

A Single-Supply Op-Amp Circuit Collection

Bruce Carter

Op-Amp Applications, High Performance Linear Products

One of the biggest problems for designers of op-amp circuitry arises when the circuit must be operated from a single supply, rather than ± 15 V. This application note provides working circuit examples.

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1 Introduction

There have been many excellent collections of op-amp circuits in the past, but all of them focus exclusively on split-supply circuits. Many times, the designer who has to operate a circuit from a single supply does not know how to do the conversion.

Single-supply operation requires a little more care than split-supply circuits. The designer should read and understand this introductory material.

1.1 Split Supply vs Single Supply

All op amps have two power pins. In most cases, they are labeled V_{CC+} and V_{CC-} , but sometimes they are labeled V_{CC} and GND. This is an attempt on the part of the data sheet author to categorize the part as a split-supply or single-supply part. However, it does not mean that the op amp has to be operated that way—it may or may not be able to operate from different voltage rails. Consult the data sheet for the op amp, especially the absolute maximum ratings and voltage-swing specifications, before operating at anything other than the recommended power-supply voltage(s).

Most analog designers know how to use op amps with a split power supply. As shown in the left half of Figure 1, a split power supply consists of a positive supply and an equal and opposite negative supply. The most common values are ± 15 V, but ± 12 V and ± 5 V are also used. The input and output voltages are referenced to ground, and swing both positive and negative to a limit of $V_{OM\pm}$, the maximum peak-output voltage swing.

A single-supply circuit (right side of Figure 1) connects the op-amp power pins to a positive voltage and ground. The positive voltage is connected to V_{CC+} , and ground is connected to V_{CC-} or GND. A virtual ground, halfway between the positive supply voltage and ground, is the reference for the input and output voltages. The voltage swings above and below this virtual ground to the limit of $V_{OM\pm}$. Some newer op amps have different high- and low-voltage rails, which are specified in data sheets as V_{OH} and V_{OL} , respectively. It is important to note that there are very few cases when the designer has the liberty to reference the input and output to the virtual ground. In most cases, the input and output will be referenced to system ground, and the designer must use decoupling capacitors to isolate the dc potential of the virtual ground from the input and output (see section 1.3).

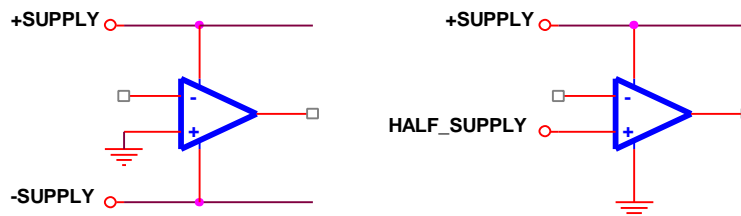


Figure 1. Split Supply (L) vs Single Supply (R) Circuits

A common value for single supplies is 5 V, but voltage rails are getting lower, with 3 V and even lower voltages becoming common. Because of this, single-supply op amps are often rail-to-rail devices, which avoids losing dynamic range. Rail-to-rail may or may not apply to both the input and output stages. Be aware that even though a device might be specified as rail-to-rail, some

specifications can degrade close to the rails. Be sure to consult the data sheet for complete specifications on both the inputs and outputs. It is the designer's obligation to ensure that the voltage rails of the op amp do not degrade the system specifications.

1.2 Virtual Ground

Single-supply operation requires the generation of a virtual ground, usually at a voltage equal to $V_{CC}/2$. The circuit in Figure 2 can be used to generate $V_{CC}/2$, but its performance deteriorates at low frequencies.

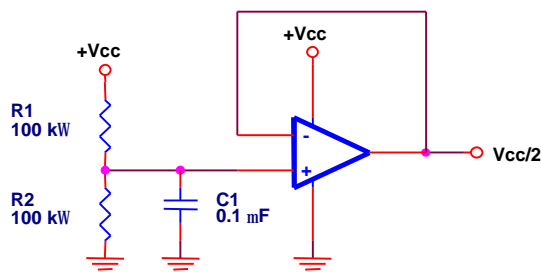


Figure 2. Single-Supply Operation at $V_{CC}/2$

R_1 and R_2 are equal values, selected with power consumption vs allowable noise in mind. Capacitor C_1 forms a low-pass filter to eliminate conducted noise on the voltage rail. Some applications can omit the buffer op amp.

In what follows, there are a few circuits in which a virtual ground has to be introduced with two resistors within the circuit because one virtual ground is not suitable. In these instances, the resistors should be 100 kW or greater; when such a case arises, values are indicated on the schematic.

1.3 AC-Coupling

A virtual ground is at a dc level above system ground; in effect, a small, local-ground system has been created within the op-amp stage. However, there is a potential problem: the input source and output load are probably referenced to system ground, and if the op-amp stage is connected to a source that is referenced to ground instead of virtual ground, there will be an unacceptable dc offset. If this happens, the op amp becomes unable to operate on the input signal, because it must then process signals at and below its input and output rails.

The solution is to ac-couple the signals to and from the op-amp stage. In this way, the input and output devices can be referenced to ground, and the op-amp circuitry can be referenced to a virtual ground.

When more than one op-amp stage is used, interstage decoupling capacitors might become unnecessary if all of the following conditions are met:

- The first stage is referenced to virtual ground.
- The second stage is referenced to virtual ground.

- There is no gain in either stage. Any dc offset in either stage is multiplied by the gain in both, and probably takes the circuit out of its normal operating range.

If there is any doubt, assemble a prototype including ac-coupling capacitors, then remove them one at a time. Unless the input or output are referenced to virtual ground, there must be an input-decoupling capacitor to decouple the source and an output-decoupling capacitor to decouple the load. A good troubleshooting technique for ac circuits is to terminate the input and output, then check the dc voltage at all op-amp inverting and noninverting inputs and at the op-amp outputs. All dc voltages should be very close to the virtual-ground value. If they are not, decoupling capacitors are mandatory in the previous stage (or something is wrong with the circuit).

1.4 Combining Op-Amp Stages

Combining op-amp stages to save money and board space is possible in some cases, but it often leads to unavoidable interactions between filter response characteristics, offset voltages, noise, and other circuit characteristics. The designer should always begin by prototyping separate gain, offset, and filter stages, then combine them if possible after each individual circuit function has been verified. Unless otherwise specified, filter circuits included in this document are unity gain.

1.5 Selecting Resistor and Capacitor Values

The designer who is new to analog design often wonders how to select component values. Should resistors be in the 1- Ω decade or the 1-M Ω decade? Resistor values in the 1-k Ω to 100-k Ω range are good for general-purpose applications. High-speed applications usually use resistors in the 100- Ω to 1-k Ω decade, and they consume more power. Portable applications usually use resistors in the 1-M Ω or even 10-M Ω decade, and they are more prone to noise. Basic formulas for selecting resistor and capacitor values for tuned circuits are given in the various figures. For filter applications, resistors should be chosen from 1% E-96 values (see Appendix A). Once the resistor decade range has been selected, choose standard E-12 value capacitors. Some tuned circuits may require E-24 values, but they should be avoided where possible. Capacitors with only 5% tolerance should be avoided in critical tuned circuits—use 1% instead.

2 Basic Circuits

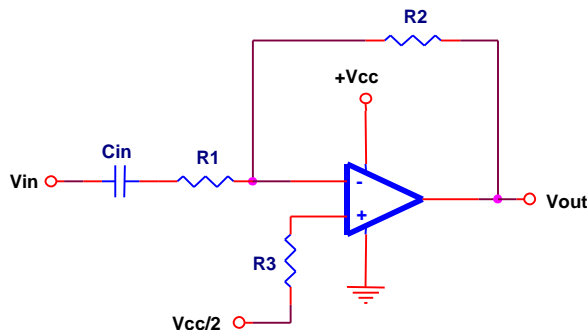
2.1 Gain

Gain stages come in two basic varieties: inverting and noninverting. The ac-coupled version is shown in Figure 3. For ac circuits, inversion means an ac-phase shift of 180°. These circuits work by taking advantage of the coupling capacitor, C_{IN}, to prevent the circuit from having dc gain. They have ac gain only. If C_{IN} is omitted in a dc system, dc gain must be taken into account.

It is very important not to violate the bandwidth limit of the op amp at the highest frequency seen by the circuit. Practical circuits can include gains of 100 (40 dB), but higher gains could cause the circuit to oscillate unless special care is taken during PC board layout. It is better to cascade two or more equal-gain stages than to attempt high gain in a single stage.

INVERTING

Gain = $-R_2/R_1$
 $R_3 = R_1 || R_2$
 for minimum error due
 to input bias current

**NONINVERTING**

Gain = $1 + R_2/R_1$
 Input Impedance = $R_1 || R_2$
 for minimum error due
 to input bias current

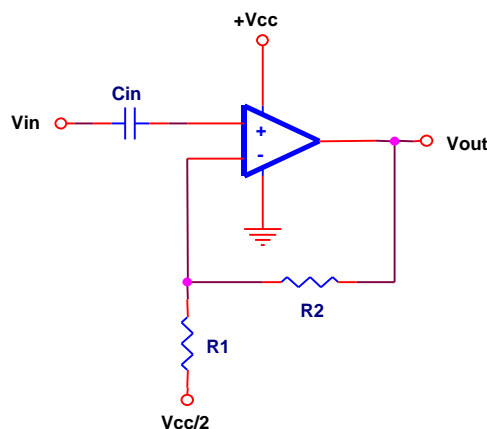


Figure 3. AC-Coupled Gain Stages

2.2 Attenuation

The traditional way of doing inverting attenuation with an op-amp circuit is shown in Figure 4, in

INVERTING

Gain = $-R_2/R_1$
 $R_3 = R_1 || R_2$
 for minimum error due
 to input bias current

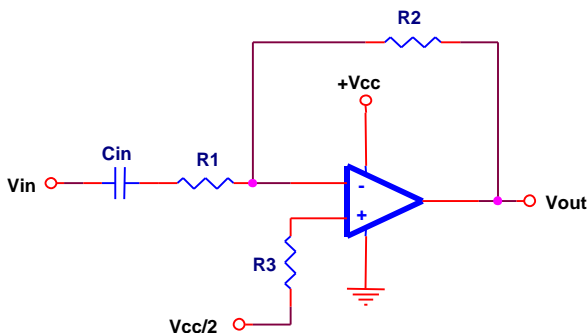


Figure 4. Traditional Inverting Attenuation With an Op Amp

which $R_2 < R_1$. This method is not recommended, because many op amps are unstable at gains of less than unity. The correct way to construct an attenuation circuit¹ is shown in Figure 5.

¹ This circuit is taken from the design notes of William Ezell

INVERTING

Component values
normalized to unity

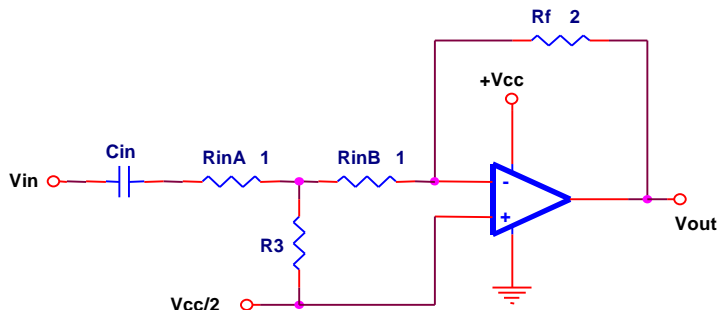


Figure 5. Inverting Attenuation Circuit

A set of normalized values of the resistor R3 for various levels of attenuation is shown in Table 1. For nontabulated attenuation values, the resistance is:

$$R_3 = \frac{V_O/V_{IN}}{2 - 2(V_O/V_{IN})}$$

To work with normalized values, do the following:

- Select a base-value of resistance, usually between 1 kΩ and 100 kΩ for Rf and Rin.
- Divide Rin in two for RinA and RinB.
- Multiply the base value for Rf and Rin by 1 or 2, as shown in Figure 5.
- Look up the normalization factor for R3 in the table below, and multiply it by the base-value of resistance.

For example, if Rf is 20 kΩ, RinA and RinB are each 10 kΩ, and a 3-dB attenuator would use a 12.1-kΩ resistor.

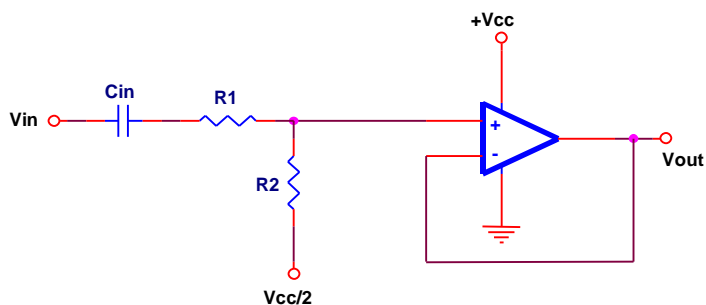
Table 1. Normalization Factors

DB Pad	Vout/Vin	R3
0	1.0000	∞
0.5	0.9441	8.4383
1	0.8913	4.0977
2	0.7943	0.9311
2	0.7079	1.2120
3.01	0.7071	1.2071
3.52	0.6667	1.000
4	0.6310	0.8549
5	0.5623	0.6424
6	0.5012	0.5024
6.02	0.5000	0.5000
7	0.4467	0.4036
8	0.3981	0.3307
9	0.3548	0.2750
9.54	0.3333	0.2500
10	0.3162	0.2312
12	0.2512	0.1677
12.04	0.2500	0.1667
13.98	0.2000	0.1250
15	0.1778	0.1081
15.56	0.1667	0.1000
16.90	0.1429	0.08333
18	0.1259	0.07201
18.06	0.1250	0.07143
19.08	0.1111	0.06250
20	0.1000	0.05556
25	0.0562	0.02979
30	0.0316	0.01633
40	0.0100	0.005051
50	0.0032	0.001586
60	0.0010	0.0005005

Noninverting attenuation can be performed with a voltage divider and a noninverting buffer as shown in Figure 6.

NONINVERTING

Component values
normalized to unity

**Figure 6. Noninverting Attenuation**

2.3 Summing

An inverting summing circuit (Figure 7) is the basis of an audio mixer. A single-supply voltage is seldom used for real audio mixers. Designers will often push an op amp up to, and sometimes beyond, its recommended voltage rails to increase dynamic range.

Noninverting summing circuits are possible, but not recommended. The source impedance becomes part of the gain calculation.

INVERTING

$$V_{out} = -R_2(V_{in1}/R_1 + V_{in2}/R_2 + V_{in3}/R_3)$$

$$= R_1A || R_1B || R_1C || R_2$$

for minimum error due to input bias current

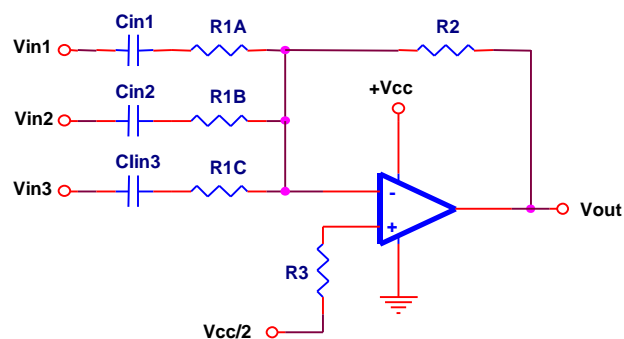


Figure 7. Inverting Summing Circuit

2.4 Difference Amplifier

Just as there are summing circuits, there are also subtracting circuits (Figure 8). A common application is to eliminate the vocal track (recorded at equal levels in both channels) from stereo recordings.

For $R_1 = R_3$ and $R_2 = R_4$:

$$V_{out} = (R_2/R_1)(V_{in2} - V_{in1})$$

$$R_1 || R_2 = R_3 || R_4$$

for minimum error due to input bias current

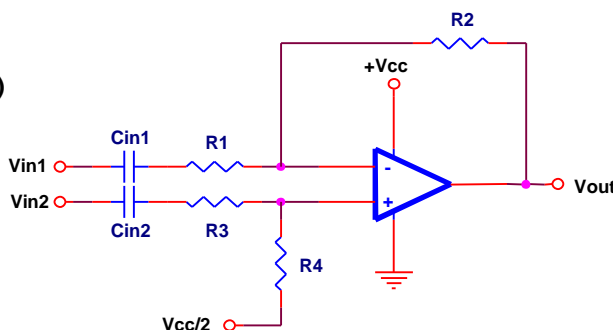


Figure 8. Subtracting Circuit

2.5 Simulated Inductor

The circuit in Figure 9 reverses the operation of a capacitor, thus making a simulated inductor. An inductor resists any change in its current, so when a dc voltage is applied to an inductance, the current rises *slowly*, and the voltage falls as the external resistance becomes more significant.

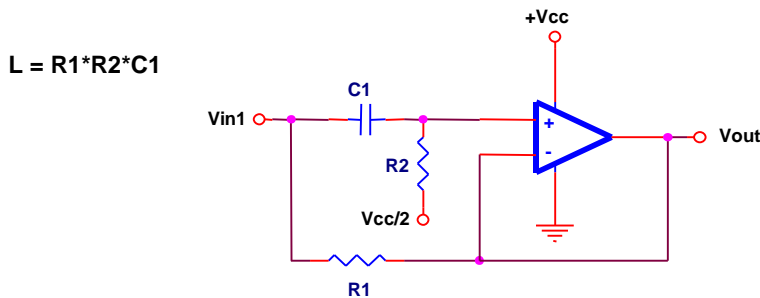


Figure 9. Simulated Inductor Circuit

An inductor passes low frequencies more readily than high frequencies, the opposite of a capacitor. An ideal inductor has zero resistance. It passes dc without limitation, but it has infinite impedance at infinite frequency.

If a dc voltage is suddenly applied to the inverting input through resistor R1, the op amp ignores the sudden load because the change is also coupled directly to the noninverting input via C1. The op amp represents high impedance, just as an inductor does.

As C1 charges through R2, the voltage across R2 falls, so the op-amp draws current from the input through R1. This continues as the capacitor charges, and eventually the op-amp has an input and output close to virtual ground ($V_{cc}/2$).

When C1 is fully charged, resistor R1 limits the current flow, and this appears as a series resistance within the simulated inductor. This series resistance limits the Q of the inductor. Real inductors generally have much less resistance than the simulated variety.

There are some limitations of a simulated inductor:

- One end of the inductor is connected to virtual ground.
- The simulated inductor cannot be made with high Q, due to the series resistor R1.
- It does not have the same energy storage as a real inductor. The collapse of the magnetic field in a real inductor causes large voltage spikes of opposite polarity. The simulated inductor is limited to the voltage swing of the op amp, so the flyback pulse is limited to the voltage swing.

2.6 Instrumentation Amplifiers

Instrumentation amplifiers are used whenever dc gain is needed on a low-level signal that would be loaded by conventional differential-amplifier topologies. Instrumentation amplifiers take advantage of the high input impedance of noninverting op-amp inputs.

The basic instrumentation amplifier topology is shown in Figure 10.

ASSUMES V_{in-} AND V_{in+}
REFERENCED TO $V_{cc}/2$

$R1 = R3$ (matched)
 $R2 = R4$ (matched)
 $R5 = R6$
Gain = $R2/R1 (1 + 2R5/R7)$

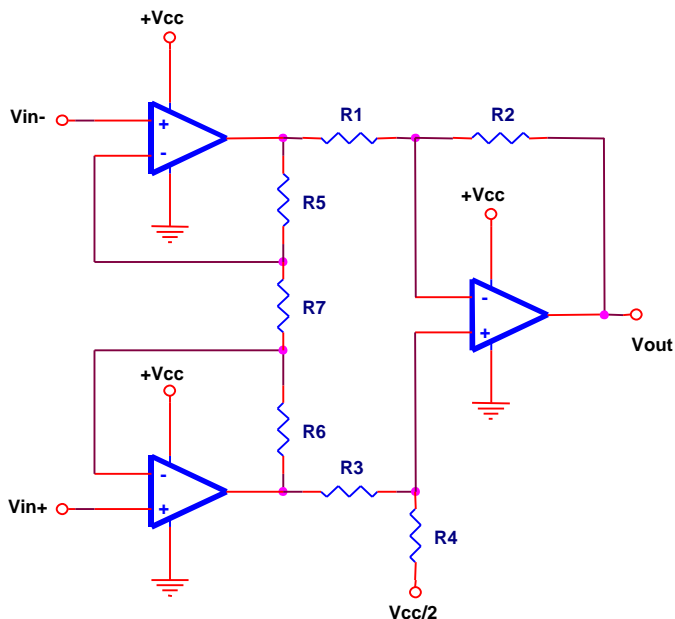


Figure 10. Basic Instrumentation-Amplifier Circuit

This circuit, and the other instrumentation amplifier topologies presented here, assume that the inputs are already referenced to half-supply. This is the case with strain gauges that are operated from V_{cc} . The basic disadvantage of this circuit is that it requires matched resistors; otherwise, it would suffer from poor CMRR (see for example, Op Amps for Everyone^[3]).

The circuit in Figure 10 can be simplified by eliminating three resistors, as shown in Figure 11.

ASSUMES V_{in-} AND V_{in+}
REFERENCED TO $V_{cc}/2$

$R1 = R3$ (matched)
 $R2 = R4$ (matched)
Gain = $R2/R1$

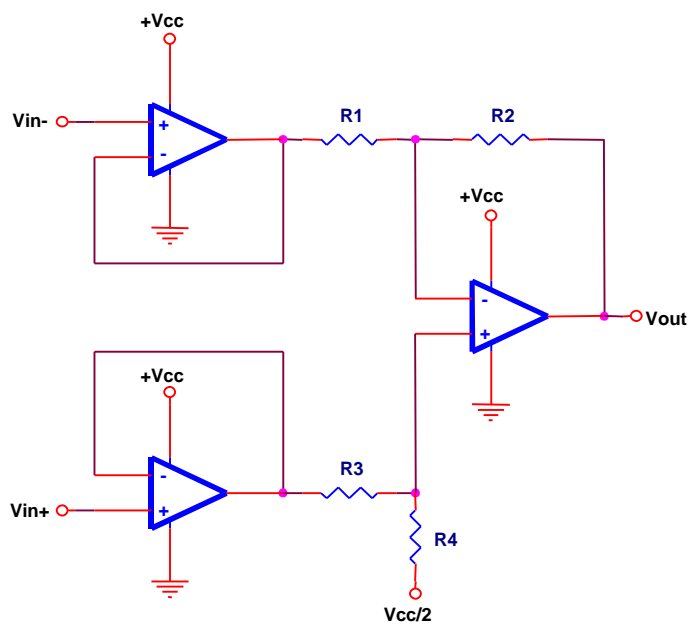


Figure 11. Modified Instrumentation-Amplifier Circuit

Here, the gain is easier to calculate, but a disadvantage is that now two resistors must be changed instead of one, and they must be matched resistors. Another disadvantage is that the first stage(s) cannot be used for gain.

An instrumentation amplifier can also be made from two op amps; this is shown in Figure 12.

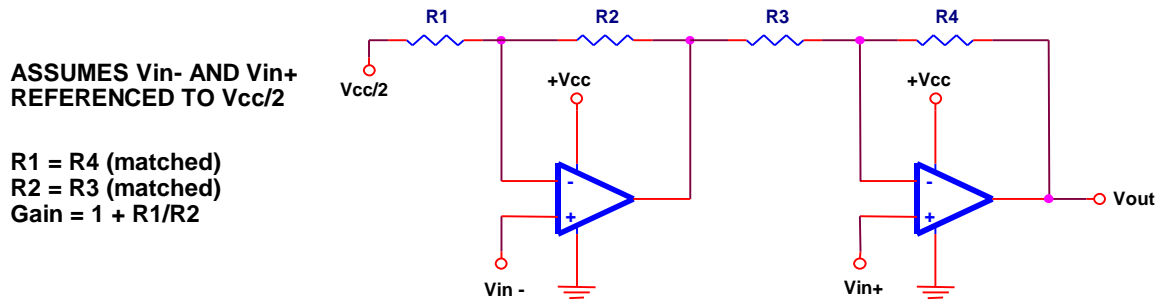


Figure 12. Instrumentation Circuit With Only Two Op Amps

However, this topology is not recommended because the first op amp is operated at less than unity gain, so it may be unstable. Furthermore, the signal from V_{in-} has more propagation delay than V_{in+} .

3 Filter Circuits

This section is devoted to op-amp active filters. In many cases, it is necessary to block dc voltage from the virtual ground of the op-amp stage by adding a capacitor to the input of the circuit. This capacitor forms a high-pass filter with the input so, in a sense, all these circuits have a high-pass characteristic. The designer must insure that the input capacitor is at least 100 times the value of the other capacitors in the circuit, so that the high-pass characteristic does not come into play at the frequencies of interest in the circuit. For filter circuits with gain, 1000 times might be better. If the input voltage already contains a $V_{cc}/2$ offset, the capacitor can be omitted.

These circuits will have a half-supply dc offset at their output. If the circuit is the last stage in the system, an output-coupling capacitor may also be required.

There are trade-offs involved in filter design. The most desirable situation is to implement a filter with a single op amp. Ideally, the filter would be simple to implement, and the designer would have complete control over:

- The filter corner / center frequency
- The gain of the filter circuit
- The Q of band-pass and notch filters, or style of low-pass and high-pass filter (Butterworth, Chebyshev, or Bessel).

Unfortunately, such is not the case—complete control over the filter is seldom possible with a single op amp. If control is possible, it frequently involves complex interactions between passive components, and this means complex mathematical calculations that intimidate many designers. More control usually means more op amps, which may be acceptable in designs that will not be produced in large volumes, or that may be subject to several changes before the design is finalized. If the designer needs to implement a filter with as few components as possible, there will be no choice but to resort to traditional filter-design techniques and perform the necessary calculations.

3.1 Single Pole Circuits

Single-pole circuits are the simplest filter circuits. They have a roll off of 20 dB per decade.

3.1.1 Low Pass Filter Circuits

Typical low-pass filter circuits are shown in Figure 13.

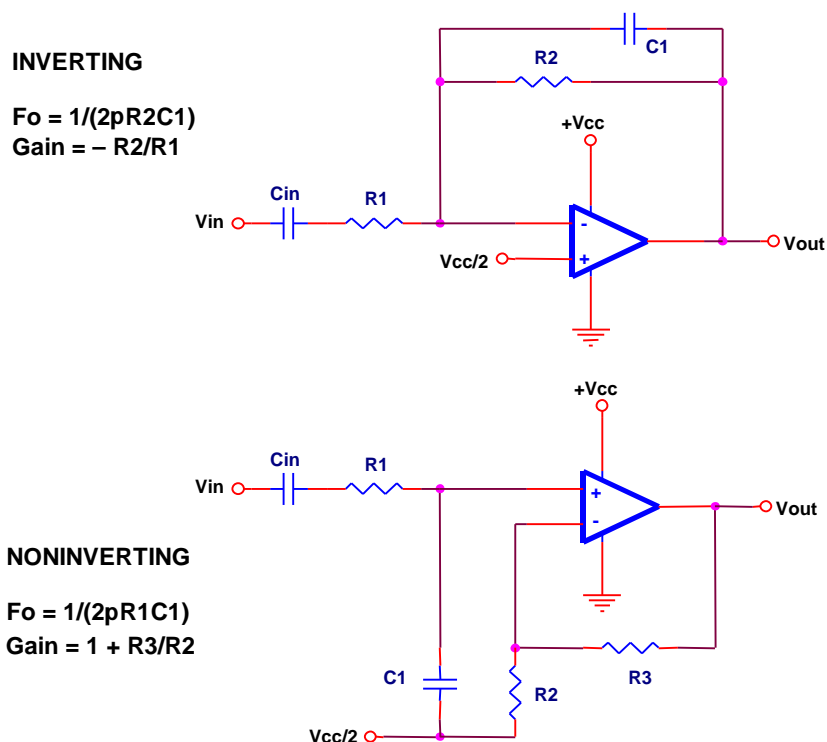


Figure 13. Low-Pass Filter Circuits

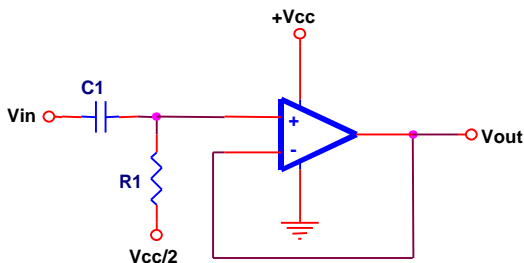
3.1.2 High Pass Filter Circuits

Typical high-pass filter circuits are shown in Figure 14.

NONINVERTING

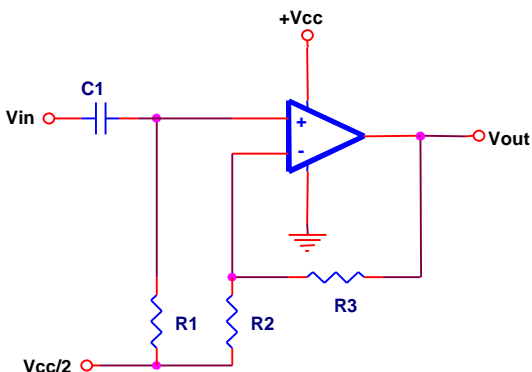
$$\text{Gain} = 1$$

$$F_o = 1/(2\pi R_1 C_1)$$

**NONINVERTING**

$$F_o = 1/(2\pi R_1 C_1)$$

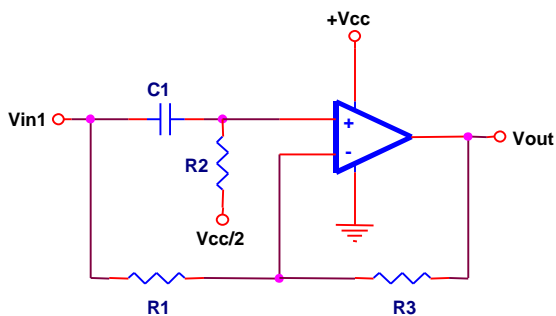
$$\text{Gain} = 1 + R_3/R_2$$

**Figure 14. High-Pass Filter Circuits****3.1.3 All-Pass Filter**

The all-pass filter passes all frequencies at the same gain. It is used to change the phase of the signal, and it can also be used as a phase-correction circuit. The circuit shown in Figure 15 has a 90° phase shift at $F(90)$. At dc, the phase shift is 0°, and at high frequencies it is 180°.

$$R_1 = R_2 = R_3 = R$$

$$F(90) = 1/(2\pi R C_1)$$

**Figure 15. All-Pass Filter Circuit**

3.2 Double-Pole Circuits

Double-pole op-amp circuit topologies are sometimes named after their inventor. Several implementations or topologies exist. Some double-pole circuit topologies are available in a low-pass, high-pass, band-pass, and notch configuration. Others are not. Not all topologies and implementations are given here: only the ones that are easy to implement and tune.

Double-pole or second-order filters have a 40-dB-per-decade roll-off.

Commonly the same component(s) adjust the Q for the band-pass and notch versions of the topology, and they change the filter from Butterworth to Chebyshev, etc. for low-pass and high-pass versions of the topology. Be aware that the corner frequency calculation is only valid for the Butterworth versions of the topologies. Chebyshev and Bessel modify it slightly.

When band-pass and notch filter circuits are shown, they are high-Q (single frequency) types. To implement a wider band-pass or notch (band-reject) filter, cascade low-pass and high-pass stages. The pass characteristics should overlap for a band-pass and *not* overlap for a band-reject filter.

Inverse Chebyshev and Elliptic filters are not shown. These are beyond the scope of a circuit collection note.

Not all filter topologies produce ideal results—the final attenuation in the rejection band, for example, is greater in the multiple-feedback filter configuration than it is in the Sallen-Key filter. These fine points are beyond the scope of an op-amp circuit collection. Consult a textbook on filter design for the merits and drawbacks of each of these topologies. Unless the application is particularly critical, all the circuits shown here should produce acceptable results.

3.2.1 Sallen-Key

The Sallen-Key topology is one of the most widely-known and popular second-order topologies. It is low cost, requiring only a single op amp and four passive components to accomplish the tuning. Tuning is easy, but changing the style of filter from Butterworth to Chebyshev is not. The designer is encouraged to read references [1] and [2] for a detailed description of this topology. The circuits shown are unity gain—changing the gain of a Sallen-Key circuit also changes the filter tuning and the style. It is easiest to implement a Sallen-Key filter as a unity gain Butterworth.

LOW PASS

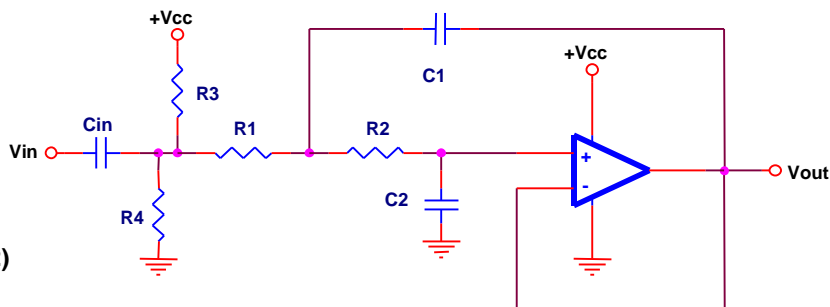
Unity Gain
Butterworth

$R_3 = R_4$ (HIGH)

$R_1 = R_2$

$C_1 = 2C_2$

$F_o = \frac{1}{\sqrt{2}} / (4pR_1C_2)$

**HIGH PASS**

Unity Gain
Butterworth

$C_1 = C_2$

$R_1 = R$

$R = 2R_1$

$F_o = \frac{1}{\sqrt{2}} / (4pR_1C_1)$

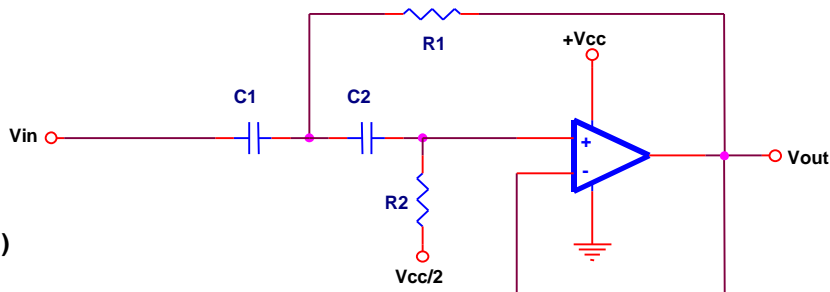


Figure 16. Sallen-Key Low- and High-Pass Filter Topologies

3.2.2 Multiple Feedback (MFB)

MFB topology is very versatile, low cost, and easy to implement. Unfortunately, calculations are somewhat complex, and certainly beyond the scope of this circuit collection. The designer is encouraged to read reference [1] for a detailed description of the MFB topology. If all that is needed is a unity gain Butterworth, then these circuits will provide a close approximation.

LOW PASS

Unity Gain Butterworth

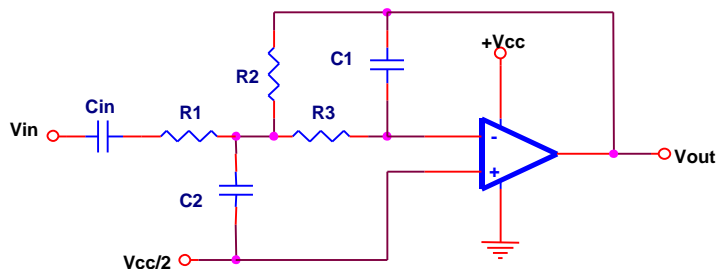
$$F_o = 1/(2\pi RC)$$

$$R1 = R2 = R/\sqrt{2}$$

$$R3 = R/(2\sqrt{2})$$

$$C1 = C$$

$$C2 = 4C$$



HIGH PASS

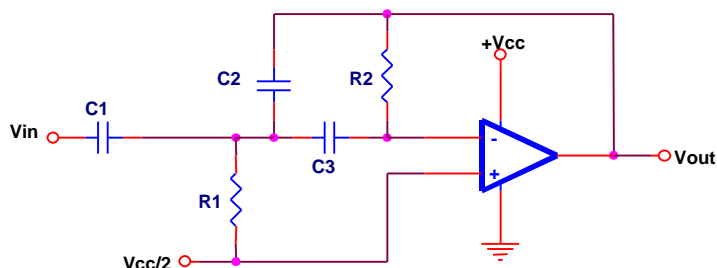
Unity Gain Butterworth

$$F_o = 1/(2\pi RC)$$

$$R1 = 0.47R$$

$$R2 = 2.1R$$

$$C1 = C2 = C3 = C$$



BAND PASS

Gain = 2.3 dB

$$F_o = 1/(2.32\pi RC)$$

$$R1 = 10R$$

$$R2 = 0.001R$$

$$R3 = 100R$$

$$C1 = 10C$$

$$C2 = C$$

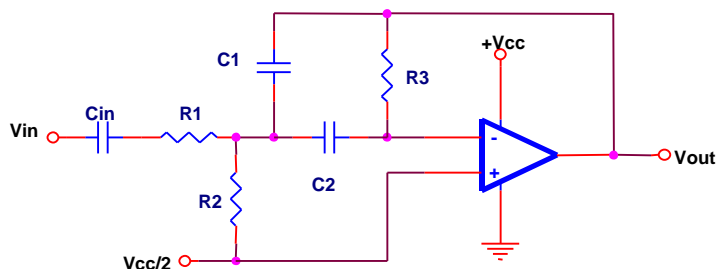


Figure 17. Multiple-Feedback Topologies

3.2.3 Twin T

The twin-T topology uses either one or two op amps. It is based on a passive (RC) topology that uses three resistors and three capacitors. Matching these six passive components is critical; fortunately, it is also easy. The entire network can be constructed from a single value of resistance and a single value of capacitance, running them in parallel to create R3 and C3 in the twin-T schematics shown in Figure. Components from the same batch are likely to have very similar characteristics.

3.2.3.1 Single Op-Amp Implementations

BAND PASS

$$R1 = R2 = R$$

$$C1 = C2 = C$$

$$R3 = R/2$$

$$C3 = 2C$$

$$F_o = 1/(2\pi RC)$$

Gain controlled by $R4$ and $R5$

$$R4 > 100 * R5$$

Q hard to control; need mismatched Resistors; also affects gain

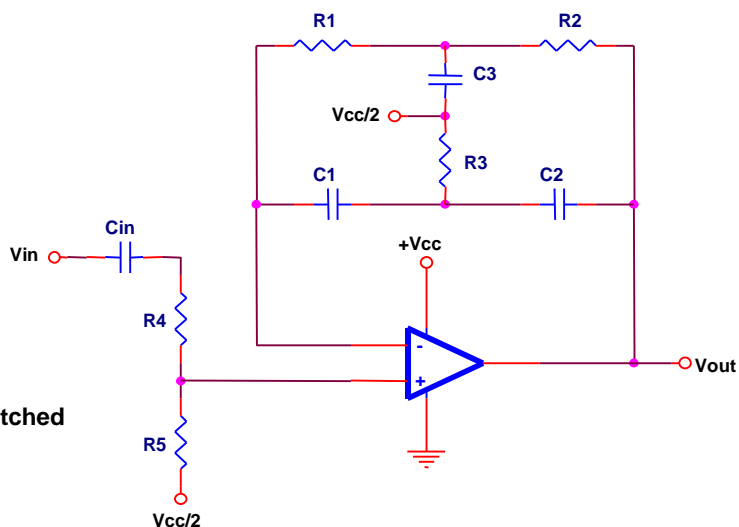


Figure 18. Single Op-Amp Twin-T Filter in Band-Pass Configuration

The bandpass circuit will oscillate if the components are matched too closely. It is best to de-tune it slightly, by selecting the resistor to virtual ground to be one E-96 1% resistor value off, for instance.

NOTCH

$$C1 = C2 = C$$

$$C3 = 2C$$

$$R1 = R2 = R$$

$$R3 = R/2$$

$$F_o = 1/(2\pi RC)$$

$$R4 = R5: \text{HIGH}$$

The only control over Q is by mismatching $R3$

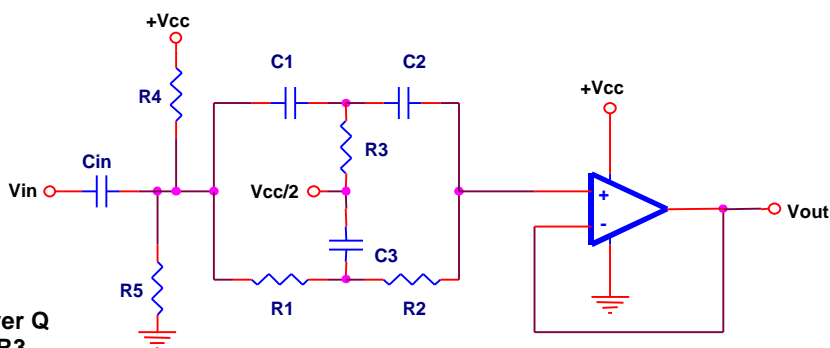


Figure 19. Single Op-Amp Twin-T Filter in Notch Configuration

3.2.3.2 Dual-Op-Amp Implementations

Typical dual op-amp implementations are shown in Figures 20 to 22

LOW PASS

$$R1 = R2 = R$$

$$C1 = C2 = C$$

$$R3 = R/2$$

$$C3 = 2C$$

$$F_o = 1/(2\pi RC)$$

Unity Gain

$$R4 < R5/2 \text{ Chebyshev}$$

$$R4 = R5/2 \text{ Butterworth}$$

$$R4 > R5/2 \text{ Bessel}$$

$$R6 = R7: \text{HIGH}$$

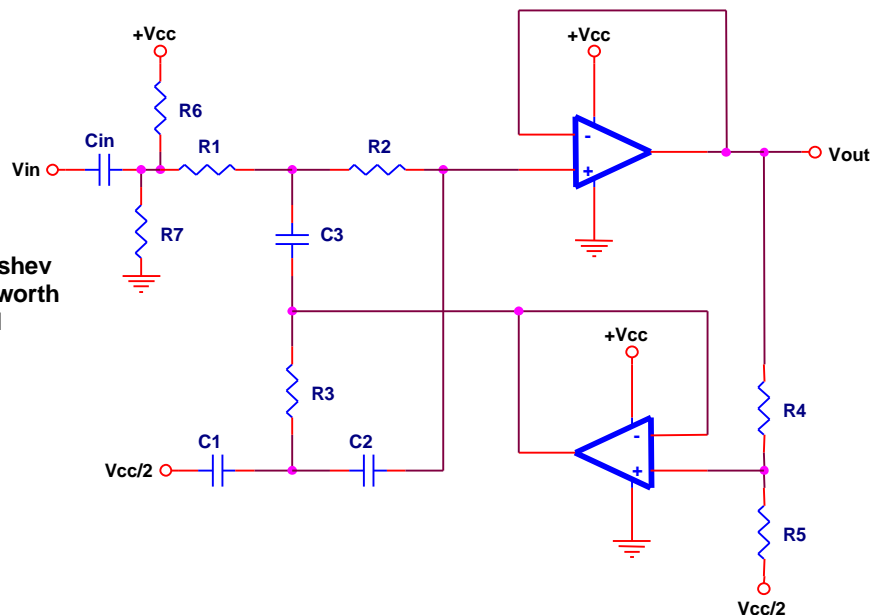


Figure 20. Dual-Op-Amp Twin-T Low-Pass Filter

HIGH PASS

$$R1 = R2 = R$$

$$C1 = C2 = C$$

$$R3 = R/2$$

$$C3 = 2C$$

$$F_o = 1/(2\pi RC)$$

Unity Gain

$$R4 < R5/2 \text{ Chebyshev}$$

$$R4 = R5/2 \text{ Butterworth}$$

$$R4 > R5/2 \text{ Bessel}$$

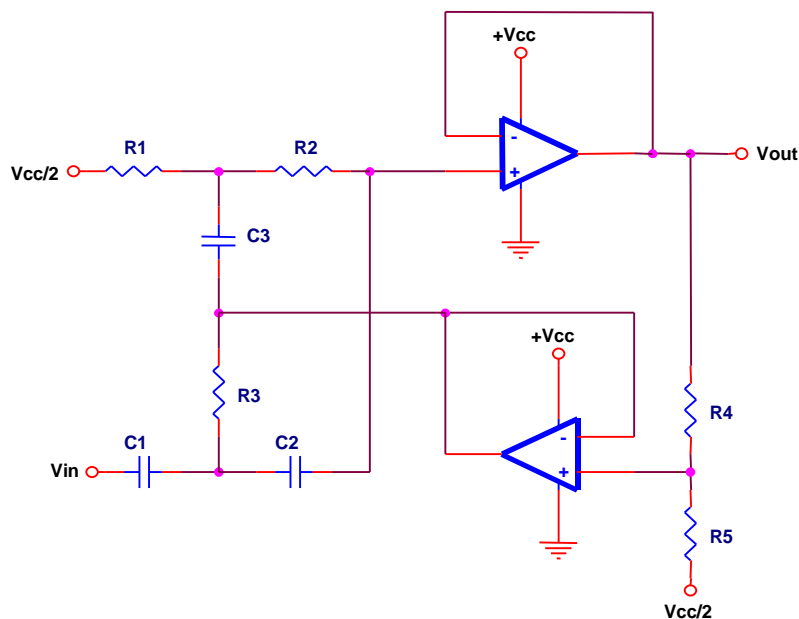


Figure 21. Dual-Op-Amp Twin-T High-Pass Filter

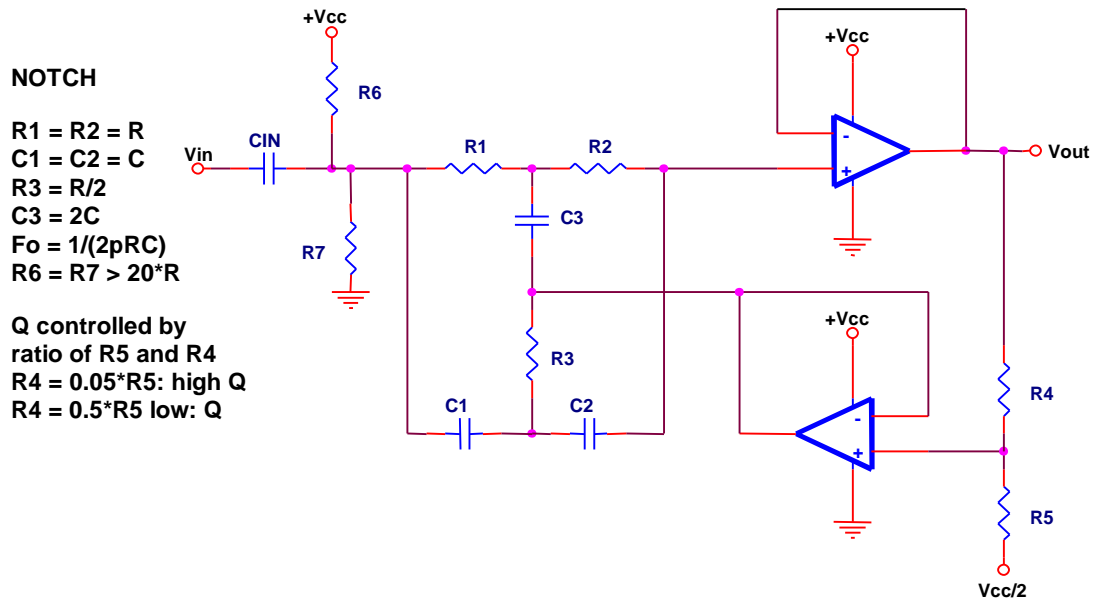


Figure 22. Dual-Op-Amp Twin-T Notch Filter

3.2.4 Fliege

Fliege is a two-op-amp topology (Figures 23–26), and therefore more expensive than one-op-amp topologies. There is good control over the tuning and the Q and style of filter. The gain is fixed at two for low-pass, high-pass, and band-pass filters, and unity for notch.

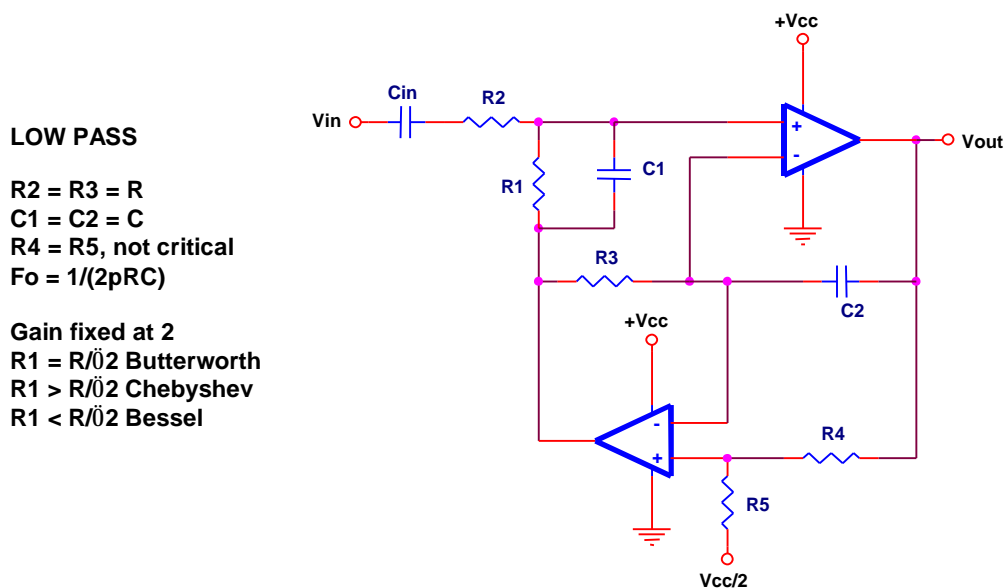


Figure 23. Low-Pass Fliege Filter

HIGH PASS

$R2 = R3 = R$
 $C1 = C2 = C$
 $R4 = R5$, not critical
 $F_o = 1/(2\pi RC)$
 Gain fixed at 2
 $R1 = R/\sqrt{2}$ Butterworth
 $R1 > R/\sqrt{2}$ Chebyshev
 $R1 < R/\sqrt{2}$ Bessel

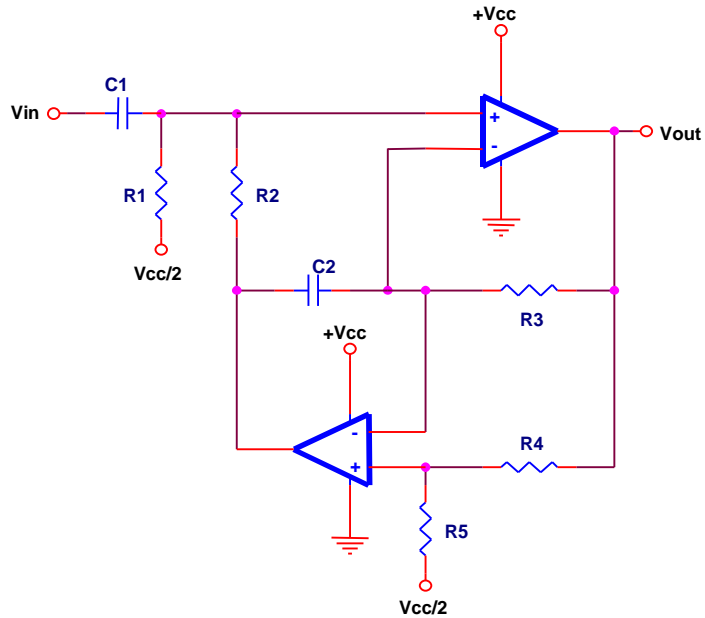


Figure 24. High-Pass Fliege Filter

BAND PASS

Gain fixed at 2
 $R1$ controls Q
 low $R1 \Rightarrow$ low Q
 high $R1 \Rightarrow$ high Q
 $R1$ should be $> R/5$
 $R2 = R3 = R$
 $C1 = C2 = C$
 $R4 = R5$, not critical
 $F_o = 1/(2\pi RC)$

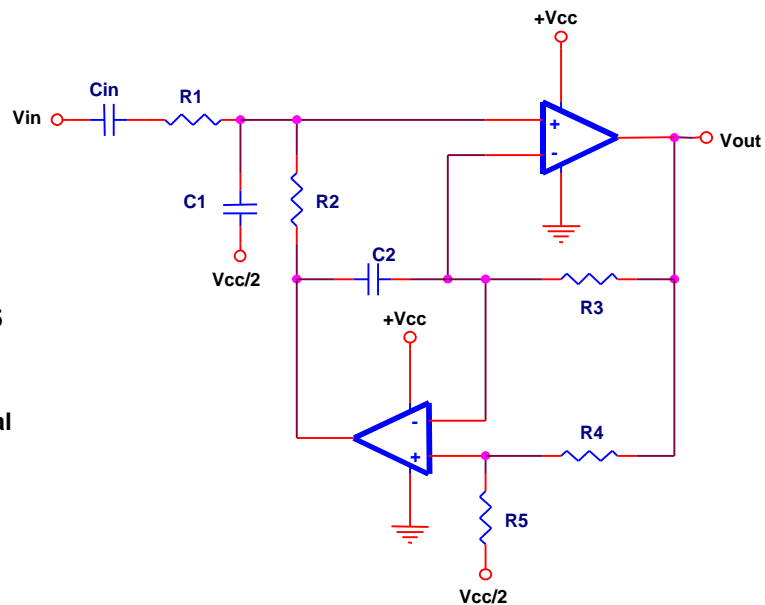


Figure 25. Band-Pass Fliege Filter

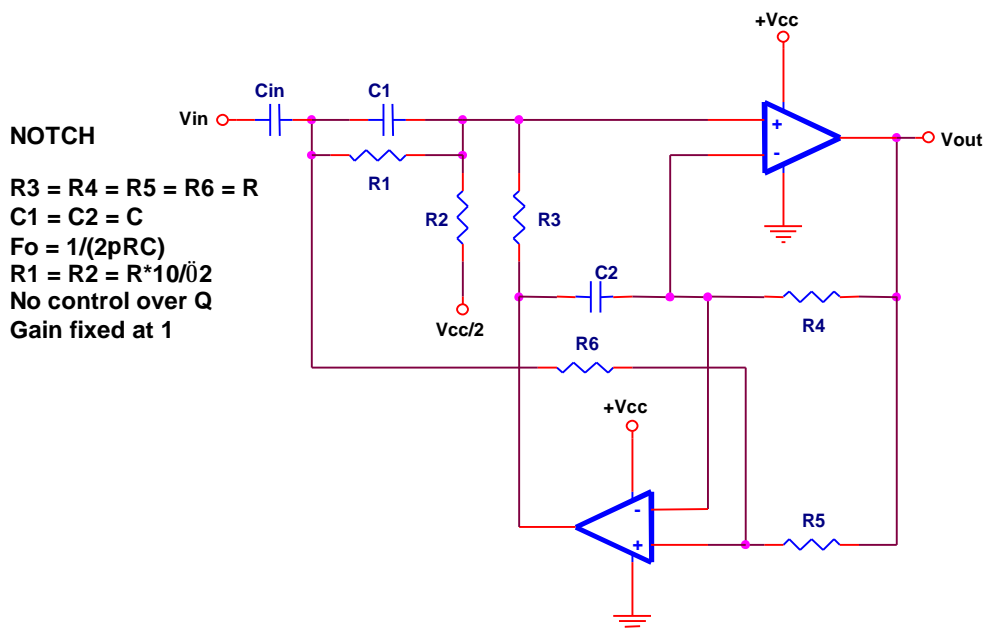


Figure 26. Notch Fliege Filter

3.2.5 Akerberg-Mossberg Filter

This is the easiest of the three-op-amp topologies to use (Figures 27–30). It is easy to change the gain, style of low-pass and high-pass filter, and the Q of band-pass and notch filters. The notch filter performance is not as good as that of the twin T notch, but it is less critical.

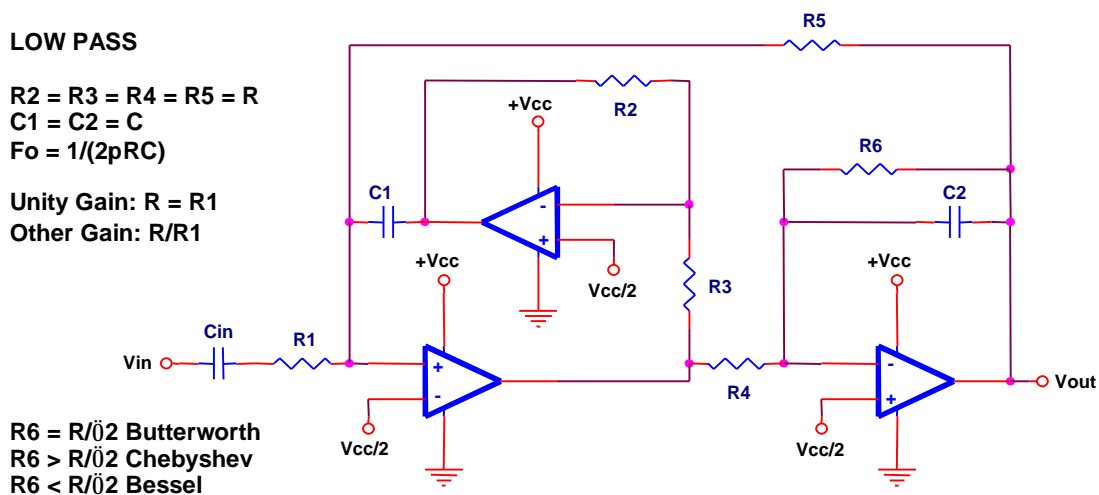


Figure 27. Akerberg-Mossberg Low-Pass Filter

HIGH PASS

$R2=R3=R4=R5=R$
 $C2=C3=C$
 $F_o=1/(2\pi RC)$

$R6 = R/\sqrt{2}$ Butterworth
 $R6 > R/\sqrt{2}$ Chebyshev
 $R6 < R/\sqrt{2}$ Bessel

Unity Gain:
 $C1=C, R1=R$
 Other Gain:
 $R1/R$ AND $C1/C$

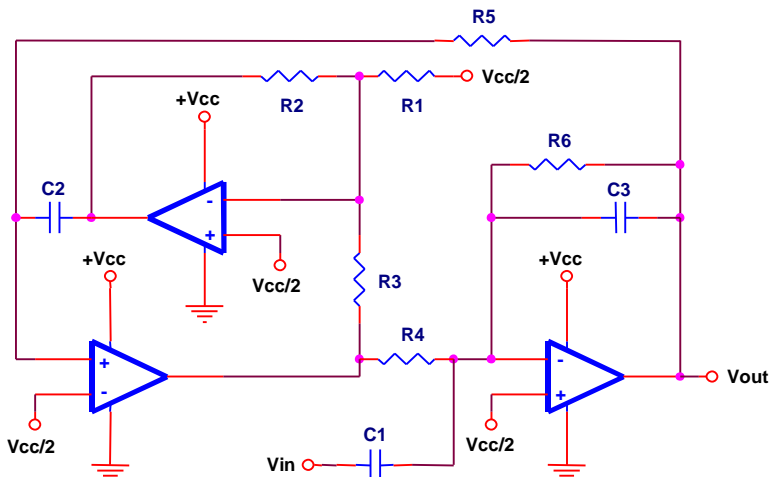


Figure 28. Akerberg-Mossberg High-Pass Filter

BAND PASS

$R2 = R3 = R4 = R5 = R$
 $C1 = C2 = C$
 $F_o = 1/(2\pi RC)$

Unity Gain:
 $R1 = R6$
 Other Gain:
 $-R6/R1$

$R1, R6$ also control Q
 low values, low Q
 high values, high Q

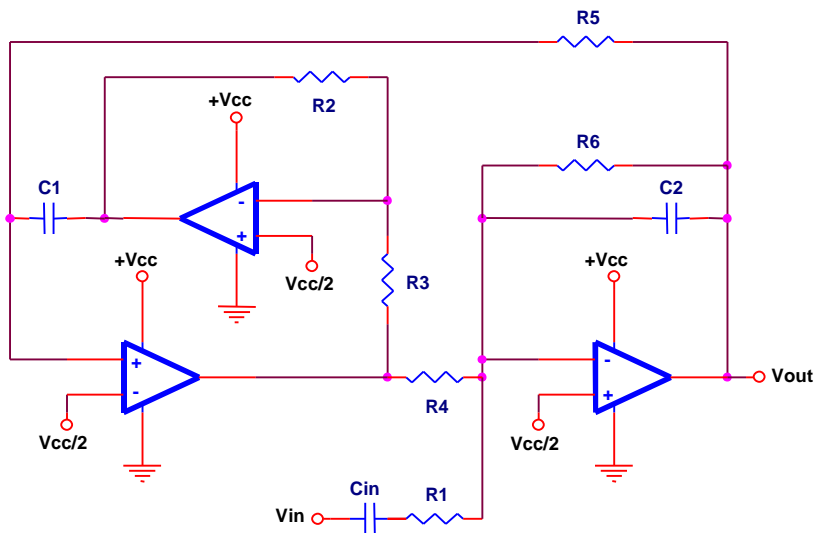


Figure 29. Akerberg-Mossberg Band-Pass Filter

NOTCH

$R1=R2=R3=R4=R5=R6=R$
 $C1 = C2 = C3 = C$
 $F_o = 1/(2RC)$

$R/2 < R7 < 2 \times R$
 $R7$ controls Q
 low value, low Q
 high value, high Q

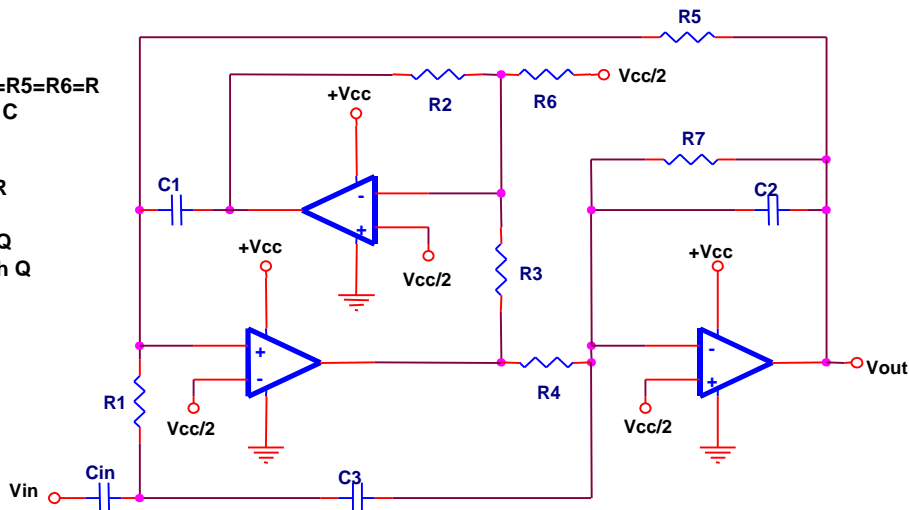
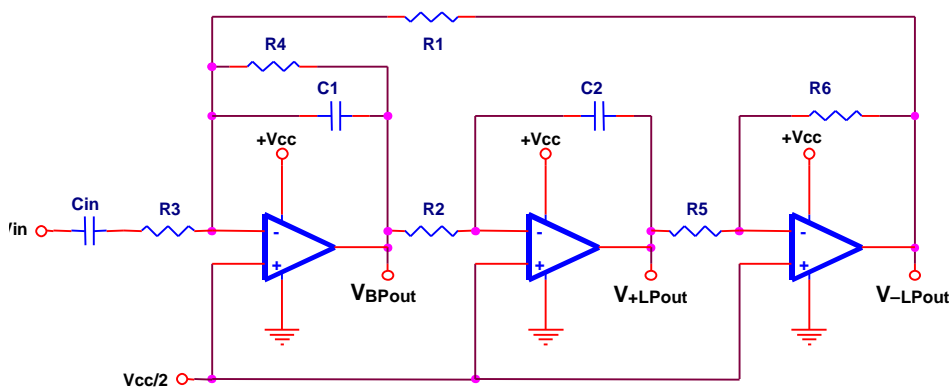


Figure 30. Akerberg-Mossberg Notch Filter

3.2.6 BiQuad

Biquad is a well know topology (Figure 31). It is only available in low-pass and band-pass varieties. The low-pass filter is useful whenever simultaneous normal and inverted outputs are needed.

**LOW PASS**

$R1 = R2 = R$
 $R5 = R6$, not critical
 $R4 = R/\sqrt{2}$
 $C1 = C2 = C$
 $F_o = 1/(2\pi RC)$

$R4 = R/\sqrt{2}$ Butterworth
 $R4 > R/\sqrt{2}$ Chebyshev
 $R4 < R/\sqrt{2}$ Bessel

Unity Gain: $R3 = R$
 Other Gain: $-R/R3$

BAND PASS

$R1 = R2 = R5 = R$
 $R6 = \text{about } R/\sqrt{2}$, not critical
 $C1 = C2 = C$
 $F_o = 1/(2\pi RC)$

$R3 = R4$ unity gain
 Gain = $-R4/R3$
 $R4$ also controls Q
 low value, low Q
 high value, high Q

Figure 31. BiQuad Low-Pass and Band-Pass Filter

3.2.7 State Variable

State variable is a three to four op-amp topology. The fourth op-amp is only required for notch filters. It is also very easy to tune, and it is easy to change the style of lowpass and highpass, and easy to change the Q of the bandpass and notch. Unfortunately, it is not as nice a topology as Akerberg-Mossberg. The same resistor is used for gain and style of filter / Q, limiting control of the filter. There is probably not a lot of reason to use this topology, unless simultaneous lowpass, highpass, bandpass, and notch outputs are required by the application.

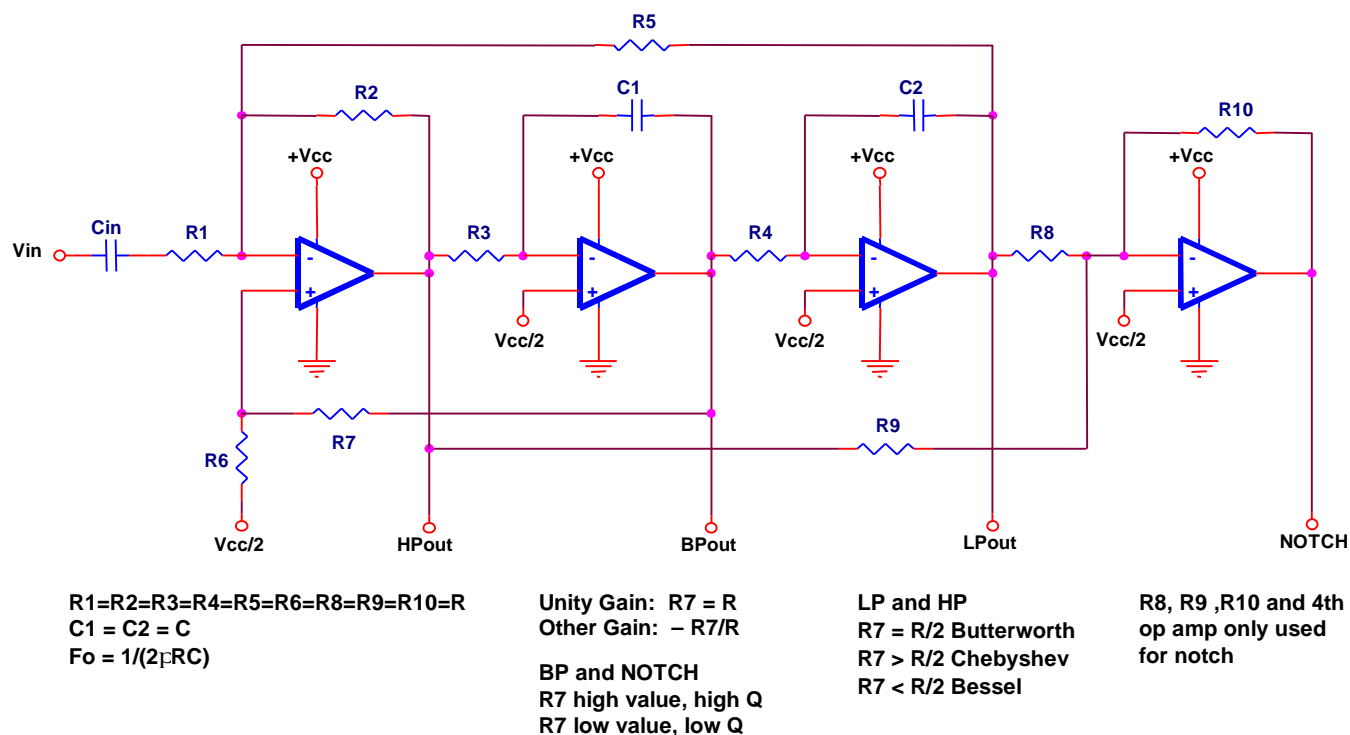


Figure 32. State-Variable Four-Op-Amp Topology

4 References

1. *Active Low Pass Filter Design*, Texas Instruments Application Report, Literature Number SLOA049
2. *Analysis Of The Sallen-Key Architecture*, Texas Instruments Application Report, Literature Number SLOA024A.
3. *Op Amps for Everyone*, Ron Mancini (Ed.), Chapter 12, Texas Instruments Literature Number SLOD006

Appendix A – Standard Resistor and Capacitor Values

E-12 Resistor / Capacitor Values

1.0, 1.2, 1.5, 1.8, 2.2, 2.7, 3.3, 3.9, 4.7, 5.6, 6.8, and 8.2; multiplied by the decade value.

E-24 Resistor / Capacitor Values

1.0, 1.1, 1.2, 1.3, 1.5, 1.6, 1.8, 2.0, 2.2, 2.4, 2.7, 3.0, 3.3, 3.6, 3.9, 4.3, 4.7, 5.1, 5.6, 6.2, 6.8, 7.5, 8.2, and 9.1; multiplied by the decade value.

E-96 Resistor / Capacitor Values

1.00, 1.02, 1.05, 1.07, 1.10, 1.13, 1.15, 1.18, 1.21, 1.24, 1.27, 1.30, 1.33, 1.37, 1.40, 1.43, 1.47, 1.50, 1.54, 1.58, 1.62, 1.65, 1.69, 1.74, 1.78, 1.82, 1.87, 1.91, 1.96, 2.00, 2.05, 2.10, 2.15, 2.21, 2.26, 2.32, 2.37, 2.43, 2.49, 2.55, 2.61, 2.67, 2.74, 2.80, 2.87, 2.94, 3.01, 3.09, 3.16, 3.24, 3.32, 3.40, 3.48, 3.57, 3.65, 3.74, 3.83, 3.92, 4.02, 4.12, 4.22, 4.32, 4.42, 4.53, 4.64, 4.75, 4.87, 4.99, 5.11, 5.23, 5.36, 5.49, 5.62, 5.76, 5.90, 6.04, 6.19, 6.34, 6.49, 6.65, 6.81, 6.98, 7.15, 7.32, 7.50, 7.68, 7.87, 8.06, 8.25, 8.45, 8.66, 8.87, 9.09, 9.31, 9.53, 9.76; multiplied by the decade value.

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