VCA-Controlled 1st Order State Variable Filter

The circuits within this application note feature THAT218x to provide the essential function of voltage-controlled amplifier (VCA). Since writing this note, THAT has introduced a new dual VCA, as well as several Analog Engines®. Analog Engines combine a VCA and an RMS detector (RMS) with optional opamps in one part. With minor modifications, these newer ICs are generally applicable to the designs shown herein, and may offer advantages in performance, cost, power consumption, etc., depending on the design requirements. We encourage readers to consider the following alternatives in addition to the 218x:

- Analog Engine (VCA, RMS, opamps): 4301
- Analog Engine with low supply voltage and low power (VCA, RMS, opamps): 4320
- Analog Engine with low cost, low supply voltage, and low power (VCA, RMS): 4315
- Analog Engine with low cost and low power (VCA, RMS): 4305
- Dual (VCA only): 2162

For more information about making these substitutions, please contact THAT Corporation's technical support group at apps_support@thatcorp.com

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VCAs are widely used throughout the audio signal chain to control amplitude, especially where dynamic gain control is desired. A less well known application of these devices is in dynamically controllable filters. VCAs ease the design of these circuits, and avoid the "zipper noise" which plagues circuits using stepped attenuators.

The Theory

Figure 1 shows the basic VCA-based first order state variable filter. For our purposes, we can state the transfer function of the VCA as being

$$\frac{VCA_{OUT}}{VCA_{IN}} = G \cdot \frac{Z_{OUT}}{Z_{IN}}$$

where G is the gain of the VCA.

This is much like the transfer function of an inverting amplifier, but without the sign change, and with an additional, variable gain term. This lack of inversion simplifies the circuit and its analysis, since U2A can now be a simple inverting summer, rather than the differential stage normally found at this location. We can state the high pass output to be

$$HighPass = Input \cdot \left(\frac{-R}{R}\right) + LowPass \cdot \left(\frac{-R}{R}\right) = -Input - LowPass \qquad \text{(equation 1)}$$

and the low pass output to be

$$LowPass = HighPass \cdot \left(\frac{\frac{1}{sC_{SET}}}{R_{SET}}\right) G = HighPass \cdot \left(\frac{G}{sC_{SET}R_{SET}}\right) \qquad \text{(equation 2)}$$

Substituting, we find that

$$HighPass = -Input - HighPass \cdot \left(\frac{G}{sC_{SET}R_{SET}}\right)$$ and

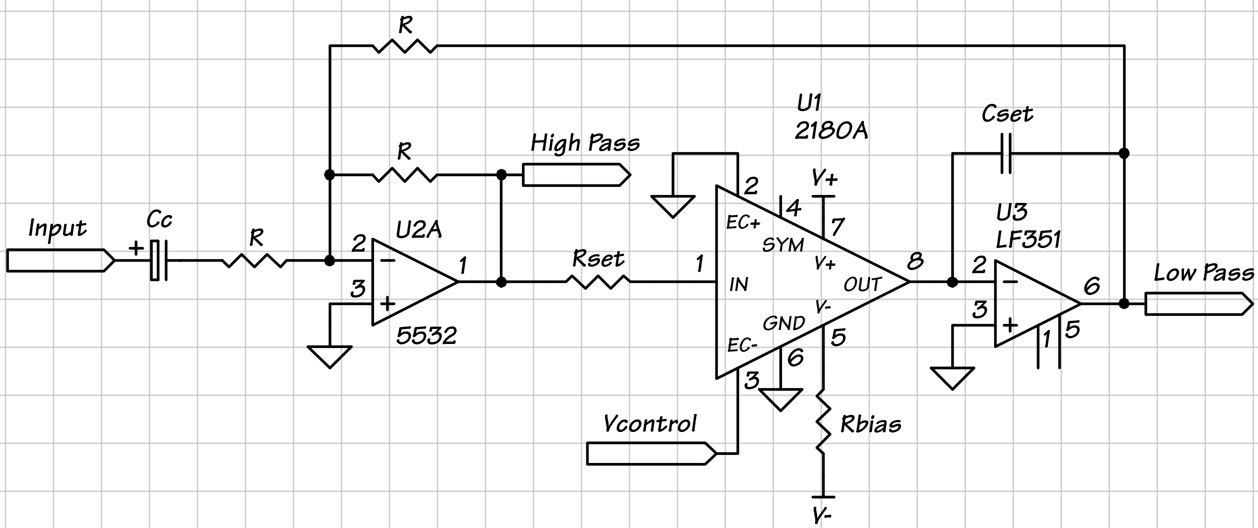

Figure 1. VCA-controlled 1st order state variable filter

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$$HighPass \cdot \left[1+\left(\frac{G}{sC_{SET}R_{SET}}\right)\right] = -Input$$

This result can be reduced to

$$\frac{HighPass}{Input} = \frac{-1}{1+\left(\frac{G}{sC_{SET}R_{SET}}\right)} = \frac{-S}{S+\left(\frac{G}{C_{SET}R_{SET}}\right)}$$

which is of the form of a typical first order high pass filter, but with the corner frequency dependent on VCA gain. Combining equations 1 and 2, we can similarly derive the lowpass output to be:

$$LowPass\left(\frac{sC_{SET}R_{SET}}{G}\right) = -Input - LowPass$$

which after cross-multiplication yields

$$LowPass = -Input \cdot \left(\frac{G}{sC_{SET}R_{SET}}\right) - LowPass \cdot \left(\frac{G}{sC_{SET}R_{SET}}\right)$$

which becomes

$$LowPass \cdot \left[1+\left(\frac{G}{sC_{SET}R_{SET}}\right)\right] = -Input\left(\frac{G}{sC_{SET}R_{SET}}\right)$$

This result may be reduced to

$$\frac{LowPass}{Input} = \frac{-\left(\frac{G}{sC_{SET}R_{SET}}\right)}{1+\left(\frac{G}{sC_{SET}R_{SET}}\right)} = \frac{-\left(\frac{G}{C_{SET}R_{SET}}\right)}{S+\left(\frac{G}{C_{SET}R_{SET}}\right)}$$

which is of the form of a typical first order low pass filter, but again with the corner frequency dependent on VCA gain.

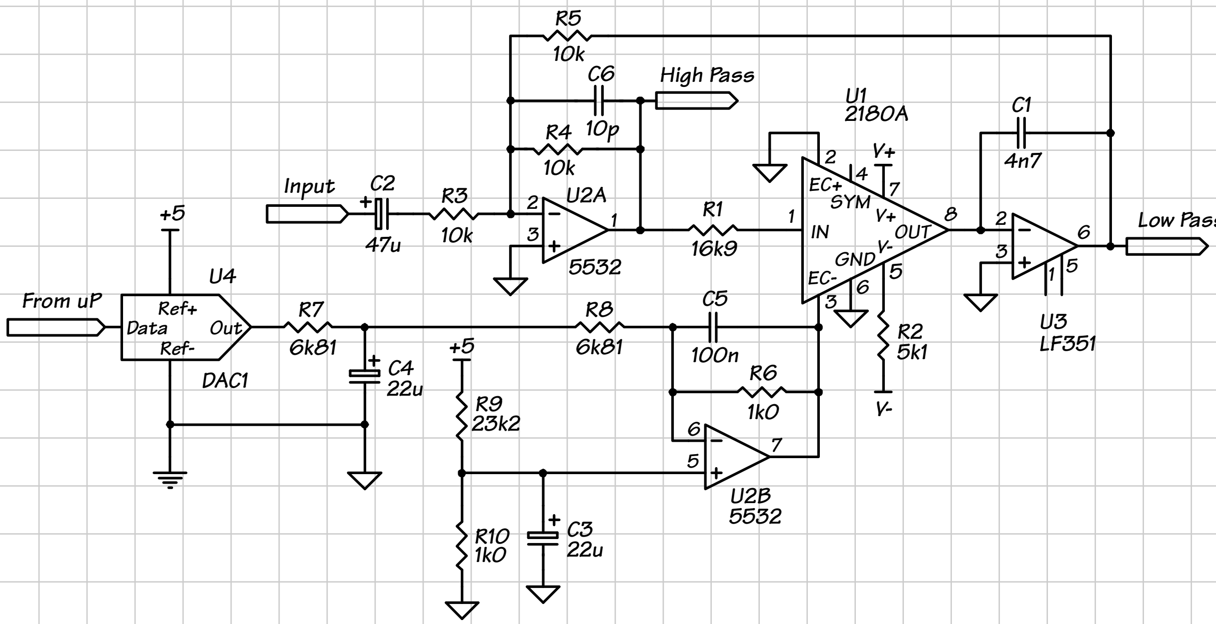

Figure 2. VCA-controlled 1st order state variable filter with conditioned control port drive

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Some Thoughts on the Actual Design

Consider the circuit in Figure 2. U2A is our highpass output, and we've arbitrarily chosen R3-R5 to be 10 kΩ. The small value of C6 does not substantially affect the calculations of the previous section, but does reduce peaking that can occur well above the audio band with some op-amps.

Since corner frequency is proportional to gain, we can sweep the entire audio band with a total gain change of 60 dB. We've chosen gains from -40 dB to 20 dB to accomplish this. U2B is used as the control port buffer. A 0-5 V_{DC} swing for the DAC is attenuated by U2A's inverting gain, which is 0.073. The divider formed by R9 and R10 is used to offset U2As swing by 40 dB to center the control signal appropriately. C3 and C4 provide filtering, both to attenuate interference from the digital section, and to mitigate the effects of the DAC's discrete steps.

You'll need to use low bias current, FET input op-amps as the VCA's output amplifier. This is because this op-amp is configured as an integrator, and its bias current must come through the VCA. High bias currents will result in a problem with "thump" (or offset-change-with-gain), and at very low VCA gains the output amplifier may clip.

The maximum bias current of the LF351 at room temperature is 200 pA, though one should note that this figure typically doubles every 10 °C. Since the bias current cannot flow through C1, it must come through the VCA input and R1. This is accomplished by a small DC correction voltage developing on C1 which, after being inverted by U2A and developing across R1, results in a correction current passing through the VCA. In essence, the DC level on output of U3, inverted by U2A and then developed across R1, "servos" out the LF353's bias current.

Assume that we wish for our circuit to exhibit less than 5 mV of offset-change-with-gain at 55°C. At 55°C, the bias current could be as high as

$$i_{bias} = 200pA \cdot 2^{\left(\frac{55-25}{10}\right)} = 1.6nA$$

which, must come from the output of the VCA. Remember that this current must pass through the input of the VCA, and that the VCA will attenuate or amplify the input current by an amount proportional to the VCA's gain. Thus, the current into the VCA input would need to be

$$i_{bias} = \frac{V_{THUMP}}{R_{IN}} \cdot A_V = \frac{5mV}{16.9k} \cdot 10^{\left(\frac{G_{MAX}}{20}\right)}$$

Isolating G_{MAX}

$$G_{MAX} = 20\log\left[I_{bias} \cdot \left(\frac{16.9k}{5mV}\right)\right] = -45dB$$

where G_{MAX} is the maximum allowable attenuation while keeping thump below 5 mV.

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In addition to the bias current of the output op-amp, other sources of leakage should be considered. For instance, pin 7 of the VCA is tied to V_{CC} which, in this circuit, is 15 V_{DC}. Being right next to the VCA output, board contamination can result in substantial leakage currents. This is best dealt with by adding a guard on both sides of the PCB. This guard should be tied to ground (pin 3 of the LF351), and should run completely around pin 8 of the VCA, pin 2 of the op-amp, and the one leg of C1 that is tied to this node. For optimal guard performance, the PCB's solder mask should be removed from the guarded area.

We want our filter to be centered in the audio band when the DAC is at half scale, which corresponds to a gain of -10 dB. The geometric center of the audio band in dB is 632 Hz. If we choose C1 to be 4.7 nF, then we can calculate

$$R1 = \frac{G}{2 \cdot \pi \cdot f \cdot C_1} = \frac{10^{\frac{-10}{20}}}{2 \cdot \pi \cdot 632 \cdot 4.7 \times 10^{-9}} \cong 17\ \mathrm{k}\Omega$$

Results

Figures 3 and 4 show the low pass and high pass responses with gain swept from -40 dB to 20 dB in 10 dB steps. Possible applications for this circuit include dynamically programmable filters, single ended noise reduction, and effects such as auto-wah's.

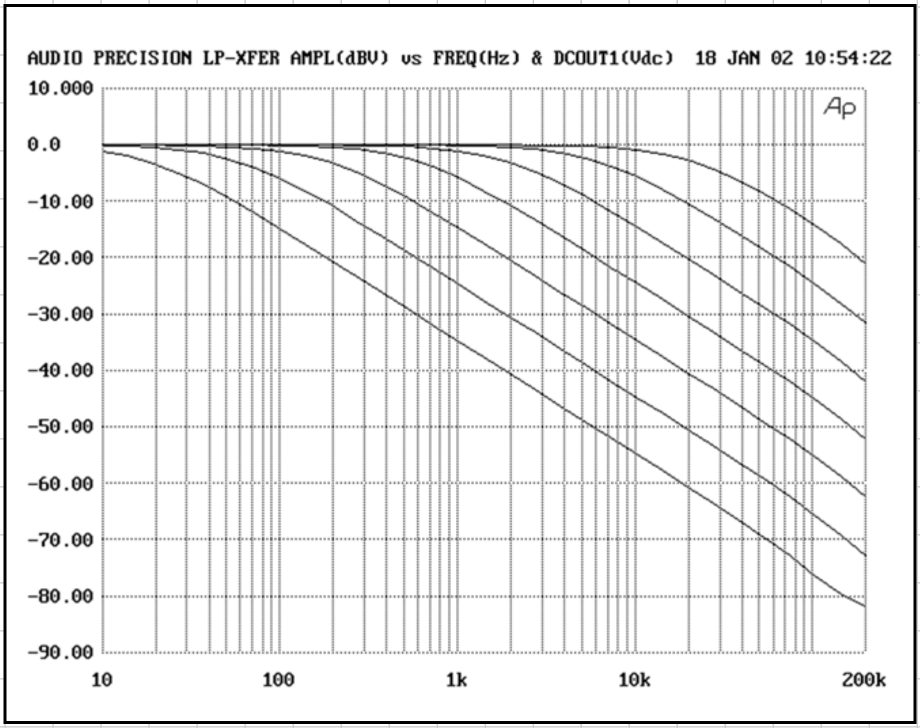

Figure 3. Frequency response of the low pass output with gain swept from -40 dB to 20 dB in 10 dB steps

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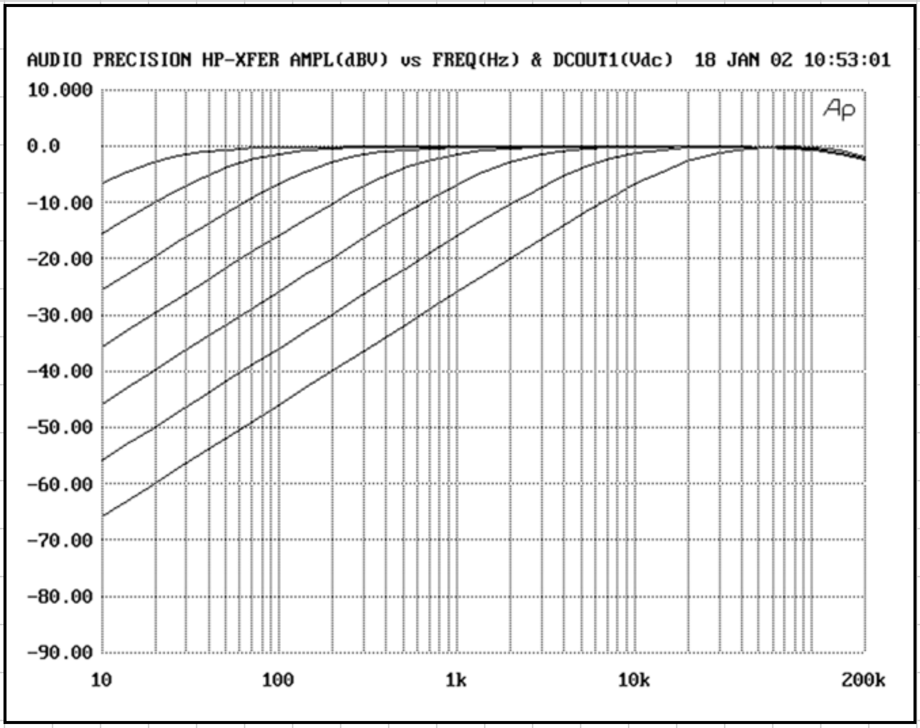

Figure 4 Frequency response of the high pass output with gain swept from -40 dB to 20 dB in 10 dB steps

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