National Semiconductor

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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 nV/$\sqrt{Hz}$ 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 V/μs.

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˚C to 85˚C. This small package makes it possible to place the LMV716 next to sensors, thus reducing external noise pickup.

Features

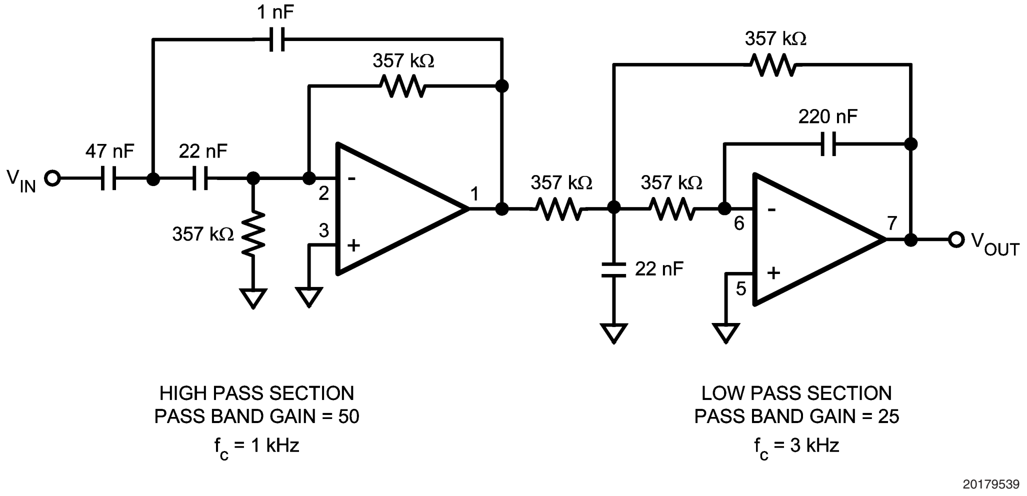
(Typical values, $V^+ = 3.3V$, $T_A = 25˚C$, unless otherwise specified)

■ Input noise voltage	12.8 nV/$\sqrt{Hz}$
■ 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˚C to 85˚C
■ 8-pin MSOP package	

Applications

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

Typical Application Circuit



High Gain Band Pass Filter

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 (θ_{JA})	
8-Pin MSOP	195˚C/W

3.3V Electrical Characteristics (Note 4)

Unless otherwise specified, all limits are guaranteed for T_J = 25˚C, V^+ = 3.3V, V^- = 0V. 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} = 1V		1.6	5 **6**	mV
I_B	Input Bias Current	(Note 8)		0.6	115 **130**	pA
I_{OS}	Input Offset Current			1		pA
CMRR	Common Mode Rejection Ratio	0 ≤ V_{CM} ≤ 2.1V	60 **50**	80		dB
PSRR	Power Supply Rejection Ratio	2.7V ≤ V^+ ≤ 5V, V_{CM} = 1V	70 **60**	82		dB
CMVR	Common Mode Voltage Range	For CMRR ≥ 50 dB	−0.2		2.2	V
A_{VOL}	Open Loop Voltage Gain	Sourcing R_L = 10 kΩ to V^+/2, V_O = 1.65V to 2.9V	80 **76**	122		dB
		Sinking R_L = 10 kΩ to V^+/2, V_O = 0.4V to 1.65V	80 **76**	122		
		Sourcing R_L = 600Ω to V^+/2, V_O = 1.65V to 2.8V	80 **76**	105		
		Sinking R_L = 600Ω to V^+/2, V_O = 0.5V to 1.65V	80 **76**	112		
V_O	Output Swing High	R_L = 10 kΩ to V^+/2	3.22 **3.17**	3.29		V
		R_L = 600Ω to V^+/2	3.12 **3.07**	3.22		
	Output Swing Low	R_L = 10 kΩ to V^+/2		0.03	0.12 **0.16**	
		R_L = 600Ω to V^+/2		0.07	0.23 **0.27**	
I_{OUT}	Output Current	Sourcing, V_O = 0V	20 **15**	31		mA
		Sinking, V_O = 3.3V	30 **25**	41		
I_S	Supply Current	V_{CM} = 1V		1.6	2.0 **3**	mA
SR	Slew Rate	(Note 9)		5.8		V/μs
GBW	Gain Bandwidth			5		MHz

3.3V Electrical Characteristics (Note 4) Unless otherwise specified, all limits are guaranteed for $T_J = 25^\circ C$, $V^+ = 3.3V$, $V^- = 0V$. $V_{CM} = V^+/2$. **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 kHz		12.8		$nV/\sqrt{Hz}$
i_n	Input-Referred Current Noise	f = 1 kHz		0.01		$pA/\sqrt{Hz}$

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 kΩ in series with 100 pF. Machine Model is 0Ω in series with 100 pF.

Note 3: The maximum power dissipation is a function of $T_{J(MAX)}$, θ_{JA} and T_A. The maximum allowable power dissipation at any ambient temperature is $P_D = (T_{J(MAX)}-T_A)/\theta_{JA}$. 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˚C to 85˚C.

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

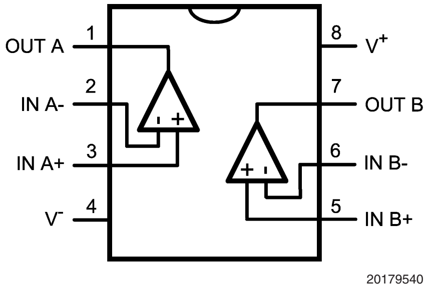
Note 7: Typical values represent the most likely parametric norm.

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

8-Pin MSOP

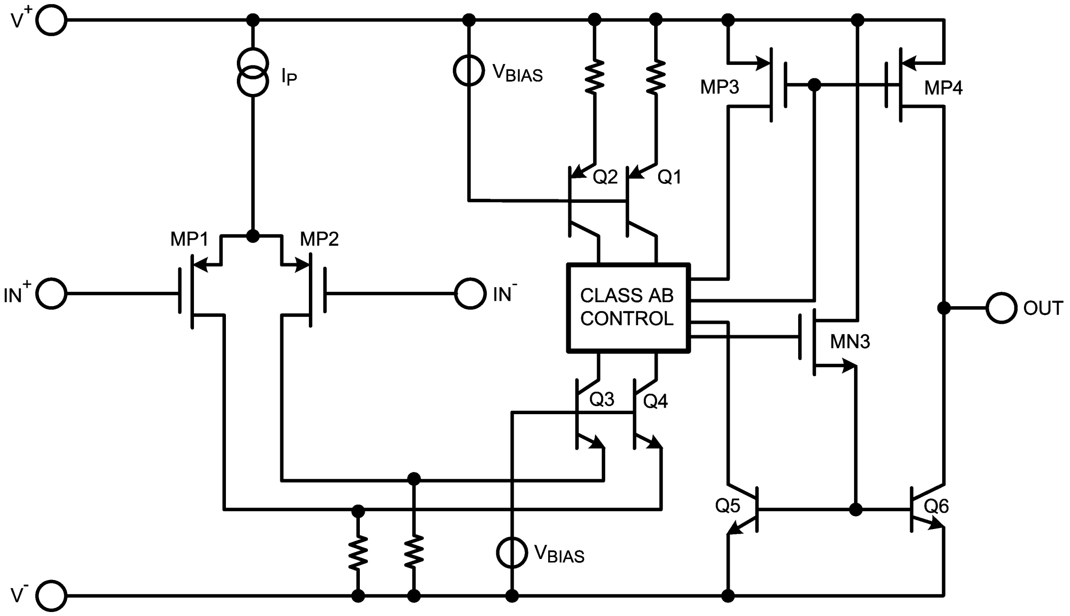


Top View

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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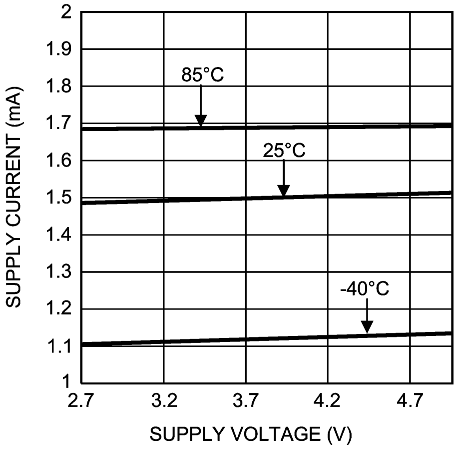
Typical Performance Characteristics

Unless otherwise specified, V^+ 3.3V, T_J = 25˚C.

Supply Current vs. Supply Voltage



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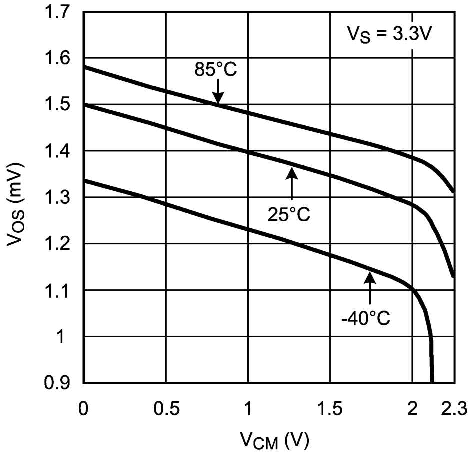
Offset Voltage vs. Common Mode



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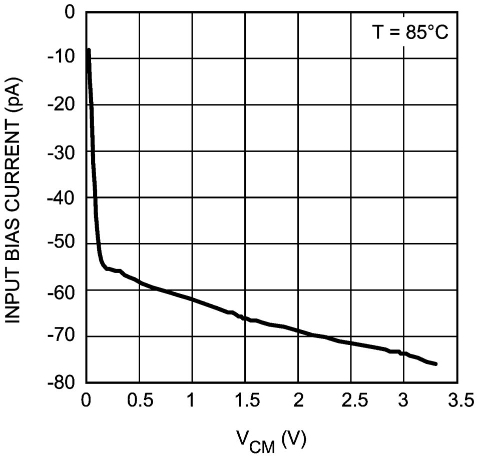
Input Bias Current vs. Common Mode



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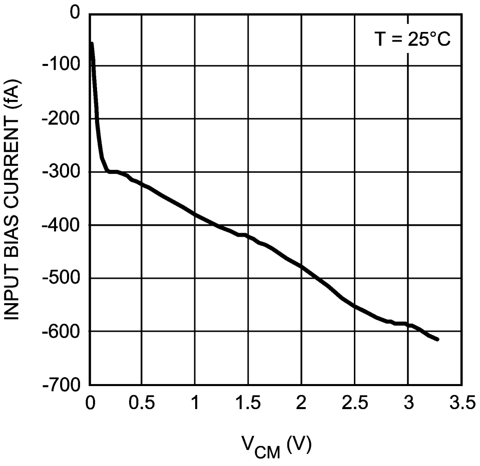
Input Bias Current vs. Common Mode



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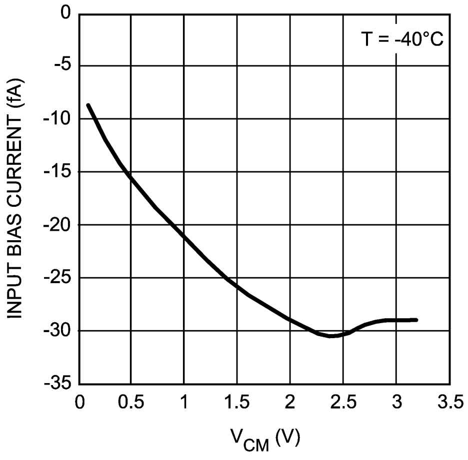
Input Bias Current vs. Common Mode



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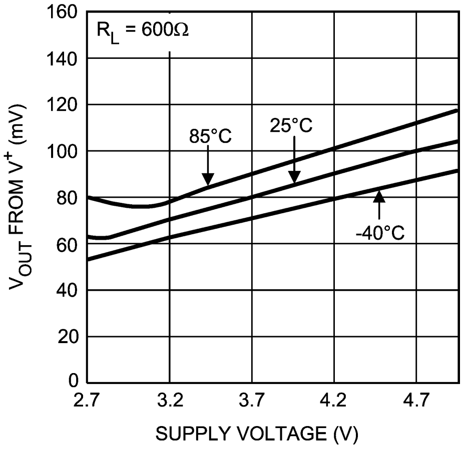
Output Positive Swing vs. Supply Voltage



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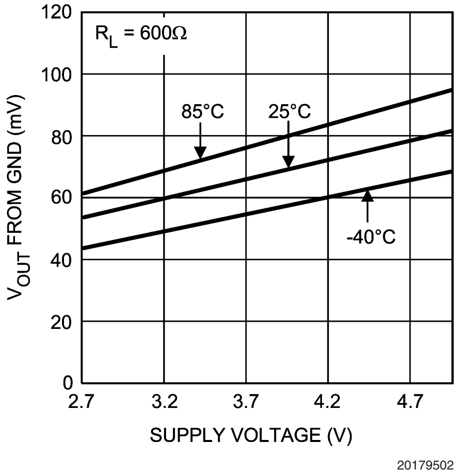
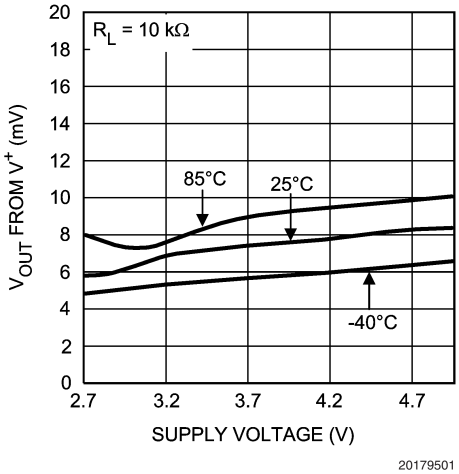
Typical Performance Characteristics Unless otherwise specified, V^+ 3.3V, T_J = 25°C. (Continued)

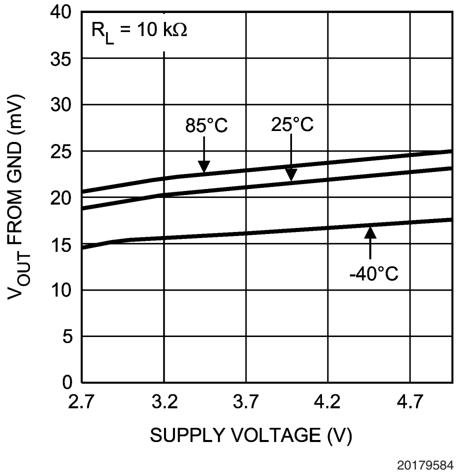
Output Negative Swing vs. Supply Voltage



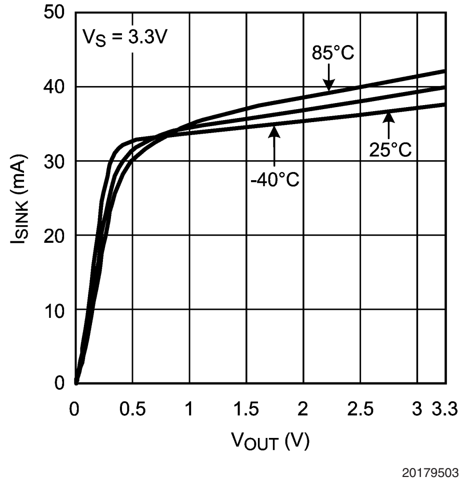
Output Positive Swing vs. Supply Voltage



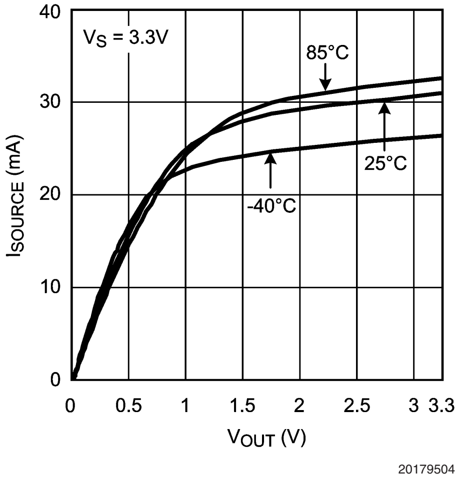
Output Negative Swing vs. Supply Voltage



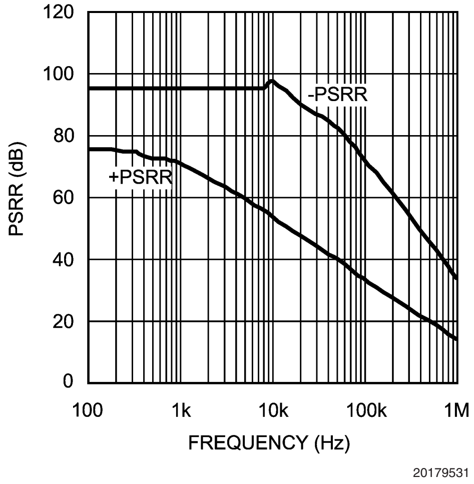
Sinking Current vs. V_{OUT}



Sourcing Current vs. V_{OUT}

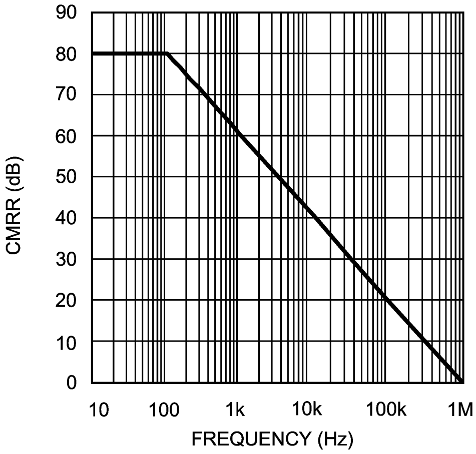


PSRR vs. Frequency



Typical Performance Characteristics Unless otherwise specified, V+ 3.3V, T_J = 25˚C. (Continued)

CMRR vs. Frequency



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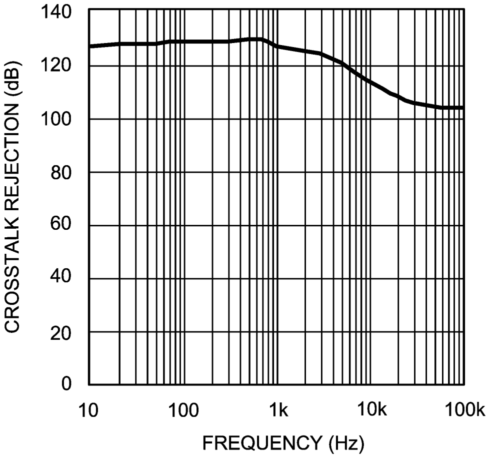
Crosstalk Rejection



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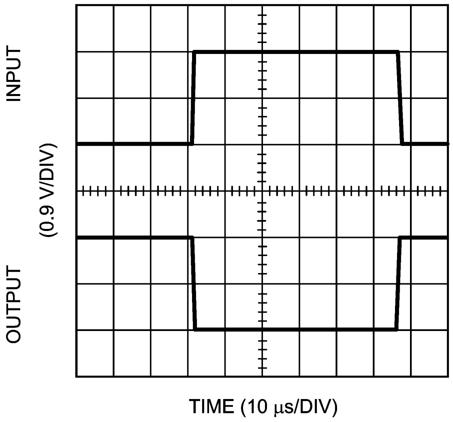
Inverting Large Signal Pulse Response



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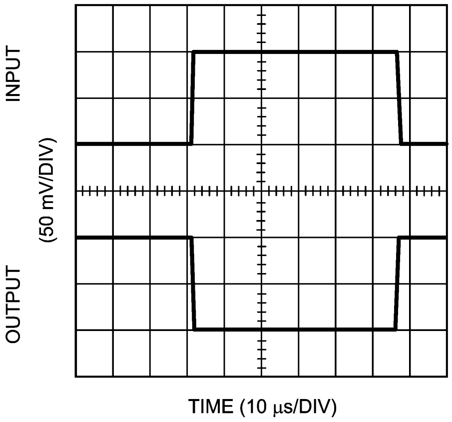
Inverting Small Signal Pulse Response



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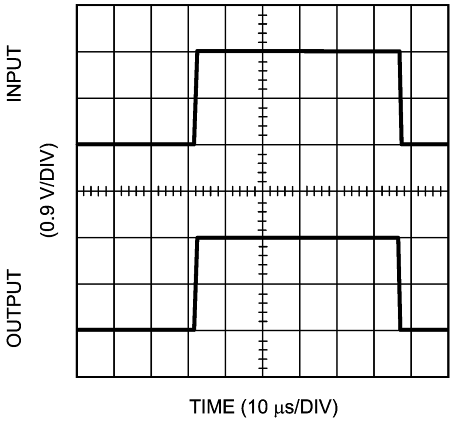
Non-Inverting Large Signal Pulse Response



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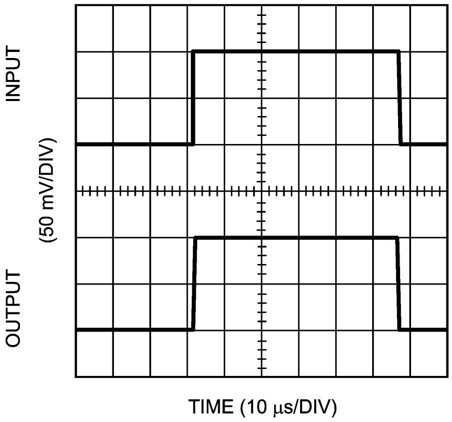
Non-Inverting Small Signal Pulse Response



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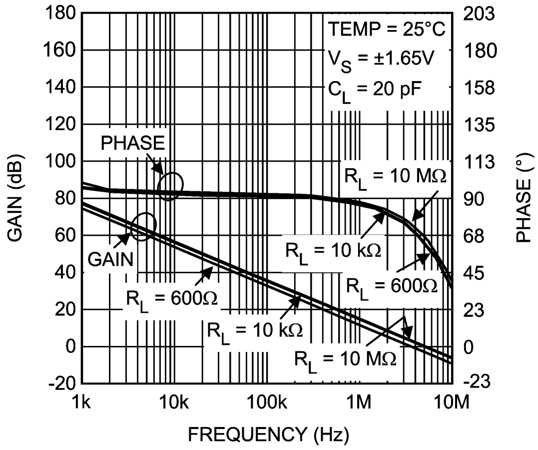
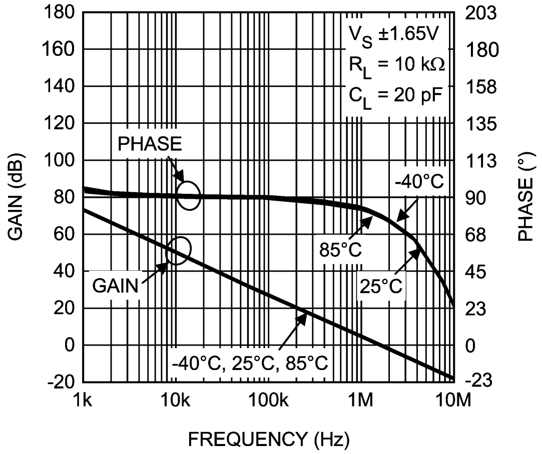
Typical Performance Characteristics Unless otherwise specified, V+ 3.3V, T_J = 25˚C. (Continued)


Open Loop Frequency vs. R_L

20179521


Open Loop Frequency Response over Temperature

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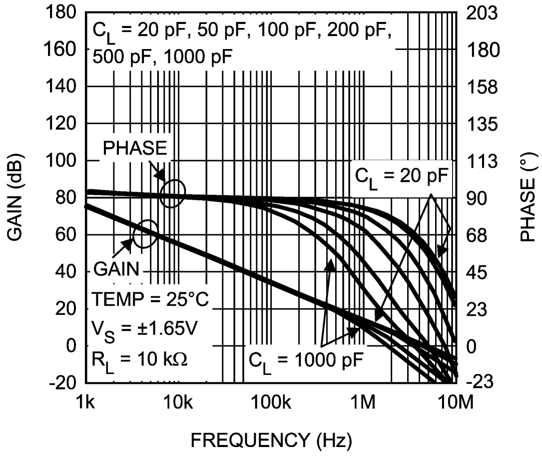
Open Loop Frequency Response vs. C_L

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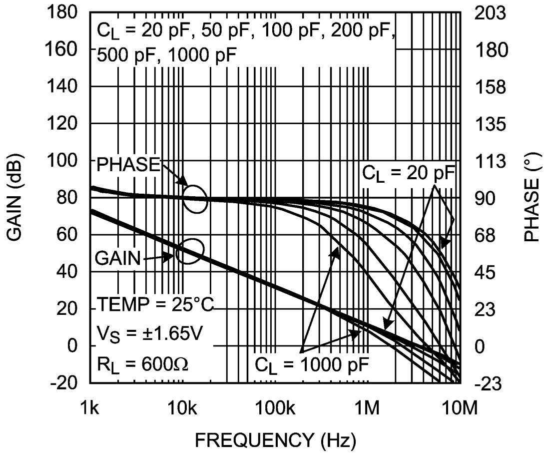
Open Loop Frequency Response vs. C_L

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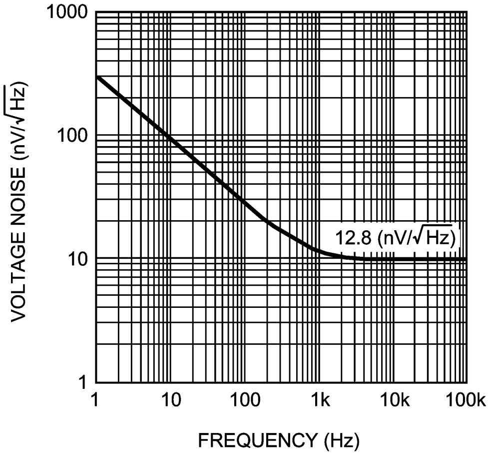
Voltage Noise vs. Frequency

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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.

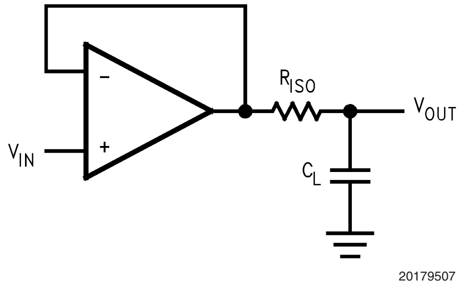


FIGURE 1. Indirectly Driving a Capacitive Load using Resistive Isolation

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.

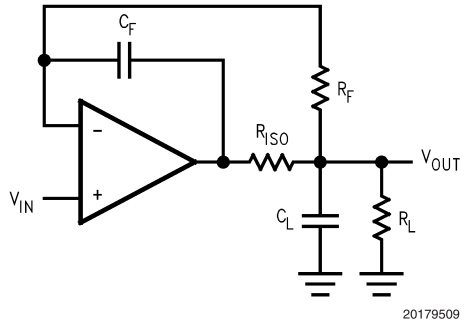
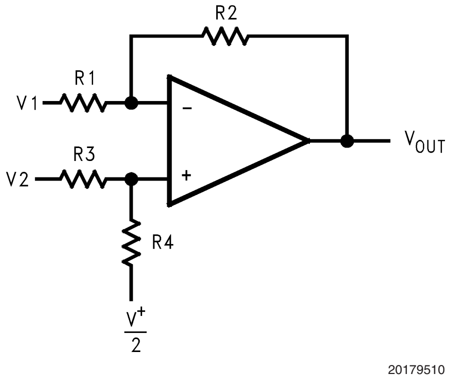


FIGURE 2. Indirectly Driving a Capacitive Load with DC Accuracy

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

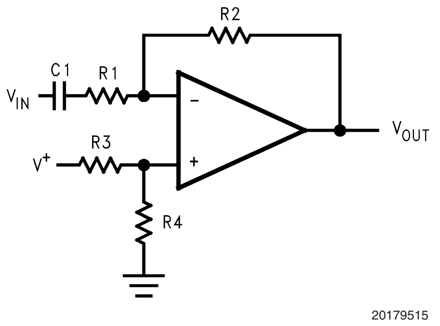
$$V_{OUT} = \frac{R2}{R1}\left(V_2 - V_1\right) + \frac{V^+}{2}$$

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, $fc = \frac{1}{2}\pi R_1C_1$. 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.



$$V_{OUT} = -\frac{R2}{R1} V_{IN}$$

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*.

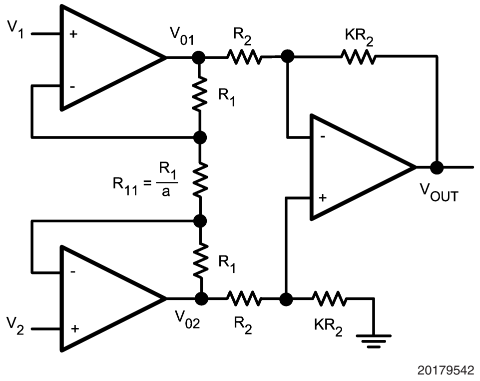


FIGURE 5. Three-Op-Amp Instrumentation Amplifier

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:

$$\text{GIVEN: } I_{R_1} = I_{R_{11}} \tag{1}$$

By Ohm's Law:

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

However:

$$V_{R_{11}} = V_1 - V_2 \tag{3}$$

So we have:

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

Application Information (Continued)

Now looking at the output of the instrumentation amplifier:

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

Substituting from *Equation (4)*:

$$V_O = -K(2a+1)(V_1 - V_2) \tag{6}$$

This shows the gain of the instrumentation amplifier to be:

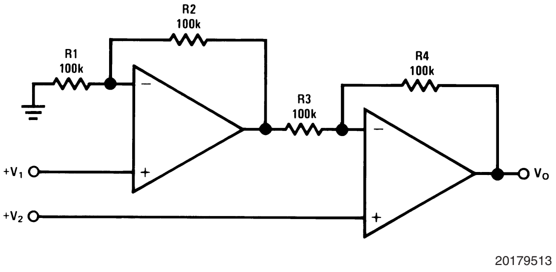
$$-K(2a+1)$$

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 kΩ, R_{11} = 1.8 kΩ, R_2 = 2.5 kΩ 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.



$$V_O = \left(1 + \frac{R4}{R3}\right)(V_2 - V_1),\ \text{where } R1 = R4 \text{ and } R2 = R3$$
$$\text{As shown: } V_O = 2(V_2 - V_1)$$

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.

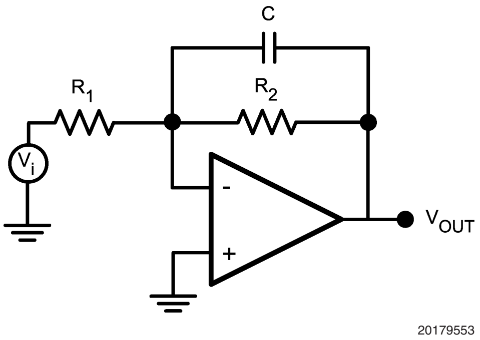


FIGURE 7. Low Pass Filter

The transfer function can be expressed as follows:

By KCL:

$$\frac{-V_i}{R_1} - \frac{V_O}{\left[\frac{1}{jwc}\right]} - \frac{V_O}{R_2} = O \tag{7}$$

Simplifying this further results in:

$$V_O = \frac{-R_2}{R_1}\left[\frac{1}{jwcR_2 + 1}\right]V_i \tag{8}$$

or

$$\frac{V_O}{V_i} = \frac{-R_2}{R_1}\left[\frac{1}{jwcR_2 + 1}\right] \tag{9}$$

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

$$H = H_O\left[\frac{1}{j2\pi fcR_2 + 1}\right] \tag{10}$$

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

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

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 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)

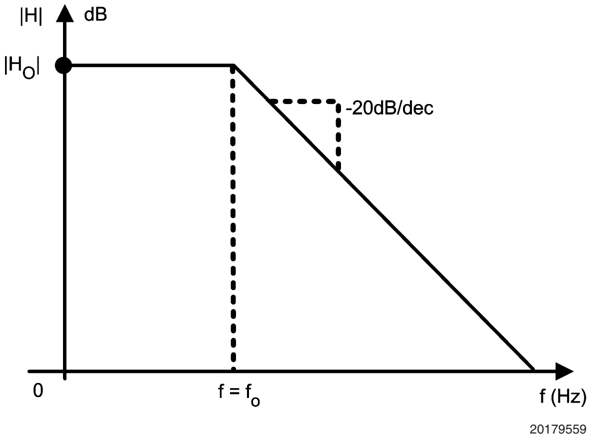


FIGURE 8. Low Pass Filter Transfer Function

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:

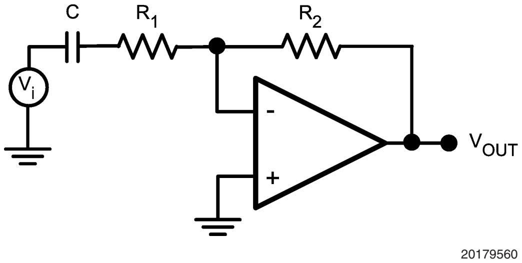


FIGURE 9. High Pass Filter

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

$$\frac{V_1 - V_i}{\frac{1}{jwC}} = \frac{V_1 - V^-}{R_1} \tag{12}$$

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

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.

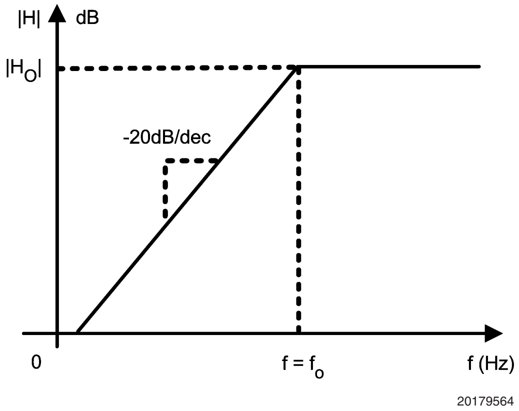


FIGURE 10. High Pass Filter Transfer Function

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.

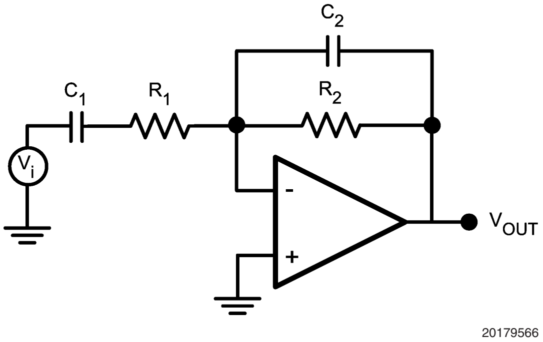


FIGURE 11. Band Pass Filter

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, $f_1 \le f \le f_2$, 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

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

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

$$H_O = \frac{-R_2}{R_1}$$

(16)

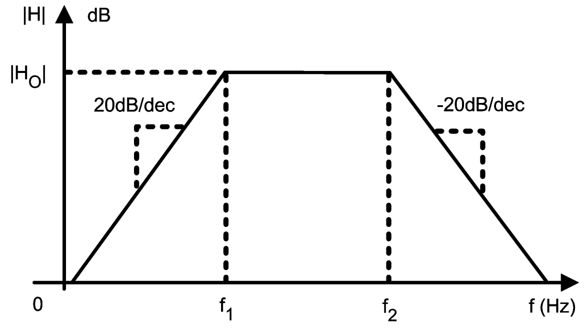


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FIGURE 12. Band Pass Filter Transfer Function

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.

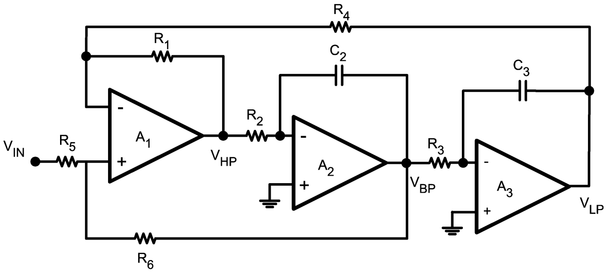

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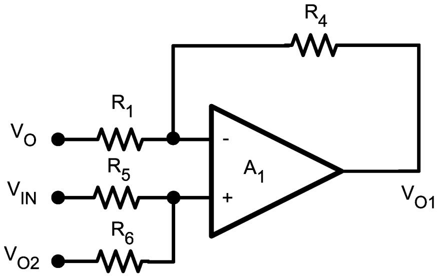
FIGURE 13. State Variable Active Filter

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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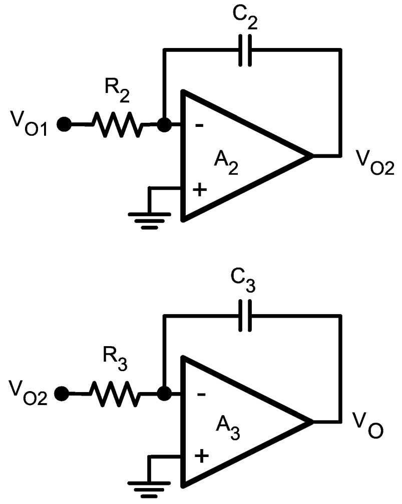

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For A_1 the relationship between input and output is:

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

(17)

This relationship depends on the output of all the filters. The input-output relationship for A_2 can be expressed as:

$$V_{O2} = \frac{-1}{s\,C_2 R_2}\; V_{O1}$$

(18)

And finally this relationship for A_3 is as follows:

$$V_O = \frac{-1}{s\,C_3 R_3}\; V_{O2}$$

(19)

Re-arranging these equations, one can find the relationship between V_O and V_{IN} (transfer function of the low pass filter), V_{O1} and V_{IN} (transfer function of the high pass filter), and V_{O2} and V_{IN} (transfer function of the band pass filter) These relationships are as follows:

Application Information (Continued)

Low Pass Filter

$$\frac{V_O}{V_{IN}} = \frac{\left[\frac{R_1 + R_4}{R_1}\right]\left[\frac{R_6}{R_5 + R_6}\right]\left[\frac{1}{C_2C_3R_2R_3}\right]}{s^2 + s\left[\frac{1}{C_2R_2}\right]\left[\frac{R_5}{R_5 + R_6}\right]\left[\frac{R_1 + R_4}{R_1}\right] + \left[\frac{1}{C_2C_3R_2R_3}\right]} \tag{20}$$

High Pass Filter

$$\frac{V_{O1}}{V_{IN}} = \frac{s^2\left[\frac{R_1 + R_4}{R_1}\right]\left[\frac{R_6}{R_5 + R_6}\right]}{s^2 + s\left[\frac{1}{C_2R_2}\right]\left[\frac{R_5}{R_5 + R_6}\right]\left[\frac{R_1 + R_4}{R_1}\right] + \left[\frac{1}{C_2C_3R_2R_3}\right]} \tag{21}$$

Band Pass Filter

$$\frac{V_{O2}}{V_{IN}} = \frac{s\left[\frac{1}{C_2R_2}\right]\left[\frac{R_1 + R_4}{R_1}\right]\left[\frac{R_6}{R_5 + R_6}\right]}{s^2 + s\left[\frac{1}{C_2R_2}\right]\left[\frac{R_5}{R_5 + R_6}\right]\left[\frac{R_1 + R_4}{R_1}\right] + \left[\frac{1}{C_2C_3R_2R_3}\right]} \tag{22}$$

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

$$\omega_c = \sqrt{\frac{1}{C_2C_3R_2R_3}}$$

and

$$Q = \sqrt{\frac{C_2R_2}{C_3R_3}}\left[\frac{R_5 + R_6}{R_6}\right]\left[\frac{R_1}{R_1 + R_4}\right] \tag{23}$$

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 $R_1 = R_4 = 30\ \text{k}\Omega$. Now values of R_5 and R_6 need to be calculated. With the chosen values for the capacitors and resistors, Q reduces to:

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

or

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

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 $R_2 = R_3 = 16\ \text{k}\Omega$.

The DC gain of this circuit is:

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

TRANSIMPEDANCE AMPLIFIER

CMOS input op amps are often used in transimpedance applications as they have an extremely high input impedance. A transimpedance amplifier converts a small input current into a voltage. This current is usually generated by a photodiode. The transimpedance gain, measured as the ratio of the output voltage to the input current, is expected to be large and wide-band. Since the circuit deals with currents in the range of a few nA, low noise performance is essential. LMV716 is a CMOS input op amps which provides wide bandwidth and low noise performance, and is therefore, ideal for transimpedance applications.

Usually, a transimpedance amplifier is designed on the basis of the current source driving the input. A photodiode is a very common capacitive current source, which requires transimpedance gain to transform its miniscule current into easily detectable voltages. The photodiode and amplifier's gain are selected with respect to the speed and accuracy required of the circuit. A faster circuit would require a photo-diode with lesser capacitance and a faster amplifier. A more sensitive circuit would require a sensitive photodiode and a high gain. A typical transimpedance amplifier is shown in *Figure 14*. The output voltage of the amplifier is given by the equation $V_{OUT} = -I_{IN}R_F$. Since the output swing of the amplifier is limited, R_F should be selected such that all possible values of I can be detected.

The LMV716 has a large gain-bandwidth product (5 MHz), which enables high gains at wide bandwidths. A rail-to-rail output swing at 5.0V supply allows detection and amplification of a wide range of input currents. A CMOS input stage with negligible input current noise and low input voltage noise allows the LMV716 to provide high fidelity amplification for wide bandwidths. These properties make the LMV716 ideal for systems requiring wide-band transimpedance amplification.

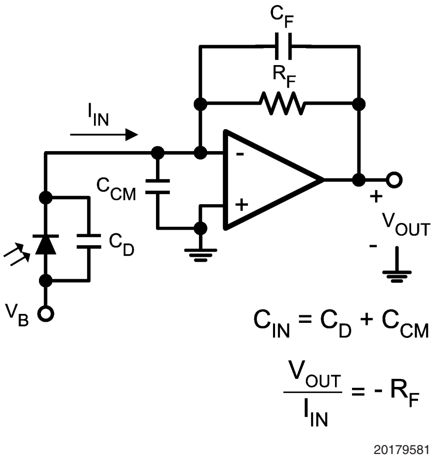


FIGURE 14. Photodiode Transimpedance Amplifier

Application Information (Continued)

As mentioned earlier, the following parameters are used to design a transimpedance amplifier: the amplifier gain bandwidth product, A_0; the amplifier input capacitance, CCM; the photodiode capacitance, CD; the transimpedance gain required, R_F; and the amplifier output swing. Once a feasible R_F is selected using the amplifier output swing, these numbers can be used to design an amplifier with the desired transimpedance gain and a maximally flat frequency response.

An essential component for obtaining a maximally flat response is the feedback capacitor, C_F. The capacitance seen at the input of the amplifier, C_{IN}, combined with the feedback capacitor, R_F, generate a phase lag which causes gain peaking and can destabilize the circuit. C_{IN} is usually just the sum of C_D and C_{CM}. The feedback capacitor C_F creates a pole, f_P in the noise gain of the circuit, which neutralizes the zero in the noise gain, f_Z, created by the combination of R_F and C_{IN}. If properly positioned, the noise gain pole created by C_F can ensure that the slope of the gain remains at 20 dB/decade till the unity gain frequency of the amplifier is reached, thus ensuring stability. As shown in *Figure 15*, f_P is positioned such that it coincides with the point where the noise gain intersects the op amp's open loop gain. In this case, f_P is also the corner frequency of the transimpedance amplifier. The value of C_F needed to make it so is given by *Equation (27)*. A larger value of C_F causes excessive reduction of bandwidth, while a smaller value fails to prevent gain peaking and instability.

$$C_F = \frac{1 + \sqrt{1 + 4\pi R_F C_{IN} A_0}}{2\pi R_F A_0} \qquad (27)$$

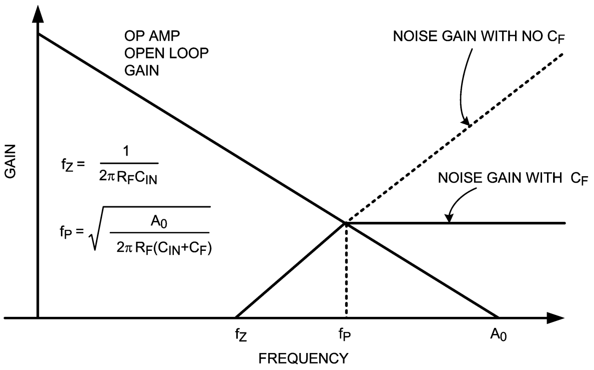


FIGURE 15. C_F Selection for Stability

Calculating C_F from *Equation (27)* can sometimes return unreasonably small values (<1 pF), especially for high speed applications. In these cases, it is often more practical to use a circuit such as the one shown in *Figure 16* in order to allow more reasonable values. In this circuit, the capacitance C_F' is (1+ R_B/R_A) multiplied by the effective feedback capacitance, C_F. A larger capacitor can now be used in this circuit to obtain a smaller effective capacitance.

For example, if a C_F of 0.5 pF is needed, while only a 5 pF capacitor is available, R_B and R_A can be selected such that $R_B/R_A = 9$. This would convert a C_F' of 5 pF into a C_F of 0.5 pF. This relationship holds as long as $R_A << R_F$.

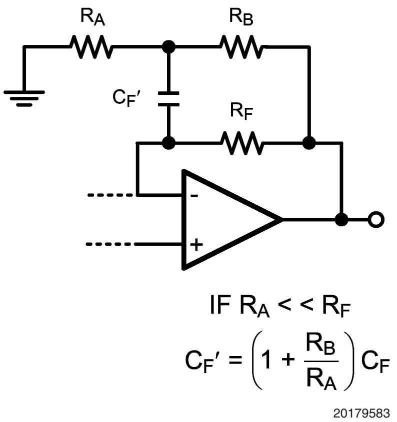
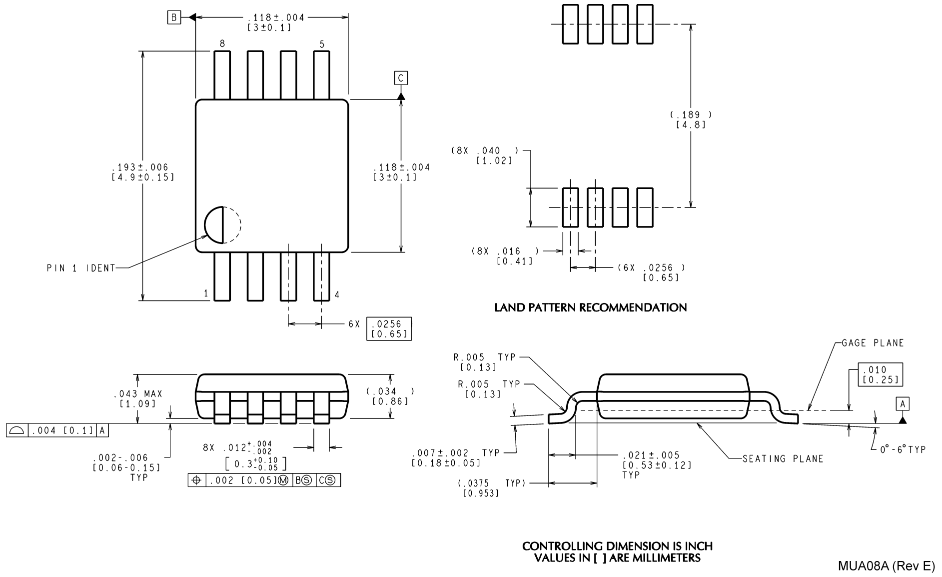


FIGURE 16. Obtaining Small C_F from Large C_F'

Physical Dimensions inches (millimeters) unless otherwise noted



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