ADDITIONAL DESIGN IDEAS FOR VOLTAGE-CONTROLLED FILTERS:

- by Bernie Hutchins

You have perhaps noticed that filters are popular - people love to know about them and use them. I have often wondered just why this is that filters seem to have a general appeal, and I am not sure, even though I share this love of filters. It perhaps has something to do with our feeling that all filters (not just electrical) are for the purpose of blocking out something unwanted, and somehow making the world a better place! We learn this when we sift sand for the first time as children. Here we will be taking a look at some more filter ideas, knowing that this will be a popular article, as all filter articles are. I wonder though if filters are always the most efficient way of getting the results we want in electronic music. It seems that when they are used as timbre modulators, there are simpler ways of doing things. We perhaps should look at timbre modulation from a more general viewpoint in a later issue.

Here, we want to tie up a few loose ends regarding four-pole filters, and to look at ways of applying the new Solid State Music SSM2040 IC. We will also be looking at VCF's from the point of view of the front panel (panel space) and from a human engineering viewpoint.

1 CORNER PEAKING OF A FOUR POLE FILTER:

The four-pole low-pass filter is an old friend by now. It is often used in modular synthesizers, and may be the only filter used in some prepatched performance oriented synthesizers. We know about all we need to know about how the basic filter works. See for example the discussion in EN#41, July 10, 1974 (reprinted in the Musical Engineer's Handbook, Chapter 5d). The conventional form of the four-pole filter consists of four single-order sections connected in cascade. This means that the total response is fourth-order and the final roll-off rate is 24db/octave. However, the fact that the sections are individually first-order means that the corner is not going to be very sharp. This is why a regenerative path is almost always added to this filter to add resonant "corner peaking". If we were completely free with our design, we would design a sharp fourth-order filter directly, but since we want to use voltage-control, we have to be restricted to sections that are easy to control. Thus, the four cascaded first-order sections with regeneration are a very reasonable engineering compromise. Here, we will want to look at corner peaking quantatively. Before, we have taken only a qualitative look, and this structure is too important to leave as a loose end.

The fourth-order filter is shown in Fig. 1. Each one of the sections has the same transfer function is the simple RC low-pass shown in Fig. 2, and a typical voltage-controlled realization is shown in Fig. 3.

Fig. 1

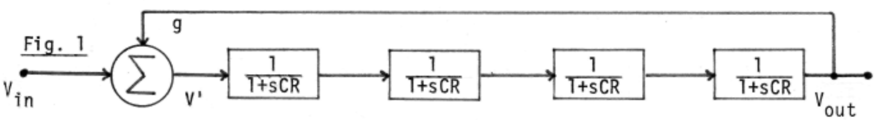


Fig. 2

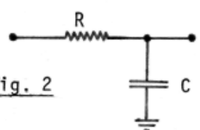
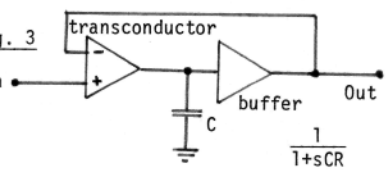


Fig. 3

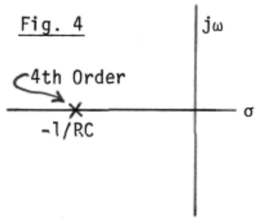


With g in Fig. 1 equal to zero, we have the easy case of four cascaded first-order sections, and the overall transfer function is just the product of the four sections. Thus we have:

$$T_0(s) = V_{out}/V_{in} = \frac{1}{(1 + sCR)^4} \tag{1}$$

This transfer function has a denominator that equals zero when s = -1/RC, and since there is a power of four in the denominator corresponding to this root, it is a fourth-order pole. That is, it is a fourth-order transfer function, and since there is only one pole position, there must be four poles all on top of each other at the position s = -1/RC. The pole-zero plot in the s-plane is shown in Fig. 4. Note that the poles are all real, since they lie on the σ-axis. We will want to take a look at the effect of the regeneration and see how the poles move as g is made to increase in magnitude. We can say that we expect that since we get a sharper corner, some of the poles are going to have to move out into the imaginary region of the s-plane, and up closer to the jω-axis.

Fig. 4



When g is no longer zero, we must allow for the feedback loop. First note that the original chain of four first-order sections is independent of this feedback path - it will just be processing a different input - and it is still true that:

$$V_{out}/V' = \frac{1}{(1 + sCR)^4} \tag{2}$$

Also, from the simple summing condition:

$$V' = V_{in} + gV_{out} \tag{3}$$

Equations (2) and (3) are easily combined to obtain the transfer function with feedback:

$$T_g(s) = \frac{1}{(1 + sCR)^4 - g} \tag{4}$$

At first sight, equation (4) seems to be a simple alteration of equation (1), but we shall see that things are actually very much different. While equation 1 gives a denominator in a nicely factored and useful form, equation (4) is not factored and we have at the moment no way of knowing where its poles are. First, we have to "unfactor" (multiply out) the power of four in the denominator of equation 4 so that we can include the -g term inside. It will also save a lot of notation to measure frequencies in terms of units of 1/RC, so we can set RC = 1. With this, the denominator of equation (1) becomes:

$$(1 + s)^4 = s^4 + 4s^3 + 6s^2 + 4s + 1 \tag{5}$$

and the denominator of equation (4) becomes:

$$(1 + s)^4 - g = s^4 + 4s^3 + 6s^2 + 4s + (1-g) \tag{6}$$

To obtain the poles of equation (4), we need the solutions to the equation

$$s^4 + 4s^3 + 6s^2 + 4s + (1-g) = 0 \tag{7}$$

Unfortunately, this is a non-trivial "quartic" (4th power) equation. Standard handbooks tell us how to solve quadratic, cubic, and even quartic equations, but the quartic soultion is fairly involved, and you could easily spend hours doing "bookkeeping" type algebra only to end up with useless results due to a mistake somewhere up the line. [Look up the procedure in a reference such as Abramowitz & Stegun, Handbook of Mathematical Functions, Section 3.8.3 if you want to see how much of a mess you might be in for]. There must be an easier way. We will use here a principle we have used before called "Never underestimate the power of knowing or suspecting the correct answer." The basic idea here is to guess what two of the complex conjugate poles are, and this gives us a quadratic factor of equation (7) if we are right. We can then divide this quadratic factor into equation (7) to give the remaining quadratic factor, which can easily be factored down to first order with the quadratic equation to give the other pair of poles. That's "guess two" and "get two". First we have to guess at least one set of answers to get some idea where the poles are. We will use something else we know about the network to do this.

We know the performance of the network when $g = 0$, and we also know the general effects of corner peaking. Happily, we also know something about the extreme case of corner peaking - the network will oscillate. In order to see how and why the network can be made to oscillate, we need to consider briefly in review the response of a first-order section. That is, we want to consider the frequency response and phase response of a section with transfer function:

$$T(s) = 1/(1 + sCR) \tag{8}$$

To get the frequency response, we substitute $j\omega$ for s in equation 8, and take the magnitude of $T(s)$, $|T(s)| = [T(j\omega)\cdot T(-j\omega)]^{1/2}$. This gives:

$$|T(s)| = \left[\frac{1}{1 + j\omega RC}\cdot\frac{1}{1 - j\omega RC}\right]^{1/2} = \left[\frac{1}{1 + \omega^2R^2C^2}\right]^{1/2} \tag{9}$$

It is clear that when $\omega = 1/RC$, $|T(s)| = 1/\sqrt{2}$. Also, the phase response is given as the inverse tangent of $\omega/(1/RC)$ which will amount to 45° at $\omega = 1/RC$. Now, it is only necessary to make again a point just made above - the string of four first-order sections is independent of the loop - it always acts as a simple four stage low-pass filter. Thus, for four stages at frequency $1/RC$, the total loss is $(1/\sqrt{2})^4 = 1/4$, and the total phase shift is $4\cdot 45° = 180°$. Hence, a gain of 4 and a 180° inversion is all we will need to sustain oscillation.

This is the one extra data point we need to get going. We know that when the gain $g = -4$, we get oscillation at $1/RC$. Since we are measuring frequencies in terms of $1/RC$ in the main part of this analysis, this means that the oscillation frequency is 1. An oscillation corresponds to two poles that set on the $j\omega$ - axis at the frequency of oscillation and its negative value. Thus we expect that a gain of -4 results in a pair of poles at +j and -j. Substituting $g = -4$ into equation 7 gives:

$$s^4 + 4s^3 + 6s^2 + 4s + 5 = 0 \tag{10}$$

Since we say there should be poles at +j and -j, then $(s - j)$ and $(s + j)$ should be factors of equation (10), and the product $(s-j)(s+j) = s^2 + 1$ also is a factor. Thus, we should be able to divide equation (10) by $s^2 + 1$ and come out even*. In fact, if we do so, the result is $s^2 + 4s + 5$. This in turn can be factor by the quadratic formula into $(s + 2 - j)(s + 2 + j)$. In summation, we can rewrite equation (10) as:

$$(s - j)(s + j)(s + 2 - j)(s + 2 + j) = 0 \tag{11}$$

We can now list all four poles of the oscillating network:

$$s = +j$$
$$s = -j$$
$$s = -2 + j$$
$$s = -2 - j$$

*If you have forgotten how to do this, see AN-57.

We can now plot the poles for two extreme cases of feedback: no feedback ($g = 0$) and feedback for $g = -4$, corresponding to oscillation. These pole positions are shown in Fig. 5. We note that as expected, there are poles on the $j\omega$ axis that cause oscillation. The new thing we learn is the position of two other poles in the case of maximum feedback. Clearly all four poles have separated from the -1 point and split to four different corners in a symmetric pattern. We have not yet determined how the poles migrated from -1 to their new positions. Again, we will want to do some guessing, but here we just show the migration by wavy lines as a confession of ignorance.

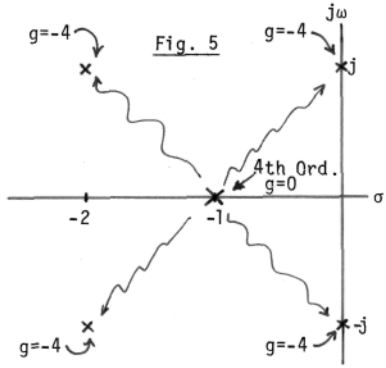


Fig. 5

So how did the poles get from -1 to the four corners? One answer we can give is "smoothly" because we know from our experience with the resonance control on four-pole filters that the response is continuous and does not jump in sudden steps. We can make guesses about the path (I guessed first that they followed circles, but that was wrong.), but there is another way available to us if needed. This is to observe that since the poles start at -1, and two of them end up on the $j\omega$-axis, then they must have crossed the line $\sigma = -0.5$ (or any similar choice). Thus, there must be a solution (for a certain value of g) that places poles at $-0.5 \pm bj$, where b is not known yet. Such a pair of poles result in a quadratic factor $s^2 + s + (0.25 + b^2)$. We can then divide this quadratic factor into equation (7). When we get to the bottom of the division, terms in s must cancel as must the constant term, and since we can adjust b as needed, and g is also a free choice, this can be done. This is illustrated below:

$$
\begin{array}{rllll}
 & s^2 + 3s + (2.75 - b^2) & & & \\
\hline
s^2 + s + (0.25+b^2) \,\big) & s^4 + 4s^3 & + 6s^2 & + 4s & + (1-g) \\
 & -s^4 - s^3 & -(0.25+b^2)s^2 & & \\
\hline
 & 3s^3 & +(5.75-b^2)s^2 & + 4s & \\
 & -3s^3 & -3s^2 & -3(0.25+b^2)s & \\
\hline
 & & (2.75-b^2)s^2 & + (3.25 - 3b^2)s & + (1-g) \\
 & & -(2.75-b^2)s^2 & - (2.75 - b^2)s & -(2.72-b^2)(0.25+b^2) \\
\hline
 & & (0) & (?) & (??)
\end{array}
$$

In order for the s term to cancel out, it is necessary that:

$$3.25-3b^2-2.75 + b^2 = 0.50 - 2b^2 = 0 \qquad \text{or} \quad b = 0.5 \tag{12}$$

In order that the constant term cancel out, it is necessary that:

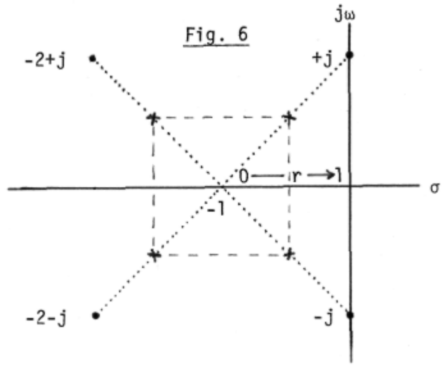
$$(1-g) - (2.75-b^2)(0.25+b^2) = (1-g) - (2.5)(0.5) \qquad \text{or } g = -0.25 \tag{13}$$

With b determined as 0.5, we place our first set of poles at $-0.5 \pm 0.5j$. Also, the second quadratic factor just determined above is $s^2 + 3s + (2.75-b^2)$ becomes $s^2 + 3s + 2.5$, which by the quadratic formula gives poles at $-1.5 \pm 0.5j$. These poles lie on the straight lines that would connect the original poles at -1 to the final poles in the corners in Fig. 5. That is, it appears that the wavy lines we drew should really have been straight. By repeating the procedure, it is possible to show that the lines (pole loci) are really straight (or that we are very unlucky!).

The migration of the poles is now determined at least in that we know the paths and the limits of these paths. What remains to be done is to find a relationship between pole position and the feedback factor g at points other than those we have just tested individually. Eventually, from the pole positions we expect to learn about the characteristics of the frequency response.

To determine the exact relationship between the feedback factor g and the pole positions, we first observe that we have shown that the poles move out from -1 in a square pattern. It is convenient to define a new "r" (resonance) variable to measure the size of the square as in Fig. 6. The value of r runs from 0 to +1 as g goes from 0 to -4. In terms of r, the four poles are located as follows:

Fig. 6



$(-1 + r + jr)$

$(-1 + r - jr)$

$(-1 - r + jr)$

$(-1 - r - jr)$

Knowing all four poles in this way allows us to reconstruct the denominator of the transfer function as the product of all (s - p) where p is the pole. Thus, the donominator is:

$$(s + 1 - r - jr)(s + 1 - r + jr)(s + 1 + r - jr)(s + 1 + r + jr) \qquad (14)$$

We can multiply this out, keeping only constant terms, and comparing with equation (6), this can be set equal to $(1 - g)$. The constant terms that are obtained in the multiplication of equation (14) are $4r^4 + 1$. Hence $1-g = 1 + 4R^4$ and:

$$g = -4r^4 \qquad (15)$$

It is the fourth power in equation (15) that accounts for the fact that in our example with poles at -0.5 (half way across), g was only -0.25, while for r = 1 (all the way across to oscillation) g had to reach -4. For convenience, a plot of the required value of g as a function of r is given in Fig. 7. Later we will want to look at this when selecting a means of controlling the value of the feedback g in a practical module.

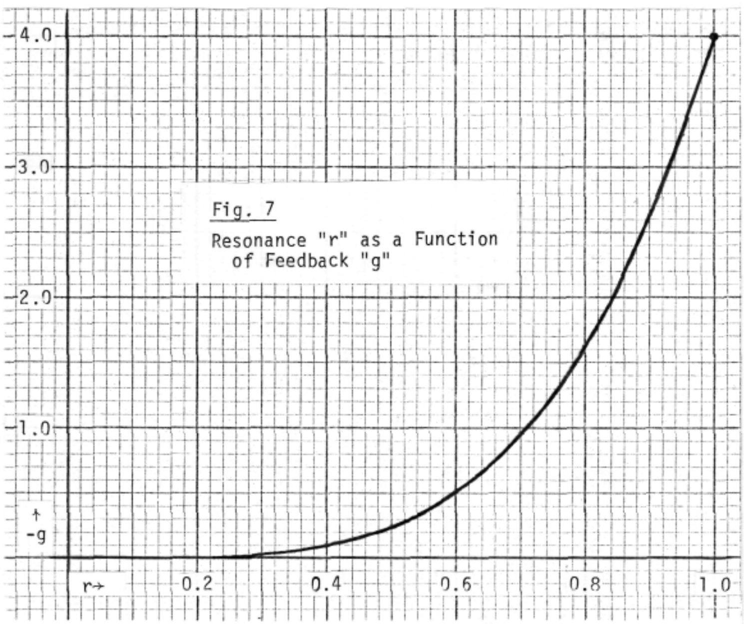


Fig. 7

Resonance "r" as a Function of Feedback "g"

The frequency response of the four-pole filter with feedback can be determined from the pole positions which we have just determined. For example, with g = -2.63, the poles are positioned at r = 0.9 as shown in Fig. 8. A graphical method can then be used to determine the frequency response. First we select any one point where we want to know the magnitude of the frequency response (for example, ω_0 in Fig. 8. We then measure the distances from the poles to the point ω_0. These are shown as dotted lines in Fig. 8. The final step is to divide 1.00 by the product of these four distances. This is repeated until enough points are obtained to sketch in the full curve. For more information on this method, see AN-45.

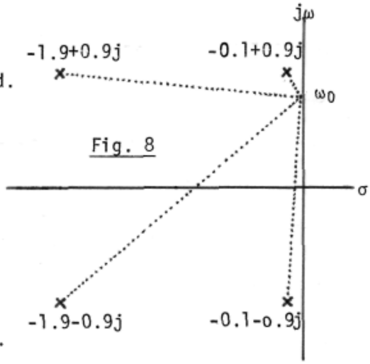


Fig. 8

Fig. 9 shows a comparison of a theoretical (graphically calculated) frequency response and an experimentally measured response corresponding to the pole positions shown in Fig. 8. Note that the high peak at a frequency of 0.9 is due mainly to the close approach of the pole at -0.1+0.9j to the $j\omega$ axis. The peak shown illustrates the meaning of "corner peaking." Note that the peaking is quite extensive (this is a log plot) and in many ways the response resembles a bandpass response.

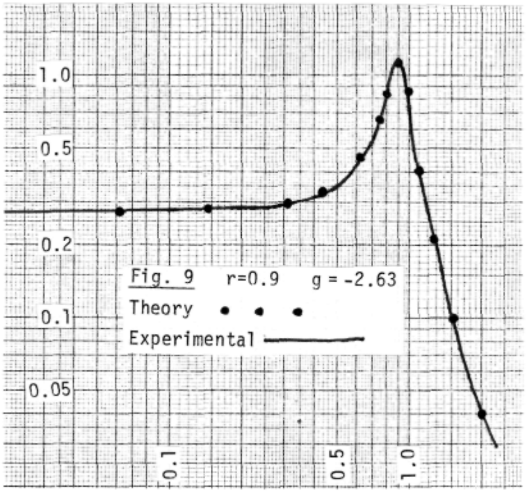


Fig. 9 r=0.9 g = -2.63

A better feeling for the effect of corner peaking can be obtained from a study of Figures 10 through 13, which show the response for increasing values of r. Fig. 10 shows the response when g = 0, and is just the expected flat roll-off with a corner that is not too impressive compared to a Butterworth corner, for example. We can begin to see the final 24db/octave from Fig. 10, but there is still a ways to go. Fig. 11 shows the effect of feedback of -0.25 corresponding to r = 0.5. While the poles have moved half the way to their oscillation positions, there is relatively little change in the curve, as can be seen by the dotted curve representing the original g = 0 curve. Note that feedback does lower the DC response level. Fig. 12 shows the curve for a value of r = 0.75 (g = -1.26) and we see a much improved corner with some ripple and overall gain loss. Probably, this reminds us of a second-order Chebyshev as much as anything, and in fact, with two poles moving to the rear, the close-in part of the curve is very similar to a two-pole system. Fig. 13 shows even more feedback (g = -2.63, r = 0.9) and a response that is looking less and less low-pass and more and more bandpass. Beyond r = 0.9, we are mainly working with a high-Q response very much like a bandpass.

From Figures 10 - 13, we can see that there is relatively little change that we might expect to become important musically until a value of r that approaches 0.75. This requires feedback of -1.26 or greater. The significance of this is that in some feedback schemes we are concerned with relatively small amounts of feedback, and therefore employ such things as log pots and exponential responses

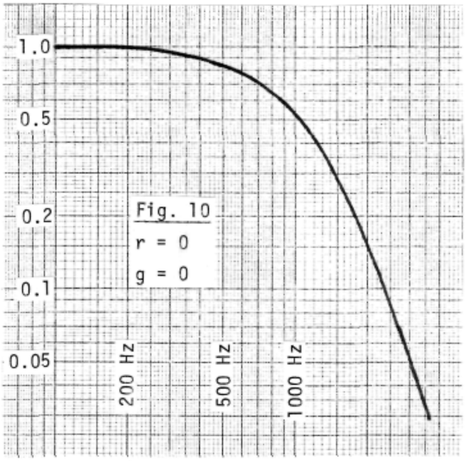


Fig. 10

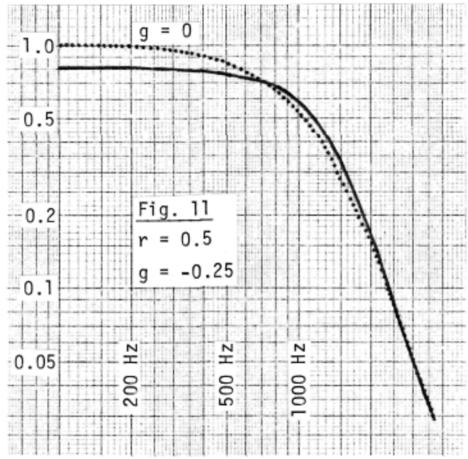


Fig. 11

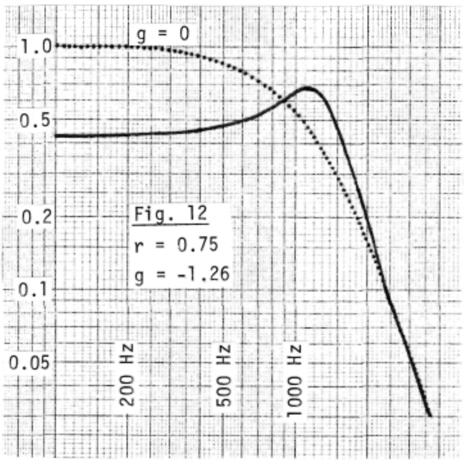


Fig. 12

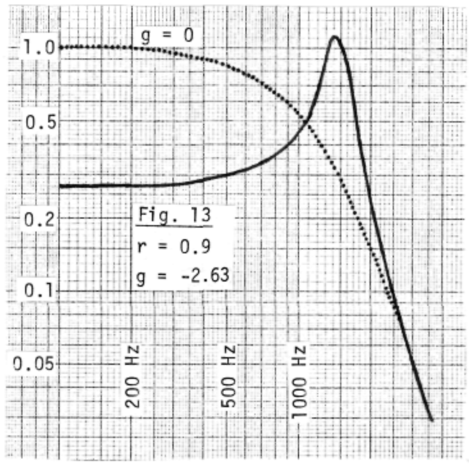


Fig. 13

to obtain very fine control. Here, we are mainly interested in a range from 25% to 100% of the maximum value. A log pot, which gives 15% resistance change for the first 50% rotation would be a waste here. In fact, even the bottom 25% of a linear pot would be wasted. Thus we might even suggest a resistor in series with the pot to raise the whole thing up some, and a "reverse audio" pot would seem ideal for controlling resonances in the range of 0.9 to 1.0 where things change very rapidly as far as the response is concerned. Some suggested feedback circuits thus appear as seen in Fig. 14. In the first, a resistor has been added to a linear pot. In the second, a bypass resistor is used (assuming feedback to a summing node). The final circuit uses a reverse audio pot to control feedback. The reverse audio pot is a backward log pot - most of the resistance change is in the first 50% of clockwise rotation.

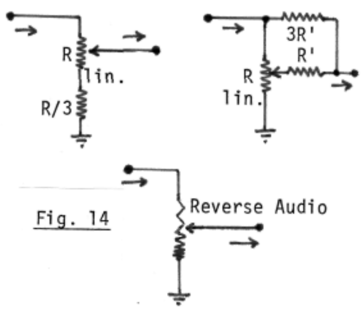


Fig. 14

SUMMARY OF FINDINGS: The effect of the feedback or resonance control of the usual form of the traditional electronic music four-pole filter is to spread the four poles, which are initially piled up at -1/RC, out in a square pattern. The frequency response of the filter is understood in terms of the new pole positions. While a small amount of feedback results in a substantial displacement of the poles (Fig. 7), this large displacement has a relatively small effect on the frequency response curves (Fig. 11). Large amounts of feedback result in relatively little additional pole displacement, but more substantial changes in the frequency response curves (Figures 12 and 13). The implication is that controls used to set the resonance of the filter should have special features to increase resolution at feedback levels between 25% and 100% maximum (just the opposite of a log pot, for example).

2 OBTAINING BANDPASS AND HIGH-PASS FUNCTIONS FROM A FOUR-POLE LOW-PASS:

One of the big "selling points" of the state-variable VCF is that it provides several modes of operation (low-pass, bandpass, high-pass, and notch) as simultaneous outputs. Here we will show that it is possible and easy to obtain a full range of functions from the basic four-pole low-pass. These functions are fourth-order in this case, although the method is general and can give third, second, or first-order as well.

It is obvious that we can "tap" any of the four first-order sections of the four-pole low-pass to obtain a first, second, third, or fourth-order low-pass response. What we will show is that it is possible to do a weighted sum of all four taps plus the input to obtain a general fourth-order numerator for the transfer function, retaining the original fourth-order denominator. The summing network is indicated in Fig. 15. From the diagram, we get:

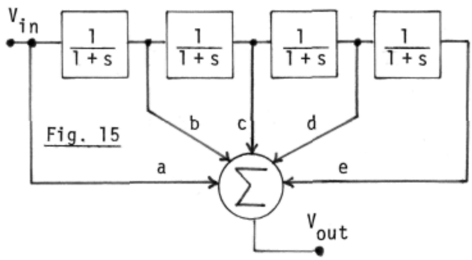


Fig. 15

$$T(s) = V_{out}/V_{in} = a + b/(1+s) + c/(1+s)^2 + d/(1+s)^3 + e/(1+s)^4 \qquad (16)$$

$$= \frac{as^4 + (4a+b)s^3 + (6a+3b+c)s^2 + (4a+3b+2c+d)s + (a+b+c+d+e)}{(1+s)^4} \qquad (17)$$

It is clear that by properly selecting values for a, b, c, d, and e, we can obtain any fourth-order numerator we want. A trivial example is when a, b, c, and d are zero and e = 1, in which case, we have our fourth-order low-pass back. We will want to look at how we can obtain the necessary weightings, and then will want to examine the effect of corner peaking of the low-pass "backbone".

Probably the most interesting case we have is the conversion to a four-pole high-pass. To get a fourth-order high-pass, we need to get an s^4 in the numerator, an all else must go. You can set up equations if you wish, but you can probably just see that a should equal 1 and 4a+b should equal zero, hence b = -4, and so on. The results for high-pass are:

$$a = 1 \qquad b = -4 \qquad c = 6 \qquad d = -4 \qquad e = 1 \qquad (18)$$

which converts equation (17) into:

$$T(s) = \frac{s^4}{(1 + s)^4} \qquad (19)$$

To test the theory, we can use a non-voltage-controlled setup of the type shown in Fig. 16. The first-order sections have transfer function -1/(1+s). It is simplest to just sum into an inverting summer, and the inversion in the transfer function automatically makes every other coefficient negative, exactly what we need.

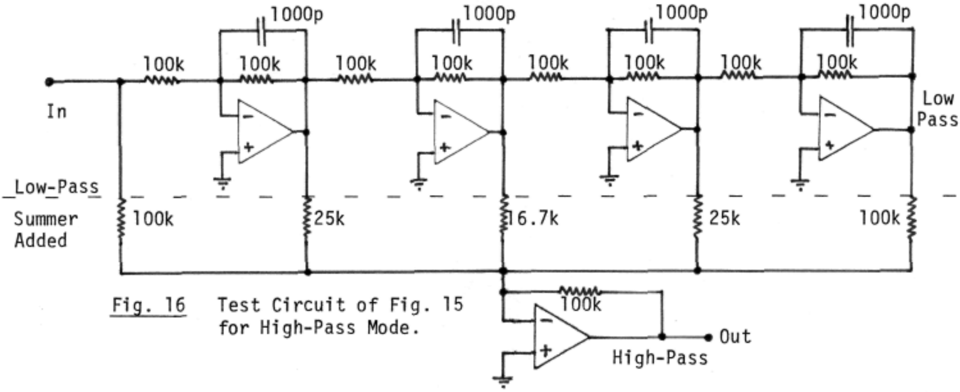


Fig. 16 Test Circuit of Fig. 15 for High-Pass Mode.

Fig. 17 shows experimental data on the filter of Fig. 16. The dotted line shows the low-pass output of the original section. The solid line is the experimental data on the high-pass output while the dashed line shows a high-pass roll-up that we would have hoped to obtain. The reason it was not obtained is almost certainly a lack of resistor precision (5% resistors were used). The solid part of the curve in itself is sufficient to confirm the basic operation of the circuit according to theory.

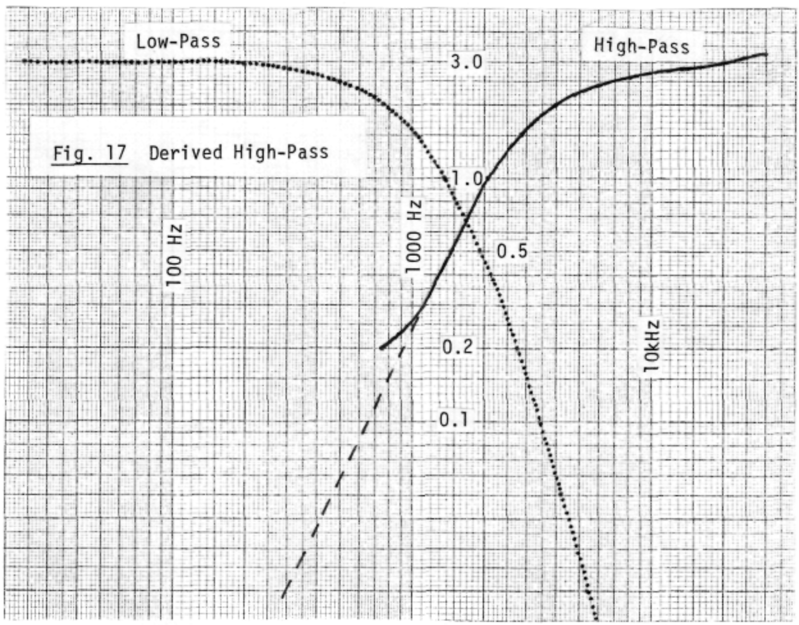


Fig. 17 Derived High-Pass

We next want to see what happens when feedback is added to corner peak the original low-pass filter. We want to know what effect this has on the derived high-pass. This was examined here mainly by experiment. Fig. 18 shows a set of experimental results for a feedback gain of -1 on the low-pass. The low-pass peaking is indicated by the dotted curve. Note that the high-pass (solid curve) is relatively unchanged from Fig. 17. Again, in Fig. 18, the dashed curve shows the results we expect if better resistor matching is used. Fig. 19 shows similar results for a gain of -3, which corresponds to pole migration in the low-pass to 93% of their oscillation positions. The low-pass curve is shown as the dotted line while the high-pass curve is the solid line. We note a sharp peaking in both curves, but the two response curves are not symmetrical. The higher shelf region of the high-pass as compared to the low-pass can be understood in terms of the zeros that have been added to the pole-zero plot. The s^4 term in the numerator of T(s) means that there is a fourth-order zero at s = 0. It is these zeros which kill off the response at low frequency. The response thus builds as the frequency increases away from zero. Once we have passed the dominant pole which causes the peaking, we still see the effect of the increasing distance from the zeros at s=0. The small circles on the solid curve of Fig. 19 indicate points that were calculated using the poles of the low-pass corner-peaked filter, and adding in the four zeros at s = 0.

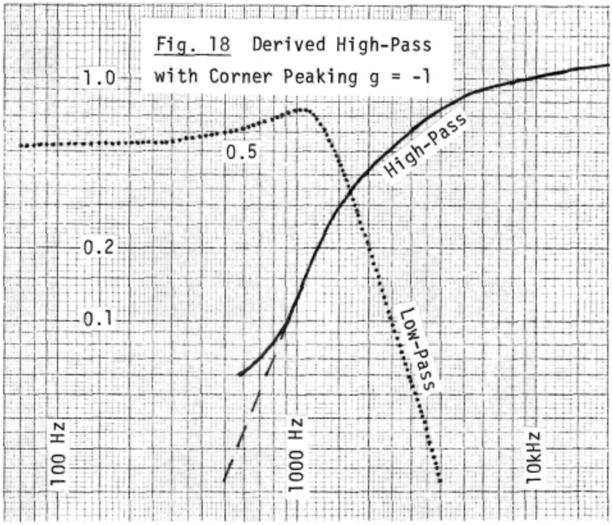


Fig. 18 Derived High-Pass with Corner Peaking g = -1

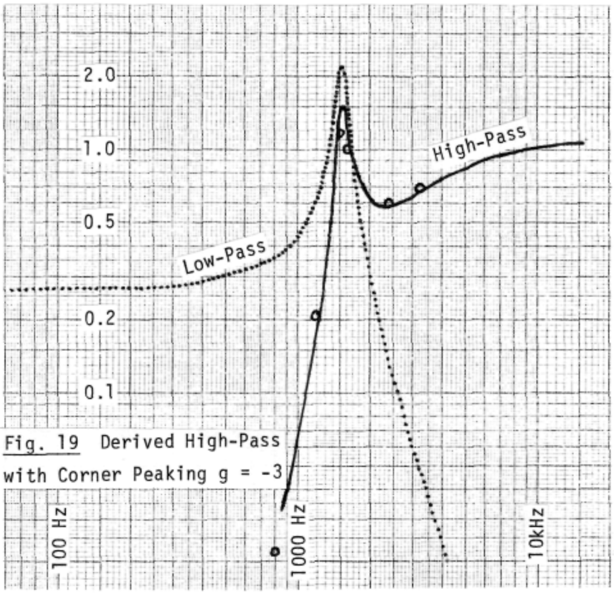


Fig. 19 Derived High-Pass with Corner Peaking g = -3

Another response function that is of interest to us is the bandpass function. This will be obtained if we can set a numerator to s^2 in the transfer function and get rid of everything else. The set of summing parameters that will achieve this are:

a = 0

b = 0

c = 1

d = -2

e = 1

It may bother you a little that a and b are zero so only three stages are tapped, and you might wonder how the filter knows it is fourth order if it only has contact with three stages, and not five. The answer to this seems to be that there are two stages of low-pass filtering that have an effect before the taps are reached. If it were the case that the first three stages were tapped, not the last three, then the filter could not be fourth order. It is a simple matter to set up a circuit similar to that in Fig. 16 to realize the necessary summing network, and experimental results are shown in Fig. 20. Ignore for the moment the g = -3 curve. The shallower curve should be a bandpass rolling off on each side at 12db/octave, because this is a fourth-order bandpass. Both sides are sharper than 6db/octave (a 45° angle), although it is a little difficult to be sure exactly what they are. The g = -3 curve shows the effect of corner peaking. Note that the lower side of the curve sharpens considerably while the upper side remains the same.

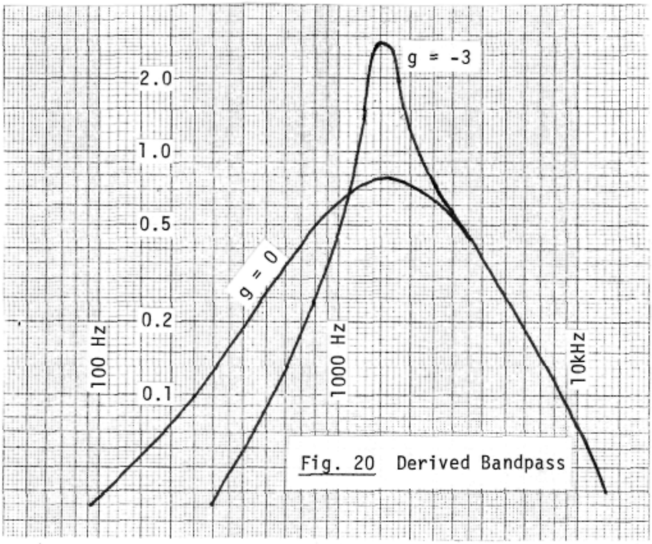


Fig. 20 Derived Bandpass

zeros

We have shown that the basic theory seems to work out in practice. In addition, a notch response and any number of other special responses can be obtained. A notch with second-order ~~poles~~ at +j and -j can be obtained with a numerator $s^4 + 2s^2 + 1$ obtained with a =1, b = -2, c = 2, d = -2, and e = 2. However, with this notch, as with most high order notches, component precision must be very high or the notch washes out. Our experiemntal data with 5% resistors gave a notch only down by a factor of 4.

We should mention that this same method works for second order as well (and can probably be extended for higher orders). For second order, there would be two low-pass stages and three summing coefficients a, b, and c. For high-pass, a = 1, b = -2, and c = 1. For bandpass, a = 0, b = 1, and c = -1. For low-pass, a = 0, b = 0, and c = 1. For notch, a = 1, b = -2, and c = 2. See AN-71, to be published.

The idea of feedback to the input from the summed output rather than from the low-pass output of course comes to mind. In the high-pass case, it does not work because as we have discussed before, all high-pass filters come down somewhere, and there is usually associated phase shift which will meet all the conditions of the low-pass filter as an oscillator. Hence we get oscillation at high frequency. In the case of the bandpass, it is possible to feed back the summed (bandpass) output to get corner peaking. Our experiments showed that the results were much the same as with the low-pass peaking.

3 PRESENT DAY VCF DESIGN OPTIONS

For many years, the four-pole low-pass and the state-variable VCF's have been the principal choices for electronic music. Filter structures that combine the main advantages of the state-variable (multi-function) and the four-pole low-pass (fourth-order response) are attractive new options. The state-variable and the four-pole low-pass are well understood (Musical Engineer's Handbook, Chapter 5d, and elsewhere in back issues of this newsletter: in particular, EN#71 on the ENS-76 VCF options). We have also demonstrated that a two-section state-variable approach can be used to combine advantages of state-variable and four-pole (EN#58). In the present report, we have shown that a four-pole approach can be used to arrive at multi-functions. In many ways, we seem to have two roughly equivalent choices for developing a new type of filter. [There is also the "dual-shift" filter that was discussed in EN# 81]. In some cases, we can get help in choosing between design options by considering different amounts of hardware required, but here, all the methods we have considered require about the same amount of hardware, so we must look for more subtle differences in making a choice. It will be simplest to just tabulate the design factors that are different for the different approaches and then discuss some of the more important differences.

	Two State-Variable	4-Pole plus Summers	Two Dual-Shift
Principal Advantage	Low Sensitivity	Parallel Outputs	Possibly less phase shift in controlled stages
Principal Drawback	Need for Switching	High Sensitivity	Need for Switching
Response Curves	Regular Shapes	Some Irregularities	Regular Shapes
Control of Resonance	Double Control	Single Control	Double Control

First, we can look at the principal drawbacks. With the state-variable, which normally has parallel outputs (simultaneous, LP, BP, HP and notch), when we try to cascade two second-order sections, we have to switch the appropriate output of the first section to the input of the second. Thus, there is a need for a special switch (probably two-pole, 7-position, see EN#84 (17)). With the 4-pole and summers, all outputs can be parallel. The principle drawback of the 4-pole is that when summers are used and when we go to fourth-order, the component sensitivity of the high-pass, bandpass, and notch outputs is quite high, and precision components may be required. The state-variable on the other hand is quite insensitive.

A lesser drawback of the 4-pole with summers is that the response shapes are not symmetrical when resonance is added. See for example, Fig. 19 of this report. A lesser drawback of the state-variable is that both sections must have individual resonance controls. Since we probably want voltage-controlled resonance with this VCF, this means another transconductor is required, or with manual control, a dual pot, which is probably more of a problem than the transconductor.

In many cases, we are very much interested in the amount of panel space that will be required for a given design. In such cases, a filter with switched modes and a single output (two positions on the panel) may be more useful than a filter with parallel outputs (up to 8 positions). This might tend to favor the two state-variable design, but of course, we can use a switch and single output with the four-pole design as well, and the switch may be simpler in the four pole case.

Another point about panel space for VCF's is that we may be able to implement our frequency controls in a manner different from what seems to be necessary for VCO designs. We may not need a fine frequency control for the VCF, although this seems essential for VCO's. Where panel space is really at a premium, a multi-turn pot may be used for frequency level, coarse and fine together.

A final point about panel space and VCF's is that the VCF is generally just going to require much more space than other modules. This is sure to be true if voltage-controlled resonance is used. Some users also find a second envelope input to a VCF useful. Thus, if you have to cut back on panel space, you must be willing to cut back on VCF features as well.

Let's take an example where you have available to build a filter four transconductors all controlled in parallel, and as many op-amps for buffers or summers as you need. This could be a SSM2040 chip and a few extra op-amps, or it could be formed from CA3080's and other individual IC's. The problem is to come up with a useful fourth-order filter. If we have no regard for panel space, there are many things we can do, and the designer probably has no problems in such a case. Here we want to look at designs that will use a small amount of panel space. Two possible setups will be described. Fig. 21 shows a two state-variable approach. Here we avoid switching

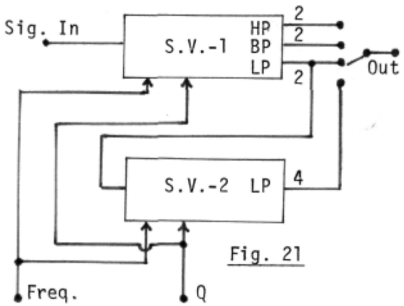


Fig. 21

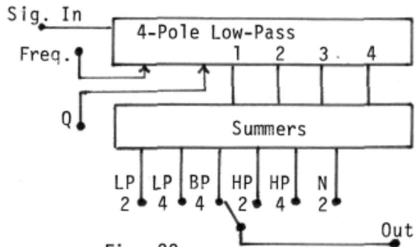


Fig. 22

on the input of the second stage by just feeding in the LP from the first stage. With this arrangement, we have a regular state-variable (second-order) LP, BP, and HP, and a fourth-order LP as well. The output panel space requires only two positions, a rotaty switch and a jack. The second example is seen in Fig. 22 and is based on the four-pole low-pass. Here we have provided summers for second and fourth-order low-pass and high-pass. A single bandpass output (either second or fourth-order) seems sufficient because the Q of the filter is variable. A second-order notch is used because the fourth-order notch is to sensitive to component tolerance. As with Fig. 21, the output panel space is only two positions, a switch and a jack.

At some point, we have to call a halt to electrical engineering to consider some musical engineering and some human engineering. What do musicians need in filters to make their music. For musicians where timberal control is not a big part of their music, relatively little is required of the filters. Musicians who do use timber variations as an important part of their music will require a lot of their filters, as will musicians doing "imitative synthesis" of traditional sounds. But regardless of their demands, they will describe their filters in terms of a "good feel" and a "good sound" and probably only resort to terms like "four-pole" as a necessary "buzz-word" to keep the engineer's attention. Things like the shape of the frequency response curve take a back seat to the feel and the sound and the ease with which the musician is able to find the right knob. This is not an easy job by any means. Probably the most important thing that new VCF designs and VCF chips can do for us is to make the electrical engineering simple enough that we can consider some of the human engineering aspects of filter design. We can begin to do things that are not <u>just</u> reasonable electronically. In using VCF chips for example, we need not exploit all the capabilities of the device (as in Fig. 21) but can use the chip because it saves effort.