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NEWSLETTER OF THE MUSICAL ENGINEERING GROUP

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GROUP ANNOUNCEMENTS:

CONTENTS OF EN#141

We are always looking for material contributed by our readers. For one thing, this makes our publishing efforts considerably easier, but perhaps more importantly, it adds substantially to the variety of information we present. From time to time various authors contribute more or less material for various understandable reasons. However, recently I heard from a potential author who did not want to submit some material because he felt that it was not up to our standards. Well, actually I was not aware that we have any official standards, but I guess with so many issues in print it is understandable that a reader could suppose that there is a certain type of material that we cover. It also might be supposed that there is a certain technical level that is required.

In a sense, we are a victim of and a slave to our past 140 issues. Our readers perhaps feel that each new issue should equal or top the previous one. They do not take this view in a demanding sort of way, but in an expecting way. We ourselves perhaps take a similar view in that, say, each six issues taken as a whole should substantially add to the main thrust of whatever it is we suppose we are doing.

While these points are understandable, we also need to point out that any degree of success we have achieved is due in large part to the interaction we enjoy with our readers. Our readers can often write and even phone us. We devote a good amount of space to reader's questions, and often follow up on reader's suggestions for articles. And, very importantly, our readers submit materials for publication. We need these infusions of new ideas by all routes.

It is important that readers feel free to submit material, regardless of whether they feel it is our usual fare, and regardless of technical level. If the reader finds his or her own material personally interesting, and also finds Electronotes interesting, it is logical that other readers will find that material worth looking at. It is the usual case that some new and different item sparks favorable comments in reaction.

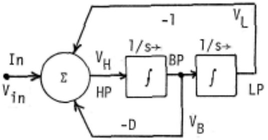
Do we have any standards? Certainly - integrity in its various forms. Material should be completely and honestly presented. Speculation is fine, noted as such. Ideas derived from the work of others should be acknowledged. Circuits presented should are considered to be built and tested unless otherwise noted. In short, try to present your ideas and findings in a manner as though you were trying to pass the information along to a best friend for his maximum benefit, transferring your mental exercises for his utilization.

Newcomer's Page: State-Variable Filters:

- by Bernie Hutchins

At first sight, the state-variable filter looks like an ideal electronic music module. It is relatively simple and inexpensive, and it does more than one thing. We always like our modules to perform several functions if possible. The state-variable filter offers three (sometimes four) filter functions simultaneously. It has a single input, and three outputs corresponding to high-pass, bandpass, and low-pass functions, all second-order. By summing the high-pass and the low-pass, a notch response can also be obtained. In addition, the state-variable filter is easily voltage-controlled, and it's Q can be controlled independently of its frequency. If you were only going to have one filter in your system at first, you would probably choose the state-variable. The only major competitor to the state-variable in electronic music is the four-pole low-pass. Low-pass filters have found a good deal of use in imitative (of acoustic instruments) synthesis. For this reason, some users use mainly a low-pass function, and they may prefer the sharper cutoff of the fourth-order four-pole low-pass.

The structure of the state variable filter is as sketched at the right. It contains three basic elements: two integrators and a summer. The summer is usually represented by a Σ sign and the integrator by an ∫ sign. Numerous variations on the basic state-variable filter are known, but in the usual form shown, the high-pass is the output of the summer while the bandpass and low-pass are integrated and double integrated versions of the summer output. There are two feedback loops around the chain of three elements. From the bandpass, there is a feedback path -D, while from the low-pass, it is a feedback -1. The center frequency of the filter is determined by the time-constants of the integrator, while the Q is determined (nominally) as 1/D. In order to make the filter voltage-controlled, we have only to make the integrator time constants voltage-controlled, and this is usually done with an operational transconductance amplifier.



Unlike with many other filter structures, we can often work with a very general model of the state-variable filter and derive performance parameters from the model. We do not always have to do a full analysis with R's, C's, and op-amps. For example, we have related above the equivalence of the Q of the filter to the reciprocal of the feedback path from bandpass to the summer output. Thus, if we can determine this path value, we can get the Q without an analysis of the structure.

An analysis of most state-variable structures, whether from a basic model or at the component level, is often similar. For example, in the basic model above, we can identify an integrator with its Laplace transform equivalent, 1/s, and from this find the output of the summer. Thus, we would have:

$$V_H = V_{in} - V_L - DV_B$$

At the same time, the 1/s representation of the integrators relates V_L and V_B to V_H, and we have:

$$V_H = V_{in} - V_H/s^2 - DV_H/s$$

This in turn gives us the so-called transfer function as:

$$T_H(s) = V_H/V_{in} = \frac{s^2}{s^2 + Ds + 1}$$

and the other functions $T_B(s)$ and $T_L(s)$ follow from the integrator. In an actual circuit, the analysis is similar, except, for example we would probably not have just 1/s for the integrators, but 1/sCR or -1/sCR, so the frequency is "denormalized" and we see how it is set and controlled.

While state-variable filters are easy to analyze in theory, in some practical cases we need to be careful of subtle effects.

Some New Results Concerning Q-Enhancement In OTA-Based VCF's:

-by Bernie Hutchins

INTRODUCTION:

For many years now we have been using the state-variable Voltage-Controlled Filter (VCF) controlled by Operational Transconductance Amplifiers (OTA's) as one of our most important VCF designs. Such state-variable filters show a phenomenon known as "Q Enhancement." This occurs due to phase shifts across various IC's in the circuit, and manifests itself as a higher Q as frequency increases. For example, the VCF might be set for a Q of 20 at some low frequency, and as the frequency is then raised into the range of a few thousand Hertz, the Q goes up to say 25 or so, and this continues until oscillation occurs at some still higher frequency. Thus it is desirable to compensate for this frequency dependant Q somehow, so that Q is constant as frequency changes. The control and compensation of Q enhancement in fixed frequency active state-variable filters is well studied. Similar compensations in variable frequency filters (VCF's) is less well understood (because the control elements also contribute to the problem), but relatively satisfactory methods of compensation for VCF's are available. Here we will take a more careful look at this problem, and will examine more critically some of the assumptions made in the past.

HISTORY OF THE STATE-VARIABLE VCF:

Our first state-variable VCF actually appeared way back in EN#15 [1], and was controlled with type 595 four-quadrant multipliers. I'm not sure exactly when the idea of using multipliers in this way came to me, but it was fairly obvious, and may well have been influenced by a copy of a publication sent in by an early reader [2]. In any event, the state-variable VCF actually became practical only when the OTA in the form of the CA3080 was employed. This application is straightforward, and was known to me from confidential disclosures from G. Wilcox, J. Scott, and M. Suchoff (among others) before we were fortunate enough to have the design from Terry Mikulic submitted and published in EN#33 [3]. At the same time, the voltage-controlled integrator based on the OTA was used in a VCO design by Sergio Franco [4] making the transfer of the ideas to the VCF all the more obvious. Various refinements to the basic OTA VCF were made, and a four-pole VCF was also developed [5]. The next important step was influenced by a 1976 paper in Electronic Design by Sergio Franco [6] which discussed the Q-enhancement problem and suggested a compensation method which we then adopted. This solution, the use of a compensating capacitor across one of the attenuating resistors of the OTA input, thus producing a phase lead, appeared in our ENS-76 series of modules [7], and is still used today. In general, the method is satisfactory, although experimentally determined.

A few more references can be mentioned. Ian Fritz called our attention to an earlier paper by Sparkes and Sedra [8], which is also referenced by Franco [6]. This earlier (1973) paper uses multiplier controlled VCF's and shows Q-enhancement and compensation results. This in turn leads to the still earlier paper by Thomas [9] where Q-enhancement in active state-variable filters is discussed. The tradition of compensating state-variable active filters, and other active filters for finite gain-bandwidth product is one which continues even today, but it is relatively well understood [10 - 13]. Other papers on VCF design using OTA's or similar include an earlier report by Franco [14] and a paper by D. Colin who was then at ARP [15]. After the ENS-76 series, a reader wrote to suggest that the compensating capacitors we used actually compensated the op-amps rather than the control elements as we had suggested [16], and referenced a paper by G. Volpe [17]. A quick check of the numbers suggested that they actually overcompensated the op-amps, compensating the OTA's as well. In fact, both points of view have their valid regions, and this we hope to demonstrate below. Finally, a recent paper by H. Malvar [18] should be mentioned as another example of OTA VCF compensation, as well as for a couple of other useful ideas.

REVIEW OF BASIC STRUCTURES COMPENSATION PRACTICES:

By way of reminder and review, we show in Fig. 1 the basic structure of a state-variable OTA-controlled VCF. It consists of two OTA-controlled integrators and a

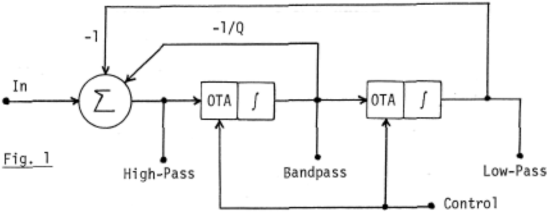


Fig. 1

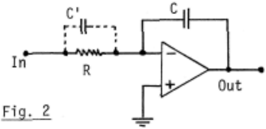


Fig. 2

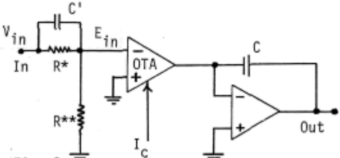


Fig. 3

summer in the feedback structure shown. Each integrator consists of an OTA such as the CA3080 and usually one op-amp. Fig. 2 shows an example of an active integrator. Such an integrator has a transfer function -1/sRC ideally. In the case of a real op-amp, which is usually modeled as single-pole compensated, the capacitor C' is added to provide the frequency compensation. The single-pole compensation model of the op-amp is $V_{out} = (V_+ - V_-)(G/s)$ where G is the gain-bandwidth product in rad./sec. In this case, the capacitor C' in parallel with R adds a zero to the response at -1/RC'. If C' is chosen so that this zero is at -G, then the overall integrator stage is properly frequency compensated. Techniques such as this example illustrates are well studied [10, 13].

Fig. 3 shows how the same technique can be applied (or at least, has been applied) to the OTA integrator. Here we show the resistor network composed of R* and R** which attenuates the input signal V_{in} to a level E_{in} needed at the actual input of the OTA chip (the signal must be small for linearity). The question we want to consider here is the effect of adding the shunt capacitor C' to the resistor R*. This we can conveniently do by examining the transfer function of the input attenuator, E_{in}/V_{in}.

$$T_a(s) = E_{in}/V_{in} = \frac{R^{**}}{R^{**} + \dfrac{R^*(1/sC')}{R^* + 1/sC'}} \approx \frac{R^{**}(1 + sC'R^*)}{R^*(1 + sC'R^{**})} \quad (R^* >> R^{**}) \qquad (1)$$

Thus we see that the input attenuator with C' added results in a zero at s = -1/R*C' and a pole at s = -1/R**C'. Looking at typical values, we might have R* = 10k, R** = 22Ω, and C' = 30 pfd. The wide spacing between R** and R* means that the pole will be very very far out on the negative real axis while the zero will just be far out. We can choose C' so that it cancels the pole of the op-amp that we want to get rid of without fearing that the extra pole we get (at -1/R**C') will be of any real importance.

Thus we set the zero to cancel the op-amp pole at -G, then we have C' = 1/R*G. At this point, we can look at some actual experimental values to see what is used in actual practice. Table 1 shows a listing of sources and compensation values. We should point out right here that these results are difficult to compare for a number of reasons, but we are just looking for a trend, so the necessary equivocations will be omitted.

TABLE 1

Source	Op-Amp Used	G of Op-Amp	Zero Placed	OTA Used
Franco [6]	MC1456	1 MHz	0.66 MHz	CA3096
ENS-76-1 [7]	CA3140	4.5 Mhz	1.6 MHz	CA3080
ENS-76-2 [7]	FET	---	0.53 MHz	CA3080
Sparkes & Sedra [8]	741	1 MHz	0.16 MHz	AD530 multiplier

From the table, we see that the general trend is that the zero is placed much closer in than would be necessary from the op-amp gain bandwidth product limitations. The implication would seem to be that the method corrects (or helps to correct) for some additional phase shift in the loop. Possible candidates for this additional phase shift are the OTA's and the input summing amplifier. That is, experimentally, the method that supposedly compensates for op-amp phase shift in the integrators also seems to compensate for additional phase shift, leveling off the Q. This is certainly useful information from a pragmatic point of view, but academically, we would also like to know exactly how this works, and if it is the best possible way to compensate the circuit.

AN EXAMPLE OF Q-ENHANCEMENT:

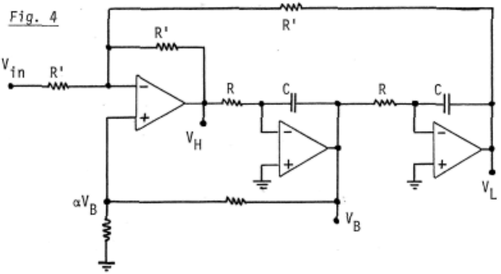
We need to take a fairly careful look at Q-enhancement in a simple active-filter state-variable filter to see what the problems are there, and how they manifest themselves. It will be convenient to look at the common state-variable structure of Fig. 4. The filter gives a transfer function, for ideal op-amps, of:

$$T_H(s) = \frac{-s^2}{s^2 + 3\alpha s/RC + 1/R^2C^2} \qquad (2)$$

where $T_H(s) = V_H/V_{in}$. Similar transfer functions are obtainable at the bandpass (V_B) *and low-pass (V_L) outputs. All of these have the denominator of equation (2), and* from the denominator, poles are found at:

$$s_{1,2} = (1/2RC)[\ -3\alpha \ \pm j\sqrt{4 - 9\alpha^2}\ ] \qquad (3)$$

Fig. 4



The Q of the filter is nominally:

$$Q = Q_o = 1/3\alpha \tag{4}$$

We can get a more realistic model, one which is useful for high-Q and high design frequency, by using the single-pole model of the op-amps:

$$V_{out} = (G/s)[V_+ - V_-] \tag{5}$$

We need to apply this to both the integrators and to the input summer. The application of equation (5) to the integrator has been worked out elsewhere [13] and is given as:

$$T_i(s) = \frac{-1}{sCR[1 + s/G + 1/RCG]} \tag{6}$$

This equation relates the output voltages V_H, V_B, and V_L as:

$$V_B = T_i(s)V_H \qquad V_L = T_i^2(s)V_H \tag{7a,7b}$$

In the ideal case, $T_i(s)$ is given by $-1/sCR$, but here, it is given by equation (6). As in the ideal case, we find the input voltages to the leftmost op-amp as:

$$V_- = \frac{V_H + V_L + V_{in}}{3} \qquad V_+ = \alpha V_B \tag{8a,8b}$$

In the ideal case, we would set $V_- = V_+$, and arrive at equation (2), but here we will relate V_- to V_+ using equation (5). The result is then:

$$T_H(s) = V_H/V_{in} = \frac{-1}{3s/G - 3\alpha T_i(s) + T_i^2(s) + 1} \tag{9}$$

Equation (9) gives the correct answer allowing for the finite gain-bandwidth product of the op-amp. For a final answer, we need to plug in for $T_i(s)$. This is a manner of considerable algebra. Using equation (6), and setting RC = 1, we get:

$$T_H(s) = \frac{-D_i^2(s)}{as^5 + bs^4 + cs^3 + ds^2 + es + f} \tag{10}$$

where $D_i(s)$ is the denominator of $T_i(s)$, and the constants a, b, c, d, e, and f are given by:

$$a = 3/G_n^2 \tag{11a}$$

$$b = 7/G_n + 6/G_n^2 \tag{11b}$$

$$c = 5 + 8/G_n + 3/G_n^2 \tag{11c}$$

$$d = 3\alpha + 2 + G_n + 1/G_n \tag{11d}$$

$$e = 3\alpha(G_n + 1) \tag{11e}$$

$$f = G_n \tag{11f}$$

with: $$G_n = G/(1/RC) = GRC \tag{12}$$

Here G_n is a normalized gain-bandwidth product, measured in units of 1/RC. Thus if G_n is large, we are designing well away from the gain-bandwidth product limitations, while if G_n is relatively small, we may need to be careful, with the distinction between large and small depending on the design Q.

Equation (10) tells us a good deal. First, note that the denominator is fifth-order, as we expect, two orders from the filter's capacitors, and three orders from the op-amp poles. Secondly, considering the coefficient given in equation (11), we see that higher powers of s are relatively weak, and the equation reverts to second order as G_n goes to infinity. Thirdly, we see that the zeros of $T_H(s)$ are given by the denominator of $T_i(s)$.

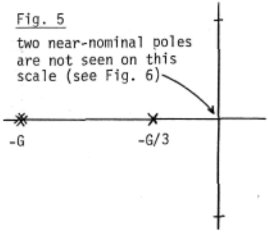


Fig. 5

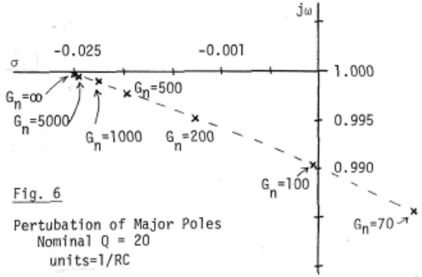


Fig. 6

Pertubation of Major Poles
Nominal Q = 20
units=1/RC

With $T_H(s)$ known, we can find poles and zeros and get a feel for any Q-enhancement and/or frequency shifts. In examining the poles and zeros, we look at two regimes. First, we want to look for any shifts in the nominal positions of the poles and zeros we expect in the ideal case. Secondly, we want to look at the positions of extra poles and zeros due to the op-amps. In the first regime, we are looking on a scale of frequency on the order of 1/RC. In the second regime, we are looking on a scale of frequency of order G, and in general, G will be somewhat or much greater than 1/RC. Our large scale look is given by Fig. 5, which shows the poles of $T_H(s)$. We will discuss the zeros in a moment. Fig. 6 shows a very restricted region in the vicinity of j/RC on the jω-axis. This shows the position of the poles that were originally present in second-order, and shows how they move as G_n decreases from infinity. From this we see that we get a rapid Q-enhancement and a small decrease in frequency (down to 99% the original) as G_n decreases to 100. This is the main manifestation of the effects of the op-amps on the circuit. Note that the Q is given as the reciprocal of twice the σ-axis magnitude to a good approximation here.

We note the position of an op-amp caused pole at -G/3 in Fig. 5. This can be understood in terms of the summer op-amp working at an effective gain of 3. It was previously shown [13] that this pole moves in from -G as the gain increases. The effective gain of 3 is due to the fact that there are two R' resistors acting as inputs to the summer on the (-) side, and these in parallel to ground give R'/2 giving a non-inverting gain of $1 + R'/(R'/2) = 3$. This is the same calculation that can lead to the value of Q for this type of circuit. The importance of this pole is that it is the closest of the undesired poles, and provides three times as much phase shift as a pole at -G. As we shall see, the cancellation of this pole is probably the most important compensation step. Further, this result seems to bare out claims [16, 17] that the phase lead networks are used to take care of phase shift associated with the input summer. Still another point relates to the need for care in designing input structures so that summers work at as small effective gains as possible. [More on this later.]

We have so far not discussed the zeros. Since the zeros of $T_H(s)$ are caused by $D_i^2(s)$, we see from equation (6) that we get two zeros at s=0, and two zeros at approximately -G. The two zeros at s = 0 are just the desired nominal zeros at s = 0 that are required for high-pass. The two at -G are fortuitous in that they cancel the two poles at -G shown in Fig. 5. Note that as we go from high-pass to bandpass (see equation 7a) that we loose two zeros, one of those at s = 0 (as we should) and one at -G. Likewise, going on to low-pass (see equation 7b) gets rid of another zero at s = 0 (as we should), and the remaining zero at s = -G. Thus pole cancellation at s = -G depends on the response function. The numerator will be different for high-pass, bandpass, and low-pass, not only in its power of s, but in that it will have different associated zeros (or lack of such zeros) in the vicinity of -G. Thus while the most significant poles move as in Fig. 6, we may expect some quite minor variations in response shape due to extra zeros way out.

MODELING OF THE OTA:

The ordinary op-amp is, for the most part, well modeled, and we can use the single pole roll-off model with good results (or often we can just consider the op-amp to be ideal). It would be useful if a similar model were available for the OTA. It appears that things are not so simple. For one thing, this is not found in the OTA literature. For another, it is clear that the OTA has no intentionally inserted compensation. [In fact, the OTA is uncompensated, and is often, if not ususally, used open loop.] Finally, it is very difficult to measure the frequency properties of the OTA with ordinary equipment due to the high frequencies needed, the small signal levels, stray capacitance, and a number of other frustrations which the reader can soon discover by trying to take open loop frequency and phase measurements. What is clear is that any roll-off poles are somewhat or well above 1 MHz.

For the most part, we will be using the limited amount of data that is available in the literature as a guide, and will hope to show that the OTA bandwidth is as stated through the success of Q-compensation based on these values. The usual literature on the CA3080 [19, 20] gives us no open-loop data except to state a 2 MHz bandwidth at a control current of 0.5 ma. Some curves are available for the CA3060 triple OTA device [21] and in the newer dual CA3280 [22] but the information is not all that clear. What does seem to be true is that the bandwidth does increase with the control current (as does the gain). The reason for this behavior is perhaps found in a tutorial paper on monolithic op-amps [23] which shows an example input that is quite similar to an OTA. Since no intentional frequency compensation is added to the OTA, it would seem reasonable that limitations are due to the input stage and possibly to some of the PNP structures. In the reference [23] it is stated that the poles of the input stage are a linear function of "tail current" which is equivalent to control current in the OTA. There are two important poles indicated there, the second pole being about three times further out than the more important one. Since we can't be sure of the exact OTA model, we will try to keep our calculations flexible.

ANALYSIS OF OTA INTEGRATOR STAGES:

Two popular forms of of the OTA integrator are seen in Fig. 7 and Fig. 8. Both integrators are non-inverting, Fig. 7 by virtue of two inverting structures in cascade. The non-inverting integrator is preferred because it simplifies the summing going back to the input, and a single inverting op-amp stage can be used there. The stages shown are standard except for the addition of the components C" and R' shown. R' is used in place of C' of Fig. 3. It is known that the addition of a series resistor R' with $R' = 1/GC$ gives the same compensating zero that was achieved with C' [10, 13]. We use this R' here for simplicity, and it is assumed that R' is chosen to exactly compensate the G of the op-amp used, and is not artificially made larger as C' is in practice (see earlier comments). Thus we isolate and remove the op-amp pole, so far assuming that C" is zero.

Now, concerning C", we have added this with the thought that it might be made to act much as C' does with the fixed resistor of Fig. 3, except here C" is across the OTA's equivalent resistor. Ultimately we will need to justify this through a more complete analysis and an experimental test, but some feeling for it may come as follows. First, assume an overall bandwidth parameter of the OTA op-amp system and call this G'(s). In analogy with Fig. 3, we would need to choose $C" = 1/RG'(s)$ and here $R = R_{eq} = 23.7/I_C$. Now, if we further assume that G'(s) has a linear multiplying factor proportional to I_C, then we can see that a constant value of C" is indicated. The limited amount of information on the OTA certainly indicates that the bandwidth is something like a linear function of I_C. While this argument is oversimplified, it may serve to aid intuition.

In the analysis of the OTA integrator of Fig. 7, we will model the op-amp according to the usual single-pole model:

$$V_{out} = (V_+ - V_-)\frac{G}{s} \tag{13}$$

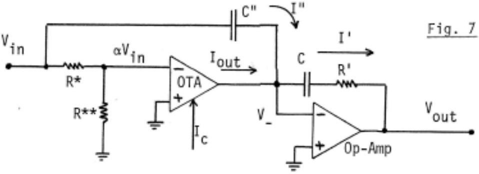


Fig. 7

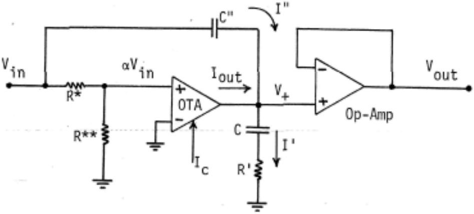


Fig. 8

where G is the gain-bandwidth product of the op-amp. We can model the OTA according to:

$$I_{out} = (V_+ - V_-)\, g_m(s) \tag{14}$$

where $g_m(s)$ is the transconductance as a function of frequency. For the moment, we will be carrying this as an unknown function. Later we will specify that it can be written as:

$$g_m(s) = \frac{g_o N(s)}{D(s)} \tag{15}$$

where N(s) and D(s) are the numerator and the denominator associated with the frequency transfer characteristics. For example, if we are going to eventually look at a single pole roll-off for $g_m(s)$, then $N(s)/D(s) = \omega_p/(s+\omega_p)$. Our interest in keeping it general at this point is to see how poles and zeros of $g_m(s)$ will be transferred to the OTA integrator. This is about all we need to know, except that we know that C" and R' are small. We cannot ignore them, but if we get a product term involving both, we will be able to throw that out.

The problem is completely set up by summing currents at the V_- node:

$$I_{out} + I'' = I' \tag{16}$$

Using equation (14) $I_{out} = -\alpha g_m(s) V_{in}$. We also have $I'' = (V_{in} - V_-)sC''$ and $I' = (V_- - V_{out})/(R' + 1/sC)$. Finally, from equation (13) for the op-amp we have $V_- = -\frac{s}{G}V_{out}$ so the rest is a matter of algebra. We get:

$$T_i(s) = \frac{G[\alpha g_m(s)(1+sCR') - sC'']}{s[s(C+C'') + CG]} \tag{17}$$

where we have thrown out two terms in R'C" along the way. The final form of $T_i(s)$ will depend on our choice of $g_m(s)$, but it is already clear that we have our desired integrator pole at s = 0, and the expected op-amp pole at approximately s = -G, since C" is small compared to C. It can also be seen that if equation (15) is substituted into equation (17), that the poles of $g_m(s)$, from D(s), will become poles of $T_i(s)$ Thus we will be looking to see how well the zeros generated by the numerator of $T_i(s)$ can be used to compensate for the poles of $T_i(s)$.

Analysis of Fig. 8 is quite similar to that of Fig. 7. We use equation (14) for the OTA, and sum currents $I'' + I_{out} = I'$, and solve for V_+. Then it is a simple matter to allow for the op-amp follower's response $V_{out}/V_+ = G/(s+G)$ [13] to arrive at $T_i(s)$ for Fig. 8:

$$T_i(s) = \frac{G[\alpha g_m(s)(1+sCR') + sC'']}{s(s+G)(C+C'')} \tag{18}$$

If we make C" much greater than C, then equations (17) and (18) are identical except for the sign in the numerator which is (-) for equation (17) and (+) for equation (18). If we represent this sign by a parameter "a", and use equation (15) for $g_m(s)$ then we can write a general form for the numerator of either case as:

$$n(s) = \alpha g_o N(s)(1 + sCR') + asC'' D(s) \tag{19}$$

Now we need to chose an actual model for $g_m(s) = g_o N(s)/D(s)$ and this we will take as the single pole, so:

$$N(s)/D(s) = \omega_p/(s + \omega_p) \tag{20}$$

This gives us a second-order numerator n(s) for the OTA integrators, and we can find the zeros by setting n(s) = 0 as:

$$n(s) = s^2 + s\omega_p\left[\frac{CR'}{aR_{eq}C''} + 1\right] + \frac{\omega_p}{aC''R_{eq}} = 0 \tag{21}$$

where $R_{eq} = 1/\alpha g_o$ is our usual ideal of an equivalent resistance for the OTA in this type of setup. Note how C" has combined with R_{eq} to form a time constant. We can plug in values into equation (21) and solve for the zeros. Here we will assume a control current to the OTA of 0.5ma, and assume that the 2 MHz bandwidth given on the OTA data sheet [20] refers to the single pole model, in which case $\omega_p = 2\pi \cdot 2 \times 10^6$ and $R_{eq} = 23.7/I_c = 47.4k$. We have already discussed that R' should be matched to the gain-bandwidth product of the op-amp used in the integrator, so if this op-amp is an LF351 with a gain-bandwidth product of 4.5 MHz, then R' = 1/GC. Typically C might be 330 pf, so R' would come out $1/(2\pi \cdot 4.5 \times 10^6)(330 \times 10^{-12}) = 107\Omega$. This leaves us with the choice of a = ±1 for the structure considered, and a value of C". We can then see how the zeros move as a function of C".

The results of solving equation (21) using the quadratic formula are plotted in Fig. 9 for the structure of Fig. 7 and in Fig. 10 for the structure of Fig. 8. It is evident that there is a significant difference between the two. We need to develop a terminology for these two non-inverting integrators, and it will be useful to refer to Fig. 7 as a "dual inverting integrator" while Fig. 8 will be called a "non-inverting integrator." Note that the dual inverting integrator has been used more recently in VCF designs [6,7], but the non-inverting integrator is also found [3,7]. A preference for the dual inverter has been suggested, based on the idea that it is better to drive a current into a constant (ground) potential rather than into a variable voltage (considering the OTA output), although no hard evicence for this preference is available.

From Fig. 9, the dual inverter, we see that the zeros that result from the use of C" are always real, although one is always in the right half-plane. On the other hand, when the non-inverting integrator is used, Fig. 10, we see that the zeros are complex and arc in as C" increases from 0. At about 5 pfd, they become real zeros. In either case (Fig. 9 or Fig. 10), the zeros arrive at s = 0 and at $s = -\omega_p$ as C" becomes very large. This can easily be understood in terms of C" effectively "shorting" out the OTA, in which case, the OTA pole at $-\omega_p$ should be gone, and also, since we now have a capacitor input leg (the large C"), the integrator is gone too (hence the loss of the pole at s = 0). The same conclusion results analytically by letting C" go to infinity in equation (21), as the reader can verify. Taking the other limit, as C" goes to zero in equation (21), we find a lone zero remaining, at s = -1/R'C. This is expected, as we have set 1/R'C = G, and it is clear that the removal of C" leaves us with our usual integrator compensation for the op-amp.

Fig. 9

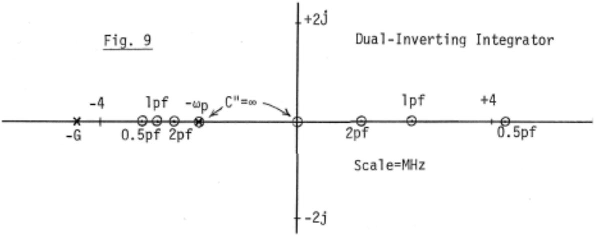
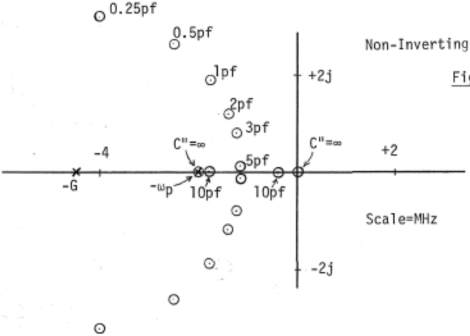
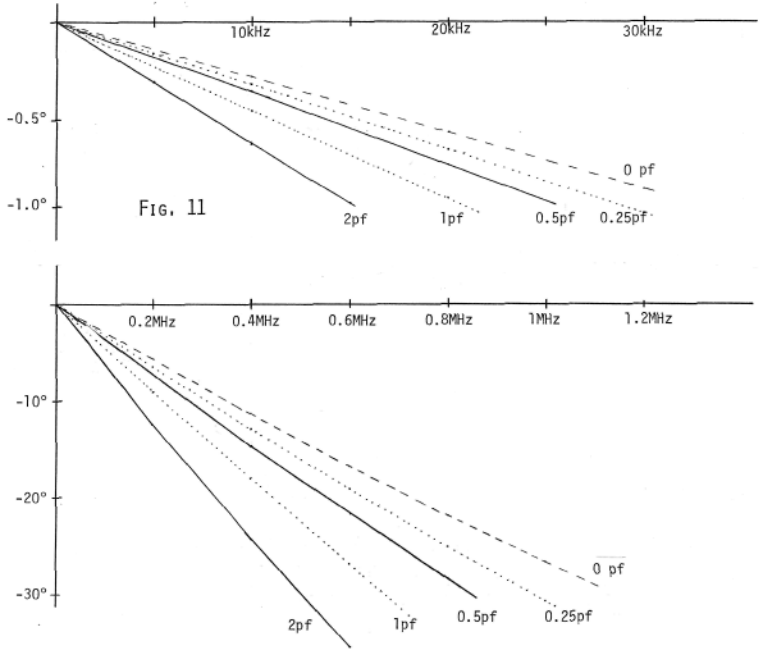


Fig. 10



It is relatively difficult to judge the effect of the zeros shown in Fig. 9 and in Fig. 10. In neither case do we see something nice like a cancellation of both the offending poles (at - G and at $-\omega_p$). In order to better examine the situation, we can look at the total phase shift due to the zeros produced by C" and the two offending poles. In this view, we are not looking for a zero to actually cancel a pole, but rather to just effectively reduce its phase shift. As a measure of the effectiveness of this process, we measure the phase angles, with respect to the positive real axis direction, of the zeros looking to the frequency of interest on the $j\omega$-axis, and then subtract the corresponding pole angles. Fig. 11 shows a set of phase response curves for the dual-inverting integrator, while Fig. 12 shows the corresponding set for the non-inverting integrator. Keep in mind that these phases are undesired additions to any desired phase from the integrator (i.e., 90° is desired). Each of Fig. 11 and Fig. 12 shows two regimes of frequency and phase, with the upper portion representing the extreme upper left portion of the bottom part, and corresponds to the region of audio interest. From the curves, it is clear that the addition of C" to the non-inverting integrator helps a lot, while the addition of C" to the dual-inverting integrator hurts the situation. In particular, a value of C" of about 1.7 pfd in the non-inverting case is an excellent compensation over the audio range (see upper portion of Fig. 12). The reason why C" helps in the non-inverter but not in the dual-inverter can be understood in terms of the input to the OTA used in each case. For the non-inverter, it is the (+) input to the OTA, and the OTA looks like a "non-inverting resistor." In the case of the dual-inverter, it is the (-) input that is used, and current through C" is fighting the normal corrective action.

FIG. 11



The placement of C" relative to the polarity of the input of the OTA being used is thus the central design principle in these and other OTA integrator structures.

It should be pointed out right here that a capacitor as small as 1.7 pfd is very difficult to work with. The reason is that some stray capacitances are this large or larger. Thus, for one thing, we may note that in a dual inverting structure, any stray capacitance, even in the absence of C", may make the Q-enhancement problem worse. On the other hand, any stray capacitance in the similar position in the non-inverter is already working to correct the problem. In addition, we can see the need for a clean and direct layout in the vicinity of the OTA output node.

It may seem strange to the reader that the more unusual of the zero patterns (Fig. 10) is in fact the more useful here. This is less of a surprise if we compare this with active compensation of integrators [10, 11, 24] which there uses complex poles to compensate for the phase of a real zero. In this case, an op-amp follower is placed in the feedback loop of the integrator. This places two poles at $-0.5G \pm \sqrt{3}jG/2$, and replaces the pole at -G with a zero there. The phase error is virtually reduced to zero for frequencies up to about 0.1G. Note that the zeros in Fig. 10 are hovering in somewhat the same region as the poles just discussed, although there are two poles to compensate here. We do understand the good results for about 1.7 pfd on this basis however.

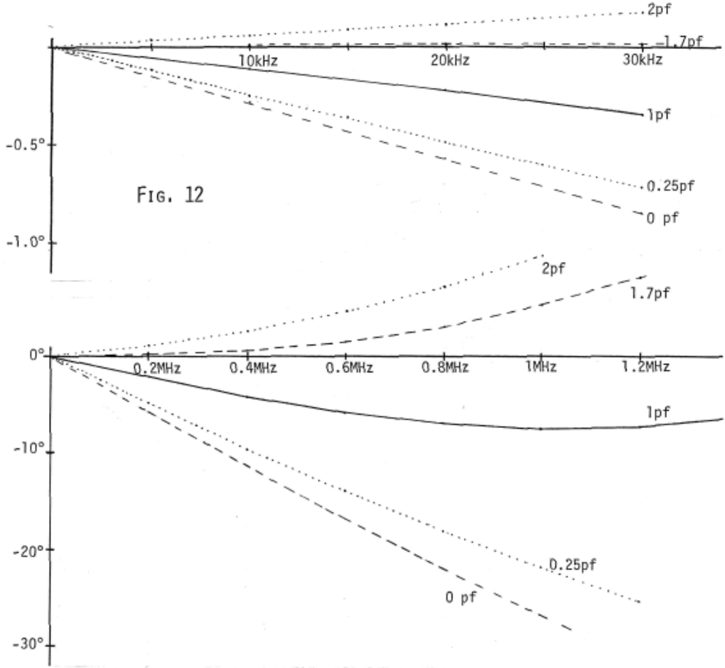


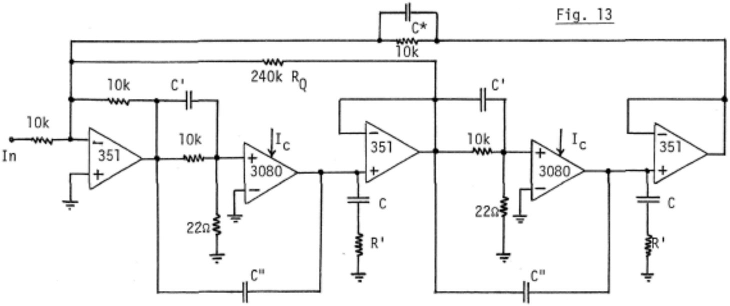
FIG. 12

EXPERIMENTAL TESTS:

At this point, it is useful to make a few tests of the ideas that we have looked at above. The experimental state-variable circuit is shown in Fig. 13, and can be seen to be built around two non-inverting integrators. There are a number of "extra" components, C', C", C*, and R' which are intended to serve to provide compensation for Q-enhancement. Not all these will actually be used in the circuit at any one time. The compensation with C' would be our "standard" compensation, the one we have used in the past. The compensation with R', as we have seen, works the same as C', but here we are assuming that it is exactly what is needed to compensate for the G of the LM351 to which it is associated. The capacitor C" is used to compensate for the finite $g_m(s)$ of the OTA in the manner discussed above.

This leaves us with the capacitor C* to be explained. As we found in the example of Q-enhancement with the active state-variable filter, the phase shift due to the input summer was very significant. Thus we can consider using a phase-lead network with the summing resistor, and this is what C* does. Note that C* in parallel with 10k should place a zero to cancel the op-amp (summer) pole. The op-amp, being an

Fig. 13



LM351, has a pole at -4.5 MHz, but here it is working at an effective gain of 3, so the pole is moved in by that factor, to -1.5 MHz. Thus $C^* = 1/2\pi(10k)(1.5MHz)$ is about 10 pfd. [Note that C', if set to cancel G of the integrator op-amps, would be only 3.5 pfd.]

Fig. 14 shows some experimental measurements of Q as a function of frequency for various compensation schemes. Each experimental point is an average of three separate Q measurements. The curves are labeled with the components that are actually used. If not listed, that component (C',C",C*, or R') was omitted for that trial. The first thing to note is that at the low frequency end, the curves all group around either Q = 22 or around Q = 17, the lower value of Q corresponding to cases where C" = 2pfd is used. We can understand the basic reason for this lowering of Q, even at low frequencies, by isolating the effect of a capacitor C' as in Fig. 2. Considering the op-amp ideal for the moment, the transfer function of Fig. 2 is:

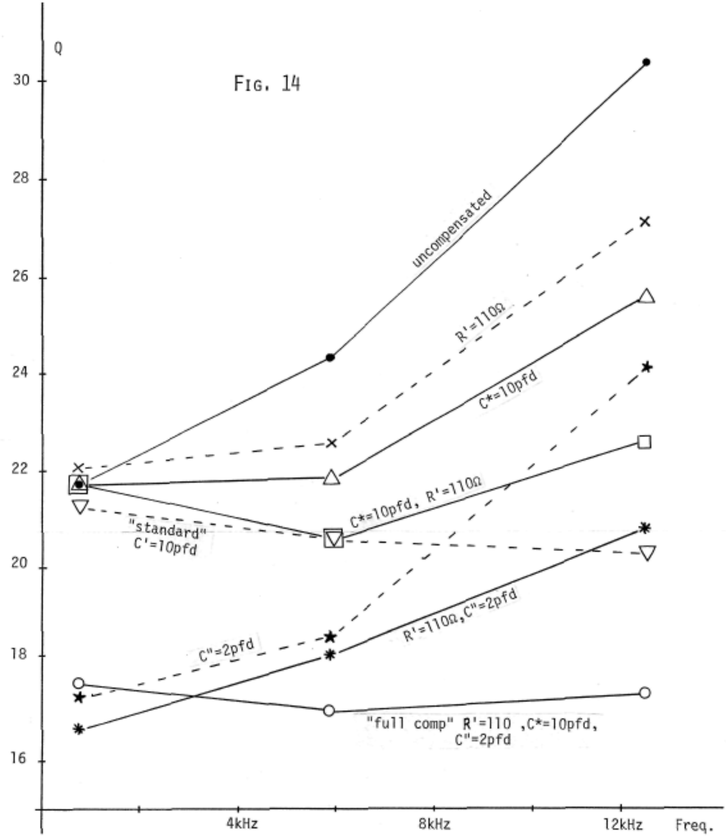
$$T_i(s) = -\frac{1 + sC'R}{sCR} \qquad (22)$$

which implies that we could find the effect of C' by substituting s/(1+sC'R) for s in a standard state-variable denominator:

$$D_{sv}(s) = s^2 + Ds/RC + 1/R^2C^2 \qquad (23)$$

Setting $C' = \beta C$, and making the substitution, we find a new damping term, $D + 2\beta$, substituting for D. If C = 330 pfd and C' = 2 pfd, with D = 1/Q = 1/24 = 0.0417, then $D + 2\beta$ corresponds to a Q of about 18.6, a lowering of about 5, which is what we saw (22 down to about 17). By the same token, the lowering from the nominal value of 24 down to 22 can perhaps be attributed to stray capacitance. The lowering of 5 is due to 2 pfd, so a lowering of 2 would be a capacitance of less than 1 pfd, which is certainly in the range of stray capacitances. [Note R' has cancelled op-amp pole here].

The uncompensated case clearly shows enhancement, reaching a Q of about 31 at the high-frequency end (●). Simply compensating the integrator op-amps (R'=110, shown by ×) helps, but does not do the whole job. We tried, but did not show in the figure, the case where R" = 330Ω, three times its value for a single op-amp, and this did work to reduce the enhancement. It was very similar to the standard enhancement (▽) which uses 10 pfd for C'. [Note that C'=10 pfd is about three times the value needed for one op-amp.] It can also be seen that the use of compensating capacitor C*=10pfd for the input summer is a help, but still leaves some enhancement (△). [We dont really need a similar capacitor across the 240k resistor, as this would we very small (0.4pfd). Finally, note that a combination of C* and R' (◻) does work fairly well. This takes care of everything except the OTA.



The curves that hover around Q = 17 at the low frequency end all have the OTA compensated by C" = 2pfd. Note that the use of C"=2pfd only (⋆) does not help much, with C" and R' being better (✱), and the full compensation case (R', C*, and C", shown by the open circle) is quite good.

The successful cases to compare are the standard compensation (C' = 10 pfd), the full op-amp compensation (R' and C*), and the full compensation (R', C*, and C"). It might be difficult to choose between these three cases from a performance basis, although the full compensation does come out the best here. Also, the standard compensation does have some droop, and we would probably want to go back and change

C' to 9 pfd or perhaps 8 pfd experimentally. Thus it may be the case that by distributing the compensation among the elements, treating each one individually and nominally, we may get a leveler Q curve with less trimming. However, this would need to be verified with more examples. In general, we would be quite satisfied with the full compensation results.

The state-variable formed from the dual-inverter integrators was studies very briefly. It was found that the addition of C" (to Fig. 7) did in fact make the Q-enhancement problem worse instead of better. A value of C" of 5 pfd gave a Q of 183 (nominally 24) at 12 kHz. A value of C" of 10 pfd was unstable for all frequencies.

CONCLUSIONS AND RELAVANCE TO VCF DESIGN OPTIONS:

Much of what has been done above has led to a better understanding of the Q-enhancement problem, its cause and cures. While a full model of the OTA based filter has not been worked out, a good deal of the parts have yielded to analysis. We can now arrive at some conclusions, and from these, makes some suggestions about design choices.

1. The standard method of Q-compensation does work in practice, but it does seem to attack the problem in the wrong place (integrators instead of summer).

2. The phase shift across the OTA's may be a relatively minor complication compared to the input summer.

3. The input summer suffers not just as an op-amp of unity gain, but as an op-amp of a somewhat higher effective gain, moving the op-amp pole in by a factor of the effective gain [13]. This is why the input summer is a major phase shift problem with regard to Q-enhancement.

DESIGN SUGGESTION 1: Try to keep the effective gain of the input summer as low as possible. Avoid multiple inputs to the summing node. If necessary, use a separate summing stage (Fig. 15a). Also, it is probably better to use a Q control pot of large resistance in the manner of Fig. 15b rather than the conventional Fig. 15c. This will help keep the effective gain down.

DESIGN SUGGESTION 2: A compensating capacitor, as C* in Fig. 13, may be preferable to the usual (C') integrator lead networks.

Fig. 15a

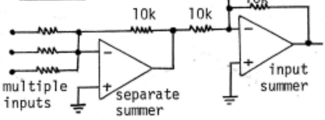


Fig. 15b

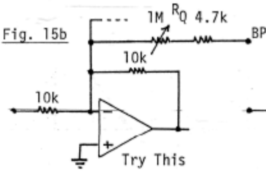
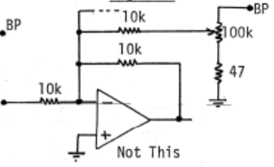


Fig. 15c



4. It is very difficult to measure the high-frequency amplitude and phase properties of the OTA. This may indicate that the OTA is not a major problem.

5. The OTA integrator can be modeled if we do not need to specify exactly the form of $g_m(s)$. Any poles of $g_m(s)$ will be transferred to the integrator's response.

6. Addition of a compensating capacitor across the OTA may be useful. This would be true where the input signal is applied to a (+) input (Fig. 8). This can produce complex zeros (Fig. 10) that reduce the effect of the phase shift of poles already present.

7. The value of C" is quite small, on the order of a few pfd. Consequently we are also interested in C" as it may exist as stray capacitance, independent of any intentional inclusion. For example, C" lowers the Q from a nominal value of 24 to about 17 (see Fig. 14). It may be that the stray C" is the value of the lowering from 24 to about 22 for the curves where no C" is intentionally included.

8. The inclusion of a capacitor C" of about 2 pfd can be useful for reducing the Q-enhancement when used in conjunction with other methods. The use of distributed compensation (individual nominal compensation at the offending elements themselves) may be better than the standard compensation which is basically an "overkill."

   DESIGN SUGGESTION 3: Consider using the non-inverting integrator configuration (Fig. 8) instead of Fig. 7. Consider using partial compensation in various places (C*, R', and C"). Try for a clean layout in the vicinity of the OTA output node. Keep in mind that any stray C" will help reduce Q-enhancement in the case of this non-inverter design.

9. Stray capacitance across resistors in integrators can cause a drop in Q even at very low-frequencies.

   DESIGN SUGGESTION 4: Be prepared for some relatively minor but still measurable variations in actual Q and Q as determined by the feedback factor from bandpass to the input summer. The resistance ratio itself will not assure the correct Q, even at very low frequencies.

FINAL REMARKS:

What we have done here is to make some steps in the direction of a better understanding of the state-variable VCF. If it is a measure of a good research effort that it asks more questions than it answers, then this is a success. While a good number of interesting results have been obtained, it still seems that there are a few missing pieces, and then a whole (rather complicated) picture to be developed. This however, must be left for another time, or perhaps for an interested reader to fill in.

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