest to the desired normalized BW of 8.22. The low-pass design of N = 3 and RC = 4.796% gives 40 dB for a normalized BW of 8.08 and 50 dB for 11.83. Therefore, a design of N = 3 and RC = 4.796% with a normalized BW of 8.22 is at an attenuation level somewhere between 40 and 50 dB. Consequently, a low-pass design based on 3 elements and a 4.796% RC will give slightly more than the desired 40 dB attenuation above 7 MHz. The next step is to calculate the C and L values of the low-pass filter using the normalized component values in Table 16.2.

From this table and for N = 3 and RC = 4.796%, C1, C3 = 0.6292 F and L2 = 0.9703 H. We calculate the scaling factors as before and use 0.608 MHz as the ripple cutoff frequency:

$$C_S = \frac{1}{2\pi Rf} = \frac{1}{2\pi \times 50 \times 0.608 \times 10^6} = 5235 \times 10^{-12}$$

$$L_S = \frac{R}{2\pi f} = \frac{50}{2\pi \times 0.608 \times 10^6} = 13.09 \times 10^{-6}$$

C1, C3 = 0.6292 × 5235 pF = 3294 pF and
L2 = 0.9703 × 13.09 µH = 12.70 µH.

The LPF (in a pi configuration) is transformed into a BPF with 3.7417-MHz center frequency by resonating the low-pass elements at the center frequency. The resonating components will take the same identification numbers as the components they are resonating.

$$L1, L3 = \frac{25330}{\left(F_c^{\,2} \times C1\right)} = \frac{25330}{(14 \times 3294)} = 0.5493 \mu H$$

$$C2 = \frac{25330}{\left(F_c^{\,2} \times L2\right)} = \frac{25330}{(14 \times 12.7)} = 142.5 \text{ pF}$$

where L, C and f are in µH, pF and MHz, respectively.

The BPF circuit after transformation (for N = 3, RC = 4.796%) is shown in **Fig 16.25**.

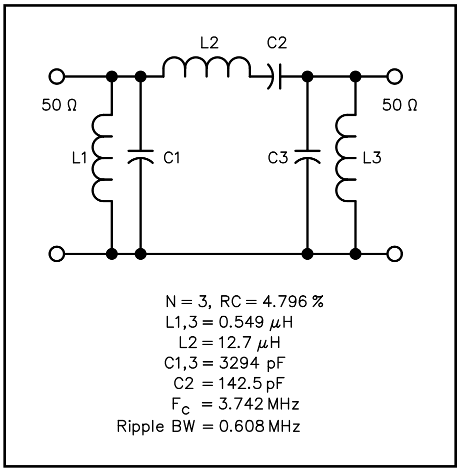


Fig 16.25—After transformation of the band-pass filter, all parallel elements become parallel LCs and all series elements become series LCs.

The component-value spread is

$$\frac{12.7}{0.549} = 23$$

and the reactance of L1 is about 13 Ω at the center frequency. For better BPF performance, the component spread should be reduced and the reactance of L1 and L3 should be raised to make it easier to achieve the maximum possible Q for these two inductors. This can be easily done by designing the BPF for an impedance level of 200 Ω and then using the center taps on L1 and L3 to obtain the desired 50-Ω terminations. The result of this approach is shown in **Fig 16.26**.

The component spread is now a more reasonable

$$\frac{12.7}{2.20} = 5.77$$

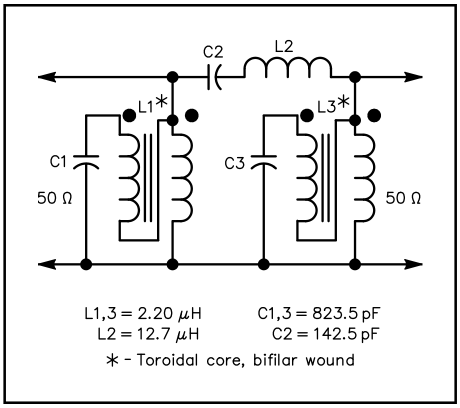


Fig 16.26—A filter designed for 200-Ω source and load provides better values. By tapping the inductors, we can use a 200-Ω filter design in a 50-Ω system.

and the L1, L3 reactance is 51.6 Ω. This higher reactance gives a better chance to achieve a satisfactory Q for L1 and L3 with a corresponding improvement in the BPF performance.

As a general rule, keep reactance values between 5 Ω and 500 Ω in a 50-Ω circuit. When the value falls below 5 Ω, either the equivalent series resistance of the inductor or the series inductance of the capacitor degrades the circuit Q. When the inductive reactance is greater than 500 Ω, the inductor is approaching self-resonance and circuit Q is again degraded. In practice, both L1 and L3 should be bifilar wound on a powdered-iron toroidal core to assure that optimum coupling is obtained between turns over the entire winding. The junction of the bifilar winding serves as a center tap.

Side-Slope Attenuation Calculations

The following equations allow the calculation of the frequencies on the upper and lower sides of a BPF response curve at any given attenuation level if the bandwidth at that attenuation level and the geometric center frequency of the BPF are known:

$$f_{lo} = -X + \sqrt{f_c^{\,2} + X^2} \quad (17)$$

$$f_{hi} = f_{lo} + BW \quad (18)$$

where
BW = bandwidth at the given attenuation level,
f_c = geometric center frequency

$$X = \frac{BW}{2}$$

For example, if f = 3.74166 MHz and BW = 5 MHz, then

$$\frac{BW}{2} = X = 2.5$$

and:

$$f_{lo} = -2.5 + \sqrt{3.74166^2 + 2.5^2} = 2.00\,\text{MHz}$$

$$f_{hi} = f_{lo} + BW = 2 + 5 = 7\text{ MHz}$$

BAND-STOP FILTERS

Band-stop filters may be classified as either narrowband or broadband. If the ratio of the upper ripple cutoff frequency to the lower cutoff frequency is greater than two, the filter is considered wideband. A wideband band-stop filter (BSF) requirement may be realized by simply paralleling the inputs and outputs of separate low-pass and high-pass filters having the same design impedance and with the low-pass filter having its cutoff frequency one octave or more below the high-pass cutoff frequency.

In order to parallel the low-pass and high-pass filter inputs and outputs without one affecting the other, it is essential that each filter have a high impedance in that portion of its stop band which lies in the passband of the other. *This means that each of the two filters must begin and end in series branches.* In the low-pass filter, the input/output series branches must consist of inductors and in the high-pass filter, the input/output series branches must consist of capacitors.

When the ratio of the upper to lower cutoff frequencies is less than two, the BSF is considered to be narrowband, and a calculation procedure similar to that of the narrowband BPF design procedure is used. However, in the case of the BSF, the design process starts with the design of a high-pass filter having the desired impedance level of the BSF and a ripple cutoff frequency the same as that of the desired ripple bandwidth of the BSF. After the HPF design is completed, every high-pass element is resonated to the center frequency of the BSF in the same manner as if it were a BPF, except that all shunt branches of the BSF will consist of series-tuned circuits, and all series branches will consist of parallel-tuned circuits—just the opposite of the resonant circuits in the BPF. The reason for this becomes obvious when the impedance characteristics of the series and parallel circuits at resonance are considered relative to the intended purpose of the filter, that is, whether it is for a band-pass or a band-stop application.

Quartz Crystal Filters

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

The quartz resonator has the equivalent circuit shown in **Fig 16.31**. L_s, C_s and R_s represent the *motional* reactances and loss resistance. C_p is the parallel plate capacitance formed by the two metal electrodes separated by the quartz dielectric. Quartz has a dielectric constant of 3.78. **Table 16.4** shows parameter values for typical moderate-cost quartz resonators. Q_U is the resonator unloaded Q.

$$Q_U = 2\pi f_s r_s \qquad (19)$$

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

A quartz resonator connected between generator and load, as shown in **Fig 16.32A**, produces the

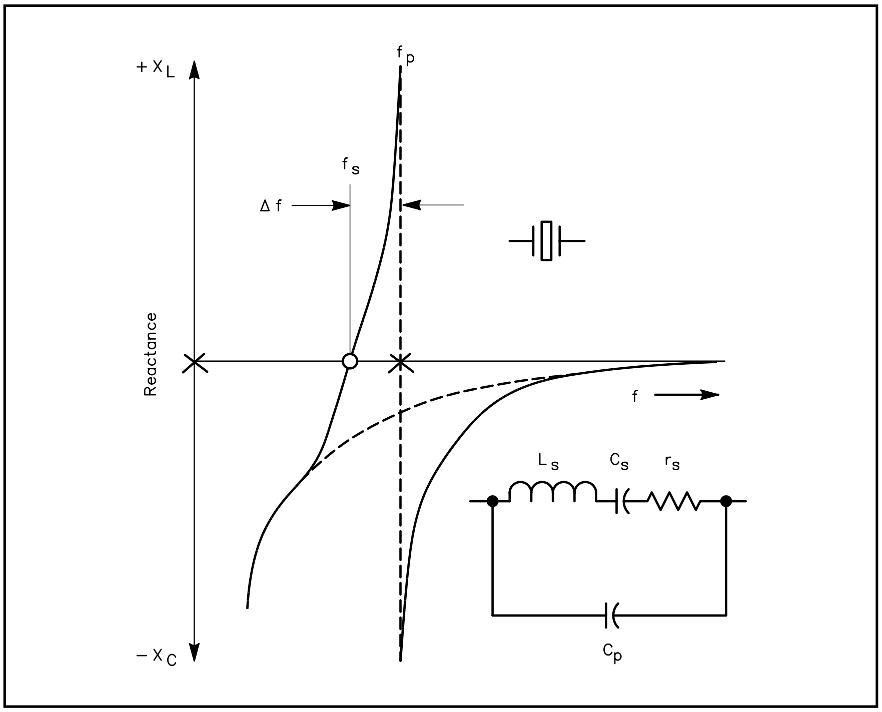


Fig 16.31—Equivalent circuit of a quartz crystal. The curve plots the crystal reactance against frequency. At f_p, the resonance frequency, the reactance curve goes to infinity.

Table 16.4
Typical Parameters for AT-Cut Quartz Resonators

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

Courtesy of Piezo Crystal Co, Carlisle, Pennsylvania

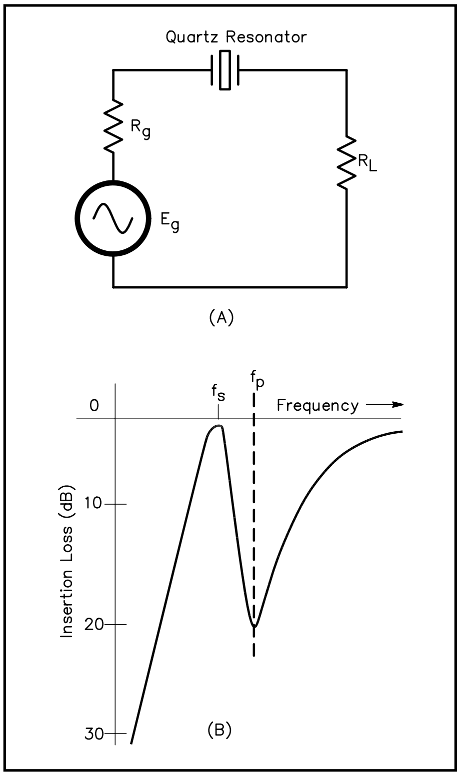


Fig 16.32—A: Series test circuit for a crystal. In the test circuit the output of a variable frequency generator, e_g, is used as the test signal. The frequency response in B shows the highest attenuation at resonance (f_p). See text.

frequency response of Fig 16.32B. There is a relatively low loss at the series resonant frequency f_s and high loss at the parallel resonant frequency f_p. The test circuit of Fig 16.32A is useful for determining the parameters of a quartz resonator, but yields a poor filter.

A crystal filter developed in the 1930s is shown in **Fig 16.33A**. The disturbing effect of C_p (which produces f_p) is canceled by the *phasing capacitor*, C1. The voltage reversing transformer T1 usually consists of a bifilar winding on a ferrite core. Voltages V_a and V_b have equal magnitude but 180° phase difference. When $C1 = C_p$, the effect of C_p will disappear and a well-behaved single resonance will occur as shown in Fig 16.33B. The band-pass filter will exhibit a loaded Q given by:

$$Q_L = \frac{2\pi f_S L_S}{R_L} \qquad (20)$$

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

The half-lattice filter shown in **Fig 16.34** is an improvement in crystal filter design. The quartz resonator parallel-plate capacitors, C_p, cancel each other. Remaining series resonant circuits, if properly offset in frequency, will produce an approximate 2-pole Butterworth or Chebyshev response. Crystals A and B are usually chosen so that the parallel resonant frequency (f_p) of one is the same as the series resonant frequency (f_s) of the other.

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

Many quartz crystal filters produced today use the ladder network design shown in **Fig 16.35**. In this configuration, all resonators have the same series resonant frequency f_s. Interresonator coupling is provided by shunt capacitors such as C12 and C23. Refs 12 and 13 provide good ladder filter design information. A test set for evaluating crystal filters is presented in the **Projects** section of this chapter.

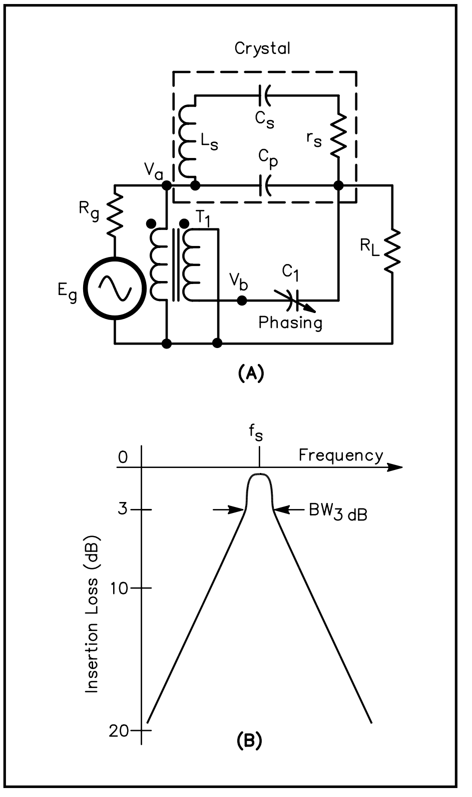


Fig 16.33—The practical one-stage crystal filter in A has the response shown in B. The phasing capacitor is adjusted for best response (see text).

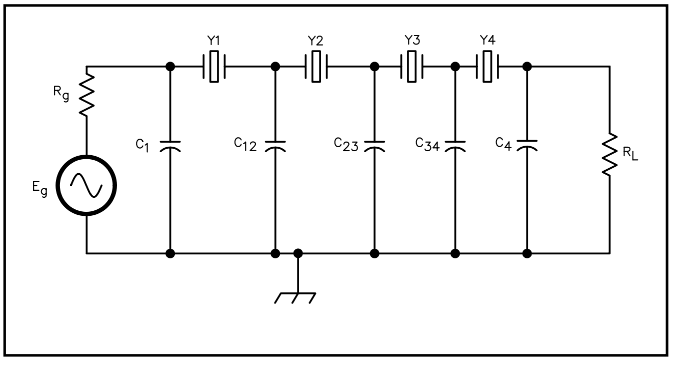


Fig 16.35—A four-stage crystal ladder filter. The crystals must be chosen properly for best response.

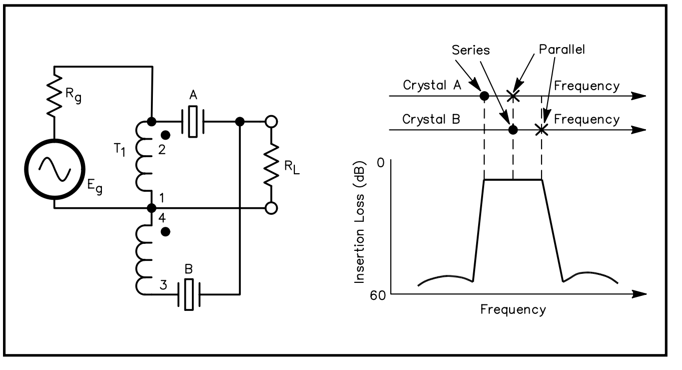


Fig 16.34—A half-lattice crystal filter. No phasing capacitor is needed in this circuit.

Monolithic Crystal Filters

A monolithic (Greek: one-stone) crystal filter has two sets of electrodes deposited on the same quartz plate, as shown in **Fig 16.36**. This forms two resonators with acoustic (mechanical) coupling between them. If the acoustic coupling is correct, a 2-pole Butterworth or Chebyshev response will be achieved. More than two resonators can be fabricated on the same plate yielding a multipole response. Monolithic crystal filter technology is popular because it produces a low parts count, single-unit filter at lower cost than a lumped-element equivalent. Monolithic crystal filters are typically manufactured in the range from 5 to 30 MHz for the fundamental mode and up to 90 MHz for the third-overtone mode. Q_L ranges from 200 to 10,000.

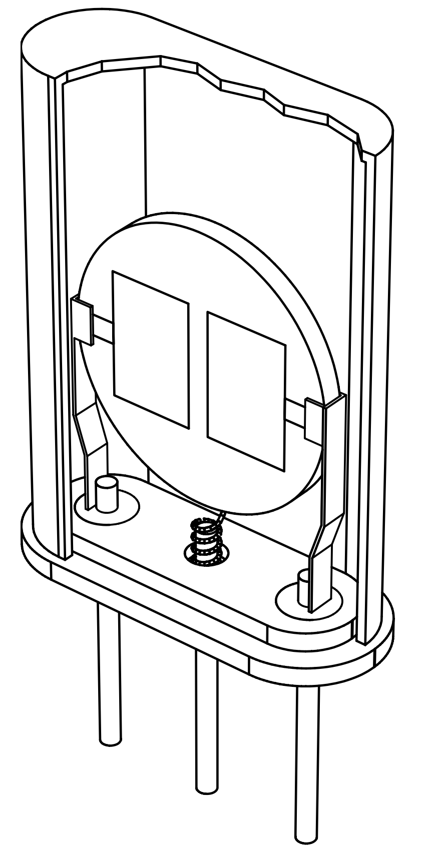


Fig 16.36—Typical two-pole monolithic crystal filter. This single small ($^1/_2$ to $^3/_4$-inch) unit can replace 6 to 12, or more, discrete components.

SAW Filters

The resonators in a monolithic crystal filter are coupled together by bulk acoustic waves. These acoustic waves are generated and propagated in the interior of a quartz plate. It is also possible to launch, by an appropriate transducer, acoustic waves which 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 16.37**. Lithium Niobate ($LiNbO_3$) is usually favored over quartz because it yields less insertion loss. The electrodes make up the filter's transducers. RF voltage is applied to the input transducer and generates electric fields between the fingers. The piezoelectric material vibrates launching an acoustic wave along the surface. When the wave reaches the output transducer it produces an electric field between the fingers. This field generates a voltage across the load resistor.

Since both input and output transducers are not entirely unidirectional, some acoustic power is lost in the acoustic absorbers located behind each transducer. This lost acoustic power produces a midband 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 midband loss can be tolerated.

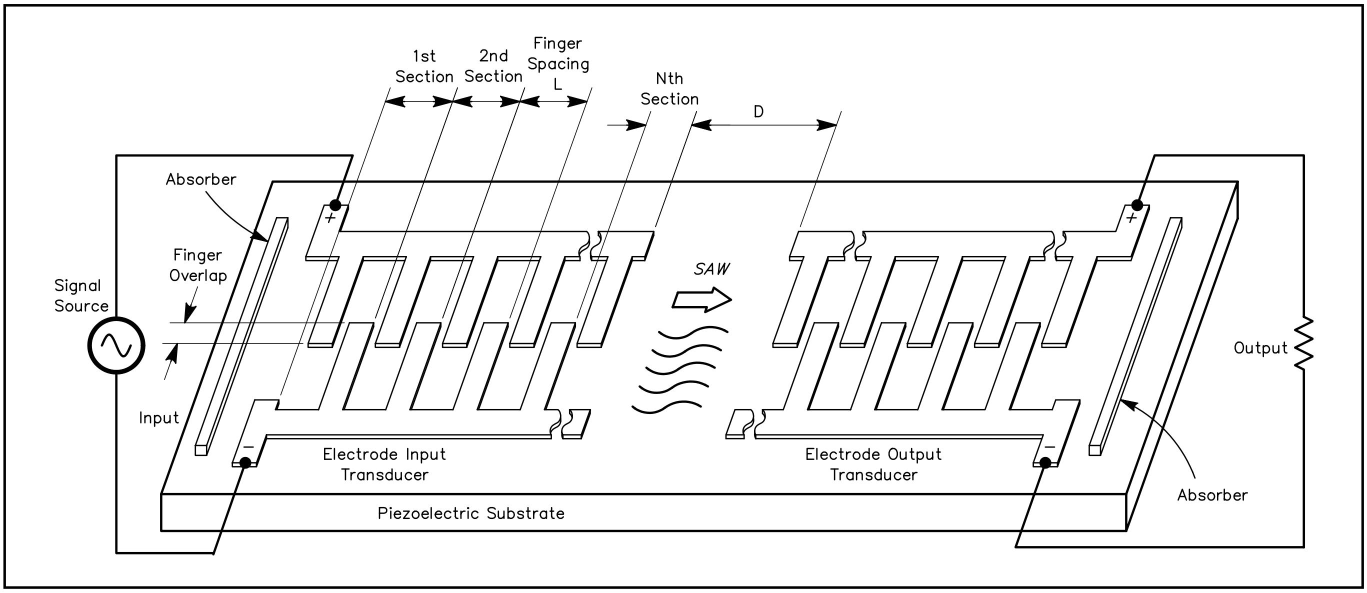


Fig 16.37—The *interdigitated* transducer, on the left, launches SAW energy to a similar transducer on the right (see text).

Transmission-Line Filters

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

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

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

Transmission Lines for Filters

Fig 16.38 shows three popular transmission lines used in transmission-line filters. The circular coaxial transmission line (coax) shown in Fig 16.38A consists of two concentric metal cylinders separated by dielectric (insulating) material. Coaxial transmission line possesses a characteristic impedance given by:

$$Z_0 = \frac{138}{\sqrt{\varepsilon}} \log\left(\frac{D}{d}\right) \qquad (21)$$

A plot of Z_0 vs D/d is shown in **Fig 16.39**. At RF, Z_0 is an almost pure resistance. If the distant end of a section of coax is terminated in Z_0, then the impedance seen looking into the input end is also Z_0 at all frequencies. A terminated section of coax is shown in **Fig 16.40A**. If the distant end is not terminated in Z_0, the input impedance will be some other value. In Fig 16.40B the distant end is short circuited and the length is less than $^1/_4\ \lambda$. The input impedance is an inductive reactance as seen by the notation $+j$ in the equation in part B of the figure.

The input impedance for the case of the open-circuit distant end, is shown in Fig 16.40C. This case results in a capacitive reactance ($-j$). Thus, short sections of coaxial line (stubs) can replace the inductors and capacitors in an LC filter. Coax line inductive stubs usually have lower loss than their lumped counterparts.

X_L vs ℓ for shorted and open stubs is shown in **Fig 16.41**. There is an optimum value of Z_0 that yields lowest loss, given by

$$Z_0 = \frac{75}{\sqrt{\varepsilon}} \qquad (22)$$

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

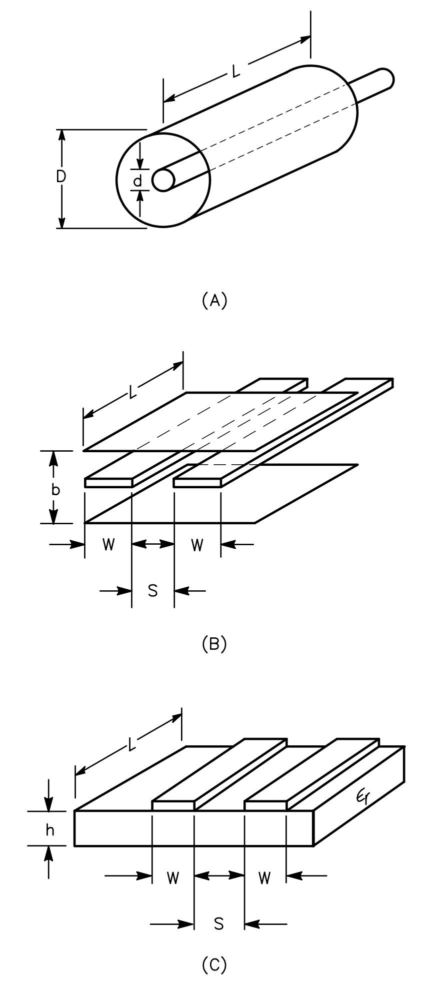


Fig 16.38—Transmission lines. A: Coaxial line. B: Coupled stripline, which has two ground planes. C: Microstripline, which has only one ground plane.

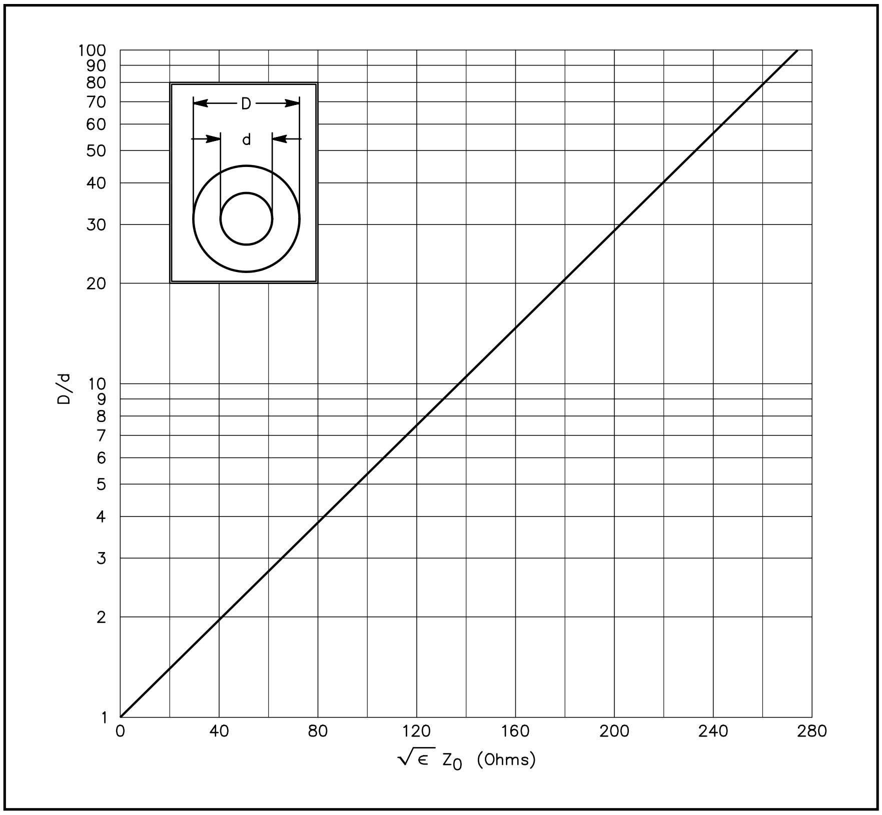


Fig 16.39—Coaxial-line impedance varies with the ratio of the inner- and outer-conductor diameters. The dielectric constant, ε, is 1.0 for air and 2.32 for polyethylene.

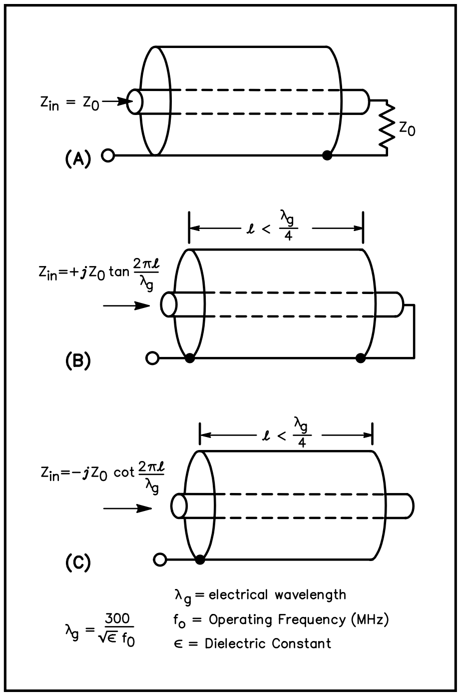


Fig 16.40—Transmission line stubs. A: A line terminated in its characteristic impedance. B: A shorted line less than 1/4-λ long is an *inductive* stub. C: An open line less than 1/4-λ long is a *capacitive* stub.

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

Fig 16.41—Stub reactance for various lengths of transmission line. Values are for Z_0 = 50 Ω. For Z_0 = 100 Ω, double the tabulated values.

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 16.38B. 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 16.38B. Stripline Z_0 vs width (*w*) is plotted in **Fig 16.42**. Air-dielectric stripline technology is best for low bandwidth (Q_L > 20) band-pass filters.

The most popular transmission line is microstrip (unshielded stipline), shown in Fig 16.38C. It can be fabricated with standard printed-circuit processes and is the least expensive configuration. Unfortunately, microstrip is the most lossy of the three lines; therefore it is not suitable for narrow band-pass filters. In microstrip the outer conductor is a single flat metal ground-plane. The inner conductor is a thin metal strip separated from the ground-plane by a solid dielectric substrate. Typical substrates are 0.062-inch G-10 fiberglass (ε = 4.5) for the 50 MHz to

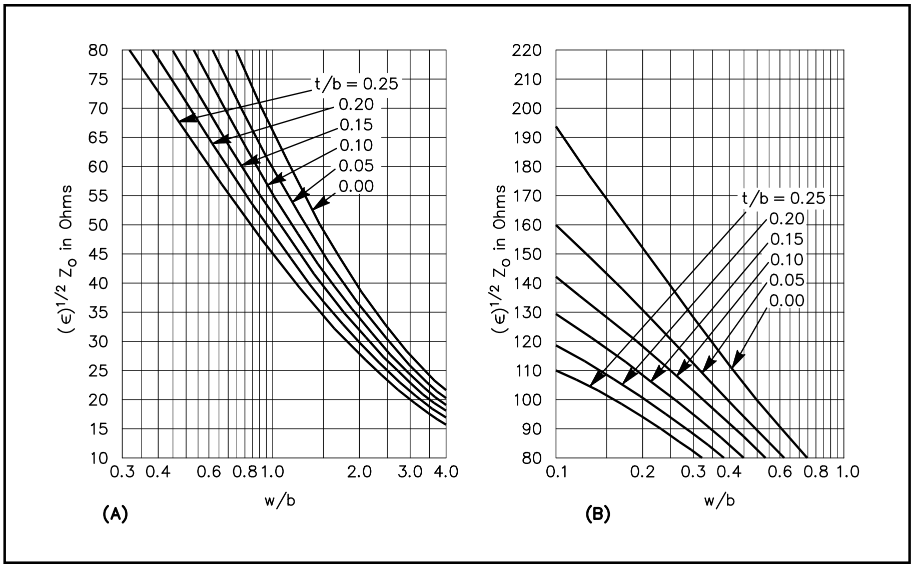


Fig 16.42—The Z_0 of stripline varies with *w*, *b* and *t* (conductor thickness). See Fig 16.38B. The conductor thickness is *t* and the plots are normalized in terms of *t/b*.

1-GHz frequency range and 0.031-inch Teflon ($\varepsilon = 2.3$) for frequencies above 1 GHz.

Conductor separation must be minimized or free-space radiation and unwanted coupling to adjacent circuits may become problems. Microstrip characteristic impedance and the effective dielectric constant (ε) are shown in **Fig 16.43**. Unlike coax and stripline, the effective dielectric constant is less than that of the substrate since a portion of the electromagnetic wave propagating along the microstrip "sees" the air above the substrate.

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

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				

Fig 16.43—Microstrip parameters (after H. Wheeler, *IEEE Transactions on MTT*, March 1965, p 132). ε_e is the effective ε.

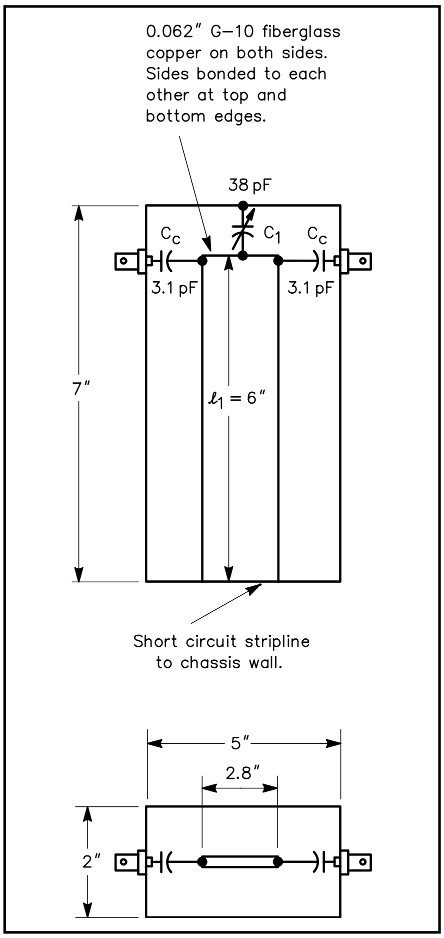


Fig 16.44—This 146-MHz stripline band-pass filter has been measured to have a Q_L of 63 and a loss of approximately 1 dB.

Transmission-Line Band-Pass Filters

Band-pass filters can also be constructed from transmission-line stubs. At VHF the stubs can be considerably shorter than a quarter wavelength yielding a compact filter structure with less midband loss than its LC counterpart. The single-stage 146-MHz stripline band-pass filter shown in **Fig 16.44** 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 midband loss = 1 dB.

Single-stage stripline filters can be coupled together to yield multistage filters. One method uses the capacitor coupled band-pass filter synthesis technique to design a 3-pole filter. Another method allows closely spaced inductive stubs to magnetically couple to each other. When the coupled stubs are grounded on the same side of the filter housing, the structure is called a "combline filter." Three examples of combline band-pass filters are shown in **Fig 16.45**. These filters are constructed in 2 × 7 × 9-inch chassis boxes.

Quarter-Wave Transmission-Line Filters

Fig 16.41 shows that when $\ell = 0.25\,\lambda_g$, the shorted-stub reactance becomes infinite. Thus, a $^1/_4$-λ shorted stub behaves like a parallel-resonant LC circuit. Proper input and output coupling to a $^1/_4$-λ resonator yields a practical band-pass filter. Closely spaced $^1/_4$-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 16.46**. Design graphs for round-rod interdigital filters are given in Ref 16. The $^1/_4$-λ resonators may be tuned by physically changing their lengths or by tuning the screw opposite each rod.

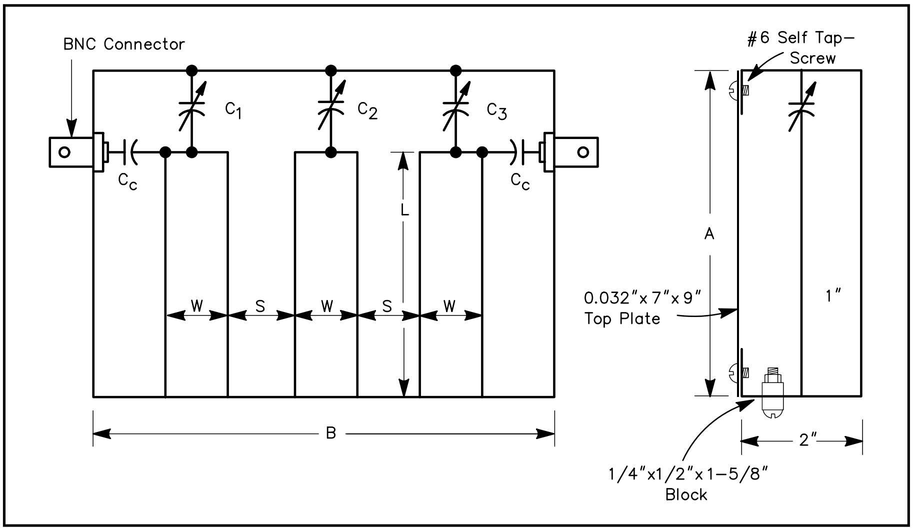


Fig 16.45—This Butterworth filter is constructed in combline. It was originally discussed by R. Fisher in December 1968 *QST*.

Dimension	***52 MHz***	***146 MHz***	***222 MHz***
A	**9"**	**7"**	**7"**
B	**7"**	**9"**	**9"**
L	**7 3/8"**	**6"**	**6"**
S	**1"**	**1 1/16"**	**1 3/8"**
W	**1"**	**1 5/8"**	**1 5/8"**
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**	**—**

If the short-circuited ends of two $^1/_4$-λ resonators are connected to each other, the resulting $^1/_2$-λ stub will remain in resonance, even when the connection to ground-plane is removed. Such a floating $^1/_2$-λ microstrip line, when bent into a U-shape, is called a "hairpin" resonator. Closely coupled hairpin resonators can be arranged to form multistage band-pass filters. Microstrip hairpin band-pass filters are popular above 1 GHz because they can be easily fabricated using photo-etching techniques. No connection to the ground-plane is required.

Transmission-Line Filters Emulating LC Filters

Low-pass and high-pass transmission-line filters are usually built from short sections of transmission lines (stubs) that emulate lumped LC reactances. Sometimes low-loss lumped capacitors are mixed with transmission-line inductors to form a hybrid component filter. For example, consider the 720-MHz 3-pole microstrip low-pass filter shown in **Fig 16.47A** that emulates the LC filter shown in Fig 16.47B. C1 and C3 are replaced with 50-Ω open-circuit shunt stubs ℓ_C long. L2 is replaced with a short section

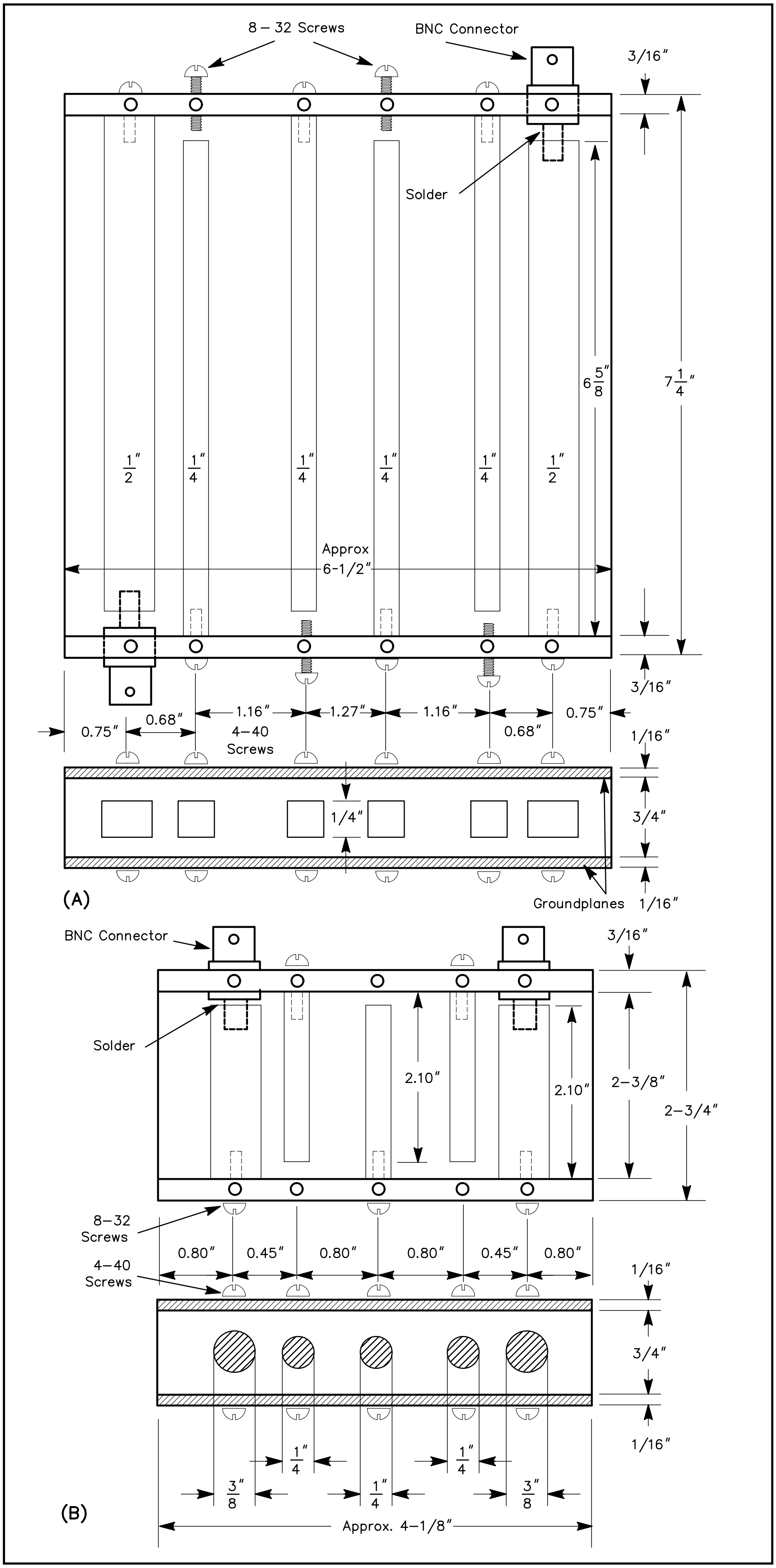


Fig 16.46—These 3-pole Butterworth filters (upper: 432 MHz, 8.6 MHz bandwidth, 1.4 dB pass-band loss; lower: 1296 MHz, 110 MHz bandwidth, 0.4 dB pass-band loss) are constructed as interdigitated filters. The material is from R. E. Fisher, March 1968 *QST.*

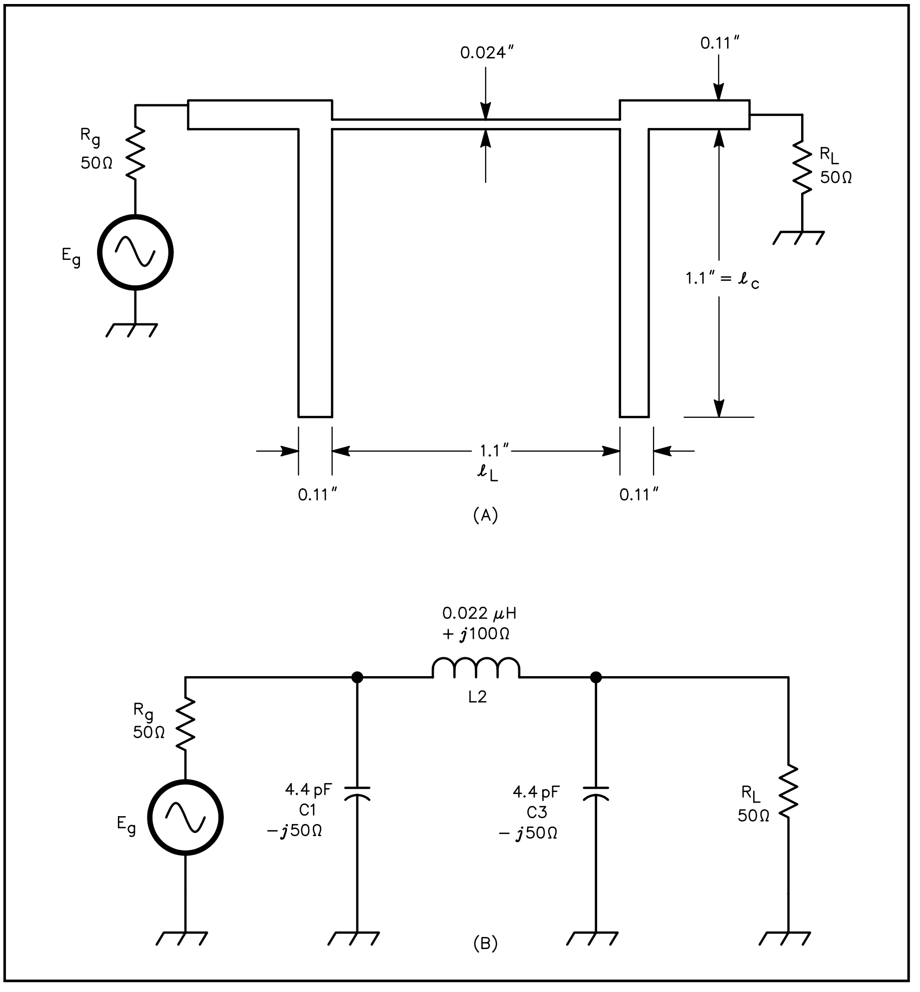


Fig 16.47—A microstrip 3-pole emulated-Butterworth low-pass filter with a cutoff frequency of 720 MHz. A: Microstrip version built with G-10 fiberglass board (ε = 4.5, h = 0.062 inches). B: Lumped LC version of the same filter. To construct this filter with lumped elements very small values of L and C must be used and stray capacitance and inductance must be reduced to a tiny fraction of the component values.

of 100-Ω line ℓ_L long. The LC filter, Fig 16.47B, was designed for f_c = 720 MHz. Such a filter could be connected between a 432-MHz transmitter and antenna to reduce harmonic and spurious emissions. A reactance chart shows that X_C is 50 Ω, and the inductor reactance is 100 Ω at f_c. The microstrip version is constructed on G-10 fiberglass 0.062-inch thick, with ε = 4.5. Then, from Fig 16.43, *w* is 0.11 inch and $l_C = 0.125\ \lambda_g$ for the 50-Ω capacitive stubs. Also, from Fig 16.43, *w* is 0.024 inch and ℓ_L is 0.125 ℓ_g for the 100-Ω inductive line. The inductive line length is approximate because the far end is not a short circuit. ℓ_g is 300/(720)(1.75) = 0.238 m, or 9.37 inches. Thus ℓ_C is 1.1 inch and ℓ_L is 1.1 inch.

This microstrip filter exhibits about 20 dB of attenuation at 1296 MHz. Its response rises again, however, around 3 GHz. This is because the fixed-length transmission-line stubs change in terms of wavelength as the frequency rises. This particular filter was designed to eliminate third-harmonic energy near 1296 MHz from a 432-MHz transmitter and does a better job in this application than the Butterworth filter in Fig 16.46, which has spurious responses in the 1296-MHz band.

Helical Resonators

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

Helical resonators are usually a better choice than $^1/_4$-λ cavities on 50, 144 and 222 MHz to eliminate these unwanted inputs. They are smaller and easier to build. In the frequency range from 30 to 100 MHz it is difficult to build high-Q inductors and coaxial cavities are very large. In this frequency range the helical resonator is an excellent choice. At 50 MHz for example, a capacitively tuned, $^1/_4$-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 W1HR in a *QST* article as a coil surrounded by a shield, but it is actually a shielded, resonant section of helically wound transmission line with relatively high characteristic impedance and low axial propagation velocity. The electrical length is about 94% of an axial $^1/_4$-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 16.48**. Although the shield may be any shape, only round and square shields will be considered here.

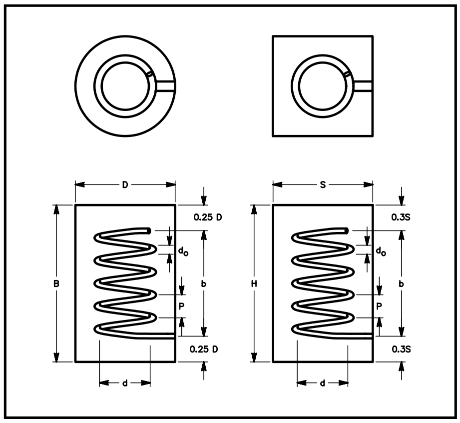


Fig 16.48—Dimensions of round and square helical resonators. The diameter, D (or side, S) is determined by the desired unloaded Q. Other dimensions are expressed in terms of D or S (see text).

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

$$Q_U = 50D\sqrt{f_o} \quad (23)$$

where

D = inside diameter of the shield, in inches

f_o = frequency, in MHz.

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

$$N = \frac{1908}{f_0 D} \quad (24A)$$

$$P = \frac{f_0 D^2}{2312} \qquad (24B)$$

$$Z_0 = \frac{99{,}000}{f_0 D} \qquad (24C)$$

$$N = \frac{1590}{f_0 S} \qquad (24D)$$

$$P = \frac{f_0 S^2}{1606} \qquad (24E)$$

$$Z_0 = \frac{82{,}500}{f_0 S} \qquad (24F)$$

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 16.49** is based on these formulas.

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

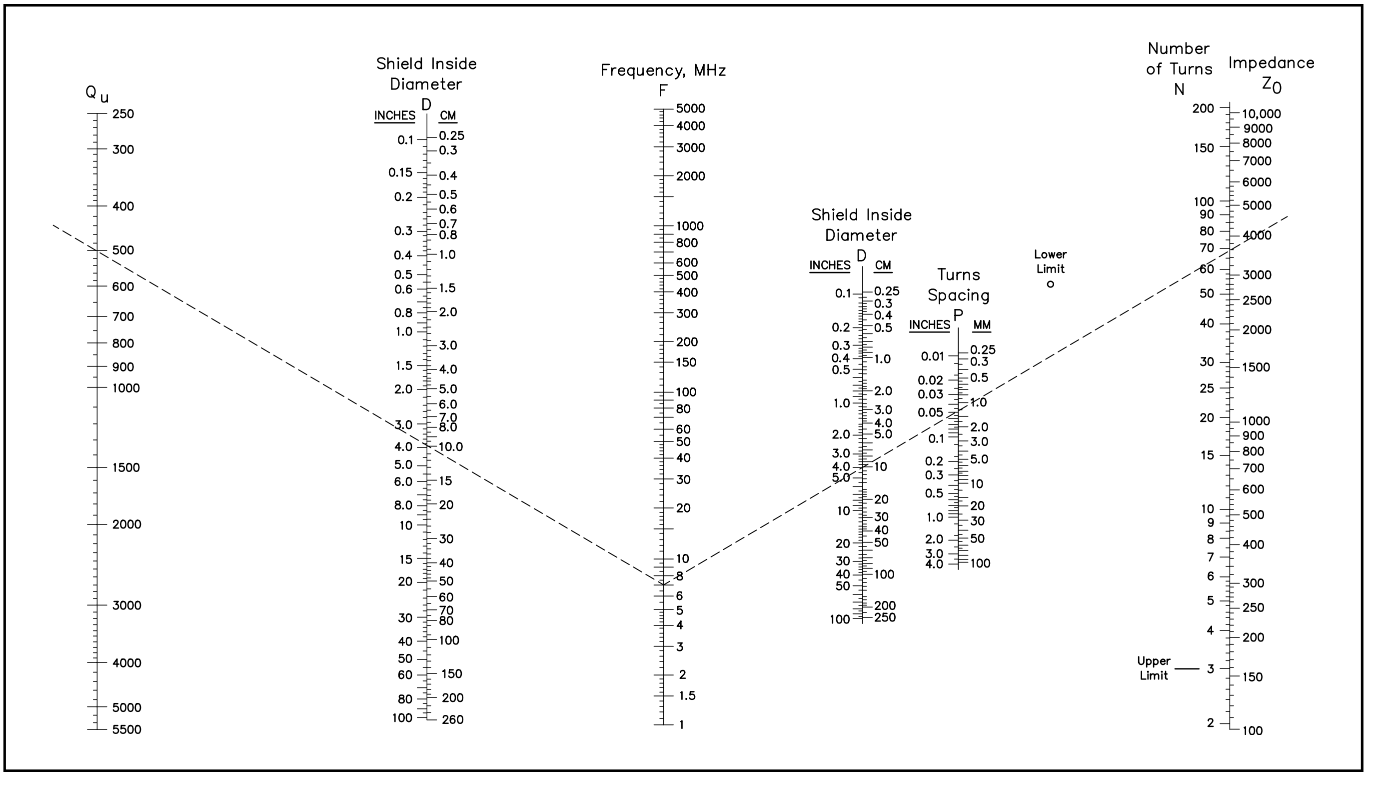


Fig 16.49—The design nomograph for round helical resonators starts by selecting Q_U and the required shield diameter. A line is drawn connecting these two values and extended to the frequency scale (example here is for a shield of about 3.8 inches and Q_U of 500 at 7 MHz). Finally the number of turns, 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 16.50 simplifies calculation of these dimensions. Note that these ratios result in a helix with a length 1.5 times its diameter, the condition for maximum Q. The shield is about 60% longer than the helix—although it can be made longer—to completely contain the electric field at the top of the helix and the magnetic field at the bottom.

The winding pitch, P, is used primarily to determine the required conductor size. Adjust the length of the coil to that given by the equations during construction. Conductor size ranges from 0.4 P to 0.6 P for both round and square resonators and are plotted graphically in **Fig 16.51**.

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 nonhelical). This condition occurs when the helix has fewer than three turns (the "upper limit" on the design nomograph of Fig 16.49).

Construction

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

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

Although the external field is minimized by

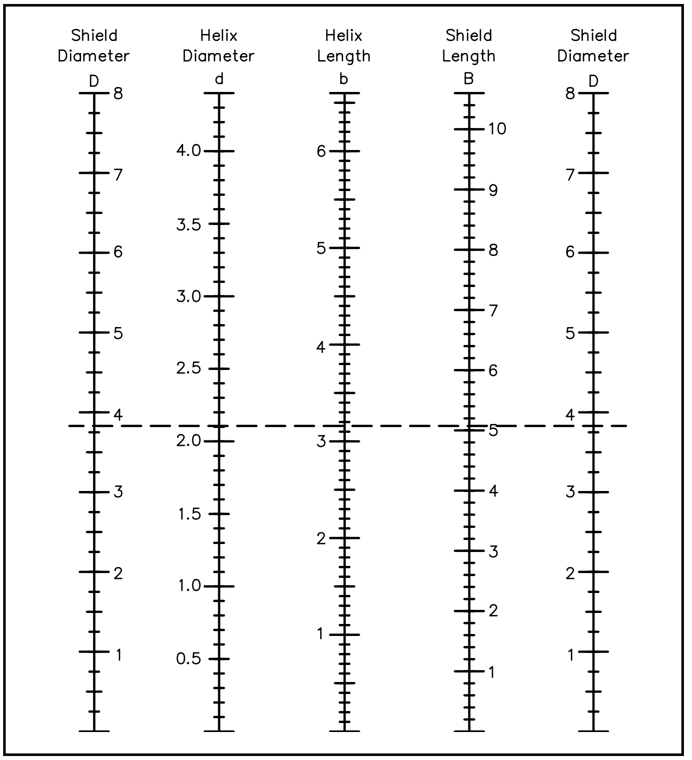


Fig 16.50—The helical resonator is scaled from this design nomograph. Starting with the shield diameter, the helix diameter, d, helix length, b, and shield length, B, can be determined with this graph. The example shown has a shield diameter of 3.8 inches. This requires a helix mean diameter of 2.1 inches, helix length of 3.1 inches, and shield length of 5 inches.

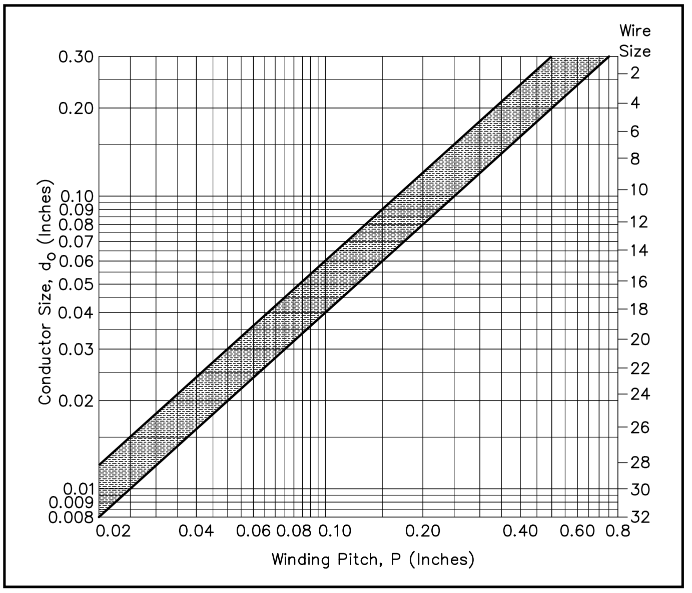


Fig 16.51—This chart provides the design information of helix conductor size vs winding pitch, P. For example, a winding pitch of 0.047 inch results in a conductor diameter between 0.019 and 0.028 inch (#22 or #24 AWG).

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

Tuning

A carefully built helical resonator designed from the nomograph of Fig 16.49 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 16.50, the resonator can be tuned by pruning the open end of the coil. However, neither of these methods is recommended for wide frequency excursions because any major deviation in helix length will degrade the unloaded Q of the resonator.

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

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

Insertion Loss

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

$$I_L = 20\log_{10}\left(\frac{1}{1-\frac{Q_L}{Q_U}}\right) \tag{25}$$

where
Q_L = loaded Q
Q_U = unloaded Q

This is plotted in **Fig 16.52**. 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.

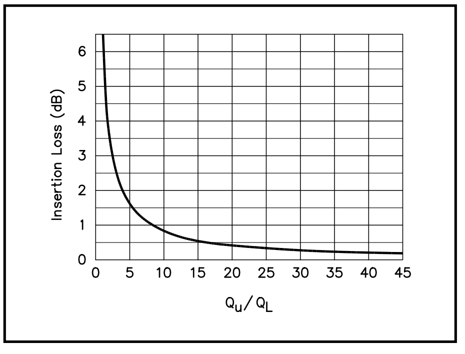


Fig 16.52—The ratio of loaded (Q_L) to unloaded (Q_U) Q determines the insertion loss of a tuned resonant circuit.

Coupling

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

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

Helical resonators with inductive coupling (loops) exhibit more attenuation to signals above the resonant frequency (as compared to attenuation below resonance), whereas resonators with capacitive coupling (probes) exhibit more attenuation below the passband, as shown for a typical 432-MHz resonator in **Fig 16.53**. Consider this characteristic when choosing a coupling method. The passband can be made more symmetrical by using a combination of coupling methods (inductive input and capacitive output, for example).

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

Since the loaded Q of a resonator is determined by the external loading, this must be considered when selecting a tap (or position of a loop or probe). The ratio of this external loading, R_b, to the characteristic impedance, Z_0, for a $^1/_4$-λ resonator is calculated from:

$$K = \frac{R_b}{Z_0} = 0.785\left(\frac{1}{Q_L} - \frac{1}{Q_U}\right) \quad (26)$$

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.

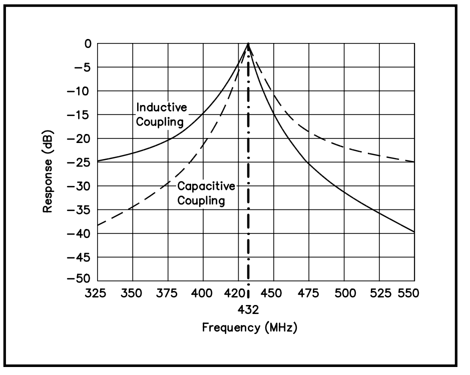


Fig 16.53—This response curve for a single-resonator 432-MHz filter shows the effects of capacitive and inductive input/output coupling. The response curve can be made symmetrical on each side of resonance by combining the two methods (inductive input and capacitive output, or vice versa).

Active Filters

Passive HF filters are made from combinations of inductors and capacitors. These may be used at low frequencies, but the inductors often become a limiting factor because of their size, weight, cost and losses. The active filter is a compact, low-cost alternative made with op amps, resistors and capacitors. They often occupy a fraction of the space required by an LC filter. While active filters have been traditionally used at low and audio frequencies, modern op amps with small-signal bandwidths that exceed 1 GHz have extended their range into MF and HF.

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

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

The design equations for various filters are shown in **Fig 16.54**. A program (ACTFIL.EXE), useful for designing 1-4 stage filters, is available from *ARRLWeb* (see page viii). **Fig 16.55** shows a typical application of a two-stage, bandpass filter. A two-stage filter is considered the minimum acceptable for CW, while three or four stages will prove more effective under some conditions of noise and interference.

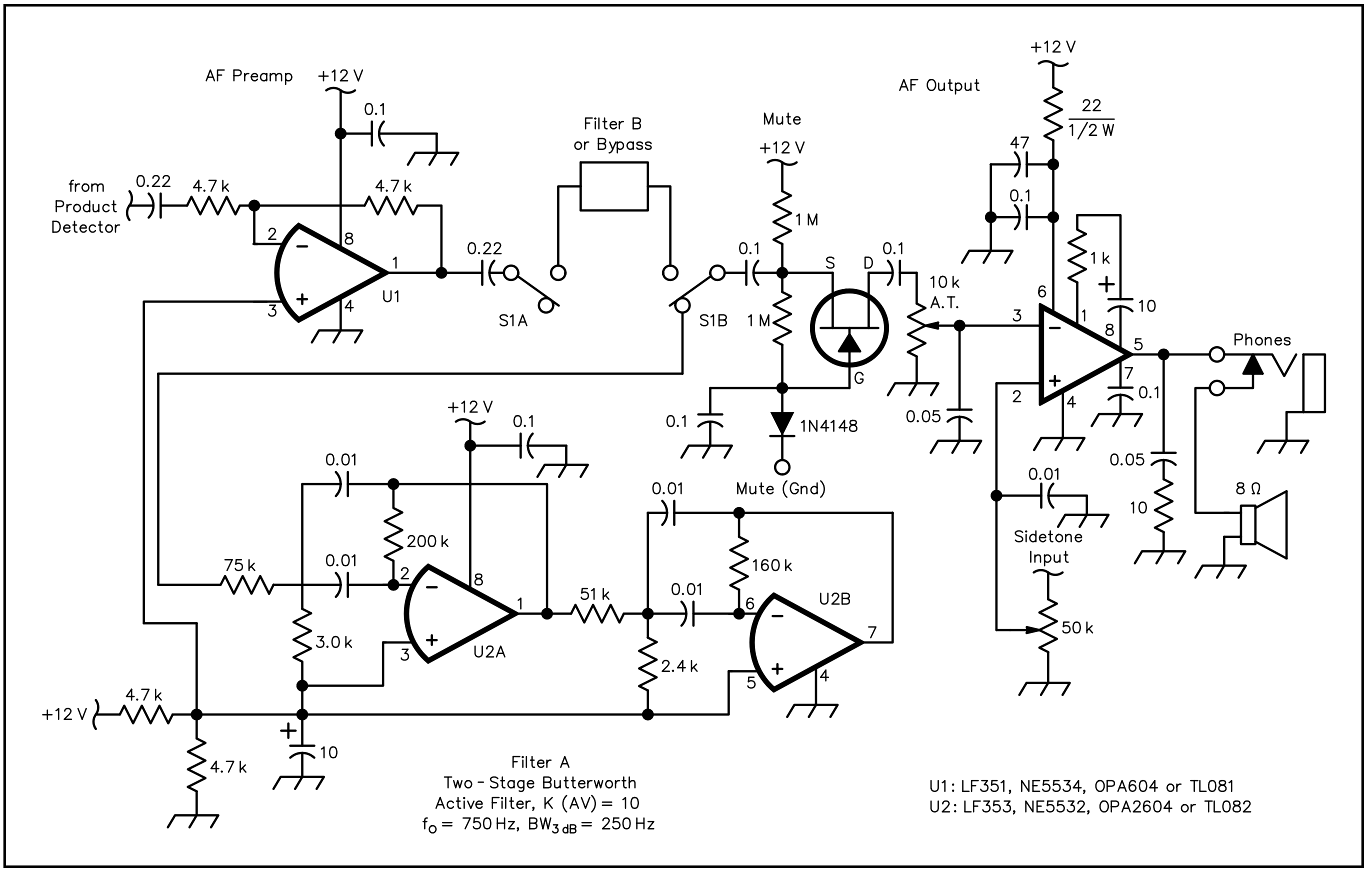
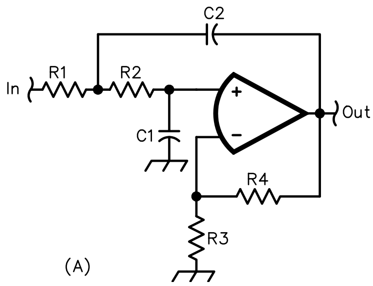
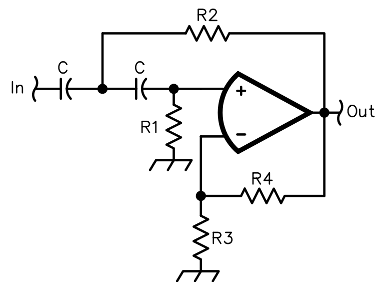


Fig 16.55—Typical application of a two-stage active filter in the audio chain of a QRP CW tranceiver. The filter can be bypassed, or another filter can be switched in by S1.



(A)



Low-Pass Filter

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

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

$$R_2 = \frac{1}{C_1C_2R_1\omega_c{}^2}$$

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

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

where
- K = gain
- f_c = –3 dB cutoff point
- $\omega_c = 2\pi f_c$
- a = 1.414 (see table, one stage)
- C_2 = a standard value near $10/f_c$ µF

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

Example:
- K = 2
- f = 2700 Hz
- ω_c = 16,965
- C_2 = 0.0033 µF
- C1 ≤ 0.00495 µF (use 0.0047 µF)
- R1 ≤ 18,147 Ω (use 18 kΩ)
- R2 = 12,293 Ω (use 12 kΩ)
- R3 = 60,880 Ω (use 62 kΩ)
- R4 = 60,880 Ω (use 62 kΩ)

High-Pass Filter

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

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

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

$$R_4 = KR_1$$

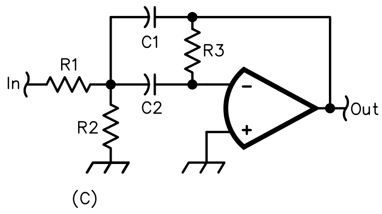
where
- K = gain
- f_c = –3 dB cutoff point
- $\omega_c = 2\pi f_c$
- a = 0.765 (see table, first of two stages)
- C = a standard value near $10/f_c$ µF

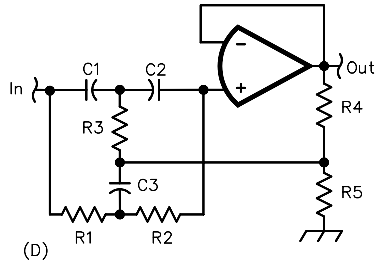
Note: For unity gain, short R4 and omit R3.

Example:
- K = 4
- f = 250 Hz
- ω_c = 1570.8
- C = 0.04 µF (use 0.039 µF)
- R1 = 11,407 Ω (use 11 kΩ)
- R2 = 23,358 Ω (use 24 kΩ)
- R3 = 15,210 Ω (use 15 kΩ)
- R4 = 45,630 Ω (use 47 kΩ)

[Continued on next page.]

Fig 16.54—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.





Band-Pass Filter

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

Then $R1 = \dfrac{Q}{K_o \omega_o C}$

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

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

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 1000 Ω)
R3 = 90.4 kΩ (use 91 kΩ or 100 kΩ)

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 7.5 kΩ)
R >> 375 Ω (1 kΩ)
R4 = 25 Ω (use 27Ω)
R5 = 975 Ω (use 1 kΩ)

Fig 16.54—Continued from previous page.

Factor "a" for Low- and High-Pass Filters

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

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

CRYSTAL-FILTER EVALUATION

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

A template, containing additional information, is available from the *ARRLWeb*, **http://www.arrl.org/notes**.

An Overview

The basic setup is shown in **Fig 16.56A**. The VCO is primarily a conventional LC-tuned Hartley oscillator with its frequency tuned over a small range by a varactor diode (MV2104 in part B of the figure). Other varactors may be used as long as the capacitance specifications aren't too different. Change the 5-pF coupling capacitor to expand the sweep width if desired.

The VCO signal goes through a buffer amplifier to the filter under test. The filter is followed by a wide-bandwidth amplifier and then a detector. The output of the detector is a rectified and filtered signal. This

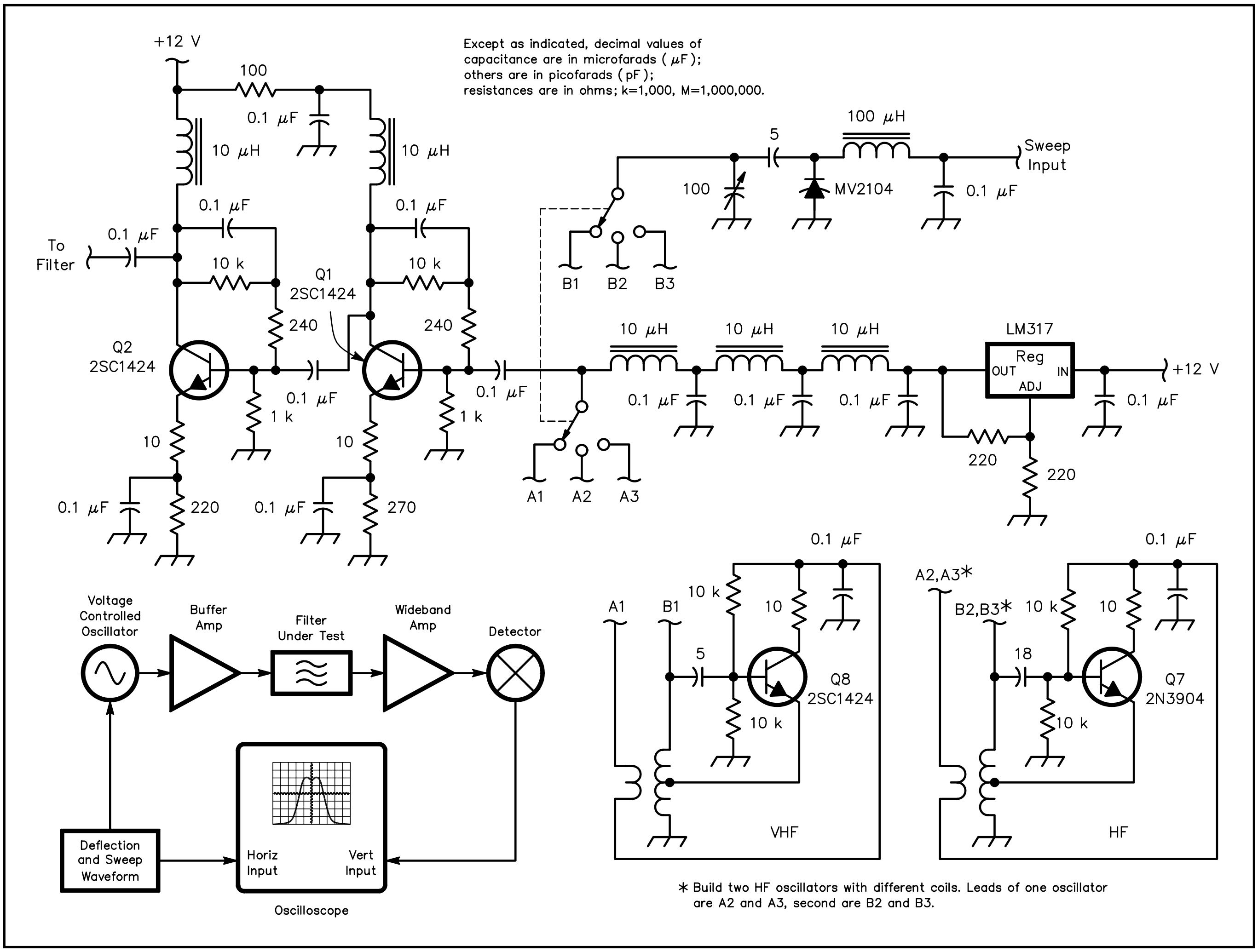


Fig 16.56—The test set block diagram, lower left, starts with a swept frequency oscillator, shown in the schematic. If a commercial swept-frequency oscillator is available, it can be substituted for the circuit shown.

varying dc voltage drives the vertical input of an oscilloscope. At any particular time, the deflection and sweep circuitry commands the VCO to "run at this frequency." The same deflection voltage causes the oscilloscope beam to deflect left or right to a position corresponding to the frequency.

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

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

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

The oscillator and buffer stage operate at low power levels to minimize frequency drift caused by component heating. Crystal filters cause large load changes as the frequency is swept in and out of the pass-band. These large changes in impedance tend to "pull" the oscillator frequency and cause inaccuracies in the passband shape depicted by the oscilloscope. Therefore a buffer amplifier is a necessity. The wide-band amplifier in **Fig 16.57** is derived from one in ARRL's *Solid State Design for the Radio Amateur*.

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

The horizontal-deflection sweep circuit uses a dual op-amp IC (see **Fig 16.58**). One section is an oscillator; the other is an integrator. The integrator output changes linearly with time, giving a uniform brightness level as the trace is moved from side to side. Increasing C1 decreases the sweep rate. Increasing C2 decreases the slope of the output waveform ramp.

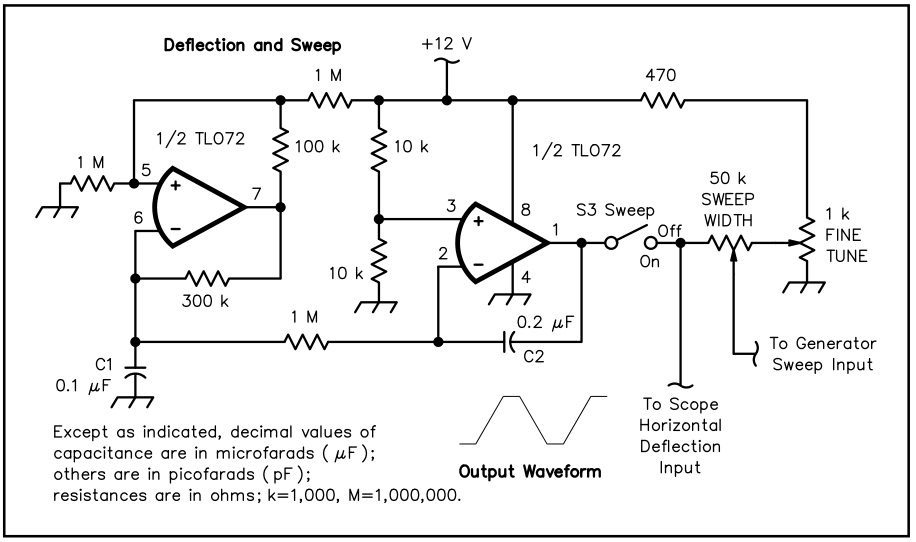


Fig 16.58—The sweep generator provides both an up and down sweep voltage (see text) for the swept frequency generator and the scope horizontal channel.

Operation

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

S3 in Fig 16.58 disables the

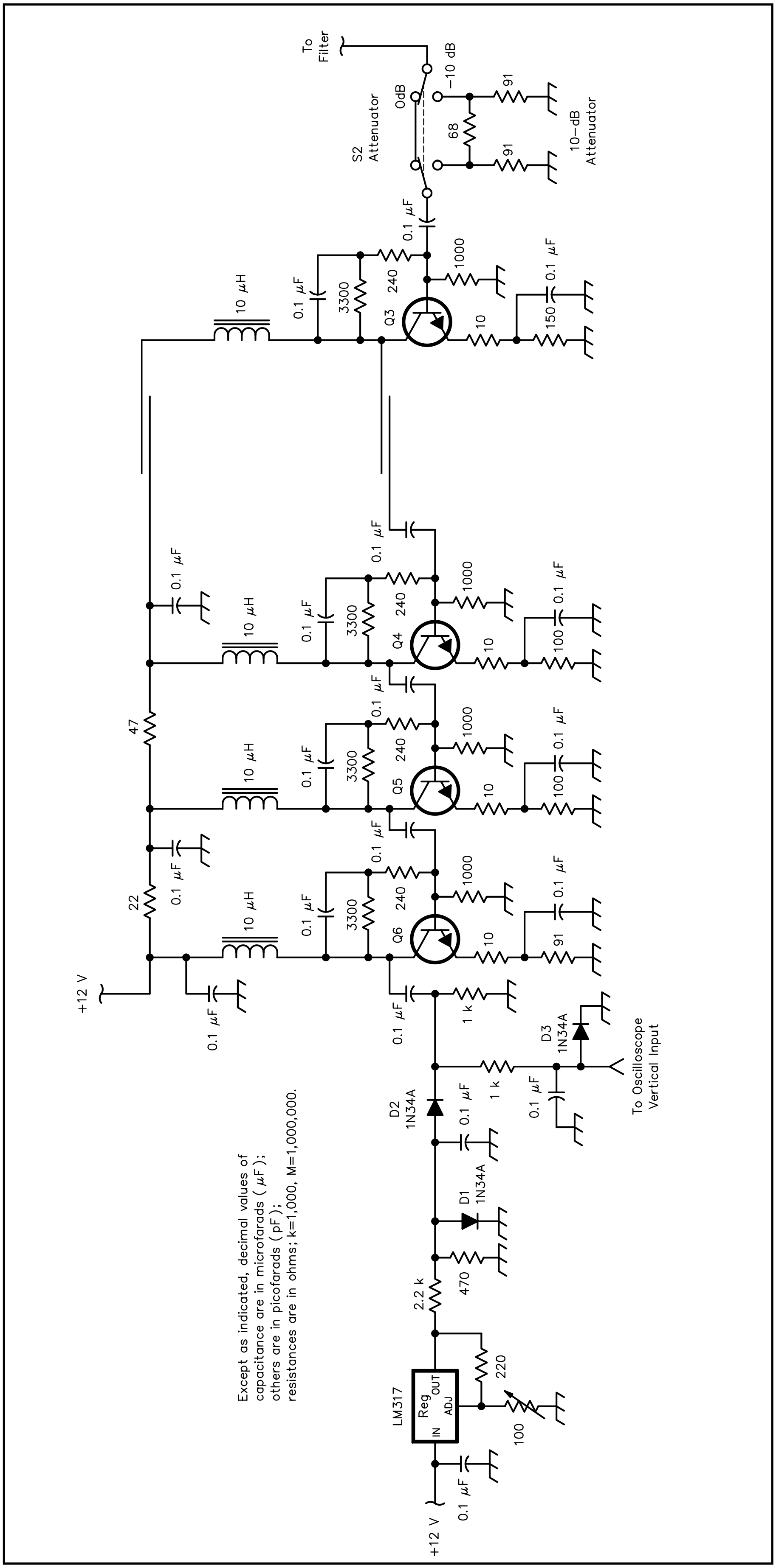


Fig 16.57—The filter under test is connected to Q3 on the right side of the schematic. The detector output, on the left side, connects to the oscilloscope vertical input. A separate voltage regulator, an LM 317, is used to power this circuit. Q3, Q4, Q5 and Q6 are 25C1424 or 2N2857.

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

Table 16.5
VCO Coils

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

The two larger coils are wound on $^{3}/_{4}$-inch PVC pipe and the smaller one on a $^{1}/_{2}$-inch drill bit. Tuning coverage for each oscillator is obtained by squeezing or spreading the turns before gluing them in place. The output windings connected to A1, A2 and A3 are each single turns of #14 wire spaced off the end of the tapped coils.

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

The buffer amplifier shown here is set up to drive a 50-Ω load and the wide-bandwidth amplifier input impedance is about 50 Ω. If the filter is not a 50-Ω unit, however, various methods can be used to accommodate the difference.

References

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

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

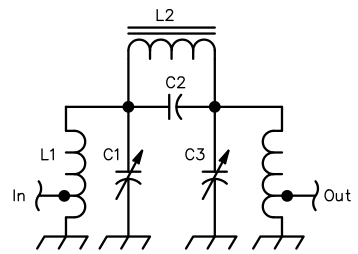
BAND-PASS FILTERS FOR 144 OR 222 MHz

Spectral purity is necessary during transmitting. Tight filtering in a receiving system ensures the rejection of out-of-band signals. Unwanted signals that lead to receiver overload and increased intermodulation-distortion (IMD) products result in annoying in-band "birdies." One solution is the double-tuned band-pass filters shown in **Fig 16.64.** They were designed by Paul Drexler, WB3JYO. Each includes a resonant trap coupled between the resonators to provide increased rejection of undesired frequencies.

Many popular VHF conversion schemes use a 28-MHz intermediate frequency (IF), yet proper filtering of the image frequency is often overlooked in amateur designs. The low-side injection frequency used in 144-MHz mixing schemes is 116 MHz and the image frequency, 88 MHz, falls in TV channel 6. Inadequate rejection of a broadcast carrier at this frequency results in a strong, wideband signal at the low end of the 2-m band. A similar problem on the transmit side can cause TVI. These band-pass filters have effectively suppressed undesired mixing products. See **Fig 16.65** and **16.66**.

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

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



Component Values

	144 MHz	*220 MHz*
C2	1 pF	1 pF
C1, C3	1-7 pf piston	1-7 pF piston
L2	27t no. 26 enam on T37-10	15t no. 24 enam on T44-10
L1, L3	7t no. 18, 1/4-in ID, tap 1 1/2t	4t no. 18, 1/4-in ID, tap 1 1/2t

Fig 16.64—Schematic of the band-pass filter. Components must be chosen to work with the power level of the transmitter.

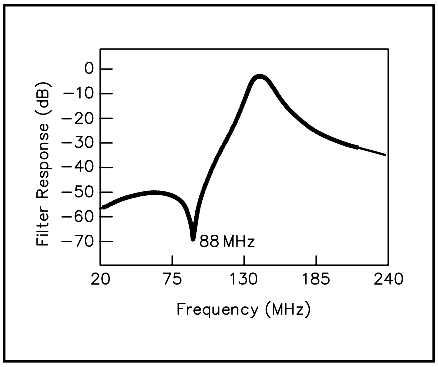


Fig 16.65—Filter response plot of the 144-MHz band-pass filter, with an image-reject notch for a 28 MHz IF.

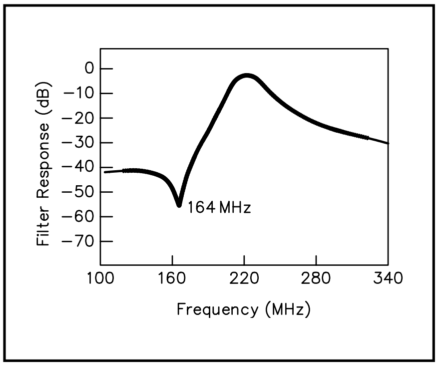


Fig 16.66—Filter response plot of the 222-MHz band-pass filter, with an image-reject notch for a 28 MHz IF.

Switched Capacitor Filters

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

A CONTINUOUSLY VARIABLE BANDWIDTH AUDIO FILTER

Active audio filters are a very popular way to enhance the selectivity, especially on CW, of many rigs. Passive LC filters, usually based on 88-mH toroids, were used for many years, followed by active filters built with one or more opamps. Switched capacitor filters, such as this one designed by Denton Bramwell, K7OWJ, offer additional performance, with fewer chips. This project was first described in the July 1995 issue of *QST.*

General Description

Maxim[1] markets a series of useful filter (and other) chips that exhibit excellent performance. The filter described here rolls off at an impressive 96 dB per octave. Best of all, the upper cutoff frequency *can be continuously varied* to accommodate any reasonable desired bandwidth! The values given here provide a 3-dB bandwidth ranging from 450 Hz for CW to 2700 Hz for voice. The filter's bandwidth is determined by merely adjusting a potentiometer.

Maxim's switched-capacitor-chip family provides Bessel, Butterworth and elliptic low-pass designs. This Butterworth version has a flat passband—which is excellent if you're cascading stages—and is based on the MAX 295 chip.

These ICs are exceptionally easy to apply (see **Table 1**). If you want but one stage (48 dB per octave) and you don't need to vary the filter's bandwidth, all that's required are ±5-V supplies and a single capacitor. Working with an oscillator internal to the IC, the capacitor sets the clock rate. Attach input and output lines, add a few inexpensive components, a single-voltage power supply and you will have a working one-stage filter.

Table 1
Selected Maxim Filter ICs

Part Number	*Filter Type*	*Unit Price ($US)*	*Rolloff Characteristics*	*Notes*
MAX291 MAX295	8th-order Butterworth	6	48 dB/octave; about 110 dB ultimate	Maximum flatness in the passband; excellent ultimate rejection
MAX293 MAX297	8th-order elliptic	6	–80 dB at 1.5 × the corner frequency; about 80 dB ultimate	Probably an excellent choice for a single-IC, fixed-frequency filter. A MAX297 can directly replace a MAX295, if steeper rolloff is needed and more in-band ripple is acceptable.

Note: For single-chip, fixed-frequency filters using the MAX295 and MAX297, the corner frequency can be set by a single capacitor connected between pin 1 and ground. The internal clock frequency, F_I, is determined by the value of the clock capacitor, C, according to the formula: $fI_{(kHz)} = 10^5 \div 3C_{(pF)}$. The corner frequency equals $^1/_{50}$ of the clock frequency.

The ’295 will not operate on voltages much higher than ±5 (split supply) or +10 V (a single supply). A potential of 12 V between the supply leads is the rated absolute maximum, so if you try to use this voltage, you may well pop the chip.

Circuit Description

Fig 16.67 shows the incoming audio from the receiver passing through a simple RC high-pass filter, and on to U1. U1’s output is coupled through another simple RC high-pass filter to the input of U2. U2’s output is filtered by two more stages of active RC high-passing using 741 op amps (U3 and U4). This fixes a –3 dB point about 300 Hz on the low-frequency side of the filter and a completely adjustable –3 dB point on the high side.

The bypassing shown uses 220-µF capacitors for decoupling the stages. Good decoupling is absolutely necessary to achieve good ultimate rejection. (Dual and quad op amps can’t be decoupled from each other because they all use the same supply!) The 220-µF capacitors have a reactance of 2.9 Ω at 250 Hz. You might get by with smaller capacitance values.

Up to 5 mV of clock feedthrough can appear with the audio at the output of the ’295. Maxim provides an op amp within the ’295 for RC active filtering to suppress clock leakage. This is accomplished by C7, C8, R7, R8 and R9, which are chosen for a corner frequency of 3.4 kHz.

For continuously variable bandwidth service, it’s necessary to use an external clock to feed the MAX295s (U1 and U2). A 555 timer (U5) is configured as an astable multivibrator. The output frequency is controlled by a potentiometer (R20). The corner frequency of the ’295 is $^{1}/_{50}$ of the clock frequency. With the components shown, U5 delivers 37 kHz to 180 kHz, for a corner frequency ranging from 740 to 3,600 Hz. The 555 output frequency range will change with supply voltage. Be prepared to adjust the values if you use a supply other than ± 5 V. The resistance of most potentiometers is usually not tightly controlled, so the actual clock range you obtain may be a bit different from the values above

The supply shown for this filter provides ±5-V supplies—the negative supply must deliver about 25 mA. If you want to operate this filter from a single 12-V supply, you can create a virtual ground by stacking two Zener diodes (see the inset of Fig 16.67). With this approach (rather than using separate positive and negative supplies, *be sure that the common (ground) line of the filter’s PC board never meets the ground for the rest of your station!* The filter’s ground must float. If you enclose your filter in a plastic box as I did (and the jacks are thus mounted in plastic), it’s okay for the incoming audio and the 12-V supply to share the same return (C1 provides dc decoupling). Since the output is a jack to a set of headphones, the actual ground is not important—unless you add an outboard audio amplifier. Then you must add an audio transformer for isolation.

Construction

Layout is generally non-critical. The author built his filter on a general-purpose prototyping board using point-to-point wiring, but PC boards are available.[2]

In Use

With S1 off, the filter receives no power and the headphones are connected directly to the receiver output. With S1 on, power is applied to the filter and the headphones are connected to the filter output. Because the filter has unity gain, the input and output audio levels are equal, so there’s no need to adjust your receiver’s audio gain control when switching the filter in and out of the line.

Summary

To test the filter on-the-air, tune to a busy region of a band and listen as the interfering signals drop into oblivion as you rotate R20, the BANDWIDTH control. With an attenuation of 96 dB per octave, only

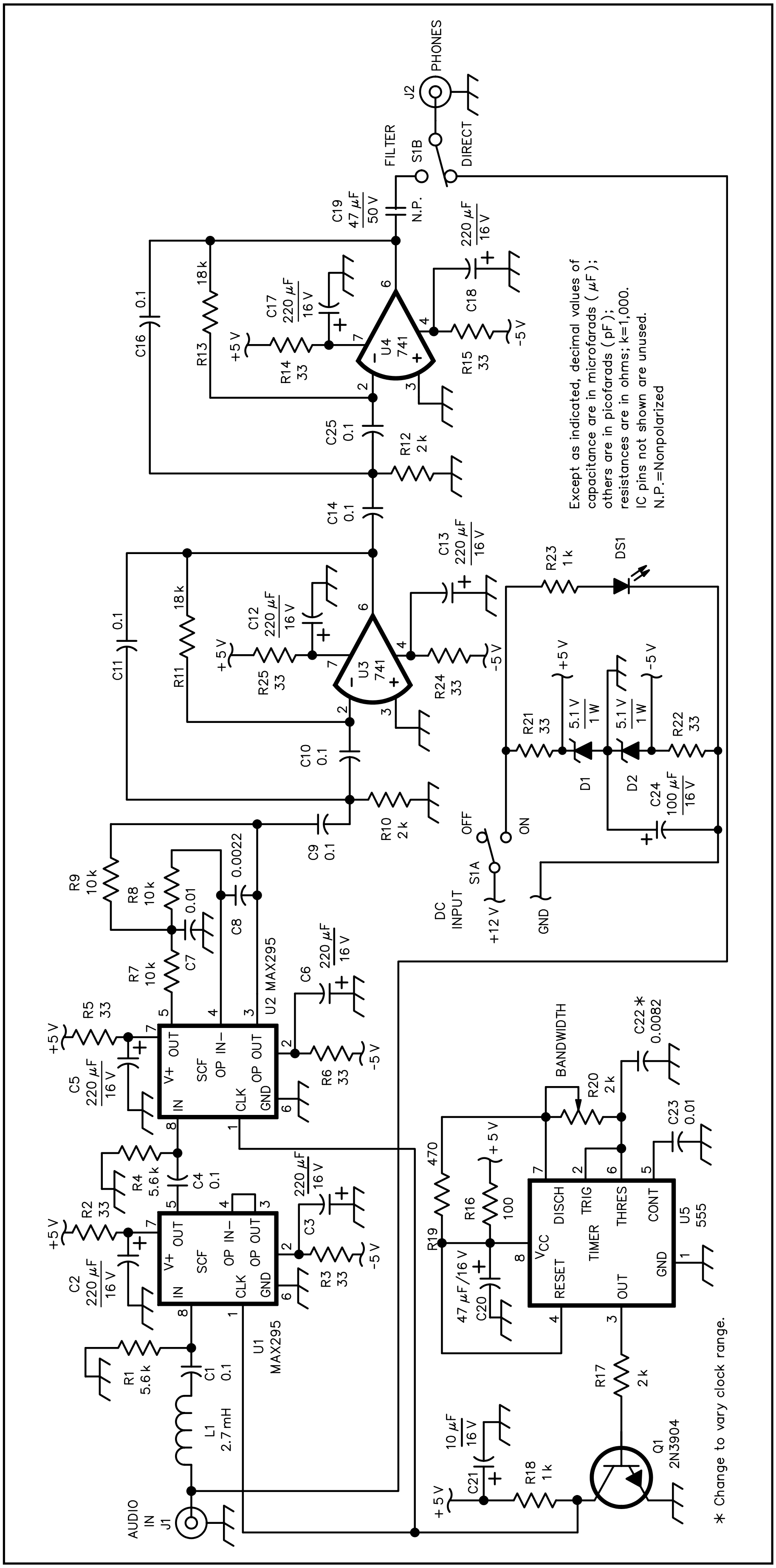


Fig 16.67—Schematic diagram of the continuously variable audio filter. Unless otherwise specified, resistors are 1/4-W, 5%-tolerance carbon-composition or film units. Equivalent parts can be substituted.

C2, C3, C5, C6, C12, C13, C17, C18—220 µF, 16 V electrolytic
C19—47 µF, 50 V nonpolarized
C20—47 µF, 16 V electrolytic or tantalum
C21—10 µF, 16 V electrolytic or tantalum
C24—100 µF, 16 V electrolytic or tantalum
D1, D2—5.1-V, 1-W Zener diode (1N4733)
DS1—LED
J1, J2—Phono jack
L1—2.7 mH (optional)
Q1—2N3904
R20—2-kΩ linear-taper potentiometer
S1—DPDT toggle
U1, U2—MAX295
U3, U4—741 op amp
U5—555 timer

a small adjustment of the potentiometer moves a signal from very bothersome to below the noise. By itself, the audio filter provides skirts as deep and as steep as better-quality IF filters.

The unit's distributed high-pass filtering makes a quite noticeable reduction in low-frequency grunts and rumbles. It also colors the atmospheric noise a bit. With the pot set for maximum width, you'll hear the apparent pitch of the background noise shift slightly upward when you turn on the filter. This shift represents noise you no longer have to cope with when copying a station.

Notes

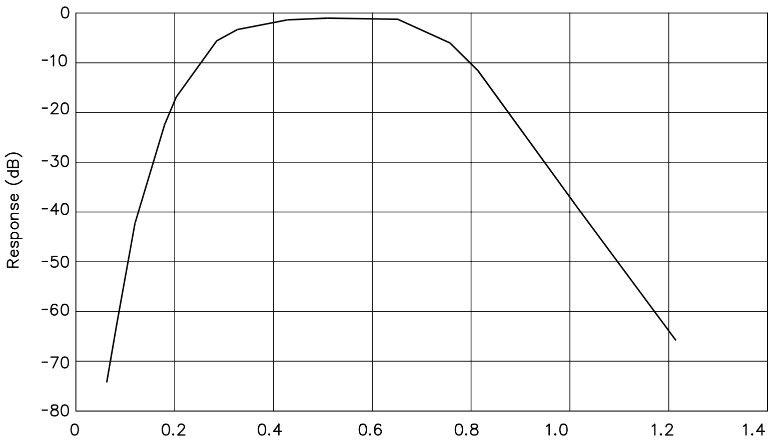
[1] Maxim ICs are available from several sources including Digi-Key Corp. See the **References** chapter.

[2] PC boards are available from FAR Circuits. The price: $6 plus $1.50 shipping. FAR Circuits is listed in the **References** chapter Address List. A PC-board template package is available on *ARRLWeb*.

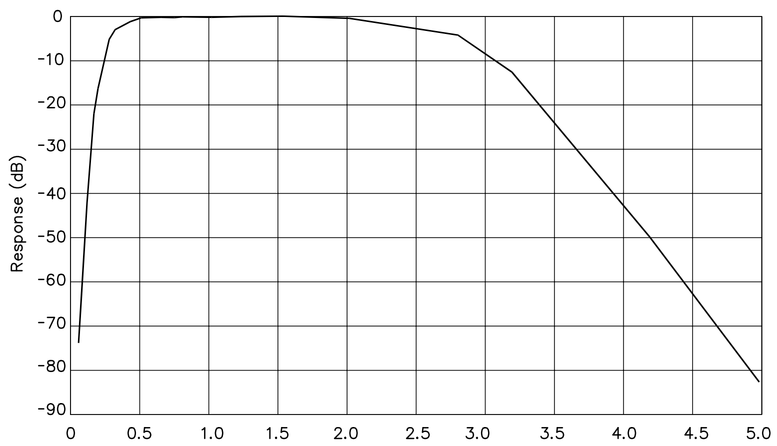
Filter Response with the Clock Set at 37.6 kHz (A) and 149.4 kHz (B)

With the clock frequency set for 37.6 kHz, curve (A) shows a 10 dB down response of roughly 260 Hz to 800 Hz, and 60 dB down response of about 100 Hz to 1180 Hz. You can calibrate the front panel control to whatever bandwidth you like—3 dB down, 6 dB down, or 60 dB down.

Curve (B), taken with the clock set to 149.4 kHz, shows about 250 Hz to 3050 Hz at 10 dB down and about 100 to 4400 Hz at 60 dB down.



(A)



(B)

A BC-BAND ENERGY-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 16.68**. The standard capacitor values listed are within 2.8% of the design values. If the attenuation peaks (f2, f4 and f6) do not fall at 0.677, 1.293 and 1.111 MHz, tune the series-resonant circuits by slightly squeezing or separating the inductor windings.

Construction of the filter is shown in **Fig 16.69**. Use Panasonic NP0 ceramic disk capacitors (ECC series, class 1) or equivalent for values between 10 and 270 pF. For values between 330 pF and 0.033 µF, use Panasonic P-series polypropylene (type ECQ-P) capacitors. These capacitors are available through Digi-Key (see the Address List in the **References**

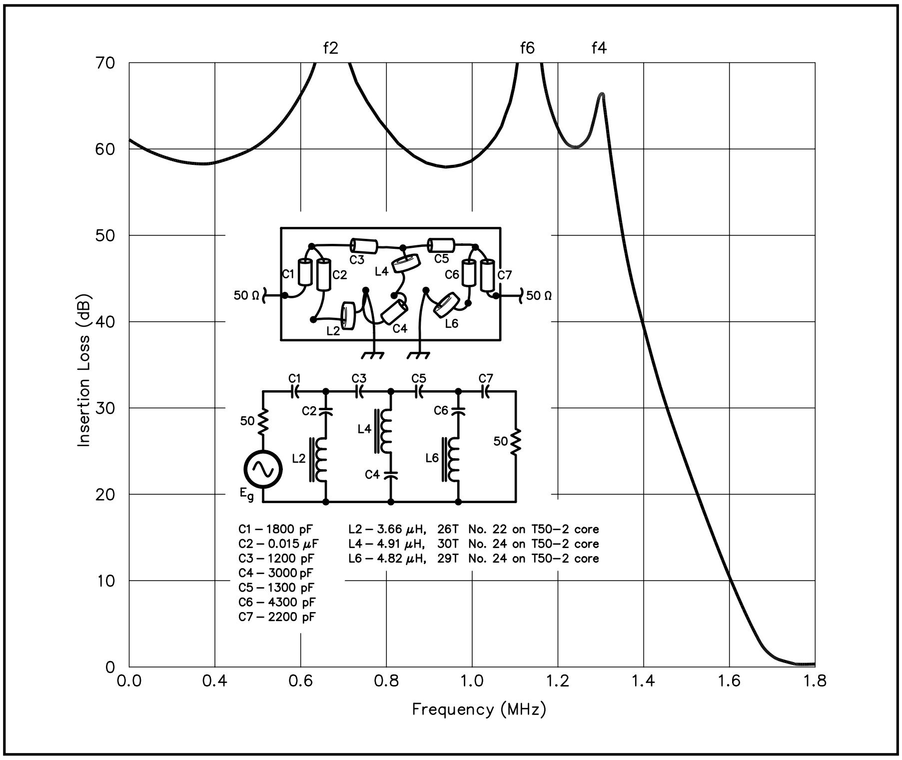


Fig 16.68—Schematic, layout and response curve of the broadcast band rejection filter.

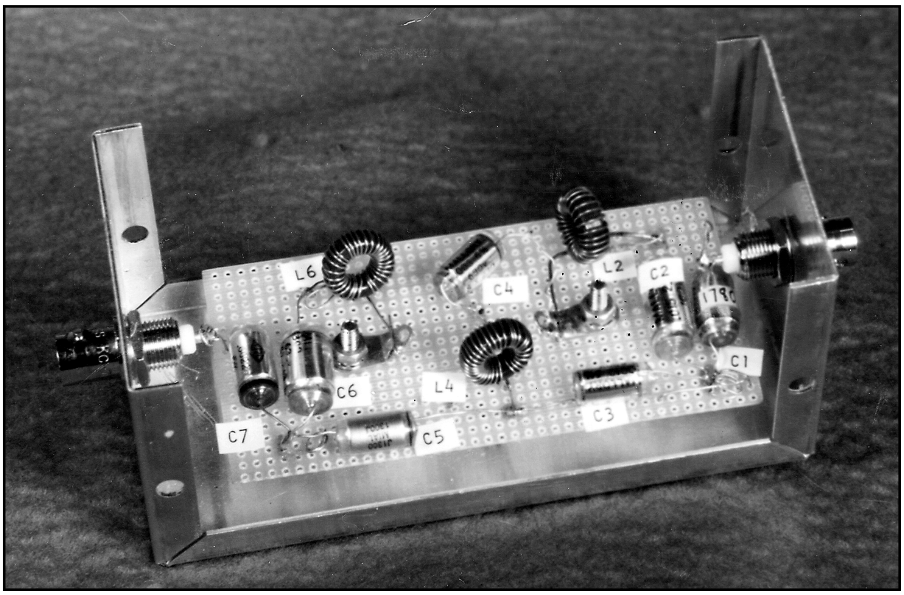


Fig 16.69—The filter fits easily in a 2 × 2 × 5-inch enclosure. The version in the photo was built on a piece of perfboard.

chapter) and other suppliers. The powdered-iron T-50-2 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 16.68.) For the 80-m version, L2 through L6 should be 20 to 25 turns each, wound on T-50-6 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 16.68. 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).

SECOND-HARMONIC-OPTIMIZED (CWAZ) LOW-PASS FILTERS

The FCC requires transmitter spurious outputs below 30 MHz to be attenuated by 40 dB or more for power levels between 5 and 500 W. For power levels greater than 5 W, the typical second-harmonic attenuation (40-dB) of a seven-element Chebyshev low-pass filter (LPF) is marginal. An additional 10 dB of attenuation is needed to assure compliance with the FCC requirement.

Jim Tonne, WB6BLD, solved the problem of significantly increasing the second-harmonic attenuation of the seven-element Chebyshev LPF while maintaining an acceptable return loss (> 20 dB) over the amateur passband. Jim's idea was presented in February 1999 *QST* by Ed Wetherhold, W3NQN. These filters are most useful with single-band, single device transmitters. Common medium-power multiband transceivers use push-pull power amplifiers because such amplifiers inherently suppress the second harmonic.

Tonne modified a seven-element Chebyshev standard-value capacitor (SVC) LPF to obtain an additional 10 dB of stop-band loss at the second-harmonic frequency. He did this by adding a capacitor across the center inductor to form a resonant circuit. Unfortunately, return loss (RL) decreased to an unacceptable level, less than 12.5 dB. He needed a way to add the resonant circuit, while maintaining an acceptable RL level over the passband.

The typical LPF, and the Chebyshev SVC designs listed in this chapter all have acceptable RL levels that extend from the filter ripple-cutoff frequency down to dc. For many Amateur Radio application, we need an acceptable RL only over the amateur band for which the LPF is designed. We can trade RL levels below the amateur band for improved RL in the passband, and simultaneously increase the stop-band loss at the second-harmonic frequency.

THE CWAZ LOW-PASS FILTER

This new eight-element LPF has a topology similar to that of the seven-element Chebyshev LPF, with two exceptions: The center inductor is resonated at the second harmonic in the filter stop band, and the component values are adjusted to maintain a more than acceptable RL across the amateur passband. To distinguish this new LPF from the SVC Chebyshev LPF, Wetherhold named it the "Chebyshev with Added Zero" or "CWAZ" LPF design.

You should understand that CWAZ LPFs are *output filters for single-band transmitters*. They provide optimum second and higher harmonic attenuation while maintaining a suitable level of return loss over the amateur band for which they're designed.

Fig 16.70 shows a schematic diagram of a CWAZ LPF design. Table 16.6 lists suggested capacitor and inductor values for all amateur bands from 160 through 10 meters. If you want to calculate CWAZ values for different bands, simply divide the first-row C and L values (for 1 MHz) by the start frequency of the desired band. For example, C1, 7 for the 160-meter design is equal to 2986/1.80 = 1659 pF. The other component values for the 160-meter LPF are calculated in a similar manner.

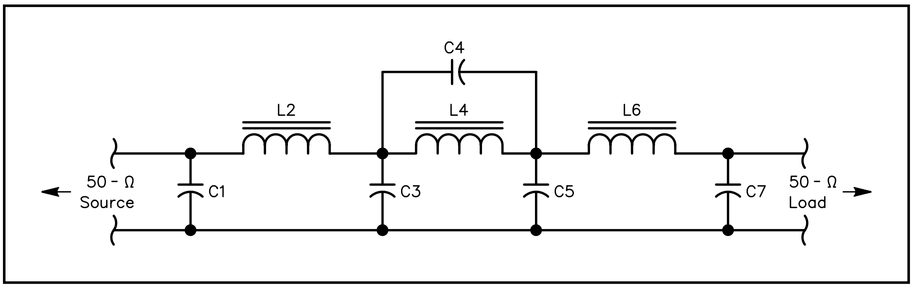


Fig 16.70—Schematic diagram of a CWAZ low-pass filter designed for maximum second-harmonic attenuation. See Table 16.6 for component values of CWAZ 50-Ω designs. L4 and C4 are tuned to resonate at the F4 frequency given in Table 16.6. For an output power of 10 W into a 50-Ω load, the RMS output voltage is $\sqrt{10 \times 50}$ = 22.4 V. Consequently, a 100 V dc capacitor derated to 60 V (for RF filtering) is adequate for use in these LPFs if the load SWR is less than 2.5:1.

Table 16.6

CWAZ 50-Ω Low-Pass Filters

Designed for second-harmonic attenuation in amateur bands below 30 MHz.

Band (m)	*Start Frequency (MHz)*	*C1,7 (pF)*	*C3,5 (pF)*	*C4 (pF)*	*L2,6 (µH)*	*L4 (µH)*	*F4 (MHz)*
—	1.00	2986	4556	680.1	9.377	8.516	2.091
		1659	2531	378			3.76
160	1.80	1450 + 220	2100 + 470		5.21	4.73	
		1500 + 150	2200 + 330	330 + 47			3.78
		853	1302	194			7.32
80	3.50		1150 + 150		2.68	2.43	
		470 + 390	1200 + 100	150 + 47			7.27
		427	651	97.2			14.6
40	7.00				1.34	1.22	
		330 + 100	330 + 330	100			14.4
		296	451	67.3			21.1
30	10.1				0.928	0.843	
		150 + 150	470	68			21.0
		213	325	48.6			29.3
20	14.0				0.670	0.608	
		220	330	47			29.8
		165	252	37.6			37.8
17	18.068				0.519	0.471	
		82 + 82	100 + 150	39			37.1
		142	217	32.4			43.9
15	21.0				0.447	0.406	
		150	220	33			43.5
		120	183	27.3			52.0
12	24.89				0.377	0.342	
		120	180	27			52.4
		107	163	24.3			58.5
10	28.0				0.335	0.304	
		100	82 + 82	27			55.6

NOTE: The CWAZ low-pass filters are designed for a single amateur band to provide more than 50 dB attenuation to the second harmonic of the fundamental frequency and to the higher harmonics. All component values for any particular band are calculated by dividing the 1-MHz values in the first row (included for reference only) by the start frequency of the selected band. The upper capacitor values in each row show the calculated design values obtained by dividing the 1-MHz capacitor values by the amateur-band start frequency in megahertz. The lower standard-capacitor values are suggested as a convenient way to realize the design values. The middle capacitor values in the 160 and 80-meter-band designs are suggested values when the high-value capacitors (greater than 1000 pF) are on the low side of their tolerance range. The design F4 frequency (see upper value in the F4 column) is calculated by multiplying the 1-MHz F4 value by the start frequency of the band. The lower number in the F4 column is the F4 frequency based on the suggested lower capacitor value and the listed L4 value.

CWAZ VERSUS SEVENTH-ORDER SVC

The easiest way to demonstrate the superiority of a CWAZ LPF over the Chebyshev LPF is to compare the RL and insertion-loss responses of these two designs. **Fig 16.71** shows a 20-meter SVC Chebyshev LPF design based on the tables in this chapter. **Fig 16.72** shows the computer-calculated return- and insertion-loss responses of the LPF shown in Fig 16.71. The plotted responses were made using Jim Tonne's *ELSIE* filter design and analysis software. This DOS-based program is available from Trinity Software.[1]

Fig 16.73 (on next page) shows the computer-calculated return- and insertion-loss responses of a CWAZ LPF intended to replace the seven-element 20-meter Chebyshev SVC LPF. The stop-band attenuation of the CWAZ LPF in the second-harmonic band is more than 60 dB and is substantially greater than that of the Chebyshev LPF. Also, the pass-band RL of the CWAZ LPF is quite satisfactory, at more than 25 dB. The disadvantages of the CWAZ design are that an extra capacitor is needed across L4, and several of the designs listed in Table 16.6 require paralleled capacitors to realize the design values. Nevertheless, these disadvantages are minor in comparison to the increased second-harmonic stop-band attenuation that is possible with a CWAZ design.

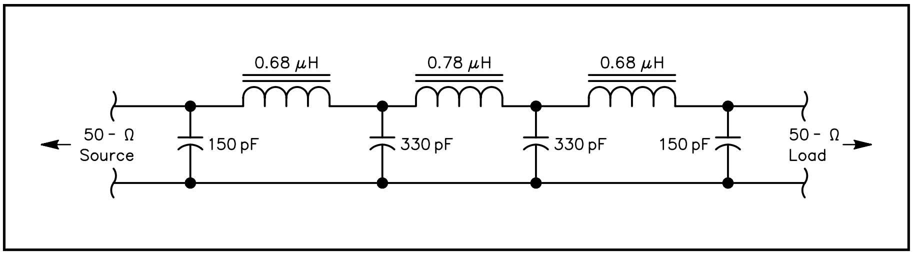


Fig 16.71—Schematic diagram of a 20-meter SVC Chebyshev LPF.

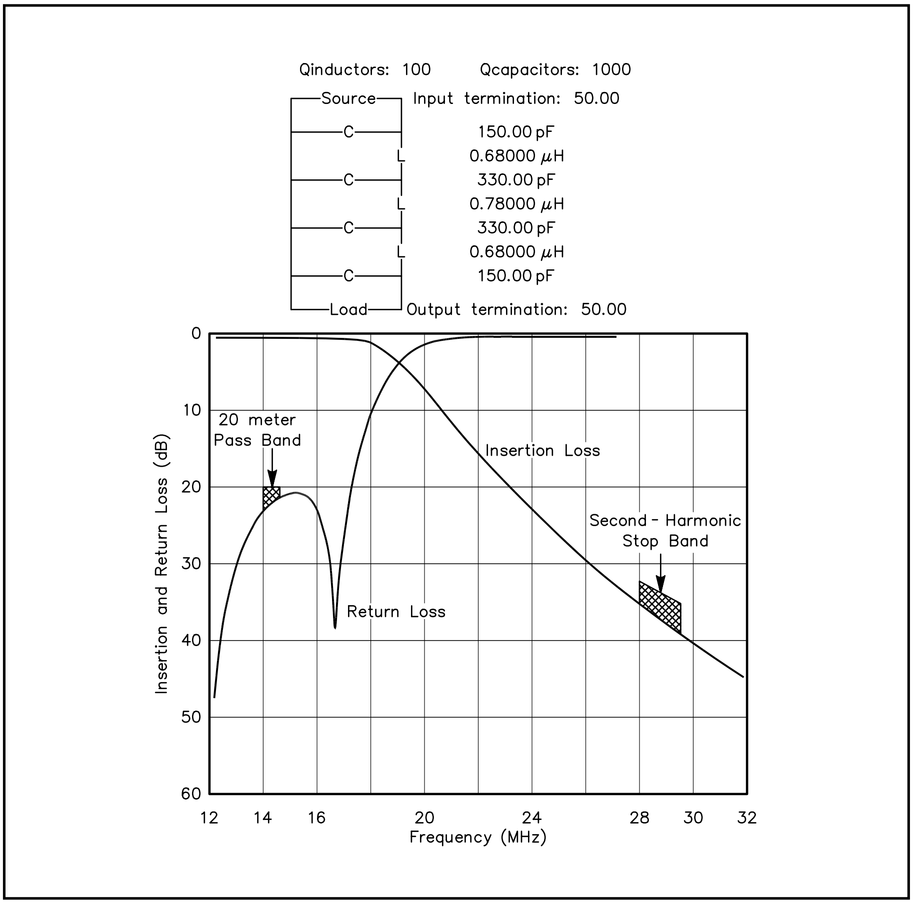


Fig 16.72—The plots show the *ELSIE* computer-calculated return- and insertion-loss responses of the seventh-order Chebyshev SVC low-pass filter shown in Fig 16.71. The 20-meter passband RL is about 21 dB, and the insertion loss over the second-harmonic frequency band ranges from 35 to 39 dB. A listing of the component values is included.

[1] Those seriously interested in passive LC filter design may experience the capabilities of *ELSIE* through a demo disk. The demo is restricted to LC filters of the third-order. Contact Trinity Software (see the **References** chapter Address List for contact information).

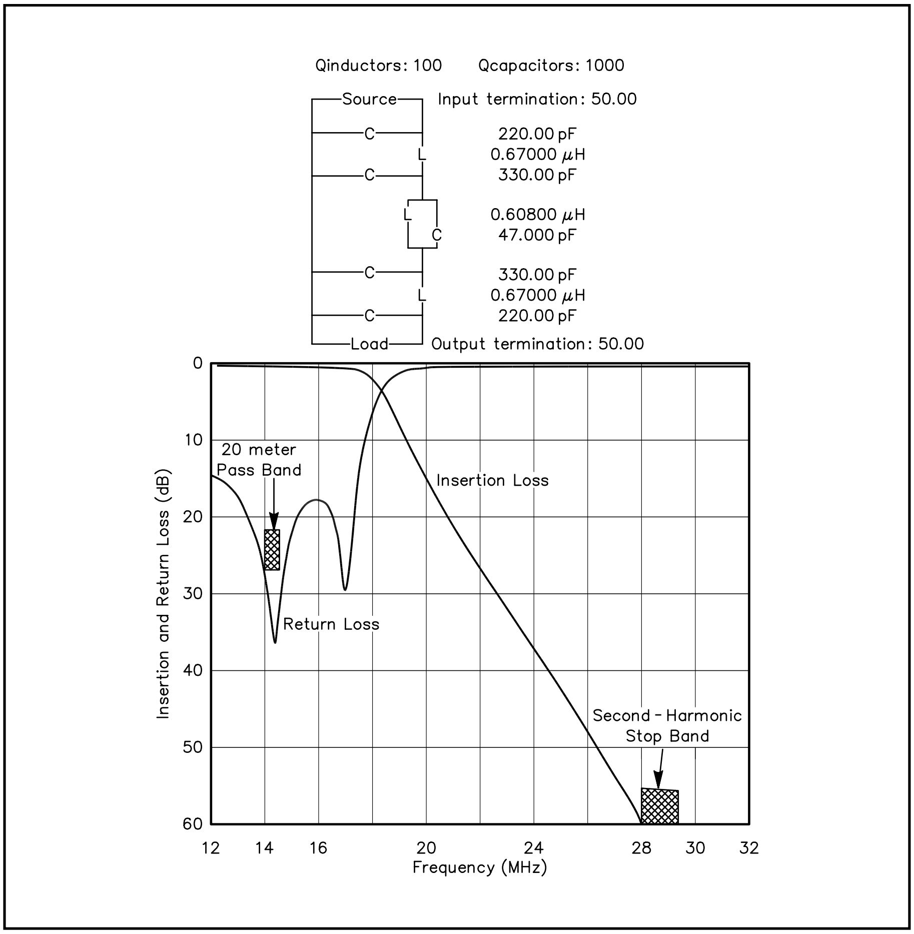


Fig 16.73—The plots show the *ELSIE* computer-calculated return- and insertion-loss responses of the eight-element low-pass filter using the CWAZ capacitor and inductor values listed in Table 16.6 for the 20-meter low-pass filter. Notice that the calculated attenuation to second-harmonic signals is greater than 60 dB, while RL over the 20-meter passband is greater than 25 dB.

THE DIPLEXER FILTER

This section, covering diplexer filters, was written by William E. Sabin, W0IYH. The diplexer is helpful in certain applications, and Chapter 15 shows them used as frequency mixer terminations.

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

Fig 16.74 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 16.75 shows the LP and HP responses of a diplexer filter for the 80 meter band. The following items are to be noted:

- The 3 dB responses of the LP and HP meet at 5.45 MHz.
- The input impedance is close to 50 Ω at all frequencies, as indicated by the high value of return loss (SWR <1.07:1).
- At and near 5.45 MHz, the LP input reactance and the HP input reactance are conjugates; therefore they cancel and produce an almost perfect 50-Ω input resistance in that region.

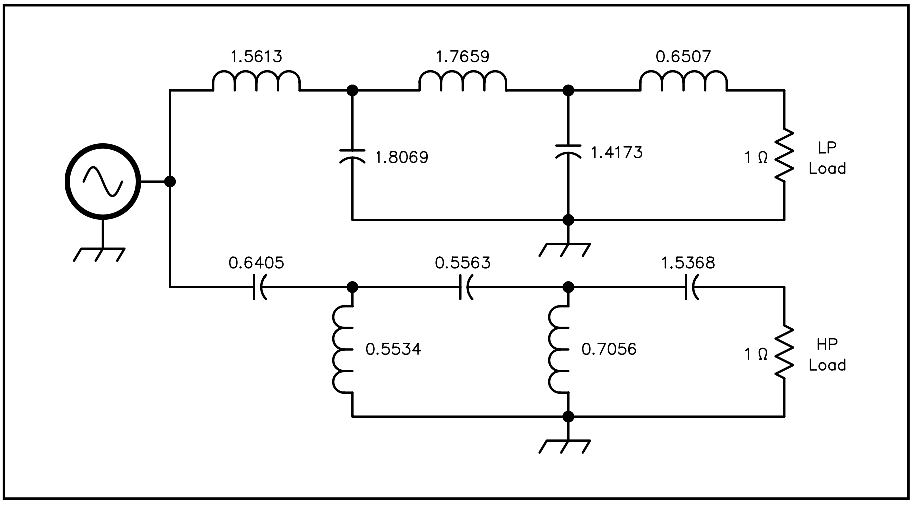


Fig 16.74—Low-pass and high-pass prototype diplexer filter design. The low-pass portion is at the top, and the high-pass at the bottom of the drawing. See text.

- Because of the way that the diplexer filter is derived from synthesis procedures, the transfer characteristic of the filter is pretty much independent of the actual value of the amplifier's dynamic output impedance.[2] This is a useful feature, since the RF power amplifier's 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.

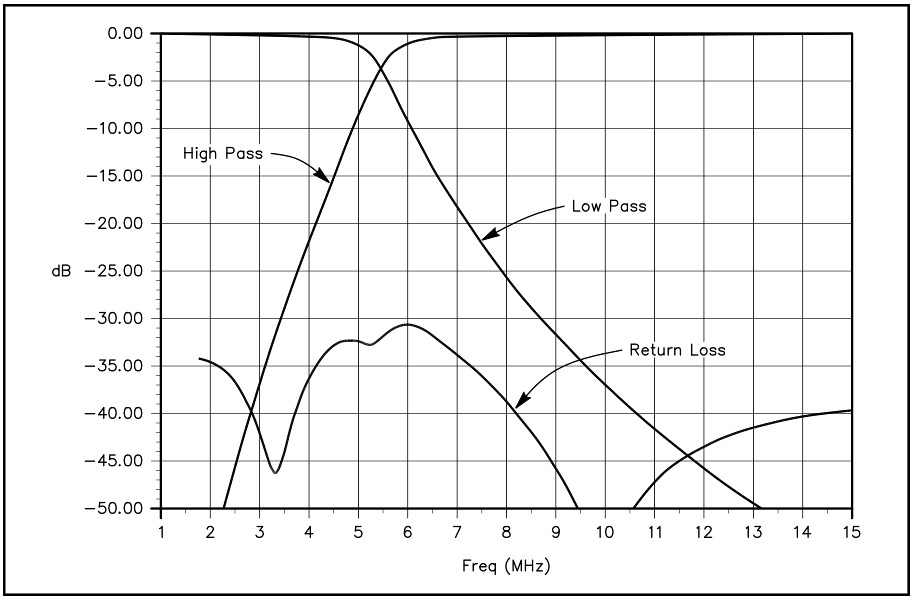


Fig 16.75—Response for the low-pass and high-pass portions of the 80-meter diplexer filter. Also shown is the return loss of the filter.

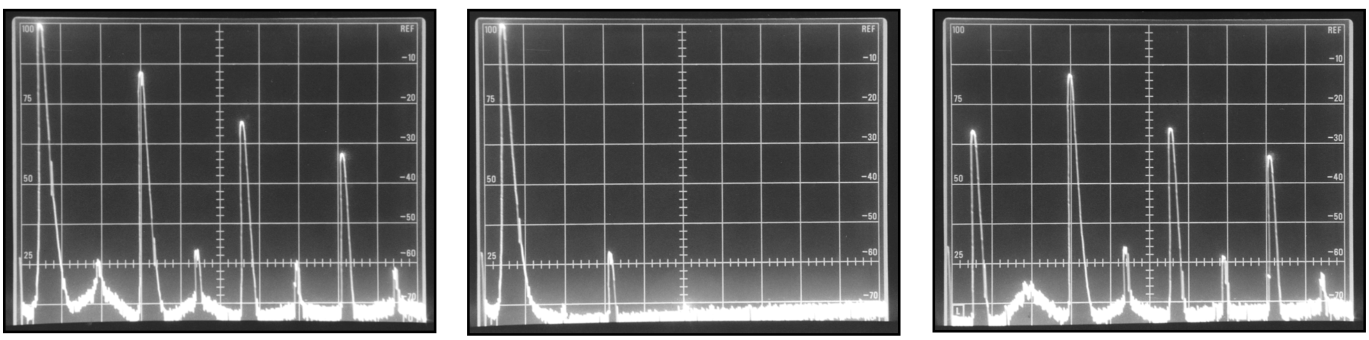
Fig 16.76A shows the unfiltered of a solid-state push-pull power amplifier for the 80-meter band. In the figure you can see that:

- The second harmonic has been suppressed by a proper push-pull design.
- The third harmonic is typically only 15 dB or less below the fundamental.

The amplifier output goes through our diplexer filter. The desired output comes from the LP side, and is shown in Fig 16.76B. In it we see that:

- The fundamental is attenuated only about 0.2 dB.
- The LP has some harmonic content; however, the levels exceed FCC requirements for a 100-W amplifier.

Fig 16.76C shows the HP output of the diplexer which 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 almost perfectly matches the 50-Ω amplifier output impedance. This is because power that would otherwise be *reflected* back to the amplifier is absorbed in the dump resistor.



(A) **(B)** **(C)**

Fig 16.76—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.

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

- The reactance of the low-pass filter outside the desired passband. This is especially true for transistors that are designed for high-frequency operation.
- Self resonance of a series inductor at some high frequency.
- A stopband 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 16.75, 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] Check *ARRLWeb* at: **http://www.arrl.org/qex/**.

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 stopband frequencies is desirable for some reason. The *ARRL Radio Designer* program is an ideal way to finalize the design before the actual construction.[7] The coil dimensions and the dump resistor wattage need to be determined from a consideration of the power levels involved, as illustrated in Fig 16.76.

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]

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 stopband 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. This publication is available from ARRL as Order no. 7253. It can be ordered at: **http://www.arrl.org/catalog**. 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. W0IYH, "Diplexer Filters for the HF MOSFET Power Amplifier," *QEX*, Jul/Aug, 1999. Also check *ARRLWeb* at: **http://www.arrl.org/qex/**.

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

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

OTHER FILTER PROJECTS

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

- Differential-mode high-pass filter for 75-Ω coax (for TV reception)
- *Brute-force* ac-line filter
- Loudspeaker common-mode choke
- LC filter for speaker leads
- Audio equipment input filter

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9. M. Dishal, "Top Coupled Band-pass Filters," *IT&T Handbook*, 4th edition, American Book, Inc, 1956, p 216. Also, P. Geffe, *Simplified Modern Filter Design*, J. F. Rider, 1963, pp 42-48.
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