

# Buffered ADC Family Eliminates Signal Conditioning Complexity

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Engineers often assume that analog-to-digital converter inputs are high impedance. Direct-sampling SAR ADC inputs will often be high impedance when not converting, but will draw “spikes” of current at the beginning of sample acquisition. On average, this behavior can be modeled as a crude nonlinear resistance that is inversely proportional to the sample capacitor size and sample rate, but instantaneously the signal chain must be able to settle completely in response to this abuse before acquisition ends and a conversion begins. Coupling a signal chain to an

ADC is an art form, often requiring a combination of theory and experimentation. By contrast, the [LTC2358](#) family of multi-channel, buffered high voltage SAR ADCs offers truly high impedance inputs that simplify or eliminate the need for signal conditioning in many cases. When signal conditioning is required, it can be directly connected to the LTC2358 inputs without regard to its ability to drive a switched capacitor. The following circuits show some applications that take advantage of the properties of the LTC2358 inputs.

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## EASY DRIVE AND OVERDRIVE

### 1. Eight Differential Channels with Sixteen Picoamp Buffered Inputs Are Easy to Drive Directly

Each channel simultaneously samples the voltage difference ( $V_{IN^+} - V_{IN^-}$ ) between its analog input pins over a wide common mode input range while attenuating unwanted signals common to both input pins by the common mode rejection ratio (CMRR) of the ADC. Wide common mode input range coupled with high CMRR (128dB at 200Hz) allows the  $IN^+/IN^-$  analog inputs to swing with an arbitrary relationship to each other, provided each pin remains between ( $V_{EE} + 4V$ ) and ( $V_{CC} - 4V$ ). This feature of the LTC2358 enables it to accept a wide variety of signal swings, including traditional classes of analog input signals such as pseudo-differential unipolar, pseudo-differential true bipolar, and fully differential, simplifying signal chain design. Figure 1 shows the typical application of the LTC2358. For conversion of signals extending to  $V_{EE}$ , the unbuffered LTC2348 ADC is recommended.

The picoamp-input CMOS buffers offer a very high degree of transient isolation from the sampling process in the ADC. This means that most sensors, signal conditioning

amplifiers and filter networks with less than 10k $\Omega$  of impedance can drive the passive 3pF analog input capacitance directly. Figure 2 shows the equivalent circuit for each differential analog input channel.

The very high input impedance of the internal unity gain buffers, typically > 1000G $\Omega$ , greatly reduces the drive requirements of the external amplifier and makes it possible to include optional RC filters with k $\Omega$  impedance and arbitrarily slow time constants for anti-aliasing or other purposes. Micropower op amps with limited drive capability are also well suited to drive the high impedance analog inputs.

As recommended in the data sheet, for source impedances greater than 10k $\Omega$ , a 680pF (or larger) capacitor at the analog input reduces the reverse transient voltage glitch through the internal input buffer to maintain the DC accuracy of the LTC2358. This very small glitch is also charge-conserving, which is to say that it is purely AC-coupled and the total charge of the glitch is zero and has no DC component. Picoamp DC input currents are maintained even in the presence of arbitrarily sharp transients and when the inputs are overdriven to (but not beyond) the  $V_{CC}$  and  $V_{EE}$  power rails.

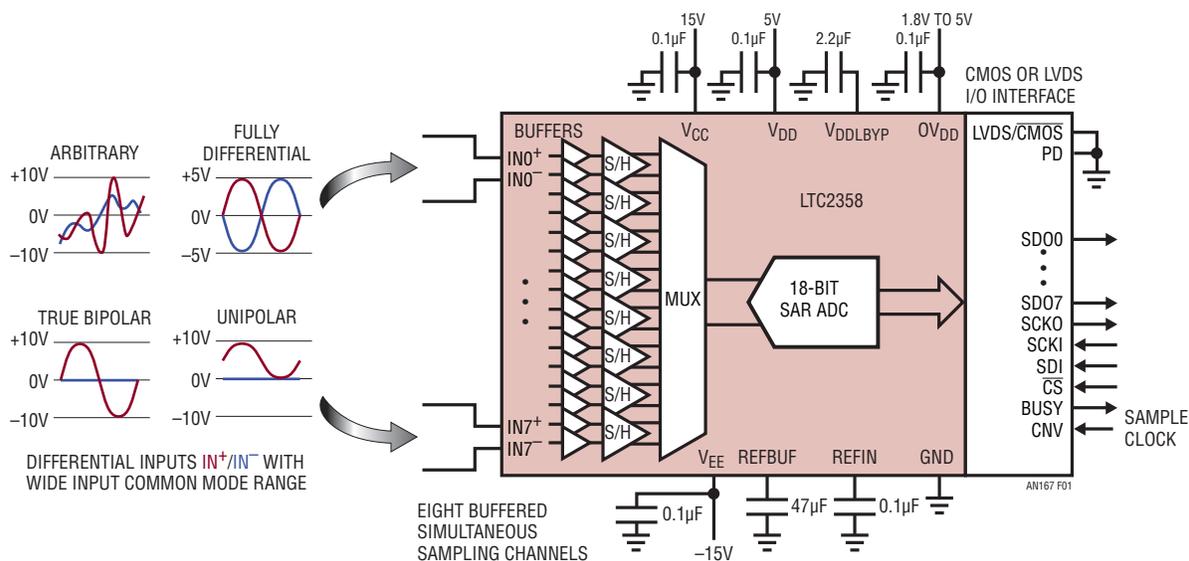


Figure 1. Eight Differential Channels with Sixteen Picoamp Buffered Inputs Are Easy to Drive Directly

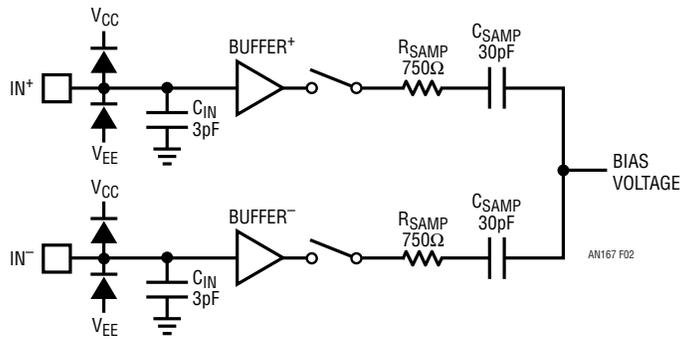


Figure 2. Equivalent Circuit for Differential Analog Inputs, Single Channel Shown

## 2. Twisted Pair of Arbitrary Length Drives LTC2358 Directly

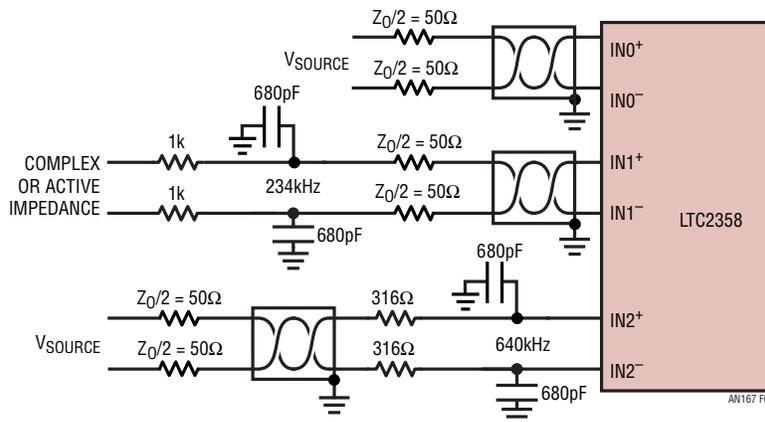
One of the simple, but very useful, applications of the buffered analog inputs is the ability to accept analog signals from twisted wire pairs of arbitrary length. It is recommended that the twisted pair be properly terminated at the driving source to minimize potential cable reflections. The twisted pair characteristic impedance is usually in the low 100Ω range. For example, CAT7 cable has 4 individually shielded twisted pairs with 100Ω differential impedance.

The level of shielding between the twisted pairs in a CAT7 cable may vary with the physical construction style. For example, flat ribbon CAT7 cable showed poor internal capacitive crosstalk isolation of only about 10dB, while a CAT7 cable with the usual round cross-section showed at least 50dB of capacitive crosstalk isolation. In this case, capacitive crosstalk isolation

is the ratio of the self-capacitance of a twisted pair to the capacitance between twisted pairs in dB units. This capacitive crosstalk is most relevant at higher frequencies and when the source impedance is much higher than the characteristic impedance of the cable.

If the source is a complex or active impedance of unknown characteristics, an additional isolation RC filter is recommended between the unknown source impedance and the source termination resistors. An RC filter, like the 640kHz filter shown in Figure 3 at IN2, may also be used at the ADC inputs to reduce RF interference that may have been picked up by the twisted pair. The analog inputs have no self-rectification mechanism that would convert RF interference into a spurious DC level at the input pin, thus making the analog inputs very robust against EMI.

The circuits shown in Figure 3 were verified with a 15-foot CAT7 cable to have no discernible effect on offset voltage or linearity.



SOURCE TERMINATION  $Z_0 = 100\Omega$  REDUCES SIGNAL REFLECTIONS.  
CAT7  $Z_0 = 100\Omega$

Figure 3. Twisted Pair of Arbitrary Length Drives LTC2358 Directly

## 3. Overdrive the Analog Inputs with Limited Current

Driving an analog input above  $V_{CC}$  on any channel up to 10mA will not affect conversion results on other channels. Approximately 70% of this overdrive current will flow out of the  $V_{CC}$  pin and the remaining 30% will flow out of  $V_{EE}$ . This current flowing out of  $V_{EE}$  will produce heat across the  $V_{CC} - V_{EE}$  voltage drop and must be taken into account for the total absolute maximum power dissipation of 500mW. Driving an analog input below  $V_{EE}$  may corrupt conversion results on other channels. The LTC2358 can handle input currents of up to 100mA below  $V_{EE}$  or above  $V_{CC}$  without latchup. Keep in mind that driving the inputs above  $V_{CC}$  or below  $V_{EE}$  may reverse the normal current flow from the external power supplies driving these pins, which may raise the externally applied supply voltages. Figure 4 illustrates the overdrive response with 2.49k external resistors up to  $\pm 40V$ .

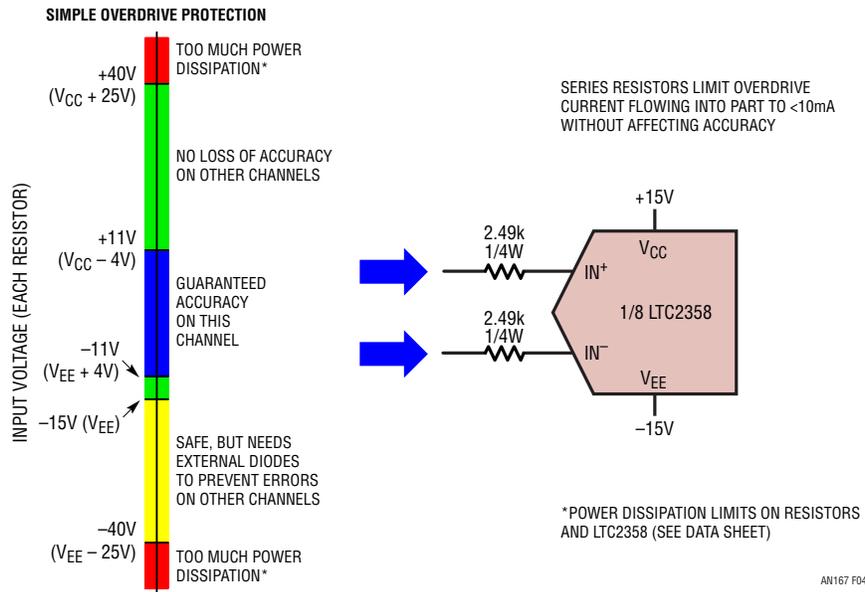
Depending on system requirements, a range of input overdrive current limiting circuits can be used as shown in Figure 5. Single external resistors up to 10k can be used to limit input current while remaining transparent to the AC and DC performance of the LTC2358 when inside the normal conversion ranges. For example, 10k, 1W input resistors will limit the input current under 10mA with  $\pm 100V$  of overdrive. If less overdrive power dissipation and wider input voltages are desired, current limiting depletion mode N-channel MOSFETs can replace the external current limiting resistors. A pair of LND150 depletion mode N-channel MOSFETs from Microchip-Supertex wired in series, in opposing direction, lowers the external peak overdrive currents to the  $I_{DSS} = 3mA$  maximum, while tolerating up to  $\pm 400V$ . Infineon also makes depletion N-channel

MOSFETs like the BSS126 with  $I_{DSS} = 7mA$  maximum up to 600V. Refer to the MOSFET manufacturer's specifications for safe operating area. If even lower overdrive circuit dissipation is desired, additional degeneration resistors may be wired between the N-channel MOSFET sources and gates to reduce peak currents to  $\pm 150\mu A$  maximum.

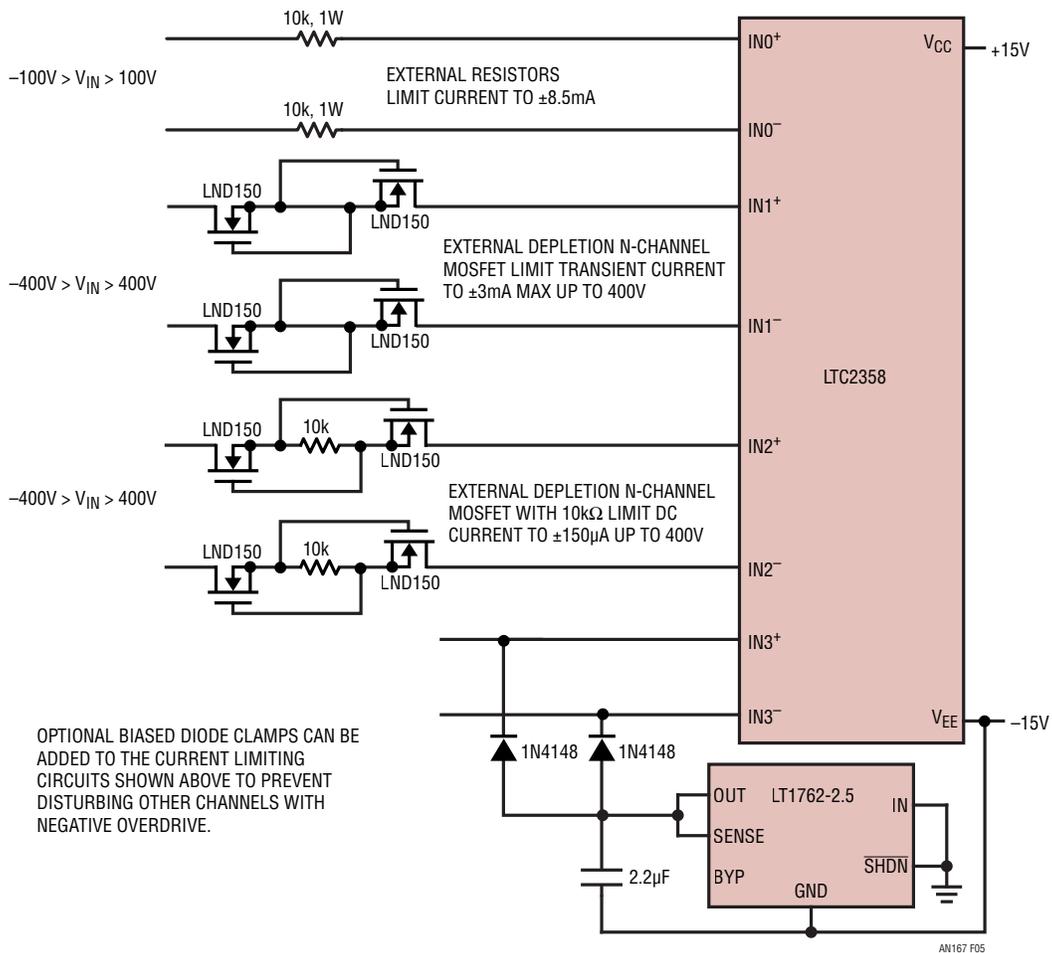
As mentioned earlier, up to 10mA overdrive above  $V_{CC}$  has no effect on the analog results of other channels. If the same level of immunity is desired for overdrive below  $V_{EE}$ , diode clamps may be added to the input current limiting circuits to limit the input negative swing. One option illustrated in IN3 of Figure 5 is to use standard small signal silicon diodes, like the 1N4148, that are tied to 2.5V above  $V_{EE}$ . When these diodes are forward biased under negative overdrive conditions, the analog input remains clamped  $2.5V - 0.7V = 1.8V$  above  $V_{EE}$ . This option takes advantage of the low leakage levels of silicon diodes like the 1N4148. Small signal Schottky diodes, like the SD101, wired from the analog inputs directly to  $V_{EE}$  improve but don't completely eliminate the crosstalk effect from the overdriven channel to the other channels, and are thus not recommended. The higher leakage currents of the Schottky diodes add directly to the analog input leakage current performance, which poses a further drawback.

The overdrive current limiting circuits may also be combined with filter or voltage range scaling circuits shown elsewhere in this note, so that the same resistive elements serve to limit overdrive current as well as filter or scale the analog input signal inside the normal conversion ranges.

The overdrive current limiting circuits shown in Figure 5 retain the overall linearity, gain and offset performance of the LTC2358.



**Figure 4. Illustration of Overdrive Behavior with Input Current Limiting Resistors**



**Figure 5. Overdrive the Analog Inputs with Limited Current**

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## 4. High Voltage Analog Supply Pins are Switcher-Friendly

The high voltage supplies of the LTC2358 ( $V_{CC}$  and  $V_{EE}$ ) have a power supply rejection ratio (PSRR) in excess of 130dB at DC, and a wide range of operating voltages. The absolute common mode input range ( $V_{EE} + 4V$  to  $V_{CC} - 4V$ ) is determined by the choice of high voltage supplies. These supplies may be biased asymmetrically around ground and include the ability for  $V_{EE}$  to be tied directly to ground. This versatility in supply voltage range and high PSRR loosens high voltage supply accuracy requirements, and allow the LTC2358 to tolerate supply ripple on  $V_{CC}$  and  $V_{EE}$ .

Because the LTC2358's PSRR is very good even at higher frequencies (90dB at 100kHz), a micropower switched DC/DC converter such as the LT3463 can be used to generate the high voltage  $V_{CC}/V_{EE}$  rails from a single 5V supply without injecting noise into the ADC output. The single 5V supply drives both  $V_{DD}$  for the LTC2358 and  $V_{IN}$  for the LT3463, simplifying power supply requirements while maintaining a small board footprint. Figure 6 shows a recommended circuit for the case with  $V_{DD} = OV_{DD} = 5V$ ,  $V_{CC} = 15V$  and  $V_{EE} = -15V$ . Consult the LTC2358 and LT3463 data sheets for other supply voltage configurations.

In order to avoid magnetically coupled interference from the inductors, the LT3463 and its associated components should be located on the digital side of the LTC2358, and

preferably away from the LTC2358. Layout should be carefully planned with local supply bypass caps to avoid coupling switching transient currents from the LT3463 to the LTC2358. A single ground plane works well with the LTC2358. Separate analog and digital ground planes are not recommended. Bypass capacitors must be placed both at the  $V_{DD}$  pin of the LTC2358 and the  $V_{IN}$  pin of the LT3463.

Figure 7 shows an FFT of the LTC2358-18 output, when configured in this single 5V supply circuit. The ripple on the  $V_{CC}$  and  $V_{EE}$  nodes were measured as  $\sim 80mV_{P-P}$  and  $\sim 50mV_{P-P}$  sawtooth waves at a few kHz, but no spectral peaks are detected on the FFT plot, and the ADC's performance is unaffected when compared to an equivalent circuit with linear regulators supplying  $V_{CC}$  and  $V_{EE}$ .

If there is a switched-mode supply elsewhere in the system, it can drive  $V_{CC}$  and  $V_{EE}$  directly or through a simple RC filter. It is possible that the ripple frequency of the existing switching power supply reaches into the MHz range. A lab experiment shows that  $50mV_{P-P}$  of square wave ripple at 1MHz at  $V_{CC}$  or  $V_{EE}$  is well rejected by the LTC2358, leaving only a  $6\mu V$  residual peak tone in the ADC output spectrum. This  $6\mu V$  tone is negligible for most applications, but it can be completely eradicated with simple  $50\Omega/4.7\mu F$  RC filters at  $V_{CC}$  and  $V_{EE}$ . The supply currents,  $|I_{VCC}| < 9.8mA$  and  $|I_{VEE}| < 9.8mA$  maximum, cause only small voltage drops under 500mV on the  $50\Omega$  resistors of this supply bypass RC filter network.

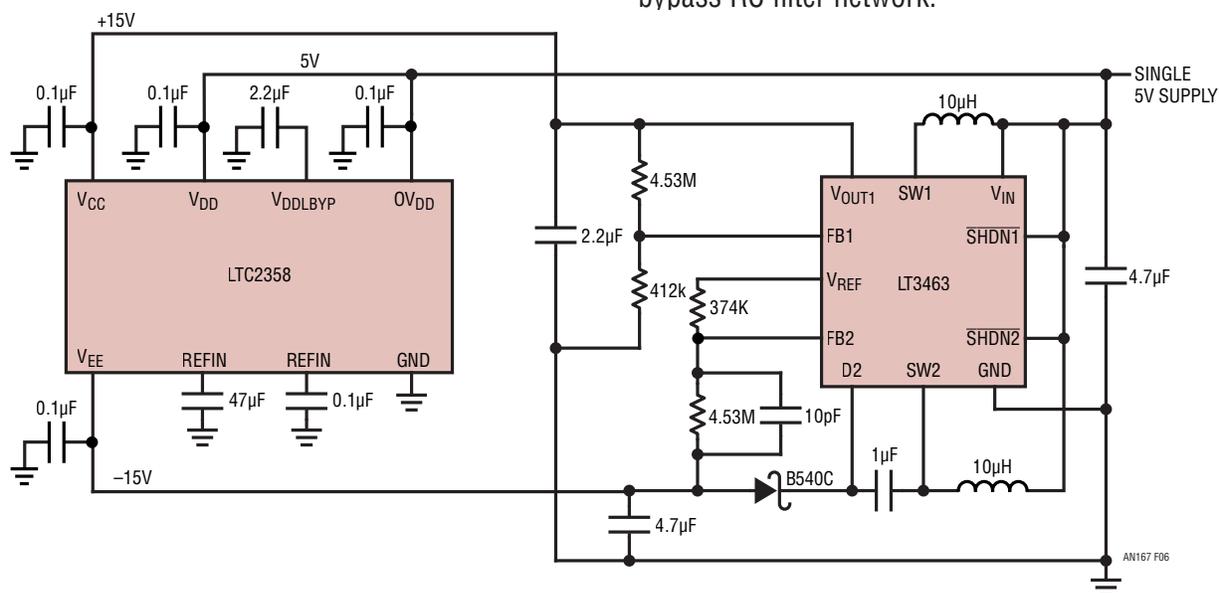


Figure 6. Single 5V Supply Operation

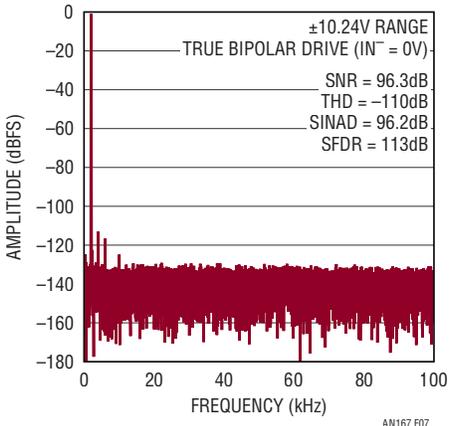


Figure 7. Full AC Performance is Maintained with Switched-Mode Supplies

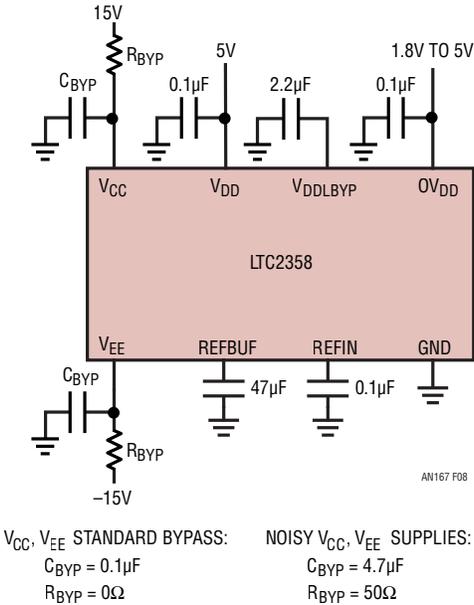


Figure 8. Flexible Bypass Rejects All Ripple from Noisy High Voltage Supplies

# Application Note 167

## 5. Amplify Sensor or Current Sense Signal Over a Wide Common Mode Voltage

The ability of the LTC2358 to accept arbitrary signal swings over a wide input common mode range with high CMRR can simplify application solutions. In practice, many sensors produce a differential voltage riding on top of a large common mode signal. Figure 9 depicts one way of using the LTC2358 to digitize signals of this type. The amplifier stage provides a differential gain of approximately 10V/V to the desired sensor signal while the unwanted common mode signal is attenuated by the ADC CMRR. The circuit employs the  $\pm 5V$  SoftSpan™ range of the ADC. The ADC inputs may swing up to  $V_{CC} - 4V = 27V$  or down to  $V_{EE} + 4V = -3V$ . Keep in mind that the 5V<sub>P-P</sub> swing at the ADC inputs consumes 5V of the available common mode voltage range. Any other combination of  $V_{CC}$  and  $V_{EE}$  voltages may be used to suit a particular application up to  $V_{CC} - V_{EE} = 38V$  maximum, with  $V_{CC} > 7.5V$  and  $-16.5V < V_{EE} < 0V$ . Figure 10 shows measured CMRR performance of this

solution, which is competitive with the best commercially available instrumentation amplifiers. Figure 11 shows measured AC performance of this solution.

The gain of the amplifier could be increased to 100 with higher valued feedback resistors, so that a  $\pm 50mV$  differential input becomes a  $\pm 5V$  swing to drive the full dynamic range of the ADC in the  $\pm 5V$  SoftSpan range. The LTC2057HV chopper-stabilized op amp has a maximum offset specification of  $4\mu V$ , which allows for the accurate measurement of small currents through the external sense resistors.

In Figure 12, another application circuit is shown which uses two channels of the LTC2358 to simultaneously sense the voltage and bidirectional current through a sense resistor over a wide common mode range. Two RC filters may also be placed between the sense resistor and the LTC2358 inputs to eliminate switching transients from the power supply or its load.

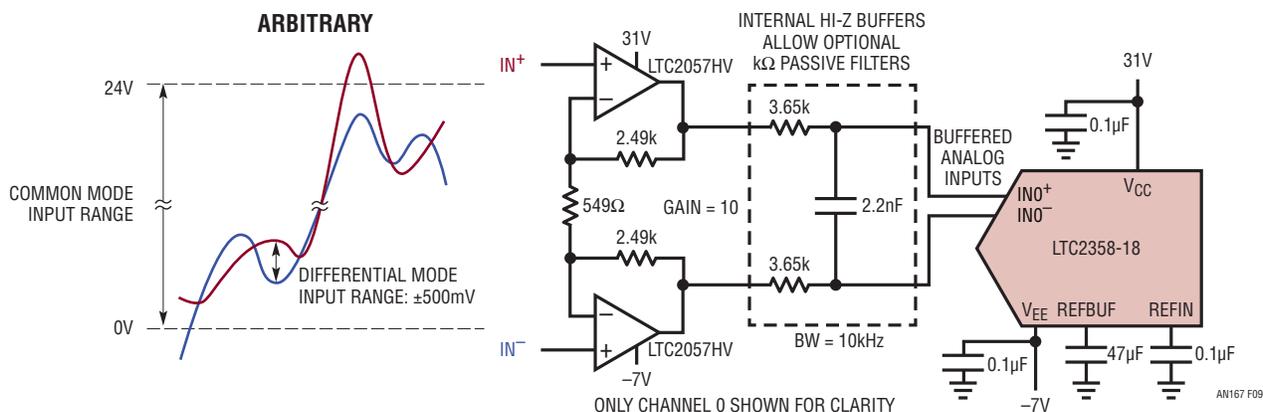


Figure 9. Amplify Differential Signals with Gain of 10 Over a Wide Common Mode Range with Buffered Analog Inputs

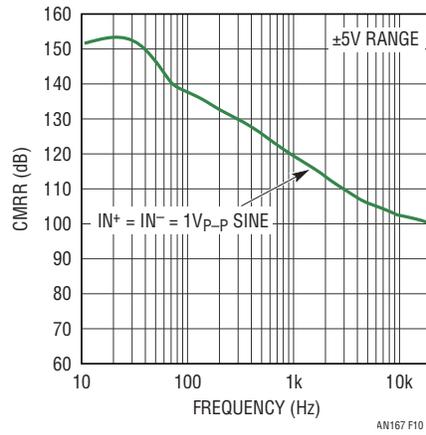


Figure 10. CMRR vs Input Frequency (Circuit Shown in Figure 9)

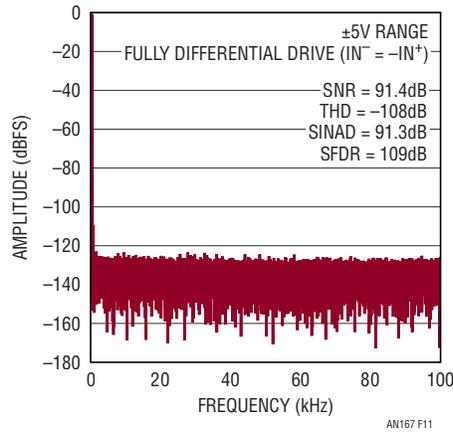
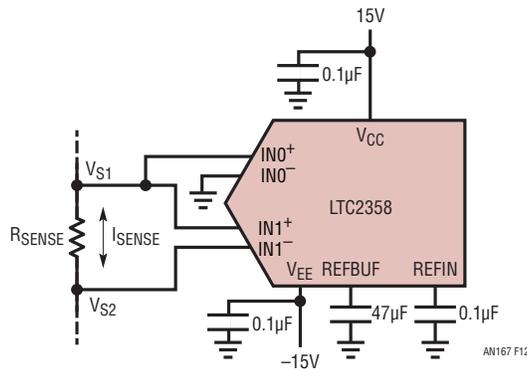


Figure 11.  $IN^+/IN^- = 450\text{mV}$  200Hz Fully Differential Sine,  $0\text{V} \leq V_{CM} \leq 24\text{V}$ , 32k Point FFT,  $f_{SMPLE} = 200\text{kps}$  (Circuit Shown in Figure 9)



ONLY CHANNELS 0 AND 1 SHOWN FOR CLARITY

$$I_{SENSE} = \frac{V_{S1} - V_{S2}}{R_{SENSE}} \quad \begin{matrix} -10.24\text{V} \leq V_{S1} \leq 10.24\text{V} \\ -10.24\text{V} \leq V_{S2} \leq 10.24\text{V} \end{matrix}$$

Figure 12. Simultaneously Sense Voltage ( $IN_0$ ) and Current ( $IN_1$ ) Over a Wide Common Mode Range

# Application Note 167

## CROSSTALK

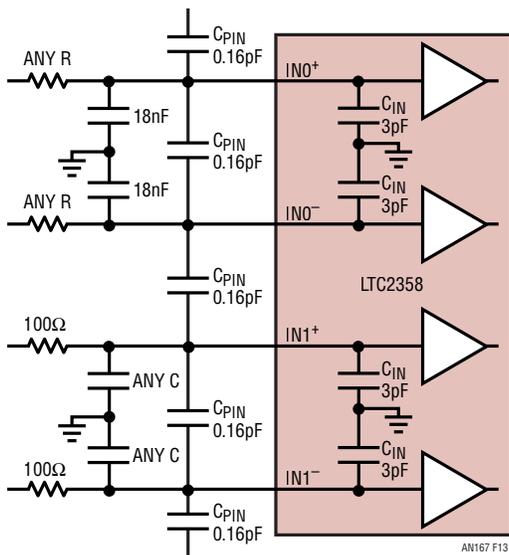
### 6. Reduce PC Board Crosstalk with Input Capacitors and Single-Ended Operation

The LTC2358 features proprietary circuitry to achieve exceptional internal crosstalk isolation between active channels (109dB typical). The PC board wiring to the analog inputs should be shielded with ground in the conductor layers above and below, as well as with adjacent ground runs to minimize external capacitive crosstalk between channels. The capacitance between adjacent package pins is 0.16pF. Low source resistance and/or high source capacitance help reduce external capacitively coupled crosstalk. For example, a 18nF capacitor at the analog input attenuates adjacent pin induced capacitive coupling of 0.16pF by 100dB, regardless of source resistance as illustrated with IN0 in Figure 13.

Low source impedance also reduces external wiring crosstalk. Channel IN1 in Figure 13 shows that 100Ω (or less) source resistance can be used independently of the input capacitance to obtain 100dB of crosstalk rejection up to 100kHz.

At high frequencies, with high source resistances and no additional input capacitors, the 0.16pF capacitance between adjacent pins forms a > 26dB voltage attenuator with the input capacitance of the adjacent channel, which includes 3pF of internal capacitance plus any PCB input trace capacitance. A further attenuation is then provided by the source resistance and the 3pF internal input capacitance with a 6dB/octave improvement toward lower frequencies until DC is reached, where the full ADC crosstalk performance of 109dB is realized. For the case of 10k source resistance, coupling is reduced with a 6dB/octave slope for frequencies below the 5MHz pole formed by the 10k source resistance and 3pF input capacitance at the input pin. At 100kHz, the pin-to-pin crosstalk rejection calculates to  $0.16\text{pF}/3\text{pF} \cdot 0.1\text{MHz}/5\text{MHz} = 0.001$  (-60dB).

These calculations must be adjusted to include additional capacitance between board input traces. Single-ended input drive also enjoys additional external crosstalk isolation because every other input pin is grounded, or is driven by a low impedance DC source, and serves as a shield between channels. Keep in mind that each input wiring connection must be fully shielded on all sides with GND right up to the input pin.

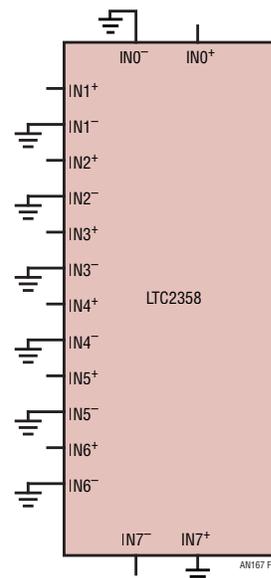


PC TRACES LEADING TO INPUTS MUST BE SHIELDED.

IN0: CAPACITORS PROTECT HIGH IMPEDANCE CIRCUITS FROM 0.16pF PIN-TO-PIN CAPACITANCE. EXTERNAL CROSSTALK < -100dB WITH ANY SOURCE RESISTANCE.

IN1: SOURCE IMPEDANCE BELOW 100Ω REDUCES EXTERNAL CROSSTALK TO LESS THAN -100dB AT 100kHz FOR ANY VALUE OF INPUT FILTER CAPACITANCE.

**Figure 13. Input Capacitors and Low Source Impedance Reduce PC Board Crosstalk**



**Figure 14. Grounded INx- Inputs Serve as External Shields**

## RANGE

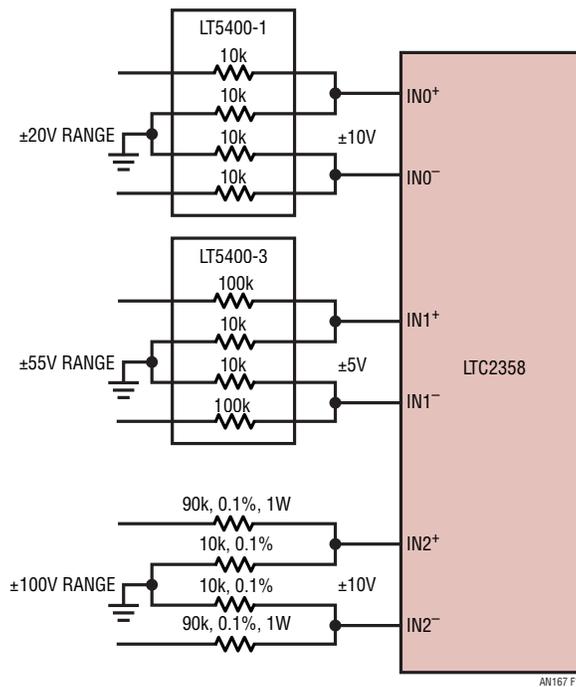
### 7. Attenuator Expands ADC Input Range

The  $> 1000G\Omega$ , picoamp analog inputs are ideally suited for external precision attenuators to realize higher voltage input ranges. For example, the LT5400 family of precision quad resistor networks with 0.01% matching accuracy can be used to achieve various analog input ranges up to the maximum operating voltage of  $\pm 75V$  for the LT5400.

The impedance of the attenuator circuits in Figure 15 is under  $10k\Omega$ , which allows for full settling of the LTC2358 inputs from the small AC-coupled transient that is fed back

through the internal CMOS buffers to the channel inputs at the start of the acquisition period.

1W resistors are recommended for discrete attenuators to minimize resistor self-heating, which would potentially change the value of the resistance due to the temperature coefficient of the resistor. A 1W resistor rating is much larger than the 90mW that is dissipated with a 100V input on the 90k resistor at channel IN2.



EXPAND THE DIFFERENTIAL AND COMMON MODE RANGES WITH EXTERNAL ATTENUATORS.

Figure 15. Input Range Expander

## 8. Automatic Gain Ranging with External Attenuators

Figure 16 shows a high impedance attenuator network applied to measure the voltage at  $A_{IN}$  with minimal loading of 1.33M in three possible voltage ranges of 0V to 100V, 0V to 200V and 0V to 400V. The three channels sample the attenuated voltage simultaneously. The correct range is then selected by the user as the smallest range that is not saturated with all-ones at the digital channel output. It is also advisable to always select the 100V range from IN0 output if no channels are saturated with all-ones. This implements a form of automatic gain ranging.

When the voltage at  $A_{IN}$  rises to 250V, the ESD protection diodes at  $IN0^+$  will start to forward bias and conduct up to  $80\mu A$  when  $A_{IN} = 400V$ . This overdrive of  $IN0^+$  has no effect on the other channels. See Section 3 for more about overdriving the analog channel inputs.

The  $M\Omega$  impedance of this network is made possible by the very low analog input leakage ( $5pA$ , typically). This

$M\Omega$  impedance was chosen for minimal loading on the external source and for minimal power dissipation at the 0.1% attenuator input resistors. The 680pF filter capacitors at the analog channel inputs suppress the internal CMOS buffer feedthrough glitch at the start of the acquisition period and filter out external noise. The room temperature leakage of the channel inputs is  $5pA$  typically, which only imparts a negligible  $1\mu V$  offset at the  $<200k\Omega$  attenuator outputs. Beware that leakage current rises exponentially with temperature up to  $500pA$  maximum at  $85^\circ C$ . A lower impedance network reduces the effects of leakage current at high temperatures. Alternatively, resistors that match the impedance of the noninverting channel inputs may also be added in series with the grounded inverting channel inputs to obtain some cancellation of the offset voltages that are induced by the elevated input bias current at high temperature. These resistors should also be bypassed with additional filter capacitors of the same value as used in the active inputs.

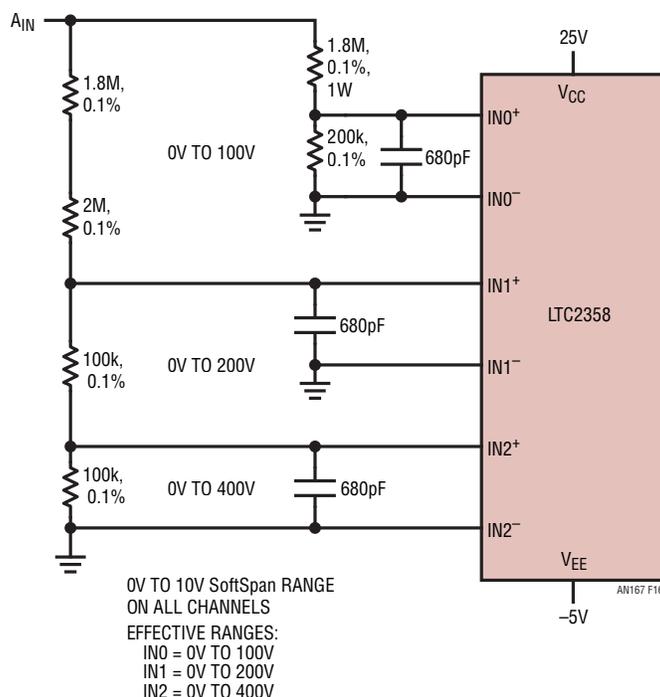


Figure 16. Automatic Gain Ranging

## 9. Double Input Range to $\pm 20V$ ( $40V_{P-P}$ ) and Increase SNR to 99dB

The very wide common mode range of the LTC2358, extending up to  $30V_{P-P}$  with  $V_{CC} - V_{EE} = 38V$ , combined with the excellent common mode rejection, 100dB minimum, make it possible to drive the analog inputs arbitrarily without degradation. For example, any two channels can be stacked in series to double the input range and improve SNR by 3dB. The output code of the two channels is added together to produce a net result with one added bit of resolution: 17 bits if LTC2358-16 is used or 19 bits if LTC2358-18 is used. Simultaneous sampling keeps the two channels synchronized at the sampling moment at the rising edge of CNV. The accuracy of the resistors does not affect the gain of the stacked combination because any extra signal presented to one channel due to a resistor match error is exactly subtracted from the other channel in the stack. The only effect of the mismatch of the  $10k + 10k$  voltage divider is that near full-scale one channel will saturate before the other. The stacked combination has a range of  $\pm 20.48V_{P-P} \cdot (1 - VDE)$ . VDE is voltage divider error in this equation, for example,  $VDE = 0.001$  for 0.1% voltage divider error. As voltage values

reach within VDE of the ideal stacked full-scale, the voltage range will have half the gain slope as one channel saturates until the summed output saturates completely as the second channel saturates. For most applications, the foreshortening of the peak-to-peak stacked range by the small attenuation error is of no consequence because the stacked gain is not affected and is the same as the average gain error of the two stacked channels.

A further expansion of the analog input range is also possible by overdriving the internal 4.096V REFBUF pin with the LTC6655-5 external 5V reference, a low noise, low drift high precision reference. The analog input range for each channel expands by the same proportion as the reference from  $\pm 10.24V$  to  $\pm 12.5V$ . With LTC2358-18 and the external 5V reference, the stacked  $\pm 25V$  range has SNR = 100dB. Refer to the Applications Information in the LTC2358 data sheet for instructions on wiring the LTC6655-5V as an external reference.

This circuit may be expanded further for wider voltage ranges by replacing the 10k resistors in Figure 17 with precision resistors, and building precision attenuators as shown in Section 8.

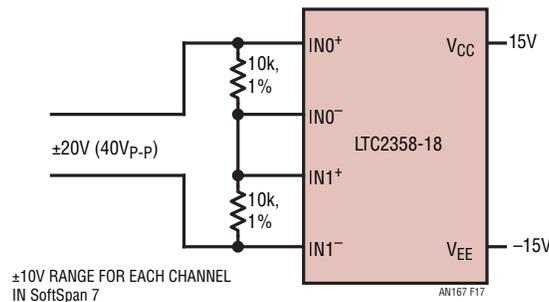


Figure 17. Double Input Range to  $\pm 20V$  ( $40V_{P-P}$ ), Increase SNR to 99dB

## FILTERS

### 10. Active or Passive Filters with $k\Omega$ Impedance Drive LTC2358 Directly

Anti-aliasing and noise filters are very common before the analog inputs of an ADC. The very high input impedance ( $>1000G\Omega$ ) picoamp analog inputs of the LTC2358 are easy to drive with a wide range of RC passive filter combinations that are easily optimized to filter the analog signal rather than to meet the stringent drive requirements of conventional unbuffered ADCs.

As recommended in the data sheet, a 680pF capacitor, or larger, at the analog input serves to absorb the very small AC transient from the sampling process that feeds back through the internal buffers at the start of the acquisition period. This maintains the DC accuracy of the LTC2358 for source impedances greater than  $10k\Omega$  that don't settle within the acquisition period. This very small glitch is also charge-conserving, which is to say that it is purely AC-coupled and the total charge of the glitch is zero and has no DC component. The external capacitor is convenient to realize simple RC filters that reduce noise from the analog signal being digitized. For example, a 33kHz low pass RC filter can be realized with  $R = 4.02k\Omega$  and  $C = 1200pF$  as seen in Figure 18. Other bandwidths can be realized with higher or lower R values, while keeping C at 680pF or higher capacitance.

When high-frequency interference in the MHz range is particularly troublesome, an additional cascaded real pole at a higher frequency can be very helpful to suppress it. The second RC filter can have a higher impedance to reduce loading on the first RC filter. The example in Figure 19

shows the first RC pole at 33kHz with 2k and 2.4nF and the second RC pole at 66kHz with 3.57k and 680nF. The loading effect pushes out the poles to 23kHz and 94kHz. An interfering 10mV tone at 1MHz is thus attenuated by 30dB with just one pole at 33kHz and by 53dB with the two poles at 33kHz and 66kHz to just 22 $\mu$ V.

When external interference frequencies approach the ADC sampling rate or when wideband sensor noise is present, a higher order filter is most effective to clean up the signal. Figure 20 compares a single pole 33kHz filter with a three-pole 33kHz Sallen-Key active filter. The steeper frequency response of the active filter very effectively eliminates the 10mV 190kHz interferer and also more effectively reduces in-band noise below 100kHz.

An active filter can also take advantage of the buffered inputs of the LTC2358. The LT1351 op amp in the Sallen-Key active filter circuit simply provides active AC feedback to shape the AC response, while staying out of the DC signal path. The op amp should therefore have good AC response near the breakpoint. Most of the contribution of the op amp feedback is near the breakpoint of the filter, such that DC voltage errors and low frequency voltage noise from the op amp stay out of the signal path. The DC input leakage current of the LT1351 op amp is 50nA maximum and drops less than 170 $\mu$ V across the 3.39k total filter resistance.

COG ceramic or film capacitors are recommended for their linearity and precision in filter applications. X7R and X5R ceramic capacitors should be avoided because they have poor tolerance and have large voltage coefficients that introduce nonlinearity and distortion into the signal path.

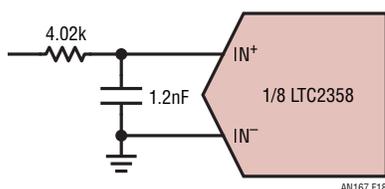


Figure 18. Single Pole 33kHz RC Filter with  $k\Omega$  Impedance

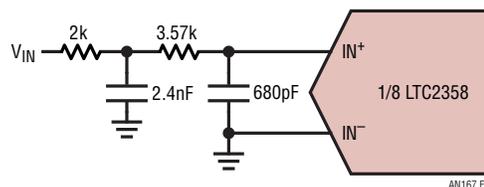


Figure 19. 2-Pole 33kHz and 66kHz RC Filter Reduces MHz Interference

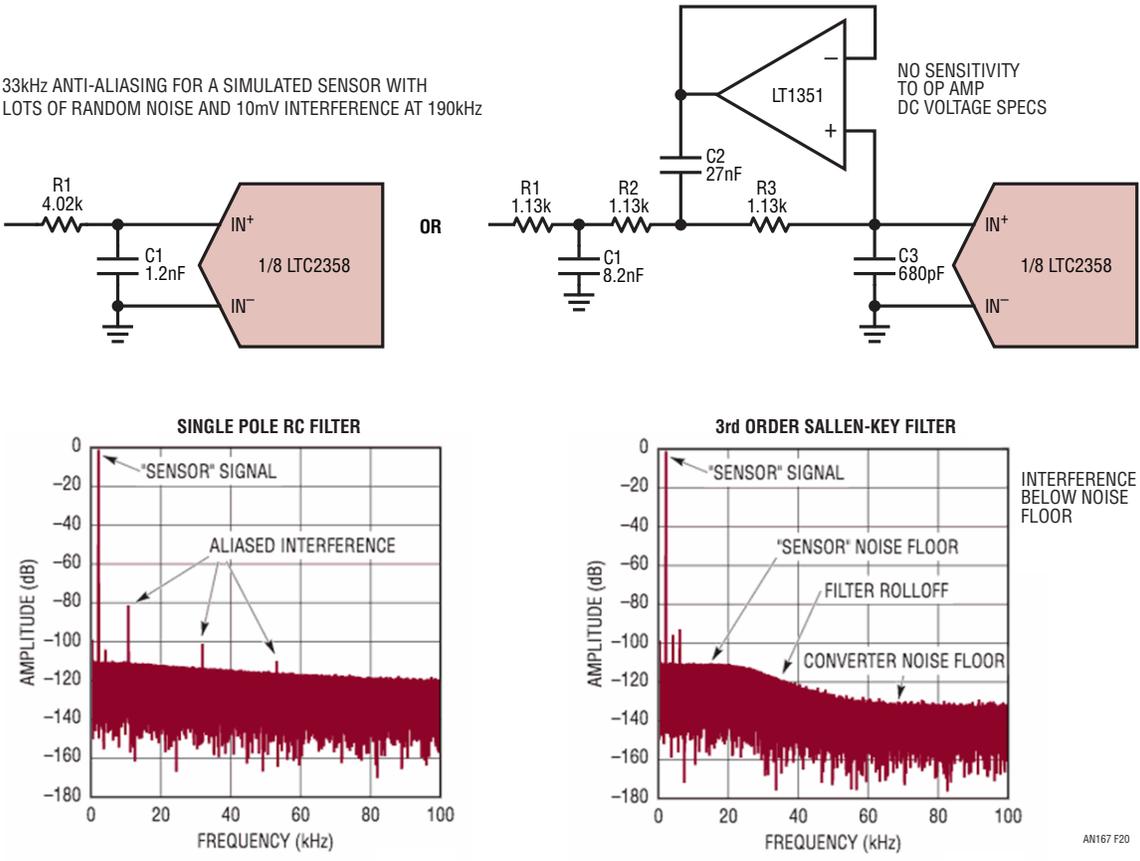


Figure 20. 3-Pole Sallen-Key Low Pass Filter Reduces Noise and Interference Above 100kHz

# Application Note 167

## 11. Flexible Inputs Simplify AC-Coupling with Single and Bipolar Supplies

The very high analog input impedance makes it easy to AC-couple signals with a classic CR high pass filter. For example, with  $V_{CC} = +15V$  and  $V_{EE} = -15V$  serving as bipolar supplies, a  $0.1\mu F$  COG or film capacitor with a ground return  $100k$  resistor realize a  $16Hz$  AC-coupling pole. At  $85^\circ C$ , the analog input bias current is  $500pA$  maximum, which contributes less than  $50\mu V$  to the analog input offset. A higher impedance CR network can be traded off for higher offset voltage from the bias current at the top end of the operating temperature range. If the application operates below  $85^\circ C$ , the analog input current is reduced approximately by a factor of  $2.2\times$  for every  $10^\circ C$  of temperature reduction. Much greater CR impedances are therefore practical at room temperature. Additional resistors may be placed in series with the analog inputs to match the net resistance seen by the plus and minus inputs to cancel out some of the effect of the increased input bias currents at high temperatures.

In single supply AC-coupled applications there is usually the need to synthesize a mid-supply node with a voltage divider and large electrolytic capacitor to bias and return the classic AC-coupling CR network resistor. The AC-coupling

circuit in Figure 21 eliminates this mid-supply node. A CR network AC-couples to  $V_{IN}^+$  and an RC network takes DC bias from the previous signal conditioning stage or sensor, which may also be powered by a single power supply. The very wide common mode voltage range, which extends from  $V_{CC} - 4V$  to  $V_{EE} + 4V$  and the very high CMRR ( $100dB$  minimum at  $200Hz$ ) make each analog channel the functional equivalent of a state-of-the-art differential instrumentation amplifier designed to handle arbitrary analog input signals with differential and/or common mode components. The  $V_{IN}^+$  and  $V_{IN}^-$  inputs of each channel may float anywhere inside the common mode voltage range ( $V_{EE} + 4V$  to  $V_{CC} - 4V$ ) without degradation. This makes it possible to implement AC-coupling in a single supply system without a mid-supply bias node.

COG ceramic or film capacitors are recommended for their linearity and precision in filter applications. X7R and X5R ceramic capacitors should be avoided because they have poor tolerance and have large voltage coefficients that introduce nonlinearity and distortion into the signal path.

Optional low pass RC filters as described in the previous note may also be placed between the AC-coupling network and the analog inputs to reduce noise and interference from the input signal.

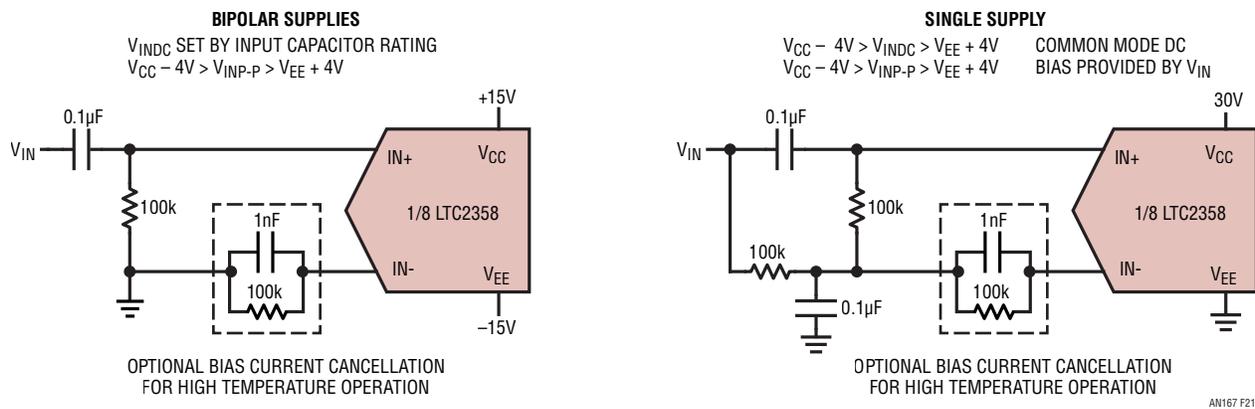


Figure 21. AC-Coupling with a 16Hz Pole

## 12. Active or Passive Notch Filters Do Not Degrade DC Parameters

One of the most common sources of interference into the analog signal path is power line hum at 50Hz or 60Hz. Notch filters can be implemented with analog circuits at the input or with digital calculations on the output data stream. The greatest advantage of the analog notch filter is to reduce large amounts of hum from the sensor or signal source down to a level that does not consume much of the input range of the ADC. Subsequently, digital filters can be applied to eliminate any remaining hum.

The very high input impedance ( $>1000G\Omega$ ) picoamp analog inputs can be driven directly by a classic passive Twin-T notch filter tuned to the hum frequency with a  $Q = 0.25$ , and relatively high impedances in the  $10k\Omega - 100k\Omega$  range. If a sharper notch with higher  $Q$  is desired, an LT1352 dual op amp may be added to sharpen the notch to a  $Q = 2.5$ . In the configuration shown in Figure 22, the low power LT1352 dual op amp provides only AC feedback within the active notch filter RC network and its effect on the DC and AC accuracy at the channel input is greatly diminished. One can think of this topology, with the op amp off to the side, as a minimally invasive active notch filter. The DC offset voltage of the op amps is completely rejected because the op amp outputs are AC-coupled to the signal path. Distortion products and noise from the op amp are also incrementally rejected by the RC filter network for frequencies away from the notch frequency. The input bias current of the first op amp causes a small offset voltage drop across the filter network.

Notch filters are generally sensitive to component values to establish the notch frequency and to attain deep rejection at the notch. This sensitivity is heightened by the higher  $Q$  of a narrower notch design. For that reason,  $Q = 2.5$  was chosen to achieve at least 20dB of rejection at the notch

frequency as a good trade-off between tolerance sensitivity and notch depth. It helps notch frequency accuracy and depth to use four identical resistors and four identical capacitors instead of a half value resistor and a double valued capacitor for the central leg of the Twin-T network. COG ceramic or film capacitors are recommended for their linearity and precision in filter applications. X7R and X5R ceramic capacitors should be avoided because they have poor tolerance and have large voltage coefficients that introduce nonlinearity and distortion into the signal path.

The free simulator LTspice® from Linear Technology can be used to help design your filter as shown in Figures 23 and 24.

Lab results from the passive and active notch filters driving the LTC2358-18 show the following results:

- Passive Notch Filter Performance
  - at 2kHz: SNR = 95.6dB THD = -108dB
  - at 61Hz (actual resonant frequency set by component tolerance) the rejection is -56dB
  - at 60Hz the rejection is -55dB
- Active Notch Filter Performance with the LT1352 Dual Buffer
  - at 2kHz: SNR = 95.6dB THD = -108dB
  - at 61Hz (actual resonant frequency set by component tolerance) the rejection is -42dB
  - at 60Hz the rejection is -35dB

The LTspice simulation in Figure 24 shows that the active filter notch is much narrower than the passive filter notch. This leaves more of the passband undisturbed with the active filter, but at the expense of potentially less rejection at the frequency of interest due to notch frequency variation from component tolerance, as evidenced by the lab results above.

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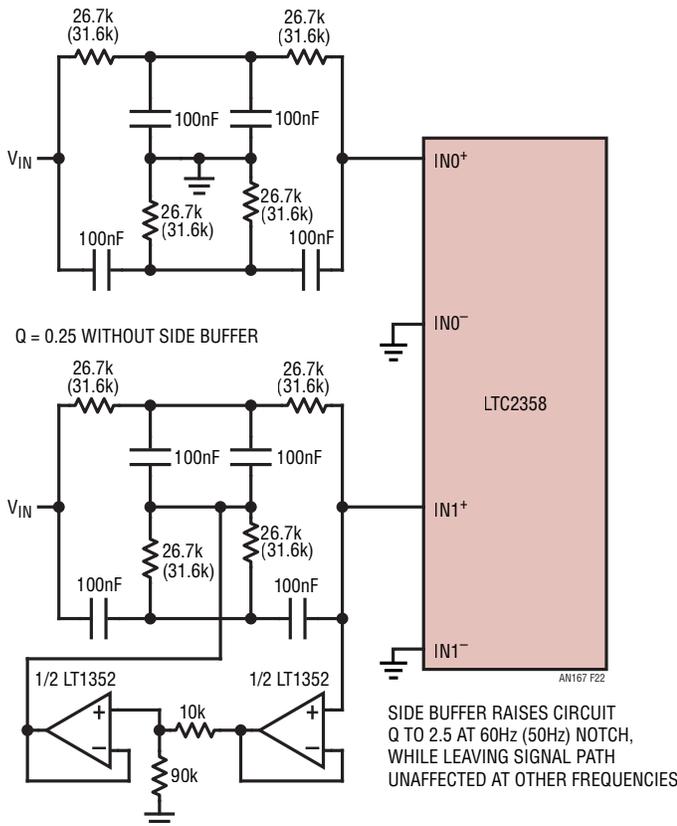


Figure 22. 60Hz (50Hz) Notch Filter with Optional Active Q Boost

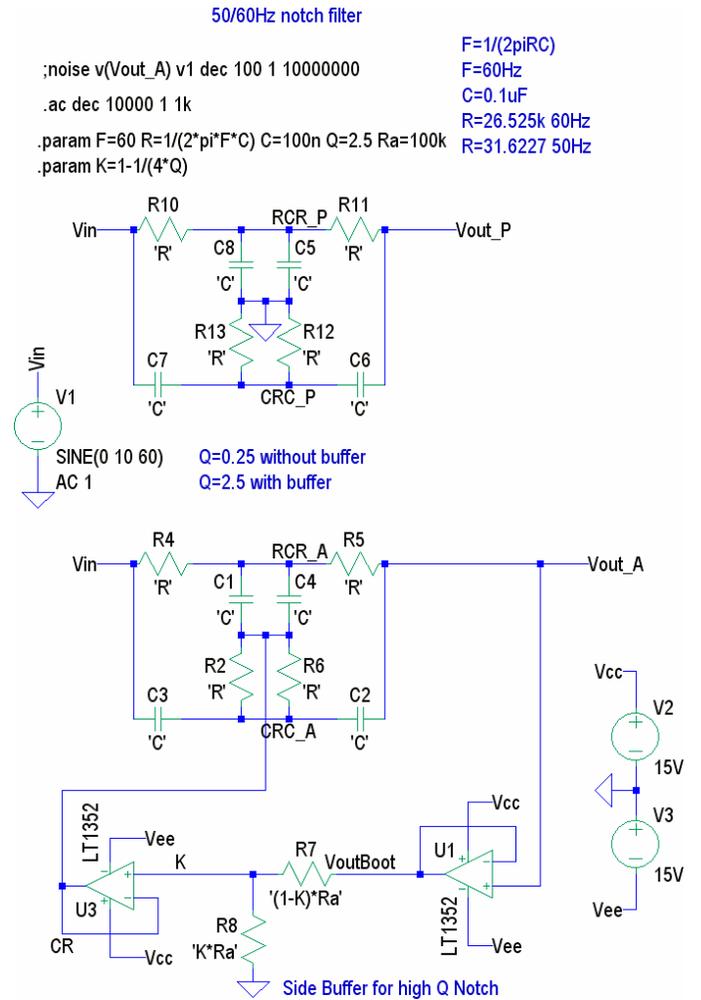


Figure 23. 60Hz (50Hz) Notch Filter with Optional Active Q Boost LTSpice Simulation Schematic

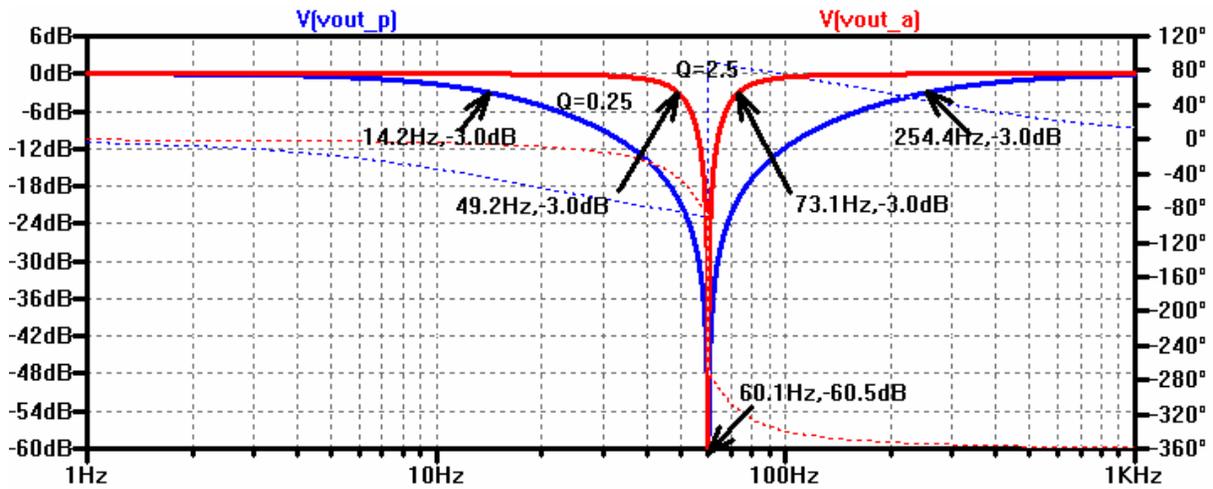


Figure 24. 60Hz (50Hz) Notch Filter with Optional Active Q Boost LTSpice Simulation Results

## SENSORS

### 13. Temperature Measurement Eliminates Thermistor Self-Heating

Thermistors can transduce temperature into relatively large current or voltage variations that are easy to digitize with little or no amplification. Thermistors with  $k\Omega$  impedance can easily drive the very high impedance buffers in the LTC2358. These relatively large currents and voltages at the thermistor also dissipate power and self-heat the thermistor, causing it to report erroneously high temperatures. For example, the Victory (VECO) 42A29 20k thermistor has a 0.013-inch diameter and a dissipation constant of  $0.09\text{mW}/^\circ\text{C}$ . This dissipation constant predicts  $2.2^\circ\text{C}$  of self-heating in still air with 2V bias. The measured self-heating was about  $2^\circ\text{C}$ .

The problem is worst for physically small thermistors that try to measure the temperature of small thermal masses, like still gasses or small objects. Victory also supplies a smaller thermistor with a 0.010-inch diameter with a dissipation constant of  $0.045\text{mW}/^\circ\text{C}$ , which doubles the expected self-heating to  $4.4^\circ\text{C}$ . Conversely, Victory's larger thermistor with 0.043-inch diameter has a dissipation constant of  $0.35\text{mW}/^\circ\text{C}$ , which predicts only  $0.6^\circ\text{C}$  of self-heating.

In typical applications, the thermistor voltage drop variation that results from temperature variations also causes different levels of dissipation in the thermistor, further corrupting the actual temperature measurement with a temperature dependent self-heating effect.

Figure 25 illustrates the self-heating effect. A simple N-channel MOSFET keeps the thermistor shorted out until the first conversion of the LTC2358. The temperature measurements were taken by the LTC2358 at a 50ksp/s rate after M1 was turned off with  $\overline{\text{READ}}$ . The self-heating effect accumulated to nearly  $2^\circ\text{C}$  over several seconds, while changing very little with each data sample every  $20\mu\text{s}$ .

The circuit in Figure 25 can also be used to make fast temperature measurements with a narrow duty cycle at  $\overline{\text{READ}}$  that greatly reduce the average thermistor self-heating effect. If  $\overline{\text{READ}}$  is kept low for  $5\mu\text{s}$  to take a measurement sample and the process is repeated every 1ms, the average self-heating effect is reduced by a factor of 200. The suggested  $5\mu\text{s}$  sampling window allows up to 40pF of parasitic capacitance at the thermistor to settle out to 18 bits with a 400ns time constant. Be sure to include any additional thermistor cable capacitance in the time constant calculation.

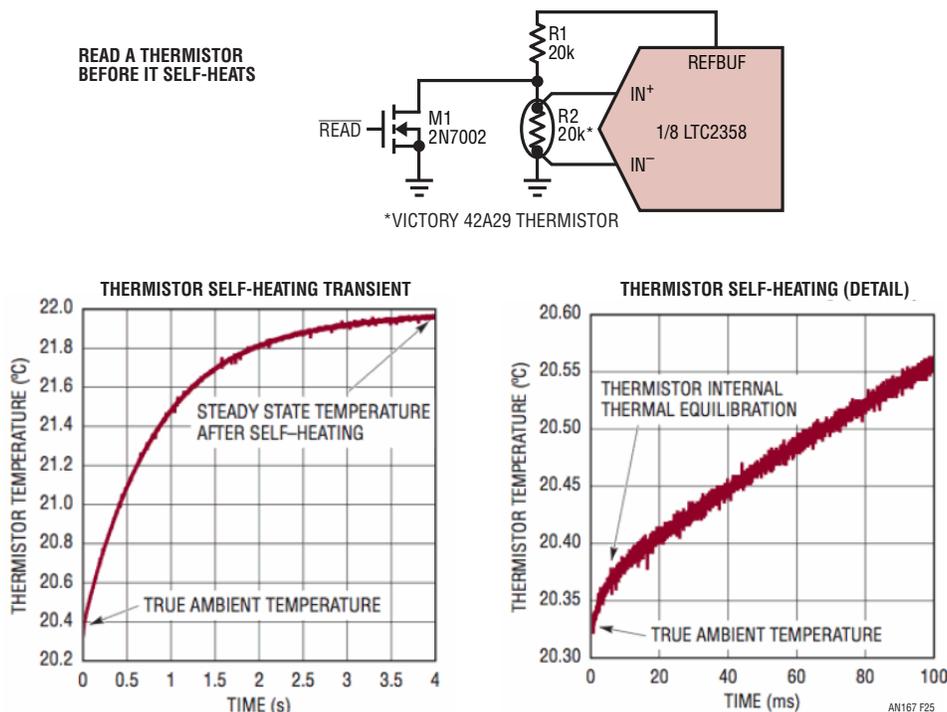


Figure 25. Temperature Measurement Eliminates Thermistor Self-Heating

# Application Note 167

## 14. Biased Photodiode Drives LTC2358 Directly

The picoamp-input analog channels of the LTC2358 can measure the photocurrent of a photodiode as a voltage drop directly across a current sense resistor that is in series with the photodiode as shown in Figure 26. A LTC6268 op amp may also be used in transimpedance configuration with the photodiode to apply a fixed voltage ( $5V - 4.096V = 0.904V$ ) across the diode. Refer to the LTC6268 data

sheet for application details in transimpedance configuration. Note that in the transimpedance circuit configuration shown, the measurement is taken directly across the current sense resistor, such that the offset voltage of the op amp is not part of the measurement. An unbuffered ADC could not be connected directly to the inverting input of the transimpedance amplifier.

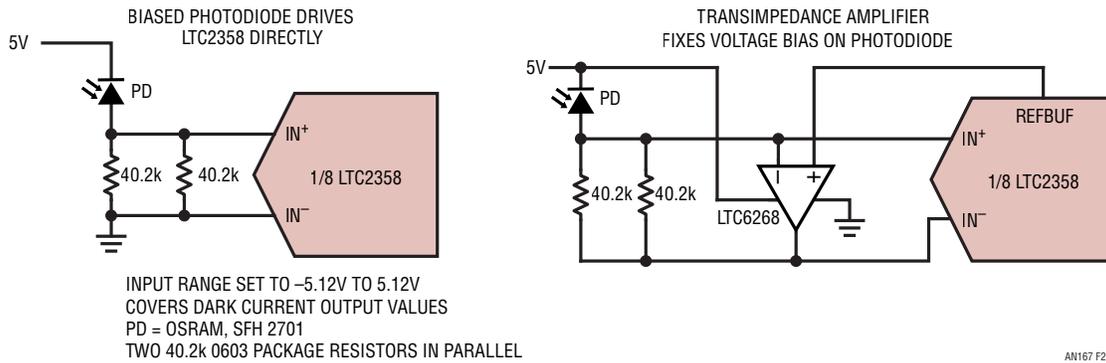


Figure 26. Biased Photodiode Drives LTC2358 Directly

## 15. Remote Sensor with Micropower Preamp Drives LTC2358 Directly

The picoamp-input analog channels of the LTC2358 can be driven directly by the lowest power micropower op amps without disturbing or loading the op amp output and without increasing the supply current of the op amp. For example, the LTC2063 op amp draws only  $1.4\mu A$  of supply current and is ideally suited for remote battery operation. Figure 27

shows the LTC2063 serving as a gain of 200 preamplifier for an oxygen sensor from City Technology. The op amp drives a twisted pair cable through an RC filter. The filter isolates the op amp from the capacitive loading of the cable and also blocks interference picked up by the cable. No RC filter is required by the ADC at its analog input, but optional RC filtering may be added between the preamp output and the ADC to further reduce external noise and interference.

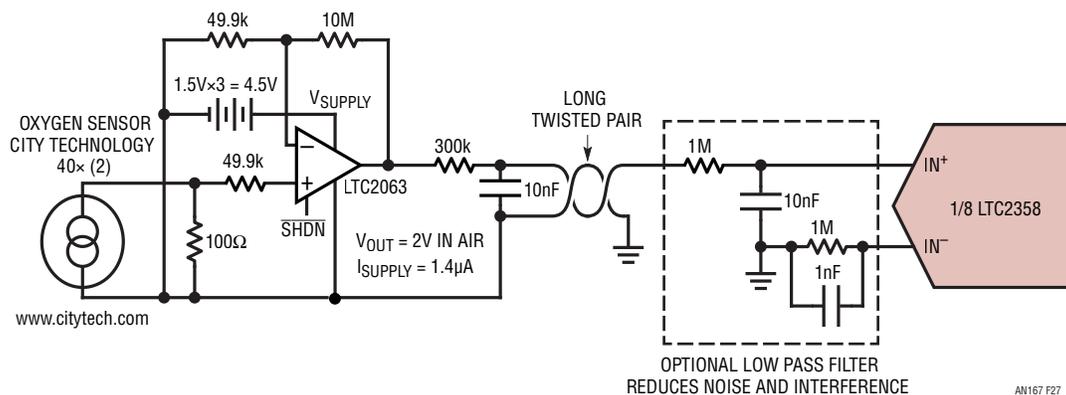


Figure 27. Remote Sensor with Micropower Preamp Drive LTC2358 Directly