Chapter V

APPLYING IN-AMPS EFFECTIVELY

Dual-Supply Operation

The conventional way to power an in-amp has been from a *split* or dual polarity power supply. This has the obvious advantage of allowing both a positive and a negative input and output swing.

Single-Supply Operation

Single-supply operation has become an increasingly desirable characteristic of a modern in-amp. Many present day data acquisition systems are powered from a single low voltage supply. For these systems, there are two vitally important characteristics. First, the in-amp's input range should extend between the positive supply and the negative supply (or ground). Second, the amplifier's output should be rail-to-rail as well, providing an output swing to within 100 mV or less of either supply rail or ground. In contrast, a standard dual-supply in-amp can only swing to within a volt or two of either supply or ground. When operated from a 5 V single supply, these in-amps have only a volt or two of output voltage swing, while a true rail-to-rail amplifier can provide a peak-to-peak output nearly as great as the supply voltage. Another important point is that a single-supply, or rail-to-rail in-amp, will still operate well (or even better) from a dual supply, and it will often operate at lower power than a conventional dual-supply device.

Power Supply Bypassing, Decoupling, and Stability Issues

Power supply decoupling is an important detail that is often overlooked by designers. Normally, bypass capacitors (values of 0.1 μF are typical) are connected between the power supply pins of each IC and ground. Although usually adequate, this practice can be ineffective or even create worse transients than no bypassing at all. It is important to consider where the circuit's currents originate, where they will return, and by what path. Once that has been established, bypass these currents around ground and other signal paths.

In general, like op amps, most monolithic in-amps have their integrators referenced to one or both power supply lines and should be decoupled with respect to the output reference terminal. This means that for each chip a bypass capacitor should be connected between each power supply pin and the point on the board where the in-amp's reference terminal is connected, as shown in Figure 5-1.

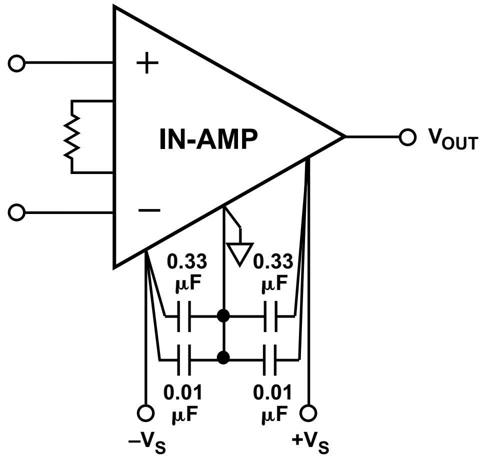


Figure 5-1. Recommended Method for Power Supply Bypassing

For a much more comprehensive discussion of these issues, refer to application note AN-202 "An IC Amplifier Users' Guide to Decoupling, Grounding, and Making Things Go Right for a Change," by Paul Brokaw, on the ADI website at www.analog.com.

THE IMPORTANCE OF AN INPUT GROUND RETURN

One of the most common applications problems that arises when using in-amp circuits is failure to provide a dc return path for the in-amp's input bias currents. This usually happens when the in-amp's inputs are capacitively coupled. Figure 5-2 shows just such an arrangement. Here, the input bias currents quickly *charge up* capacitors C1 and C2 until the in-amp's output "rails", either to the supply or ground.

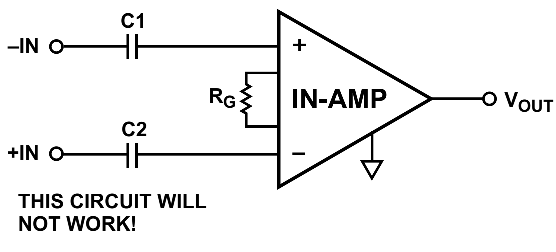


Figure 5-2. An AC-Coupled In-Amp Circuit Without an Input Ground Return

The solution is to add a high value resistance (R1, R2) between each input and ground, as shown in Figure 5-3.

The input bias currents can now flow freely to ground and do not build up a large input offset as before. In the vacuum tube circuits of years past, a similar effect occurred, requiring a *grid leak* resistance between the grid (input) and ground to drain off the accumulated charge (the electrons on the grid).

AC Input Coupling

Referring again to Figure 5-3, practical values for R1 and R2 are typically 1 MΩ or less. The choice of resistor value is a trade-off between offset errors and capacitance value. The larger the input resistor, the greater the input offset voltage due to input offset currents. Offset voltage drift will also increase.

With lower resistor values, higher value input capacitors must be used for C1 and C2 to provide the same –3 dB corner frequency

$$F_{-3\ dB} = (1/(2\pi R1C1))\ \textit{where}\ R1 = R2\ \textit{and}\ C1 = C2$$

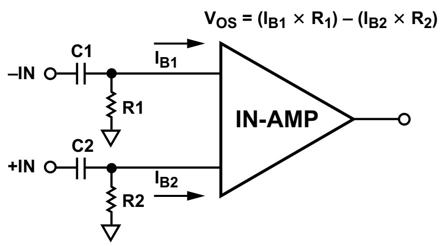


Figure 5-3. A High Value Resistor between Each Input and Ground Provides an Effective DC Return Path

Unless there is a large dc voltage present on the input side of the ac coupling capacitor, nonpolarized capacitors should be used. Therefore, in the interest of keeping component size as small as possible, C_1 and C_2 should be 0.1 μF or less. Generally, the smaller the capacitor value the better, because it will cost less and be smaller in size. The voltage rating of the input coupling capacitor needs to be high enough to avoid breakdown from any high voltage input transients that might occur.

RC Component Matching

Since

$$(I_{B1}\ R1) - (I_{B2}\ R2) = \Delta V_{OS}$$

any mismatch between R1 and R2 will cause an input offset imbalance (I_{B1}–I_{B2}) which will create an input offset voltage error.

A good guideline is to keep I_B R < 10 mV.

The input bias currents of Analog Devices in-amps vary widely, depending on input architecture. However, the vast majority have maximum input bias currents between 1.5 nA and 10 nA. Table 5-1 gives typical R and C cookbook values for ac coupling using 1% metal film resistors and two values of input bias current.

Figure 5-4 shows the recommended dc return for a transformer-coupled input.

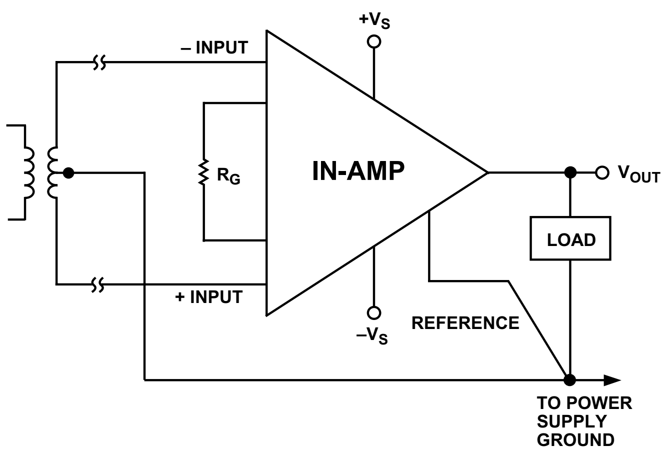


Figure 5-4. Recommended DC Return Path for a Transformer-Coupled Input

Table 5-1. Recommended Component Values for AC Coupling In-Amp Inputs

–3 dB BW	RC Coupling Components C1, C2	RC Coupling Components R1, R2	Input Bias Current (IB)	V_{OS} at Each Input	V_{OS} Error for 2% R1, R2 Mismatch
2 Hz	0.1 μF	1 MΩ	2 nA	2 mV	40 μV
2 Hz	0.1 μF	1 MΩ	10 nA	10 mV	200 μV
30 Hz	0.047 μF	115 kΩ	2 nA	230 μV	5 μV
30 Hz	0.1 μF	53.6 kΩ	10 nA	536 μV	11 μV
100 Hz	0.01 μF	162 kΩ	2 nA	324 μV	7 μV
100 Hz	0.01 μF	162 kΩ	10 nA	1.6 mV	32 μV
500 Hz	0.002 μF	162 kΩ	2 nA	324 μV	7 μV
500 Hz	0.002 μF	162 kΩ	2 nA	324 μV	7 μV

CABLE TERMINATION

When in-amps are used at frequencies above a few hundred kilohertz, properly terminated 50 Ω or 75 Ω coaxial cable should be used for input and output connections. Normally, cable termination is simply a 50 Ω or 75 Ω resistor connected between the cable center conductor and its shield at the end of the coaxial cable. Note that a buffer amplifier may be required to drive these loads to useful levels.

INPUT PROTECTION BASICS FOR ADI IN-AMPS

Input Protection from ESD and DC Overload

As interface amplifiers for data acquisition systems, instrumentation amplifiers are often subjected to input overloads, i.e., voltage levels in excess of their full scale for the selected gain range or even in excess of the supply voltage. These overloads fall into two general classes: steady state and transient (ESD, etc.), which occur for only a fraction of a second. With 3-op amp in-amp designs, when operating at low gains (10 or less), the gain resistor acts as a current-limiting element in series with their resistor inputs. At high gains, the lower value of R_G may not adequately protect the inputs from excessive currents.

Standard practice is to place current-limiting resistors in each input, but adding this protection also increases the circuit's noise level. A reasonable balance needs to be found between the protection provided and the increased resistor (Johnson) noise introduced. Circuits using in-amps with a relatively high noise level can tolerate more series protection without seriously increasing their total circuit noise.

Of course, the less added noise the better, but a good guideline is that circuits needing this extra protection can easily tolerate resistor values that generate 30% of the total circuit noise. For example, a circuit using an in-amp with a rated noise level of 20 nV/$\sqrt{\text{Hz}}$ can tolerate an additional 6 nV/$\sqrt{\text{Hz}}$ of Johnson noise.

Use the following cookbook method to translate this number into a practical resistance value. The Johnson noise of a 1 kΩ resistor is approximately 4 nV/$\sqrt{\text{Hz}}$. This value varies as the square root of the resistance. So, a 20 kΩ resistor would have $\sqrt{20}$ times as much noise as the 1 kΩ resistor, which is 17.88 nV/$\sqrt{\text{Hz}}$ (4.4721 times 4 nV/$\sqrt{\text{Hz}}$). Because *both* inputs need to be protected, two resistors are needed and their combined noise will add as the square root of the number of resistors (the root sum of squares value). In this case, the total added noise from the two 20 kΩ resistors will be 25.3 nV/$\sqrt{\text{Hz}}$ (17.88 times 1.414).

Figure 5-5 provides details on the input architecture of the **AD8221** in-amp. As shown, it has internal 400 Ω resistors that are in series with each input transistor junction.

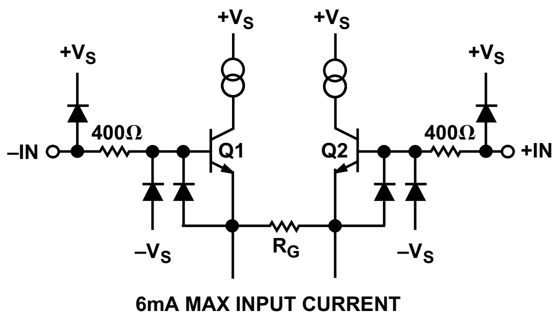


Figure 5-5. AD8221 In-Amp Input Circuit

The AD8221 was designed to handle maximum input currents of 6 mA steady state (or dc). Its internal resistors and diodes will protect the device from input voltages 0.7 V above the positive supply, or 2.4 V more negative than the minus supply (6 mA × 0.4 kΩ). Therefore, for ±15 V supplies, the maximum safe input level is +15.7 V, –17.4 V. Additional external series resistors can be added to increase this level considerably, at the expense of higher circuit noise level.

The AD8221 in-amp is a very low noise device, with a maximum (e_{NI}) 8 nV/$\sqrt{Hz}$. A single 1 kΩ resistor will add approximately 107 nV/$\sqrt{Hz}$ of noise. This would raise the maximum dc level to approximately 22.5 V above each supply or ±37.5 V with 15 V supplies.

Figure 5-6 shows the input section of the AD620 in-amp. This is very similar to that of the AD8221: both use a 400 Ω resistor in series with each input, and both use diode protection. The chief differences are the four additional AD8221 diodes. One set is tied between each input and the positive supply, and the other set is connected between the base of each input transistor and the negative supply. The AD620 uses its 400 Ω internal resistor and a single set of diodes to protect against negative input voltages. For positive voltage overloads, it relies on its own base-emitter input junction to act as the clamping diode.

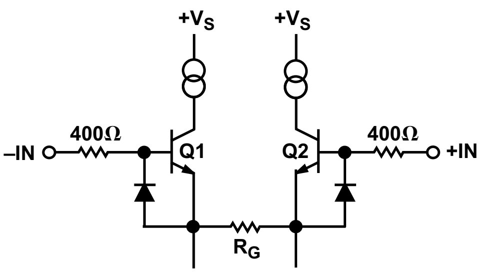


Figure 5-6. AD620 Series (AD620, AD621, AD622) In-Amp Input Circuit

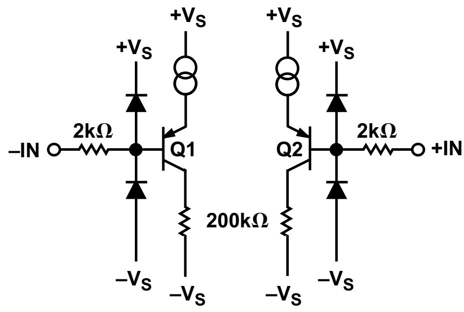


Figure 5-7. AD627 In-Amp Input Circuit

The AD627 can tolerate 20 mA transient input currents (Figure 5-7). In addition, it has built-in 2 kΩ resistors and can handle input voltages 40 V higher than its supply lines (20 mA times 2 kΩ). This level of protection is quite beneficial. Because of its low power, many of the AD627's applications will use a low voltage single power supply. If even more protection is needed, quite large external resistors can be added without seriously degrading the AD627's 38 nV/$\sqrt{Hz}$ noise level. In this case, adding two 5 kΩ resistors will raise the circuit's noise approximately 13 nV/$\sqrt{Hz}$ (30 percent), but would provide an additional ±100 V of transient overload protection.

Figure 5-8 shows the input architecture of the AD623 in-amp. In this design, the internal (ESD) diodes are located *before* the input resistors, and as a consequence provide less protection than the other designs. The AD623 can tolerate 10 mA maximum input current, but in many cases, some external series resistance will be needed to keep input current below this level.

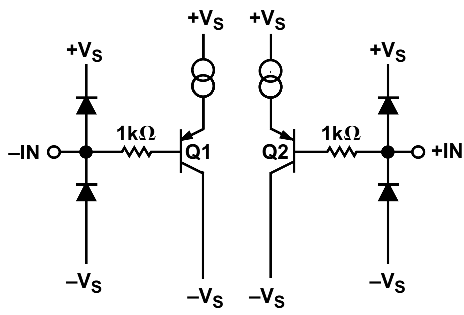


Figure 5-8. AD623 In-Amp Input Circuit

Since the AD623's device noise is approximately 35 nV/$\sqrt{Hz}$, up to 5 kΩ of external resistance can be added here to provide 50 V of dc overload protection, while only increasing input noise to 38 nV/$\sqrt{Hz}$ total.

Table 5-2 provides recommended series protection resistor values for a 10% or 40% increase in circuit noise.

Table 5-2. Recommended Series Protection Resistor Values

Device	In-Amp Noise (e_{NI})	Max Input Overload Current	Recommended External Protection Resistors Adding Additional Noise* of 10%	of 40%
AD8221	8 nV/√Hz	6 (mA)	340 Ω	2.43 kΩ
AD8225	8 nV/√Hz	6 (mA)	340 Ω	2.43 kΩ
AD620	9 nV/√Hz	6 (mA)	348 Ω	2.49 kΩ
AD621	9 nV/√Hz	6 (mA)	348 Ω	2.49 kΩ
AD622	9 nV/√Hz	6 (mA)	348 Ω	2.49 kΩ
AD623	35 nV/√Hz	10 (mA)	8.08 kΩ	40.2 kΩ
AD627	38 nV/√Hz	20 (mA)	8.87 kΩ	43.2 kΩ

*This noise level is for two resistors, one in series with each resistor.

Adding External Protection Diodes

Device input protection may be increased with the addition of external clamping diodes as shown in Figure 5-9. As high current diodes, are used, input protection is increased, which allows the use of much lower resistance input protection resistors which, in turn, reduces the circuit's noise.

Unfortunately, most ordinary diodes (Schottky, silicon, etc.) have high leakage currents that will cause large offset errors at the in-amp's output; this leakage increases exponentially with temperature. This tends to rule out the use of external diodes in applications where the in-amp is used with high impedance sources.

Specialty diodes with much lower leakage are available, but these are often difficult to find and are expensive. For the vast majority of applications, limiting resistors alone provide adequate protection for ESD and longer duration input transients.

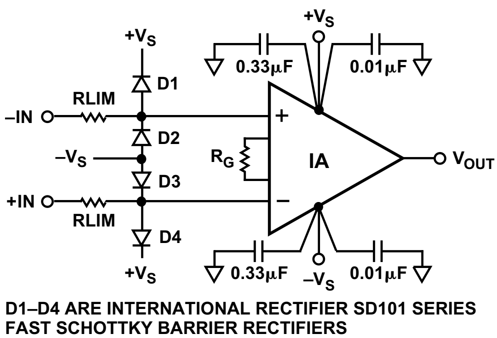


Figure 5-9. Using External Components to Increase Input Protection

Despite their limitations, external diodes are often required in some special applications, such as electric shock defibrillators, which utilize short duration, high voltage pulses. The combination of external diodes and very large input resistors (as high as 100 kΩ) may be needed to adequately protect the in-amp.

It is a good idea to check the diodes' specifications to ensure that their conduction begins well before the in-amp's internal protection diodes start drawing current. Although they provide excellent input protection, standard Schottky diodes can have leakage up to several mA. However, as in the example of Figure 5-9, fast Schottky barrier rectifiers, such as the international rectifier type SD101 series, can be used; these devices have 200 nA max leakage currents and 400 mW typical power dissipation.

ESD and Transient Overload Protection

Protecting in-amp inputs from high voltage transients and ESD events is very important for a circuit's long-term reliability. Power dissipation is often a critical factor as input resistors, whether internal or external, must be able to handle most of the power of the input pulse without failing.

ESD events, while they may be very high voltage, are usually of very short duration and are normally one-time events. Since the circuit has plenty of time to cool down before the next event occurs, modest input protection is sufficient to protect the device from damage.

On the other hand, regularly occurring short duration input transients can easily overheat and burn out the input resistors or the in-amps input stage. A 1 kΩ resistor, in series with an in-amp input terminal drawing 20 mA, will dissipate 0.4 W, which can easily be handled by a standard 0.5 W or greater surface-mount resistor. If the input current is doubled, power consumption quadruples as it increases as the square of the input current (or as the square of the applied voltage).

Although it is a simple matter to use a higher power protection resistor, this is a dangerous practice, as the power dissipation will also increase in the in-amp's input stage. This can easily lead to device failure (see the preceding section on input protection basics for input current limitations of ADI in-amps). Except for ESD events, it is always best to adopt a conservative approach and treat all transient input signals as full duration inputs.

Designs that are expected to survive such events over long periods of time must use resistors with enough resistance to protect the in-amp's input circuitry from failure and enough power to prevent resistor burnout.

DESIGN ISSUES AFFECTING DC ACCURACY

The modern in-amp is continually being improved, providing the user with ever-increasing accuracy and versatility at a lower price. Despite these improvements in product performance, there remain some fundamental applications issues that seriously affect device accuracy. Now that low cost, high resolution ADCs are commonly used, system designers need to ensure that if an in-amp is used as a preamplifier ahead of the converter, the in-amp's accuracy matches that of the ADC.

Designing for the Lowest Possible Offset Voltage Drift

Offset drift errors include not just those associated with the active device being used (IC in-amp or discrete in-amp design using op amps), but also thermocouple effects in the circuit's components or wiring. The in-amp's input bias and input offset currents flowing through unbalanced source impedances also create additional offset errors. In discrete op amp in-amp designs, these errors can increase with temperature unless precision op amps are used.

Designing for the Lowest Possible Gain Drift

When planning for gain errors, the effects of board layout, the circuit's thermal gradients, and the characteristics of any external gain setting resistors are often overlooked. A gain resistor's absolute tolerance, its thermal temperature coefficient, its physical position relative to other resistors in the same gain network, and even its physical orientation (vertical or horizontal) are all-important design considerations if high dc accuracy is needed.

In many ADC preamp circuits, an external user-selected resistor sets the gain of the in-amp, so the absolute tolerance of this resistor and its variation over temperature, compared to that of the IC's on-chip resistors, will affect the circuit's gain accuracy. Resistors commonly used include through-hole 1% 1/4 W metal film types and 1% 1/8 W chip resistors. Both types typically have a 100 ppm/°C temperature coefficient. However, some chip resistors can have TCs of 200 ppm/ °C or even 250 ppm/ °C.

Even when a 1% 100 ppm/°C resistor is used, the gain accuracy of the in-amp will be degraded. The resistor's initial room temperature accuracy is only ±1%, and the resistor will drift another 0.01% (100 ppm/°C) for every °C change in temperature. The initial gain error can easily be subtracted out in software, but to correct for the error versus temperature, frequent recalibrations (and a temperature sensor) would be required.

If the circuit is calibrated initially, the overall gain accuracy is reduced to approximately 10 bits (0.1%) accuracy for a 10°C change. An in-amp with a standard 1% metal film gain resistor should never be used ahead of even a 12-bit converter: it would destroy the accuracy of a 14-bit or 16-bit converter.

Additional error sources associated with external resistors also affect gain accuracy. The first are variations in resistor heating caused by input signal level. Figure 5-10, a simple op amp voltage amplifier, provides a practical example.

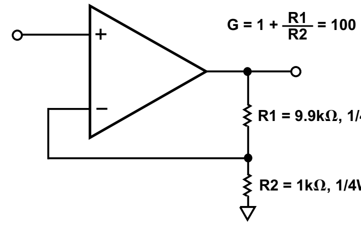


Figure 5-10. An Example of How Differences in Input Signal Level Can Introduce Gain Errors

Under zero signal conditions, there is no output signal and no resistor heating. When an input signal is applied, however, an amplified voltage appears at the op amp output. When the amplifier is operating with gain, Resistor R1 will be greater than R2. This means that there will be more voltage across R1 than across R2. The power dissipated in each resistor equals the square of the voltage across it divided by its resistance in ohms. The power dissipated and, therefore, the internal heating of the resistor will increase in proportion to the value of the resistor.

In the example, R1 is 9.9 kΩ and R2 is 1 kΩ. Consequently, R1 will dissipate 9.9 times more power than R2. This leads to a gain error that will vary with input level. The use of resistors with different temperature coefficients can also introduce gain errors.

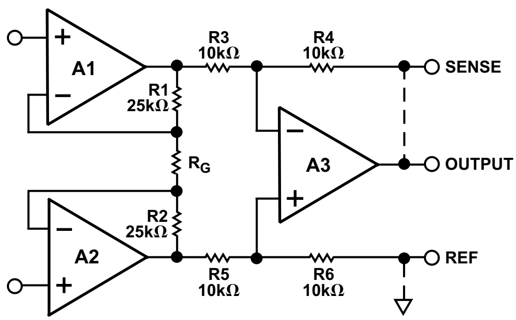


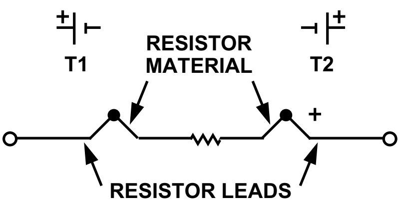
Figure 5-11. A Typical Discrete 3-Op Amp In-Amp Using Large Value, Low TC Feedback Resistors

Even when resistors with matched temperature coefficients (TC) are used, gain errors that vary with input signal level can still occur. The use of larger (i.e., higher power) resistors will reduce these effects, but accurate, low TC power resistors are expensive and hard to find.

When a discrete 3-op amp in-amp is used, as shown in Figure 5-11, these errors will be reduced. In a 3-op amp in-amp, there are two feedback resistors, R1 and R2, and one gain resistor, R_G. Since the in-amp uses two feedback resistors while the op amp uses only one, each of the in-amp's resistors only needs to dissipate half the power (for the same gain). Monolithic in-amps, such as the AD620, offer a further advantage by using relatively large value (25 kΩ) feedback resistors. For a given gain and output voltage, large feedback resistors will dissipate less power (i.e., $P = V^2/R_F$). Of course, a discrete in-amp can be designed to use large value, low TC resistors as well, but with added cost and complexity.

Another less serious but still significant error source is the so-called thermocouple effect, sometimes referred to as thermal EMF. This occurs when two different conductors, such as copper and metal film, are tied together. When this bimetallic junction is heated, a simple thermocouple is created. When using similar metals such as a copper-to-copper junction, a thermoelectric error voltage of up to 0.2 mV/°C may be produced. An example of these effects is shown in Figure 5-12.

A final error source occurs when there is a thermal gradient across the external gain resistor. Something as simple as mounting a resistor on end to conserve board space will invariably produce a temperature gradient across the resistor. Placing the resistor flat down against the PC board will cure this problem unless there is air flowing along the axis of the resistor (where the air flow cools one side of the resistor more than the other side). Orienting the resistor so that its axis is perpendicular to the airflow will minimize this effect.



TYPICAL RESISTOR THERMOCOUPLE EMFs

• CARBON COMPOSITION	400μV/°C
• METAL FILM	20μV/°C
• EVENOHM OR MANGANIN WIRE-WOUND	2μV/°C
• RCD COMPONENTS HP-SERIES	0.05μV/°C

Figure 5-12. Thermocouple Effects Inside Discrete Resistors

Practical Solutions

As outlined, a number of dc offset and gain errors are introduced when external resistors are used with a monolithic in-amp. Discrete designs tend to have even larger errors. There are three practical solutions to this problem: use higher quality resistors, use software correction, or, better still, use an in-amp that has *all* of its gain resistors on-chip, such as the AD621.

Option 1: Use a Better Quality Gain Resistor

As a general rule, only 12-bit or 13-bit gain performance is possible using commonly available 1% resistors, which assumes that some type of initial calibration is performed.

A practical solution to this problem is to simply use a better quality resistor. A significant improvement can be made using a 0.1% 1/10 W surface-mount resistor. Aside from having a 10× better initial accuracy, it typically has a TC of only 25 ppm/°C, which will provide better than 13-bit accuracy over a 10°C temperature range.

If even better gain accuracy is needed, there are specialty houses that sell resistors with lower TCs, but these are usually expensive military varieties.

Option 2: Use a Fixed-Gain In-Amp

By far, the best overall dc performance is provided by using a monolithic in-amp, such as the AD621 or AD8225, in which all the resistors are contained within the IC. Now all resistors have identical TCs, all are at virtually the same temperature, and any thermal gradients across the chip are very small, and gain error drift is guaranteed and specified to very high standards.

At a gain of 10, the AD621 has a guaranteed maximum dc offset shift of less than 2.5 μV/°C and a maximum gain drift of ±5 ppm/°C, which is only 0.0005 %/°C.

The AD8225 is an in-amp with a fixed gain of 5. It has a maximum offset shift of 2 μV/°C and a maximum drift of 0.3 μV/°C.

RTI and RTO Errors

Another important design consideration is how circuit gain affects many in-amp error sources such as dc offset and noise. An in-amp should be regarded as a two stage amplifier with both an input and an output section. Each section has its own error sources.

Because the errors of the output section are multiplied by a fixed gain (usually 2), this section is often the principal error source at low circuit gains. When the in-amp is operating at higher gains, the gain of the input stage is increased. As the gain is raised, errors contributed by the input section are multiplied, while output errors are not. So, at high gains, the input stage errors dominate.

Since device specifications on different data sheets often refer to different types of errors, it is very easy for the unwary designer to make an inaccurate comparison between products. Any (or several) of four basic error categories may be listed: input errors, outputs errors, total error RTI, and total error RTO. Here follows an attempt to list, and hopefully simplify, an otherwise complicated set of definitions.

Input errors are those contributed by the amplifier's input stage alone; output errors are those due to the output section. Input related specifications are often combined and classified together as a referred to input (RTI) error, while all output related specifications are considered referred to output (RTO) errors.

For a given gain, an in-amp's input and output errors can be calculated using the following formulas:

$$\textit{Total Error, RTI} = \textit{Input Error} + (\textit{Output Error}/\textit{Gain})$$

$$\textit{Total Error, RTO} = (\textit{Gain} \times \textit{Input Error}) + \textit{Output Error}$$

Sometimes the spec page will list an error term as RTI or RTO for a specified gain. In other cases, it is up to the user to calculate the error for the desired gain.

Offset Error

Using the **AD620A** as an example, the total voltage offset error of this in-amp when operating at a gain of 10 can be calculated using the individual errors listed on its specifications page. The (typical) input offset of the AD620 (V_{OSI}) is listed as 30 μV. Its output offset (V_{OSO}) is listed as 400 μV. Thus, the total voltage offset referred to input, RTI, is equal to

$$\textit{Total RTI Error} = V_{OSI} + (V_{OSO}/G) = 30\ \mu V + (400\ \mu V/10) = 30\ \mu V + 40\ \mu V = 70\ \mu V$$

The total voltage offset referred to the output, RTO, is equal to

$$\textit{Total Offset Error RTO} = (G\ (V_{OSI})) + V_{OSO} = (10\ (30\ \mu V)) + 400\ \mu V = 700\ \mu V$$

Note that the two error numbers (RTI versus RTO) are 10× in value and logically they should be, as at a gain of 10, the error at the output of the in-amp should be 10 times the error at its input.

Noise Errors

In-amp noise errors also need to be considered in a similar way. Since the output section of a typical 3-op amp in-amp operates at unity gain, the noise contribution from the output stage is usually very small. But there are 3-op amp in-amps that operate the output stage at higher gains and 2-op amp in-amps regularly operate the second amplifier at gain. When either section is operated at gain, its noise is amplified along with the input signal.

Both RTI and RTO noise errors are calculated the same way as offset errors, except that the noise of two sections adds as the root mean square. That is

$$\textit{Input Noise} = eni, \textit{Output Noise} = eno$$

$$\textit{Total Noise RTI} = \sqrt{(eni)^2 + (eno/Gain)^2}$$

$$\textit{Total Noise RTO} = \sqrt{(Gain(eni))^2 + (eno)^2}$$

For example, the (typical) noise of the AD620A is specified as 9 nV/$\sqrt{\text{Hz}}$ eni and 72 nV/$\sqrt{\text{Hz}}$ eno. Therefore, the total RTI noise of the AD620A operating at a gain of 10 is equal to

$$\textit{Total Noise RTI} = \sqrt{(eni)^2 + (eno/Gain)^2} = \sqrt{(9)^2 + (72/10)^2} = 11.5\,\text{nV}/\sqrt{\text{Hz}}$$

REDUCING RFI RECTIFICATION ERRORS IN IN-AMP CIRCUITS

Real world applications must deal with an ever increasing amount of radio frequency interference (RFI). Of particular concern are situations in which signal transmission lines are long and signal strength is low. This is the classic application for an in-amp since its inherent common-mode rejection allows the device to extract weak differential signals riding on strong common-mode noise and interference.

One potential problem that is frequently overlooked, however, is that of radio frequency rectification inside the in-amp. When strong RF interference is present, it may become rectified by the IC and then appear as a dc output offset error. Common-mode signals present at an in-amp's input are normally greatly reduced by the amplifier's common-mode rejection.

Unfortunately, RF rectification occurs because even the best in-amps have virtually no common-mode rejection at frequencies above 20 kHz. A strong RF signal may become rectified by the amplifier's input stage and then appear as a dc offset error. Once rectified, no amount of low-pass filtering at the in-amp output will remove the error. If the RF interference is of an intermittent nature, this can lead to measurement errors that go undetected.

Designing Practical RFI Filters

The best practical solution is to provide RF attenuation *ahead* of the in-amp by using a differential low-pass filter. The filter needs to do three things: remove as much RF energy from the input lines as possible, preserve the ac signal balance between each line and ground (common), and maintain a high enough input impedance over the measurement bandwidth to avoid loading the signal source.

Figure 5-13 provides a basic building block for a wide number of differential RFI filters. Component values shown were selected for the AD8221, which has a typical –3 dB bandwidth of 1 MHz and a typical voltage noise level of 7 nV/$\sqrt{Hz}$. In addition to RFI suppression, the filter provides additional input overload protection, as resistors R1a and R1b help isolate the in-amp's input circuitry from the external signal source.

Figure 5-14 is a simplified version of the RFI circuit. It reveals that the filter forms a bridge circuit whose output appears across the in-amp's input pins. Because of this, any mismatch between the time constants of C1a/R1a and C1b/R1b will unbalance the bridge and reduce high frequency common-mode rejection. Therefore, resistors R1a and R1b and capacitors C1a and C1b should always be equal.

As shown, C2 is connected across the bridge output so that C2 is effectively in parallel with the series combination of C1a and C1b. Thus connected, C2 very effectively reduces any ac CMR errors due to mismatching. For example, if C2 is made 10 times larger than C1, this provides a 20× reduction in CMR errors due to C1a/C1b mismatch. Note that the filter does not affect dc CMR.

The RFI filter has two different bandwidths: differential and common mode. The differential bandwidth defines the frequency response of the filter with a differential input signal applied between the circuit's two inputs, +IN and –IN. This RC time constant is established by the sum of the two equal-value input resistors (R1a, R1b), together with the differential capacitance, which is C2 in parallel with the series combination of C1a and C1b.

The –3 dB *differential* bandwidth of this filter is equal to

$$BW_{DIFF} = \frac{1}{2\pi R(2C2 + C1)}$$

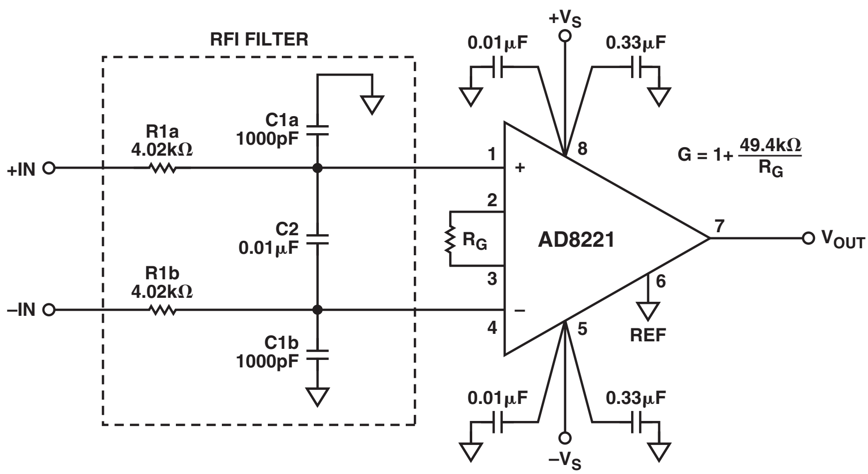


Figure 5-13. LP Filter Circuit Used to Prevent RFI Rectification Errors

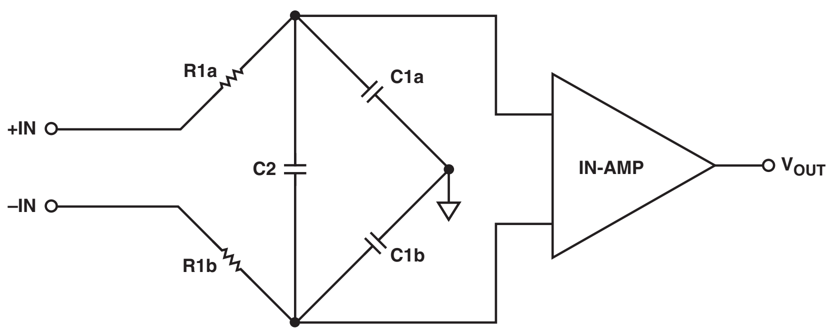


Figure 5-14. Capacitor C2 Shunts C1a/C1b and Very Effectively Reduces AC CMR Errors Due to Component Mismatching

The common-mode bandwidth defines what a common-mode RF signal *sees* between the two inputs tied together and ground. It's important to realize that C2 does not affect the bandwidth of the common-mode RF signal, as this capacitor is connected between the two inputs (helping to keep them at the same RF signal level). Therefore, common-mode bandwidth is set by the parallel impedance of the two RC networks (R1a/C1a and R1b/C1b) to ground.

The –3 dB *common-mode* bandwidth is equal to

$$BW_{CM} = \frac{1}{2\pi R1C1}$$

Using the circuit of Figure 5-13, with a C2 value of 0.01 µF as shown, the –3 dB differential signal bandwidth is approximately 1,900 Hz. When operating at a gain of 5, the circuit's measured dc offset shift over a frequency range of 10 Hz to 20 MHz was less than 6 µV RTI. At unity gain, there was no measurable dc offset shift.

The RFI filter should be built using a PC board with ground planes on both sides. All component leads should be made as short as possible. The input filter common should be connected to the amplifier common using the most direct path. Avoid building the filter and the in-amp circuits on separate boards or in separate enclosures, as this extra lead length can create a loop antenna. Instead, physically locate the filter right at the in-amp's input terminals. A further precaution is to use good quality resistors that are both noninductive and nonthermal (low TC). Resistors R1 and R2 can be common 1% metal film units. However, all three capacitors need to be reasonably high Q, low loss components. Capacitors C1a and C1b need to be ±5% tolerance devices to avoid degrading the circuit's common-mode rejection. The traditional 5% silver micas, miniature size micas, or the new Panasonic ±2% PPS film capacitors (Digi-key part # PS1H102G-ND) are recommended.

Selecting RFI Input Filter Component Values Using a Cookbook Approach

The following general rules will greatly ease the design of an RC input filter.

1. First, decide on the value of the two series resistors while ensuring that the previous circuitry can adequately drive this impedance. With typical values between 2 kΩ and 10 kΩ, these resistors should not contribute more noise than that of the in-amp itself. Using a pair of 2 kΩ resistors will add a Johnson noise of 8 nV/$\sqrt{\text{Hz}}$; this increases to 11 nV/$\sqrt{\text{Hz}}$ with 4 kΩ resistors and to 18 nV/$\sqrt{\text{Hz}}$ with 10 kΩ resistors.
2. Next, select an appropriate value for capacitor C2, which sets the filter's differential (signal) bandwidth. It's always best to set this as low as possible without attenuating the input signal. A differential bandwidth of 10 times the highest signal frequency is usually adequate.
3. Then select values for capacitors C1a and C1b, which set the common-mode bandwidth. For decent ac CMR, these should be 10% the value of C2 or less. The common-mode bandwidth should always be less than 10% of the in-amp's bandwidth at unity gain.

Specific Design Examples

An RFI Circuit for AD620 Series In-Amps

Figure 5-15 is a circuit for general-purpose in-amps such as the AD620 series, which have higher noise levels (12 nV/$\sqrt{\text{Hz}}$) and lower bandwidths than the AD8221.

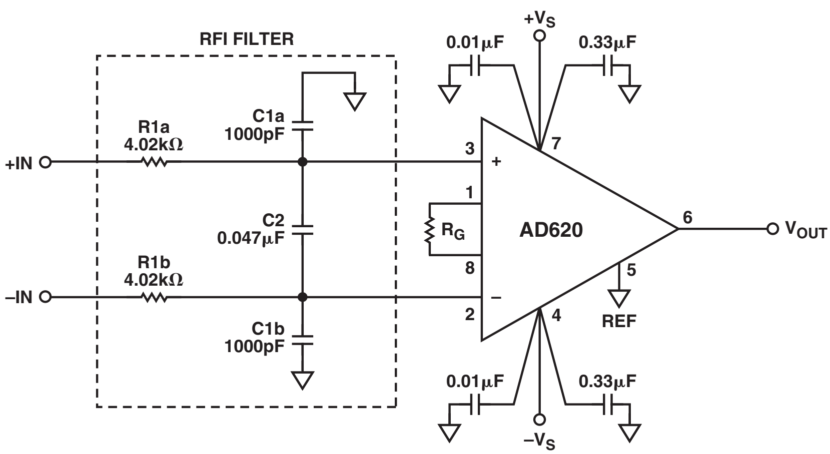


Figure 5-15. RFI Circuit for AD620 Series In-Amp

Accordingly, the same input resistors were used but capacitor C2 was increased approximately five times to 0.047 μF to provide adequate RF attenuation. With the values shown, the circuit's –3 dB bandwidth is approximately 400 Hz; the bandwidth may be increased to 760 Hz by reducing the resistance of R1 and R2 to 2.2 kΩ. Note that this increased bandwidth does not come free. It requires the circuitry preceding the in-amp to drive a lower impedance load and results in somewhat less input overload protection.

An RFI Circuit for Micropower In-Amps

Some in-amps are more prone to RF rectification than others and may need a more robust filter. A micropower in-amp, such as the AD627, with its low input stage operating current, is a good example. The simple expedient of increasing the value of the two input resistors, R1a/R1b, and/or that of capacitor C2, will provide further RF attenuation, at the expense of a reduced signal bandwidth.

Since the AD627 in-amp has higher noise (38 nV/$\sqrt{\text{Hz}}$) than general-purpose ICs such as the AD620 series devices, higher value input resistors can be used without seriously degrading the circuit's noise performance. The basic RC RFI circuit of Figure 5-13 was modified to include higher value input resistors, as shown in Figure 5-16.

The filter bandwidth is approximately 200 Hz. At a gain of 100, the maximum dc offset shift with a 1 V p-p input applied is approximately 400 μV RTI over an input range of 1 Hz to 20 MHz. At the same gain, the circuit's RF signal rejection (RF level at output/RF applied to the input) will be better than 61 dB.

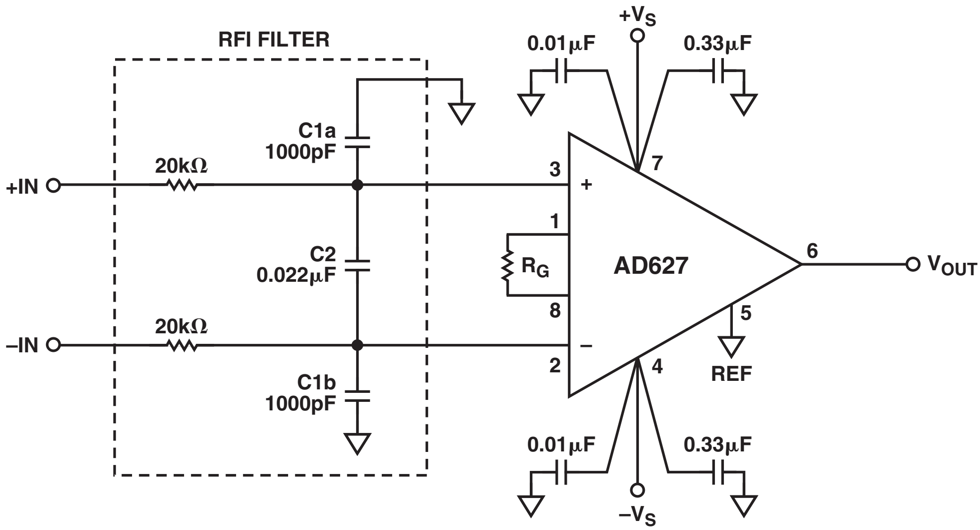


Figure 5-16. RFI Suppression Circuit for the AD627

An RFI Filter for the AD623 In-Amp

Figure 5-17 shows the recommended RFI circuit for use with the AD623 in-amp. Because this device is less prone to RFI than the AD627, the input resistors can be reduced in value from 20 kΩ to 10 kΩ; this increases the circuit's signal bandwidth and lowers the resistors' noise contribution. Moreover, the 10 kΩ resistors still provide very effective input protection. With the values shown, the bandwidth of this filter is approximately 400 Hz. Operating at a gain of 100, the maximum dc offset shift with a 1 V p-p input is less than 1 μV RTI. At the same gain, the circuit's RF signal rejection is better than 74 dB.

AD8225 RFI Filter Circuit

Figure 5-18 shows the recommended RFI filter for this in-amp. The AD8225 in-amp has a fixed gain of 5 and a bit more susceptibility to RFI than the AD8221. Without the RFI filter, with a 2 V p-p, 10 Hz to 19 MHz sine wave applied, this in-amp measures about 16 mV RTI of dc offset. The filter used provides a heavier RF attenuation than that of the AD8221 circuit by using larger resistor values: 10 kΩ instead of 4 kΩ. This is permissible because of the AD8225's higher noise level. Using the filter, there was no measurable dc offset error.

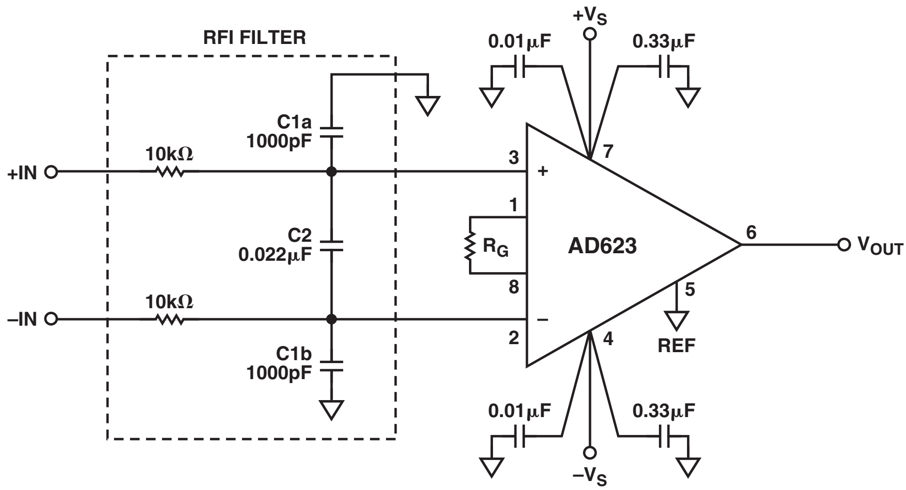


Figure 5-17. AD623 RFI Suppression Circuit

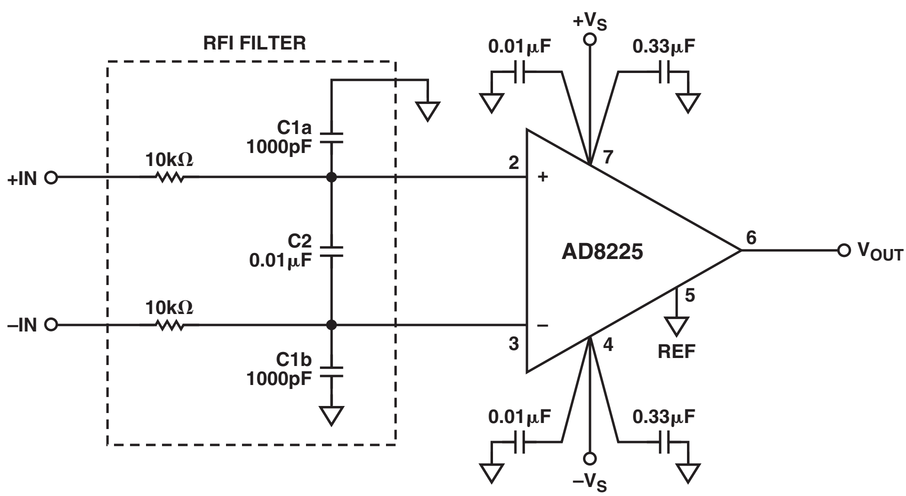


Figure 5-18. AD8225 RFI Filter Circuit

Common-Mode Filters Using X2Y® Capacitors*

Figure 5-19 shows the connection diagram for an X2Y capacitor. These are very small, three terminal devices with four external connections—A, B, G1, and G2. The G1 and G2 terminals connect internally within the device. The internal plate structure of the X2Y capacitor forms an integrated circuit with very interesting properties. Electrostatically, the three electrical nodes form two capacitors that share the G1 and G2 terminals. The manufacturing process automatically matches both capacitors very closely. In addition, the X2Y structure includes an effective autotransformer/common-mode choke. As a result, when these devices are used for common-mode filters, they provide greater attenuation of common-mode signals above the filter's corner frequency than a comparable RC filter. This usually allows the omission of capacitor C2, with subsequent savings in cost and board space.

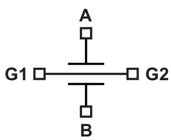


Figure 5-19. X2Y Electrostatic Model

Figure 5-20a illustrates a conventional RC common-mode filter, while Figure 5-20b shows a common-mode filter circuit using an X2Y device. Figure 5-21 is a graph contrasting the RF attenuation provided by these two filters.

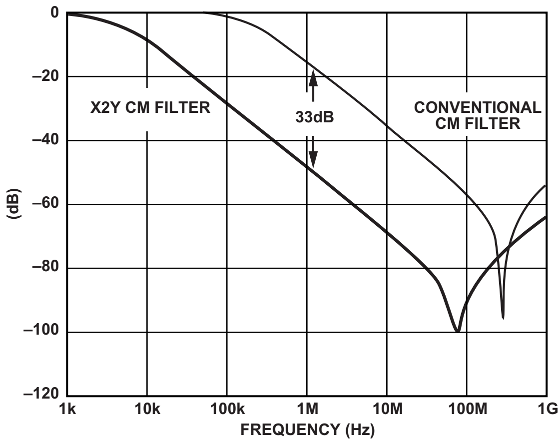


Figure 5-21. RFI Attenuation, X2Y vs. Conventional RC Common-Mode Filter

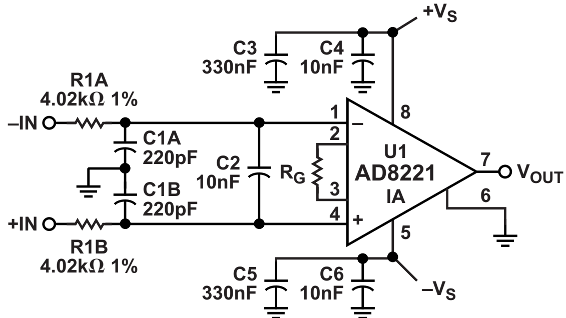


Figure 5-20a. Conventional RC Common-Mode Filter

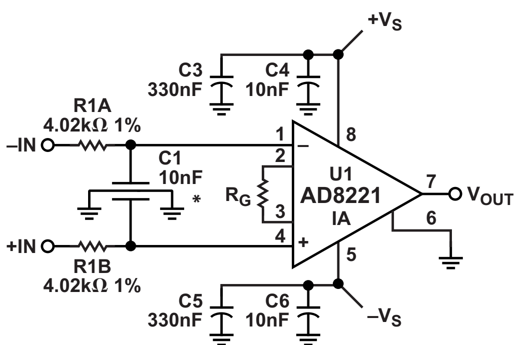


Figure 5-20b. Common-Mode Filter Using X2Y Capacitor

*C1 is part number 500X14W103KV4. X2Y components may be purchased from Johanson Dielectrics, Sylmar, CA 91750, (818) 364-9800. For a full listing of X2Y manufacturers visit: http://www.x2y.com/manufacturers.

Using Common-Mode RF Chokes for In-Amp RFI Filters

As an alternative to using an RC input filter, a commercial common-mode RF choke may be connected in front of an in-amp, as shown in Figure 5-22. A common-mode choke is a two-winding RF choke using a common core. Any RF signals that are common to both inputs will be attenuated by the choke. The common-mode choke provides a simple means for reducing RFI with a minimum of components and provides a greater signal pass band, but the effectiveness of this method depends on the quality of the particular common-mode choke being used. A choke with good internal matching is preferred. Another potential problem with using the choke is that there is no increase in input protection as is provided by the RC RFI filters.

Using an AD620 in-amp with the RF choke specified, at a gain of 1,000, and a 1 V p-p common-mode sine wave applied to the input, the circuit of Figure 5-22 reduces the dc offset shift to less than 4.5 μV RTI. The high frequency common-mode rejection ratio was also greatly improved, as shown in Table 5-3.

Table 5-3. AC CMR vs. Frequency Using the Circuit of Figure 5-22

Frequency	CMRR (dB)
100 kHz	100
333 kHz	83
350 kHz	79
500 kHz	88
1 MHz	96

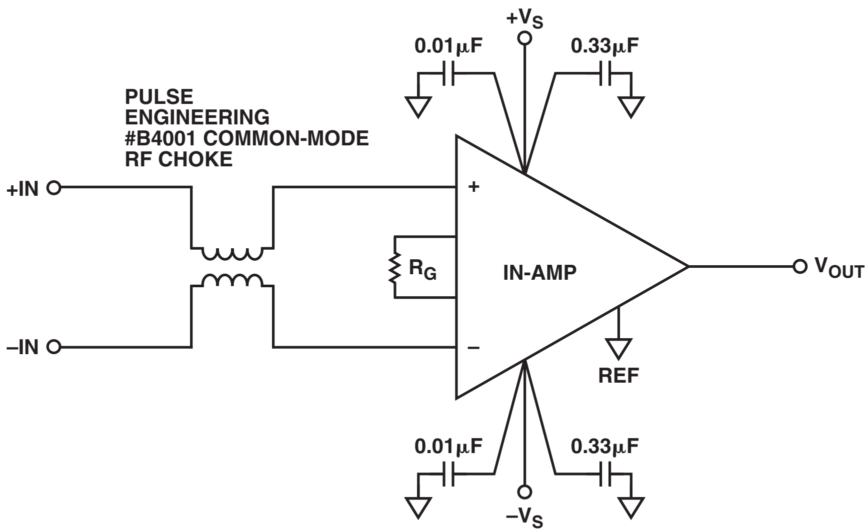


Figure 5-22. Using a Commercial Common-Mode RF Choke for RFI Suppression

Because some in-amps are more susceptible to RFI than others, the use of a common-mode choke may sometimes prove inadequate. In these cases, an RC input filter is a better choice.

RFI TESTING

Figure 5-23 shows a typical setup for measuring RFI rejection. To test these circuits for RFI suppression, connect the two input terminals together using very short leads. Connect a good quality sine wave generator to this input via a 50 V terminated cable.

Using an oscilloscope, adjust the generator for a 1 V peak-to-peak output at the generator end of the cable. Set the in-amp to operate at high gain (such as a gain of 100). DC offset shift is simply read directly at the in-amp's output using a DVM. For measuring high frequency CMR, use an oscilloscope connected to the in-amp output by a compensated scope probe and measure the peak-to-peak output voltage (i.e., feedthrough) versus input frequency. When calculating CMRR versus frequency, remember to take into account the input termination ($V_{IN}/2$) and the gain of the in-amp.

$$CMRR = 20\log \frac{\left(\dfrac{V_{IN}}{2}\right)}{\left(\dfrac{V_{OUT}}{Gain}\right)}$$

USING LOW-PASS FILTERING TO IMPROVE SIGNAL-TO-NOISE RATIO

To extract data from a noisy measurement, low-pass filtering can be used to greatly improve the signal-to-noise ratio of the measurement by removing all signals that are not within the signal bandwidth. In some cases, band-pass filtering (reducing response both below and above the signal frequency) can be employed for an even greater improvement in measurement resolution.

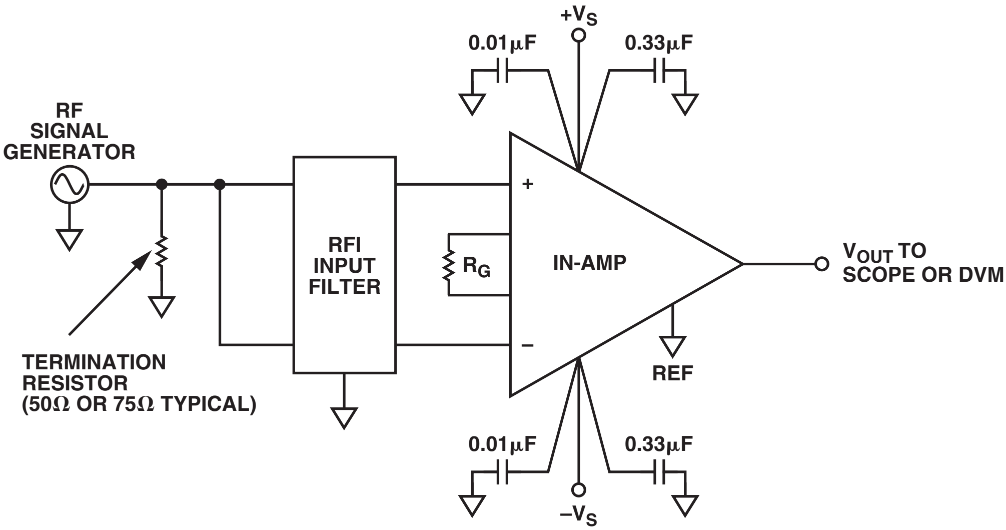


Figure 5-23. Typical Test Setup for Measuring an In-Amp's RFI Rejection

The 1 Hz, 4-pole active filter of Figure 5-24 is an example of a very effective low-pass filter that normally would be added after the signal has been amplified by the in-amp. This filter provides high dc precision at low cost while requiring a minimum number of components.

Note that component values can simply be scaled to provide corner frequencies other than 1 Hz (see Table 5-4). If a 2-pole filter is preferred, simply take the output from the first op amp.

The low levels of current noise, input offset, and input bias currents in the quad op amp (either an AD704 or OP497) allow the use of 1 MΩ resistors without sacrificing the 1 μV/°C drift of the op amp. Thus, lower capacitor values may be used, reducing cost and space.

Furthermore, since the input bias current of these op amps is as low as their input offset currents over most of the MIL temperature range, there is rarely a need to use the normal balancing resistor (along with its noise-reducing bypass capacitor). Note, however, that adding the optional balancing resistor will enhance performance at temperatures above 100°C.

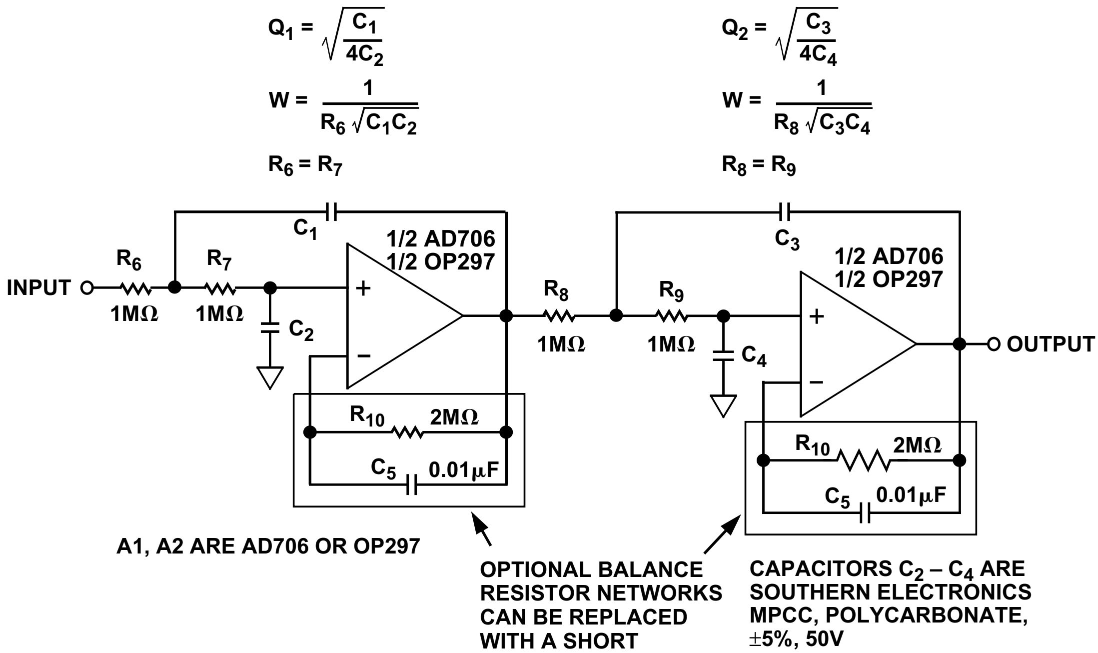


Figure 5-24. A 4-Pole Low-Pass Filter for Data Acquisition

Table 5-4. Recommended Component Values for a 1 Hz, 4-Pole Low-Pass Filter

	Section 1		Section 2					
Desired Low-Pass Response	**Frequency (Hz)**	**Q**	**Frequency (Hz)**	**(Q)**	**C1 (μF)**	**C2 (μF)**	**C3 (μF)**	**C4 (μF)**
Bessel	1.43	0.522	1.60	0.806	0.116	0.107	0.160	0.0616
Butterworth	1.00	0.541	1.00	1.31	0.172	0.147	0.416	0.0609
0.1 dB Chebychev	0.648	0.619	0.948	2.18	0.304	0.198	0.733	0.0385
0.2 dB Chebychev	0.603	0.646	0.941	2.44	0.341	0.204	0.823	0.0347
0.5 dB Chebychev	0.540	0.705	0.932	2.94	0.416	0.209	1.00	0.0290
1.0 dB Chebychev	0.492	0.785	0.925	3.56	0.508	0.206	1.23	0.0242

Specified values are for a –3 dB point of 1.0 Hz. For other frequencies, simply scale capacitors C1 through C4 directly; i.e., for 3 Hz Bessel response, C1 = 0.0387 μF, C2 = 0.0357 μF, C3 = 0.0533 μF, and C4 = 0.0205 μF.

EXTERNAL CMR AND SETTLING TIME ADJUSTMENTS

When a very high speed, wide bandwidth in-amp is needed, one common approach is to use several op amps or a combination of op amps and a high bandwidth subtractor amplifier. These discrete designs may be readily tuned-up for best CMR performance by external trimming. A typical circuit is shown in Figure 5-25. The dc CMR should always be trimmed first, since it affects CMRR at all frequencies.

The $+V_{IN}$ and $-V_{IN}$ terminals should be tied together and a dc input voltage applied between the two inputs and ground. The voltage should be adjusted to provide a 10 V dc input. A dc CMR trimming potentiometer would then be adjusted so that the outputs are equal and as low as possible, with both a positive and a negative dc voltage applied.

AC CMR trimming is accomplished in a similar manner, except that an ac input signal is applied. The input frequency used should be somewhat lower than the –3 dB bandwidth of the circuit.

The input amplitude should be set at 20 V p-p with the inputs tied together. The ac CMR trimmer is then nulled-set to provide the lowest output possible. If the best possible settling time is needed, the ac CMR trimmer may be used, while observing the output wave form on an oscilloscope. Note that, in some cases, there will be a compromise between the best CMR and the fastest settling time.

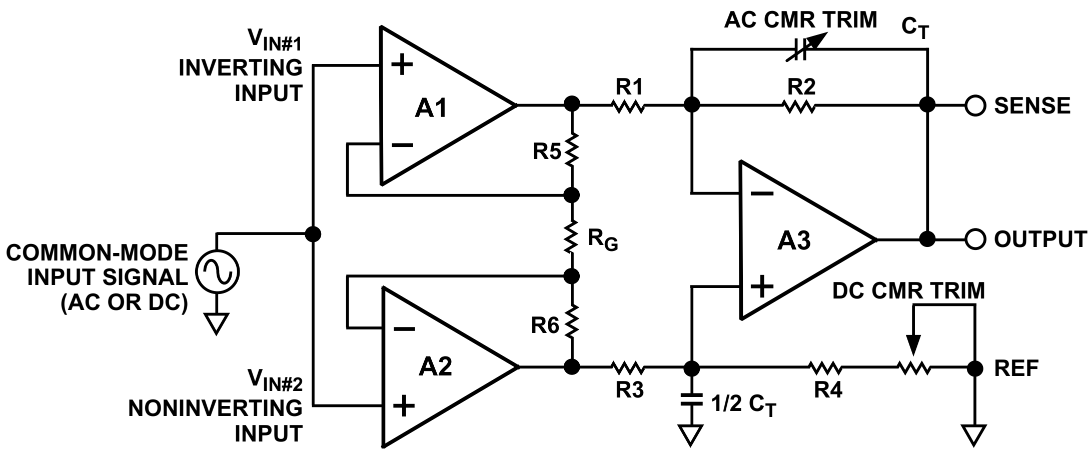


Figure 5-25. External DC and AC CMRR Trim Circuit for a Discrete 3-Op Amp In-Amp