

## LMV716

# 5 MHz, Low Noise, RRO, Dual Operational Amplifier with CMOS Input

### General Description

The LMV716 is a dual operational amplifier with both low supply voltage and low supply current, making it ideal for portable applications. The LMV716 CMOS input stage drives the  $I_{BIAS}$  current down to 0.6 pA; this coupled with the low noise voltage of  $12.8 \text{ nV}/\sqrt{\text{Hz}}$  makes the LMV716 perfect for applications requiring active filters, transimpedance amplifiers, and HDD vibration cancellation circuitry.

Along with great noise sensitivity, small signal applications will benefit from the large gain bandwidth of 5 MHz coupled with the minimal supply current of 1.6 mA and a slew rate of  $5.8 \text{ V}/\mu\text{s}$ .

The LMV716 provides rail-to-rail output swing into heavy loads. The input common-mode voltage range includes ground, which is ideal for ground sensing applications.

The LMV716 has a supply voltage spanning 2.7V to 5V and is offered in an 8-pin MSOP package that functions across the wide temperature range of  $-40^\circ\text{C}$  to  $85^\circ\text{C}$ . This small package makes it possible to place the LMV716 next to sensors, thus reducing external noise pickup.

### Features

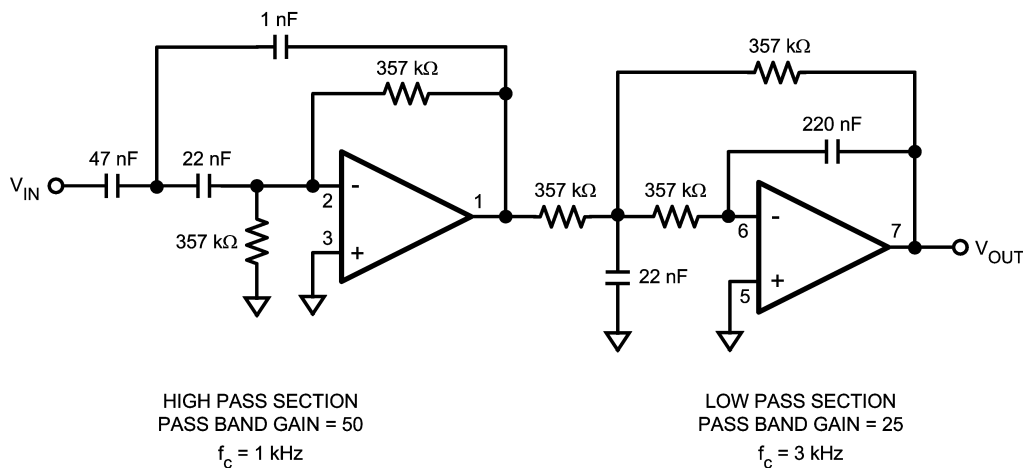
(Typical values,  $V^+ = 3.3\text{V}$ ,  $T_A = 25^\circ\text{C}$ , unless otherwise specified)

■ Input noise voltage	12.8 nV/√Hz
■ Input bias current	0.6 pA
■ Offset voltage	1.6 mV
■ CMRR	80 dB
■ Open loop gain	122 dB
■ Rail-to-rail output	
■ GBW	5 MHz
■ Slew rate	5.8 V/μs
■ Supply current	1.6 mA
■ Supply voltage range	2.7V to 5V
■ Operating temperature	$-40^\circ\text{C}$ to $85^\circ\text{C}$
■ 8-pin MSOP package	

### Applications

- Active filters
- Transimpedance amplifiers
- Audio preamp
- HDD vibration cancellation circuitry

### Typical Application Circuit



High Gain Band Pass Filter

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**Absolute Maximum Ratings** (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

ESD Tolerance (Note 2)	
Human Body Model	2000V
Machine Model	200V
Supply Voltage ( $V^+ - V^-$ )	5.5V
Storage Temperature Range	-65°C to 150°C
Junction Temperature (Note 3)	150°C max

Mounting Temperature

Infrared or Convection (20 sec)

260°C

**Operating Ratings** (Note 1)

Supply Voltage	2.7V to 5V
Temperature Range	-40°C to 85°C
Thermal Resistance ( $\theta_{JA}$ )	
8-Pin MSOP	195°C/W

**3.3V Electrical Characteristics** (Note 4)

Unless otherwise specified, all limits are guaranteed for  $T_J = 25^\circ\text{C}$ ,  $V^+ = 3.3\text{V}$ ,  $V^- = 0\text{V}$ .  $V_{CM} = V^+/2$ . **Boldface** limits apply at the temperature extremes (Note 5).

Symbol	Parameter	Condition	Min (Note 6)	Typ (Note 7)	Max (Note 6)	Units
$V_{OS}$	Input Offset Voltage	$V_{CM} = 1\text{V}$		1.6	5 <b>6</b>	mV
$I_B$	Input Bias Current	(Note 8)		0.6	115 <b>130</b>	pA
$I_{OS}$	Input Offset Current			1		pA
CMRR	Common Mode Rejection Ratio	$0 \leq V_{CM} \leq 2.1\text{V}$	60 <b>50</b>	80		dB
PSRR	Power Supply Rejection Ratio	$2.7\text{V} \leq V^+ \leq 5\text{V}$ , $V_{CM} = 1\text{V}$	70 <b>60</b>	82		dB
CMVR	Common Mode Voltage Range	For CMRR $\geq 50$ dB	-0.2		2.2	V
$A_{VOL}$	Open Loop Voltage Gain	Sourcing $R_L = 10\text{ k}\Omega$ to $V^+/2$ , $V_O = 1.65\text{V}$ to $2.9\text{V}$	80 <b>76</b>	122		dB
		Sinking $R_L = 10\text{ k}\Omega$ to $V^+/2$ , $V_O = 0.4\text{V}$ to $1.65\text{V}$	80 <b>76</b>	122		
		Sourcing $R_L = 600\Omega$ to $V^+/2$ , $V_O = 1.65\text{V}$ to $2.8\text{V}$	80 <b>76</b>	105		
		Sinking $R_L = 600\Omega$ to $V^+/2$ , $V_O = 0.5\text{V}$ to $1.65\text{V}$	80 <b>76</b>	112		
$V_O$	Output Swing High	$R_L = 10\text{ k}\Omega$ to $V^+/2$	3.22 <b>3.17</b>	3.29		V
		$R_L = 600\Omega$ to $V^+/2$	3.12 <b>3.07</b>	3.22		
	Output Swing Low	$R_L = 10\text{ k}\Omega$ to $V^+/2$		0.03	0.12 <b>0.16</b>	
		$R_L = 600\Omega$ to $V^+/2$		0.07	0.23 <b>0.27</b>	
$I_{OUT}$	Output Current	Sourcing, $V_O = 0\text{V}$	20 <b>15</b>	31		mA
		Sinking, $V_O = 3.3\text{V}$	30 <b>25</b>	41		
$I_S$	Supply Current	$V_{CM} = 1\text{V}$		1.6	2.0 <b>3</b>	mA
SR	Slew Rate	(Note 9)		5.8		V/ $\mu\text{s}$
GBW	Gain Bandwidth			5		MHz

### 3.3V Electrical Characteristics

(Note 4) Unless otherwise specified, all limits are guaranteed for  $T_J = 25^\circ\text{C}$ ,  $V^+ = 3.3\text{V}$ ,  $V^- = 0\text{V}$ .  $V_{CM} = V^+/2$ . **Boldface** limits apply at the temperature extremes (Note 5). (Continued)

Symbol	Parameter	Condition	Min (Note 6)	Typ (Note 7)	Max (Note 6)	Units
$e_n$	Input-Referred Voltage Noise	$f = 1\text{ kHz}$		12.8		$\text{nV}/\sqrt{\text{Hz}}$
$i_n$	Input-Referred Current Noise	$f = 1\text{ kHz}$		0.01		$\text{pA}/\sqrt{\text{Hz}}$

**Note 1:** Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but specific performance is not guaranteed. For guaranteed specifications and the test conditions, see the Electrical Characteristics.

**Note 2:** Human Body Model is  $1.5\text{ k}\Omega$  in series with  $100\text{ pF}$ . Machine Model is  $0\Omega$  in series with  $100\text{ pF}$ .

**Note 3:** The maximum power dissipation is a function of  $T_{J(\text{MAX})}$ ,  $\theta_{JA}$  and  $T_A$ . The maximum allowable power dissipation at any ambient temperature is  $P_D = (T_{J(\text{MAX})} - T_A)/\theta_{JA}$ . All numbers apply for packages soldered directly into a PC board.

**Note 4:** Electrical Table values apply only for factory testing conditions at the temperature indicated. Factor testing conditions result in very limited self-heating of the device such that  $T_J = T_A$ . No guarantee of parametric performance is indicated in the electrical tables under conditions of internal self-heating where  $T_J > T_A$ . Absolute Maximum Ratings indicate junction temperature limits beyond which the device maybe permanently degraded, either mechanically or electrically.

**Note 5:** Boldface limits apply to temperature range of  $-40^\circ\text{C}$  to  $85^\circ\text{C}$ .

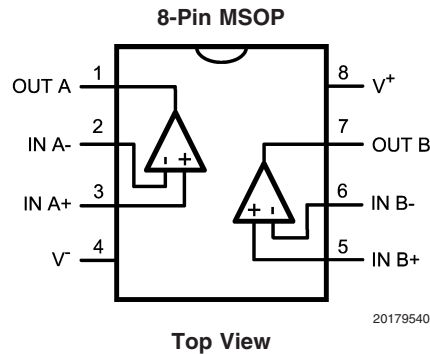
**Note 6:** All limits are guaranteed by testing or statistical analysis.

**Note 7:** Typical values represent the most likely parametric norm.

**Note 8:** Input bias current is guaranteed by design.

**Note 9:** Number specified is the lower of the positive and negative slew rates.

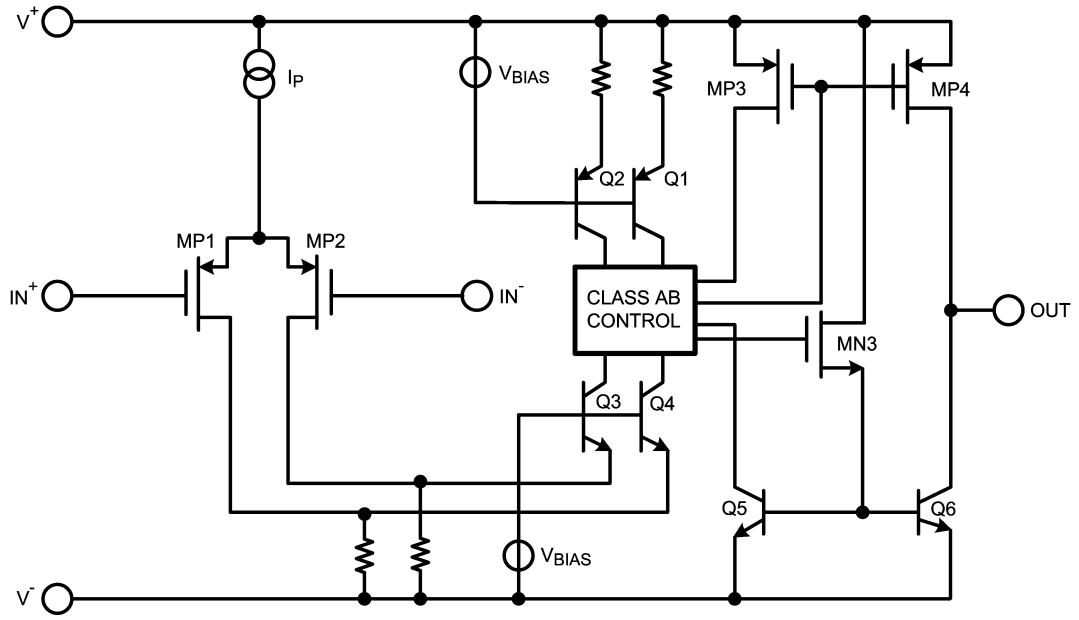
## Connection Diagram



## Ordering Information

Package	Part Number	Package Marking	Transport Media	NSC Drawing
8-Pin MSOP	LMV716MM	AR3A	1k Units Tape and Reel	MUA08A
	LMV716MMX		3.5k Units Tape and Reel	

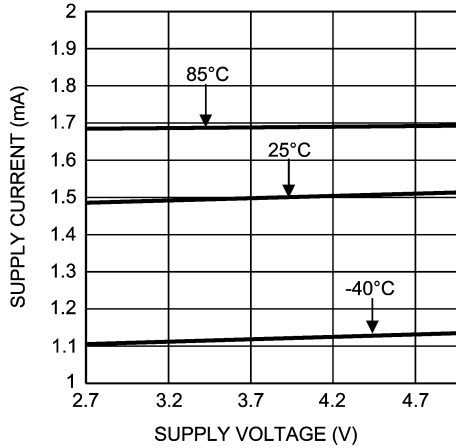
### Simplified Schematic



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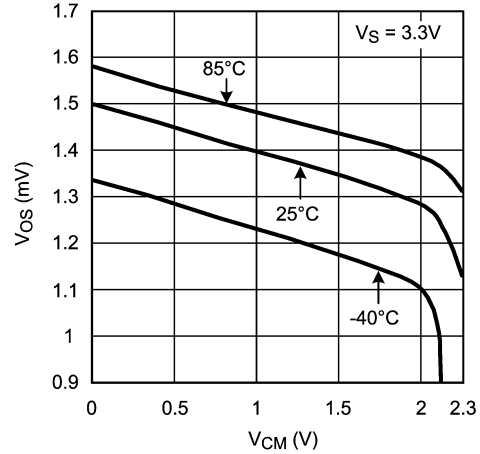
**Typical Performance Characteristics** Unless otherwise specified,  $V^+ = 3.3V$ ,  $T_J = 25^\circ C$ .

**Supply Current vs. Supply Voltage**



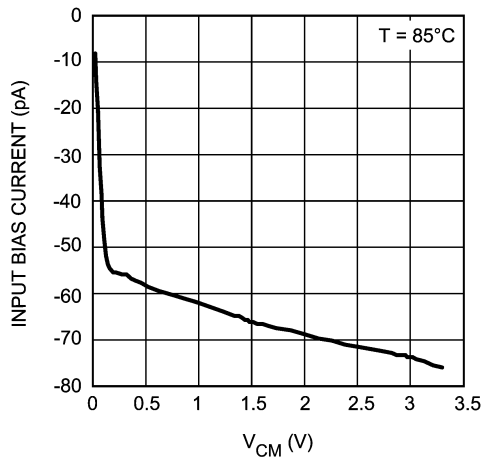
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**Offset Voltage vs. Common Mode**



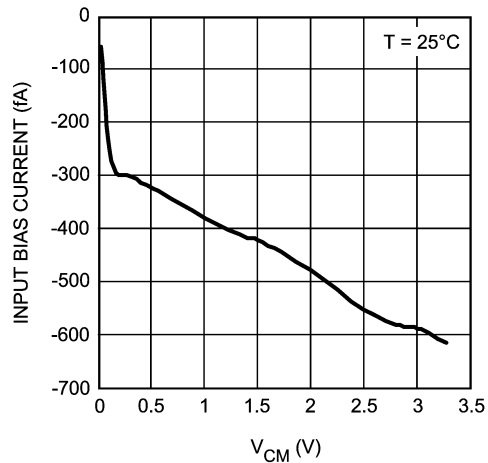
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**Input Bias Current vs. Common Mode**



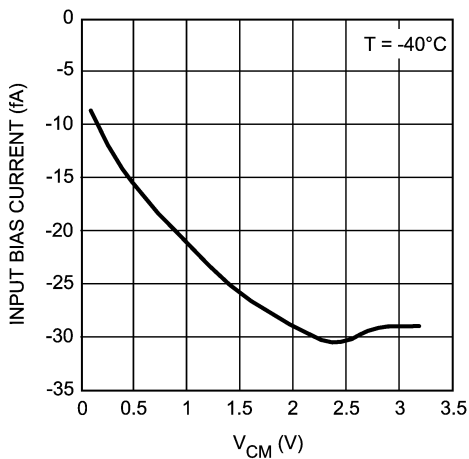
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**Input Bias Current vs. Common Mode**



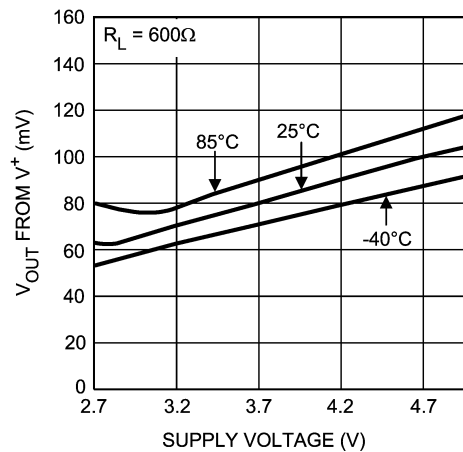
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**Input Bias Current vs. Common Mode**



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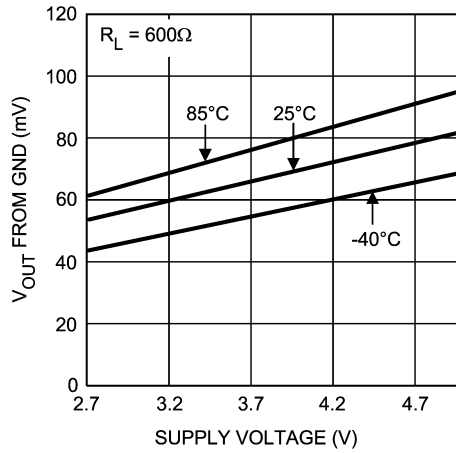
**Output Positive Swing vs. Supply Voltage**



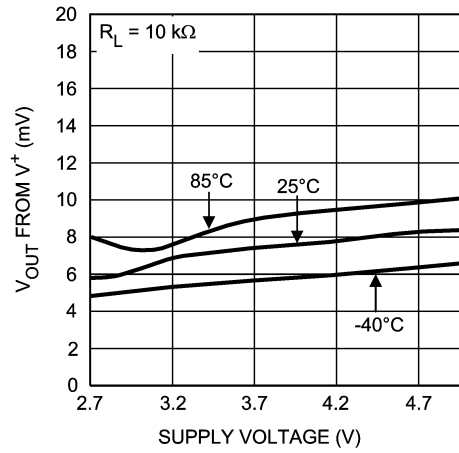
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**Typical Performance Characteristics** Unless otherwise specified,  $V^+ = 3.3V$ ,  $T_J = 25^\circ C$ . (Continued)

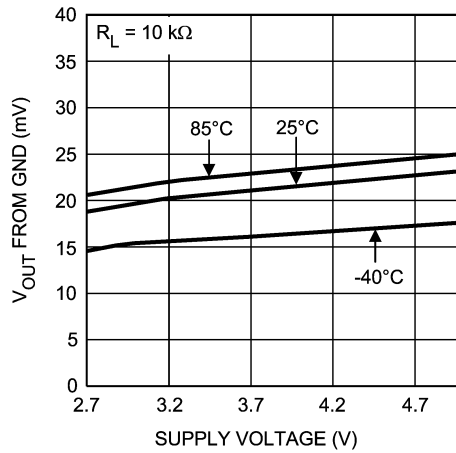
**Output Negative Swing vs. Supply Voltage**



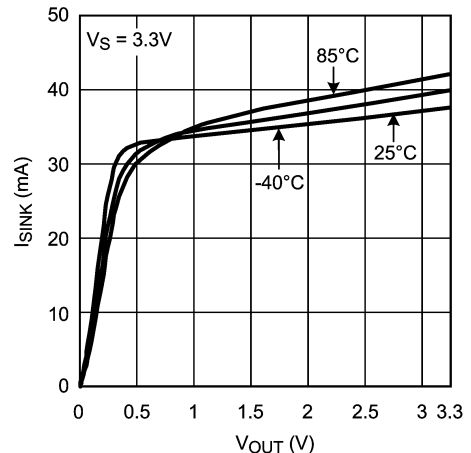
**Output Positive Swing vs. Supply Voltage**



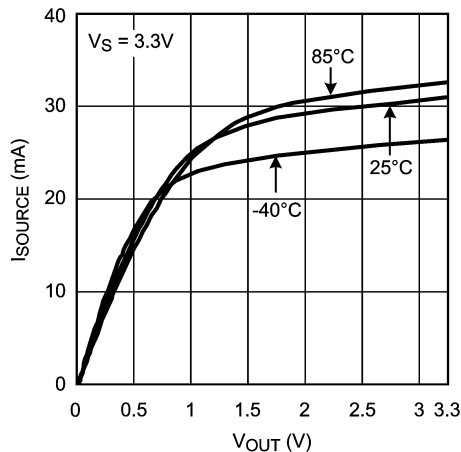
**Output Negative Swing vs. Supply Voltage**



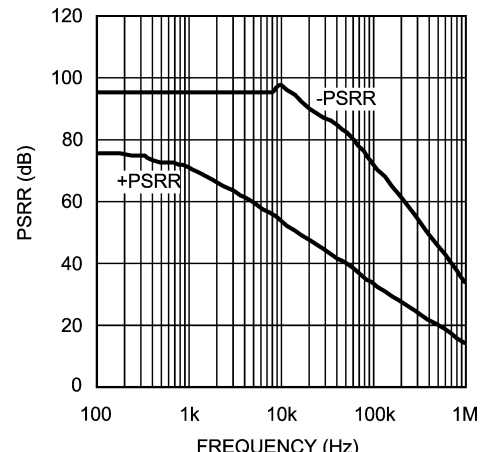
**Sinking Current vs.  $V_{OUT}$**



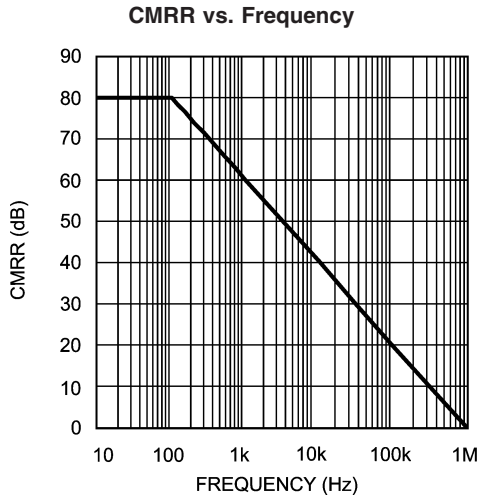
**Sourcing Current vs.  $V_{OUT}$**



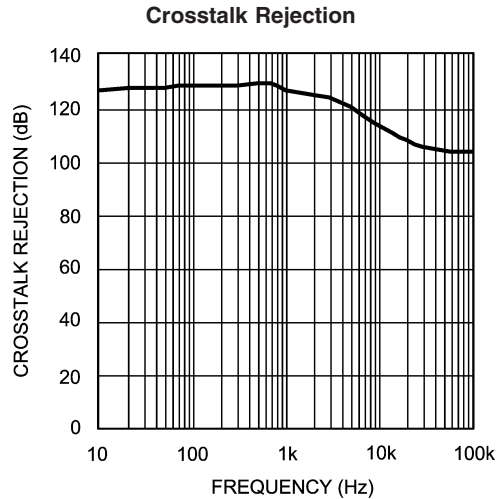
**PSRR vs. Frequency**



**Typical Performance Characteristics** Unless otherwise specified,  $V^+ = 3.3V$ ,  $T_J = 25^\circ C$ . (Continued)

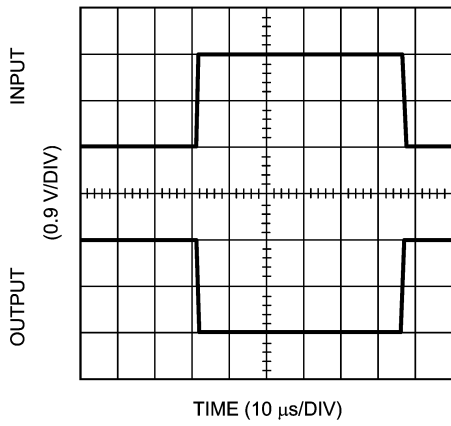


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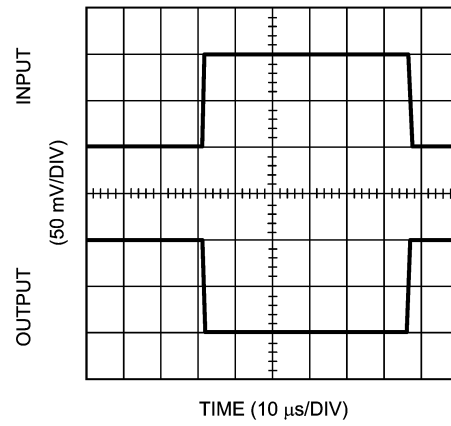
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**Inverting Large Signal Pulse Response**



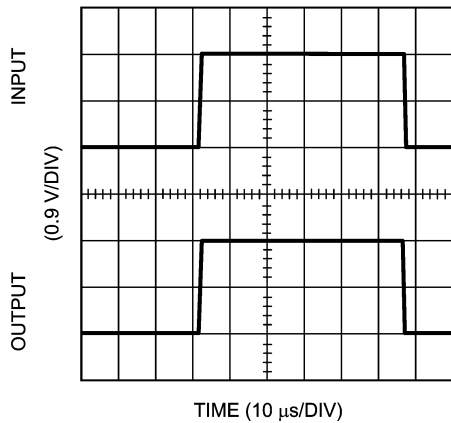
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**Inverting Small Signal Pulse Response**



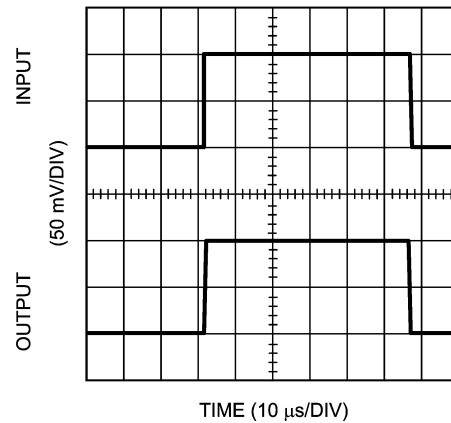
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**Non-Inverting Large Signal Pulse Response**



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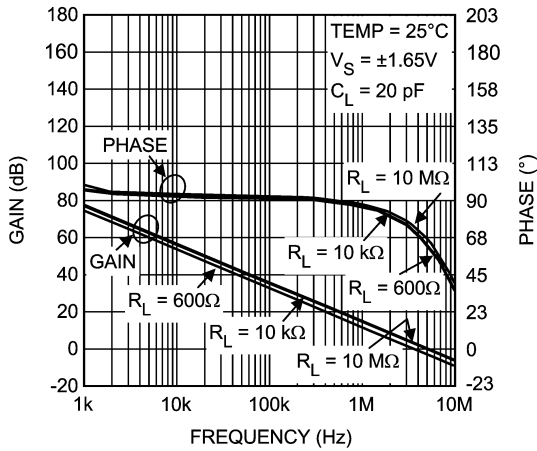
**Non-Inverting Small Signal Pulse Response**



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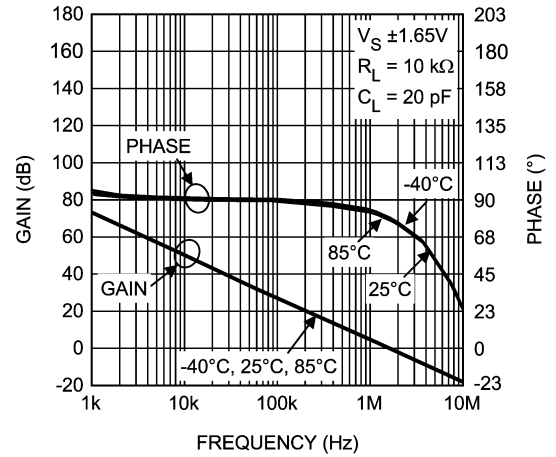
**Typical Performance Characteristics** Unless otherwise specified,  $V^+ = 3.3V$ ,  $T_J = 25^\circ C$ . (Continued)

**Open Loop Frequency vs.  $R_L$**



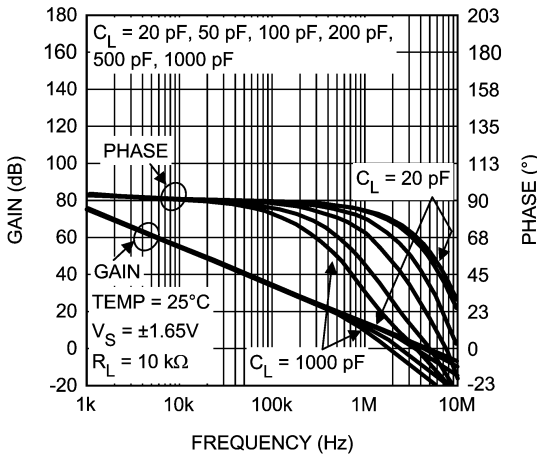
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**Open Loop Frequency Response over Temperature**



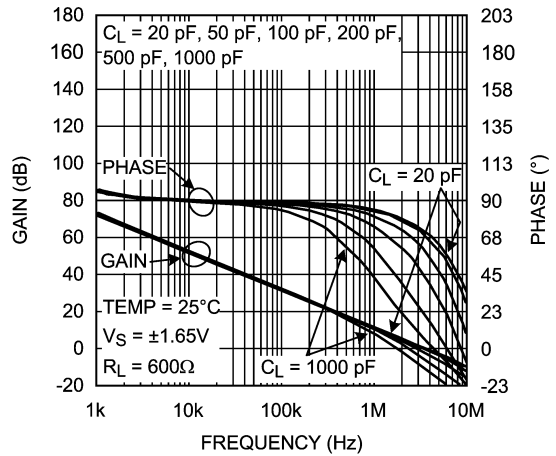
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**Open Loop Frequency Response vs.  $C_L$**



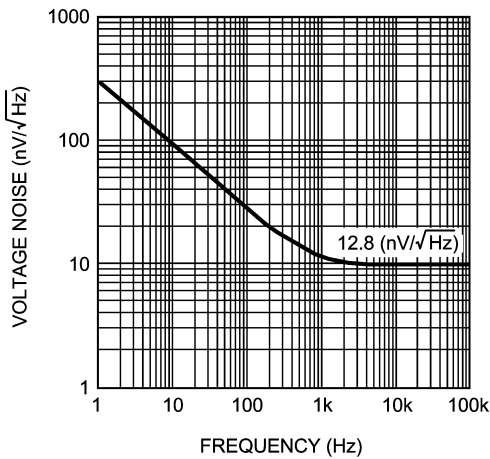
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**Open Loop Frequency Response vs.  $C_L$**



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**Voltage Noise vs. Frequency**



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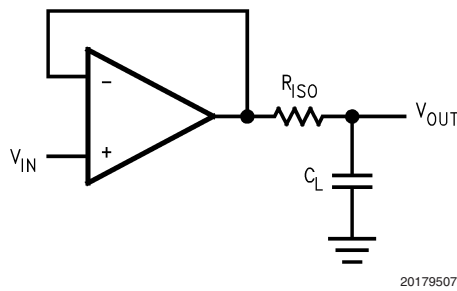


## Application Information

With the low supply current of only 1.6 mA, the LMV716 offers users the ability to maximize battery life. This makes the LMV716 ideal for battery powered systems. The LMV716's rail-to-rail output swing provides the maximum possible dynamic range at the output. This is particularly important when operating on low supply voltages.

### CAPACITIVE LOAD TOLERANCE

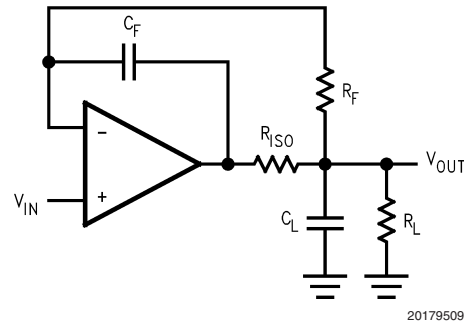
The LMV716, when in a unity-gain configuration, can directly drive large capacitive loads in unity-gain without oscillation. The unity-gain follower is the most sensitive configuration to capacitive loading; direct capacitive loading reduces the phase margin of amplifiers. The combination of the amplifier's output impedance and the capacitive load induces phase lag. This results in either an underdamped pulse response or oscillation. To drive a heavier capacitive load, the circuit in *Figure 1* can be used.



**FIGURE 1. Indirectly Driving a Capacitive Load using Resistive Isolation**

In *Figure 1*, the isolation resistor  $R_{ISO}$  and the load capacitor  $C_L$  form a pole to increase stability by adding more phase margin to the overall system. The desired performance depends on the value of  $R_{ISO}$ . The bigger the  $R_{ISO}$  resistor value, the more stable  $V_{OUT}$  will be.

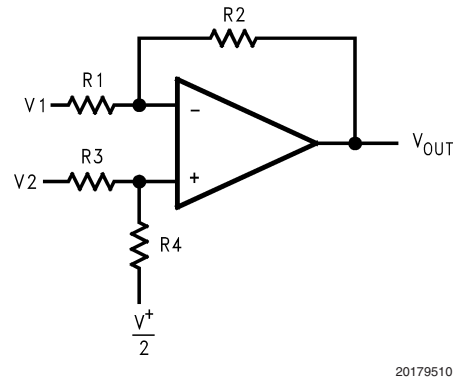
The circuit in *Figure 2* is an improvement to the one in *Figure 1* because it provides DC accuracy as well as AC stability. If there were a load resistor in *Figure 1*, the output would be voltage divided by  $R_{ISO}$  and the load resistor. Instead, in *Figure 2*,  $R_F$  provides the DC accuracy by using feed-forward techniques to connect  $V_{IN}$  to  $R_L$ . Due to the input bias current of the LMV716, the designer must be cautious when choosing the value of  $R_F$ .  $C_F$  and  $R_{ISO}$  serve to counteract the loss of phase margin by feeding the high frequency component of the output signal back to the amplifier's inverting input, thereby preserving phase margin in the overall feedback loop. Increased capacitive drive is possible by increasing the value of  $C_F$ . This in turn will slow down the pulse response.



**FIGURE 2. Indirectly Driving a Capacitive Load with DC Accuracy**

### DIFFERENCE AMPLIFIER

The difference amplifier allows the subtraction of two voltages or, as a special case, the cancellation of a signal common to two inputs. It is useful as a computational amplifier in making a differential to single-ended conversion or in rejecting a common mode signal.



$$V_{OUT} = \left( \frac{R1 + R2}{R3 + R4} \right) \frac{R4}{R1} V_2 - \frac{R2}{R1} V_1 + \left( \frac{R1 + R2}{R3 + R4} \right) \frac{R3}{R1} \cdot \frac{V^+}{2}$$

for  $R1 = R3$  and  $R2 = R4$

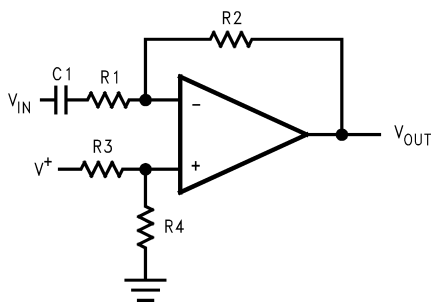
$$V_{OUT} = \frac{R2}{R1} (V_2 - V_1) + \frac{V^+}{2}$$

**FIGURE 3. Difference Amplifier**

## Application Information (Continued)

### SINGLE-SUPPLY INVERTING AMPLIFIER

There may be cases where the input signal going into the amplifier is negative. Because the amplifier is operating in single supply voltage, a voltage divider using  $R_3$  and  $R_4$  is implemented to bias the amplifier so the inverting input signal is within the input common voltage range of the amplifier. The capacitor  $C_1$  is placed between the inverting input and resistor  $R_1$  to block the DC signal going into the AC signal source,  $V_{IN}$ . The values of  $R_1$  and  $C_1$  affect the cutoff frequency,  $f_c = 1/2\pi R_1 C_1$ . As a result, the output signal is centered around mid-supply (if the voltage divider provides  $V^+/2$  at the non-inverting input). The output can swing to both rails, maximizing the signal-to-noise ratio in a low voltage system.



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$$V_{OUT} = -\frac{R_2}{R_1} V_{IN}$$

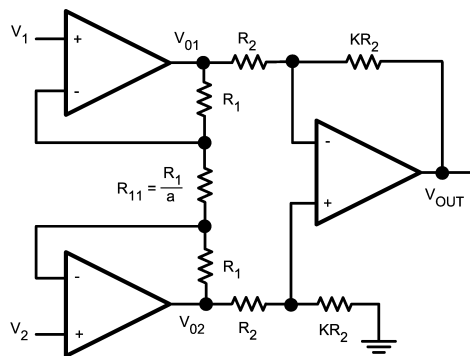
FIGURE 4. Single-supply Inverting Amplifier

### INSTRUMENTATION AMPLIFIER

Measurement of very small signals with an amplifier requires close attention to the input impedance of the amplifier, the overall signal gain from both inputs to the output, as well as, the gain from each input to the output. This is because we are only interested in the difference of the two inputs and the common signal is considered noise. A classic solution is an instrumentation amplifier. Instrumentation amplifiers have a finite, accurate, and stable gain. Also they have extremely high input impedances and very low output impedances. Finally they have an extremely high CMRR so that the amplifier can only respond to the differential signal.

### Three-Op-Amp Instrumentation Amplifier

A typical instrumentation amplifier is shown in Figure 5.



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FIGURE 5. Three-Op-Amp Instrumentation Amplifier

There are two stages in this configuration. The last stage, the output stage, is a differential amplifier. In an ideal case the two amplifiers of the first stage, the input stage, would be set up as buffers to isolate the inputs. However they cannot be connected as followers due to the mismatch of real amplifiers. The circuit in Figure 5 utilizes a balancing resistor between the two amplifiers to compensate for this mismatch. The product of the two stages of gain will be the gain of the instrumentation amplifier circuit. Ideally, the CMRR should be infinite. However the output stage has a small non-zero common mode gain which results from resistor mismatch.

In the input stage of the circuit, current is the same across all resistors. This is due to the high input impedance and low input bias current of the LMV716. With the node equations we have:

$$\text{GIVEN: } I_{R_1} = I_{R_{11}} \quad (1)$$

By Ohm's Law:

$$\begin{aligned} V_{O1} - V_{O2} &= (2R_1 + R_{11}) I_{R_{11}} \\ &= (2a + 1) R_{11} \cdot I_{R_{11}} \\ &= (2a + 1) V_{R_{11}} \end{aligned} \quad (2)$$

However:

$$V_{R_{11}} = V_1 - V_2 \quad (3)$$

So we have:

$$V_{O1} - V_{O2} = (2a + 1) (V_1 - V_2) \quad (4)$$

## Application Information (Continued)

Now looking at the output of the instrumentation amplifier:

$$V_O = \frac{KR_2}{R_2} (V_{O2} - V_{O1}) = -K(V_{O1} - V_{O2}) \quad (5)$$

Substituting from Equation (4):

$$V_O = -K(2a + 1)(V_1 - V_2) \quad (6)$$

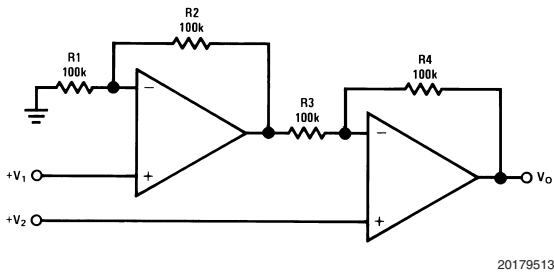
This shows the gain of the instrumentation amplifier to be:  $-K(2a+1)$

Typical values for this circuit can be obtained by setting:  $a = 12$  and  $K = 4$ . This results in an overall gain of  $-100$ .

Three LMV716 amplifiers are used along with 1% resistors to minimize resistor mismatch. Resistors used to build the circuit are:  $R_1 = 21.6 \text{ k}\Omega$ ,  $R_{11} = 1.8 \text{ k}\Omega$ ,  $R_2 = 2.5 \text{ k}\Omega$  with  $K = 40$  and  $a = 12$ . This results in an overall gain of  $-K(2a+1) = -1000$ .

### Two-Op-Amp Instrumentation Amplifier

A two-op-amp instrumentation amplifier can also be used to make a high-input impedance DC differential amplifier *Figure 6*. As in the three op amp circuit, this instrumentation amplifier requires precise resistor matching for good CMRR.  $R_4$  should be equal to  $R_1$ , and  $R_3$  should equal  $R_2$ .



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$$V_O = \left(1 + \frac{R_4}{R_3}\right)(V_2 - V_1), \text{ where } R_1 = R_4 \text{ and } R_2 = R_3$$

$$\text{As shown: } V_O = 2(V_2 - V_1)$$

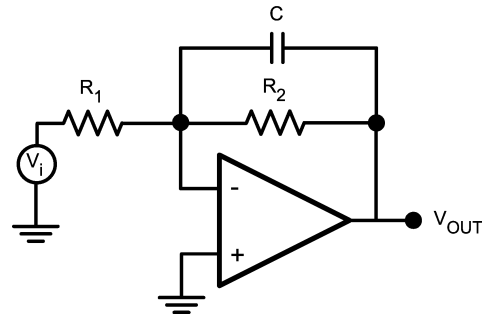
FIGURE 6. Two-Op-Amp Instrumentation Amplifier

### ACTIVE FILTERS

Active filters are circuits with amplifiers, resistors, and capacitors. The use of amplifiers instead of inductors, which are used in passive filters, enhances the circuit performance while reducing the size and complexity of the filter. The simplest active filters are designed using an inverting op amp configuration where at least one reactive element has been added to the configuration. This means that the op amp will provide "frequency-dependent" amplification, since reactive elements are frequency dependent devices.

### Low Pass Filter

The following shows a very simple low pass filter.



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FIGURE 7. Low Pass Filter

The transfer function can be expressed as follows:

By KCL:

$$\frac{-V_i}{R_1} - \left[\frac{1}{j\omega C}\right] \frac{V_O}{R_2} = 0 \quad (7)$$

Simplifying this further results in:

$$V_O = \frac{-R_2}{R_1} \left[\frac{1}{j\omega C R_2 + 1}\right] V_i \quad (8)$$

or

$$\frac{V_O}{V_i} = \frac{-R_2}{R_1} \left[\frac{1}{j\omega C R_2 + 1}\right] \quad (9)$$

Now, substituting  $\omega = 2\pi f$ , so that the calculations are in  $f(\text{Hz})$  rather than in  $\omega(\text{rad/s})$ , and setting the DC gain  $\left[\frac{-R_2}{R_1} = H_O\right]$  and  $H = \frac{V_O}{V_i}$

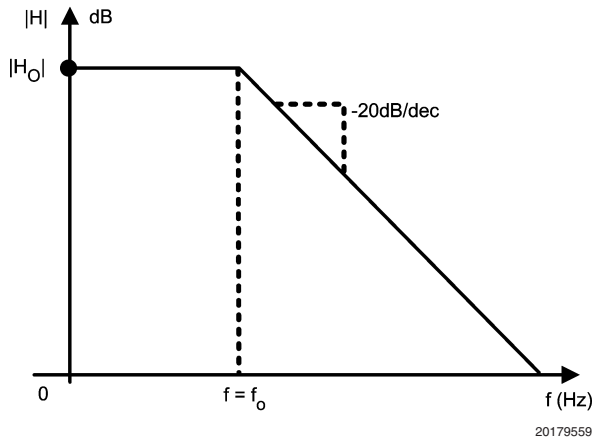
$$H = H_O \left[\frac{1}{j2\pi f C R_2 + 1}\right] \quad (10)$$

$$\text{set: } f_O = \frac{1}{2\pi R_1 C}$$

$$H = H_O \left[\frac{1}{1 + j(f/f_O)}\right] \quad (11)$$

Low pass filters are known as lossy integrators because they only behave as integrators at higher frequencies. The general form of the bode plot can be predicted just by looking at the transfer function. When the  $f/f_O$  ratio is small, the capacitor is, in effect, an open circuit and the amplifier behaves at a set DC gain. Starting at  $f_O$ , which is the  $-3 \text{ dB}$  corner, the capacitor will have the dominant impedance and hence the circuit will behave as an integrator and the signal will be attenuated and eventually cut. The bode plot for this filter is shown in *Figure 8*.

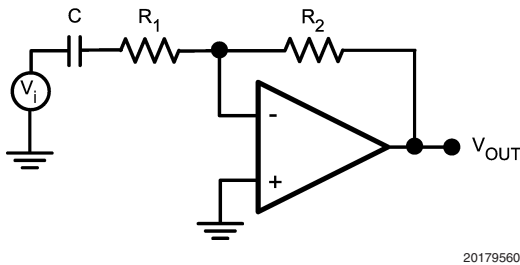
**Application Information** (Continued)



**FIGURE 8. Low Pass Filter Transfer Function**

**High Pass Filter**

The transfer function of a high pass filter can be derived in much the same way as the previous example. A typical first order high pass filter is shown below:



**FIGURE 9. High Pass Filter**

Writing the KCL for this circuit :  
( $V_1$  denotes the voltage between C and  $R_1$ )

$$\frac{V_1 - V_i}{\frac{1}{j\omega C}} = \frac{V_1 - V^-}{R_1} \tag{12}$$

$$\frac{V^- + V_1}{R_1} = \frac{V^- + V_O}{R_2} \tag{13}$$

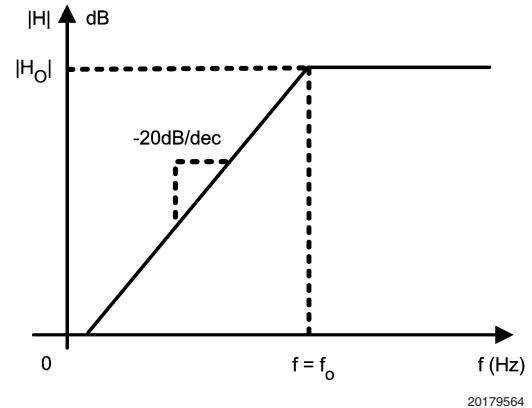
Solving these two equations to find the transfer function and using:

$$f_O = \frac{1}{2\pi R_1 C}$$

(high frequency gain)  $H_O = \frac{-R_2}{R_1}$  and  $H = \frac{V_O}{V_i}$   
Which gives:

$$H = H_O \frac{j(f/f_O)}{1 + j(f/f_O)} \tag{14}$$

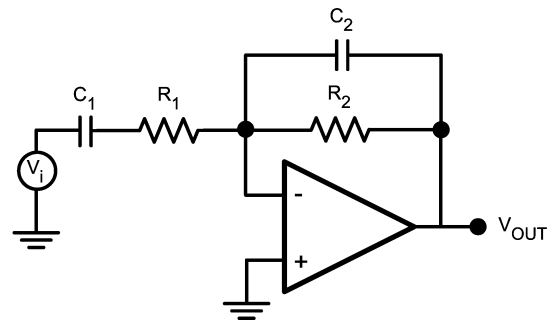
Looking at the transfer function, it is clear that when  $f/f_O$  is small, the capacitor is open and therefore, no signal is getting to the amplifier. As the frequency increases the amplifier starts operating. At  $f = f_O$  the capacitor behaves like a short circuit and the amplifier will have a constant, high frequency gain of  $H_O$ . Figure 10 shows the transfer function of this high pass filter.



**FIGURE 10. High Pass Filter Transfer Function**

**Band Pass Filter**

Combining a low pass filter and a high pass filter will generate a band pass filter. Figure 11 offers an example of this type of circuit.



**FIGURE 11. Band Pass Filter**

In this network the input impedance forms the high pass filter while the feedback impedance forms the low pass filter. If the designer chooses the corner frequencies so that  $f_1 < f_2$ , then all the frequencies between,  $f_1 \leq f \leq f_2$ , will pass through the filter while frequencies below  $f_1$  and above  $f_2$  will be cut off. The transfer function can be easily calculated using the same methodology as before and is shown in Figure 12.

$$H = H_O \frac{j(f/f_1)}{[1 + j(f/f_1)][1 + j(f/f_2)]} \tag{15}$$

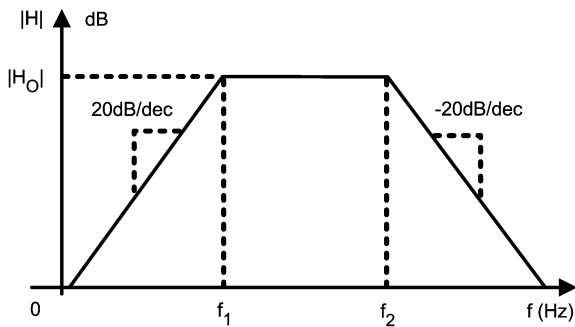
## Application Information (Continued)

Where

$$f_1 = \frac{1}{2\pi R_1 C_1}$$

$$f_2 = \frac{1}{2\pi R_2 C_2}$$

$$H_0 = \frac{-R_2}{R_1} \tag{16}$$

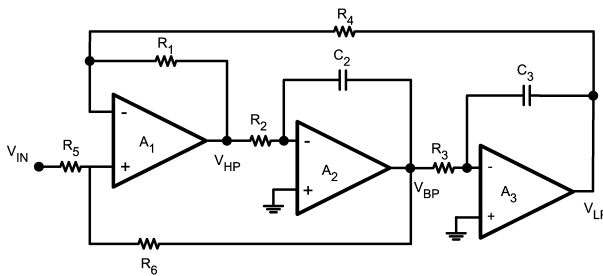


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FIGURE 12. Band Pass Filter Transfer Function

### STATE VARIABLE ACTIVE FILTER

State variable active filters are circuits that can simultaneously represent high pass, band pass, and low pass filters. The state variable active filter uses three separate amplifiers to achieve this task. A typical state variable active filter is shown in Figure 13. The first amplifier in the circuit is connected as a gain stage. The second and third amplifiers are connected as integrators, which means they behave as low pass filters. The feedback path from the output of the third amplifier to the first amplifier enables this low frequency signal to be fed back with a finite and fairly low closed loop gain. This is while the high frequency signal on the input is still gained up by the open loop gain of the first amplifier. This makes the first amplifier a high pass filter. The high pass signal is then fed into a low pass filter. The outcome is a band pass signal, meaning the second amplifier is a band pass filter. This signal is then fed into the third amplifiers input and so, the third amplifier behaves as a simple low pass filter.

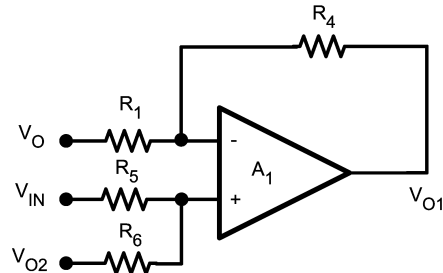


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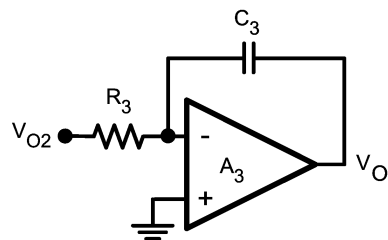
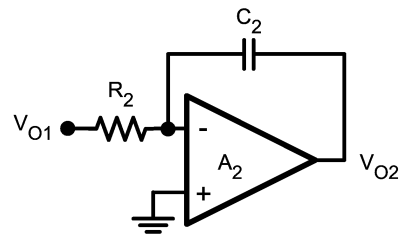
FIGURE 13. State Variable Active Filter

The transfer function of each filter needs to be calculated. The derivations will be more trivial if each stage of the filter is shown on its own.

The three components are:



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For A<sub>1</sub> the relationship between input and output is:

$$V_{O1} = \frac{-R_4}{R_1} V_0 + \left[ \frac{R_6}{R_5 + R_6} \right] \left[ \frac{R_1 + R_4}{R_1} \right] V_{IN} + \left[ \frac{R_5}{R_5 + R_6} \right] \left[ \frac{R_1 + R_4}{R_1} \right] V_{O2} \tag{17}$$

This relationship depends on the output of all the filters. The input-output relationship for A<sub>2</sub> can be expressed as:

$$V_{O2} = \frac{-1}{s C_2 R_2} V_{O1} \tag{18}$$

And finally this relationship for A<sub>3</sub> is as follows:

$$V_O = \frac{-1}{s C_3 R_3} V_{O2} \tag{19}$$

Re-arranging these equations, one can find the relationship between V<sub>O</sub> and V<sub>IN</sub> (transfer function of the low pass filter), V<sub>O1</sub> and V<sub>IN</sub> (transfer function of the high pass filter), and V<sub>O2</sub> and V<sub>IN</sub> (transfer function of the band pass filter) These relationships are as follows:

## Application Information (Continued)

### Low Pass Filter

$$\frac{V_O}{V_{IN}} = \frac{\left[ \frac{R_1 + R_4}{R_1} \right] \left[ \frac{R_6}{R_5 + R_6} \right] \left[ \frac{1}{C_2 C_3 R_2 R_3} \right]}{s^2 + s \left[ \frac{1}{C_2 R_2} \right] \left[ \frac{R_5}{R_5 + R_6} \right] \left[ \frac{R_1 + R_4}{R_1} \right] + \left[ \frac{1}{C_2 C_3 R_2 R_3} \right]} \quad (20)$$

### High Pass Filter

$$\frac{V_{O1}}{V_{IN}} = \frac{s^2 \left[ \frac{R_1 + R_4}{R_1} \right] \left[ \frac{R_6}{R_5 + R_6} \right]}{s^2 + s \left[ \frac{1}{C_2 R_2} \right] \left[ \frac{R_5}{R_5 + R_6} \right] \left[ \frac{R_1 + R_4}{R_1} \right] + \left[ \frac{1}{C_2 C_3 R_2 R_3} \right]} \quad (21)$$

### Band Pass Filter

$$\frac{V_{O2}}{V_{IN}} = \frac{s \left[ \frac{1}{C_2 R_2} \right] \left[ \frac{R_1 + R_4}{R_1} \right] \left[ \frac{R_6}{R_5 + R_6} \right]}{s^2 + s \left[ \frac{1}{C_2 R_2} \right] \left[ \frac{R_5}{R_5 + R_6} \right] \left[ \frac{R_1 + R_4}{R_1} \right] + \left[ \frac{1}{C_2 C_3 R_2 R_3} \right]} \quad (22)$$

The center frequency and Quality Factor for all of these filters is the same. The values can be calculated in the following manner:

$$\omega_c = \sqrt{\frac{1}{C_2 C_3 R_2 R_3}}$$

and

$$Q = \sqrt{\frac{C_2 R_2}{C_3 R_3} \left[ \frac{R_5 + R_6}{R_6} \right] \left[ \frac{R_1}{R_1 + R_4} \right]} \quad (23)$$

Designing a band pass filter with a center frequency of 10 kHz and Quality Factor of 5.5

To do this, first consider the Quality Factor. It is best to pick convenient values for the capacitors.  $C_2 = C_3 = 1000$  pF. Also, choose  $R_1 = R_4 = 30$  k $\Omega$ . Now values of  $R_5$  and  $R_6$  need to be calculated. With the chosen values for the capacitors and resistors, Q reduces to:

$$Q = \frac{11}{2} = \frac{1}{2} \left[ \frac{R_5 + R_6}{R_6} \right] \quad (24)$$

or

$$\begin{aligned} R_5 &= 10R_6 \\ R_6 &= 1.5 \text{ k}\Omega \\ R_5 &= 15 \text{ k}\Omega \end{aligned} \quad (25)$$

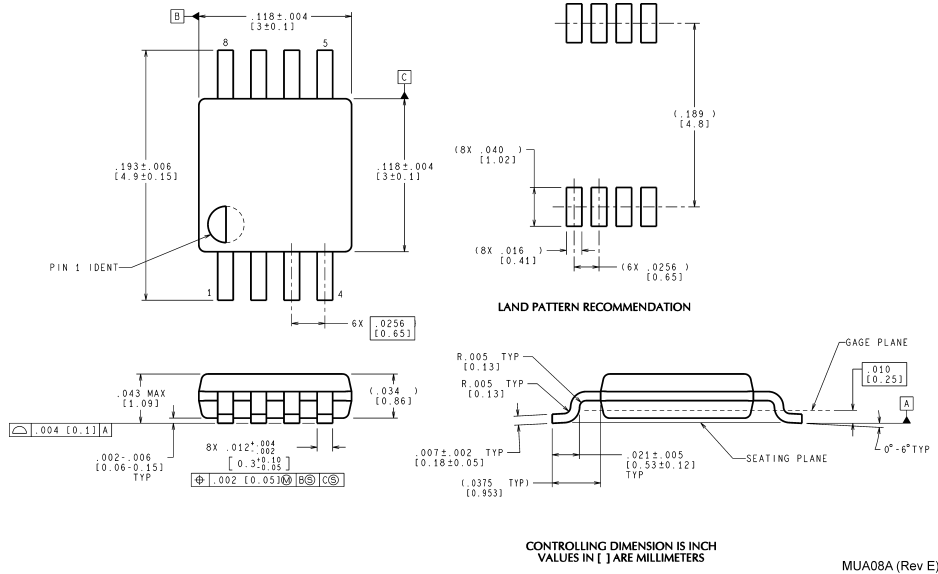
Also, for  $f = 10$  kHz, the center frequency is  $\omega_c = 2\pi f = 62.8$  kHz.

Using the expressions above, the appropriate resistor values will be  $R_2 = R_3 = 16$  k $\Omega$ .

The DC gain of this circuit is:

$$\text{DC GAIN} = \left[ \frac{R_1 + R_4}{R_1} \right] \left[ \frac{R_6}{R_5 + R_6} \right] = -14.8 \text{ dB} \quad (26)$$

**Physical Dimensions** inches (millimeters) unless otherwise noted



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