Using Bipolar Preamplifiers in an LTC2411-Based Correlated Double Sampling Scheme Produces a Noise Floor of Under 100nV$_{RMS}$

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Introduction

Correlated double sampling is a common term referring to the practice of alternating applied excitation every other sample and performing a mathematical function to average the pairs of samples of opposite sign. The polarity of the samples is correlated with the polarity of applied excitation. This scheme has the benefit of canceling any fixed DC error components in the Wheatstone bridge, amplifiers and the converter, as these will alternate in polarity relative to the signal. Offset voltages and currents, thermocouple voltages at junctions of dissimilar metals and the lower frequency components of

1/f noise are virtually eliminated. Correlated double sampling has been used in high end weigh scale applications for many years.

A low noise bipolar amplifier may seem like the obvious choice as a preamplifier for low source-impedance applications such as load cells or strain gauge sensing, but their use requires special techniques. Many years ago, AC excitation of bridge circuits was the norm, as the DC accuracy of vacuum tube or early transistors amplifiers was simply inadequate. The availability of inexpensive, high precision op amps largely eliminated the practice of using

AC excitation. Putting preamplification in front of 24-bit delta sigma converters such as the LTC2410 or LTC2411 now uncovers the limitations of precision op amps. The most serious of these limitations are offset voltage drift, 1/f noise and thermocouple voltages that, in the presence of turbulent airflow, produce noise.

The use of correlated double sampling, a close cousin of AC excitation, circumvents most of the problems associated with the use of a low noise bipolar amplifier, but can often become complicated, or can easily be compromised to the extent that its use is not worthwhile.
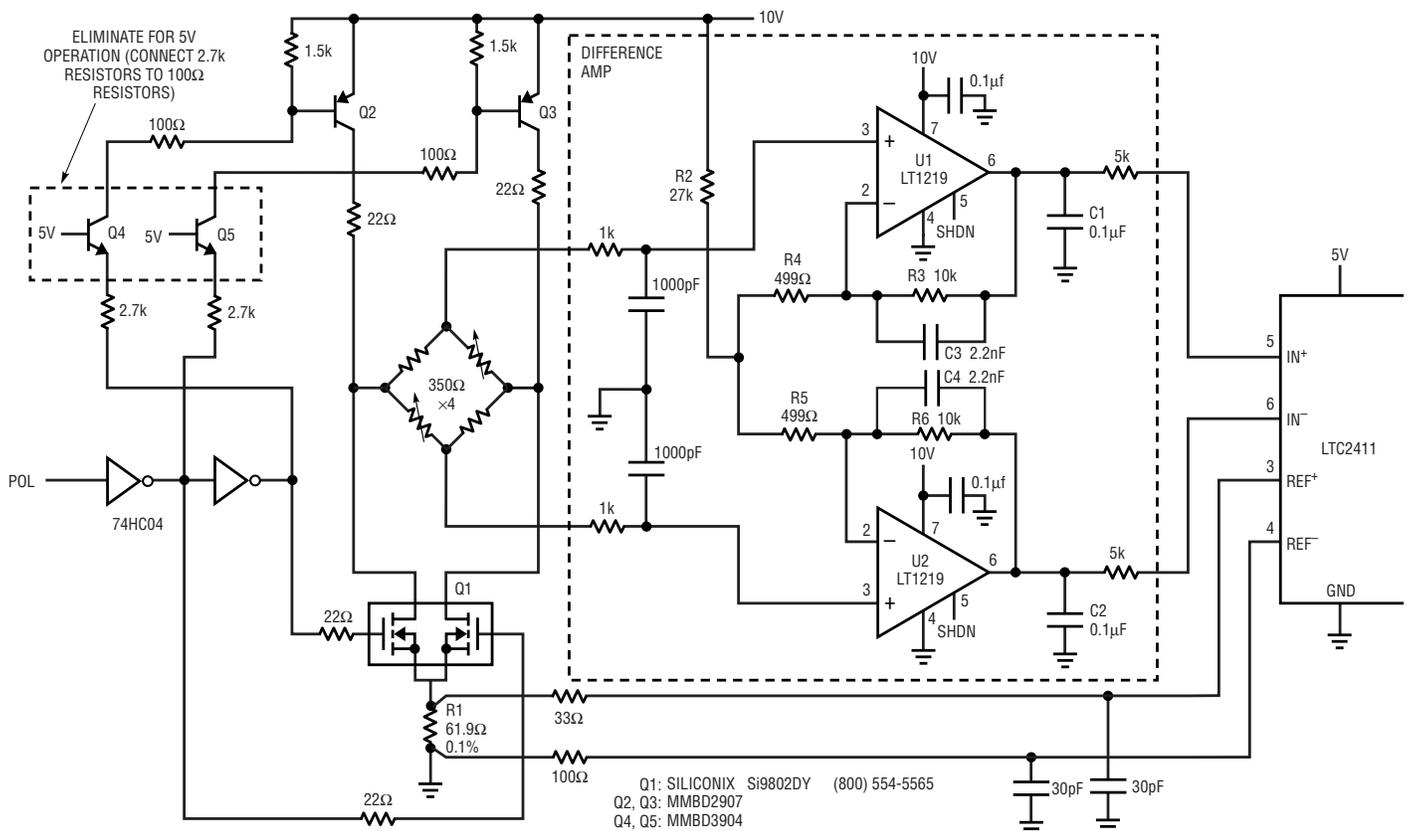

Figure 1. Correlated double sampling resolves 100nV

The circuit in Figure 1 shows a simple way of implementing correlated double sampling with the LTC2411 differential input 24-bit delta sigma converter. The LTC2411 has the virtue of being able to digitize an input voltage that is outside the range defined by the reference, thereby providing a simple means to implement a ratiometric example of correlated double sampling.

The use of a bipolar op amp as a preamplifier introduces 1/f noise, offset voltage drift, input offset current, gain error and nonlinearity. In addition, bipolar amplifiers exhibit envelope detection in the presence of the conversion spikes that are common in switched capacitor sampling converters. Bipolar amplifiers also produce envelope detection of external RFI, translating a varying RFI envelope into an erratic DC offset.

Extending the sensitivity of the front end of these converters brings into play the effects of thermocouple voltages as a source of noise due to turbulent airflow. Temperature gradients across pairs of junctions can produce tens to hundreds of $\mu V/°C$.

Correlated double sampling reduces the effect of slowly varying error components; hence, thermal baffling or encapsulation will effectively eliminate thermocouple voltages as a concern. Radio frequency interference, however, can vary abruptly, so the use of RFI suppression at the inputs of bipolar op amps is necessary at these levels. C3 and C4 in Figure 1 provide RFI suppression, as well as suppressing transient voltages during bridge commutation.

The maximum practical gain for preamplifier circuits is limited by gain accuracy/stability requirements, as closed-loop gain is approximately the product/sum of amplifier open-loop gain and the closed-loop gain. The gain stability is determined by open-loop gain variation over temperature and with supply voltage. Linearity also is affected, as open-loop gain will vary as a function of output swing. Thus, at higher closed-loop gain, there is more sensitivity to open-loop gain variation.

If the op amp's minimum open-loop gain is one million (120dB), the gain error in parts per million (ppm) happens to be approximately the same as the closed-loop gain. For example, a closed-loop gain of ten will have about 10ppm of error relative to the ideal and a closed-loop gain of 500 will have about 500ppm of gain error.

A Practical Implementation

The circuit in Figure 1 uses a bipolar amplifier (LT1219—U1 and U2) that has neither the lowest noise nor the highest gain. It does, however, have an output stage that can effectively suppress the conversion spikes. The LT1219 is a C_{LOAD}^{TM} stable amplifier that, by design, needs at least 0.1µF output capacitance to remain stable. The 0.1µF ceramic capacitors at the outputs (C1 and C2) should be placed and routed to minimize lead inductance or their effectiveness in preventing envelope detection in the input stage will be reduced. Alternatively, several smaller capacitors could be placed so that lead inductance is further reduced. This is a consideration because the frequency content of the conversion spikes extends to 50MHz or more. The output impedance of most op amps increases dramatically with frequency, but the effective output impedance of the LT1219 remains low, determined by the ESR and inductance of the capacitors above 10MHz. The conversion spikes that remain at the output of other bipolar amplifiers pass through the feedback network and often overdrive the input of the amplifier, producing envelope detection.

The wideband noise level of the LT1219 is 33nV/$\sqrt{Hz}$, seemingly much noisier than the lowest noise amplifiers. However, in the region just below the 1/f corner that is not well suppressed by the correlated double sampling, the average noise density is similar to the noise density of many low noise amplifiers. If the amplifier is rolled off below about 1500Hz, the total noise bandwidth is determined by the converter's sinc[4] filter at about 12Hz. The use of correlated double sampling involves averaging even numbers of samples; hence, in this situation, two samples would be averaged to give an input-referred noise level of about 100nV$_{RMS}$.

The amplifier bandwidth can extend to 5kHz or more without significantly changing total noise bandwidth, as long as the amplifier gain is rolled off by 6dB prior to the first multiple of the sample frequency in the LTC2411.

The LT1219 has a shutdown feature that can also be used to multiplex several channels of a similar nature into the LTC2411 with little compromise in performance. See the LT1219 data sheet for information on this feature.

The example circuit uses two inexpensive MMBD2907 PNP transistors as high-side excitation switches in an H-bridge drive scheme, and Q1, a dual MOSFET (Si9802DY), for the low-side switches. The bridge-excitation current is sensed across R1, a 61.9Ω precision resistor. The use of a MOSFET with low gate-to-body leakage allows the current through the bridge to pass through the sense resistor without significant error associated with the gate drive.

Level-shift transistors Q4 and Q5 are not required for 5V operation. If they are included they allow higher excitation voltages up to the maximum recommended for the bridge.

In the case shown, if a 10V supply is used, the excitation voltage to the bridge is 8.5V and the outputs of the bridge are above the supply rail of the ADC. These amplifiers (U1 and U2) are also used to produce a level shift to bring the outputs within the input range of the converter. This instrumentation amplifier topology does not require well-matched resistors in order to produce good CMRR. However, the use of R2 requires that R3 and R6 match well, as the common mode gain is approximately –12dB. If the bridge is composed of four equal 350Ω resistors, the differential component associated with mismatch of R3 and R6 is nearly constant with either polarity of excitation and, as with offset, its contribution is canceled due to correlated double

sampling. Note that R3–R6 must be low TC resistors if gain stability is required. If the bridge has temperature characteristics that are more significant than the temperature coefficient of the resistive elements themselves, the effective value of R1 can be modified with a thermistor or, if the temperature is measured, via a DAC.

Although this example uses 10V excitation and a 10V supply for the amplifier, the components have been chosen to allow 5V operation. If 5V operation is not a requirement, this approach can also be used with a lower noise amplifier such as the LT1124, which requires higher supply voltages. The LT1124 will not tolerate the capacitive load of C3 and C4, and these must be replaced with snubber circuits consisting of 33Ω and 1µF in series to ground. Consult the factory for recommendations on the use of other amplifiers in this circuit. The noise performance achievable with the LT1124 is on the order of 45nV$_{RMS}$.

Conclusion

The above example is just one of many ways that the LTC2411 can be used in a correlated double sampling scheme. This is a simple example but is capable of impressive results, although the implications of sensing the current through the bridge as opposed to the voltage across the bridge must be understood. Success is dependent on careful construction, although it is not as sensitive to temperature gradients as a conventional DC excitation scheme. R1 is very critical. The prototype in our lab can resolve 1 penny (2.5 grams) on a 350lb scale—in other words, 1 part in 140,000 on a load cell with a sensitivity of 1.5mV/V. *⊿7*