Digitally-Controllable Audio Filters and Equalizers

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Filters are commonplace elements in audio signal processing. Most well-known filter topologies generate fixed responses with fixed element values, or use variable resistors to alter their responses under user control. This paper explores solutions to varying the response of filters and equalizers using digital control techniques suitable for computer-controlled systems. High-pass, low-pass, band-pass, and parametric topologies are covered using VCAs, resistor ladders, and multiplying DACs as control elements.

1. INTRODUCTION

While Digital-Signal-Processing (DSP) -based audio systems offer the promise of infinite flexibility with regard to computer-controlled filtering, the cost of professional-quality analog-to-digital (A/D) and digital-to-analog (D/A) converters and digital signal processors is currently prohibitive in many applications. There are also many applications where computer control over a well defined set of filtering and/or equalization functions is desirable. For these applications, digitally-controlled *analog* filters can be a high-performance, cost-effective alternative.

This paper describes several approaches for implementing familiar audio filtering and equalization functions under digital control. It is hoped that these examples will serve to illustrate the important issues and offer a starting point for the engineer seeking to design computer-controlled audio filters for a wide range of specific applications.

2. DIGITAL CONTROL OF ANALOG FILTERS

2.1 Resistor-Switch Arrays

Historically, the characteristics of analog audio filters have been varied using potentiometers, either as variable resistances or variable voltage dividers. The most obvious approach to digital control of these components is to provide a string of resistors and electronic switches that can be selected via a digital code (Figure 1). Integrated devices such as this exist [1], and, when appropriate, they can be substituted directly for potentiometers in analog circuits. However, currently available devices are somewhat limited in their application to professional audio

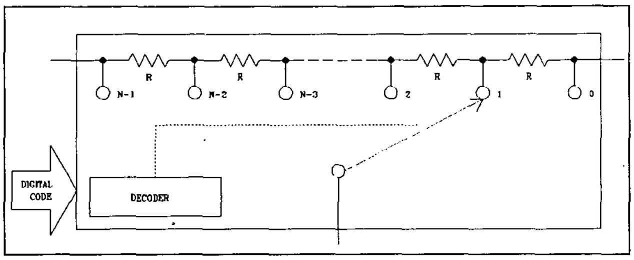

Figure 1 - Digitally Controlled "Pot"

products in that they are restricted to operation on maximum power supply voltages of +/-5 V. Limiting signal levels to within these extremes entails sacrificing approximately 10 dB of dynamic range with respect to systems using traditional +/-15 V (or greater) power supply voltages. Available "tapers" are either linear (equal resistance steps) or two-slope piecewise-linear approximations to a logarithmic taper. An additional limitation of this approach shows up when a system requires filter parameters to be varied while the circuit is processing program material. (Sometimes the circuit is varied *in response to* the program material.) In such cases, the discrete nature of the changes in position along the resistor string of the "wiper" can lead to zipper noise.

Other commercially available resistor-switch array integrated circuits (ICs) are intended specifically for implementing graphic equalizer functions [2, 3]. These devices are aimed at replacing the potentiometers that control boost and cut in a traditional analog graphic equalizer. The resistor segments are chosen to implement exponential steps in boost or cut at the filter center frequency, typically one or two dB per step over a +/-12 dB range. These devices have seen application in professional audio equalizers as well as in the consumer market where they were originally targeted. Zipper noise is less a concern with graphic equalizers since they are usually set once and not varied (or at least not varied much) while program material is present. The best of these graphic equalizer ICs will accept +/-7.5 V power supply voltages, which gains about 3 dB in dynamic range over +/5 V devices. The main objection to their wider use in professional audio graphic equalizers is that 1 dB resolution in boost and cut is considered inadequate for the most demanding applications.

2.2 Multiplying D/A Converters

Another approach to digital control of analog filters involves the use of multiplying D/A converters (MDACs). An MDAC is a specific type of D/A converter (DAC) with an externally accessible reference input that will accept bipolar signals (Figure 2). Its output is the product of the reference input and the digital input code according to the equation:

$$V_{out} = -V_{ref}\left(\frac{CODE}{2^N - 1}\right) \quad (1)$$

where V_{ref} is the reference input voltage, V_{out} is the DAC output voltage, CODE is the decimal value of the digital input, and N is the number of input bits.

An MDAC can thus implement two-quadrant multiplication in discrete steps. As shown in Figure 2, most MDACs are intended to be used as current output devices fed to the virtual ground of an opamp current-to-voltage converter. Often a feedback resistor which will track with the resistors internal to the converter is integrated into the device. In other devices, the entire current-to-voltage converter is included on the same substrate. Note that the current output is non-inverting with respect to the reference-voltage input, but that the voltage output is inverted.

A wide range of integrated MDACs is currently available in resolutions from 8 to 16 bits, with devices up to 12 bits available in multiple device-per-package configurations. The number of bits required for any application is a function of the range and resolution of parameter variation desired. The steps are typically linear, yet many filter parameters that must be adjusted in audio applications are more suitably adjusted in exponential fashion. For example, frequency is most naturally varied in octaves, or fractions thereof, and amplitude is most often adjusted in dB. Implementing exponential control with a linear DAC inevitably wastes bits at one end of the scale, thus requiring a higher resolution DAC than the desired number of control steps would indicate.

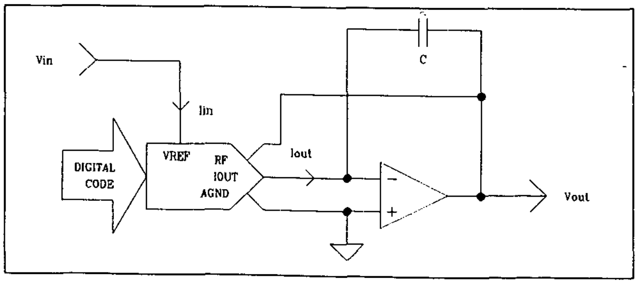

Figure 2 - Typical MDAC

There are MDACs with logarithmically spaced steps intended for use as audio attenuators. While many of these have non-uniform dB-step-size, at least one implements 3/8 dB steps over an 88.5 dB range [4]. As an illustration, this device provides 8 bits of exponential control resolution but requires an internal 17-bit MDAC to implement it. Additionally, because the control steps are discrete, the possibility of zipper noise exists in dynamic control applications where parameters are varied during the processing of program material.

2.3 Voltage-Controlled Amplifiers with DACs

Another form of two quadrant multiplier, the voltage-controlled amplifier (VCA), has been used to provide electronic control over analog circuits for many years now. As implied by the name, the gain or attenuation of this device is a function of an externally applied control voltage. The VCAs used most often in audio applications have control port characteristics of the form:

$$G = e^{\frac{V_{control}}{V_{scale}}} \quad (2)$$

where G is the VCA current gain, $V_{control}$ is the gain control voltage and V_{scale} is a constant that depends on the device type.

Equation (2) may be rearranged as follows:

$$V_{control} = V_{scale}\ln(G). \quad (3)$$

The control voltage, $V_{control}$, exponentially controls the VCA gain. This produces the desirable result that the control voltage linearly affects gain *in decibels*. Using a conventional linear DAC to generate this control voltage (Figure 3) provides digital control of gain or attenuation in equal, exponentially-spaced steps. This produces a constant step size in decibels per step, or dB/bit. As shown in Figure 3, a typical VCA has a virtual-ground current input and a current output intended for connection to the virtual ground of an opamp current-to-voltage converter. In contrast to the MDAC, this VCA inverts the current output with respect to the input, but the eventual voltage output is in phase with respect to the input.

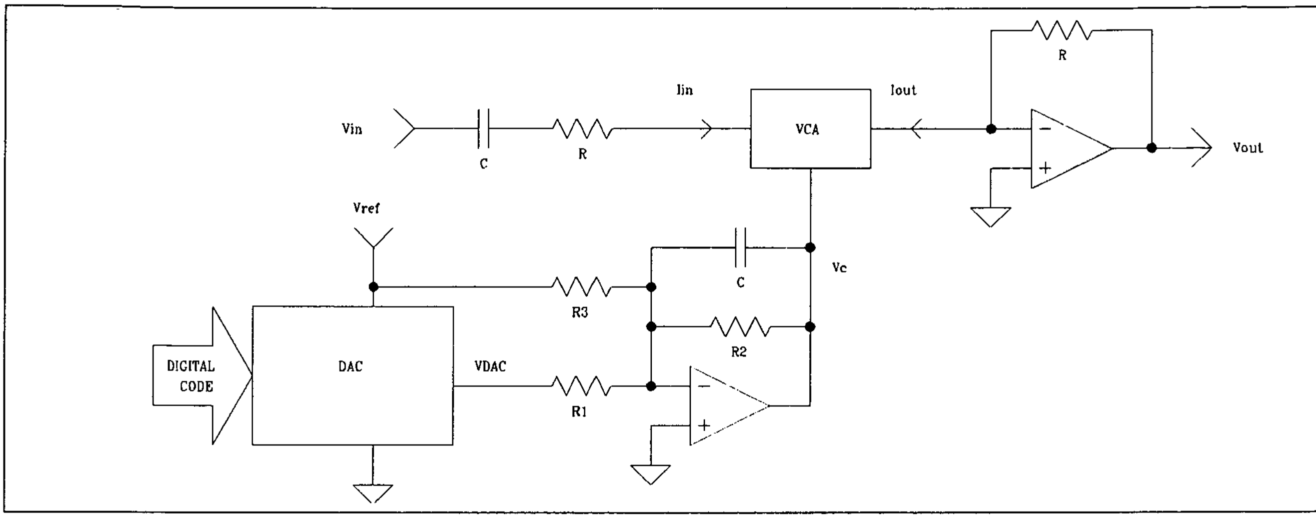

Figure 3 - VCA with Linear Control DAC

The combination of a VCA and DAC offers some unique advantages to the audio designer in addition to the exponential control mentioned above. As shown in Figure 3, the output of the controlling DAC may be lowpass filtered (R_2 and C) before application to the VCA control port. This serves to attenuate any glitch energy that the DAC may generate during transitions, and also serves to make the transitions between discrete gain steps continuous, thus eliminating the source of zipper noise. A further practical advantage is that the DAC may be physically distant from the VCA (and close to the controlling processor), easing the task of keeping digitally generated noise out of the audio signal path.

Figure 3 also illustrates a typical method of scaling and offsetting the DAC output to suit the desired range and resolution of VCA gain. The control voltage V_c may be expressed as:

$$V_c = -V_{ref}\left[\left(\frac{CODE}{2^N - 1}\right)\frac{R_2}{R_1} - \frac{R_2}{R_3}\right], \qquad (4)$$

where CODE is the decimal value of the digital input code and N is the number DAC input bits.

The voltage V_{scale} referred to in equations (2) and (3) is usually temperature dependent, most commonly proportional to absolute temperature (PTAT). This causes the gain of the VCA to vary with temperature at gains other than unity. While this temperature dependence is relatively mild and may be ignored in many applications, for the most demanding designs a convenient method of compensating is to make the DAC reference voltage V_{ref} vary in a similar manner. (For a PTAT control port characteristic, V_{ref} must vary .33%/degree Celsius around 27 degrees Celsius.)

2.4 Tradeoffs

Applying the resistor-switch array devices in audio circuits typically involves simply replacing potentiometers with their digitally-controlled equivalents in existing circuits. If their performance in the areas of dynamic range, resolution, zipper noise, and distortion are adequate for the application, their use is a relatively straightforward exercise.

The following sections concentrate on the use of two-quadrant multipliers (MDACs or VCA-DAC combinations) as building blocks to implement digitally controlled filter and equalization functions. As will be shown, using these building blocks requires some revision of familiar circuits. It is hoped that these examples will serve to illustrate the general approach to adapting existing analog circuits to digital control.

The choice of which of these two building blocks to use is very application dependent. The MDAC will typically (though not in all cases) offer very low distortion and noise. As mentioned above, the discrete steps can lead to audible zipper noise if parameters are adjusted while program material is being processed. If exponential control is desired, extra cost will be incurred both in the requirement for a DAC with higher resolution than the number of control steps would imply, as well as in the means to calculate the correct codes to implement exponentially-spaced steps. This typically takes the form of a ROM-based lookup table used by the controlling processor.

The use of a VCA-DAC combination will typically result in slightly higher noise and distortion levels, though the best currently available devices offer distortion levels below .03%, and dynamic range of over 115 dB [5]. As mentioned above, real-time variation of filter parameters while processing program material can be implemented without zipper noise problems. Controlling software (and

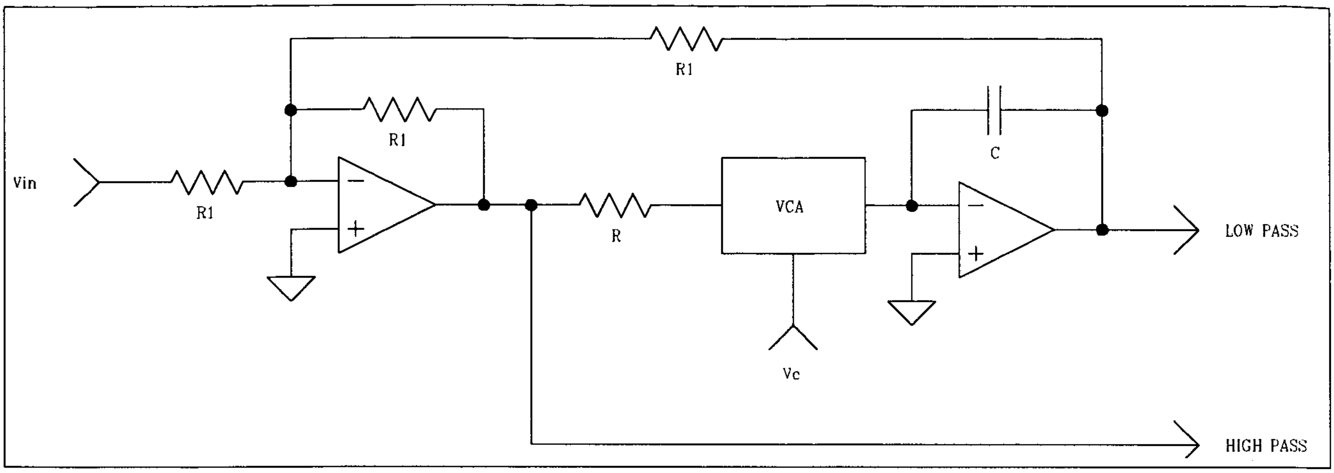

Figure 4 - VCA-Controlled First-Order State-Variable Filter

hardware) is often simplified by the exponential control characteristic of the VCA.

Though the VCA-DAC combination is used in most of the examples below, it is generally possible to substitute an MDAC for each instance of a VCA and control DAC. The designer will need to adjust the circuit for the inverted input-to-output transfer of the MDAC relative to that of the VCA, and to add additional gain external to the MDAC in instances where the VCA provides gain as well as attenuation.

3. FIRST ORDER FILTERS AND EQUALIZERS

3.1 Highpass and Lowpass Filters

The circuit shown in Figure 4 is a first-order state variable filter offering both highpass and lowpass outputs [6]. The transfer functions at the two outputs are:

$$\frac{V_{lp}}{V_{in}} = -\frac{\frac{G}{RC}}{s + \frac{G}{RC}} \qquad (5)$$

and

$$\frac{V_{hp}}{V_{in}} = -\frac{s}{s + \frac{G}{RC}} \qquad (6)$$

where G is the VCA current gain, V_{in} is the input voltage, V_{lp} is the lowpass output voltage, and V_{hp} is the highpass output voltage. With the resistor ratios shown, the passband gain at both outputs is unity, and the cutoff frequency of each filter is controlled by the VCA gain. Using a VCA with a control characteristic as described by equations (2) and (3) along with a linear control DAC causes each bit of the DAC input word to change the cutoff frequency by a constant fraction of an octave.

3.2 First Order Shelving Filters

The circuit shown in Figures 5 is a first-order high-frequency equalizer. This circuit operates somewhat differently from its more familiar potentiometer-based counterpart. The transfer function of the circuit of Figure 5 is:

$$\frac{V_{out}}{V_{in}} = -\frac{\left(s + \frac{G+1}{[G(R_1+R_2)+R_2]C}\right)[G(R_1+R_2)+R_2]}{\left(s + \frac{G+1}{[R_1+(G+1)R_2]C}\right)[R_1+(G+1)R_2]} \qquad (7)$$

where G is, again, the VCA current gain.

While this at first looks uninsightful, inspection reveals that, at low frequencies (as s approaches 0), the transfer function approaches -1. At high frequencies (as s approaches infinity), the transfer function approaches:

$$\left.\frac{V_{out}}{V_{in}}\right|_{s\to\infty} = -\left[\frac{G\left(\frac{R_1+R_2}{R_2}\right)+1}{G+\frac{R_1+R_2}{R_2}}\right]. \qquad (8)$$

To gain further insight, examine a few cases of values of G. When G=1, the transfer function reduces to -1. Thus, when the VCA gain is set to unity, the circuit's response is flat. For the case where:

$$G >> \frac{R_1+R_2}{R_2},$$

equation (8), the expression for high-frequency gain approaches:

$$\left.\frac{V_{out}}{V_{in}}\right|_{s\to\infty} = -\frac{R_1+R_2}{R_2} \text{ for } G >> \frac{R_1+R_2}{R_2}. \qquad (9)$$

Similarly, for the case where:

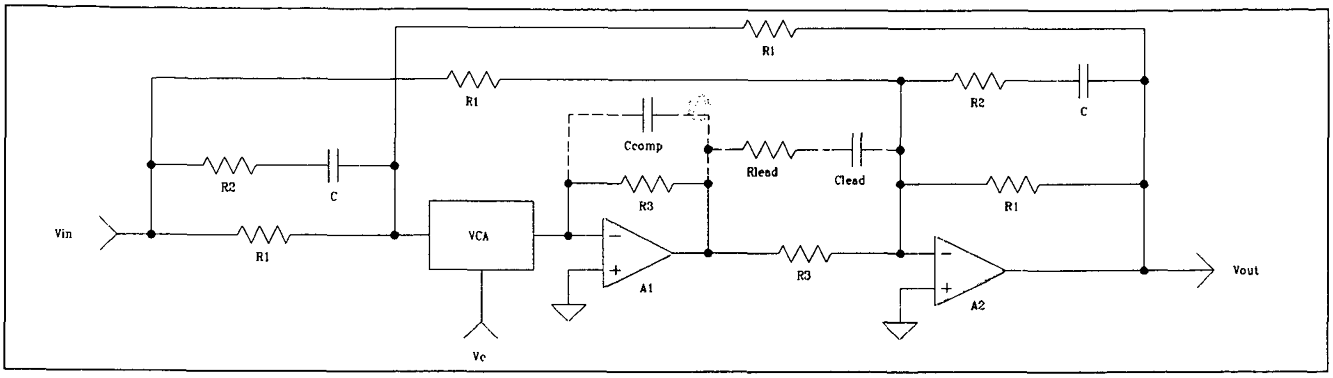

Figure 5 - VCA-Controlled High Frequency Shelving Filter

$$G << \frac{R_2}{R_1 + R_2},$$

equation (8) approaches:

$$\left.\frac{V_{out}}{V_{in}}\right|_{s \to \infty} = -\frac{R_2}{R_2 + R_1} \text{ for } G << \frac{R_2}{R_2 + R_1}. \qquad (10)$$

An intuitive understanding of the circuit can be gained by noting that when the VCA gain is much less than 1, the VCA and A_1 are effectively out of the circuit and the circuit performance is governed entirely by the resistor from the input to the summing junction of A_2 and the feedback components around A_2. These components make up a familiar high-frequency deemphasis network. Conversely, when the VCA gain is very large, the VCA, along with A_1 and A_2 form a block of high open loop gain. Under this condition, the circuit behavior is governed by the components between the input and the summing junction of the VCA (a high-frequency pre-emphasis network) and the feedback resistor from the circuit output to the VCA summing junction.

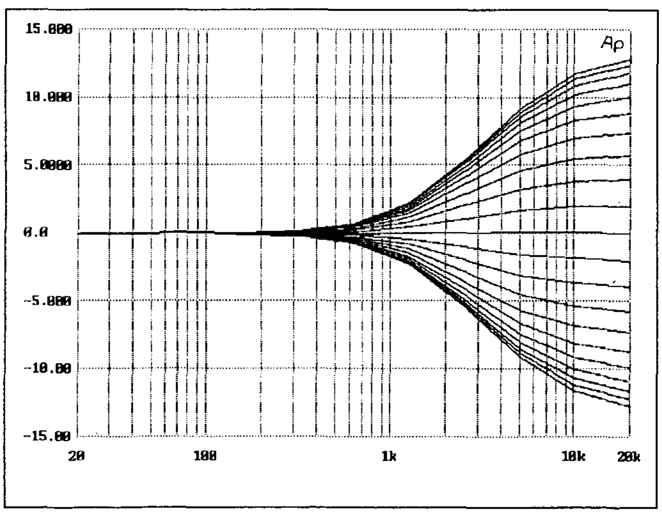

Figure 6 - High-Frequency Shelving Responses (VCA Gain Varied 3 dB/Step)

The use of the VCA as a variable open-loop gain block results in a control characteristic that is not directly proportional to VCA gain, and thus this circuit does not offer direct exponential (dB) control over boost and cut. The control characteristic is nevertheless quite useful, as the example in Figure 6 illustrates. The individual frequency response curves are for VCA gain stepped from -30 dB to +30 dB in 3 dB steps. The component values (referring to Figure 5) are R_1=45.3 kΩ, R_2=10 kΩ, R_3= 45.3 kΩ, and C=2.2 nF. The curves asymptotically approach limits set by the passive preemphasis and deemphasis networks. However, for boost and cut commands substantially less than these asymptotic limits (to about +/-10 dB in Figure 6), the boost and cut are nearly proportional to the VCA gain. This suggests that the designer should choose preemphasis and deemphasis networks for more ultimate boost and cut than is desired for the application to yield the most useful control characteristic.

The low-frequency shelving circuit shown in Figure 7 has a transfer function:

$$\frac{V_{out}}{V_{in}} = -\frac{s + \left[\dfrac{G + \left(\frac{R_1 + R_2}{R_1}\right)}{(G+1)R_2 C}\right]}{s + \left[\dfrac{G\left(\frac{R_1 + R_2}{R_1}\right) + 1}{(G+1)R_2 C}\right]} \qquad (11)$$

It may be analyzed in a similar fashion. Figure 8 shows a family of curves taken from a circuit with the following component values (referring to Figure 7): R_1 = 19.6 kΩ, R_2 = 84.5 kΩ, R_3 =40.2 kΩ, C=22 nF. VCA gain is stepped from -30 dB to +30 dB in 3 dB steps.

Both of these shelving circuits exhibit THD less than .01% throughout the audio band and 20kHz-bandwidth noise floors below -95 dBV.

A few practical notes are in order. The capacitor C_{comp} (typically 10 pF to 47pF) shown with dotted connections around A_1 in both Figure 5 and Figure 7 is typically required to mitigate phase shift caused by the combination

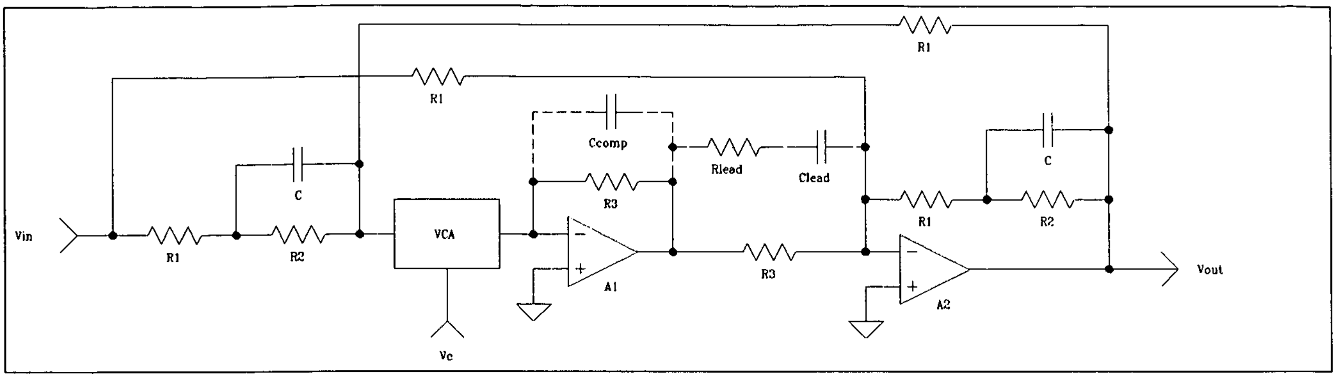

Figure 7 - VCA-Controlled Low-Frequency Shelving Filter

of A_1's feedback resistor and the VCA's (or MDAC's) output capacitance. In the case of high VCA gain the R_3/C_{comp} pole becomes the dominant pole in the loop. Since the VCA has a finite gain-bandwidth product, it begins to contribute excess phase shift at high frequencies as gain is increased. Depending on the device used and the maximum boost required, the lead network C_{lead}/R_{lead} may be required to maintain stability.

4. HIGHER ORDER FILTERS

4.1 State-Variable Filters

One of the most versatile filter building blocks is the familiar state-variable filter. Figure 9 shows this circuit adapted for VCA control [6]. The transfer functions for each of the three outputs (highpass, bandpass, and lowpass) are:

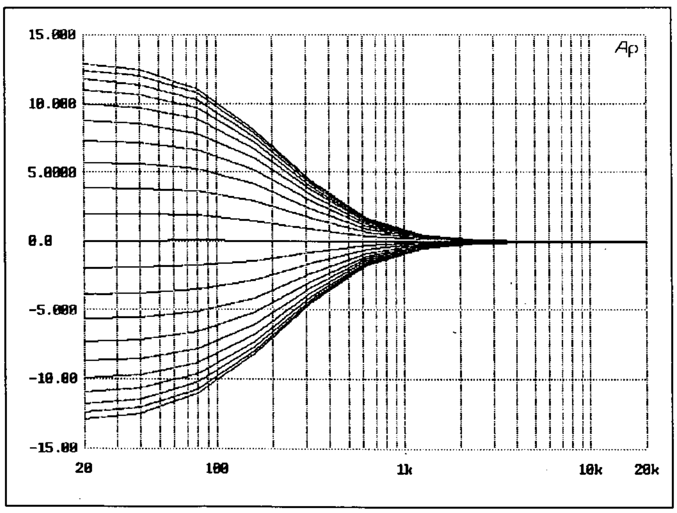

Figure 8 - Low-Frequency Shelving Responses (VCA Gain Varied 3 dB/Step)

$$\frac{V_{hp}}{V_{in}} = \frac{s^2}{s^2 + \frac{G_1}{R_2C}s + \left(\frac{\sqrt{G_1G_2}}{R_2C}\right)^2}, \tag{12}$$

$$\frac{V_{bp}}{V_{in}} = \frac{\frac{G_1}{R_2C}s}{s^2 + \frac{G_1}{R_2C}s + \left(\frac{\sqrt{G_1G_2}}{R_2C}\right)^2}, \tag{13}$$

and

$$\frac{V_{lp}}{V_{in}} = \frac{\frac{G_1G_2}{(R_2C)^2}}{s^2 + \frac{G_1}{R_2C}s + \left(\frac{\sqrt{G_1G_2}}{R_2C}\right)^2}. \tag{14}$$

where:

$G_1 = e^{\frac{V_{c1}}{V_{scale}}}$ = the current gain of VCA_1 and

$G_2 = e^{\frac{V_{c2}}{V_{scale}}}$ = the current gain of VCA_2.

The resistor ratios around the circuit were chosen for unity passband gain at all three outputs to simplify the equations, but this is not required. It should also be noted that a bandreject output can be created by summing the lowpass and highpass outputs.

By inspection, the natural frequency and Q for all three outputs are:

$$\omega_n = \frac{\sqrt{G_1G_2}}{R_2C} = \frac{e^{\frac{V_{c1}+V_{c2}}{2V_{scale}}}}{R_2C} \tag{15}$$

and

$$Q = \sqrt{\frac{G_2}{G_1}} = e^{\frac{V_{c2}-V_{c1}}{2V_{scale}}}. \tag{16}$$

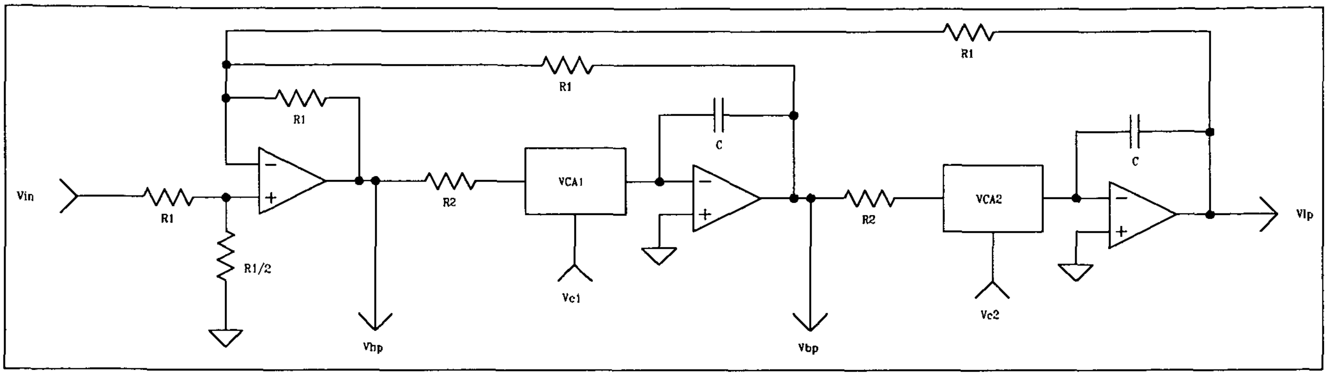

Figure 9 - VCA-Controlled State Variable Filter

At first glance it appears that independent control over natural frequency and Q (one of the main attractions of the state-variable topology) will be rather awkward with this circuit. However, this is another instance where the exponential control characteristic of the VCA is very handy. Adding sum and difference amplifiers to process the control voltages to the VCAs as shown in Figure 10 yields the following:

$$V_{c1} = -\left(\frac{R_4}{R_3}\right)V_{freq} + \left(\frac{R_6}{R_5+R_6}\right)\left(\frac{R_3+R_4}{R_3}\right)V_Q \quad (17)$$

and

$$V_{c2} = -\left(\frac{R_9}{R_8}\right)V_{freq} - \left(\frac{R_9}{R_7}\right)V_Q \quad (18)$$

Choosing resistor values such that:

$$\frac{R_4}{R_3} = \frac{R_9}{R_8} = A \text{ and } \frac{R_9}{R_7} = \left(\frac{R_6}{R_5+R_6}\right)\left(\frac{R_3+R_4}{R_3}\right) = B$$

yields:

$$V_{c1} = -AV_{freq} + BV_Q \quad (19)$$

$$V_{c2} = -AV_{freq} - BV_Q. \quad (20)$$

Substituting these expressions into equations (15) and (16) yields:

$$\omega_n = \frac{\sqrt{G_1G_2}}{R_2C} = \frac{e^{\frac{-AV_{freq}}{V_{scale}}}}{R_2C} \quad (21)$$

and

$$Q = \sqrt{\frac{G_2}{G_1}} = e^{\frac{-BV_Q}{V_{scale}}}. \quad (22)$$

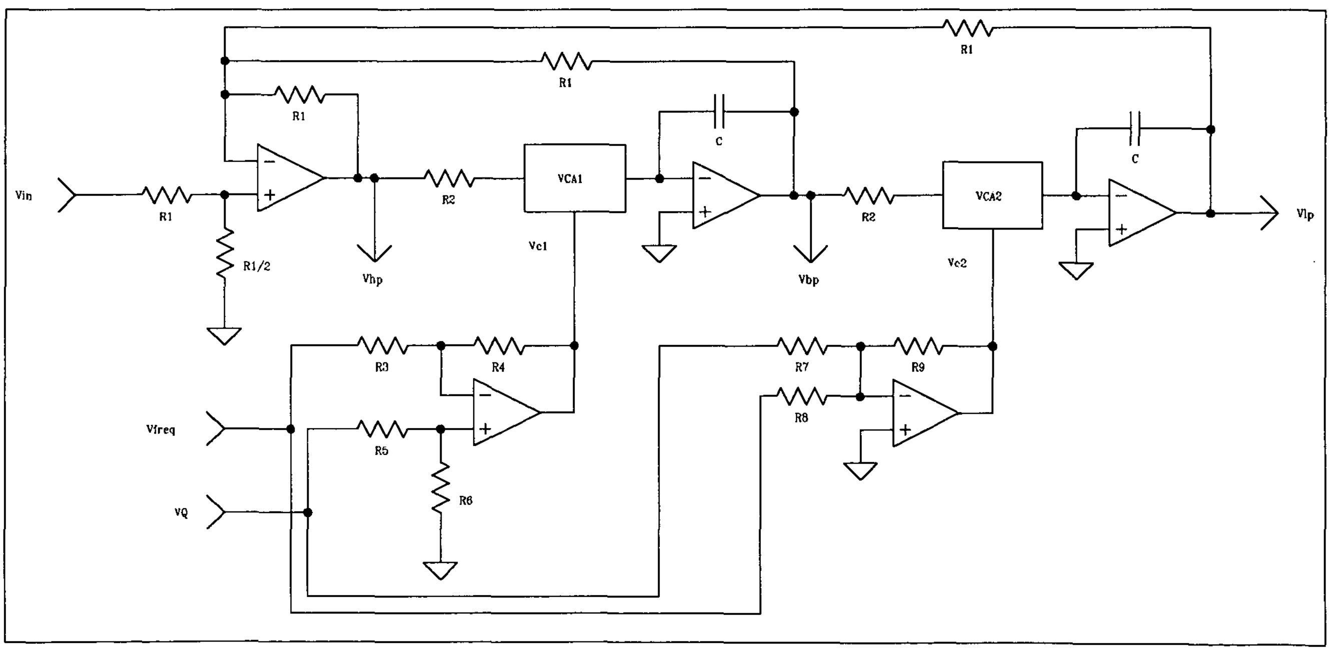

Figure 10 - VCA-Controlled State Variable Filter With Control-Voltage Conditioning Circuitry

The natural frequency and Q are now independent functions of V_{freq} and V_q, respectively. The natural frequency will again be conveniently varied in a "volts-per-octave" manner with V_{freq}. Varying the Q in this fashion is less familiar, but not difficult. For example, setting V_q to 0 volts (0 dB) sets Q=1, setting V_q to the voltage corresponding to 6 dB VCA gain sets Q=2, etc.

The values of R_2 and C should be chosen so that the VCAs are used primarily in attenuation mode, rather than requiring high VCA gain. This will maximize the bandwidth of the VCAs and minimize "Q-enhancement" problems. A useful approach is to choose R_2 and C such that:

$$\frac{1}{2\pi R_2 C} = f_{\max} \tag{23}$$

where f_{max} is the maximum natural frequency desired.

Figure 11 shows the state-variable filter adapted for MDAC control. Note that the feedback paths have been changed to reflect the fact that the integrators are now inverting rather than non-inverting. Independent control of natural frequency and Q is performed in the same manner, though exponential control is more computationally complex.

4.2 Parametric Equalizers

Figure 12 shows a parametric equalizer circuit utilizing the state-variable filter described above. Noting equation (13) (the transfer function to the bandpass output of the filter), the transfer function of the parametric equalizer with switch S_1 in the "boost" position is:

$$\frac{V_{out-boost}}{V_{in}} = 1 + \frac{(G_3 - 1)\frac{G_1}{R_2 C}s}{s^2 + \frac{G_1}{R_2 C}s + \left(\frac{\sqrt{G_1 G_2}}{R_2 C}\right)^2}$$

$$= \frac{s^2 + G_3\left(\frac{G_1}{R_2 C}\right)s + \left(\frac{\sqrt{G_1 G_2}}{R_2 C}\right)^2}{s^2 + \frac{G_1}{R_2 C}s + \left(\frac{\sqrt{G_1 G_2}}{R_2 C}\right)^2}. \tag{24}$$

where G_3= the current gain of VCA_3.

With switch S_1 in the cut position, the transfer function becomes:

$$\frac{V_{out-cut}}{V_{in}} = \frac{s^2 + \left(\frac{G_1}{R_2 C}\right)s + \left(\frac{\sqrt{G_1 G_2}}{R_2 C}\right)^2}{s^2 + G_3\left(\frac{G_1}{R_2 C}\right)s + \left(\frac{\sqrt{G_1 G_2}}{R_2 C}\right)^2}. \tag{25}$$

Evaluating the magnitude of both of these expressions at the natural frequency of the filter, $s = j\omega_n$, yields:

$$\left|\frac{V_{out-boost}}{V_{in}}\right|_{s=j\omega_n} = G_3 = e^{\frac{V_{c3}}{V_{scale}}} \tag{26}$$

and

$$\left|\frac{V_{out-cut}}{V_{in}}\right|_{s=j\omega_n} = \frac{1}{G_3} = e^{\frac{-V_{c3}}{V_{scale}}}. \tag{27}$$

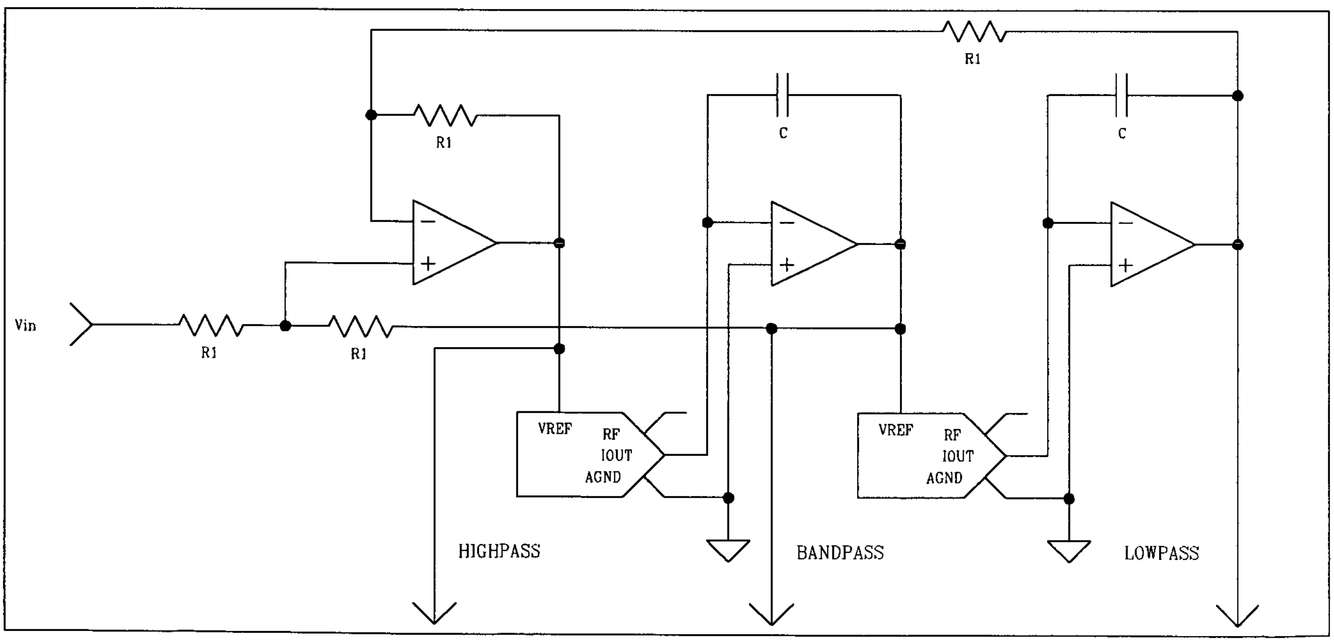

Figure 11 - MDAC-Controlled State Variable Filter

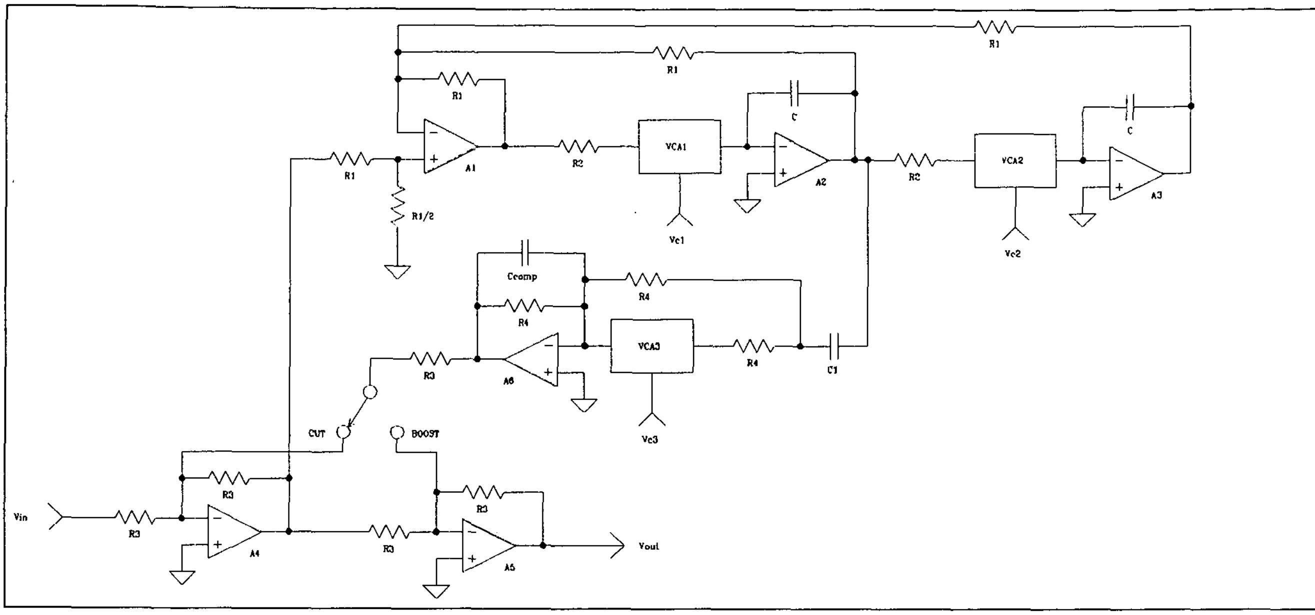

Figure 12 - VCA-Controlled Symmetrical Parametric Equalizer

Thus, the boost and cut at the center frequency are directly proportional to the gain of VCA_3 and can be controlled in a "volts-per-dB" manner via its control voltage. The gain of VCA_3 is always varied between 0 dB and the gain corresponding to the desired maximum boost -- even for cut operation. Switch S_1 would typically be implemented with a solid state switch, and the controller would be programmed to change the switch setting while the gain of VCA_3 is set to unity (G_3=1). Under these conditions there is no signal present at the output of A_6 due to the cancellation of signal currents at the summing junction of A_6.

This type of parametric equalizer seems to be the most prevalent, and produces symmetrical boost/cut characteristics as illustrated in Figures 13, 14, and 15. These measurements were made on a prototype of this circuit with the following values (referring to Figure 12): R_1=10 kΩ, R_2=4.99 kΩ, R_3=10 kΩ, R_4=20 kΩ, and C=1.5 nF. The measured 20kHz-bandwidth noise floor of this circuit was typically below -96 dBV when set for flat response, below -88 dBV when set for -15 dB cut, and below -85 dBV when set for +15 dB boost. Distortion was below .05% in the audio band under all modes.

If switch S_1 is kept in the boost position, and VCA_3's control voltage is varied to produce attenuation as well as gain, the characteristics revert to those of what are commonly known as asymmetric parametric equalizers. This type is preferred by some users. Figure 16 shows measurements from the same prototype illustrating this mode of

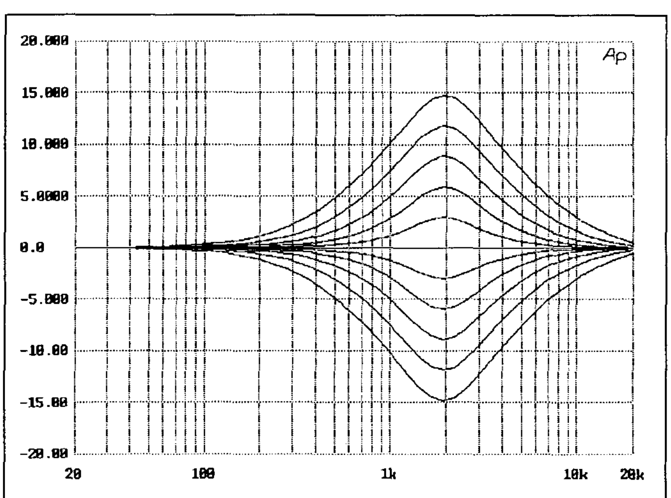

Figure 13 - Symmetrical Parametric EQ Responses for Boost and Cut (G_3 varied in 3 dB steps)

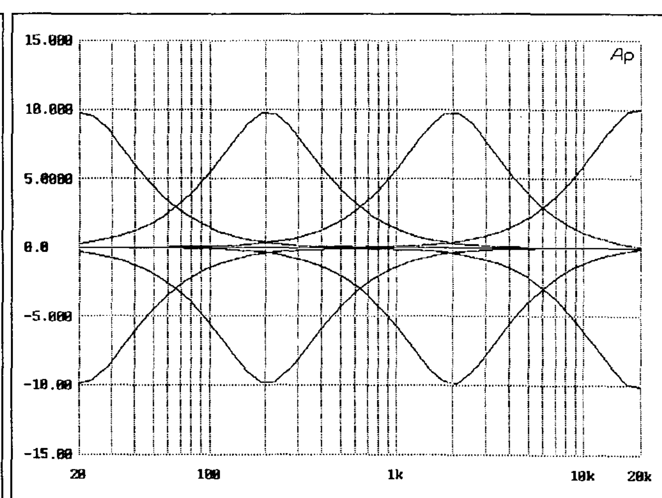

Figure 14 - Symmetrical Parametric EQ Responses (V_{freq} varied in "20 dB" steps)

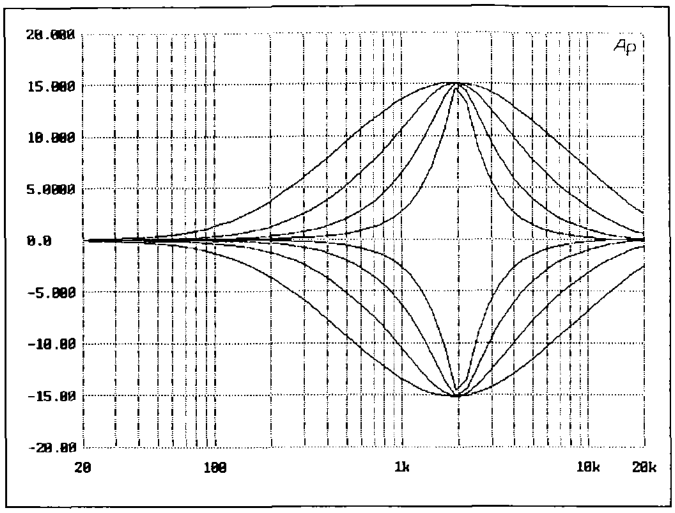

Figure 15 - Symmetrical Parametric EQ Responses (V_q varied in 6 dB steps - Q=.5, 1, 2, 4)

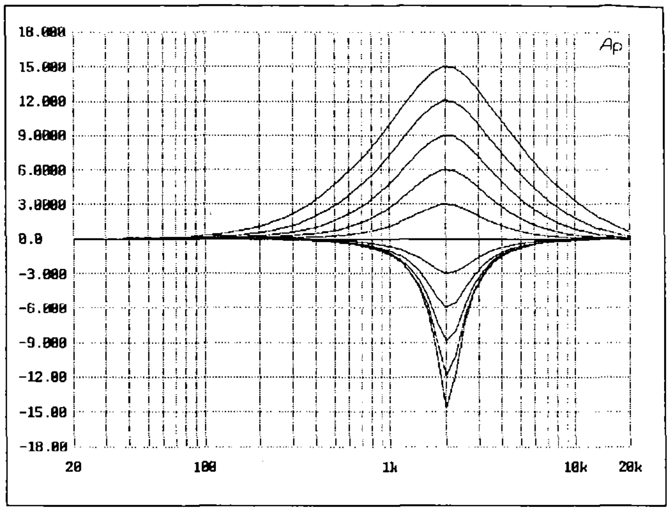

Figure 16 - Asymmetrical Parametric EQ Responses (G_3 varied from -15 dB to +15 dB in 3 dB steps)

operation. If only this mode of operation is desired, the circuit may be simplified as shown in Figure 17.

4.3 Variable Q Sallen and Key Filters

The circuit in Figure 18 is a Sallen and Key highpass filter adapted for VCA-controlled Q. If G is the current gain of the VCA, the transfer function of this circuit is:

$$\frac{V_{out}}{V_{in}} = \frac{s^2}{s^2 + \left(\frac{1-G}{R_1} + \frac{2}{R_2}\right)s + \frac{1}{R_1 R_2 C^2}}. \tag{28}$$

From this we derive the natural frequency (ω_n) and Q to be:

$$\omega_n = \frac{1}{\sqrt{R_1 R_2}\, C} \tag{29}$$

and

$$Q = \frac{\frac{1}{2}\sqrt{\frac{R_2}{R_1}}}{1 + (1-G)\frac{R_2}{2R_1}}. \tag{30}$$

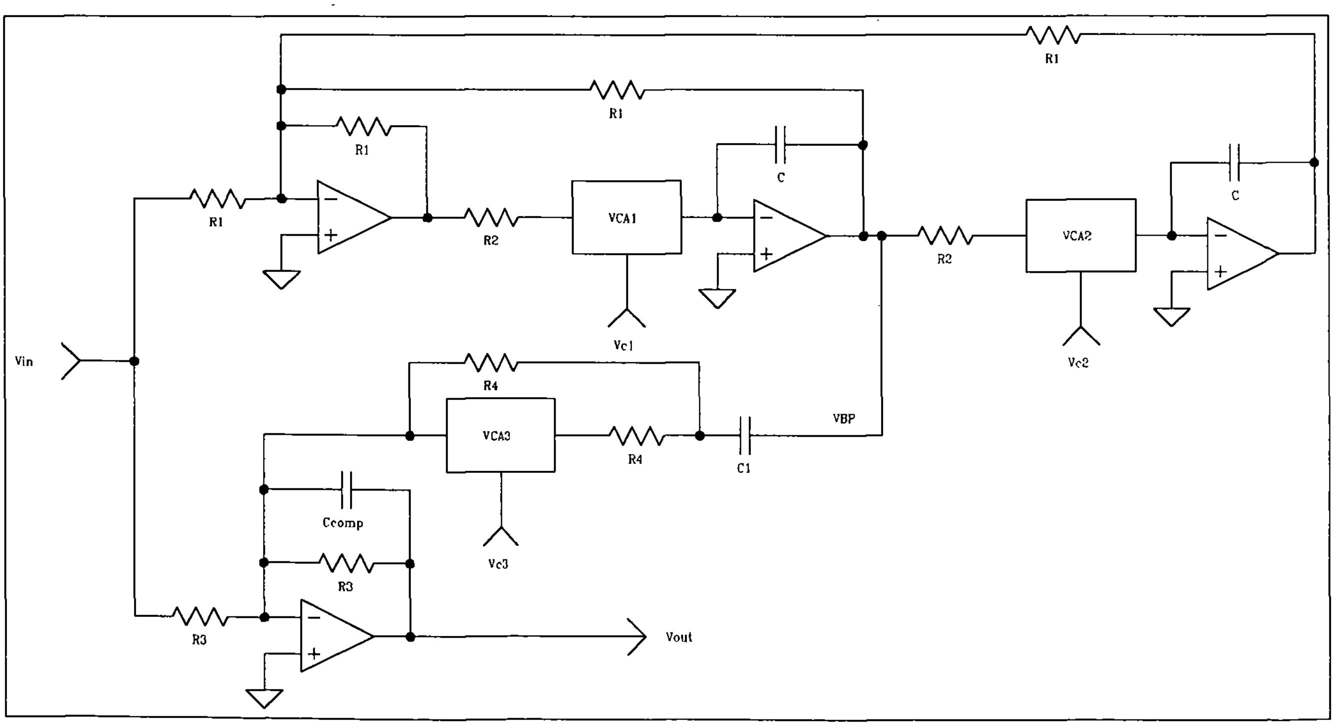

Figure 17 - Simplified VCA-Controlled Asymmetrical Parametric Equalizer

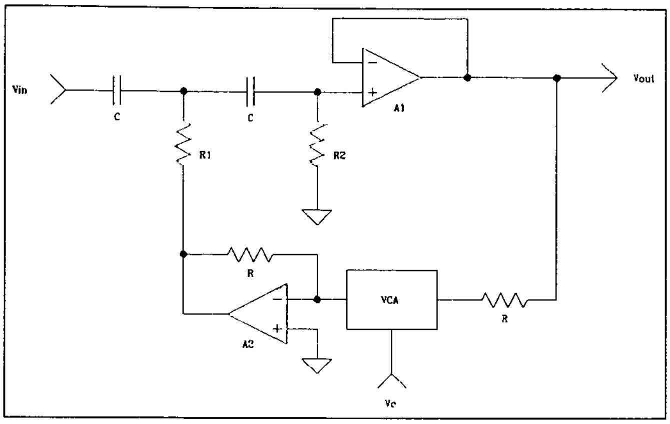

Figure 18 - Variable-Q Sallen and Key Highpass

For G=1, this reduces to the familiar expression for the Q of a Sallen and Key highpass filter:

$$Q = \frac{1}{2}\sqrt{\frac{R_2}{R_1}} \quad (31)$$

The curves in Figure 19 illustrate how Q varies with VCA gain. For these curves R_1 and R_2 were chosen to be 3.32 kΩ and 53.6 kΩ respectively, setting the initial Q (at 0 dB VCA gain) to 2. The VCA gain is stepped from 0 dB to -4 dB is 1 dB steps. Q can also, of course, be increased by increasing VCA gain. Caution should be observed, however, since if G exceeds $\frac{2R_1}{R_2} + 1$ the circuit becomes an oscillator.

A corresponding lowpass version is, of course, also possible.

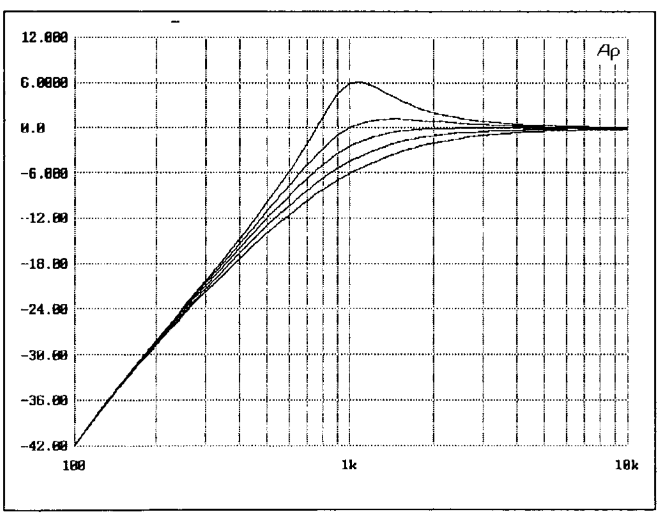

Figure 19 - Variable-Q Highpass Responses (VCA gain varied from 0 dB to -4 dB in 1 dB steps)

5. SUMMARY AND CONCLUSIONS

The authors have presented several examples illustrating the use of two-quadrant multipliers to implement digital control of analog audio filters. It has been shown that, with minor modifications, many of the familiar equalization and filter circuits that have become a standard part of the audio engineer's toolbox can be adapted to digital control.

Two common versions of the two-quadrant multiplier were discussed. The MDAC, which performs its multiplication in discrete steps, yields minimal performance degradation due to its essentially passive nature. This comes at the cost of the potential for zipper noise, and, in applications requiring fine exponential control of amplitude or frequency parameters, higher cost for high-resolution DACs and more complex control firmware.

The combination of VCAs and inexpensive control DACs offers continuous transitions between discrete parameter settings, thus avoiding zipper noise. VCAs with an exponential gain-control characteristic offer direct decibel control, yielding high resolution control of filter parameters over a wide range. This also simplifies controller firmware for frequency and amplitude parameters that are traditionally dealt with in exponential fashion. The performance available with today's VCAs rivals or exceeds current DSP-based systems at much lower cost.

It is hoped that these examples will serve as a starting point for engineers wishing to enhance the functionality of their designs by adding the flexibility that digital control brings.

6. REFERENCES

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