Variable Active Filter Designs

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Midterm Practical Design

EECE226L-001, Spring 2023

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Background:

As a self-described “degree-holding audio enthusiast”, I have long been interested in the technical aspects of filtering and amplification in modern designs. Collecting operational amplifiers of wide-ranging performance (and equally varied reputation) has been a longtime hobby, and this seems an excellent occasion to test and quantify characteristics of some of the more notable op-amps widely used in the audio industry, or in some cases simply tumbling across the electronics wasteland that is my desk. Active filters present a unique opportunity – they may be designed for any gain behavior across the frequency spectrum – a variable design is even better. Immediately I considered some of the more unusual effects pedals utilized by musicians, those implemented in legendary synthesizers, and employed by the most creative recording engineers. I would want something able to provide consistent, narrow bandwidth amplification, and be adjustable across the audible frequencies occupied by typical synthesizers; roughly 50Hz through 4200Hz, though the entire audible spectrum would be a plus. Such a filter might be able to be used in conjunction with a noise generator for the input, and some gain normalization on the output, to create a synthesizer with a unique sound.

A filter with an adjustable center frequency seems ideal. The Sallen-Key topology is able to support setting the Q without affecting the center frequency. Although Q and the gain at the center frequency cannot be changed independently, band-pass designs allow varying the center frequency with a single resistor. A design such as this would allow building two such filters into a single “dual” op-amp package, while varying both center frequency resistors with a single 2-gang potentiometer. This would comprise a stereo filter with identical performance and characteristics for both channels and could take advantage of some of the higher-end audio op-amps’ crosstalk rejection capability. A state-variable filter topology, such as the Kerwin–Huelsman–Newcomb design, would be another design worth investigating, as Q can be precisely set and controlled while the center frequency may be varied independently, although gain is again linked to Q. These designs sum a signal with its integral and double-integral, and being biquad designs, require multiple op-amps to build, but are more tolerant of lower-performance ICs. A stereo filter design using this topology would require 3 dual op-amps and a 4-gang potentiometer. The integrator op-amps may be replaced with control voltage sources, as is usually the case with modular synthesizers – this would be a desirable option to pursue at a later date. Regardless, as the filters would need to sweep the desired range and provide consistent gain, testing a handful of op-amps for consistent center frequency gain across the sweep seems like a good first step.

Op-amps on-hand to be tested and/or simulated include:

Texas Instruments:	LM833N
	LM4562NA
	LME49720NA (die identical to 4562)
Japan Radio Company:	NJM/JRC4556AD
	NJM/JRC4580D
Burr-Brown:	OPA2132P
	OPA2277P

Design Specifications:

High-Q, (30 to 50, option to adjust with potentiometer a plus)

High gain, (+20 to +40 dB, option to adjust with potentiometer a plus, linked to Q acceptable)

Variable center frequency (80Hz through 4200Hz, 50Hz through 20kHz a plus)

Consistent Q and gain across variable frequency (less than +/- 5 dB variation from min. to max.)

Preliminary Testing:

Op-amps are known to vary in performance and may not be able to maintain consistent output across these frequencies. A testing scheme was devised wherein each op-amp was simulated in a unity-gain low-pass Sallen-Key filter circuit [1, fig. 3.23] with a high-Q setting and low center frequency. The filter characteristics can be calculated and compared against the simulation result to make a determination of the op-amps' performance. If an op-amp should fall short of the expected response, or if the center frequency varies significantly, then the op-amp may be unsuitable for the design specifications. If any should perform better than the rest by a wide margin, then they would likely be best suited for building the practical circuit.

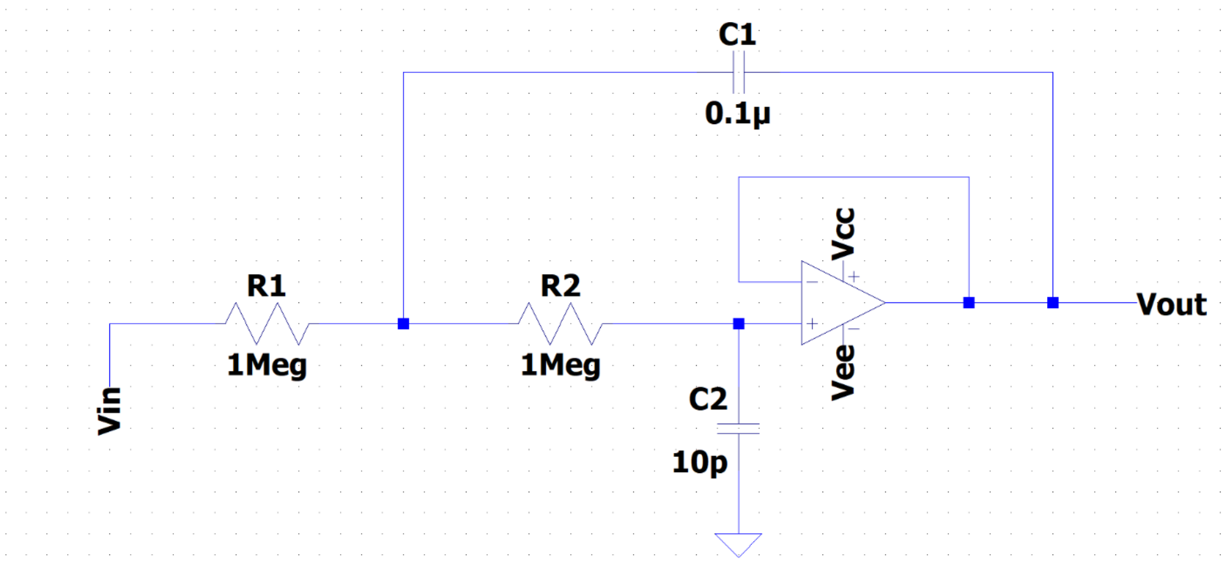

Figure 1: Simulation test circuit. Sallen-Key low-pass with low-frequency, high-Q setting.

Here, $K = 1$, and $R_1 = R_2 = R$. To achieve a high Q setting from on-hand values, C_1 and C_2 are set quite disparately. For a low-pass Sallen-Key, the center frequency is given by:

$$\omega_0 = \left(\sqrt{R_1C_1R_2C_2}\right)^{-1} \rightarrow \omega_0 = \left(R\sqrt{C_1C_2}\right)^{-1} \qquad \text{[1, eqn. 3.60b]}$$

$$\omega_0 = \left(1M\sqrt{(0.1\mu)(10p)}\right)^{-1} = 1000\, rad/s = 159.155\, Hz$$

Q is given by:

$$Q = \left((1-K)\sqrt{R_1C_1/R_2C_2} + \sqrt{R_1C_2/R_2C_1} + \sqrt{R_2C_2/R_1C_1}\right)^{-1} \rightarrow Q = \left(2\sqrt{C_2/C_1}\right)^{-1} \qquad \text{[1, eqn. 3.60c]}$$

$$Q = \left(2\sqrt{(0.1\mu)(10p)}\right)^{-1} = 50$$

And the transfer function itself is given by:

$$\frac{V_o}{V_i}(\boldsymbol{j}\omega) = \frac{K}{\left(R_1C_1R_2C_2(\boldsymbol{j}\omega)^2 + \boldsymbol{j}\omega\left((1-K)R_1C_1 + R_1C_2 + R_2C_2\right) + 1\right)} \qquad \text{[1, pg. 143]}$$

$$\frac{V_o}{V_i}(\boldsymbol{j}\omega) = \left(R^2C_1C_2(\boldsymbol{j}\omega)^2 + \boldsymbol{j}\omega(2RC_2) + 1\right)^{-1}$$

$$\frac{V_o}{V_i}(\boldsymbol{j}1000) = \left((1M)^2(0.1\mu)(10p)(\boldsymbol{j}1000)^2 + \boldsymbol{j}1000\left(2(1M)(10p)\right) + 1\right)^{-1} = (50\angle -90°)\, V/V$$

For a gain (in dB) of:

$$20\log(50) = 33.979\, dB$$

We should expect to see a center frequency response near 159.155 Hz, Q approaching 50, and magnitude approaching 33.979 dB.

Spice directives were applied to directly find measurements of interest for each op-amp simulated. Namely: center frequency, bandwidth, Q, and magnitude. The circuit was duplicated in LTSpice to allow plotting and measuring of all op-amps in the same simulation, providing a visual comparison of responses. The results of this simulation are presented below, and on the next page.

Op-Amp	f_0 (Hz)	BW (Hz)	Q (f/BW)	V_o/V_i	dB
LM833	**159.074**	3.34386	47.5721	47.5887	33.5501
LM4562	157.543	3.22595	**48.8362**	48.8509	**33.7774**
NJM4556	158.782	8.03165	19.7695	19.7885	25.9282
NJM4580	158.855	10.7173	14.8223	14.8482	23.4335
OPA2132	159.001	3.31353	47.9854	48.0165	33.6278
OPA2277	138.420	3.19061	43.3837	43.4074	32.7513

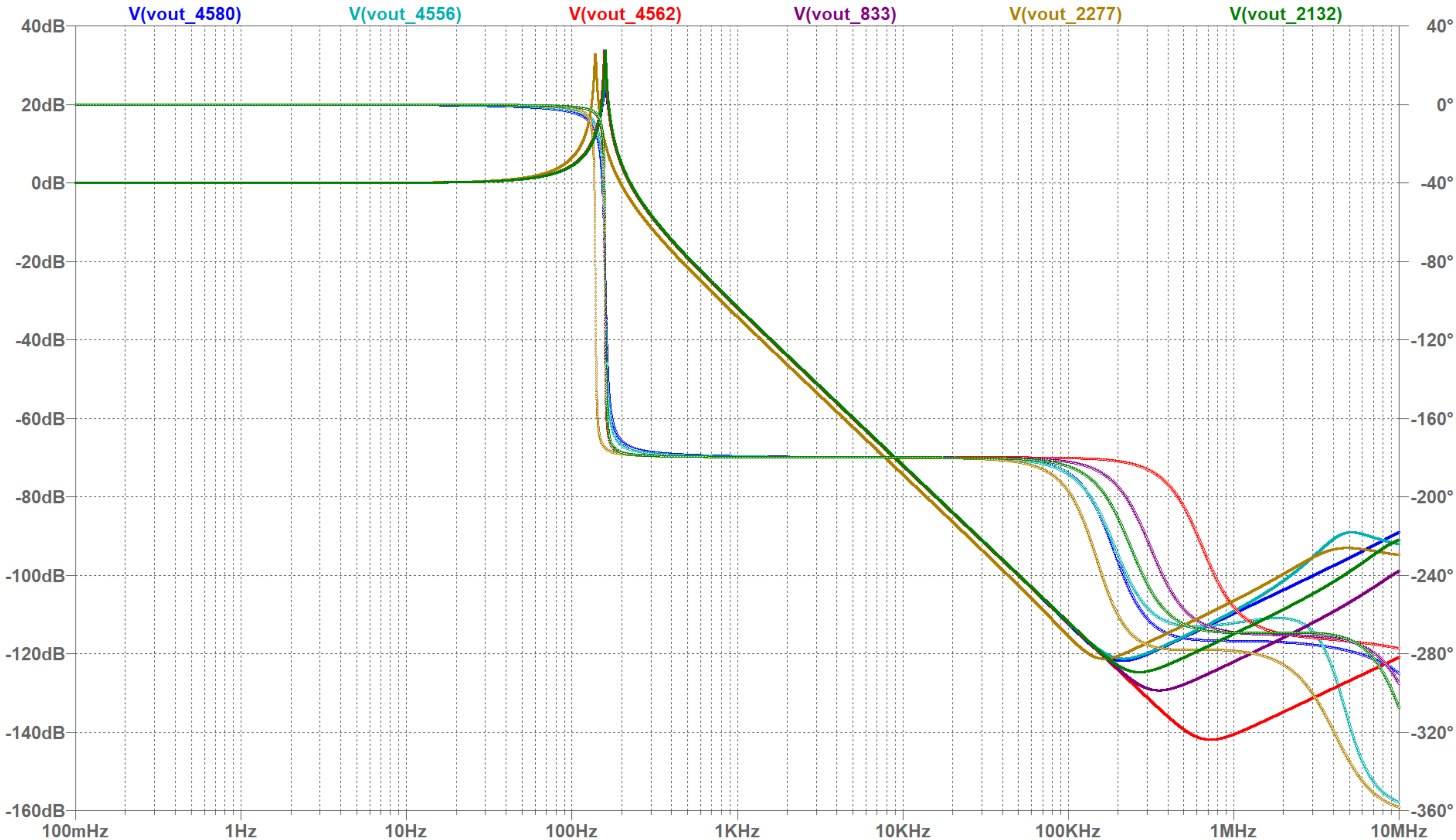

Figure 2: LTSpice bode plot of op-amps under test, showing low-pass high-Q characteristics, with variation in center frequency, Q, peak magnitude, and phase.

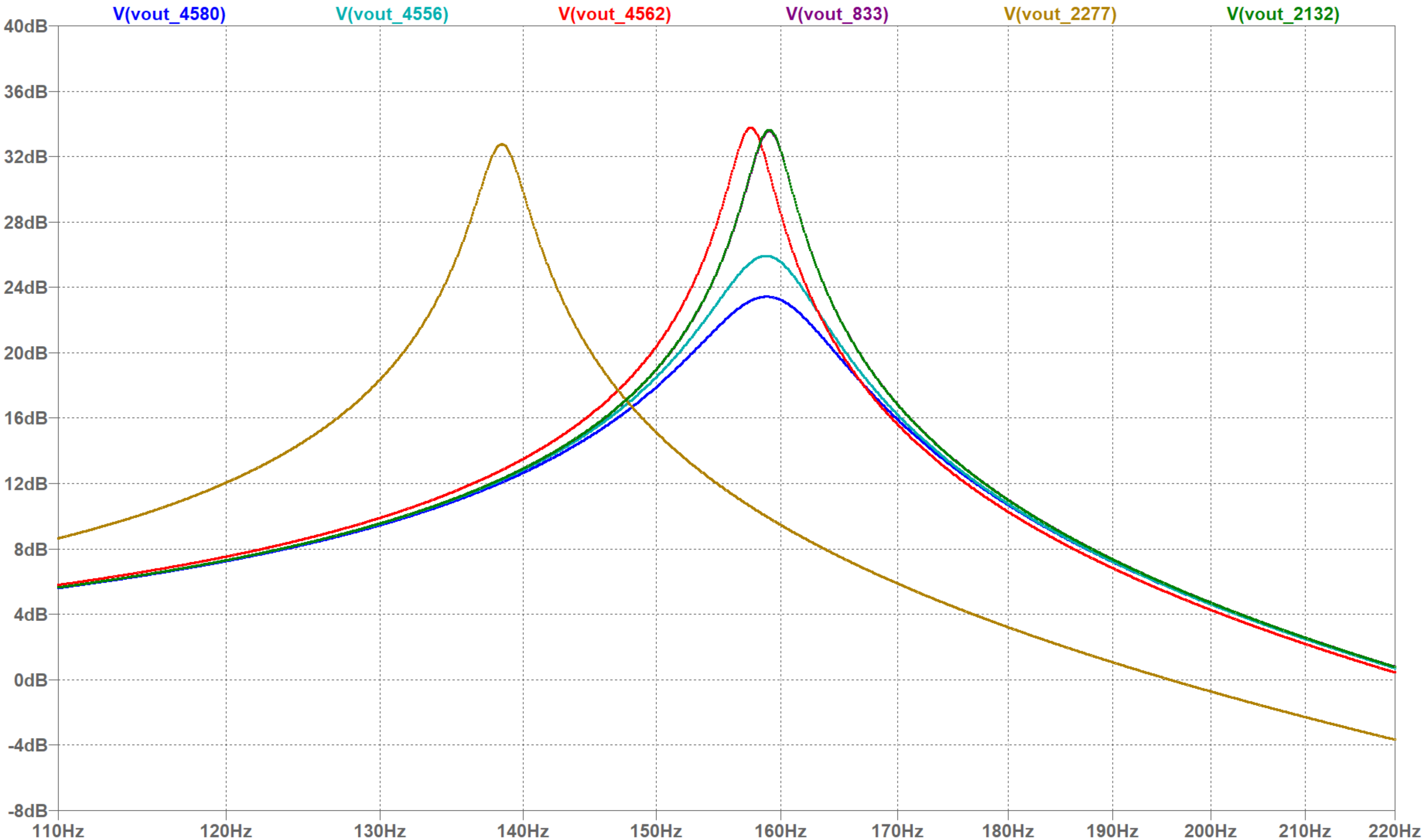

Figure 3: The same simulation plot, with attention to the peak response at the calculated center frequency. [Note: LM833 obscured by nearly identical OPA2132 trace]

Low frequency, high-Q circuits appear to present a challenge for op-amps. It is clear that NJM4556 and NJM4580 are not capable of meeting the Q specification, reaching only 19.7695 and 14.8223 respectively, although they were able to produce an accurate center frequency peak response. Also note OPA2277's unusual 138.420Hz peak response. The same circuit was tested with $R = 6.8\ k\Omega$ to observe high frequency performance, however all op-amps had little issue achieving $Q \cong 50$, although OPA2277 and NJM4556 both showed lower center frequency peak responses. This seems reason enough to strongly consider the remaining LM833, LM4562/LME49720, and OPA2132 op-amps for continued testing, but perhaps a detailed comparison of the high frequency results will show a singular best-in-class selection.

For $R = 6.8\ k\Omega$, the expected results are:

$$\omega_0 = \left(6.8k\sqrt{(0.1\mu)(10p)}\right)^{-1} = 147.059\ krad/s = 23.405\ kHz$$

$$Q = \left(2\sqrt{(0.1\mu)(10p)}\right)^{-1} = 50$$

$$\frac{V_o}{V_i}(\boldsymbol{j}147.059k) = \left((6.8k)^2(0.1\mu)(10p)(\boldsymbol{j}147.059k)^2 + \boldsymbol{j}147.059k\big(2(6.8k)(10p)\big) + 1\right)^{-1}$$

$$\frac{V_o}{V_i}(\boldsymbol{j}147.059k) = (50\angle -90°)\ V/V$$

For a gain (in dB) of:

$$20\log(50) = 33.979\ dB$$

We should expect to see a center frequency response near 23.405 kHz, Q approaching 50, and magnitude approaching 33.979 dB. The results are again presented below and on the next page, and show LM4562 as nearest to the majority of calculated values. LM4562/LME49720 will be selected for continued design and built circuits.

Op-Amp	f_0 (kHz)	BW (Hz)	Q (f/BW)	V_o/V_i	dB
LM833	21.518	414.146	51.9573	51.9658	34.3144
LM4562	**22.699**	447.797	50.6896	50.7114	**34.1021**
NJM4556	19.934	400.240	**49.8057**	49.7801	33.9411
NJM4580	21.787	426.795	51.0483	51.0138	34.1538
OPA2132	21.301	401.064	53.1112	53.1073	34.5031
OPA2277	18.450	346.409	53.2611	53.2733	34.5302

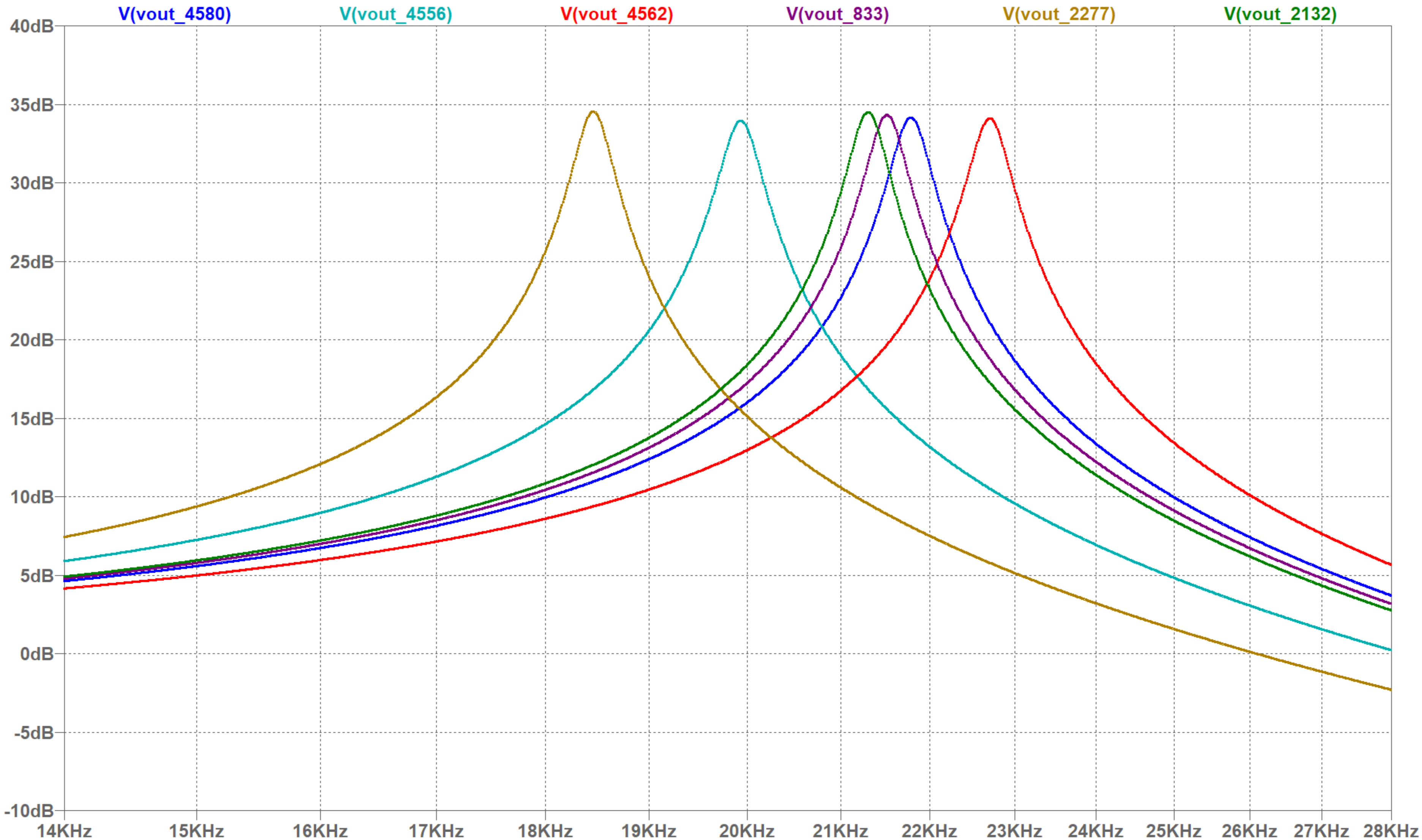

Figure 4: High frequency simulation plot with attention to peak response, showing LM4562/LME49720 nearest to the calculated center frequency of 23.405 *kHz.*

Design:

The low-pass Sallen-Key circuit was acceptable for testing op-amps, but was unable to meet the design specifications without resorting to a very large range on the input resistor network, making it more difficult to implement as adjustable with high Q, when using a dual-gang potentiometer. Band-pass variations were attempted, but varying frequency proved unwieldy and again required a wide range of resistor network values. Some improvement was found with a multiple-feedback design, however the resistor network required vastly different values, out of proportion to one-another, to adjust the center frequency without affecting all other characteristics, rendering it difficult to implement with any kind of multi-gang potentiometer. Ultimately, a simplified state-variable Kerwin–Huelsman–Newcomb filter [2][3] was found to provide the specified characteristics, similar to the multiple-feedback design, while being able to be controlled with a dual-gang potentiometer per channel. The multiple integrator topology allowed for discrete variation of filter characteristics, which afforded rapid progress in achieving specifications.

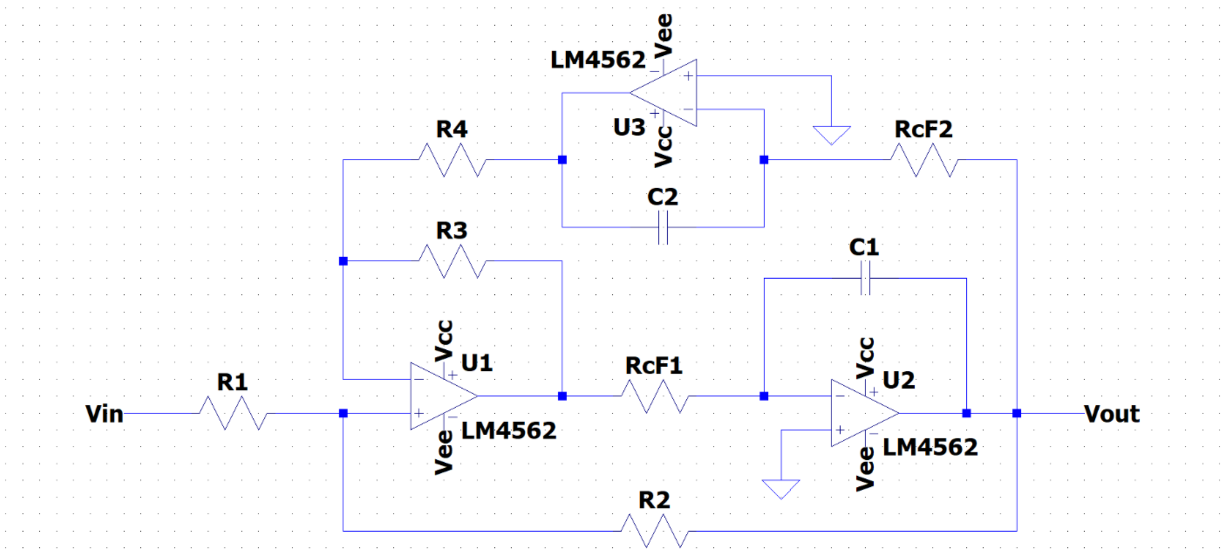

Figure 5: Simulation design circuit. Simplified state-variable filter.

Note that the first stage, U_1, is a summing amplifier with the outputs from the second stage integrator, U_2, and V_{in} as non-inverting inputs, and the output from the third stage integrator, U_3, as an inverting input. Essentially, U_1 sums V_{in} with the band-pass response from U_2, and subtracts the low-pass response from U_3. The two integrators may be designed with the same time constant, which provides a single center frequency rather than two corners – this comprises the band-pass characteristic by integrating the high-pass response of the summed signals.

This also means that each integrator stage has its corner frequency set independently from the rest of the circuit. If they are set the same as each other, where $C_1 = C_2 = C$ and $R_{f_{c1}} = R_{f_{c2}} = R_{f_c}$, then the circuit's center frequency can be determined. This means that the state-variable filter has the same formula for center frequency as an integrator filter's corner frequency:

$$\omega_c = \left(R_{f_c} C\right)^{-1} \qquad \text{[3, img. 4]}$$

Consider the audible frequency range of 50Hz to 20kHz, and say $C = 0.1\ \mu F$:

$$\omega_{c1} = 2\pi 50 = 314.159\ rad/s$$

$$\omega_{c2} = 2\pi 20k = 125.664\ rad/s$$

The range of R_{f_c} needed for this filter can be determined:

$$\omega_c = \left(R_{f_c} C\right)^{-1} \rightarrow R_{f_c} = (\omega_c C)^{-1}$$

$$R_{f_{c1}} = \left((314.159)(0.1u)\right)^{-1} = 31830.99\ \Omega$$

$$R_{f_{c2}} = \left((125.664k)(0.1u)\right)^{-1} = 79.577\ \Omega$$

Already this is an improvement over the Sallen-Key filter, as the resistance values required for the entire audible spectrum are much lower. Consider the narrower range, occupied by synthesizer instruments, 80Hz through 4200Hz:

$$R_{f_{c1}} = \left((502.655)(0.1u)\right)^{-1} = 19894.37\ \Omega$$

$$R_{f_{c2}} = \left((26389.378)(0.1u)\right)^{-1} = 378.94\ \Omega$$

4-gang potentiometers with a range up to 20 $k\Omega$ are available, so this design seems worth pursuing with $C = 0.1\ \mu F$, but perhaps it can be improved. Selecting $C = 0.33\ \mu F$ may work out even more conveniently:

$$R_{f_{c1}} = \left((314.159)(0.33u)\right)^{-1} = 9645.76\ \Omega$$

$$R_{f_{c2}} = \left((125.664k)(0.33u)\right)^{-1} = 24.11\ \Omega$$

This allows varying from 50Hz through 20kHz with a $10k\Omega$ potentiometer, which are inexpensive and commonly available with good precision in a variety of different packages and tapers, including linear and audio tapers.

The Q-factor is related to the damping coefficient as:

$$2\xi = Q^{-1}$$

Where 2ξ is represented by the resistor network as:

$$2\xi = \frac{1}{Q} = \frac{R_1(R_3 + R_4)}{R_4(R_1 + R_2)}\sqrt{\frac{R_3}{R_4}} \qquad \text{[3, img. 21]}$$

And set the two feedback resistors $R_3 = R_4 = R_{3.4}$:

$$\frac{1}{Q} = \frac{R_1 2R_{3.4}}{R_{3.4}(R_1 + R_2)} \rightarrow \frac{1}{Q} = \frac{R_1 R_{3.4}}{R_1 + R_2}$$

If $R_{3.4}$ is set to $R_{3.4} = 10\ k\Omega$:

$$\frac{1}{Q} = 2\frac{R_1}{R_1 + R_2} \rightarrow Q = \frac{R_1 + R_2}{2R_1}$$

In terms of the individual R values:

$$R_2 = R_1(2Q - 1)$$

$$R_1 = R_2/(2Q - 1)$$

If we set $Q = 40$, the middle of the specified range:

$$R_2 = R_1(2(40) - 1) \rightarrow R_2 = R_1 79$$

$$R_1 = R_2/(2(40) - 1) \rightarrow R_1 = R_2/79$$

Now these values can be set for the required Q based on available resistor values. If $R_1 = 130\ \Omega$, available in the E24 5% and 1% tolerance preferred value group, then $R_2 = 10270\ \Omega \cong 10\ k\Omega$.

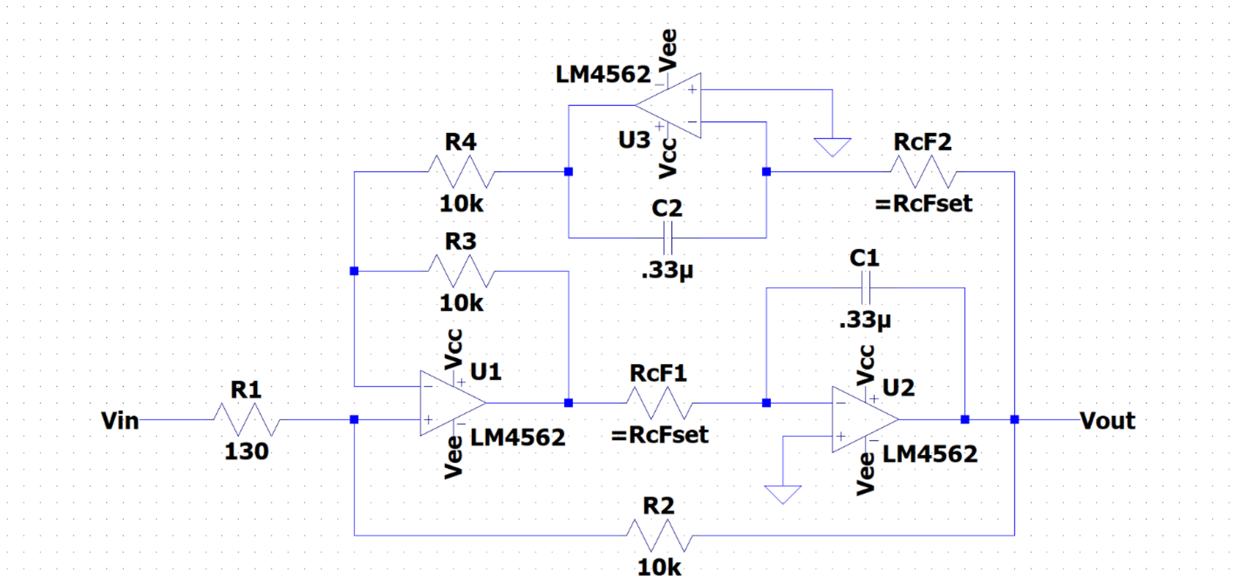

Figure 6: Simplified state-variable filter with values determined to meet design specifications.

The transfer function for the state-variable filter's DC passband gain is given by:

$$A_0 = \frac{\frac{1}{RC} \cdot \frac{R_2(R_3 + R_4)}{R_3(R_1 + R_2)}}{\frac{R_3}{R_4 RC} + \frac{1}{\omega RC}\left(\frac{R_1(R_3 + R_4)}{R_4(R_1 + R_2)}\right) + \frac{1}{\omega RC}} \qquad \text{[3, img. 15]}$$

$$A_0 = \frac{\omega R_2 R_4 (R_3 + R_4)}{R_3(\omega R_3(R_1 + R_2) + R_4(2R_1 + R_2) + R_1 R_3)}$$

$$A_0 = \frac{2\omega R_2 {R_{3.4}}^2}{{R_{3.4}}^2(\omega(R_1 + R_2) + 3R_1 + R_2)}$$

For peak magnitude at ω, multiply by the Q:

$$\frac{V_{out}}{V_{in}}(\omega) = \frac{2\omega Q R_2 {R_{3.4}}^2}{{R_{3.4}}^2(\omega(R_1 + R_2) + 3R_1 + R_2)}$$

Simulation Results:

For the initial simulation, set $R_{cFset} = 482\ \Omega$ for a peak response at:

$$\omega_c = ((482)(0.33u))^{-1} = 6286.94\ rad/s = 1000.60\ Hz$$

Q has been set by R_1 and R_2:

$$Q = \frac{(130) + (10k)}{2(130)} = 38.96$$

And the expected DC passband gain at $\omega = 6286.94\ rad/s$:

$$\frac{V_{out}}{V_{in}}(\omega) = \frac{2(6286.94)(38.96)(10k)(10k)^2}{(10k)^2\Big((6286.94)\big((130) + (10k)\big) + 3(130) + (10k)\Big)} = 76.9075\ V/V$$

For a gain (in dB) of:

$$20\log(76.907) = 37.7194\ dB$$

Several other center frequencies were calculated for comparison against the simulation:

R_{cFset} (Ω)	f_0 (Hz)	Q (f/BW)	V_o/V_i	dB
24.11	20003.6378	38.96	76.9194	37.7207
48.2	10005.9690	38.96	76.9188	37.7206
482	1000.5969	38.96	76.9075	37.7457
4820	100.0597	38.96	76.7948	37.7066
9645.76	49.9999	38.96	76.6697	37.6925

The results of the simulations are presented below, and on the next page.

R_{cFset} (Ω)	f_0 (Hz)	BW (Hz)	Q (f/BW)	V_o/V_i	dB
24.11	19998.6	476.775	41.9456	82.8245	38.3632
48.2	10004.6	248.572	40.2483	79.4660	38.0036
482	1000.46	25.6111	39.0636	77.1408	37.7457
4820	100.046	2.56789	38.9604	76.9401	37.7231
9645.76	50.0035	1.28323	38.9669	76.9322	37.7222

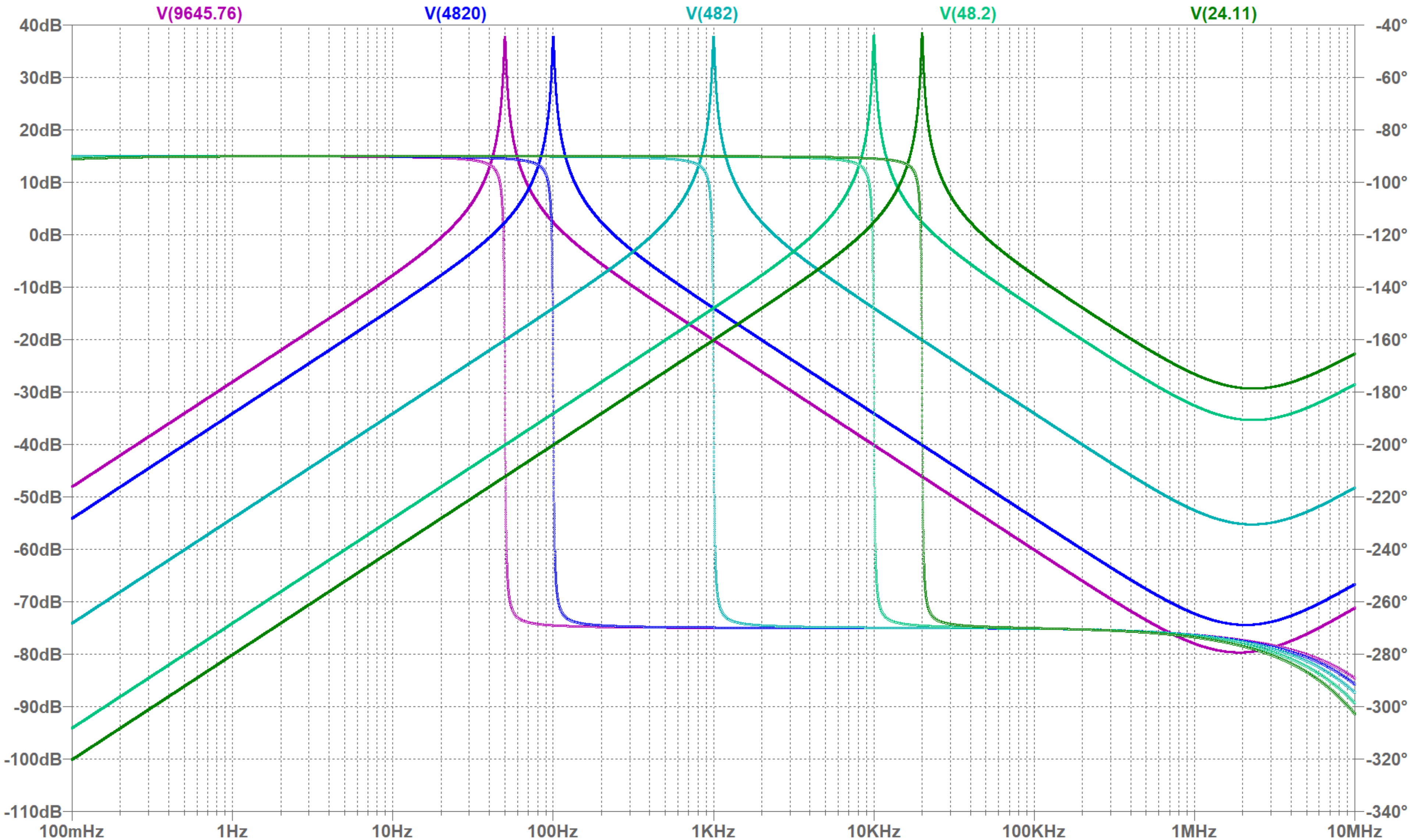

Figure 7: Bode plot of state-variable filter at various settings, showing band-pass high-Q characteristics, with variation in center frequency.

Conclusion:

The simplified state-variable topology was observed to meet the design specifications and could likely exceed them as needed when high quality op-amps are used. The center frequency was able to be varied, and could be controlled using a dual-gang $10k\Omega$ potentiometer. The formula for Q was also shown to be dependent on R_1 and R_2, and so could easily be varied across a range using a single potentiometer as well. The peak response and Q was consistent across the frequency range tested, with minor variance. The design would be easy to build using the op-amps tested, and uses readily available components. Extensibility of the state-variable topology is also quite good, as the integrator stages can be replaced by control voltages to automate or otherwise selectively control the filter's characteristics.

The practical lab portion of this experiment should include building the circuit with measured component values and creating a table of actual response measurements under conditions and settings as near to the simulated cases as possible. Using the actual component values, recalculate the expected results using the formulas in this report, and compare against the measured responses. If possible, connect the filter to an audio source and an output to observe the filters effects. If necessary, create a passive attenuator resistor network to adjust the source to a decibel level appropriate for the expected actual filter response. Likewise, apply an output buffer to the output if needed. Does the filter behave as expected? Make a statement on the performance of the filter and suggest improvements or adjustments. If time permits, consider building the filter circuit using DIP-8 sockets, and switch out the LM4562 for other op-amps. Take brief measurements on the oscilloscope and listen to the audio source again, and make a subjective statement on whether the differences are audible.

References:

[1] S. Franco, Design with Operation Amplifiers and Analog Integrated Circuits, 4th Ed. McGraw Hill, 2015.

[2] A. M. Soliman and A. H. Madian, "MOS-C KHN Filter Using Voltage Op Amp, CFOA, OTRA and DCVC," *Journal of Circuits, Systems, and Computers*, Vol. 18, No. 4 733–769, 2009.

[3] "State Variable Filter," *Electronics Tutorials* [website] Available: Electronics Tutorials, https://www.electronics-tutorials.ws/filter/state-variable-filter.html