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:

\[ \text{Total Error, RTI} = \text{Input Error} + (\text{Output Error/Gain}) \]
\[ \text{Total Error, RTO} = (\text{Gain} \times \text{Input Error}) + \text{Output Error} \]

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 \(\mu\)V. Its output offset (\(V_{OSO}\)) is listed as 400 \(\mu\)V. Thus, the total voltage offset referred to input, RTI, is equal to

\[ \text{Total RTI Error} = V_{OSI} + (V_{OSO}/G) = 30 \, \mu\text{V} + (400 \, \mu\text{V}/10) \]
\[ = 30 \, \mu\text{V} + 40 \, \mu\text{V} = 70 \, \mu\text{V} \]

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

\[ \text{Total Offset Error RTO} = (G \times V_{OSI}) + V_{OSO} = (10 \times 30 \, \mu\text{V}) + 400 \, \mu\text{V} = 700 \, \mu\text{V} \]

Note that the two error numbers (RTI versus RTO) are 10 times 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

\[ \text{Input Noise} = \text{eni}, \text{Output Noise} = \text{eno} \]
\[ \text{Total Noise RTI} = \sqrt{(\text{eni})^2 + (\text{eno}/\text{Gain})^2} \]
\[ \text{Total Noise RTO} = \sqrt{[\text{Gain}(\text{eni})]^2 + (\text{eno})^2} \]

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

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

**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/√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

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

\[ BW_{CM} = \frac{1}{2\pi RC} \]

Using the circuit of Figure 5-13, with a C2 value of 0.01 \( \mu \)F 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 \( \mu \)V 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 k\( \Omega \) and 10 k\( \Omega \), these resistors should not contribute more noise than that of the in-amp itself. Using a pair of 2 k\( \Omega \) resistors will add a Johnson noise of 8 nV/\( \sqrt{\text{Hz}} \); this increases to 11 nV/\( \sqrt{\text{Hz}} \) with 4 k\( \Omega \) resistors and to 18 nV/\( \sqrt{\text{Hz}} \) with 10 k\( \Omega \) 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](image-url)
Accordingly, the same input resistors were used but capacitor C2 was increased approximately five times to 0.047 \( \mu \)F 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 \( \mu \)V 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](image)
**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.
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 5-20a. 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>
<thead>
<tr>
<th>Frequency</th>
<th>CMRR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 kHz</td>
<td>100</td>
</tr>
<tr>
<td>333 kHz</td>
<td>83</td>
</tr>
<tr>
<td>350 kHz</td>
<td>79</td>
</tr>
<tr>
<td>500 kHz</td>
<td>88</td>
</tr>
<tr>
<td>1 MHz</td>
<td>96</td>
</tr>
</tbody>
</table>

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

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.

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

**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](image-url)
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.

\[
Q_1 = \sqrt{\frac{C_1}{4C_2}} \quad W = \frac{1}{R_6 \sqrt{C_1 C_2}} \quad R_6 = R_7
\]

\[
Q_2 = \sqrt{\frac{C_3}{4C_4}} \quad W = \frac{1}{R_8 \sqrt{C_3 C_4}} \quad R_8 = R_9
\]

![Figure 5-24. A 4-Pole Low-Pass Filter for Data Acquisition](image)

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

<table>
<thead>
<tr>
<th>Desired Low-Pass Response</th>
<th>Section 1</th>
<th>Section 2</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Frequency (Hz)</td>
<td>Q</td>
</tr>
<tr>
<td>Bessel</td>
<td>1.43</td>
<td>0.522</td>
</tr>
<tr>
<td>Butterworth</td>
<td>1.00</td>
<td>0.541</td>
</tr>
<tr>
<td>0.1 dB Chebychev</td>
<td>0.648</td>
<td>0.619</td>
</tr>
<tr>
<td>0.2 dB Chebychev</td>
<td>0.603</td>
<td>0.646</td>
</tr>
<tr>
<td>0.5 dB Chebychev</td>
<td>0.540</td>
<td>0.705</td>
</tr>
<tr>
<td>1.0 dB Chebychev</td>
<td>0.492</td>
<td>0.785</td>
</tr>
</tbody>
</table>
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, \( C_1 = 0.0387 \, \mu F, \quad C_2 = 0.0357 \, \mu F, \quad C_3 = 0.0533 \, \mu F, \) and \( C_4 = 0.0205 \, \mu F. \)

**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](image-url)