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# Crystal Parameter Measurements Simplified

*The author describes a procedure to make very accurate measurements on quartz crystals. You can do this with a simple fixture using four resistors, a capacitor and some RF connectors.*

This is a technique I derived in 1961 for measuring crystal parameters in a laboratory as an undergraduate student. Fifty years later as radio amateurs, we have much better equipment available on our workbench to do this.

Besides the fixture, the additional equipment needed consists of:

- A digital RF signal generator.
- A frequency counter.
- An RF voltmeter or RF probe.

## Crystal Parameters

The quartz crystal unit, in an HC-49U package, consists of a circular quartz disc with aluminum or gold plating on opposite surfaces. The crystal is mounted vertically inside the case. It is held by two supports on the edges of the crystal. Two leads exit the base to secure the crystal in a circuit. Figure 1 is a photo of the internal structure of an HC-49U crystal unit on the left, and the unit in the case on the right.

The quartz crystal unit is electrically represented by a series resistor, R, an inductor, L, and a capacitor, C. A parallel capacitor, $C_0$, is needed because of the plating and the leads. Figure 2 is the equivalent circuit diagram.

The parameters R, L, and C are referred to in technical publications and books as $R_m$, $L_m$, and $C_m$. The inductance, capacitance and resistance are referred to as the motional parameters of the quartz crystal, thus the subscript *m*.

## Derivation of the Resonant Frequency Formulas

The admittance, $Y_{AB}$, between the terminals A and B in the schematic of Figure 2 is given by Equation 1.

$$Y_{AB} = \frac{1}{Z_{AB}} = \frac{1}{R + j(\omega L - 1/\omega C)} + j\omega C_0 \quad \text{[Eq 1]}$$

where $\omega = 2\pi f$.

By combining the two terms on the right, we get Equation 2.

$$Y_{AB} = \frac{(1 - \omega^2 LC_0 + C_0/C) + j\omega RC_0}{R + j(\omega L - 1/\omega C)} \quad \text{[Eq 2]}$$

and inverting both sides gives us Equation 3.

$$Z_{AB} = \frac{R + j(\omega L - 1/\omega C)}{(1 - \omega^2 LC_0 + C_0/C) + j\omega RC_0} \quad \text{[Eq 3]}$$

Multiplying both the numerator and the denominator by the complex conjugate of the denominator gives us Equation 4.


**Figure 1 — The internal structure of a quartz crystal unit is shown on the left. The complete package in the metal HC-49U case is shown on the right.**

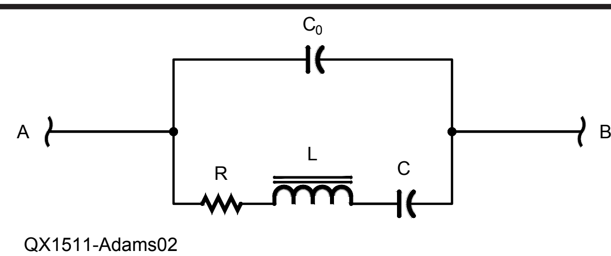


**Figure 2 — This is an equivalent circuit for a quartz crystal unit.**

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$$Z_{AB} = \frac{Real + j(\omega L - 1/\omega C - \omega^3 L^2 C_0 + 2\omega L C_0 / C - C_0/\omega C^2 - \omega R^2 C_0)}{C_0^2/C^2 + 1 - 2\omega^2 L C_0 + 2C_0/C + \omega^2 R^2 C_0^2 - 2\omega L C_0^2/C + \omega^4 L^2 C_0^2} \quad \text{[Eq 4]}$$

where *Real* is a real number with too many terms to fit on the line. We are not going to use it anyway.

At resonance, the complex component of the above equation is zero. That is the term following the *j*. We can simplify Equation 4 by expressing the complex component as Equation 5.

$$\omega L - 1/\omega C - \omega^3 L^2 C_0 + 2\omega L C_0/C - C_0/\omega C^2 - \omega R^2 C_0 = 0 \quad \text{[Eq 5]}$$

Multiplying both sides of the equation by $\omega C^2$ to remove the fractions gives us Equation 6.

$$\omega^2 L C^2 - C - \omega^4 L^2 C^2 C_0 + 2\omega^2 L C C_0 - C_0 - \omega^2 R^2 C^2 C_0 = 0 \quad \text{[Eq 6]}$$

The last term is much smaller than the other terms combined, so we eliminate it. The result is given in Equation 7.

$$\omega^4 L^2 C^2 C_0 - \omega^2 (LC^2 + 2LCC_0) + (C + C_0) = 0 \quad \text{[Eq 7]}$$

We can solve this equation by finding the roots of the quadratic equation with $\omega^2$ as the independent variable. There are any number of good mathematical software packages that can do this easily. *Wolfram Alpha* is a free online calculator.[1]

There are two resulting resonant frequencies. The series resonant frequency, $f_s$, is given by Equation 8.

$$f_s = \frac{1}{2\pi}\sqrt{\frac{1}{LC}} \quad \text{[Eq 8]}$$

The parallel resonant, or antiresonant frequency, $f_a$, is given by Equation 9.

$$f_a = \frac{1}{2\pi}\sqrt{\frac{1}{LC} + \frac{1}{LC_0}} \quad \text{[Eq 9]}$$

We can see that $f_a$ is always greater than $f_s$.

The crystal is always connected to an external circuit, and $C_0$ has additional capacitance in parallel with it. We will call that additional capacitance $C_p$. This will modify Equation 9, and the parallel resonant frequency will be given by Equation 10.

$$f_a = \frac{1}{2\pi}\sqrt{\frac{1}{LC} + \frac{1}{LC_t}} \quad \text{[Eq 10]}$$

where $C_t = C_0 + C_p$. Crystal manufacturers specify the resonant frequency of a crystal at this frequency with a particular load capacitance, $C_p$.

## Impedance at Resonant Frequencies

At the series resonant frequency, $f_s$, we get $\omega_s^2 = 1/LC$, and by plugging this expression into Equation 3 we get Equation 11.

$$Z_{AB} = \frac{R}{1 - (\omega R C_0)^2} \approx R \quad \text{[Eq 11]}$$

We use the approximation because $R$ is on the order of 10 to 100 Ω, and $\omega$ is on the order of $10^6$, but $C_0$ is just a few picofarads, and on the order of $10^{-12}$. This makes the term very small compared to 1.

For the the parallel or antiresonant frequency we have Equation 12.

$$\omega_a^2 = \frac{1}{LC} + \frac{1}{LC_t} \quad \text{[Eq 12]}$$

Substituting the $\omega^2$ value into Equation 3, and using the impedance of the capacitor, $X_C$, we obtain Equation 13.

$$Z_{AB} = \frac{1}{\omega^2 C_t^2 R} = \frac{X_c^2}{R} \quad \text{[Eq 13]}$$

The impedance for the parallel or antiresonant frequency is also pure resistance and much greater than the series resonant impedance, with a value typically between 100 kΩ and 1 MΩ.

In order to obtain $C_m$ and $L_m$, we need only to measure the series resonant frequency and the antiresonant frequency, and the capacitance, $C_0$. We then use the numbers in Equations 8 and 10 to solve for $L_m$ and $C_m$. We need a stable and accurate signal generator, an accurate and precise frequency counter and a fairly sensitive RF voltmeter or RF probe. The frequency counter should be able to measure and display frequencies to within 1 Hz. The frequency counter may be built into the signal generator.

The output level from the test fixture at the parallel resonant point is going to be down as much as 110 dB from the peak voltage. This makes this measurement very difficult. Let's find an easier way.

## Crystal in Series With A Capacitor

Let's examine a crystal in series with a capacitor. Figure 3 shows the schematic for this model.

The impedance between terminals A and B of the circuit is given by Equation 14.

$$Z_{AB} = \frac{R + j(\omega L - 1/\omega C)}{1 - \omega^2 L C_0 + C_0/C + j\omega R C_0} + \frac{1}{j\omega C_X} \quad \text{[Eq 14]}$$

We wade through some lengthy arithmetic to find the two resonant frequencies. This is more tedious than the previous derivation, resulting in an expression with more than 20 terms. I will not bore you with the details and leave it as an exercise, if you are interested in a challenge. The two resulting resonant frequencies are given by Equations 15 and 16.

$$\omega_c = \sqrt{\frac{1}{LC} + \frac{1}{LC_t}} \quad \text{[Eq 15]}$$

$$\omega_a = \sqrt{\frac{1}{LC} + \frac{1}{LC_0}} \quad \text{[Eq 16]}$$

where $C_t = C_0 + C_X$ and $\omega = 2\pi f$.

We have shifted the previous series resonant point, now represented as $\omega_c$, up in frequency. The antiresonant frequency remains exactly the same.

We now use Equations 8 and 15 to determine $L_m$ and $C_m$ of the crystal. This is a system of two equations with three unknowns. We measure $C_0$ directly with an L/C meter. Take Equation 8 and rewrite it as Equation 17.

$$\omega_s^2 = \frac{1}{LC} \quad \text{[Eq 17]}$$

[1]Notes appear on page 26

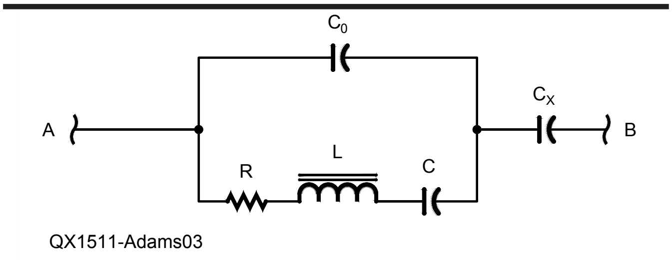


**Figure 3 — This schematic diagram is the model circuit for a crystal in series with capacitor $C_X$.**

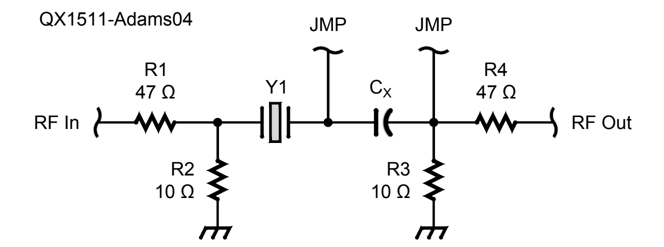


**Figure 4 — Here is the schematic diagram for the author's crystal test fixture. R1 = R4 = 47 Ω, R2 = R3 = 10 Ω, Y1 is the crystal under test, and $C_X$ = 47 pF. JMP is a jumper to short out $C_X$.**

We will also rewrite Equation 15 as Equation 18.

$$\omega_c^2 = \frac{1}{LC} + \frac{1}{LC_t} \quad \text{[Eq 18]}$$

The subscript c indicates that this is a measurement with $C_X$ in series with the crystal.

Subtracting Equation 17 from Equation 18 gives us Equation 19.

$$\omega_c^2 - \omega_s^2 = \frac{1}{LC_t} \quad \text{[Eq 19]}$$

This now becomes Equation 20.

$$4\pi^2 (f_c^2 - f_s^2) = \frac{1}{LC_t} \quad \text{[Eq 20]}$$

At this point, everyone wants to make an approximation for the difference of the two squares. Let's use the equation $x^2 - y^2 = (x + y)(x - y)$ and get more precise results. This will give us Equation 21, solved for $L_m$.

$$L_m = \frac{1}{4\pi^2 (f_c + f_s)(f_c - f_s)(C_0 + C_X)} \quad \text{[Eq 21]}$$

We can measure the two resonant frequencies using a signal generator and frequency counter. Measure $C_0$ and $C_X$ using a capacitance meter, and then crunch the numbers.

## Test Fixture

In order to make the measurements we use a test fixture. Other test measurements in publications and on the Internet use more complex circuits. This test circuit is very simple. Figure 4 shows the schematic diagram for the circuit. You can see how simple and inexpensive it can be.

The input and output impedance of the fixture is close to 50 Ω, but is not critical. R2 and R3 should be kept small to reduce the loaded *Q* on the crystal, but not too small to attenuate the output RF voltage of the fixture to a very small value. The resonant frequencies are not affected by these values. If the values are large, the resonant peak spreads out and it is more difficult to home in on the exact peak. The small values of R2 and R3 also serve to swamp any effects of stray capacitance in the fixture.

Figure 5 is a photograph of the test fixture that I use for this procedure.

Here are the steps to measure the data needed.

1) With the RF generator voltage connected to the fixture, find the series resonant frequency with capacitor $C_X$ shorted. Start the frequency generator a few kilohertz below the marked frequency of the crystal and slowly increase the frequency while watching the RF output level. Write down the frequency at which the peak output voltage occurs. This is $f_s$.

2) Remove the short across capacitor $C_X$. Find the new output voltage peak at a slightly higher frequency. This will be $f_c$.

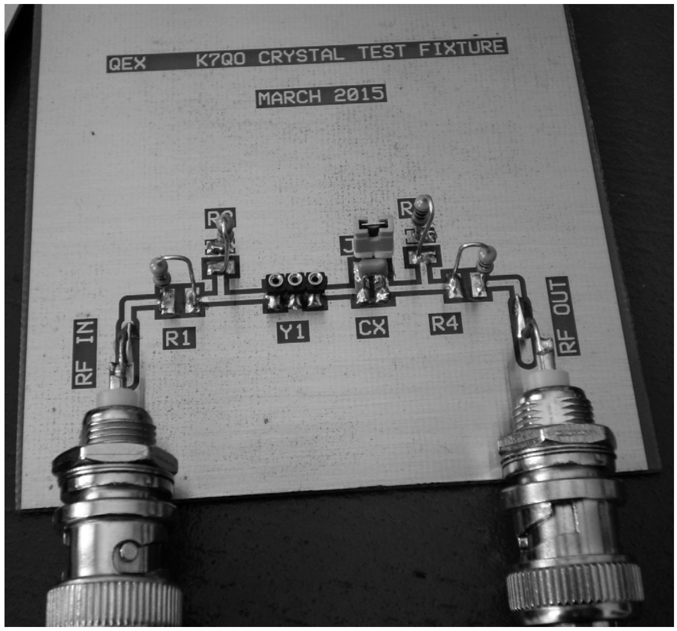


**Figure 5 — This photo shows the author's crystal test fixture. It was built using a circuit board layout with parts labeled. The center pin of the crystal socket is grounded to reduce the socket capacitance across the crystal.**

3) When you build the fixture measure $C_X$ before installing it and again in the circuit without a crystal in the socket to determine the extra stray capacitance caused by the shorting terminals. I used a 47 pF disc capacitor, but any value near this should do nicely. I recommend an NPØ capacitor.

4) Measure $C_0$ of the crystal using an accurate L/C meter, such as the Almost All Digital Electronics (AADE) L/C Meter II.[2]

Now you have all the data needed to calculate $L_m$. We obtain $C_m$ from the series resonant frequency, Equation 8, by using the $L_m$ value and $f_s$.

## Crystal Motional Resistance

Here are the steps to measure the motional resistance, $R_m$, or the effective series resistance (ESR) of the crystal.

1) Short $C_X$ and again find the series resonant frequency. Make note of the output voltage as accurately as possible.

2) Remove the crystal, leaving everything else as is.

3) Replace the crystal with a variable resistor of 100 Ω. I use a 25 turn variable resistor that has 0.2 inch lead spacing, to fit the crystal socket. Adjust the resistor for the exact same RF voltage output we had with the crystal in the socket.

4) Remove the variable resistor and measure the resistance. This is $R_m$.

Congratulations. You have all four crystal parameters. You have $L_m$, $C_m$, $R_m$, and $C_0$. These values may be used to determine the circuit for a crystal IF filter with a specific bandwidth.

You can determine the quality factor, $Q$, of the crystal by taking the series resonant frequency, $f_s$, motional inductance, $L_m$, and series resistance, $R_s$, and use Equation 22.

$$Q = \frac{X_L}{R_m} = \frac{\omega_s L_m}{R_m} = \frac{2\pi f_s L_m}{R_m} \qquad \text{[Eq 22]}$$

This is the inductive reactance at the resonant frequency divided by the motional resistance, $R_m$, of the crystal.

I have written some *Python* code to perform the calculations for the parameters of the crystal under test. This code carries out the computations to the full 64 bit precision of the computer processor. My code is available for download from the ARRL *QEX* files web page.[3]

## Procedure Verification

In order to verify that this procedure is both useful and accurate I picked at random nine crystals from my collection. I then sent these to Tom Thomson, WØIVJ, in Colorado to measure their characteristics by using an AIM Model 4170 Vector Network Analyzer. He also had Larry Benko, WØQE, do the same measurements with another 4170 VNA. Table 1 shows the results, with their measurements and mine. As you can see, the agreement on the crystal parameters is excellent.

## *SPICE* Simulation

As a check of all my theoretical work, I ran a *SPICE* simulation using *ngspice*. I set up an input RF voltage of 1.00 V and swept a crystal model from 4.190 MHz to 4.210 MHz. The voltage output was plotted in dB to show the null depth.

The important thing to note is that the null, corresponding to the parallel resonant mode, remains at the same frequency, but varies in magnitude. This agrees with the theoretical derivation and resulting formula.

## Matching Crystals

Using the technique discussed to match crystals is going to be a long and tedious task. One of the things that I want to demonstrate is how a Colpitts crystal oscillator can be used to match crystals and get excellent results.

Suppose you have a number of crystals and you are looking for four crystals for a four pole crystal filter. You want the crystals to match within 10 Hz of each other. Then, using the oscillator and a frequency counter you plug the crystals in and measure the output frequency of each, and keep them ordered. Also note the output voltage from the oscillator. If we have two crystals with the same frequency, we will take the one with the higher output from the oscillator because it will have the lowest $R_m$.

I have a few hundred 4.096 MHz crystals that I won at an auction on eBay. I found four of them that matched within 5 Hz of each other in the oscillator. I then used the procedure outlined in this paper to measure their crystal parameters. My results are given in Table 2.

Depending upon what program you use to generate the component values for your filters, you can match crystals using the Colpitts crystal oscillator, and then measure the parameters of just one crystal for use in the program. You could also measure a few of the matched crystals and average their parameter values to use in the program. Experimentation will determine which is the fastest method and just how well it meets your criteria for the resulting filter(s).

## Conclusion

You now know how to measure crystal parameters accurately and how to easily match a set of crystals for a filter. The test fixture is simple and easy to contruct using any of a number of building techniques. I hope that you will find this test fixture and procedure to be a useful addition to your workbench, and that it will simplify the construction of many successful projects.

*Chuck Adams, K7QO, was first licensed as KN5FJZ in the mid 1950s, during the greatest sunspot cycle in recorded history. He has held the calls K5FJZ, K5FO, and now K7QO. He is a retired professor of computer sciences, electrical engineering, and physics. He holds a PhD in physics, with a specialization in radiative transfer and electromagnetics. He now spends his time experimenting and building his own equipment. From time to time, he even gets on the air.*

## Notes

[1] *Wolfram Alpha* is a free online calculator that will calculate a large variety of quantities from your input values. Go to **www.wolframalpha.com**.

[2] The Almost All Digital Electronics website has had information about an array of kits, including the L/C Meter II at **www.aade.com**. [Unfortunately, when I checked this link prior to publication, the website home page has a note informing us that Neil Heckt passed away on August 19, 2015. The note further indicates that we should be patient while his family determines the future of the company. — *Ed.*]

[3] The author's *Python* code for computing the crystal parameters from the measured data is available for download from the ARRL *QEX* files web page. Go to **www.arrl.org/qexfiles** and look for the file **1x16_Adams.Zip**

**Figure 6 — SPICE simulation for sweeping a crystal. The left-most curve is with no series capacitor and then two more curves for two different values for $C_X$.**

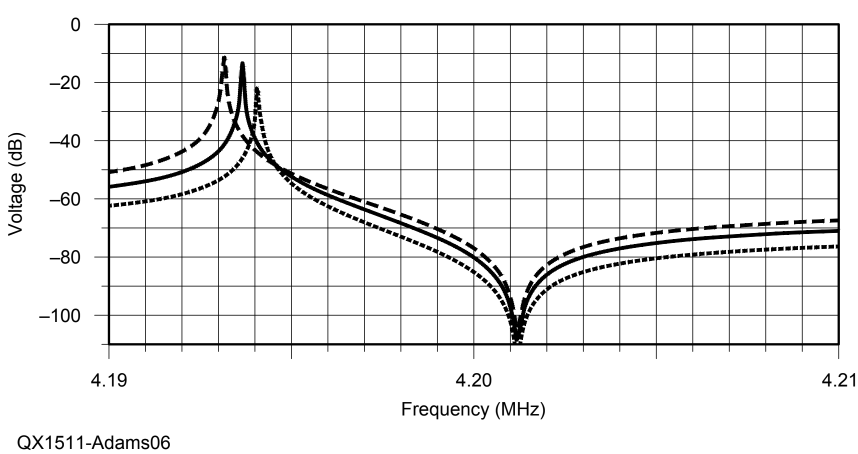

**Table 1**
**Crystal Measurement Procedure Verification**

| *Lab Tech* | *Crystal Number* | *$F_{Series}$* | *$F_{Parallel}$* | *$R_s$* | *$L_s$ (mH)* | *$C_s$ (pf)* | *$C_p$ (pf)* | *$Q_s$* | *Measuring Instrument* |
|---|---|---|---|---|---|---|---|---|---|
| WØIVJ | 1 | 3.578426 | 3.585154 | 49.822 | 139.418 | 0.0141885 | 3.780 | 65801 | AIM 4170 VNA |
| WØQE | 1 | 3.578427 | 3.585256 | 49.700 | 142.449 | 0.0138866 | 3.638 | 64443 | AIM 4170 VNA |
| K7QO | 1 | 3.578426 | -------- | 49.6 | 141.624 | 0.013968 | 3.65 | 64199 | K7QO Fixture |
| WØIVJ | 2 | 4.193154 | 4.200966 | 16.943 | 111.185 | 0.0129572 | 3.484 | 180122 | AIM 4170 VNA |
| WØQE | 2 | 4.193163 | 4.200894 | 17.159 | 115.719 | 0.0124495 | 3.376 | 177674 | AIM 4170 VNA |
| K7QO | 2 | 4.193159 | -------- | 15.4 | 114.234 | 0.012611 | 3.22 | 195432 | K7QO Fixture |
| WØIVJ | 3 | 4.031548 | 4.036547 | 40.962 | 309.640 | 0.0050332 | 2.032 | 211124 | AIM 4170 VNA |
| WØQE | 3 | 4.031552 | 4.036428 | 40.046 | 340.051 | 0.0045830 | 1.895 | 215101 | AIM 4170 VNA |
| K7QO | 3 | 4.031553 | -------- | 39.1 | 326.544 | 0.004773 | 1.57 | 211552 | K7QO Fixture |
| WØIVJ | 4 | 4.193152 | 4.201202 | 18.176 | 107.122 | 0.0134487 | 3.509 | 165524 | AIM 4170 VNA |
| WØQE | 4 | 4.193157 | 4.201100 | 18.432 | 112.633 | 0.0127907 | 3.376 | 160993 | AIM 4170 VNA |
| K7QO | 4 | 4.193154 | -------- | 17.5 | 112.514 | 0.012804 | 3.44 | 169390 | K7QO Fixture |
| WØIVJ | 5 | 4.094814 | 4.102963 | 23.238 | 134.404 | 0.0112398 | 2.830 | 147508 | AIM 4170 VNA |
| WØQE | 5 | 4.094819 | 4.103052 | 23.469 | 134.440 | 0.0112368 | 2.794 | 147386 | AIM 4170 VNA |
| K7QO | 5 | 4.094819 | -------- | 21.8 | 130.161 | 0.0130161 | 2.74 | 153616 | K7QO Fixture |
| WØIVJ | 6 | 3.998939 | 4.005005 | 22.484 | 132.716 | 0.0119351 | 3.940 | 153331 | AIM 4170 VNA |
| WØQE | 6 | 3.998953 | 4.005015 | 22.619 | 136.166 | 0.0116326 | 3.837 | 151261 | AIM 4170 VNA |
| K7QO | 6 | 3.998947 | -------- | 21.6 | 133.566 | 0.0133566 | 3.80 | 155370 | K7QO Fixture |
| WØIVJ | 7 | 11.055203 | 11.079818 | 7.407 | 11.640 | 0.0178059 | 4.007 | 109648 | AIM 4170 VNA |
| WØQE | 7 | 11.055211 | 11.079788 | 7.337 | 11.755 | 0.0176319 | 3.965 | 111282 | AIM 4170 VNA |
| K7QO | 7 | 11.055188 | -------- | 7.1 | 11.449 | 0.018102 | 3.99 | 112009 | K7QO Fixture |
| WØIVJ | 8 | 4.094873 | 4.102956 | 24.034 | 130.270 | 0.0115962 | 2.943 | 144651 | AIM 4170 VNA |
| WØQE | 8 | 4.094876 | 4.103023 | 24.476 | 134.745 | 0.0112110 | 2.818 | 141641 | AIM 4170 VNA |
| K7QO | 8 | 4.094880 | -------- | 24.1 | 132.374 | 0.011412 | 2.90 | 161414 | K7QO Fixture |
| WØIVJ | 9 | 13.499968 | 13.529170 | 4.063 | 5.074 | 0.0273900 | 6.345 | 100390 | AIM 4170 VNA |
| WØQE | 9 | 13.499973 | 13.529200 | 4.129 | 5.064 | 0.0274465 | 6.339 | 104041 | AIM 4170 VNA |
| K7QO | 9 | 13.499920 | -------- | 4.10 | 4.739 | 0.029329 | 6.10 | 98042 | K7QO Fixture |

| *Number* | *Form Factor* | *Crystal Identification Printed on Each Unit* |
|---|---|---|
| 1 | HC-49U | MPCO 3.579545 |
| 2 | HC-49U | HOSONIC 4.1943 B603 |
| 3 | HC-49S | 4.032 |
| 4 | HC-49U | HOSONIC 4.1943 B603 |
| 5 | HC-49U | MMD A18BA1 4.096JHz 9942G |
| 6 | HC-49U | ABRACON 4.000 AB 0443 |
| 7 | HC-49U | FOX115-20 11.0592 |
| 8 | HC-49U | MMD A18BA1 4.096MHz |
| 9 | HC-49U | 78941-1 13.500 KDS 5K |

**Table 2**
**Sample Crystal Measurements**

| *Crystal* | *$f_s$ (Hz)* | *$f_c$ (Hz)* | *$L_m$ (mH)* | *$C_m$ (fF)* | *$C_0$ (pF)* |
|---|---|---|---|---|---|
| 1 | 4094849 | 4095292 | 132.12 | 11.43 | 2.97 |
| 2 | 4094849 | 4095287 | 133.94 | 11.28 | 2.85 |
| 3 | 4094846 | 4095294 | 130.82 | 11.54 | 2.90 |
| 4 | 4094849 | 4095301 | 129.62 | 11.65 | 2.92 |

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**QST Issue:** Nov 2002
**Title:** A High Performance, Low Cost 1.8 to 54 MHz Low Pass Filter
**Author:** Bill Jones, K8CU

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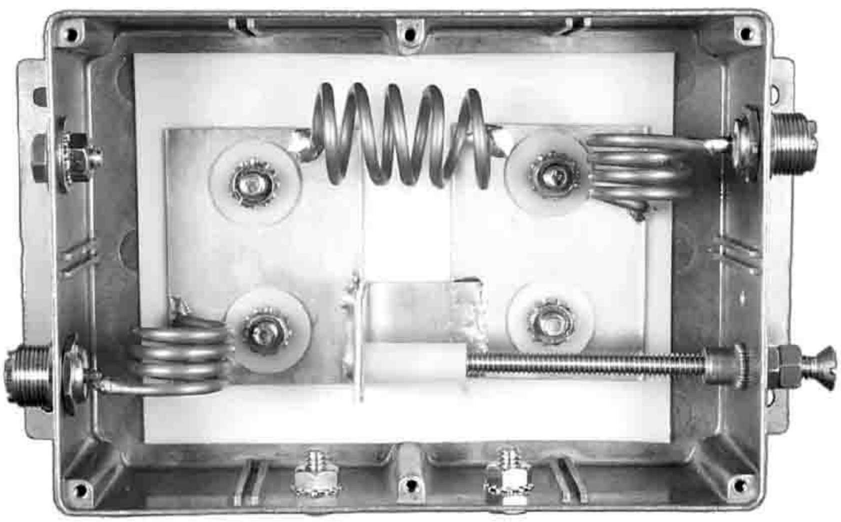
**By Bill Jones, K8CU**

# A High Performance, Low Cost 1.8 to 54 MHz Low Pass Filter

This high performance low pass filter, designed with software available to radio amateurs at no cost, is easily constructed with common hand tools.



This low pass filter design project for 6 meters started with goals of low insertion loss, mechanical simplicity, easy construction and operation on all HF amateur bands including 6 meters. Originally built as an accessory filter for a 1500 W 6-meter amplifier, the filter easily handles legal limit power. It attenuates harmonic radiation in the VHF and higher frequency bands and is made with low cost commonly available materials. No complicated test equipment is necessary for alignment.

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

## Electrical Design

The software tool used to design this low pass filter is named *Elsie*. Jim Tonne, WB6BLD, of Trinity Software has made a student/demo version of his *Elsie* filter design software available at no charge. The program is a professional design tool for those interested in filter design/network analysis. The student/demo version is limited to seven stages. This limitation does not affect the usefulness of this program for many Amateur Radio filter requirements. In addition, there is no time limit on how long this student version will remain active on your computer. This program may be downloaded from Jim's site, at **www.qsl.net/wb6bld/** (go to *Software*, and then select *ELSIE.ZIP*). Program documentation and example data files are included. The *Elsie* format data file for this filter, *DC54.lct*, may be downloaded for your own evaluation from the author's Web site at **www.realhamradio.com**.

The *Elsie* menu options and intuitive program design make it relatively easy to get started. The user has a choice of manual filter design or design assisted by the computer. I used a low pass filter design with inductor input and having five poles. After making other filter choices, such as design frequency, the program can calculate all performance parameters and display the predicted filter response. You can use keyboard arrow keys to select an

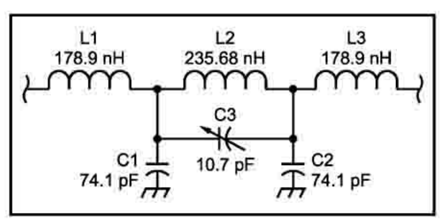



**Figure 1—The low pass filter schematic.**
**C1, C2—74.1 pF. 2 inch by 2.65 inch brass plate sandwiched with 0.03125 inch thick Teflon sheet. The metal enclosure is the remaining grounded terminal of this capacitor.**
**L1, L3—178.9 nH. Wind with 1/8 inch OD soft copper tubing, 3.5 turns, 0.75 inch diameter form, 0.625 inch long, 1/4 inch lead length for soldering to brass plate. The length of the other lead to RF connector as required.**
**L2—235.68 nH. Wind with 1/8 inch OD soft copper tubing, 5 turns, 0.75 inch diameter form, 1.75 inches long. Leave 1/4 inch lead length for soldering.**

item, tune it and immediately see the result. A variety of program options are available for fine-tuning the initial design to allow specific design goals to be realized. The data files may be exported into such other applications as *Touchstone* and *Pspice*. The *Elsie* software has auxiliary tools that help in filter design. These tools run within the program and do not require exiting the software and then restarting again.

I've found that my existing external scientific graphing software could take advantage of the *Elsie* standard two-column format export option for all charts. This helps when adding an *Elsie* chart into a document already using a standardized plotting format. For most uses, the *Elsie* internal video screen and hard copy printer outputs are fine. Figure 1 is a schematic diagram of the filter.

## The Components

See the Figure 1 caption for the parts list. The use of low self-inductance capacitors with Teflon dielectric easily allows legal limit high power operation and aids in the ultimate stop band attenuation of this filter. Capacitors with essentially zero lead length will not introduce significant series inductance that upsets filter operation. This filter also uses a trap that greatly attenuates second harmonic frequencies of the 6-meter band.

## Mechanical Design, Assembly, and Construction

One design goal of this filter was easy tuning with modest home test equipment. See Figure 2. To realize this, build the coils carefully according to the component values table. The homemade coils solder directly to the top surface of the brass capacitor plates. The capacitors are made using a brass to Teflon to aluminum case sandwich. An easy to make variable capacitor is made from two pieces of 0.032-inch thick brass plate and a Teflon insulator. The filter inductors are mounted at right angles to each other to help maintain good stop band attenuation.

One *Elsie* software tool will calculate the details of each inductor. Inductors L1 and L3 are designed with a half turn winding. This allows short connections to the brass capacitor plate and the RF connectors mounted on the enclosure

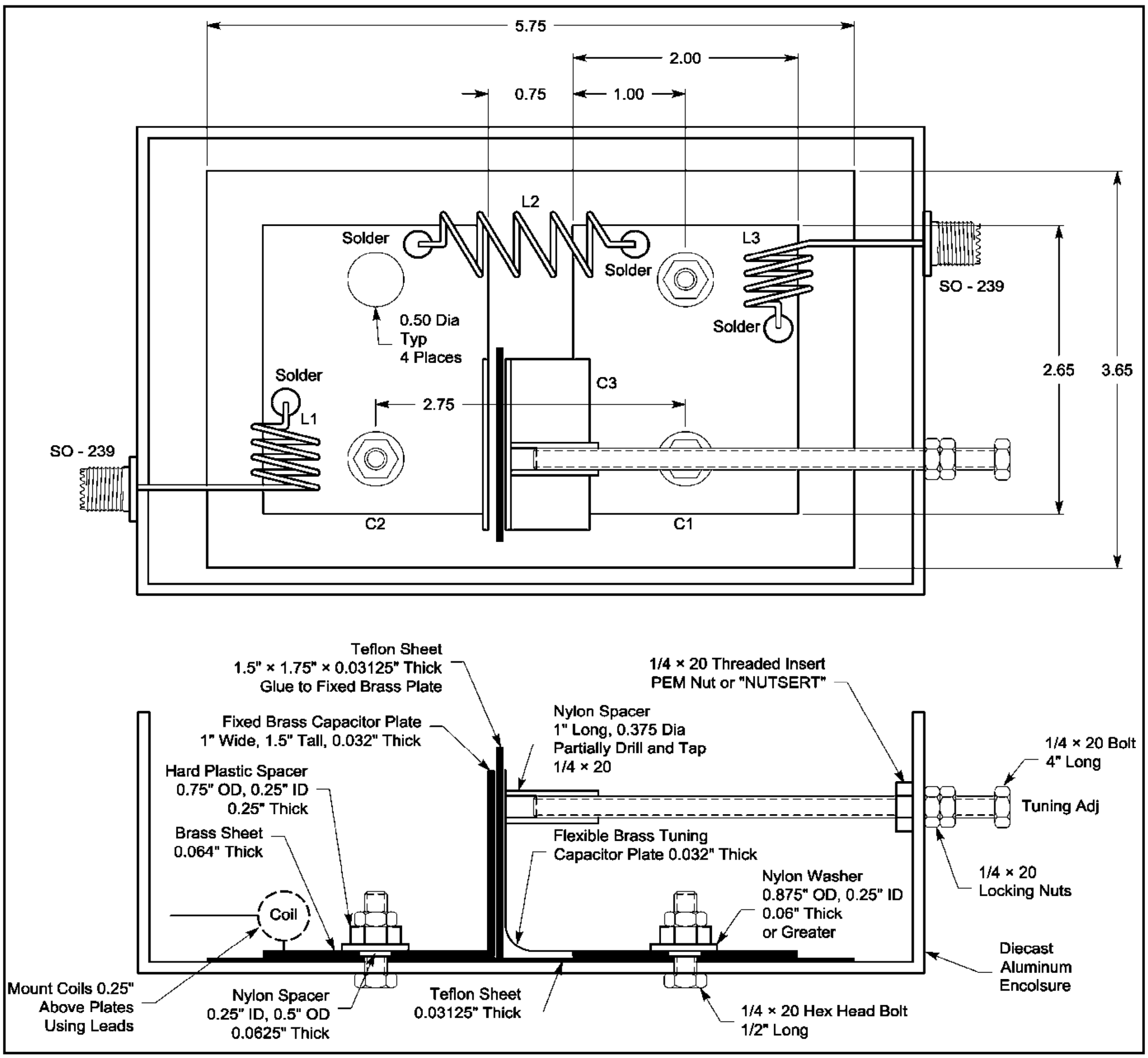



**Figure 2—Assembly drawing for the low pass filter. All dimensions are in inches.**

walls. The coils are physically spaced with 1/4-inch lead lengths, and then soldered to the brass plates.

Many of the parts required to make this filter are available at hardware stores. In particular, the 1 and 2 inch wide brass strips (sold as Hobby or Miniature brass), 1/8 inch diameter soft copper tubing, nuts, bolts, and nylon spacers and washers are commonly available at low cost. It is important that the specified 0.03125-inch thickness of Teflon be used since another size will result in a different capacitor value. If you have another Teflon thickness available, you will need to calculate the specific capacitor values depending upon the new thickness and brass plate sizes.

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

A separate Teflon sheet, also used in the variable capacitor, is glued to the stationary brass plate. This insulator is used to prevent a short circuit in case the tuning screw is tightened too much. Teflon is extremely slick, and doesn't glue well unless chemically prepared. One way to get acceptable glue joint performance between the brass support plate and the insulator is to scuff the Teflon and brass surfaces with 240-grit sandpaper. The intention is to increase the available surface area as much as possible, and provide more places for the glue to fasten to. Glue the Teflon in place with a bead of RTV or epoxy. After drying, the Teflon sheet can be intentionally peeled from the brass plate, but it appears to hold reasonably well. Special Teflon that has been treated to allow good adhesion is available, but the expense isn't justified for this simple application. This Teflon variable capacitor insulator sheet measures 1.5 inches wide by 1.75 inches tall and is larger than the two brass plates. This gives an outside edge insulation some safety margin.

### Calculating Capacitance

The 0.064-inch thick brass capacitor plates have two 0.5-inch holes in them for the mounting bolts and washers. The surface area of each hole is $\pi R^2$, so the two holes combined have a total surface area of 0.3925 square inch. The brass plate size is 2 inches by 2.65 inches. This equals 5.3 square inches of surface area. Subtracting the area of the two holes gives a total surface area of 4.9 square inches. The formula for capacitance[1] is:

$C = 0.2248\ (kA/d)$

where

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

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

The brass sheet material acts like a large heat sink, so an adequate soldering iron is required. A large chisel point 125-W iron will work well. The soldering heat does not affect the Teflon material. However, beware of the temptation to use a small propane torch. Two bolts in each capacitor hold the Teflon sheet and brass plates firmly together. The bolts are insulated from the brass plates by nylon spacers the same thickness as the brass. The nylon plunger for the tuning capacitor needs to be drilled and tapped to accept the 1/4 × 20 thread of the adjustment bolt. A threaded insert or PEM nut in the enclosure provides support for the tuning bolt.

### Tuning Capacitor and Input SWR Adjustment

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

### Six-Meter Alignment Procedure

If you are not concerned with 6-meter operation, ignore this procedure. Simply set the variable capacitor plates 0.1-inch apart and disregard the following steps. If you wish to use this filter on the HF amateur bands from 1.8 to 30 MHz only, the adjustable tuning capacitor adjustment is not critical and does not affect HF SWR performance. However, don't eliminate the capacitor entirely. The software predicts degraded VHF response with it missing. For use on the HF bands only, the tuning screw and associated nylon plunger may be omitted.

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

#### *Step One*

After the filter is constructed, adjust the variable capacitor until the top plate spacing is about 0.1 inch apart. Using a variable SWR analyzer, sweep the 6-meter band area, searching for a very low SWR null anywhere in the vicinity of about 45 to 60 MHz or so. If a low SWR value (near 1:1) can be found, even though the frequency of the low SWR isn't where you want it, proceed to step two. Otherwise, adjust the input coil L1 by expanding or compressing the turns until a low SWR can be obtained anywhere in the range of about 45 to 60 MHz. If you have a way to measure the notch response at 100.2 MHz, proceed to step two. Otherwise, proceed to step three.

#### *Step Two*

Now apply 100.2 MHz to the filter input. Adjust the variable capacitor until the 6 meter second harmonic at 100.2 MHz is nulled on the filter output. Hook up the SWR analyzer again, and sweep the 6-meter band with the SWR analyzer. If the low SWR frequency is too low, adjust middle coil L2 for less inductance (expand turns apart), and then readjust the variable capacitor to bring the notch back on frequency. Continue these iterations until the SWR null is where you want, and the notch frequency is correctly set. Alternately, if the desired SWR low spot is too high in frequency, adjust L2 for more inductance (compress the coil turns), and then readjust the variable capacitor for the second harmonic notch. Continue this until both the low SWR frequency location and the notch null are set where you want. You may need to unsolder one end of coil L2 to allow the adjustment for a longer or shorter coil length as you expand or compress turns. Just solder the end again after you make your length correction. Note that you will probably need to install the enclosure lid during the very final tuning steps. I was able to reduce the second har-

[1] *The ARRL Handbook*, 72nd Ed., Newington, CT, ARRL, 1995, p 6.9.

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

### *Step Three*

This step is only performed if you don't have a way to generate the 100.2 MHz input signal, and then detect a null on the filter's output terminal. The variable capacitor will become your SWR adjustment to move the SWR null spot to the portion of the 6-meter band you desire. If you run out of adjustment range on the variable capacitor (turned all the way in), just compress the L2 turns together, and try again. Alternately, if the variable capacitor is backed completely off, just expand the L2 turns, and try again. After your SWR is set, you are finished. Although the second harmonic notch probably isn't exactly on frequency, you will still have good (but not optimum) suppression since the notch is very deep. Table 1 is a list of parts.

**Table 1**
**Parts List**

| *Qty* | *Description* |
|---|---|
| 1 | Miniature brass strip, 1 × 12″, 0.032″ thick (variable tuning capacitor) |
| 1 | Miniature brass strip, 2 × 12″, 0.064″ thick (main filter capacitors) |
| 5′ | 1/8 inch diameter soft copper tubing |
| 4 | ¼ × 20 × ½″ long hex head bolt |
| 4 | Plastic spacer or washer, 0.5″ OD, 0.25″ ID, 0.0625″ thick |
| 6 | 1/4 × 20 hex nut with integral tooth lock washer |
| 1 | 1/4 × 20 × 4″ long bolt |
| 1 | 1/4 × 20 threaded nut insert, PEM nut, or "Nutsert" |
| 1 | 1 × 0.375″ diameter nylon spacer. ID smaller than 0.25″ (used for variable capacitor plunger). |
| 4 | Nylon spacer, 0.875″ OD, 0.25 to 0.34″ ID, approximately 0.065″ or greater thickness (used to attach brass capacitor plates). |

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

## Performance Discussion

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

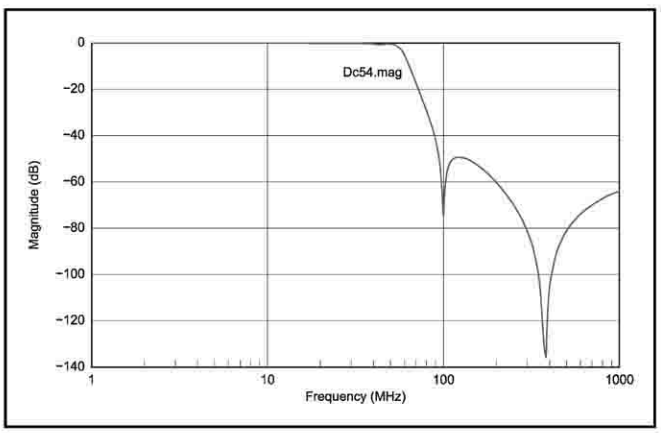



**Figure 3—The filter response from 1 to 1000 MHz.**

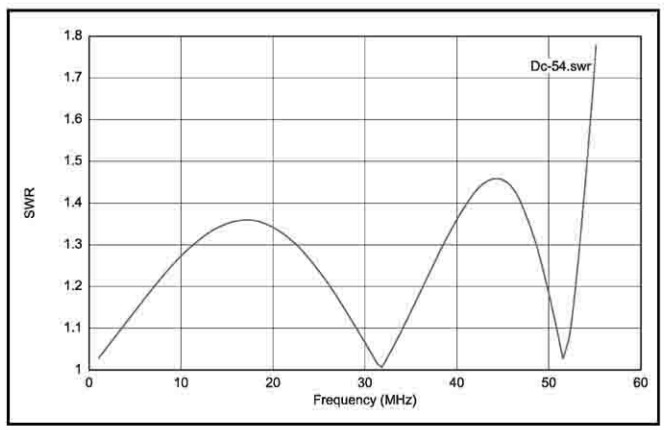



**Figure 4—The filter SWR from 1 MHz to 55 MHz.**

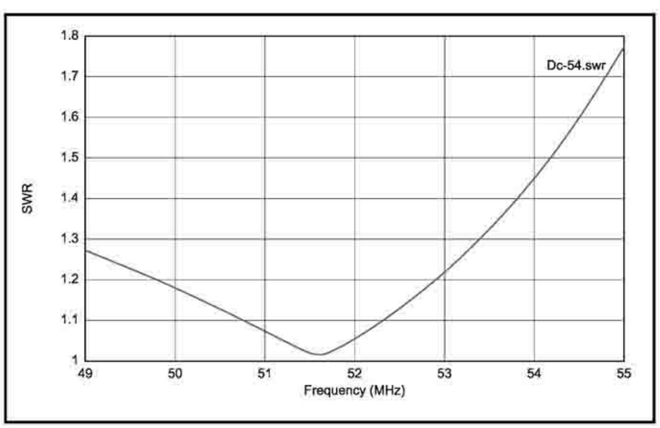



**Figure 5—Six-meter filter SWR.**

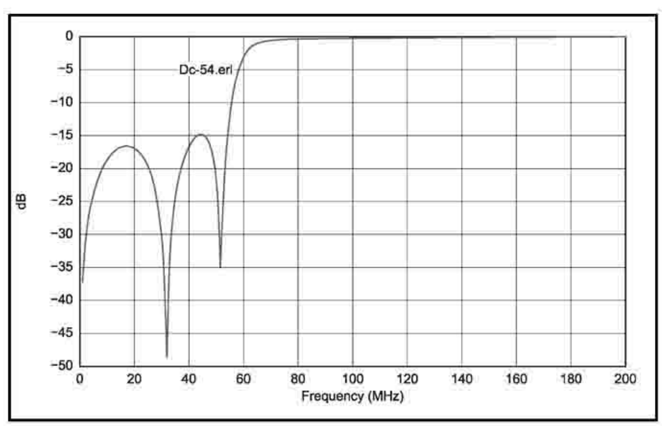



**Figure 6—Calculated filter return loss.**

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

Figure 3 shows the filter response from 1 to 1000 MHz. The impressive notch near 365 MHz is because of these inherent stray capacitances across each of the coils. Slight variations in each coil will make slightly different tuned traps. This will introduce a stagger-tuned effect that results in a broader notch.

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

The SWR across the HF bands and 6 meters is shown in Figure 4. Figure 5 shows only the 6-meter band SWR.

Calculated return loss of the filter across 1 to 200 MHz is shown in Figure 6. Notice that the 10-meter region has particularly good return loss. Component values in this filter were adjusted so that this return loss spike was moved from about 40 MHz to around 28-30 MHz.

## Summary

This filter meets the original design objectives. Since I use 6 meters as well as the regular HF bands, this project has produced a doubly useful station accessory. Low insertion loss makes this filter useful for receiving applications also. The *Elsie* filter software tool made the electrical design portion of this project fun. Thanks to Jim Tonne, WB6BLD, for the *Elsie* design software and for his informal consultation and helpful comments about this filter. Jim also suggested this filter topology and offered component values to consider.

*Bill Jones, K8CU, has been an active radio amateur, CW DXer and home project builder since first receiving his license in 1966 at age 17. Bill is an electronics professional with experience ranging from radar maintenance in Vietnam, analog and digital telephone PBX design, to embedded controller implementation. He spent 18 years at Optek, Inc, designing hardware and writing assembly language software for embedded microcontrollers used in specialized electro-optical industrial process controls. He is currently employed by the Ohio Department of Transportation, where he works in radio communications. Some recent interests include small gas engines, 6-meter DXing and maintaining his personal Web site,* **www.RealHamRadio.com**. *A member of a nearby sportsman's club, Bill can sometimes be found fishing with his wife, Bonnie. You can contact Bill at* **k8cu@realhamradio.com**.

# NEW PRODUCTS

## MFJ'S 'DX BEACON' REAL-TIME PROPAGATION MONITOR

◊ Want up-to-the-minute information about worldwide band conditions on 20 through 10 meters? MFJ's new Model 890 DX Beacon Monitor can do just that, thanks to its ability to listen to transmitted information from the 18 stations around the globe that comprise the IARU International Beacon Network.

The DX Beacon Monitor lets you instantly see which propagation beacons are being received: An LED lights up on the unit's built-in world map to show you the beacon location and where to point your antenna. As data from each beacon is received, the corresponding LED lights to indicate that the radio path on that frequency is open between the beacon and your station.

Because the worldwide transmitter network uses precise timing intervals, the DX Beacon Monitor has a built-in atomic clock receiver and microprocessor control to maintain precise synchronization with the beacon network. The Model 890 is a standalone unit that requires no connections to your transceiver or receiver. Measuring about 5×7×3 inches HWD, the monitor requires 9-12 V dc.


Price: $99. For more information, see your favorite Amateur Radio products dealer or contact MFJ at 300 Industrial Park Rd, Starkville, MS 39759; tel 800-647-1800, fax 662-323-6551, **mfj@mfjenterprises.com**, **www.mfjenterprises.com**.

## INTERNET-BASED HAM RADIO INSTRUCTION

◊ Boston-based J. Cunningham & Associates has just launched **www.hamtestonline.com**, a Web site aimed at helping new and experienced hams prepare for US Amateur Radio written exams. According to the publisher, this is the first Web site to incorporate computer-based training (CBT) technology for ham radio education.

HamTestOnline site tracks the questions you've seen, the questions you've learned and the questions you get right and wrong. It asks you questions based on your personal needs, concentrating on the areas where you are weak.

The easy-to-use service operates entirely online and there is no software to download or install. The system includes all questions from the latest Technician, General and Amateur Extra question pools.

HamTestOnline offers a free trial, which includes 20% of the questions from each question pool. A paid subscription costs $19.95 and provides access to all questions in all three question pools for a period of two years. The site offers a money-back guarantee if you are dissatisfied for any reason.

For more information, point your Web browser to **www.hamtestonline.com**.

## TINY VFO KIT FROM DATAK MANUFACTURING

◊ Aimed at home-brewers and QRP enthusiasts, DATAK's new Model 80-1410 wide-range VFO kit can be made to cover frequencies from 100 kHz to 33 MHz by swapping a resistor and a capacitor. Using the new LTC 1799 precision oscillator IC chip, the kit is essentially a resistance-controlled oscillator that can be set up to cover multiple 200-kHz tuning ranges from VLF to VHF. Stability is rated at ±40 ppm per °C. Use it as a VFO or as an adjustable "crystal replacement" oscillator to drive mixers, front ends, etc. The VFO kit can be assembled in about 20 minutes, runs on 6 to 14 V dc (at 140 mA) and ships with parts for tuning 40 meters (6.9 to 7.2 MHz).

Price: $15.99 (includes complete instructions and a schematic for a companion 40-meter, 5-W, CW transmitter). For more information, contact DATAK at PO Box 6386, 3660 Publishers Dr, Rockford, IL 61125; tel 800-645-2262. To find a distributor in your area, point your web browser to **www.philmore-datak.com**.

Next New Products

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

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

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

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

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

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

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

## DESIGNS AND INTERFACING

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

## CONSTRUCTION

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

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

**Table 12.25**
**CW Filter Using One 85-mH Inductor Stack and Two Modified 85-mH Inductors**

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

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

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

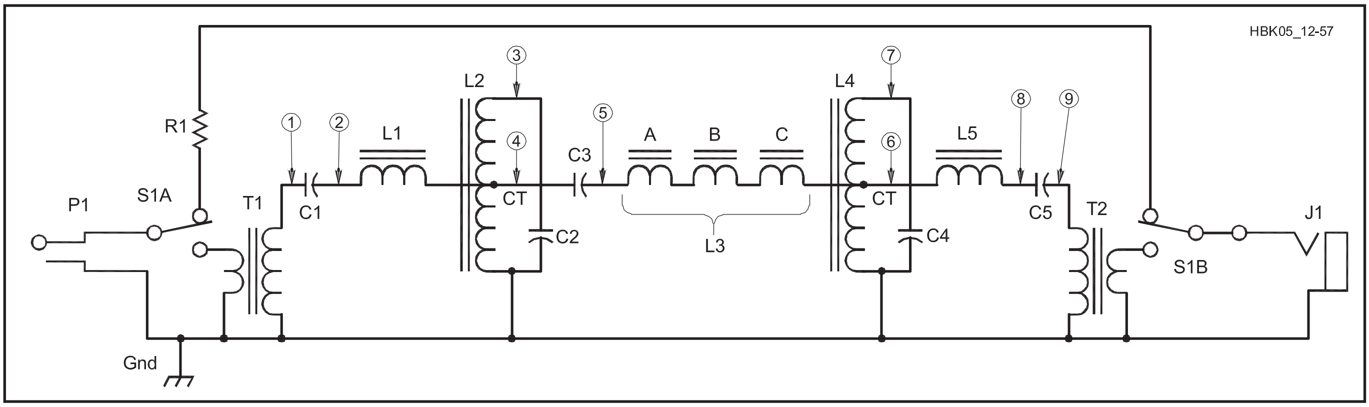



**Fig 12.89 — Schematic diagram of the five-resonator CW filter. See Table 12.25 for capacitor and inductor values to build a filter with a center frequency of 546, 600, 700, 750 or 800 Hz.**

**P1 — Phone plug to match your receiver audio output jack.**
**J1 — Phone jack to match your headphone.**
**R1 — 6.8 to 50 Ω, ¼-W, 10% resistor (see text).**
**S1 — DPDT switch.**
**T1, T2 — 200 to 8-Ω impedance-matching transformers, 0.4-W, Miniature Core Type EI-24, Mouser No. 42TU200.**
**Note: The circled numbers identify the circuit nodes corresponding to the same nodes labeled in the pictorial diagram in Fig 12.90.**

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

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

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

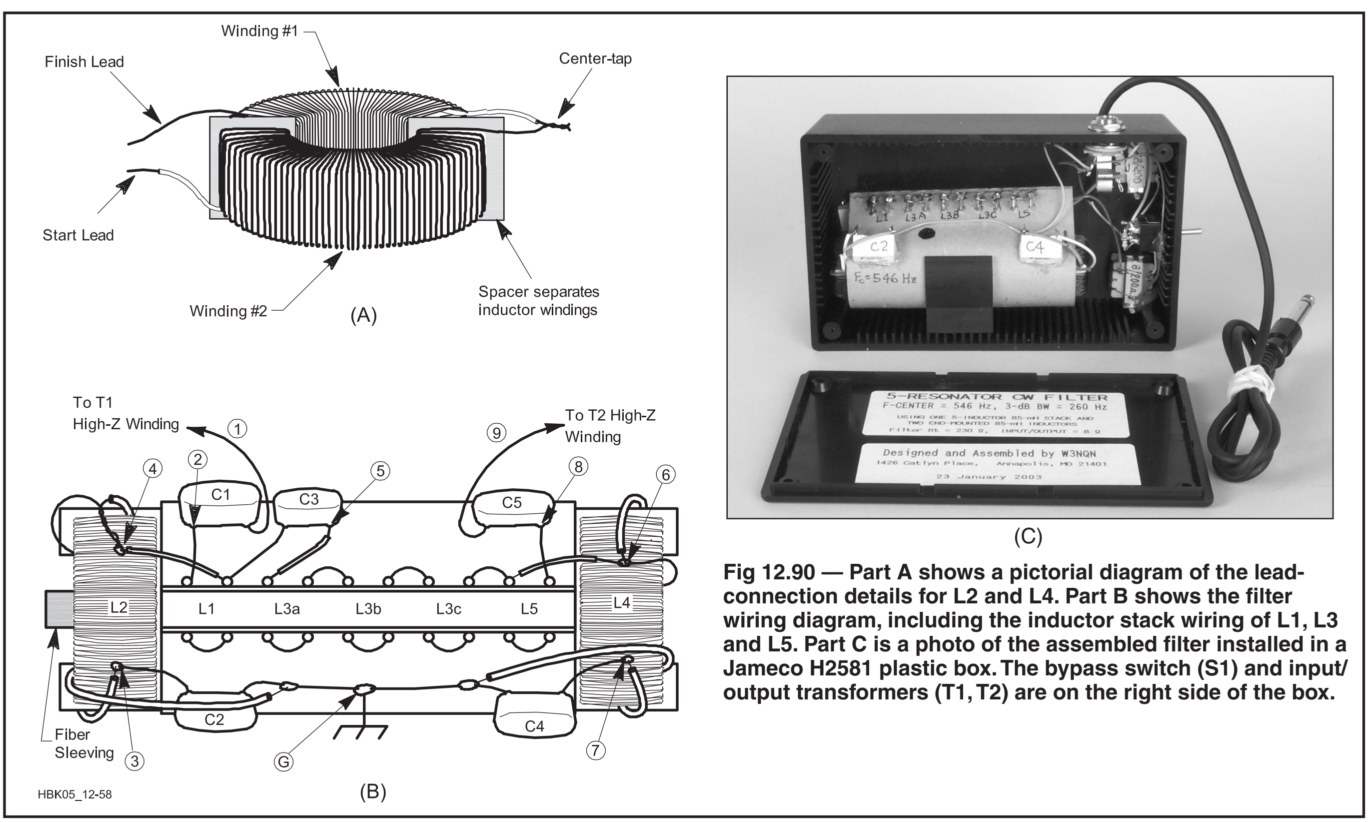



**Fig 12.90 — Part A shows a pictorial diagram of the lead-connection details for L2 and L4. Part B shows the filter wiring diagram, including the inductor stack wiring of L1, L3 and L5. Part C is a photo of the assembled filter installed in a Jameco H2581 plastic box. The bypass switch (S1) and input/output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

**Table 12.26**
**Node-to-Node Resistances for the 546-Hz CW Filter**

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

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

## INTERFACING TO SOURCE AND LOAD

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

## PERFORMANCE

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

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

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

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

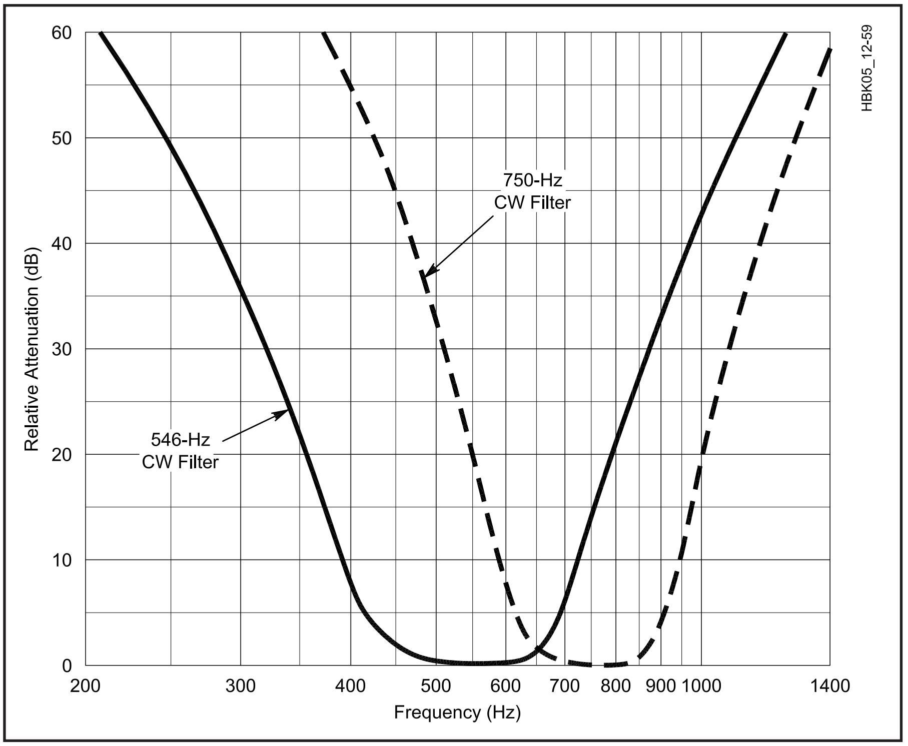



**Fig 12.91 — Measured attenuation responses of the 546- and 750-Hz filters. The responses are plotted relative to the zero dB attenuation levels at the center frequencies of the filters. The other filter response curves are similar, but centered at their design frequency.**

## ORDERING PARTS/CONTACTING THE AUTHOR

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

## Notes


[1]Private correspondence from Steve Root, KØSR.
[2]R. Schetgen, Ed., 1994 *ARRL Handbook*, pp 28.1-28.2 (Simple High-Performance CW Filter)
[3]W. Orr, Ed., *Radio Handbook*, 23rd edition, Howard W. Sams & Co., 1987, pp 13.4-13.6 (1-Stack CW Filter).
[4]Wetherhold, "Modern Design of a CW Filter using 88- and 44-mH Surplus Inductors," *QST*, Dec 1980, pp 14-19 and Feedback, *QST*, Jan 1981, p 43.
[5]Wetherhold, "High-Performance CW Filter," *Ham Radio*, Apr 1981, pp 18-25.
[6]Wetherhold, "CW and SSB Audio Filters Using 88-mH Inductors," *QEX*, Dec 1988, pp 3-10.
[7]Wetherhold, "A CW Filter for the Radio Amateur Newcomer," *Radio Communication* (Radio Society of Great Britain), Jan 1985, pp 26-31.

# Using Active Filter Design Tools

*Editor's note: Section and figure references in this article are from the 2010-2013 editions of the* ARRL Handbook. *This material was originally contributed to the* Handbook *by Dan Tayloe, N7VE.*

Sophisticated active filter circuits are more easily designed using filter-design software. Follow the same general approach described in Chapter 11 of *The ARRL Handbook* to determine the filter's performance requirements and then the filter family. You can then enter the values or make the necessary selections for the design software. Once a basic design has been calculated, you can then "tweak" the design performance, use standard value components and make other adjustments. The design example presented in this document shows how a real analog design is assembled by understanding the performance requirements and then using design software to experiment for a "best" configuration.

This design example makes use of Texas Instrument's "freeware" filter design software, *FilterPro*. This package is extremely useful in designing active RC filters. This package allows filter parameters to be adjusted so as to "tweak" the design close to standard component values. (Go to **www.ti.com** and search for "FilterPro".) The reader is encouraged to follow along and experiment with *FilterPro* as a means of becoming familiar with the software so that it can be used for other filter design tasks.

## DESIGN EXAMPLE: 750-HZ HIGH-PERFORMANCE DIRECT CONVERSION RECEIVER FILTER

This example illustrates the design of the front-end filter for a high-performance direct-conversion I-Q phasing CW receiver. Along with the design of the filter, there will be some discussion of how much gain to assign to each circuit in a sequence of stages. This is included to illustrate some of the processes by which performance requirements for a circuit are established.

A unity-gain filter with a fixed-gain block in front of it means that the fixed-gain block will have no out-of band signal rejection and thus may be subject to overload due to strong out-of-band signals. Conversely, a unity-gain filter followed by a fixed-gain block could degrade signal sensitivity due to the noise internal to the unity-gain filter section. A filter with distributed gain is better than either of these two situations as the signals can be both filtered and amplified stage by stage.

The input signal for this example will be assumed to be straight from the detector with a 100-Ω output impedance and may be as large as 4 $V_{P-P}$ (+16 dBm). This filter is assumed to operate from 12-V supplies so that up to 8 $V_{P-P}$ can easily be handled by commonly available op amps such as a LM5532. (Unless an op amp is capable of rail-to-rail output, its output voltage can approach no closer than 2 V from either the positive or negative supply rails, thus 12 – 2 – 2 = 8 V of total output swing.)

A typical receiver requires approximately 80 to 90 dB of total gain for adequate headphone output levels. Speaker-level output usually requires an additional 20 dB of gain. If the volume control is placed too early in the gain chain (at the antenna input, for example), the audio stage will always be running at maximum gain and the result will be a high level of unwanted internally generated receiver hiss, noise that is not affected by the volume control. On the other hand, if the volume control is too late in the gain chain (such as at the receiver output), all the signals being received will be amplified by the maximum receiver gain of 80 to 90 dB. This can cause strong signals in the passband to saturate the audio chain, causing unwanted distortion unless the receiver uses AGC to reduce the gain.

A reasonable compromise is to place the volume control roughly halfway along the chain of gain stages. Thus, if 80 dB of total gain is desired, roughly 40 dB would occur before the volume control and 40 dB after. This keeps the gain after the volume control low enough that unwanted receiver hiss will be largely eliminated when the volume control is turned all the way down as long as a low-noise amplifier chain is used. This implies that the first 40 dB of gain will be before the volume control, which in this design will be rolled into the active RC filter.

The design objectives at this point are:

1) To provide 40 dB of gain in the desired 750 Hz passband, but 0 dB of gain at 2 kHz and higher.

2) At every stage in the filter, the signal at 2 kHz should not be allowed to exceed 8 $V_{P-P}$ for a 4-$V_{P-P}$ input. This means gain at 2 kHz should be 20 log (8/4) or 6 dB or less out of each stage.

3) The filter should be designed to minimize the impact on receiver sensitivity. It should not add unnecessary noise that would mask weak signals or be susceptible to overload from strong signals.

The first and second goals are attempts to ensure that no signal out of the detector at 2 kHz or higher will overload the filter section. This enables the creation of a very high performance direct-conversion receiver. The second goal is to allow that receiver to have high sensitivity which in turn has noise implications on the filter design.

### Audio Filter Q Implications

When using the *FilterPro* software, Q must be specified for each filter section. In experimenting with the filter types of the Bessel, Butterworth, and Chebyshev and observing the frequency responses, it can be quickly seen that higher Q is associated with sharper filter frequency rolloff. From this observation, the conclusion could be drawn that high Q in an active RC filter is a good thing. This conclusion is not entirely true.

From a receiver design perspective, the goal is to reject undesired signals to the highest degree possible. An ideal 750-Hz low-pass filter would pass all signals at and below 750 Hz while completely eliminating all signals 751 Hz and higher. A high-order active RC filter with high-Q filter sections comes closest to this ideal. However, this sharp frequency rolloff does not come without a price.

The first problem with high-Q filter sections is ringing. **Fig 11.47** was generated using *FilterPro* for a 5th-order Chebyshev filter with 1 dB of ripple, a cutoff frequency of 750 Hz, and 40 dB of gain. This filter provides 20 dB of attenuation at 2 kHz exceeding our design goal of 0 dB. However, notice in particular the sharp peak filter group delay response at the 750 Hz cutoff frequency. This sharp group delay peak is associated with audio ringing.

The effect of ringing in a filter is much the same as ringing a bell. Strike a bell with a hammer, and the bell "rings" at a certain frequency. Likewise, when noisy, static-filled band noise hits a high-Q filter such as the one shown above, this impulse noise tends to produce an audible "ring" sound at the frequency of the delay spike. The effect of the filter ringing is that it actually creates audible interference that interferes with and can mask the desired signal.

It should be noted that simple crystal ladder filters used in many simple superheterodyne or "superhet" receivers have a band-pass characteristic. Thus, there is both a high and a low band-pass edge where group delay peaks occur. That means the typical narrow 300- to 500 Hz-wide CW crystal filter tends to ring badly at both a high (top end of the band-pass response) and a low frequency (bottom end of the band-pass response) at the same time, which makes the ringing audio artifacts twice as bad.

The second problem with high-Q filters is an effect that is not at all obvious. It is simply the fact that our ears do not like them. This is caused by both the high phase and delay varia-

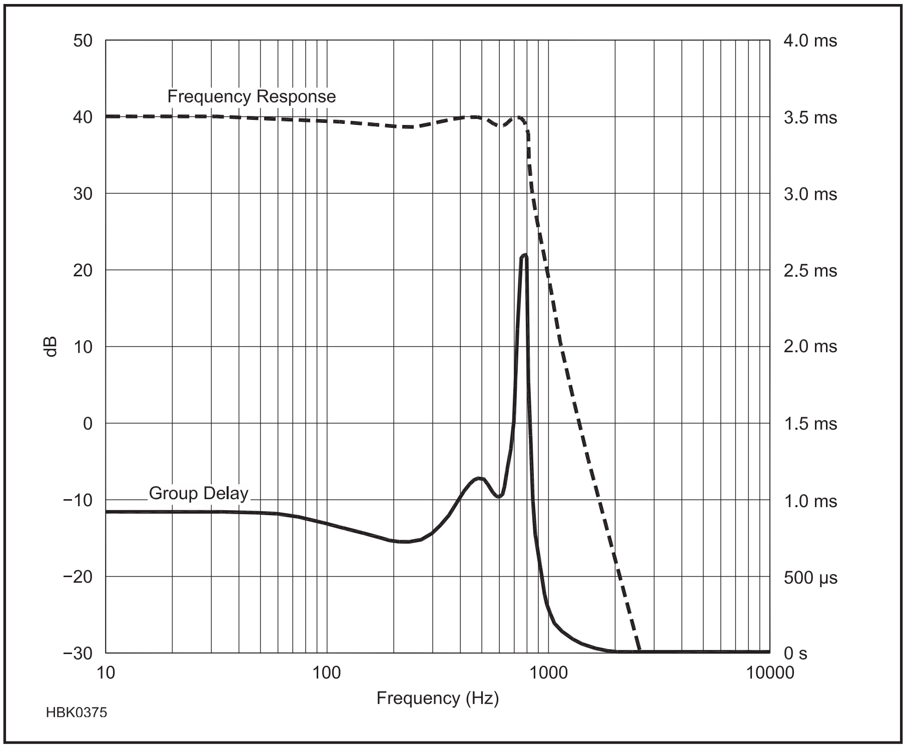



**Fig 11.47 — Frequency response and group delay of 5th order Chebyshev, 1 dB passband ripple, 40 dB of gain.**

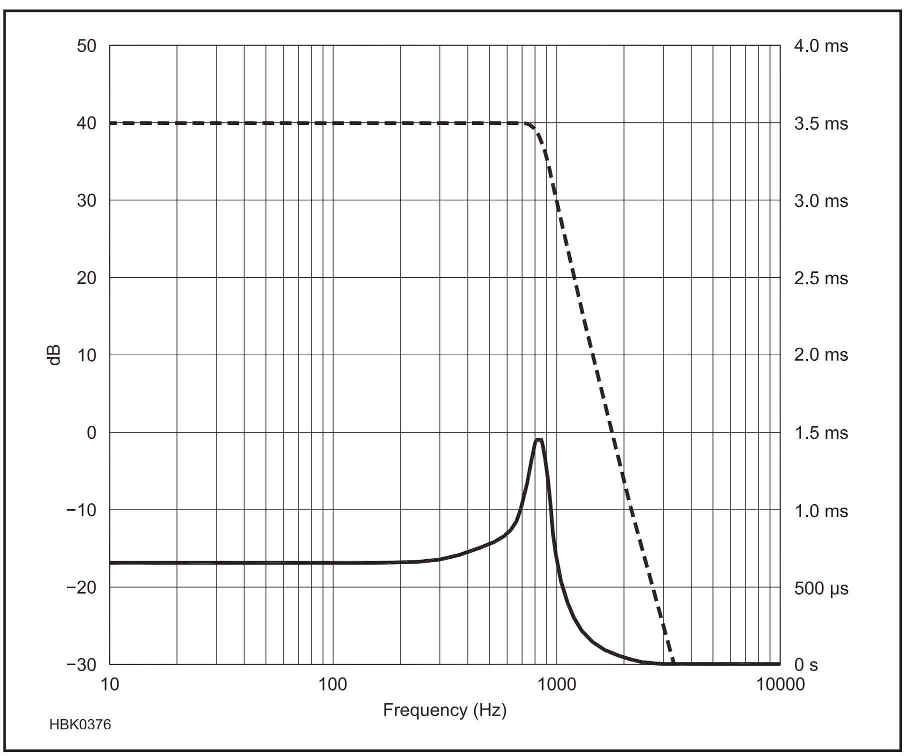



**Fig 11.48 — Frequency response (dashed line) and group delay (solid line) of 5th order Chebyshev, 0.06 dB passband ripple.**

tions near the edge of the filter. Something in our brain "notices" these variations and objects to them. To our ears, a filter than has lower phase and delay variations "sounds" better, even outside the ringing issue.

In practice, the problems associated with both ringing and phase or delay variations can be reduced by limiting the highest Q filter section to a maximum Q of around 3. When using a design tool like *FilterPro*, this can be done by selecting a filter type of Chebyshev, and then reducing the specified allowed ripple dB value until the Q is reduced to roughly 3. For a 5th-order Chebyshev, this means reducing the allowable ripple from 1 dB to 0.06 dB. The group delay and frequency response of such a filter are shown in **Fig 11.48**.

Notice that the delay peak (the lower line) is now both smaller in amplitude and much broader than that in the previous example, which is exactly what we are after. Notice also that the frequency rolloff (the top curve) is not as sharp either, but it is still 8 dB better than our target of 0 dB at 2 kHz and above.

What does this mean? In active RC filter design, there is a tradeoff between simple and useful vs. more complex and better sounding. A higher-Q active RC filter may require only three filter sections while a lower-Q, better-sounding active RC filter with similar rolloff characteristics may require four filter sections. It is very valuable to realize that a tradeoff is possible and that an active RC filter can be built which is both sharp and sounds good (low phase/delay variations) at the same time.

### Noise Implications

Resistors create noise that can mask the small signals we are trying to filter. If this filter is being used at the high-signal end of an audio chain, high-value resistors can be selected without any real harm. Resistance values as high as 1 MΩ can be useful in allowing the selection of small value capacitors which are readily and cheaply available in 2% or 5% tolerance values. However, if this filter is to be used in the front end of a receiver chain, the resistor values need to be much lower. (Receiver noise is also discussed in the **Receivers** chapter.)

A 50-Ω resistor creates about $0.85\ \text{nV}/\sqrt{\text{Hz}}$ of noise. Think of this as the noise generated by a 50-Ω antenna system. This noise voltage varies with the square root of the resistance change. Thus, a 1-MΩ resistor produces $0.85\sqrt{1{,}000{,}000/50}$ or $120\ \text{nV}/\sqrt{\text{Hz}}$ of noise. Thus, using a 1-MΩ resistor in the first stage of an active RC filter would reduce the sensitivity of an ideal receiver by 20 log (120/0.85) or 43 dB which is not good for receiver sensitivity.

If each stage of the active RC filter has gain, the effect of that gain is to lessen the impact of the noise contributed by the resistors in each

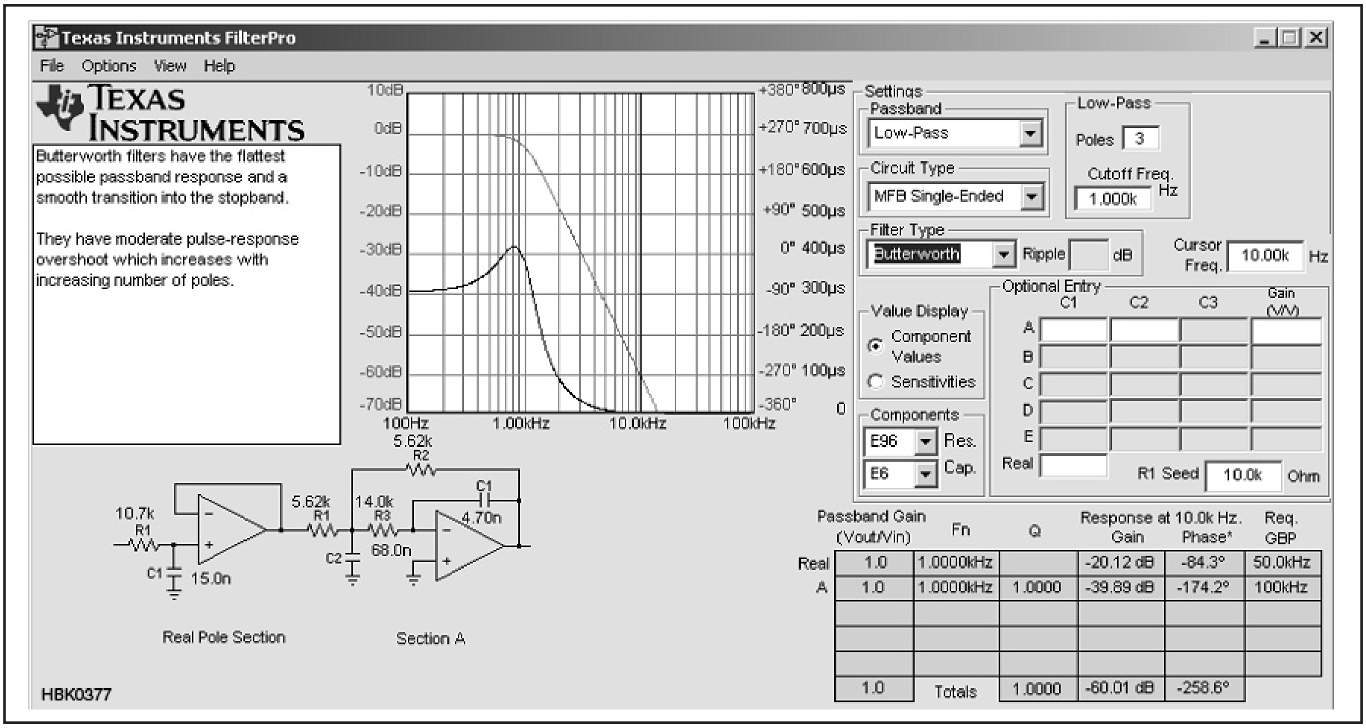



**Fig 11.49 — *FilterPro* startup view.**

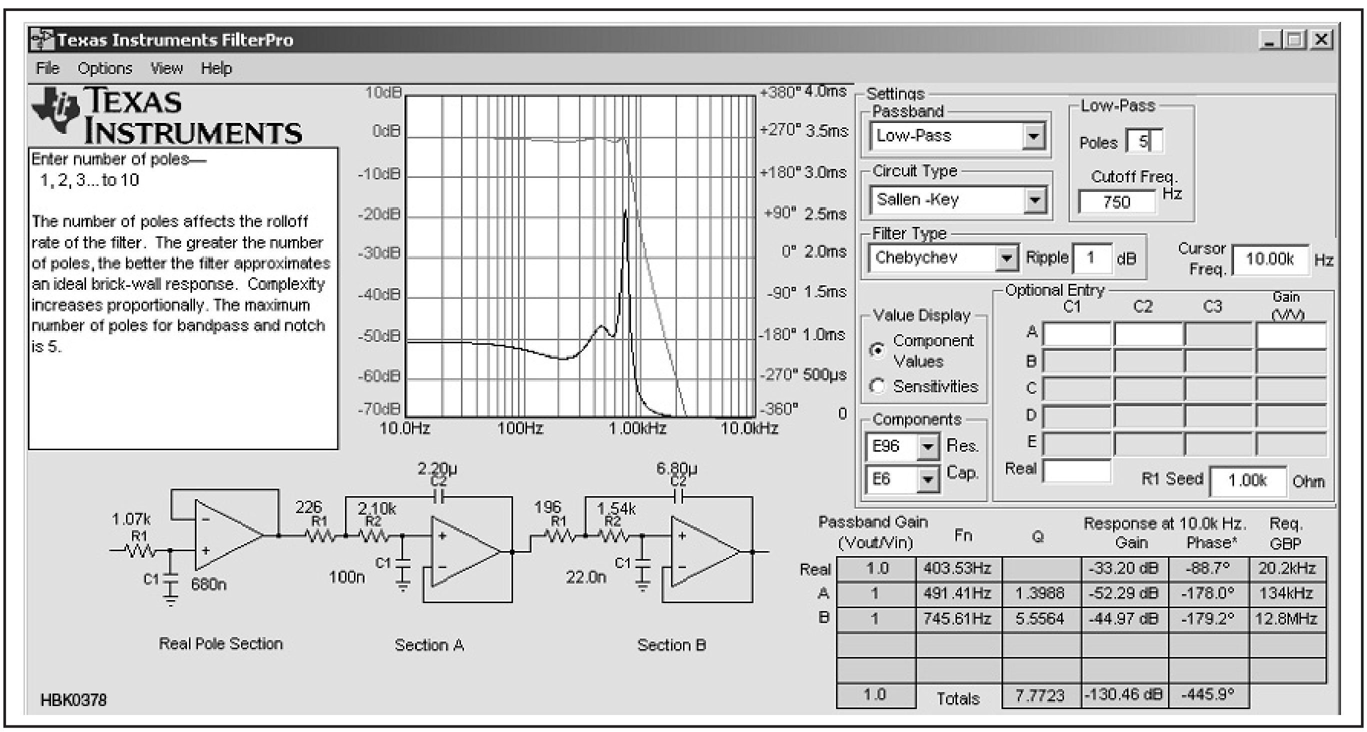



**Fig 11.50 — First pass at designing a 5-pole 750-Hz low-pass Chebyshev filter.**

succeeding stage. For example, if we want 40 dB (100x voltage gain) in three filter stages, then if the gain is evenly distributed across all the stages, each stage will have a gain of the cube root of 100 or a gain of 4.6 in each of the three stages. In this case, an input signal at $0.85\ \text{nV}/\sqrt{\text{Hz}}$ into the first stage will be 4.6× larger ($3.9\ \text{nV}/\sqrt{\text{Hz}}$) into the second stage and 4.6× larger ($18.3\ \text{nV}/\sqrt{\text{Hz}}$) into the third stage. One goal would be to use resistors than generate at most half the noise voltage of the desired signal. Thus the second stage ($3.9\ \text{nV}/\sqrt{\text{Hz}}$) should have resistors no larger than $3.9 = 0.85\sqrt{x/50}$ or $50 \times (3.9/0.85) \times (3.9/0.85) = 1052\ \Omega$. It should be no problem in the second and third stages to restrict resistor values to the 500-1000-Ω range in order to minimize their noise contribution.

## FILTER DESIGN AND COMPONENT VALUE OPTIMIZATION

With an understanding of the gain and component values, we can use *FilterPro* to design our filter. We want a gain of around 4.6× per

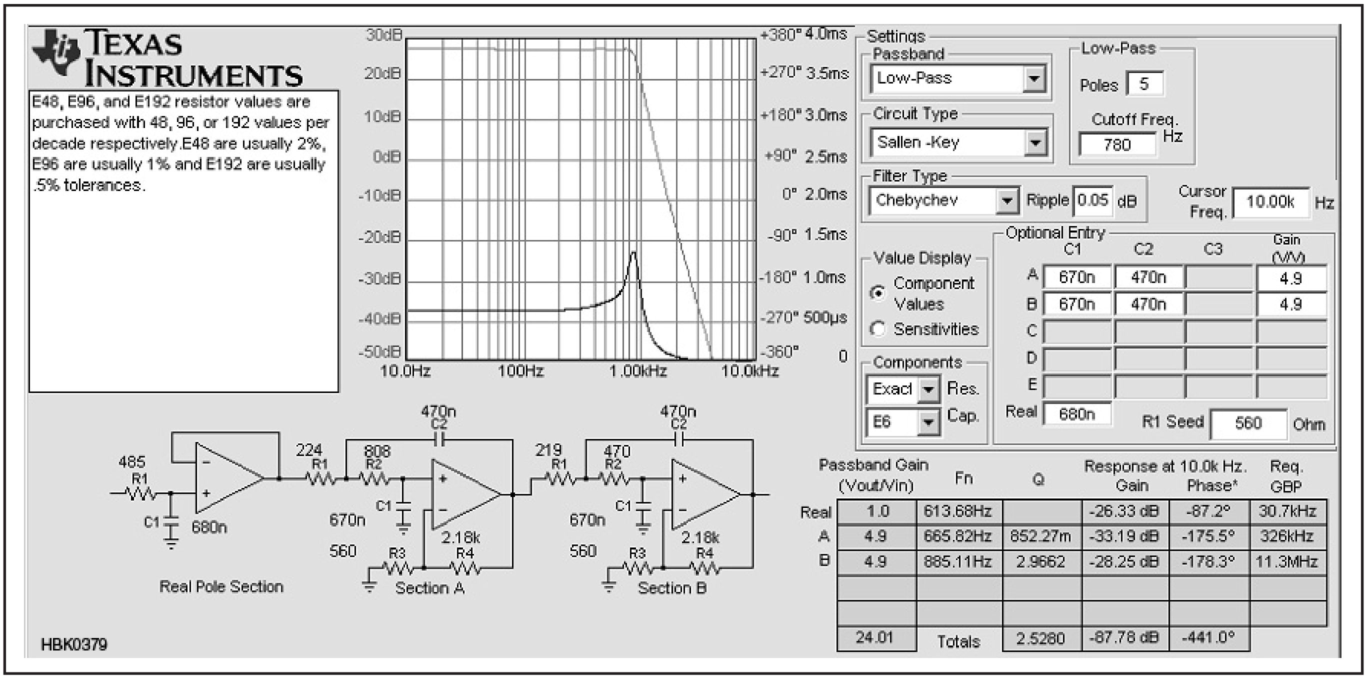



Fig 11.51 — Final result — a 5-pole 780-Hz low-pass Chebyshev filter after component optimization.

stage, resistors in the under-1000 Ω range, a filter Q of 3 or less, and a cutoff frequency of 750 Hz since this is a CW filter. The initial *FilterPro* screen appears as in **Fig 11.49** with a default Passband setting of Low-Pass.

Note the filter frequency response in the graph. Although we have not yet added any gain to any of the stages (we want to add 40 dB or 100×), we can see that the attenuation at 2 kHz is only 20 dB. If we were to add 40 dB of gain, we would still have 20 dB of gain at 2 kHz. Thus we need a sharper filter than a three-pole Butterworth.

We will now change the Filter type to "Chebychev" (one of several spellings), and the Cutoff Frequency to 750 Hz. Under Components, change the resistor setting to "Exact."

In the particular example of a direct-conversion receiver post-detector low-pass filter, it is best to use an odd number of poles so that the odd one-pole section can be readily configured as a receiver post detector preamplifier stage. (One-pole low- and high-pass filters were introduced in Fig 11.45.) Experimenting with *FilterPro*, it can be seen that even numbered pole filters produce more complex stages. Thus, when a three-pole filter is not good enough, the next step up should be a five-pole filter in this particular application.

When using a five-pole Chebyshev filter with a 750 Hz cutoff frequency and an initial value for R1 of 1 kΩ (remember, we want resistances of 1 kΩ or below), *FilterPro* gives the result shown in **Fig 11.50.**

Notice that at 2 kHz, the filter attenuation is now 60 dB. When we create 40 dB of gain in these filter stages, the filter attenuation will still be 20 dB better than needed. However, in the bottom right hand corner, the Q of each section is given, and the highest Q (section B) is 5.55 — higher than the desired maximum Q of 3. The Q of these stages can be adjusted by changing the passband ripple specification, which is set to 1 dB by default for the Chebyshev response. Manually lower the allowable ripple until the Q goes below 3. Experimentally you will find that lowering the passband ripple from 1 dB down to 0.05 dB produces the desired Q of a bit less than 3.

Next set the stages up for the proper gain. You will notice in Fig 11.50 that Circuit Type has been changed to Sallen-Key. This configuration tends to work a bit better as properly biasing the first stage to half the supply voltage will also dc bias all the stages after it and this configuration tends to produce values that are easier to work with. Earlier it was calculated that each stage needed a gain of 4.6× (4.6 × 4.6 × 4.6 = ~100×). This is not a precise value. We want to use component

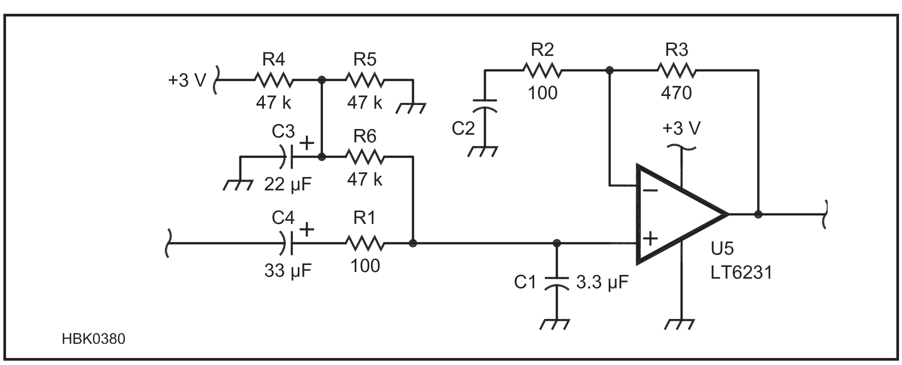



Fig 11.52 — Active filter circuit that implements the "Real Pole" section with biasing and ac decoupling components (see text).

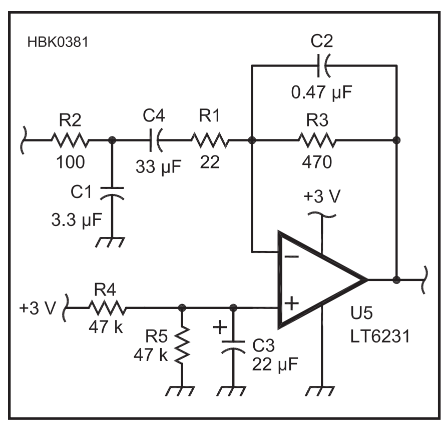



Fig 11.53 — Alternate "Real Pole" circuit with higher performance than Fig 11.52.

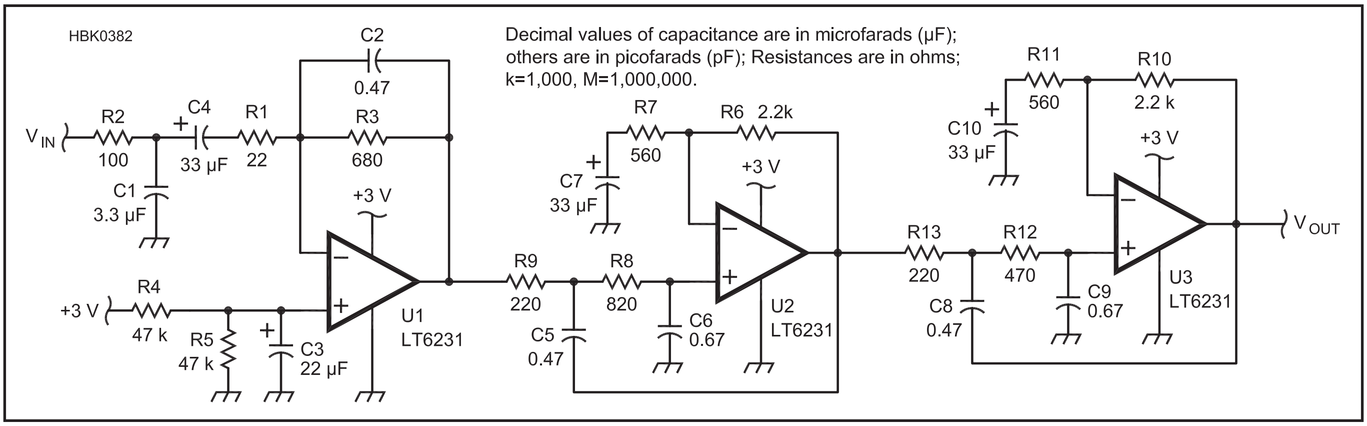



**Fig 11.54 — Schematic of the complete filter, including dc biasing components.**

values that are close to standard values. Thus we can play with the gain per stage (staying close to 4.6×), and set the "R1 seed" value to obtain values for R3 and R4 close to standard values, then play with C1 and C2 and perhaps adjust the cutoff frequency a bit to get values close to standard values for the other resistors in the two stages. The final result is shown in **Fig 11.51** — a 5-pole, 780-Hz low-pass filter. The software makes the process of "cut and try" much easier than a manual design!

The result comes about from tweaking the cutoff frequency a bit (750 Hz to 780 Hz) and playing with the capacitor values. Capacitor values of 0.67 µF were used instead of 0.68 µF as that value is expensive and hard to find. The 0.67 µF caps are composed of one more-available 0.47 µF capacitor and two very commonly available 0.1 µF capacitors in parallel. This allows the main two sections (section A and B) of the filter to use a total of four 0.47 µF and four 0.1 µF capacitors. In optimizing for component values, section B with higher Q is most sensitive to component value, so section B was optimized to get very close to 220 and 470 Ω (standard values), while the lower-Q section will not be as close when using 220 and 820 Ω.

A note about the gain-setting resistors R3 and R4: The goal was to keep resistors under 1 kΩ in these two stages, but R4 is 2.2 kΩ. As far as noise contributions are concerned, R4 and R3 "look" like they are in parallel to the input of the op amps. Thus, together they look like 1/(1/560 + 1/2180) or 445 Ω, which is indeed much less than 1 kΩ.

This still leaves the one-pole "real pole section" to be configured with the proper gain. There are two approaches that can be taken. The simplest circuit is presented in **Fig 11.52** with R1 = 485 Ω and C1= 680 nF. The input connects to C4 and the output is taken from the output pin of U5.

To match the receiver detector output impedance of 100 Ω, we scale R1 from 485 to 100 Ω. We must also scale C1 up by the same amount (4.85×) to make C1 = 3.3 µF. R1 can actually be eliminated as a separate component, replacing it with the 100-Ω output impedance of the detector. If R1 is eliminated, C1 should be moved to the input side of C4. C1 should be a ceramic type capacitor, not an electrolytic.

If the detector produces its own 1.5-V bias from the 3-V supply (like a Tayloe detector), the bias components R4, R5, R6 and C3 can also be eliminated along with the dc isolation capacitor, C4. U5 is shown as an LT6231 low-noise 3-V op amp. If a less expensive 12 V device is used, such as an LM5532, the bias network will be used to set the bias voltage to 6 V (1/2 the power-supply voltage) and all these parts will be needed.

Another higher-performance variation of the "real pole" section is shown in **Fig 11.53.** Again, R2 can be eliminated if the output impedance of the detector is 100 Ω. In this configuration, C1 and C2 work together to provide rolloff. This configuration provides around 22 dB of attenuation at 10 kHz compared to 12 dB in the first implementation above. The gain of the stage is now R3/(R1+R2) or about 5.5×. Both circuits were simulated using *LTSpice* (free circuit simulation software from Linear Technologies at **www.linear.com/software**), and R3 and C2 were selected (using trial and error) to provide a similar gain peak (13.7 dB vs 13.5 dB) and

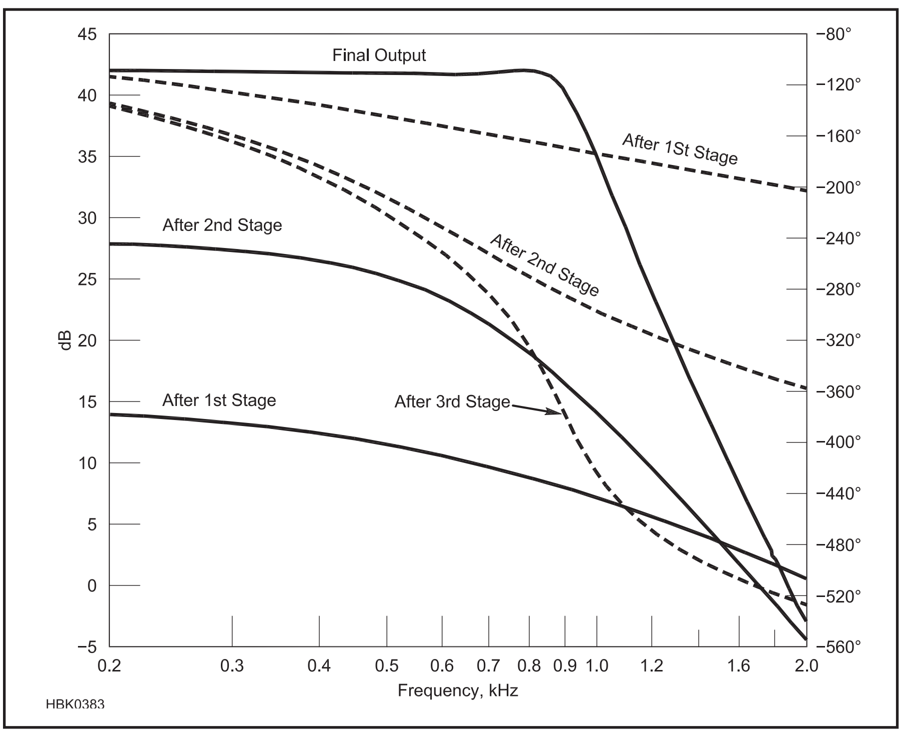



**Fig 11.55 — Frequency response of all stages added one at a time. Dashed lines show phase response.**

a similar gain at the filter cutoff frequency of 780 Hz. (8.6 dB vs 9.2 dB).

**Fig 11.54** is the complete five-pole filter with all necessary biasing and dc isolation, using standard value parts. If a 12-V op amp is used, the 3-V supply voltages shown above will be replaced with 12-V supplies. Remember that the 670 nF capacitors are composed of one 0.47 µF and two common 0.1 µF capacitors in parallel.

The resulting frequency plot (as modeled in *LT Spice)* is shown in **Fig 11.55**. The lowest gain curve is the frequency response of the initial "real pole" section. The gain almost reaches 0 dB by 2 kHz as desired. The next highest gain curve is the frequency response at the output of the second stage (the net response of the first and second stages). Although the total gain is higher, the gain still drops below 0 dB before 2 kHz. The highest gain line shows the total filter response and shows a slight rise near the cutoff frequency. The gain is almost 42 dB (125× voltage gain), close to the desired 40 dB target gain and also drops below 0 dB of gain before 2 kHz.

With this filter placed just after the receiver detector, no signal out of the detector (4 to 5 $V_{P\text{-}P}$) at 2 kHz or higher (just 1.22 kHz above the filter cutoff of 780 Hz) is capable of overloading the front end of the receiver.

### Stage Order

In the example above, the stages were ordered from lowest Q (the real section) to the highest Q (section B). This order gives the best protection from inter-stage overload, but is not necessarily the best order for best receiver sensitivity. Fig 11.55 was generated showing the net frequency response adding one stage at a time. However, **Fig 11.56** shows the frequency response of each of the three stages separately.

Notice that the section responses labeled "Real Pole" and "Section A (Q=0.85)" both lose a lot of gain approaching the filter edge at 780 Hz. This loss of gain means that the resistor noise will have more impact than expected, in effect reducing the receiver sensitivity somewhat. Notice that the highest-Q section, labeled "Section B (Q = 3)" actually has a gain peak near the band edge. A peak like this will be present in any filter stage with a Q higher than 1. The higher the Q, the higher and sharper this peak will be.

The "real pole" section has a simpler configuration that makes it easy to use as the first pre-amp stage of the filter, so it comes first. However, moving Section B from the last stage to the second stage will help overcome the stage gain reduction of the first stage and provide for better receiver sensitivity overall as shown in **Fig 11.57**.

When the three stages are ganged in this

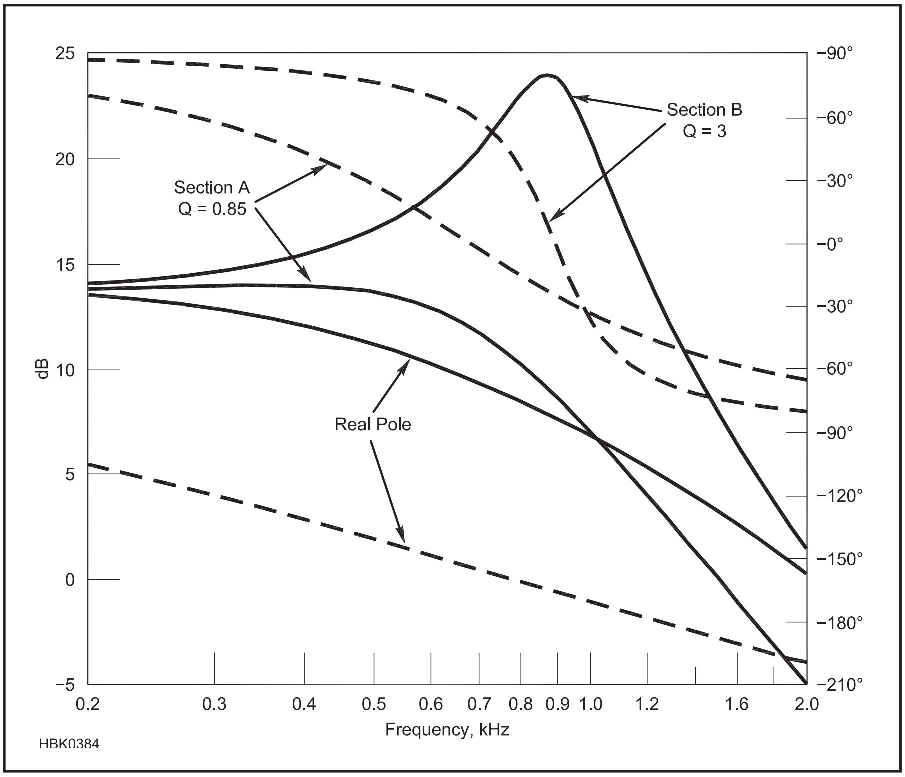



**Fig 11.56 — Frequency response of each filter stage shown separately. Dashed lines show phase response.**

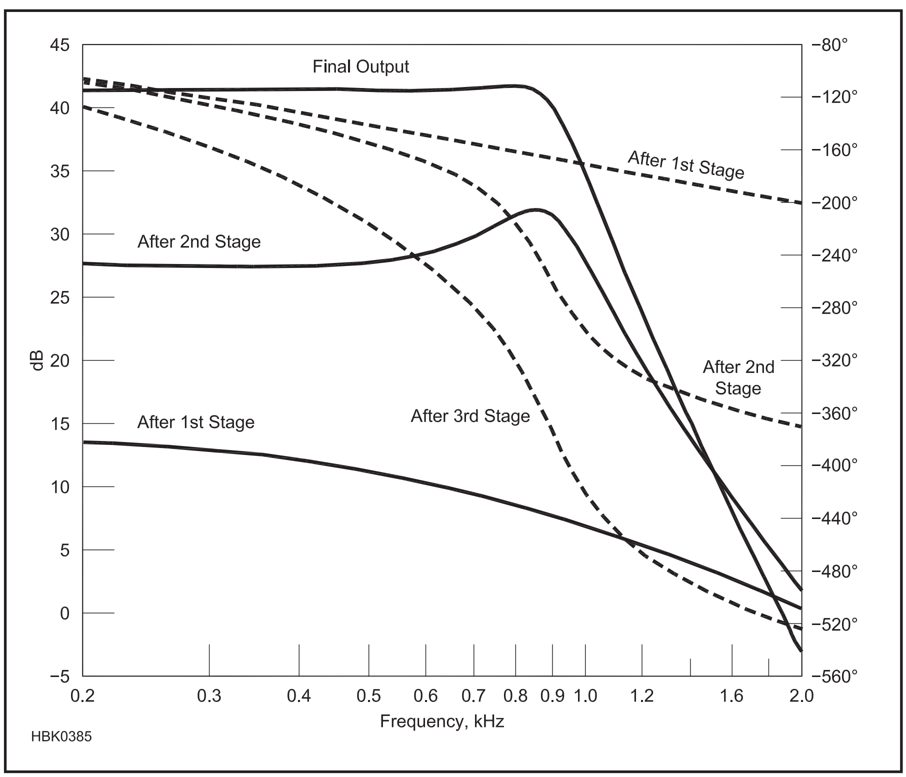



**Fig 11.57 — Cumulative frequency response with the high-Q stage in the middle. Dashed lines show phase response.**

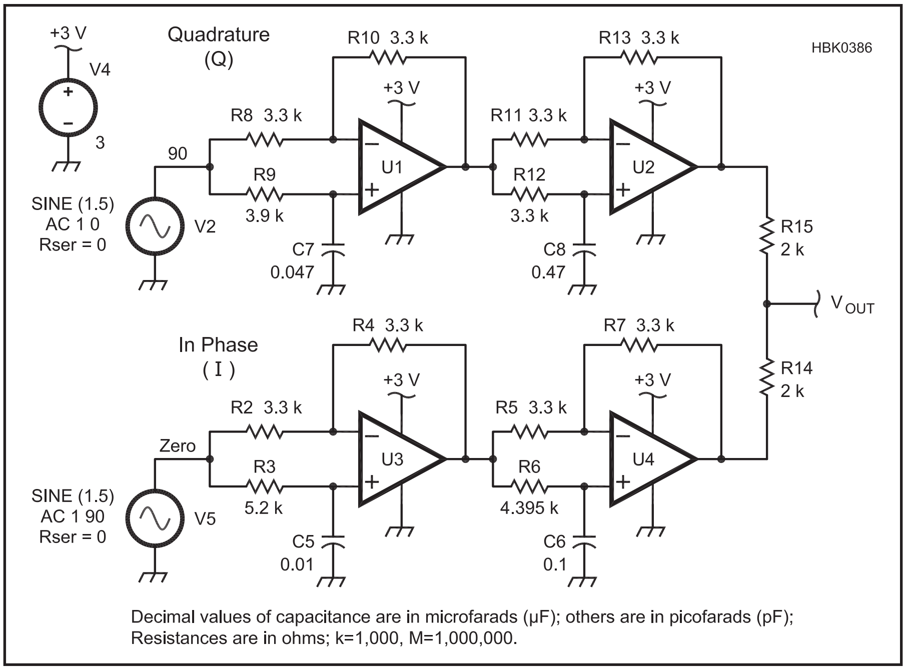



**Fig 11.58 — Two all-pass filters create SSB direct-conversion receiver audio from 300 to 1000 Hz.**

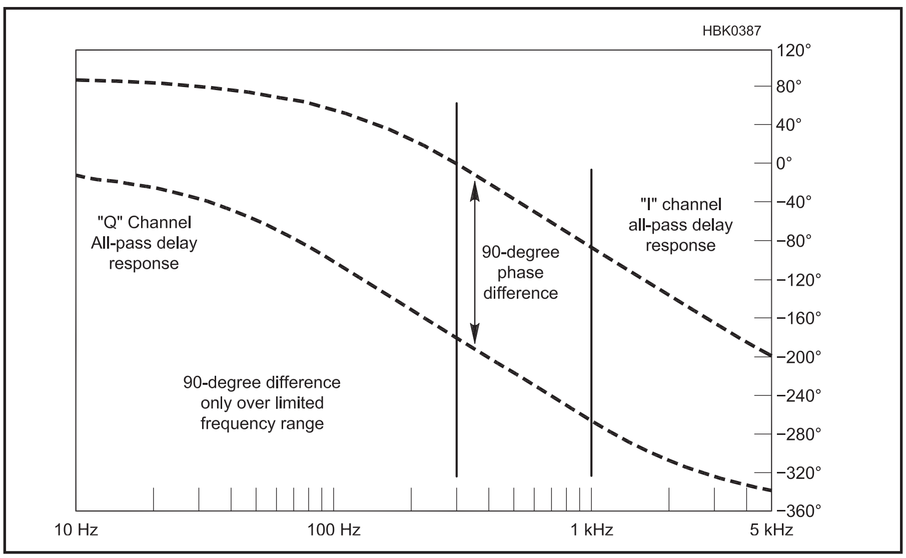



**Fig 11.59 — Differential phase delay for two all-pass phasing sections.**

order — Real Pole, Section B, Section A — the signal out of the second section is higher near the filter cutoff of 780 Hz, allowing for better low-noise performance. As can be seen in the final output curve, the stage ordering does not change the overall filter frequency response.

A drawback of this configuration could be that the second stage of the filter might now be more susceptible to overload (due to the gain peak) from large signals nearer the edge of the filter than before. This is not really the case in this filter, however. Since the filter was designed with roughly equal gain in each stage, an overload to the second stage is automatically also an overload to the last stage, since it applies additional gain to the second filter section output. However, if this filter had been designed with unity gain per section, the ordering of the stages (and stage overload) could be more of a concern.

In summary, for best ultra-low-noise results (which is only important with very small signals) the stages should be ordered:

- Highest Q sections to lowest Q sections
- If odd order, odd Real Section goes first

For best signal overload protection within the filter bandpass (which is only important if there is little or no gain per section), the sections should be ordered from lowest Q (a real pole section) first (if any) to the highest Q section.

If the filter has many sections and has little or no gain per stage, an optimum balance between good low noise results and good internal passband filter overload protection might be to alternate the sections between the lowest Q and the highest Q sections. For example if a 9-pole filter has a real section, and four other filter sections with Qs of 0.6, 0.9, 1.8, and 5.7, the best section order might be: Real Pole section, followed by Q=5.7, Q=0.6, Q=1.8, and lastly Q=0.9. Simulating and studying the combined frequency response using a program like *LTSpice* is very useful in understanding what is going on.

## ALL-PASS ACTIVE FILTERS

A filter that is often seen in phasing-type receivers is the all-pass filter. This filter passes signals of all frequencies without affecting their amplitudes, but creates a controlled phase shift that varies with frequency. This phase shift is used in quadrature direct-conversion receivers for creating single frequency reception by creating a 90° phase difference used to cancel out an unwanted sideband. (See the **Receivers** and the **DSP and Software Radio Design** chapters for more information on direct-conversion receivers.)

The circuit of **Fig 11.58** takes the quadrature I and Q outputs of a direct-conversion quadrature detector (like a Tayloe detector) and adds two all-pass filters designed to create 90° of phase difference between the top and bottom two-stage all-pass sections. U1 and U2 form one phase delay section while U3 and U4 form the other. The 90° difference in phase between these two unity gain all-pass sections is shown in **Fig 11.59.**

In a direct-conversion quadrature receiver, the I and Q outputs are 90° apart from each other and contain the audio from both sidebands. For one of the sideband signals, the I output is 90° ahead of the Q output and for the other sideband signal, Q is 90° ahead of I. Adding in an additional 90° of delay will cause I and Q to both have the same delay on one sideband (–90 + 90 = 0) and 180° of phase difference (90 + 90 = 180) on the other sideband. Since the output is taken via the sum of R14 and R15, a 180° difference will cause the signals from one sideband to cancel, while a 0° difference will allow the signals of the other sideband to add together.

The 90° difference holds well only over a limited range, so suppression of the opposite sideband gets worse at the high and low end of the “sweet spot” where the signals are

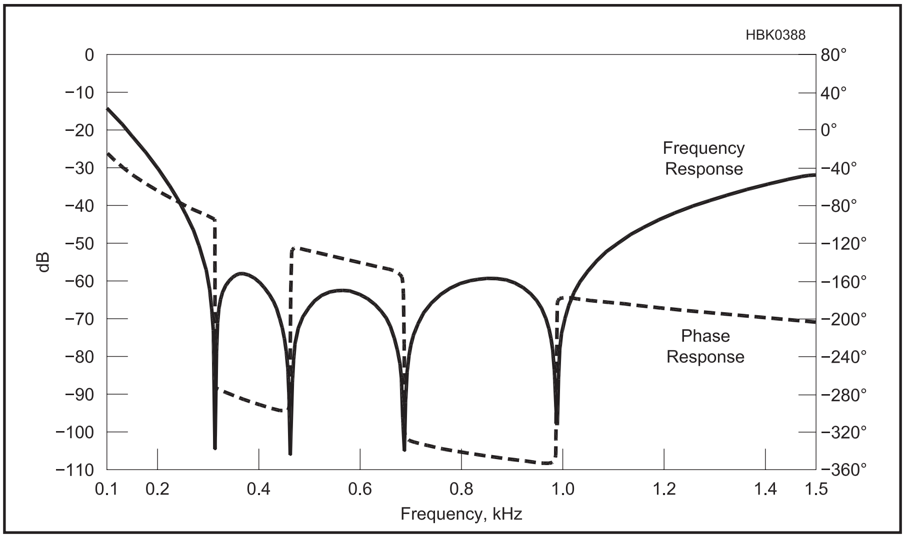



**Fig 11.60 — More than 50 dB of opposite sideband suppression is obtained by using the differential phase shift between two all-pass phasing stages.**

precisely 90° apart and suppression of the unwanted sideband is best. When adjusted properly, opposite sideband suppression can be excellent over a limited range as shown in Fig **11.60** This was a phasing section designed for CW covering 300 Hz to 1 kHz with well over 50 dB of opposite sideband rejection. A 300-Hz high-pass filter and a 1000-Hz low-pass filter can be used to attenuate the high and low frequency ranges in which sideband suppression drops below 50 dB.

It is difficult to hand-pick components to the 0.1% precision needed to get 60 dB of suppression. Small trimmer resistors can be placed in series with fixed values at R3 and R6 in order to allow the filter to be tuned. The book *Experimental Methods in RF Design* discusses all-pass phasing sections in great detail. The program *QuadNet* for designing and analyzing active quadrative networks is included on the *Handbook* CD-ROM.