AN1030

Weigh Scale Applications for the MCP3551

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INTRODUCTION

There are many different types of sensors whose underlying realization is based on a Wheatstone bridge. Strain gauges are one such sensor. As a material is strained, there is a corresponding change in resistance. In many cases, each side of the Wheatstone bridge may respond to the strain by lowering or increasing in resistance (see Figure 1).

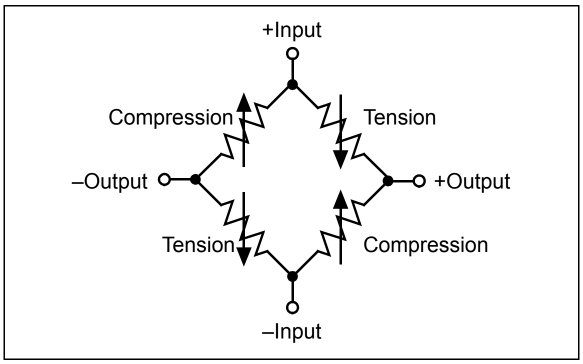

FIGURE 1: *Wheatstone Bridge of a Typical Strain Gauge.*

In the case of Figure 1, the bridge is said to be fully active. In some cases, only half of the bridge may be active (half active). For some sensors, only a single element of the bridge may change in response to the stimulus.

This application note will focus specifically on load cells, a type of strain gauge that is typically used for measuring weight. Even more specifically, the focus will be on fully active, temperature compensated load cells whose change in differential output voltage with a rated load is 2 mV to 4 mV per volt of excitation (the excitation voltage being the difference between the +Input and the –Input terminals of the load cell).

The goal is to develop a variety of circuits that can quantify this change via an analog-to-digital converter (ADC), which will be a MCP3551, 22-bit Delta-Sigma ADC. The analysis for each circuit should be applicable to other resistive bridge sensors. The different circuits will allow cost versus performance trade-offs.

The circuits presented in this application note have been realized in the MCP355X Sensor Application Developer's Board whose block diagram is shown in Figure 2. This board includes two microcontrollers. The PIC16F877 performs the basic weigh scale function while the PIC18F4550 sends data to a personal computer (PC) for analysis and debugging. The board includes a display as well as input switches that are used for calibrating the zero point and full-scale point of the load cell and for setting various processing options. Conversion results from the currently selected ADC are communicated to the PC over the USB bus. This data can be viewed on a PC using the DataView software that comes with the reference design. All of the testing and results shown in this application note were done with an MCP355X Sensor Application Developer's Board, the DataView software, and various load cells and/or load cell simulators that are either described in this document or that can be easily purchased.

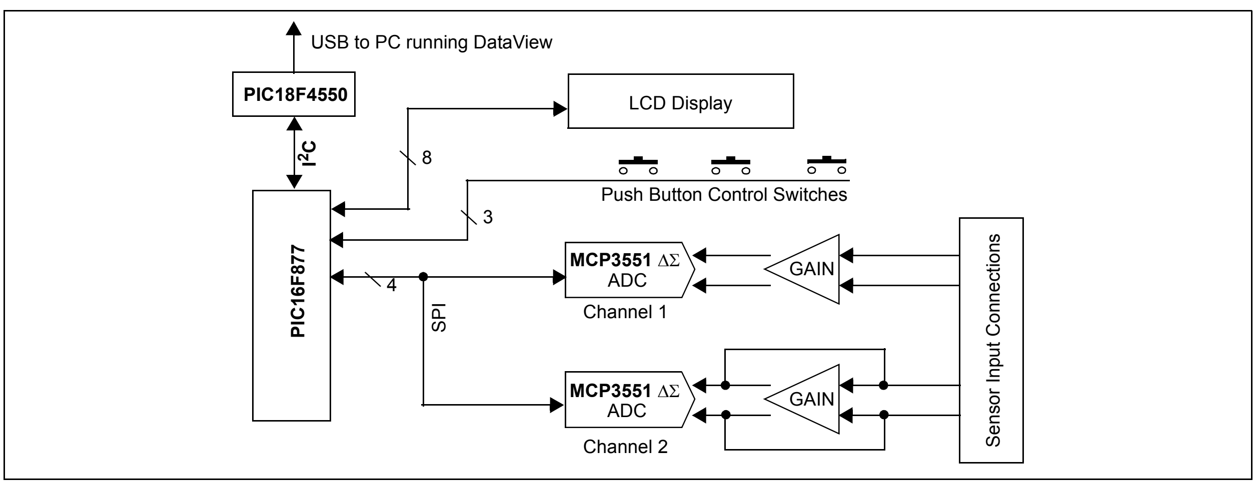

FIGURE 2: *MCP355X Sensor Application Developer's Board Functional Block Diagram.*

LOAD CELLS

Load cells come in a variety of shapes, sizes, capacities, and costs. For this application note, the focus will be on a fairly small sub-class of load cells that are fully active and temperature compensated. A temperature compensated load cell has a configuration slightly more complicated than that of Figure 1. In some cases, this means the addition of a complex series resistance at the top of the bridge that affects the voltage across the bridge as the temperature changes. The actual implementation is not important. However, it is important to realize that some load cells have definite inputs and outputs and that the input impedance may be different than the output impedance.

There are a variety of important parameters for load cells. As mentioned, the input impedance is important as well as the output impedance. In addition, it is critical to know the change in output voltage per volt of excitation, the change in output voltage versus temperature with no load, and the change in output voltage versus temperature with a full load.

Load cells have additional parameters that are critical to the final application but that are of less importance in regards to this application note. For example, load cells have a safe overload limit and a maximum overload limit. If the load exceeds the maximum overload, then the load cell may be permanently damaged.

In addition, load cells have (or may have) a linearity error specification, a hysteresis specification, a repeatability specification and a creep specification. Of course, all of these are important to the final application and define the ultimate limit of the load cell's accuracy. These parameters are only important in this application note in that they help determine the ultimate resolution required from the ADC.

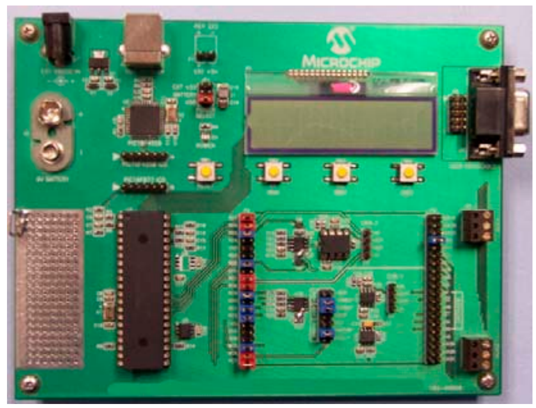

FIGURE 3: Photo of MCP355X Sensor Application Developer's Board.

Table 1 provides some specifications for a typical beam load cell intended for electronic weigh scale applications. This family of load cells has a rated capacity (RC) of 3 kg to 100 kg — the specifications are the same for all family members. Also included are the specifications for a load cell with a rated capacity of 10 kg and an excitation voltage of 5V.

TABLE 1: EXAMPLE SPECIFICATIONS FOR A LOAD CELL

Specification Description	Specification Value	10 kg Example
Safe Overload	150 %RC	15 kg
Absolute Maximum Overload	200 %RC	20 kg
Rated Output (RO)	2 mV/V ± 0.2 mV/V	9 mV to 11 mV
Non-linearity	0.015 %RO	±1.5g
Hysterisis	0.015 %RO	±1.5g
Repeatability	0.02 %RO	±2g
Creep	0.02 %RO/20 minutes	±2g
Creep Recovery	0.02 %RO/20 minutes	±2g
Excitation	12V or less	5V
Absolute Maximum Excitation	20V	—
Zero balance	±0.1 mV/V	±0.5 kg
Input Resistance	420Ω ± 30Ω	—
Output Resistance	350Ω ± 5Ω	—
Compensated Temperature Range	–10°C to 50°C	—
Temperature Effect on Zero Balance	0.04 %RO/10°C	±0.4 g/°C
Temperature Effect on Output	0.012 %LOAD/10°C	±0.12 g/°C

The specifications and values shown in Table 1 are common for temperature compensated load cells. Keep in mind that this load cell is intended for fairly precise applications and is not inexpensive. However, more expensive and more precise load cells as well as cheaper and less precise load cells are certainly available.

There are a couple of items to point out in Table 1. With a 5V excitation, the ideal full-scale output range of the load cell would be from 0V to 10 mV. This assumes the load cell is used to measure weight versus possible uses in measuring force or strain, where the output might range from -10 mV to +10 mV.

The worst-case output range would be from –0.5 mV to +22 mV. This assumes the load cell would be used in a scale that could measure up to 200% of the rated capacity of the scale. (It is recommended that the scale has an over capacity similar to that of the load cell.) It is probably not a good idea to display results up to 200% of the scale's capacity as this would encourage users to weigh items that might damage the scale. So, the maximum displayed value can be limited in software, but the circuitry should be designed to support at least 150% of full-scale and possibly even 200%.

Another consideration regarding the output range of the load cell is that the weigh scale may incorporate a pan or platform. This additional weight will always be present on the load cell. Thus, the output of the load may be several millivolts or more with no weight present. The maximum output still remains at 22 mV (200% of the rated output). The additional weight of the pan or platform will not increase the maximum output, it will simply limit the weight range of the scale (again, any load greater than 200% of the rated output may damage the scale).

It is interesting to consider some of the specifications in Table 1 in a slightly different manner (see Table 2). Rather than percent of rated output, these specifications can be given in “bits”. As an example, consider a scale that must weigh a maximum of 5 kg and display the weight in 1g increments. The resolution of the scale is 1/5000 of the maximum weight. This precision will require at least 13-bits of resolution from the analog-to-digital converter (ADC) that converts the load cell output to a digital value. While a 13-bit ADC can provide even higher resolution than is needed (1 part in 8,192), the extra resolution can be used to provide for variation in the load cell and, possibly, the weight of the pan or platform. There are reasons to consider an even higher resolution converter that will be covered later.

TABLE 2: KEY SPECIFICATIONS FROM TABLE 1 GIVEN IN TERMS OF BITS

Specification Description	Specification Value
Non-linearity	12.7 bits
Hysteresis	12.7 bits
Repeatability	12.3 bits
Creep	12.3 bits
Creep Recovery	12.3 bits
Temperature Effect on Zero Balance	14.6 bit “level” per °C
Temperature Effect on Output	16.3 bit “level” per °C

Another item of interest is that the load cell has an inherent non-linearity of approximately 13-bits. In other words, about 1 part in 8,000 (the non-linearity specification of 0.015% is 1 part in 6,667). This is also true regarding the load cell's hysteresis and slightly better than the cell's repeatability and creep (which are about 1 part in 5,000). Effectively, the load cell offers about 12-bits of performance, perhaps even a little less depending on how these errors combine. The main point here is that if we can digitize the output of this load cell to a resolution of about 13-bits to 14-bits, then the load cell will be the main limitation in the design.

There are reasons for going with even higher resolution ADCs. For example, the non-linearity of the load cell generally takes the form of a “smooth” deviation from a straight line drawn between the unloaded output voltage of the load cell and the fully loaded output voltage. Once known, this deviation can be corrected, but the mathematics involved will generally require values with resolutions greater than 13-bits.

Other specifications, such as hysteresis and repeatability, may have less concern for the final design. Hysteresis is the error that results from approaching a known weight from a lesser or greater weight. The error occurs because a greater weight may temporarily “change” the load cell more than a lesser weight. This change may be due to mechanical deformation of the load cell and/or heating induced by mechanical stress. So, when the target weight is reached (after removing some of a heavier load), the reading is different than if the weight had simply been placed on the scale (or added to the scale slowly in the case of multiple weights). This specification may not be as much of a concern for a scale where the weight will almost always be placed on the scale and then completely removed. Repeatability is similar to hysteresis and describes the variability of the scale’s reading when a known weight is measured multiple times.

Creep and creep recovery are more clearly defined specifications. A weight left sitting on the scale will result in the load cell's output voltage changing over time. The change in output voltage would ideally be zero, but practical load cells will show a small change in output voltage over many minutes (generally, the specification is given over 10 minutes or 20 minutes). For most scales, the item being weighed rarely remains on the scale for a long period of time. However, one of the reasons for the creep specification is to ensure that the load cell is "well behaved." If the load cell is not constructed properly, it is possible for the creep to be quite large and even possible for the load cell's output to never fully stabilize. Imagine a load cell made of very cheap, easily deformable material. Even after a very long period of time, the load cell may continue to deform. After the weight has been removed, the load cell might not fully recover for hours or days (if ever). The creep specifications are mainly intended to make sure that this doesn't happen.

Figure 4 provides an example of creep recovery and perhaps even hysteresis/repeatability (since these all seem to share a common root cause). A 200 kg scale, built with a 200 kg load cell, was monitored with the high-precision weigh scale circuit that will be described later in this application note. With no load, the output of the weigh scale circuit (the actual output of the MCP3551 ADC) was found to average around code 7,575. A 100 kg load was placed on the scale for 1 minute and then removed. The graph shown in Figure 4 plots the output of the load cell (as digitized by the weigh scale circuit) over the course of one hour. It takes another hour before the load cell appears to completely recover. The error shown in the graph is consistent with the specification for this particular load cell.

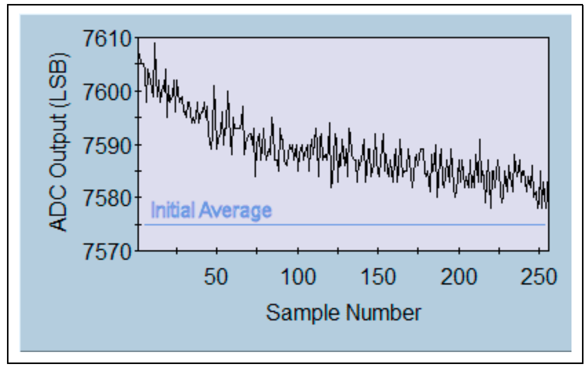

FIGURE 4: *Recovery of a 200 kg Load Cell after a 100 kg Weight was Placed On It for 1 Minute and then Removed (average output prior to weight was 7,575).*

While this specification is not typically provided for a load cell, there is a concern regarding the load cell's output noise. The reason that there is no specification for noise is that the load cell is simply a passive device and the noise is essentially the noise of a low impedance resistor (350Ω for the load cell whose specifications are in Table 1). This is such a small value that it can typically be ignored — the noise of the system will be limited by the active circuitry. For other resistive bridge sensors, the output impedance can be much higher and noise would be a concern in those cases.

THE MCP3551

There are various ways to obtain a digital value from a resistive bridge sensor and many different types of circuits have been used through the years. Recently, low-speed, high-resolution, auto calibrating delta-sigma ADCs have become popular for a variety of sensor applications, including weigh scales.

There are a number of advantages concerning delta-sigma ADCs. These include very low linearity error, low power consumption, automatic internal gain and offset calibration, ability to work with low reference voltages, and operation over a wide power supply range. In addition, delta-sigma ADCs can often be used to digitize low level signals directly, without the need for amplification of the signal.

Here are the MCP3551 Key Specifications:

Resolution	22 bits
Output Noise	2.5 µVrms
Differential Input Range	$-V_{REF}$ to $+V_{REF}$
Common-mode Input Range	$-0.3V$ to V_{DD} + 0.3V
Conversion Time	72.37 ms to 73.09 ms
Maximum Integral Non-linearity (V_{REF} = 2.5V)	6 ppm
Maximum Offset Error (25°C)	–12 µV to +12 µV
Offset Drift	0.04 ppm/°C (400 nV for V_{REF} = 5V)
Positive Full-scale Error (25°C)	–10 ppm to +10 ppm
Negative Full-scale Error (25°C)	–10 ppm to +10 ppm
Positive/Negative Full-scale Error Drift	0.028 ppm/°C (280 nV for V_{REF} = 5V)
Power Supply Voltage Range	2.7V to 5.5V
Supply Current (V_{DD} = 5V)	120 µA
Supply Current (V_{DD} = 2.7V)	100 µA

The converter's continuous auto calibration of its end-points (with no penalty in throughput) provides very low drift for both offset error and gain errors. The drift is much lower than would be seen in a successive approximation register (SAR) ADC. The linearity is better than that of a 17-bit converter and the converter's integral non-linearity (INL) is very “smooth”. This is shown in Figure 5. The fact that the INL is smooth means that over a small input range, the converter's non-linearity will be much better than the typical specification (this is not true for a SAR ADC). In addition, it is possible to characterize the non-linearity and correct for it.

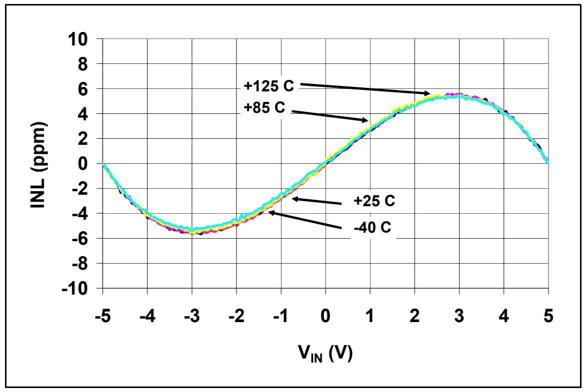

FIGURE 5: MCP3551 INL Error vs. Input Voltage (V_{DD} = 5.0V, V_{REF} = 5V).

MCP3551 Linearity

Figure 5 provides the typical INL for the MCP3551 ADC. One of the options that will be covered in detail in this application note is the possibility of using the MCP3551 for converting the output voltage of a load cell directly, with no amplification between the output of the load cell and the input of the ADC.

It was previously determined that the worst-case differential output voltage range of a load cell might be –0.5 mV to 22 mV. As an investigation, it was decided it might be of interest to measure the linearity of the MCP3551 from -6 mV to 26 mV. This span was chosen because, with a reference voltage of 4.096V, the ideal output codes for this span are from -3,072 to 13,312 for a total range of 16,384 codes or least significant bits (LSBs). So, in essence, we are looking at the MCP3551 over a 32 mV input range as though it were a 14-bit converter. The INL results are given in Figure 6 and are represented in terms of an LSB size.

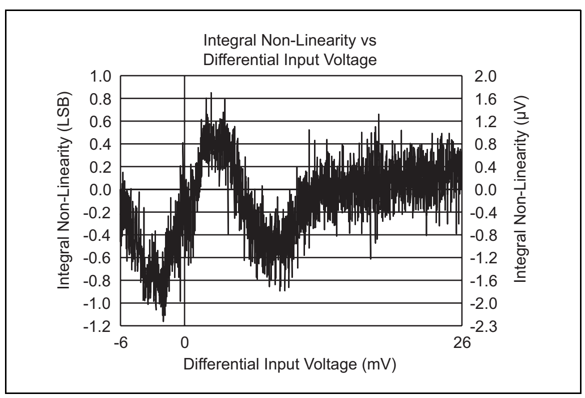

FIGURE 6: MCP3551 INL from -6 mV to 26 mV with a 4.096V Reference.

The results are “noisy” because the voltages that are being tested are very small, an LSB represents just under two microvolts. It should also be noted that the results are from a number of averages at each point that was tested.

If the only consideration was non-linearity, the results of Figure 6 show that it would be possible to use the MCP3551 as a “14-bit” converter with an input range of -6 mV to 26 mV. As will be seen, this does not make a direct connection between the MCP3551 and the load cell the best possible solution for a weigh scale. However, for some applications, it might be an acceptable solution.

As an interesting side note, the MCP3551 is a 22-Bit Delta-Sigma ADC but even higher resolution converters are available. The reader might wonder if these converters might offer better linearity than the MCP3551. Figure 7 provides the result for a 24-bit converter from another manufacturer over the -6 mV to 26 mV span. As can be seen, the results are only slightly better than those for the MCP3551. This particular device has an input range that is equal to the reference voltage, while the MCP3551 has an input range equal to two times the reference voltage. For this reason, the 24-bit device actually has 3 additional bits of resolution over the MCP3551 for the range being tested. Even with this higher resolution, the converter offers nothing extra in regards to non-linearity error for a direct conversion of the voltage output of the load cell.

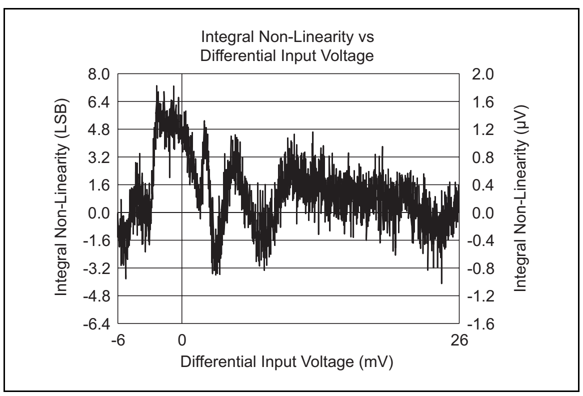

FIGURE 7: *24-Bit Converter INL from -6 mV to 26 mV with a 4.096V Reference.*

MCP3551 Input Bandwidth

The digital filter of the MCP3551 attenuates higher frequency input frequencies as shown in Figure 8.

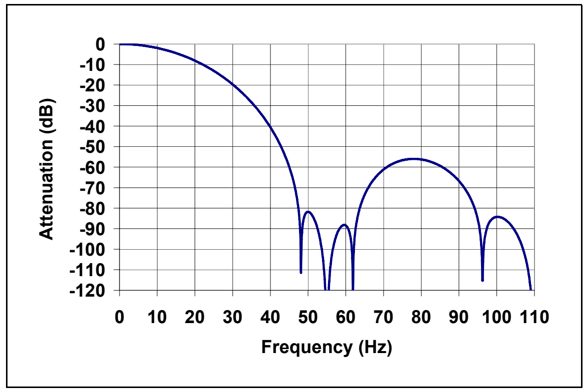

FIGURE 8: *MCP3551 Digital Filter Response.*

For a resistive bridge application, the frequency response of the ADC is usually not of great importance. The voltage produced by the sensor is mainly dependent on the excitation which also drives the reference of the ADC. If the circuit were perfectly ratiometric, it would not matter what frequencies were present. However, external signals can couple into the sensor cabling via various methods, contaminating the sensor’s output. For example, 50 Hz or 60 Hz signals from nearby power lines might couple into the signal from the sensor. As can be seen in Figure 8, the MCP3551 will reject these frequencies very effectively.

There is one very important concern regarding the input bandwidth of the MCP3551. If a signal appears at the input of the ADC that is very close to the sampling rate of the modulator, then it will alias back into the pass band of the digital filter and appear in the ADC’s output data. For the MCP3551, the modulator operates at a nominal frequency of 28,160 Hz, ±1%. Any signal that lies in this frequency range, or an integer multiple of this range, might not be fully rejected by the ADC. Fortunately, a single-pole low-pass filter with a cutoff frequency of 100 Hz to 1 kHz will generally provide enough attenuation to reject these signals.

MCP3551 Analog Inputs

A important consideration for any ADC application is the characteristics of the ADC's input circuitry. In some cases, ADCs can be difficult to drive. Their input capacitance can be large or their input impedance relatively low. Charge injection from the ADC's sampling switch can also cause the driving amplifier to ring.

Fortunately, the MCP3551 is very easy to drive. No external capacitors, either between the differential inputs or from each input to ground, are required. The differential input impedance is 2.4 MΩ, which is such a large value that a bridge sensor can typically be connected directly to the converter's inputs (though an op-amp may still be required in order to provide gain and/or filtering).

MCP3551 Output Noise

Typically, the differential output voltage of a load cell is so small that noise is a major consideration and drives a number of key decisions in regards to digitizing the sensor output. The ADC's output noise is a key factor in this.

The MCP3551's output noise is 2.5 µV RMS. This value is the internal thermal noise of the converter and is independent of reference voltage. Thus, if a "noise-free" and stable DC voltage is provided to the input of the MCP3551, we would expect to see a distribution of output codes around a mean value which represents the actual voltage input. Over a number of conversions, a histogram can be built up that represents how often each output code was observed.

Figure 9 provides a histogram of the MCP3551's output results over 16,384 conversions. This data was taken with a reference voltage of 2.5V, which means that the least significant bit (LSB) of the ADC is 1.19 µV. As a rule-of-thumb, you can multiply the converter's output noise by 6.6 in order to arrive at the number of different output codes that should be observed in a histogram derived from several thousand conversion results. This span, 16.5 µV, should have produced at least 13 to 14 different output codes. Figure 9 shows a span of 14 output codes.

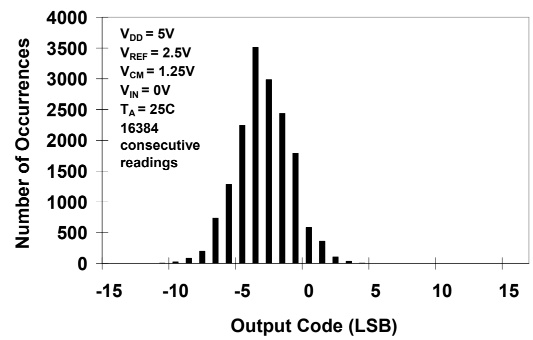

FIGURE 9: *MCP3551 Output Noise Histogram.*

The histogram of Figure 9 also provides some key information regarding the MCP3551's noise – if it is correlated or uncorrelated (random) noise. Uncorrelated or random noise should produce a Gaussian or normal distribution. Correlated noise will generally result in a different distribution – with the shape dependent on the type of noise.

Since the distribution of noise shown in Figure 9 appears to be uncorrelated, any single conversion should not be dependent on the previous result. This fact can be exploited to reduce the output noise through averaging. If two conversions are averaged, the output noise will drop by the square root of two. If four conversions are averaged, the output noise will drop by half. In general, the output noise will be:

EQUATION 1:

$$MCP3551\ Output\ Noise = \frac{2.5\mu V\ RMS}{\sqrt{N}}$$

Where: N = the number of conversions

This fact is very helpful, particularly for load cell applications. The MCP3551 is capable of 13.5 conversions per second and it is unlikely that a weigh scale will need to update its display at this rate. Two or three updates per second would probably be more than adequate. In that case, at least four consecutive conversions could be averaged, dropping the output noise of the MCP3551 to 1.25 µV RMS.

As will be shown later in this application note, this reduction in noise will apply just as well to other random sources of noise. Thus, the averaging will reduce not only the MCP3551's output noise, but noise from resistors and operational amplifiers that might be used to gain up the sensor's signal.

Ultimately, there is a limit to the possible reduction of the MCP3551's output noise. At some point, the dominant noise sources will become the correlated sources within the converter. Where that point lies is unknown – it becomes very difficult to hold the DC input steady for a long number of conversions in order to accomplish the necessary testing. In addition, there is no such thing as a “noise-free” DC voltage that can be applied to the inputs of the converter. This is true even if the inputs are tied together and directly to “ground.”

While the point where correlated noise might become a concern is unknown, it is certainly possible to consider averaging 16 or even 32 conversions to reduce the output noise of the converter and have the results match those predicted by Equation 1 very closely. Sixteen averages would probably be the limit for any weigh scale applications as the display would be updated just over once per second. However, updating the display with intermediate results while building up 32 or even 64 conversions to average for a final “settled” reading is certainly a possibility.

MCP3551 Reference Input

Assuming a non-ratiometric application, the reference input of the MCP3551 does not reject low frequency signals below 10 Hz. These simply pass through the converter as though the signal was present (at twice the amplitude) across the converter’s inputs. However, for ratiometric applications, low frequency signals on the reference will also impact the differential output of the sensor and will not impact the converter’s results.

For higher frequency signals at the reference input of the ADC, there are two important considerations. One is reference feedthrough associated with signals and noise in the 1 kHz to 10 kHz (and above) frequency range. This will be discussed in the next section. The other is noise in the frequency range of 10 Hz to 100 Hz that is not being cancelled by the ratiometric configuration for one reason or another (there is also concern for any signals or noise whose frequencies are near integer multiples of the modulator rate as these alias back into the pass band of the digital filter).

In a ratiometric application, the lower frequency noise will generally cancel. It will be much more difficult for higher frequency noise to cancel due to various phase shifts associated with the sensor such as cabling capacitance. However, even low frequency signals and noise will not cancel completely.

The main consideration for noise in the 10 Hz to 100 Hz range is that any noise that is not cancelled by the ratiometric configuration will impact the output result only as percentage of the output reading.

For example, consider a very low cost application where the MCP3551 reference input will be connected to the +5V USB Bus power on a personal computer (PC). This power will also drive the bridge sensor (this actual application will be looked at in more detail later in this application note). Anyone with any experience with PC power supplies would expect the USB Bus power to be very noisy. However, the ratiometric application will help cancel a good deal of the noise.

The low frequency noise that’s left (mostly below 100 Hz) will affect the conversion result of the ADC only as a percentage of the input voltage. The ADC has a differential input range that is $\pm V_{REF}$. If the input voltage is half of V_{REF}, then less than half the noise on V_{REF} will appear on the output data (the noise would be half and then there is some rejection by the digital filter). If the input voltage at the ADC’s inputs is 0V, then there will be no impact on the output result of the ADC regardless of the amount of noise (within reason).

This fact has an important impact on the overall design of the weigh scale. If noise may be present on the reference input of the ADC, then the impact of this noise on the performance of the system can be minimized by using the smallest possible input range of the ADC and making sure this range is located near 0V.

So, if the voltage output of the sensor is small and must be gained up, then the smallest amount of gain should be used and no more. If the signal is gained up too much, then there is increasing risk that other noise sources may contribute errors. Obviously, this risk can also be lessened by using a very low-noise source to drive the reference and bridge. However, that may increase the cost of the final design.

MCP3551 Reference Feedthrough

The reference input of the MCP3551 differs from the ADC input in yet another way – it does not completely reject higher frequency signals. On first consideration, this might not seem that important, and, in general, it is not. The component providing the MCP3551's reference voltage should offer good performance, be located nearby, and should be reasonably immune from potential contaminating signals such as 50 Hz or 60 Hz power and even higher frequency sources of noise.

However, it turns out that references and regulators may produce fairly significant noise in the 1 kHz to 10 kHz frequency range. The total RMS voltage of this is typically not significant, but it might be as much as several hundred microvolts. The reference of the MCP3551 will not completely reject this noise as can be seen in Figure 10. This graph shows the feedthrough of signals on the MCP3551 reference input to the digital output results over the frequency range of 100 Hz to 10 kHz.

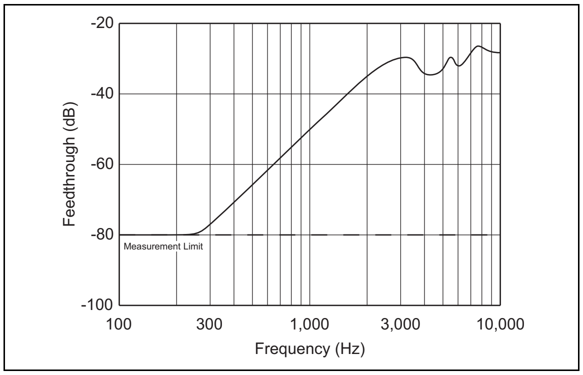

FIGURE 10: *MCP3551 Reference Feedthrough.*

An example is in order to fully explain the issues implied by the graph of Figure 10. Assume that a 3 kHz, 100 µV RMS signal is present, along with the reference voltage, at the reference input of the MCP3551. The 3 kHz signal would be attenuated by approximately 30 dB. This attenuated signal does not alias down into the pass band of the ADC. That is, a power spectrum of the converter's output data will not show a discrete tone present. Instead, the signal simply results in an increase in the converter's overall noise floor. Thus, a discrete 3 kHz, 100 µV RMS signal will add an additional 3.16 µV RMS noise to the total output noise of the MCP3551, increasing it from 2.5 µV RMS to 4.03 µV RMS.

Thus, higher frequency signals and noise present at the reference input of the MCP3551 will result in an overall increase in the converter's output noise. This can present a particularly difficult situation to debug during the development of a bridge sensor application.

It is also important to keep in mind that the reference feedthrough shown in Figure 10 occurs regardless of the voltage at the input of the ADC. As was described in the previous section, MCP3551 Reference Input, lower frequency signals or noise on the reference voltage (those in the 10 Hz to 100 Hz range) only impact the output of the converter as a percentage of the input voltage (and only for that portion of the signal that gets through the digital filter). For reference feedthrough, this is not the case. Feedthrough will occur even if the input voltage is 0V (there is a very small change in the feedthrough as a result of the input voltage, but the overall shape of the graph is not substantially affected by it).

Figure 10 provides important information for making either an informed decision regarding the source of the reference voltage or important design decisions about how to handle the issue. If the reference voltage for the MCP3551 is sourced by a very low-noise, well-behaved source, then there should not be enough noise in the 1 kHz to 10 kHz range to matter. However, such devices are typically more expensive. Another solution is to filter the reference voltage and to eliminate the higher frequency noise. This works extremely well but causes other considerations, particularly regarding a ratiometric application. The problems introduced by filtering the reference voltage will be covered later in this application note.

One final comment regarding Figure 10 is that this issue is not unique to the MCP3551. The lack of rejection of higher frequency signals appears to be a limitation of the typical delta-sigma design used throughout the industry. Figure 11 provides the reference feedthrough for a competing 24-bit delta-sigma ADC.

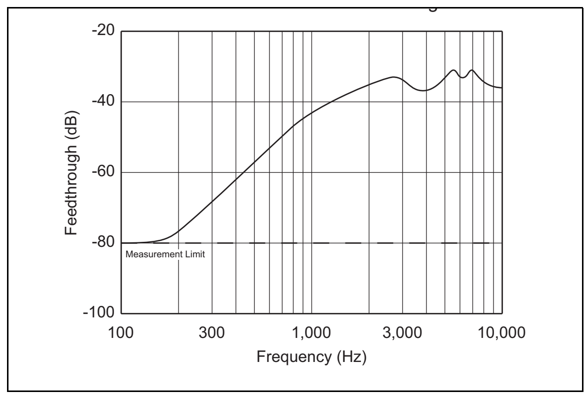

FIGURE 11: *Reference Feedthrough for a Competing 24-bit ADC.*

A BASIC RATIOMETRIC WEIGH SCALE

Figure 12 provides a block diagram of the basic weigh scale circuit that will be discussed in detail in this application note. This is not necessarily the recommended circuit, but simply serves as a starting point.

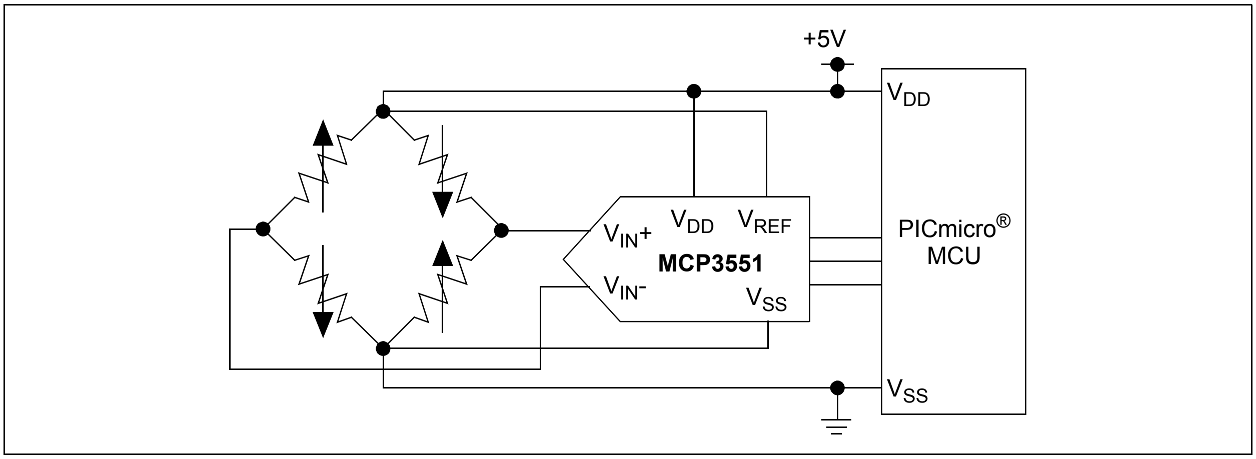

FIGURE 12: *Block Diagram of a Basic Weigh Scale.*

In the block diagram of Figure 12, a 5V source is used to provide power to a PICmicro MCU, the load cell, and the MCP3551. This 5V source also provides the reference voltage to the MCP3551. The LCD display and USB interface to the PC that is present on the MCP355X Sensor Application Developer's Board is not shown.

The diagram also shows that both the converter's ground pin (V_{SS}) and V_{REF} pin should be connected across the load cell as directly as possible. Cabling may make this difficult but some load cells contain sense connections that can be used to make the connection as is shown in the diagram.

We can start a basic analysis of this circuit by looking at what is meant by "ratiometric." The goal of a ratiometric circuit is to ensure that the output of interest (in this case, the output voltage of the load cell) is a strict ratio of the excitation. As the excitation changes, the output changes as well in order to maintain the ratio.

For Figure 12, this concept includes the ADC by making sure the excitation voltage is also the converter's reference voltage. In this way, the ADC is offering a digital value that represents that ratio of its input voltage as compared to its reference voltage.

As an example, assume that the load cell output is 1/5 of the excitation voltage or 1V differential. Ideally, for this input voltage and with V_{REF} = 5V, the MCP3551 would output a digital value that is 1/5 of its full-scale digital value or 419,430.

If the 5V power source were changed to 6V, the output of the load cell would change to 1.2V. This would still be 1/5 of V_{REF} and the MCP3551 would still output the result 419,430. This is the beauty of a ratiometric circuit—a stable reference voltage is not necessary as it would be for many analog-to-digital converter circuits.

This discussion can be expanded to also look at the elegance of the bridge itself. Not only does it provide an output voltage that directly scales with excitation voltage but the common-mode output also scales. For example, if the load cell is under no stress, then both outputs are typically at 2.5V with a 5V excitation voltage. With a 6V excitation, both outputs are at 3V. In both cases, the outputs are at half of the excitation voltage.

Even if the MCP3551 V_{DD} supply did not change with excitation voltage, the converter has more than enough common-mode rejection to reject a change on both its inputs from 2.5V to 3V without a resulting change in the digital output code (common mode rejection at DC is typically -135 dB). However, since its V_{DD} supply will also change, the common-mode voltage at the input of the ADC remains at 1/2 of V_{DD}.

Thus, the ratiometric configuration of the ADC and the load cell provide excellent common-mode and normal-mode rejection when considering what actually happens at the input of the ADC.

THE DIRECT-CONNECT WEIGH SCALE

At this point, there has been enough discussion of the various aspects of the load cell, the MCP3551, and the basic ratiometric weigh scale circuit to actually try it out. Figure 13 provides a slightly expanded circuit over that of Figure 12.

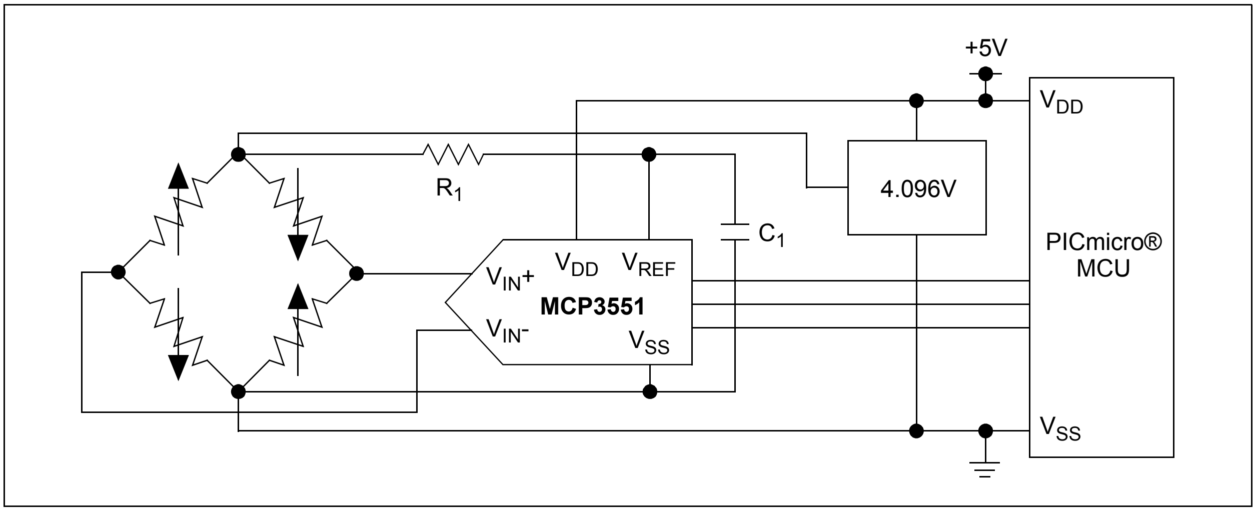

FIGURE 13: A Direct-connect Weigh Scale.

The circuit shown in Figure 13 was actually tested with two different 4.096V references: a National Semiconductor® LM4140 and an Analog Devices REF198. All of the tests that follow were done on these two variations of Figure 13 as well as the circuit configurations shown in Figures 14 and 15.

The circuit of Figure 13 can be implemented on the MCP355X Sensor Application Developer's Board when connected to the PC using USB power. Since the USB interface provides +5V power, there was interesting opportunity to compare the performance of several options regarding this circuit. One option was to connect the load cell directly across the +5V power from the USB interface (see Figure 14). Another variation was to drive the load cell from one or two pins of the PICmicro MCU that were configured as outputs and set high (see Figure 15).

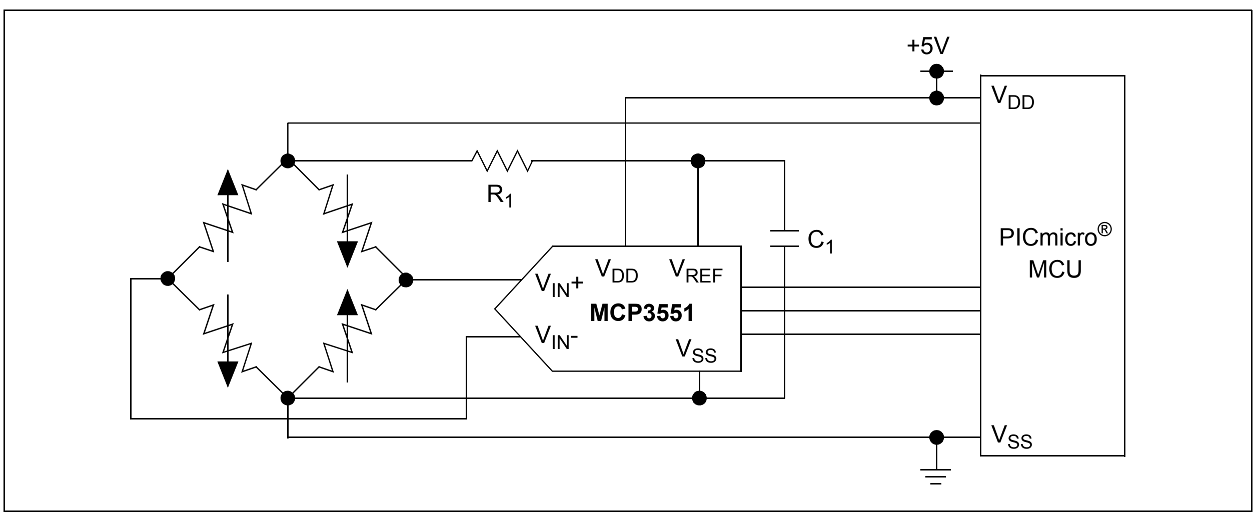

FIGURE 14: A Direct-connect Weigh Scale with the Load Cell Driven by +5V USB Power.

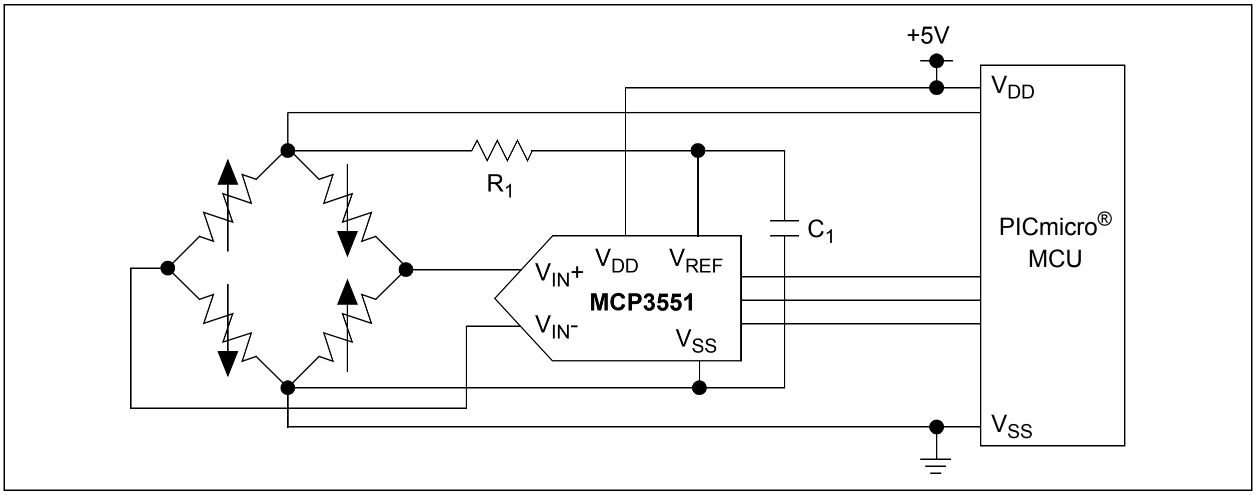

FIGURE 15: *A Direct-connect Weigh Scale with the Load Cell Driven by the PICmicro MCU.*

The circuit of Figure 15 allows for a microcontroller to easily turn the power off to the load cell in order to reduce power consumption. The power consumed by a load cell is not trivial. With a 350Ω bridge configuration and a 5V excitation, the power consumed would be 70 mW (the load cell requires 14 mA of current).

Note that the MCP3551 power is not supplied by the PICmicro MCU. Instead, it is simply connected to the 5V source directly. Such a connection is definitely recommended as the MCP3551 powers down to less than 1 µA of current when not converting, so it is not necessary to turn off its power. In addition, there is a possibility that the load cell voltage might be as low as 4V due to the PICmicro MCU's internal output impedance. While the MCP3551 could easily operate from such a voltage, other digital outputs associated with the serial interface could potentially turn on the ESD diodes inside the converter.

At first, it might seem a little unusual to drive both the load cell and the converter's reference voltage from the digital output pin of a microcontroller. What is really happening is that the load cell and MCP3551's V_{REF} pin are being connected to the 5V supply through a FET switch whose on-resistance is typically in the 30 to 50Ω range. The on-resistance of this switch will change with temperature and so the output voltage of the pin will also change. However, this is a ratiometric application and the change should not be a concern, though testing will reveal if that is true.

The next step is to consider the practicality of digitizing the output of the load cell directly with the MCP3551. The goal is to use conservative numbers without going overboard. From Table 1, the smallest output range of the load cell will be from 0.5 mV to 9 mV for no load to a full-scale load, respectively. The FET switch at the digital output pin of the microcontroller should have no more than 50Ω of on-resistance (if it does, it is possible to use two pins in order to get half the on-resistance). The resistance of the load cell will vary only a few percent or less, so the typical input impedance of the load cell is good enough. This means that 5V will drive 400Ω total for a current of 12.5 mA. Thus, the MCP3551 will see a reference voltage at its V_{REF} pin of 4.375V.

The LSB size of the MCP3551 will then be approximately 2.1 µV. The output span of the load cell covers 4,074 codes. With this simple analysis, it appears we could digitize the output of the load cell to roughly 12-bits and the INL data shown in Figure 6 provides enough information to be comfortable that the result will be within ±1 LSB of the correct number (based on calibration of both the zero and the full-scale points of the scale).

Unfortunately, the output noise of the ADC predicts that any single conversion would only be within ±4 LSBs. This has reduced a single result to something closer to 10-bits of precision. If four consecutive conversion results could be averaged, then the result would be in error by only ±2 LSBs, a gain of 1-bit to roughly 11-bits of precision (see the **“MCP3551 Output Noise discussion”** for more information regarding averaging).

A similar analysis can be done for circuits shown in Figures 13 and 14. In the case of Figure 13, the V_{REF} pin of the ADC will see a voltage of 4.096V which will produce an LSB size of 2.0 µV. For Figure 14, the reference voltage will be at approximately 5V and the LSB size will be 2.4 µV. These values will not result in substantial changes to the error analysis that has just been done for Figure 15.

The main point of the discussion so far is not that any of circuits shown in Figures 13 14 and 15 are necessarily a good starting point for a weigh scale, but to simply go through the exercise of considering the performance of such circuits. A reasonable estimate of the performance of Figure 15 has been developed, but will the actual results match? In addition, is there a penalty to be paid for driving the load cell and MCP3551 reference input with the digital pin of a microcontroller or will using a good reference produce better results?

As a starting point for testing, it should be noted that R_1 was set to 10Ω and C_1 was set to 0.1 µF (these two form a low-pass filter on the reference voltage with a cutoff frequency of 160 kHz). These values were chosen as “typical” values that might be used as starting point by someone unfamiliar with the intricacies of weigh scale design but reasonably familiar with mixed-signal design.

A Note About Testing

There are a few items that help greatly in the development and testing of a weigh scale circuit. First, it is essential to be able to get the raw ADC data directly into a PC for analysis. For the testing involved with this application note, the test board included not only a PICmicro MCU but also another microcontroller that communicated the raw ADC data to a PC via the USB bus. This data was analyzed and displayed by Microchip's DataView software using the MCP355X Sensor Application Developer’s Board. Nearly all of the tests results shown in this application note were generated by this software.

Second, it is very good idea to buy or build a “load cell simulator.” For the testing involved with this application note, two different load cell simulators were built, each on a small printed circuit board that plugged onto the test board. One that simulated a 350Ω load cell with no load (0V differential output) and another that simulated a 350Ω load cell with a worst-case load 25 mV to simulate a load of 250% of rated output.

It would a big mistake to build these simulators from standard resistors. The temperature coefficient of resistance (TCR) matching between the resistors of a high quality load cell is incredibly good-on the order of 0.1 to 0.01 parts per million. Making a simulator out of resistors with a 100 ppm TCR will allow only the most rudimentary testing. At the very least, use resistors with a TCR of 25 ppm and be prepared to cover the test board with a towel. Resistors with TCRs as low as 0.2 ppm are available. While such resistors can not be obtained very easily or cheaply, the extra effort and expense may well be worth it in the end.

Finally, it would seem that testing the weigh scale circuit with an actual load cell would be ideal. Unfortunately, load cells, particularly those in the 10 kg range or less, tend to act as excellent seismic detectors. Any bumps or even air currents will cause the output to show significant variations, making it impossible to determine the actual performance of the underlying circuit. Testing with the actual load cell is certainly necessary at some point, but get the kinks worked out first with the load cell simulators.

Before testing the direct-connect weigh scale circuit, it is necessary to define some test methodology and standardize on a manner for presenting the results. The DataView software reports noise in terms of parts-per-million (PPM) RMS of the converter's full-scale digital output range (2^{22}). Thus, output noise is really given in terms of LSBs, where one PPM = 4.2 LSBs.

Unfortunately, the DataView software does not know what the actual LSB size of the converter is because it does not know the value of the MCP3551's reference voltage. However, a result given in terms of PPM of digital full-scale is actually very useful. It makes it easy to compare precision (or resolution) regardless of the reference voltage.

On the other hand, having a result in terms µV RMS is also very useful when trying to track down noise sources and analyze results. In general, both results will be presented. Simply keep in mind that it is necessary to know the value of the MCP3551's reference voltage in order to convert from one unit to the other.

As a quick review, there were four variations of the circuits shown in Figures 13 14 and 15. In the circuit shown in Figure 13, the National Semiconductor LM4140 4.096V reference is used to source the excitation voltage for the load cell and the MCP3551 reference input. The same circuit was also used but with an Analog Devices REF198, which is also a 4.096V reference. Both of these references are good, reasonably inexpensive references. The third configuration ties the excitation voltage and the MCP3551 reference input to the 5V source directly (see Figure 14). This 5V source is the USB power from a laptop computer. This source is moderately low-noise for a computer supply but has significantly higher noise than either of the two references. It should be noted that a higher noise USB power supply was found on a desktop computer and that point will be discussed in another possible circuit configuration later in this application note. The final circuit matches the configuration of Figure 15, with the excitation voltage of the load cell and MCP3551 reference input coming from the PICmicro MCU. Again, the 5V source was USB power from the same laptop computer.

In some cases, a result will be shown that really is much more qualitative than quantitative, but is still very interesting. For many of the test configurations a 5g step will be shown. This test was done with the actual load cell and shows the output data of the ADC when 5g was placed on a 5 kg load cell. This step would be one-thousandth of full-scale. Note that the step always occurs in the center of the data display.

Now, on to the testing. It should be noted that in all four test results that follow, the PICmicro MCU was present and active but was otherwise not involved in collecting data (data was being collected by the USB microcontroller). In most cases, testing involved using a load cell simulator whose differential output was 0V (the common-mode voltage of the two outputs was approximately one-half the difference of the voltage across the load cell).

The LM4140 device is a 4.096V reference and its actual output voltage measured approximately 4.09V. With no other sources of noise, the DataView software should have reported an output noise of 0.31 PPM. The REF198 output voltage was closer to 4.096 but the resulting output noise would still be 0.31 PPM. The USB power was not exactly 5V, but close enough that DataView should have reported an output noise of close to 0.25 PPM for the last two tests. Table 3 provides the quantitative test results.

TABLE 3: RESULTS OF TESTING THE DIRECT-CONNECT WEIGH SCALE WITH R_1 = 10Ω AND C_1 = 0.1 µF.

Load Cell & MCP3551 V_{REF} Source	Output Noise	
	PPM of FS	µV RMS
LM4140	0.83	6.8
REF198	1.23	10.0
USB +5V Power	3.12	31.2
PICmicro MCU (powered by USB +5V Power)	3.23	32.3

Well, that certainly is not very good at all! Even with the 4.096V references, the results are not nearly as good as predicted. Still, there is a clue in the data that perhaps noise is playing a role, assuming that the USB power has more noise than either of the references.

An audio spectrum analyzer was used to measure the noise of the two references and the USB power. This revealed some interesting results. The USB power certainly showed higher noise than either of the references, but both references showed higher noise and at higher frequencies than was expected. Various bypassing schemes were attempted for the references, but the noise could not be lowered. These schemes also did nothing to address the USB power issue.

The power spectrums of the references and the USB power were analyzed in terms of the reference feedthrough shown in Figure 10. It was certainly possible that the noise on the MCP3551 V_{REF} pin could be affecting the digital data. It was decided that substantially decreasing the cutoff frequency of the lowpass filter on the V_{REF} input of the MCP3551 might help decrease the noise.

Filtering the V_{REF} input creates two potential problems. In one case, it introduces a phase delay between the excitation voltage of the load cell and the reference input of the MCP3551, potentially reducing the ratio-metric cancellation achieved by deriving both from a common source. In addition, variation in R_1 with temperature can create a gain error because the reference input has an equivalent input impedance of approximately 2.4 MΩ (this value also changes with temperature). The load cell has a finite gain error associated with it, so the goal is to make sure that gain error due to R_1 is similar to or even smaller than the load cell's gain error.

On the other hand, the cutoff frequency of the filter must be low enough that noise at the reference input of the MCP3551 in the 1 kHz range and above will not contribute significantly to the converter's output noise. Since the filter is a single pole filter, it must start to roll off significantly below 1 kHz in order to offer any substantial attenuation of noise above 1 kHz.

As a first pass, it was decided that R_1 would be changed to 332Ω and C_1 would be changed to 10 µF. The cutoff frequency of the modified lowpass filter is now 48 Hz. Hopefully, this is high enough that the ratiometric relationship between V_{REF} and the load cell's excitation voltage will not be broken while still offering good attenuation of higher frequency noise at V_{REF} pin of the MCP3551. Worst-case analysis shows that a 332Ω resistor for R_1 will produce less gain error with temperature than that of the load cell even assuming we were to use the full-scale input range of the converter. (The goal was to come up with a circuit that would be usable for all configurations, not just the direct-connect case.)

Table 4 provides the results for the modified circuit – a substantial improvement for all configurations.

TABLE 4: RESULTS OF TESTING THE DIRECT-CONNECT WEIGH SCALE WITH R_1 = 332Ω AND C_1 = 10 µF.

Load Cell & MCP3551 V_{REF} Source	Output Noise	
	PPM of FS	µV RMS
LM4140	0.28	2.3
REF198	0.27	2.2
USB +5V Power	0.23	2.3
PICmicro MCU (powered by USB +5V Power)	0.26	2.6

The results are really quite clear. In all cases, higher frequency noise present at V_{REF} was raising the output noise of the converter. Lowering of the cutoff frequency of the V_{REF} low-pass filter dramatically improved the results.

Note that the output noise is actually slightly below the predicted value. This is not surprising. The results from DataView are based on 256 samples from the ADC and the output noise is very close to the actual LSB size of the converter. These two items conspire to create a small uncertainty in the test results. In fact, the slightly higher noise provided in the last test, where the PICmicro microcontroller sources the load cell's excitation voltage and the MCP3551 reference voltage, may not be meaningful.

A more qualitative analysis can be made by comparing Figures 16 and 17. The results in Figure 16 are from the direct-connect circuit with R_1 equal to 10Ω, C_1 equal to 0.1 µF, and USB bus power driving the load cell and the reference input of the MCP3551. The results in Figure 17 are with the improved filter, R_1 equal to 332Ω and C_1 equal to 10 µF. Note that this load cell had a rated output of 4 mV/V for a 5 kg load – a 5g step was 0.1% of RO and caused a 20 µV change in output voltage.

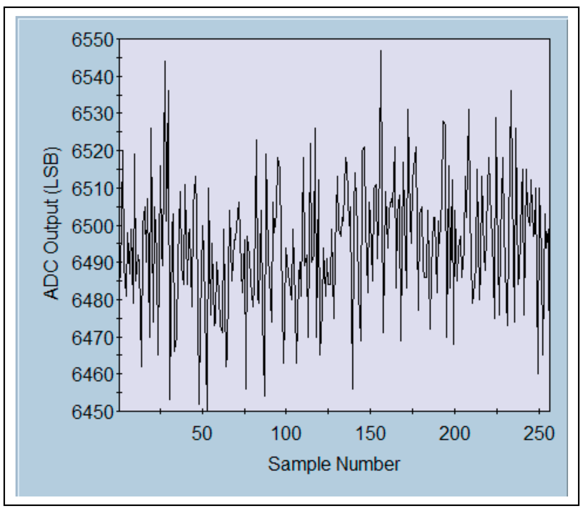

FIGURE 16: *5g Change (0.1%) on Direct-connect Weigh Scale with R_1 = 10Ω and C_1 = 0.1 µF (change occurs in approximately the middle of the graph).*

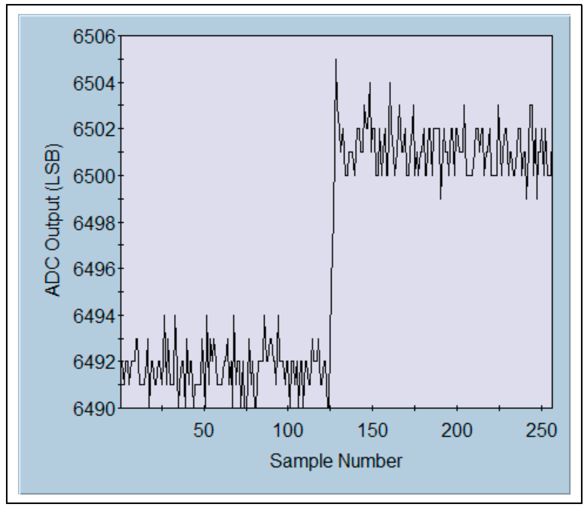

FIGURE 17: *5g Change (0.1%) on Direct-connect Weigh Scale with R_1 = 332Ω and C_1 = 10 µF.*

A HIGH-PRECISION WEIGH SCALE

While the result shown in Figure 17 does not look bad, this particular load cell has a larger rated output than most load cells (4 mV/V compared to the typical value of 2 mV/V). It is also obvious from Figure 17 that changes much smaller than 0.1% would be difficult to discern. The direct-connect weigh scale is limited to the 10-bit (1 part in a thousand) to 11-bit (1 part in two thousand) level for most load cells.

For higher precision weigh scales, the circuit shown in Figure 18 would be a more suitable starting point. This circuit gains up the output of the load cell with a high-precision, low-drift, 5V operational amplifier from Cirrus Logic (the CS3002). The gain of 101, implemented by the differential configuration of two op-amps, increases the resolution of the weigh scale by 7-bits, creating a scale capable of 17-bits to 18-bits of resolution.

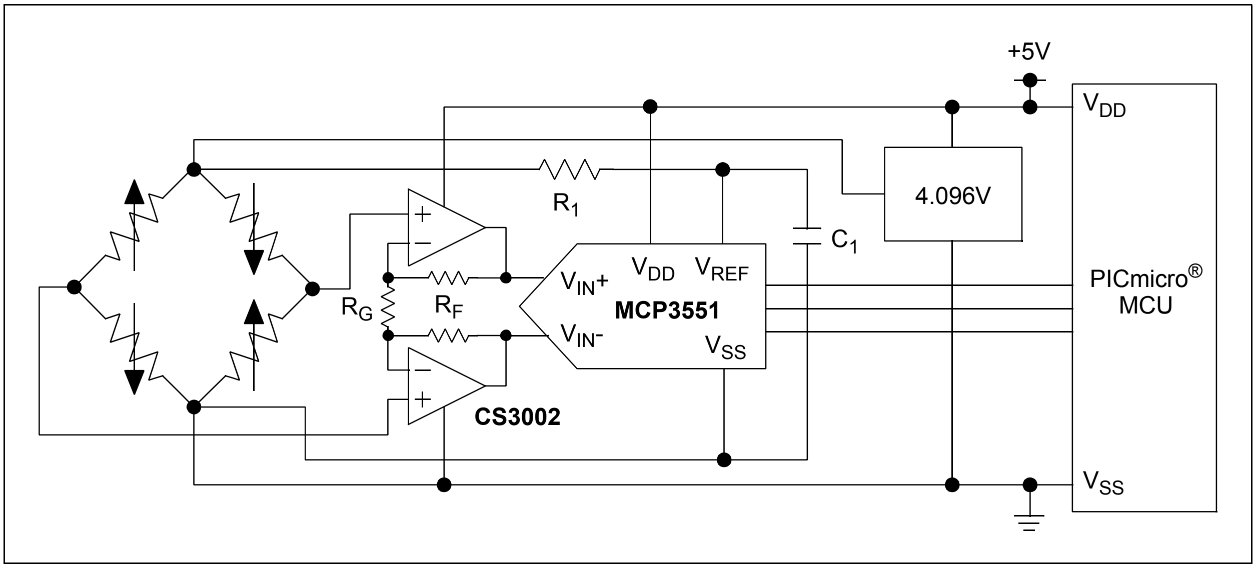

FIGURE 18: *A High-Precision Weigh Scale.*

The CS3002 was chosen for three reasons-it is a dual amplifier, it has very low noise of 125 nV peak-to-peak in a 0.1 Hz to 10 Hz bandwidth, and it has a very low maximum offset drift of ±0.05 µV/°C. The offset drift specification means that the offset drift of the amplifier will be below the offset drift of the load cell-making the load cell the primary contributor to offset drift.

The differential amplifier gain of 103 was chosen so that the amplifier noise would be similar to or greater than the noise of the analog-to-digital converter. This was done using R_F = 5.1 kΩ and R_G = 100Ω. The gain for the circuit is $2R_F/R_G+1$. This maximizes the resolution of the circuit. However, the gain should not be too large or the amplifier may clip or start to run into headroom problems near the power supply rails (+5V and ground). In this case, the circuit was given enough headroom to handle a wide variety of load cells without distortion or clipping while still providing a noise level similar to that of the MCP3551. The noise of the two amplifiers should be approximately 2.7 µV RMS compared to 2.5 µV RMS for the ADC.

As a side note to this discussion, there is often a question when designing weigh scales as to the proper specification to use for the noise analysis of the amplifier stage. Typically, the noise of an amplifier is specified as input noise voltage density (usually given in nanovolts per root hertz) and also a total peak-to-peak input noise voltage over a 0.1 Hz to 10 Hz bandwidth. The specification that best matches the weigh scale application and the weigh scale results is the input voltage noise over a 0.1 Hz to 10 Hz bandwidth. Simply use the value as given in the amplifier's data sheet as a starting point for the noise analysis. In some cases, the actual noise results may be slightly higher and, in others, it may be slightly lower. The noise results obtained with the circuit shown in Figure 18 were slightly higher than expected, but not substantially so (20% to 30% higher). It is unclear exactly why this was the case. However, it is possible that all of the noise sources were not completely accounted for or reduced as much as anticipated.

The circuit of Figure 18 was implemented on a printed circuit board that was connected to a computer via a USB interface. Since the USB interface provides +5V power, there was interesting opportunity to compare the performance of several options regarding Figure 18. One option was to connect the load cell directly across the +5V power from the USB interface (see Figure 19). Another variation was to drive the load cell from one or two pins of the PICmicro MCU that were configured as outputs and set high (see Figure 20).

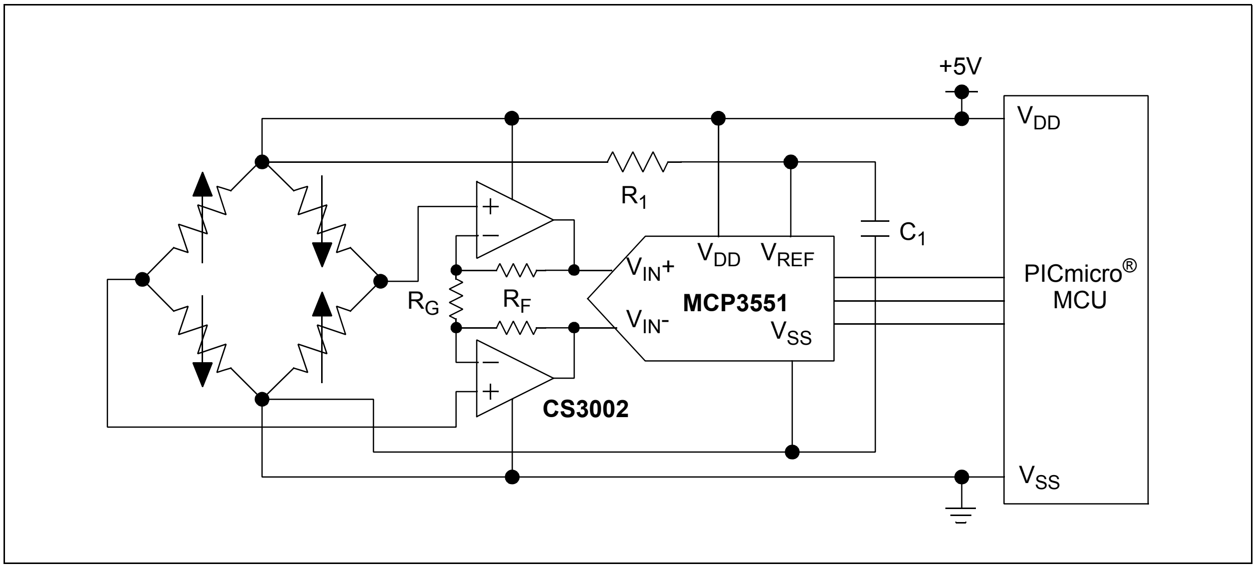

FIGURE 19: *A High-Precision Weigh Scale with the Load Cell Driven by +5V USB Power.*

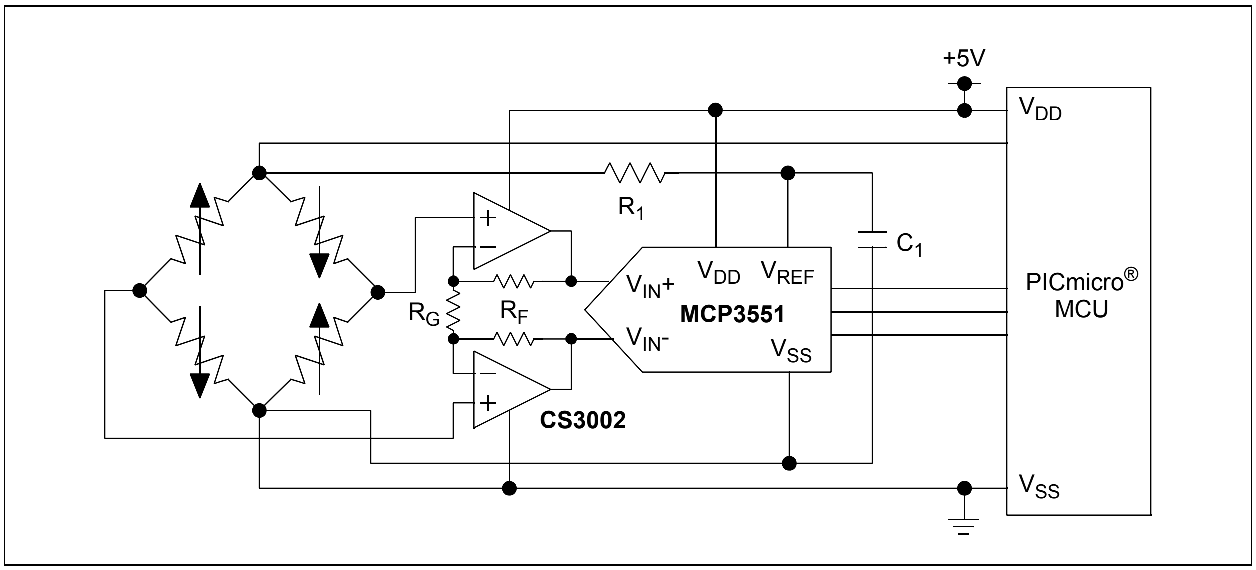

FIGURE 20: *A High-Precision Weigh Scale with the Load Cell Driven by the PICmicro MCU.*

The circuit shown in Figure 18 was actually tested with two different 4.096V references: a National Semiconductor LM4140 and an Analog Devices REF198. The following tests were done on these two variations of Figure 18 as well as the circuit configurations shown in Figures 19 and 20.

The same sequence of tests can be done on this circuit as was done for the direct-connect weigh scale. This means that the lowpass filter for V_{REF} must be restored to R_1 = 10Ω and C_1 = 0.1 µF. With the LM4140 reference, the DataView software should have reported an output noise of 0.49 PPM. The REF198 output voltage was closer to 4.096 but the resulting output noise would still be 0.49 PPM. The USB power was not exactly 5V, but close enough that DataView should have reported an output noise of close to 0.39 PPM for the last two tests. Table 5 provides the quantitative test results.

TABLE 5: RESULTS OF TESTING THE HIGH-PRECISION WEIGH SCALE WITH R_1 = 10Ω AND C_1 = 0.1 µF.

Load Cell & MCP3551 V_{REF} Source	Output Noise	
	PPM of FS	µV RMS
LM4140	1.04	8.5
REF198	1.27	10.4
USB +5V Power	3.27	32.7
PICmicro MCU (powered by USB +5V Power)	4.21	42.1

These results are actually in good agreement with the results obtained for the direct-connect weigh scale. Here again, the cutoff frequency for the lowpass filter on V_{REF} is reduced to 48 Hz by making R_1 = 332Ω and C_1 = 10 µF. Note that the concern regarding potential affects on gain error by setting R_1 to a larger value has a lot more impact on the circuit of Figure 18 than it did for the circuit of Figure 13. For the circuit of Figure 18, much more of the converter's full-scale range is used and any change in V_{REF} will have a greater impact on the conversion results. The 332Ω resistor should have less affect on gain error with temperature than the actual drift of the load cell.

Table 6 provides the results for the modified circuit – a substantial improvement for all configurations.

TABLE 6: RESULTS OF TESTING THE HIGH-PRECISION WEIGH SCALE WITH R_1 = 332Ω AND C_1 = 10 µF.

Load Cell & MCP3551 V_{REF} Source	Output Noise	
	PPM of FS	µV RMS
LM4140	0.62	5.1
REF198	0.59	4.8
USB +5V Power	0.53	5.3
PICmicro MCU (powered by USB +5V Power)	0.53	5.3

As with the direct-connect weigh scale, a more qualitative analysis can be made by comparing Figures 21 and 22. The results in Figure 21 are from the circuit with R_1 equal to 10Ω and C_1 equal to 0.1 µF, and USB bus power driving the load cell and the reference input of the MCP3551. The results in Figure 22 are with the improved filter, R_1 equal to 332Ω and C_1 equal to 10 µF. Note that this load cell had a 5 kg rated output and an output of 4 mV/V of excitation – a 5g step was 0.1% of RO and caused a 20 µV change in output voltage.

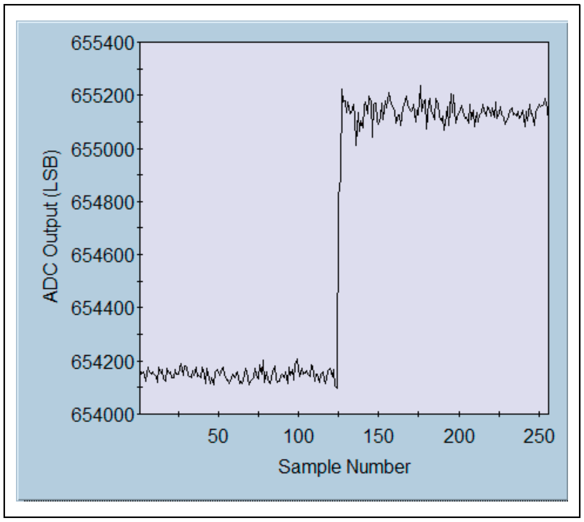

***FIGURE 21:** 5g Change (0.1%) on High-Precision Weigh Scale with R_1 = 10Ω and C_1 = 0.1 µF (change occurs in approximately the middle of the graph).*

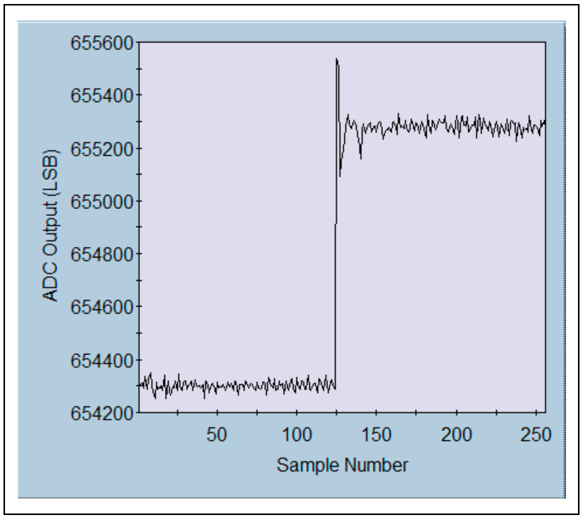

FIGURE 22: *5g Change (0.1%) on High-Precision Weigh Scale with R_1 = 332Ω and C_1 = 10 µF.*

While the difference between Figures 21 and 22 is not nearly as dramatic as between Figures 16 and 17, there is still some noticeable difference between the two. In addition, it is interesting to compare Figures 17 and 22.

There is an additional issue with the high-precision circuit that was not a concern for the direct-connect weight scale. The lowpass filter on V_{REF} has affected the ratiometric nature of the weigh scale. On the direct-connect weigh scale, no issues were discovered with this filter. However, on the high precision weigh scale, testing showed an additional problem-increased output noise when the load cell output was near full-scale, but this only occurred when using USB Bus power.

Table 7 provides test results that were obtained using a load cell simulator with a differential output voltage of 25 mV (to simulate 250% of RO) with two different computers and compares these results to those obtained with a simulator whose differential output voltage was 0V. Again, this data was taken with USB Bus power driving the load cell and MCP3551 reference input pin.

TABLE 7: COMPARISON OF OUTPUT NOISE FOR THE HIGH-PRECISION WEIGH SCALE WITH SIMULATED NO LOAD AND 250% OF RATED OUTPUT

Simulated Load	Output Noise	
	PPM of FS	µV RMS
No Load (0V), Laptop USB +5V Power	0.53	5.3
250% of Rated Output (25 mV), Laptop USB +5V Power	0.87	8.7
250% of Rated Output (25 mV), Desktop USB +5V Power	18.22	182.2

Again, it should be stressed that the output noise only increased when using the USB +5V power to drive the load cell and MCP3551 reference input (this was also true when using the PICmicro MCU which is essentially the same case). This fact provided a clue to help improve the test results. It was theorized that the low-pass filter on the MCP3551 V_{REF} pin was causing a phase delay for signals whose frequencies were in the 10 Hz to 50 Hz range, so the circuit was no longer ratiometric at these frequencies. By adding a capacitor across the load cell outputs, a delay would be added to these signals that should match the delay through the low-pass filter of the reference input – restoring the ratiometric balance of the circuit.

Tables 8 and 9 provide the test results with various capacitors across the output of the load cell for the laptop USB power and for the desktop USB power.

TABLE 8: COMPARISON OF OUTPUT NOISE FOR THE HIGH-PRECISION WEIGH SCALE WHEN USING A LAPTOP COMPUTER'S USB +5V POWER AND VARIOUS CAPACITORS ACROSS THE OUTPUT OF THE LOAD CELL

Capacitor across the Load Cell	Output Noise	
	PPM of FS	µV RMS
None	0.87	8.7
3.0 µF	0.54	5.4
3.6 µF	0.55	5.5
5.8 µF	0.79	7.9
7.2 µF	1.02	10.2

TABLE 9: COMPARISON OF OUTPUT NOISE FOR THE HIGH-PRECISION WEIGH SCALE WHEN USING A DESKTOP COMPUTER'S USB +5V POWER AND VARIOUS CAPACITORS ACROSS THE OUTPUT OF THE LOAD CELL

Capacitor across the Load Cell	Output Noise	
	PPM of FS	μV RMS
None	18.22	182.2
2.8 μF	3.45	34.5
3.0 μF	1.63	16.3
3.6 μF	3.77	37.7
5.8 μF	18.77	187.7
7.2 μF	29.39	293.9

The point of this exercise is not to imply that a capacitor should be placed across the output of the load cell in order to improve the results for the weigh scale circuit of Figure 18. Rather, this information should be used as a design consideration. The low-pass filter placed on V_{REF} does cause some ratiometric issues when noise is present. If the noise is too severe (as with the desktop supply), the resulting performance may be unacceptable. For other cases, a small capacitor with a moderately wide tolerance may solve the problem. Yet another solution is to use a good reference to drive the load cell and the MCP3551 reference input. This is probably the most definitive solution, particularly for a high-precision application.

A LOW-COST WEIGH SCALE

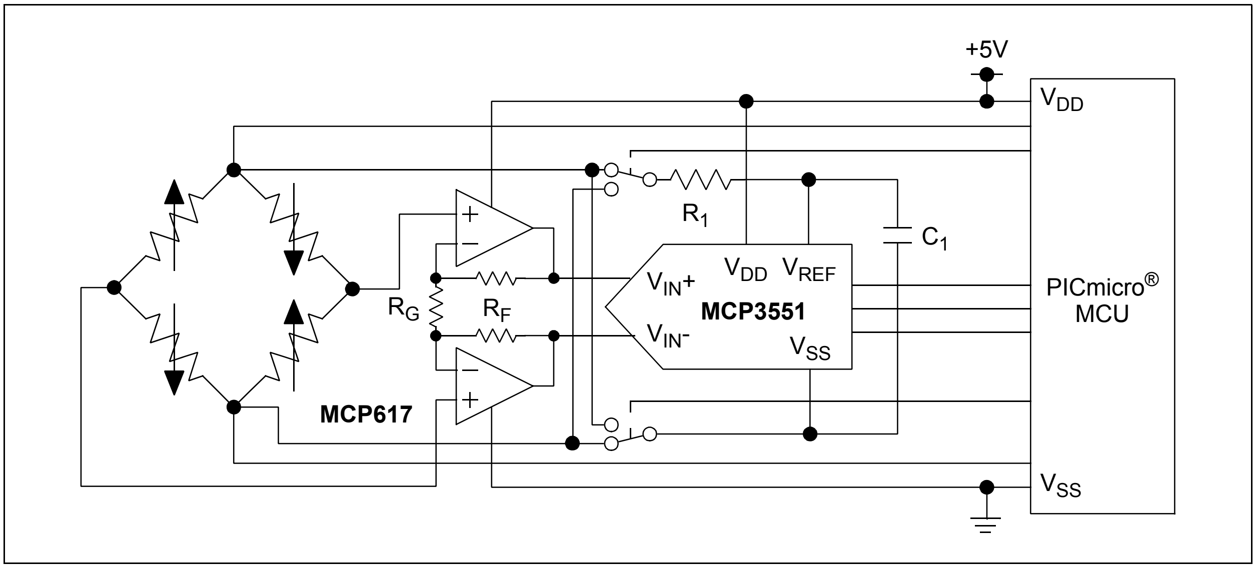

FIGURE 23: *A Low-Cost Weigh Scale.*

The goal of the circuit shown in Figure 23 is to allow the use of an operational amplifier with higher offset drift, which will generally mean a lower cost amplifier. In this case, the MCP617 was chosen and configured to provide a differential gain of 21. The MCP617 is a dual amplifier with an offset drift of ±2.5 µV/°C and an input voltage noise of 2.2 µV peak-to-peak in a 0.1 Hz to 10 Hz bandwidth. The gain of 21 means the voltage noise will be approximately 10 µV RMS at the input of the MCP3551. This noise is substantially above the 2.5 µV RMS noise of the ADC itself, so higher gains will not provide any additional improvement.

The trick to using a higher drift amplifier is to swap the sources driving the load cell. One conversion is done with the load cell driven "normally" and a second while it is driven in an "inverted" configuration. The result of the second conversion is inverted and added to the result from the first and an average of the two is computed (computing the average is a simple shift operation for the microcontroller). This technique effectively eliminates the offset error and offset drift of the amplifier as well as the offset error and offset drift of the ADC.

A more detailed description of this process is as follows:

Step 1. Configure both outputs of the PICmicro MCU that drive the load cell as low.

Step 2. Switch the ground of the MCP3551 to the "bottom" of the load cell and the reference of the MCP3551 to the "top" of the load cell.

Step 3. Configure the output of the PICmicro MCU that drives the "top" of the load cell as high.

Step 4. Perform a conversion and save the result.

Step 5. Configure both outputs of the PICmicro MCU that drive the load cell as low.

Step 6. Switch the ground of the MCP3551 to the "top" of the load cell and the reference of the MCP3551 to the "bottom" of the load cell.

Step 7. Configure the output of the PICmicro MCU that drives the "bottom" of the load cell as high.

Step 8. Perform a conversion, invert the result, add to the first conversion, divide by two, and save the result as the actual reading.

As with any circuit configuration, there are several potential pitfalls. First, the on-resistance of the switches that connect the MCP3551 ground and reference pins to the load cell must have low on-resistance. If the on-resistance is too high, then the variation in on-resistance with temperature will become a source of gain error. The desired on-resistance for the switches of Figure 23 should be 10Ω or less. Another concern is that the potential temperature change of the amplifier and ADC must be low over the time period of two conversions. This time will be at least 150 milliseconds.

For the circuit shown in Figure 23, analog SPDT switches from Fairchild Semiconductor were chosen. These devices, part number FSA4157, have an on-resistance of approximately 1Ω. This is 1/10 of the target on-resistance.

Figure 23 was tested in three different configurations: normal, inverted, and switching. In the first two configurations, the circuit was operating with either the top of the load cell driven high or low, respectively (and the opposite for the bottom). The results were simply collected and processed as normal. In the switching configuration, the load cell excitation was swapped over the course of two conversions and the results were averaged. The resulting average was sent back for analysis with DataView.

Most of these results are not of importance considering the results that have already been seen so far. The results are very similar to those obtained with the CS3002, only with more noise (about 2.5 times higher noise, as expected). In addition, more drift could be observed in the normal and inverted configurations. As with previous testing, the results confirmed the need for the 48 Hz low-pass filter prior to the V_{REF} pin of the MCP3551 (R_1 = 332Ω, C_1 = 10 µF).

The main interest lies in confirming that there is actually reduction in the offset drift of the amplifier. This reduction can be seen in the following figures, which are screen shots from DataView. In Figure 24, the circuit of Figure 23 is configured normally and no switching is going on. The figure shows the result of a "finger test" – that is, a finger is simply placed on the amplifier package to warm it up. Care is taken not to touch the leads of the amplifier. In this case, the device was in a DIP package and it was very easy to avoid touching the package leads.

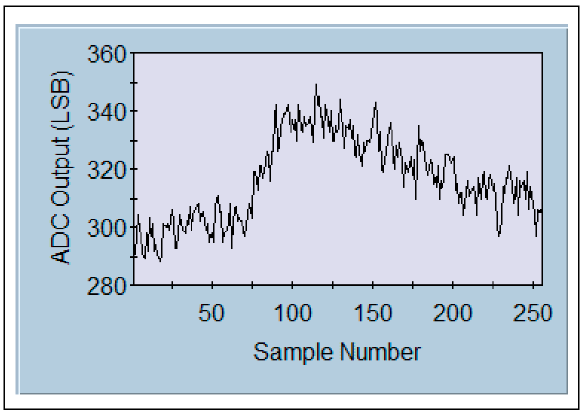

FIGURE 24: *Result of warming up the MCP617 without any switching to cancel offset drift (finger applied at sample 75 and removed at sample 125).*

In Figure 25, the circuit of Figure 23 is configured to switch the excitation voltage to the load cell and to average two conversion results. Notice that while offset drift is completely cancelled so is a good deal of the offset error of the amplifier (compare to Figure 24). The offset error is not completely canceled because the excitation voltages across the load cell are not identical in each case (normal versus inverted). The results in a small residue from the amplifier offset due to the MCP3551's full-scale mismatch between the two configurations. This "residue" offset is actually of some small concern because the excitation voltage for each configuration can drift with temperature. However, for the MCP617, the change in the value of the residue will be less than a few microvolts per degree Celsius, less than the drift of the load cell with temperature (which is gained up by a factor of 21).

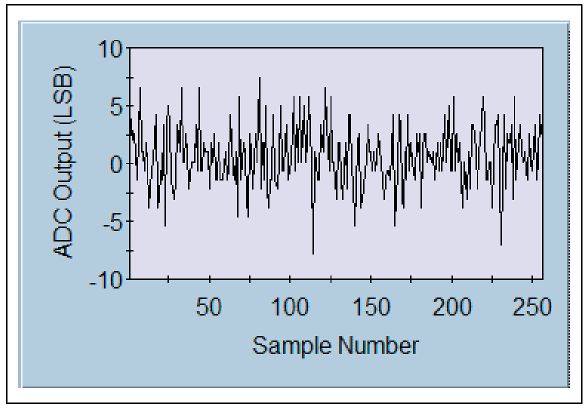

FIGURE 25: *Result of warming up the MCP617 with switching to cancel offset drift (finger applied at sample 75 and removed at sample 125).*

Using the switching technique with the MCP617 actually results in less drift than is seen with the CS3002. Figure 26 shows the result of a "finger test" done on the CS3002. (In this case, additional care was needed when touching the small SOIC package of the CS3002 in order to avoid touching the package leads). Keep in mind that the gain of this circuit was 101 compared to a gain of 21 for the MCP617.

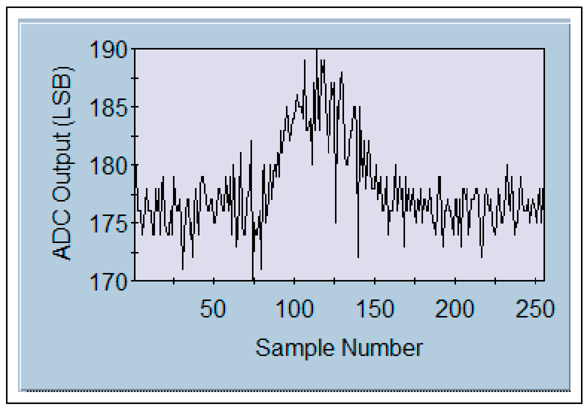

FIGURE 26: *Result of warming up the CS3002 (finger applied at sample 75 and removed at sample 125).*

As a final note regarding the circuit of Figure 23, the V_{SS} pin of the MCP3551 should be bypassed to ground with a 1 µF to 10 µF capacitor. While this may seem like an odd thing to do, the V_{SS} pin of the MCP3551 is not really at ground potential. Because of the current flow through the PICmicro MCU, the V_{SS} pin of the MCP3551 is actually several hundred millivolts above ground. Brief current draw by the MCP3551 during normal operation can produce a change in voltage at the V_{SS} pin as well as a change in voltage across the load cell. A bypass capacitor eliminates the problem.

NOISE DISCUSSION

One of the topics that has not been covered in detail is the overall noise of the various weigh scale circuits. This noise includes the noise from the analog-to-digital conversion process as well as that from the amplifiers, resistors, and the load cell.

As a quick review, the noise for the direct connect weigh scale is mainly determined by the ADC noise, which is 2.5 µV RMS. For the high precision weigh scale circuit using the CS3002, the noise is a combination of the ADC noise, the gained-up amplifier noise, and a small amount of resistor noise, for a total noise of about 4 µV RMS. For the low-cost weigh scale circuit using the MCP617, the noise is mainly from the amplifier and is about 10 µV RMS after a gain of 21.

In all three cases, the noise appears to display a Gaussian distribution. For the MCP617, the noise appears to have a component that may be related to temperature drift (as a general rule, it is very difficult to separate low frequency noise from possible drift due to temperature).

Since the noise displays a Gaussian distribution, it can easily be reduced by averaging multiple conversions, as has been previously discussed. The noise will be reduced by one over the square root of the number of results that are averaged (see Equation 1). So, four averages will result in one-half the noise while sixteen averages will result in one-quarter the noise.

In some applications, averaging may also impact the amplitude of rapidly changing signals since it acts as a very simple digital filter. However, for a typical weigh scale application, the main concern regarding averaging is the resulting update rate of the scale's display. Too much averaging will result in a display that changes too slowly.

The MCP3551 is fast enough that averaging can easily be considered. With a conversion time of approximately 73 milliseconds, the MCP3551 can perform 13.5 conversions per second. A typical weigh scale will update its display roughly once or twice per second. This means that four or eight averages could easily be accommodated. This would reduce the noise by a factor of two or three, respectively.

Averaging two, four, eight, or even sixteen consecutive results is trivial for most microcontrollers because the final division is simply a shift operation. Figure 27 provides a DataView graph of the CS3002 high precision weigh scale circuit (Figure 18) with no averaging up through sixteen averages (the number of averages is doubled every 50 samples).

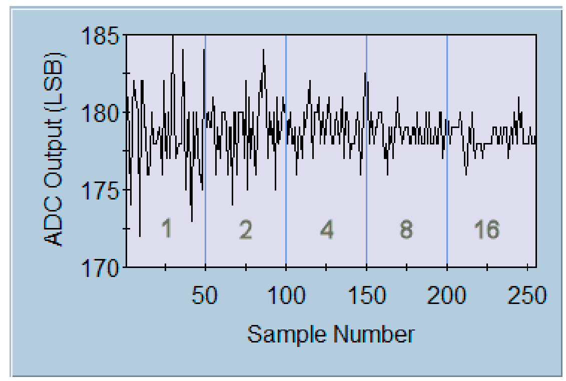

FIGURE 27: *Averaging with the CS3002 High Precision Weigh Scale Circuit (averaging starts at 1 and doubles every 50 samples).*

Figure 28 provides a similar result for the MCP617 low-cost weigh scale circuit of Figure 23. In this case, the results are from the switched configuration which means that the averaging actually starts out at two and doubles every 50 samples to a final value of 32 averages. This number would result in the weigh scale's display being updated once every 5 seconds – which is much too slow. Since the switched configuration already utilizes averaging, only two or four additional averages are possible.

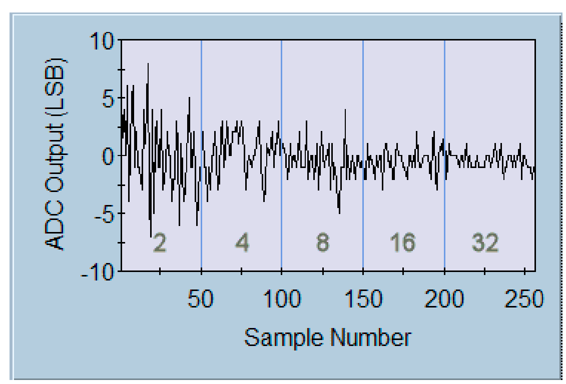

FIGURE 28: *Averaging with the MCP617 Low-cost Weigh Scale Circuit (averaging starts at 2 and doubles every 50 samples).*

The inherent averaging of the switched configuration also has another benefit. Figure 29 shows the noise of the circuit when used without switching (in the “normal” configuration) but still averaging every two conversion results to produce each sample. Figure 30 shows the noise of the circuit when used with switching but no additional averaging (so each sample is also the result of two averages).

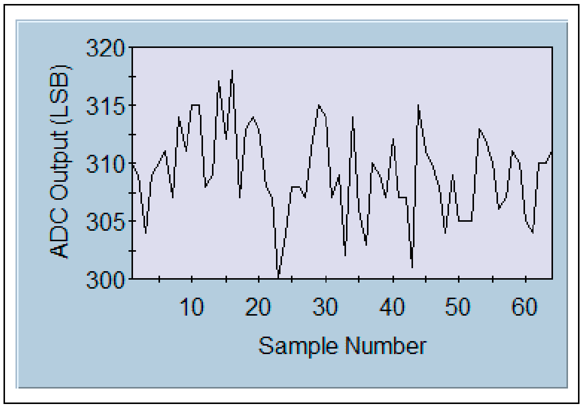

FIGURE 29: *MCP617 Low-Cost Weigh Scale Circuit in the “Normal” Configuration and Two Averages per Sample (one sigma noise = 8 µV).*

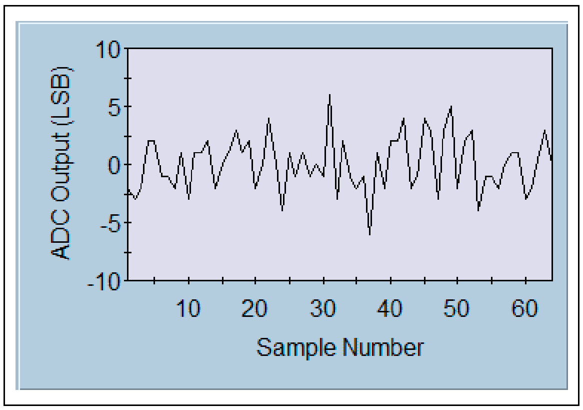

FIGURE 30: *MCP617 Low-Cost Weigh Scale Circuit in the “Switched” Configuration and No Averaging (one sigma noise = 5 µV).*

The switched configuration has reduced the overall noise more than would be expected due to averaging alone. What has happened is that some of the low frequency noise (in the 0.1 Hz to 0.5 Hz band) has actually been cancelled by the switching. As was mentioned previously, some of this “noise” may actually be drift due to temperature (such as offset drift). This is a very welcome side benefit of switching the excitation of the load cell.

To summarize the noise discussion, the noise performance of the three basic circuit configurations is compared. This comparison assumes that four results from the ADC are averaged. In addition, the results are presented in three different ways: signal-to-noise resolution in bits (also called equivalent number of bits or ENOB), “noise-free” ENOB (ENOB divided by 6.6), and noise-free dynamic range. This comparison assumes the load cell has a full-scale output of 10 mV. It also takes into account the slightly higher noise that was observed on the circuit using the CS3002 (about 30% greater than anticipated) as well as the slightly lower noise that was observed when using the MCP617 along with switching the excitation voltage to the load cell (which reduces noise about 30%).

TABLE 10: NOISE COMPARISON OF THE THREE WEIGH-SCALE CIRCUITS

Weigh-scale Circuit	ENOB	“Noise-free” ENOB	“Noise-free” Resolution
Direct Connect	13	10.4	1,350-to-1
High Precision using CS3002	18.8	16.1	70,000-to-1
Low-cost using MCP617	15.8	13.1	8,800-to-1

Figures 31 through 33 provide a final qualitative view of the capabilities of the various weigh scale circuits. Each was connected to a 200 kg load cell and DataView was used to graph the output of the MCP3551 while a small weight was added (which occurs in the middle of each graph).

For the direct-connect weigh scale circuit, 100g was added to the load cell producing a change in output voltage equal to 1/2,000th of its rated output (see Figure 31). The upper portion of the noise prior to adding the weight is just below the lower portion of the noise after adding it, which seems to indicate slightly better performance than the “noise-free” resolution for the circuit given in Table 10 of 1,350-to-1. Still, the quantization of the signal by the ADC is clearly visible and this may be hiding some noise.

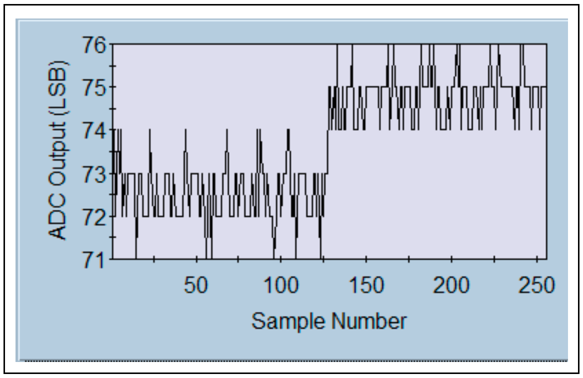

FIGURE 31: *Direct-connect Weigh Scale Circuit: 100g Change in Weight for a 200 kg Load Cell (change occurs in the middle of the graph).*

For the high precision weigh scale circuit, 2g was added to the load cell producing a change in output voltage equal to 1/100,000th of its rated output (see Figure 32). The upper portion of the noise prior to adding the weight slightly overlaps the lower portion of the noise after adding the weight, which correlates quite well with the "noise-free" resolution for the circuit provided in Table 10 of 70,000-to-1.

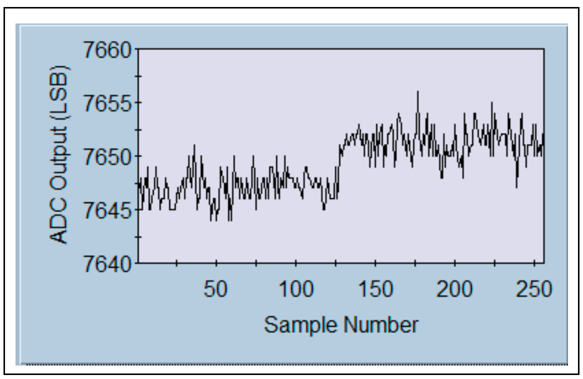

FIGURE 32: *A 2g Change in Weight for a 200 kg Load Cell (change occurs in the middle of the graph).*

For the low-cost weigh scale circuit, 16g was added to the load cell producing a change in output voltage equal to 1/12,500th of its rated output (see Figure 33). Again, the upper portion of the noise prior to adding the weight slightly overlaps the lower portion of the noise after adding the weight, which correlates quite well with the "noise-free" resolution for the circuit provided in Table 10 of 8,800-to-1.

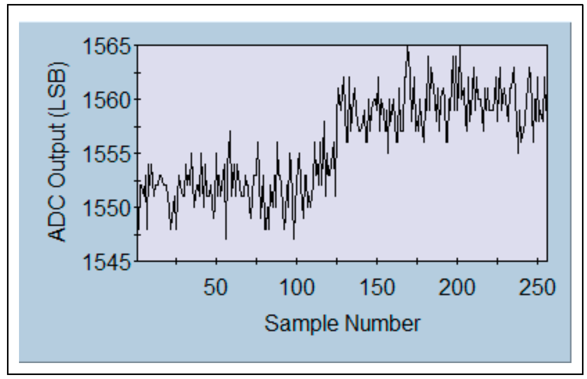

FIGURE 33: *A 16g Change in Weight for a 200 kg Load Cell (change occurs in the middle of the graph).*

CONCLUSION

The MCP3551 is an ideal ADC for a variety of resistive bridge applications. It can be connected directly to the sensor or it can be used along with other components to provide increased resolution and precision. With the addition of a PICmIcro MCU and a couple of switches, less expensive operational amplifiers can be used while still achieving excellent results.

Specifically, this application note has looked at three different circuits for use with load cells, a type of resistive bridge sensor. Collectively, the circuits provide performance ranging from 10-bits of "noise-free" resolution up to 16-bits. This resolution is available for a sensor whose differential output voltage ranges from 0V to 10 mV. For a sensor with a larger output voltage range, even higher resolution can be achieved.

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MCP616/7/8/9 Data Sheet, *"2.5V to 5.5V Micropower Bi-CMOS Op Amps"*, DS21613, 2005, Microchip Technology Inc.

"MCP355X Sensor Application Developer's Board User's Guide", DS51609A, 2006, Microchip Technology Inc.

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