Weight Scale Design with the 71M6511H/71M6511

The excellent accuracy and generous on-chip features make the 71M6511H and 71M6511 chips ideal candidates for implementing electronic weight scales for both commercial and industrial applications. This application note explains a few concepts and design examples for implementing weight scales.

The 71M6511H and 71M6511 - Good Candidates for Weight Scale Implementations

The 71M6511/71M6511H energy meter ICs consist of a multiple-input high-resolution ADC supported by an ultra-stable voltage reference, a 32-bit computation engine, an 8-bit MPU (similar to the industry standard 8051) with 64K flash memory, multiple I/O pins, two UARTs, a temperature sensor, LCD drivers, and on-chip support circuitry such as power-on reset, watchdog timer, brown-out detection, and a battery-buffered real-time clock. A block diagram is shown in Figure 1.

Even though initially designed for energy meter applications, the 71M6511/71M6511H ICs equally suitable for implementing an electronic scale. The ADC offers generous resolution and accuracy, the CE supports filtering and temperature compensation algorithms, the MPU implements general housekeeping functions, the LCD drivers interface to numerical and symbolic displays, the EEPROM interface can be used to save and recall data of any kind (previous measurements, product pricing per weight unit) and the battery mode preserves RTC and configuration data in between uses, if necessary.

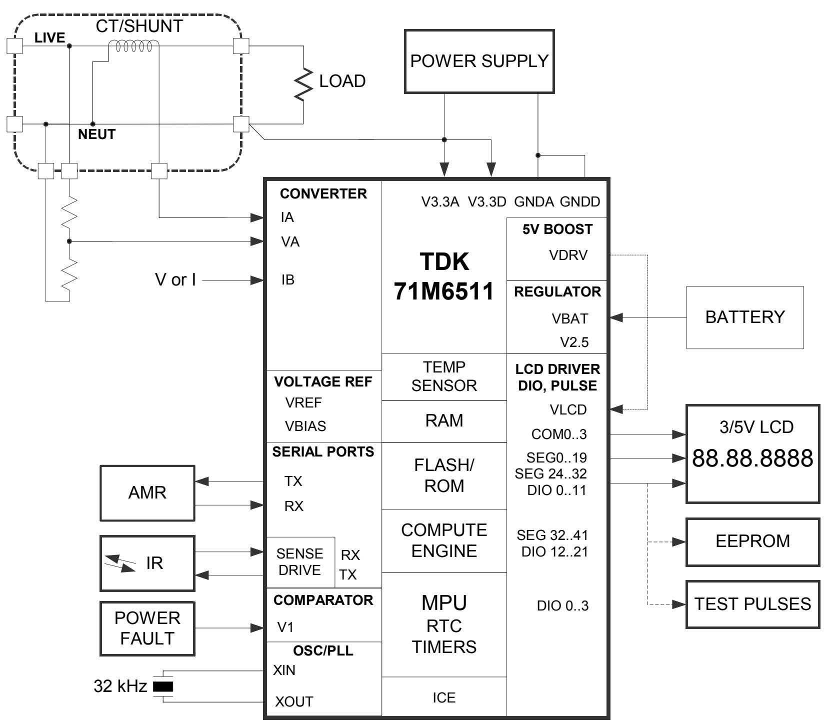

Figure 1: 71M6511 Used in an Energy Meter Application

This means that by adding only a few external components an electronic scale can be implemented.

Typical Sensors Used in Scales and how to Interface to them

The typical "weight" sensor is in reality a force sensor made up of strain gages and some fairly solid material. This combination is usually called a load cell. The mass of the object to be weighed generates a force (using the earth's gravitation) which distorts the material of the load cell. The strain gages attached to the load cell change their resistance in proportion to the distortion. The strain gages are usually connected in a bridge configuration, as shown in Figure 2, with RA and RD increasing resistance (expanding) when weight is applied and RB and RC decreasing resistance (compressing). A configuration of this type yields four times the output of a single strain gage while being very immune to temperature changes. If the distortion stays in the Hooke's region (i.e. no permanent deformation of the solid material occurs), the bridge differential voltage Vd will be proportional (or at least proportional for all practical purposes) to the weight.

Load cell sensitivity is typically specified in as mV per V at full scale. The mV number refers to the bridge differential voltage Vd while the V number refers to the excitation voltage Vexc. It is clear that the higher the excitation, the more differential output at a given weight will be generated. Excitation voltages are typically 5V to 20V. For example, if a load cell is specified as 1.875mV/V and 20Kg full scale, the output at half scale load (10Kg) will be 4.6875mV (5V * 0.5 * 1.875mV/V) if the load cell is operated at 5V excitation voltage.

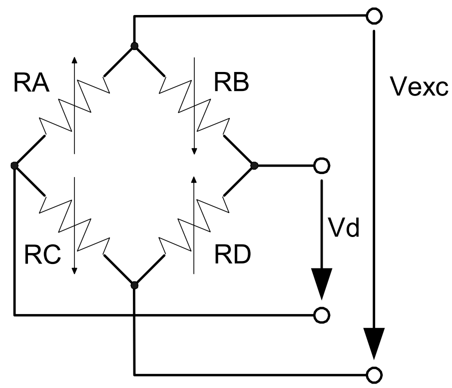

Figure 2: Sensor Bridge Configuration

The load cell depicted in Figure 2 generates the differential output given by:

$Vd = V_{RARC} – V_{RBRD} = s * f * V_{exc}$

Where s is the sensitivity in mV/V, f is the force (weight) applied as a fraction of the full scale, and Vexc being the excitation voltage. In absolute terms, the common mode voltage Vexc/2 rides on both signals V_{RARC} and V_{RBRD} that form Vd when the load cell is at zero load. This common mode voltage has to be eliminated by the signal conditioning circuitry which has to act as a differential amplifier.

Another aspect is that due to minute inaccuracies of the load cell or due to mechanical structures added to the scale (“tare”, such as a bowl for containing the weighed material or parts), a residual differential voltage (“offset at zero load”) will be present that will have to be eliminated by either the signal conditioner or by the application following the signal conditioner. The process of removing the zero offset is usually referred to as auto-zeroing.

Since Vd will typically be in the micro-Volt or low milli-Volt range, amplification or gain will have to applied in conjunction with the elimination of the common mode voltage in order to lift the output signal into the input signal range suitable for the ADC. If we consider the relatively narrow input voltage range of the 71M6511/71M6511H which is ±250mV, gains of 10 to 100 will have to be applied.

The standard approach in electronics to interfacing to a bridge-type sensor such as the load cell is the instrumentation amplifier (IA). This solution typically offers advantages over simpler solutions, such as high common mode rejection ratio, easy setting of circuit gain, good accuracy, and no requirements for high-precision resistors. However, with some restrictions, differential amplifiers formed with simple op-amps can be a good and cost-effective solution.

A Sample Amplifier Circuit

A simple differential amplifier can be implemented, as shown in Figure 3, with an operational amplifier (op-amp) and a few resistors. An important consideration for selecting the op-amp is the input offset voltage which must be low. We cannot allow an input offset voltage of more than a few mV, since after amplification with gains between 10 and 100, this would result in an offset output voltage that would restrict the useable signal range (±250mV).

The OP07 by Texas Instruments was selected for very low input offset and low offset voltage drift with temperature, another important consideration, especially for industrial applications. While the commercial scale might operate in constant temperature, an industrial scale will often be exposed to the elements. While the auto-zero mechanism will provide for a good reading at zero load, it may take considerable time for the load to be applied (think of a truck that has to be driven onto a truck scale, which may take minutes), and consequently temperature drift will occur.

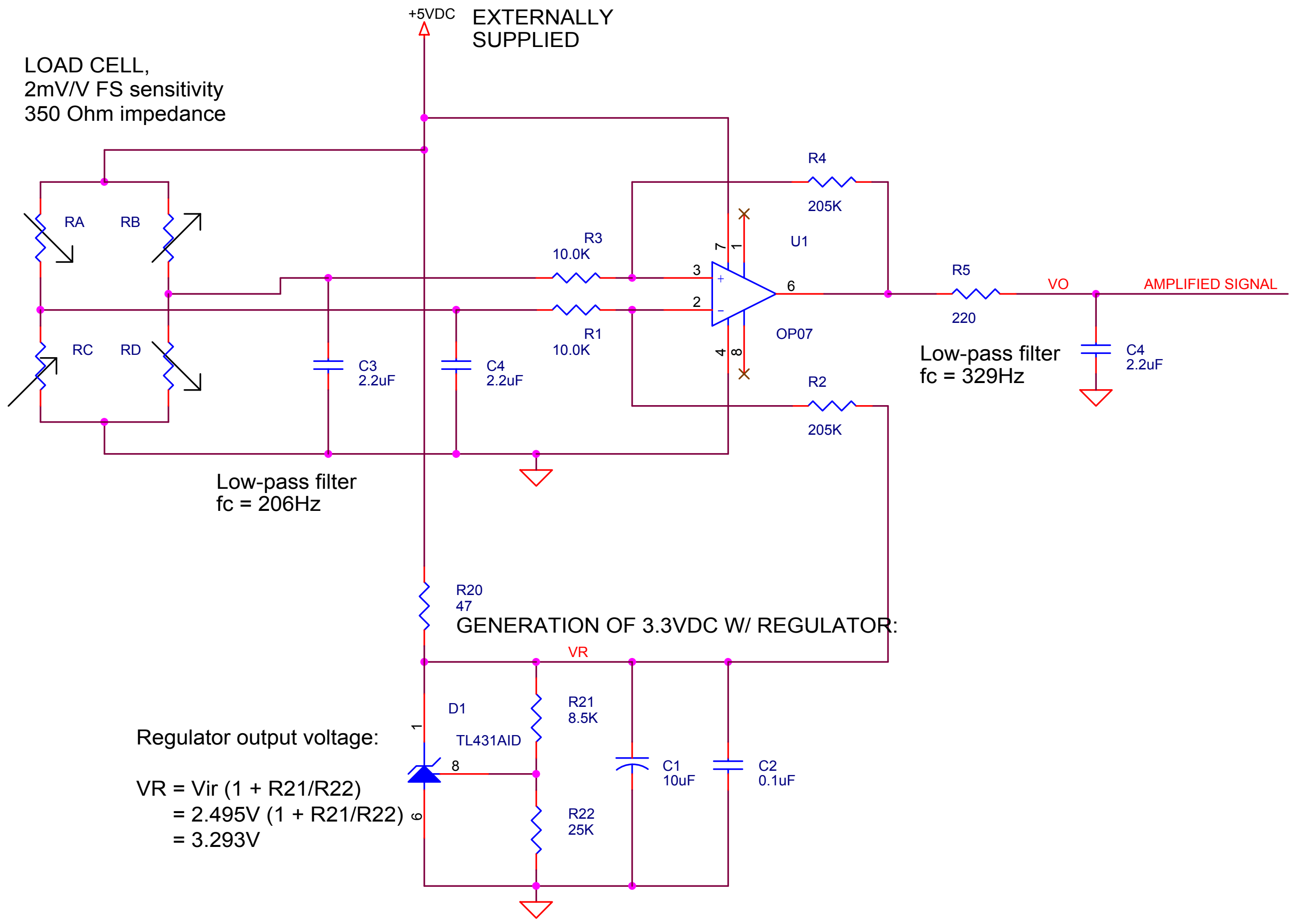

Figure 3: Simple Differential Amplifier and Reference Voltage Generator

Both the load cell and op-amp are operating on a +5VDC power supply, the combinations R4/R3 and R2/R1 provide the necessary gain, in this case 20.5. It is important that the ratio R1/R2 is equal to R3/R4. It is recommended to use 0.1% tolerance resistors. This is necessary in order to increase common-mode rejection, which is necessary if the signal picks up AC noise (especially 50 or 60Hz).

Since the inputs of the 71M6511/71M6511H are referenced to 3.3V (the V3P3 net on the Demo Boards), the output of the amplifier must be shifted to 3.3V. This is done by connecting R2 to a stable 3.3V source, implemented with a Texas Instruments TL431 regulator.

Filtering is implemented in three places: First, noise is removed at the op-amp inputs by the capacitors C3 and C4, which form a low-pass filter with the internal impedance of the load cell. Many load cells have impedances of around 350Ω, which means that the capacitors have to be large. Ceramic capacitors, which are preferred, are available with up to 10µF. The second filter is at the op-amp output, and due to the limited input impedance of the ADC in the 71M6511/71M6511H (60kΩ min.), the resistor needs to be fairly low-ohmic, again requiring a large capacitor. Of course, if a buffer stage is used, any filter can be implemented.

The cut-off frequencies in the sample amplifier are around 200Hz and 400Hz, but lower frequencies would be preferred. The effective sample frequency of the ADC in the 71M6511/71M6511H is at 2.520KHz, which means that about 3 ½ octaves remain between 200Hz and the sample frequency. The first low-pass filter (LPF), starting at 200Hz will generate around 20dB, the second LPF, starting at 400Hz will add another 14dB of signal suppression at 2.520KHz. The total number of 34dB may have to be improved for practical implementations.

Let us look at the characteristics we can expect from the circuit in Figure 3. Of primary interest is what resolution can be achieved when connecting this circuit to the ADC in the 71M6511/71M6511H, and what the temperature drift will be.

Let us first examine Table 1, which shows the basic characteristics of the load cell. At a full scale (FS) weight of 20Kg, the manufacturer states a calibrated sensitivity of 1.8538 mV/V. The multiplication of this value with the 5V excitation gives us the expected output at full scale. This is less than 2% of the ±250mV range of the ADC in the 71M6511/71M6511H, resulting in loss of resolution if used without a gain stage (of course, some sort of gain stage is needed anyway to transition from the differential to a single-ended output). Another entry in Table 1 states that the FS output is only 0.19% of the excitation voltage. This just confirms the fact that strain gages are used in a very narrow band around zero load. Loading the sensor beyond 20Kg would introduce significant non-linearity (besides breaking it).

FS weight (mass)	[Kg]	20
Load Cell FS Sensitivity	[mV/V]	1.8538
Excitation voltage	[V]	5
Load Cell FS output	[mV]	9.269
FS output percentage of excitation	[%]	0.19%

Table 1: Basic Load Cell Characteristics

Table 2 shows the gain stage and its output, compared to the 71M6511/71M6511H input. As we can see, the gain of 20.5 expands the FS signal now to 190mV, and a weight of 0.105Kg (105g) is equivalent to one milli-Volt signal amplitude.

R1/R2 at op-amp		10000
R2/R4 at op-amp		205000
Circuit gain		20.5
Circuit FS output voltage	[mV]	190.01
71M651X maximum input	[mV]	250
weight (mass) per mV	[Kg]	0.105

Table 2: Basic Amplifier Characteristics

Table 3 shows how the values from Table 2 compare in terms of resolution. Since the ADC in the 71M6511/6511H has a resolution of 21 bits and the reference is 1.195V, the achievable voltage granularity is 1.14µV ($0.5*1.195V/2^{21}$). If we compare this to the 190mV signal range, we find that this range corresponds to 166,732 steps or LSBs. This is equivalent to a resolution of 17 bits in the 0 to 190mV range. One LSB is effectively equivalent to 0.12g or 120mg. For a scale with a FS of 20Kg, this seems more than sufficient.

ADC nominal resulution	[bits]	21
ADC reference	[V]	1.195
ADC max voltage resolution	[µV]	1.1396
ADC effective resolution	[LSBs]	166732
ADC effective resolution	[bits]	17.35
ADC effective resolution	[Kg]	1.20E-04
	[g]	0.12
	[mg]	119.95
	[steps/g]	8.34

Table 3: Basic Amplifier Characteristics

Table 4 shows the offset calculation for the OP07. The maximum input offset of 0.25mV will be amplified by the applied gain to 5.13mV. This is equivalent to a weight of 0.5g or 2% of the 250mV input range of the ADC. The maximum input offset number applies to the full temperature range per the OP07 data sheet. Consequently, we can feel fairly assured that input offset will not be a problem.

OP-amp or IA used		OP07
Input offset, max	[mV]	0.25
Output offset	[mV]	5.13
Offset equivalent weight (mass)	[g]	0.54
Offset percentage of useable signal	[%]	2.05%

Table 4: Amplifier Offset

Using Table 5, we can examine the drift to be expected from our circuit. With an input offset drift of 2.5µV/°C, the output will drift at 51.25µV/°C. Assuming for the moment a temperature change of 10°C, our output will change by 512µV, which is only 0.2% of the full scale or the equivalent to 0.05Kg.

Input offset drift, max	[µV/°C]	2.5
Drift of output voltage, max	[µV/°C]	51.25
Temperature change (assumed)	[°C]	10
Output drift at assumed temperature change	[µV]	512.5
Drift equivalent weight (mass)	[Kg]	0.05
FS percentage of drift	[%]	0.27%
FS drift in PPM @ temperature change	[PPM]	2697

Table 5: Amplifier Drift

A circuit according to Figure 3 was built up and tested in the lab with the single-point load cell type SPS2-20 manufactured by AmCells (shown in Figure 4), confirming the theoretical calculations.

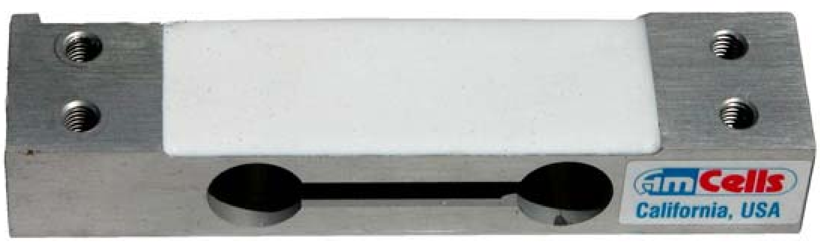

Figure 4: Single-Point Load Cell

Figure 5a shows the results of the measurements with loads of 0 to 3.5Kg applied (the difference between the constant 3.3VDC and the op-amp output is shown). A linear regression over the data points yields a slope of 108g/mV which corresponds to the 105g/mV calculated in Table 2. Figure 5a also shows the expected dependence of the output from the excitation voltage (increasing excitation will increase the output voltage).

Figure 5b shows the output with various temperatures, measured with thermal stream equipment heating the op-amp only. The slope did not change with temperature, but the offset changed from –0.2mV at room to –0.3mV at +50°C and to –0.45mV at +80°C. This is equivalent to 250µV over 50°C, or 5µV/°C, and much better than the 2.5µV/°C * 20.5 = 51µV/°C that we would expect from the maximum drift figures stated in the OP07 data sheet. As expected, there was no change in the signal when the regulator TL431 for the 3.3VDC was heated up. Since the output of the TL431 is not amplified at all, the drift of the op-amp itself is dominating all other deviations.

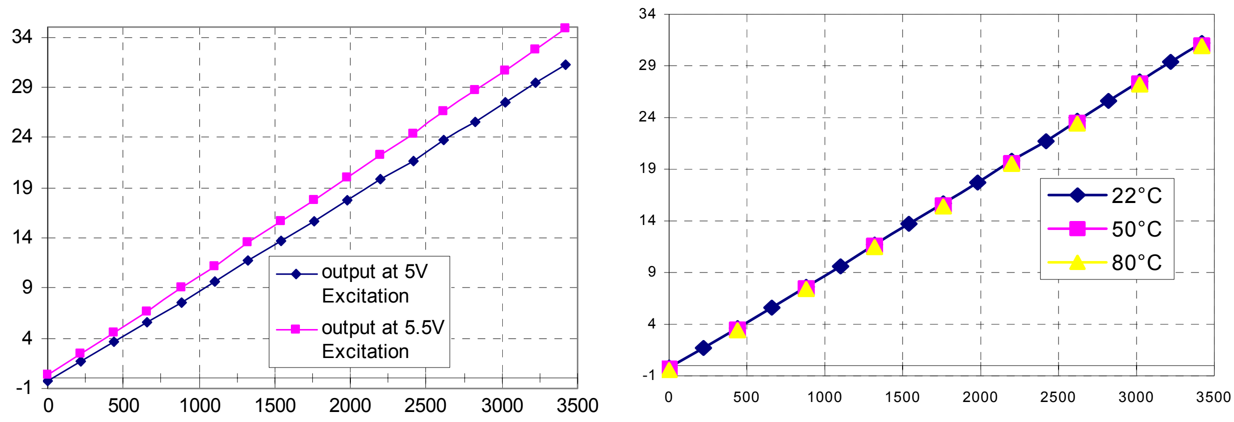

Figure 5: Sample Amplifier Output (left: Constant Temperature. Right: Variable Temperature)

A Load Cell Measurement System

So far, we have considered the amplifier as an isolated entity. We will now have to take a broader approach and look at the whole system of load cell, amplifier, power supply and ADC for two reasons: First, we have ignored the considerable drift introduced by the generator of the 3.3VDC (any deviation of the 3.3VDC with temperature will directly affect the measurement). The second reason is that it will be very difficult to guarantee that the +5VDC supply is stable over time and temperature.

A large number of implemented systems take advantage of the ratiometric measurement principle, which is illustrated in Figure 6. If the ADC can resolve the range from 0 to V_{ref} into N steps (LSBs) its digital output n

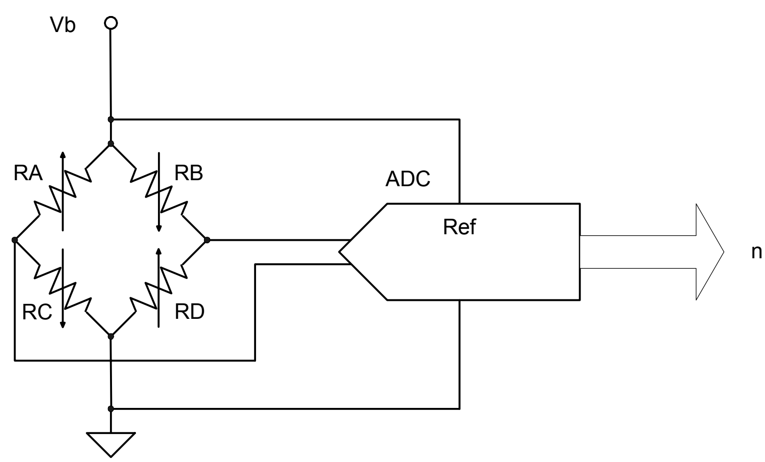

Figure 6: Ratiometric Measurement

will be, when a differential input V_d is applied:

$$n = V_d * N / V_{ref}$$

Since we use V_b to feed both the load cell and the ADC, we get:

$$n = (s * V_b * f) * N / V_b$$

(s = sensitivity, f = applied force, $N = 2^R - 1$, with R = ADC resolution)

This term is independent of V_b. Since the internal reference voltage of the 71M6511H/71M6511 should not be disabled, we use the I_B input to create the ratiometric behavior.

This would leave only one voltage independent from V_b, the 3.3VDC required to bias the amplifier and also needed to power the 71M6511H/71M6511. Scaling V_b properly with a resistor divider, and buffering it with a simple op-amp (U2), so changes in load will not affect stability, achieves the goal. R25 decouples the output of U2 sufficiently to prevent oscillations due to capacitive coupling. The resulting schematic is shown in Figure 7.

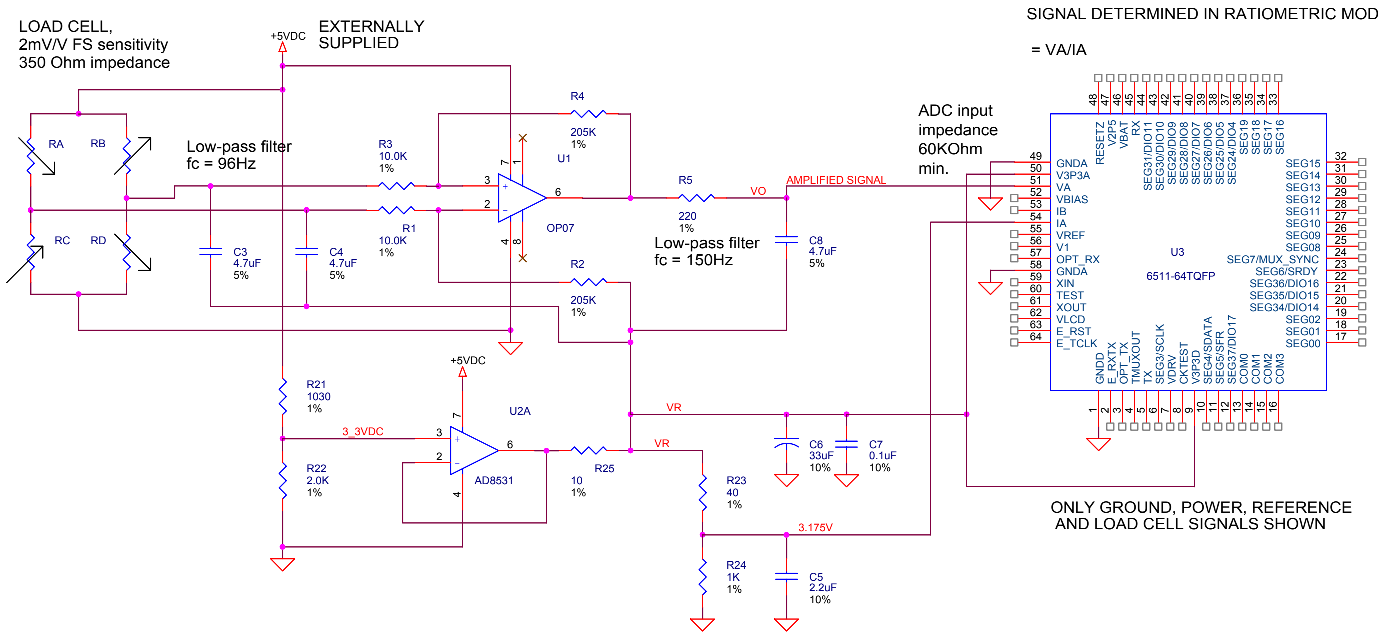

Figure 7: Ratiometric Measurement System

This circuit is ratiometric, since the displayed signal for the scale is derived by dividing the value obtained at V_A by the value obtained at I_A. As long as I_A does not reach zero, the division (provided by the MPU of the 71M6511) will eliminate the influence of variations of the 5V power supply. The voltage drop at R23 keeps the voltage supplied to I_A safely at around 3.175V.

Since the VR net is loaded by the supply current for the 71M6511H/6511 and the currents through the filter capacitors, care should be taken to make the net VR low-ohmic. This can be achieved by providing enough by-pass capacitance in C6 and C7. U2 must be stable with large capacitive loads which is the case with the AD8531 selected for this application (and other op-amps capable of driving capacitive loads at high currents).

The circuit in Figure 7 assumes only the +5VDC (= V_b) to be externally supplied. The resistor divider R21/R22 divides V_b down to 3.3V. Op-amp U2 buffers the 3.3V and supplies it to both op-amp U1 via R4 and to the power supply pins of the 71M6511H/71M6511.

The filter capacitors have been increased in value and referenced to the 3.3V supply.

In one more step of refinement, we will now eliminate the op-amp U2 and replace it with a voltage reference. The ratiometric measurement is maintained by deriving the displayed signal from V_A / (I_B – I_A). Figure 8 shows the resulting schematic.

D1 is the voltage reference used in Figure 3. R11 and R12 generate the voltage for I_A that is safely below 3.3V. R6 and R7 generate a voltage that reaches 3.55V when V_b is 15V and 3.3V when V_b is at +5V. This means that the load cell can be operated with any voltage between 5V and 15V. U1 can be supplied with up to +18V.

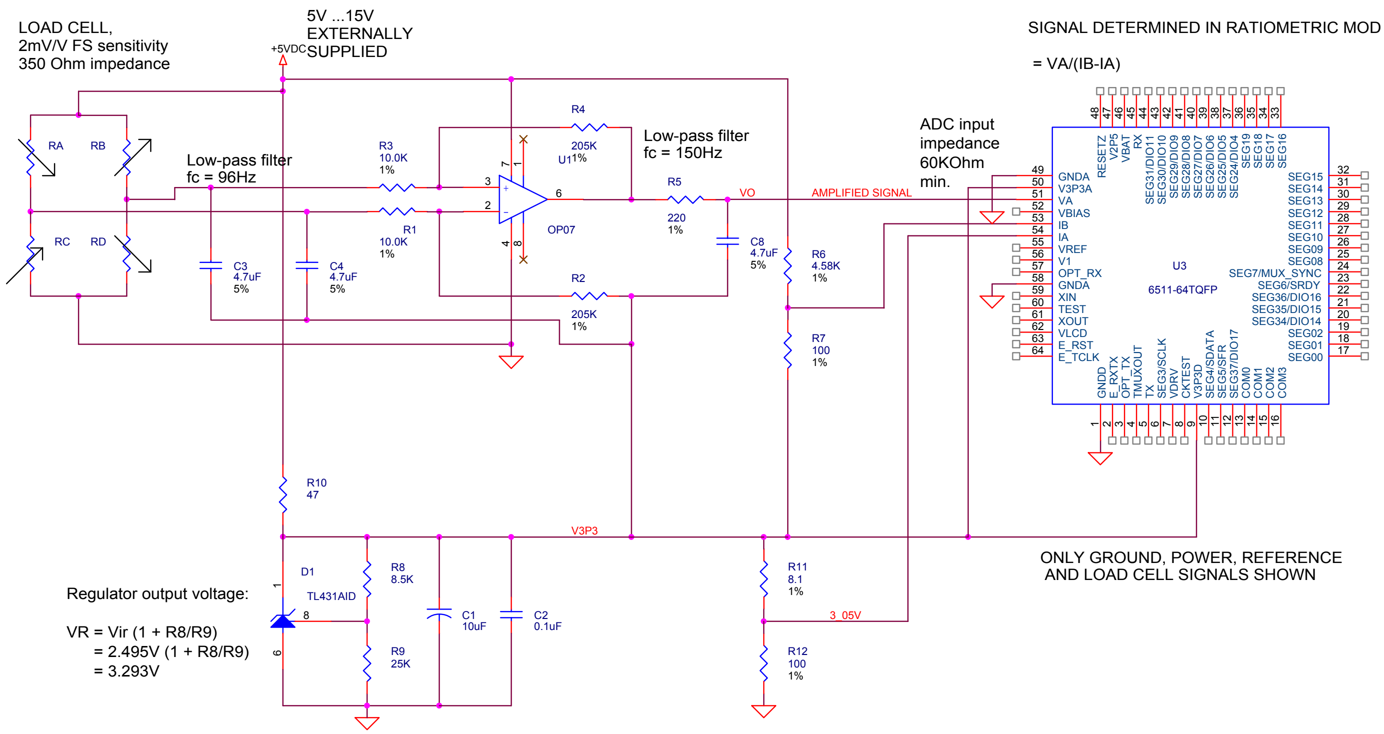

Figure 8: Ratiometric Measurement System

Calibration and Temperature Compensation

There are a number of effects that can be taken care of with calibration:

- Variations in load cell sensitivity
- Sensor non-linearity
- Amplifier non-linearity

Variations in load cell sensitivity will occur in production from unit to unit. By applying a standard weight to each scale in production test, sensitivity deviations can easily be taken care of by implementing a calibration factor. The MPU of the 71M6511H/71M6511 has the capability of performing that task.

Sensor and amplifier non-linearity must first be characterized, before a compensation mechanism can be employed.

The inherent non-linearity of a bridge-type sensor is shown in Figure 9. It graphs the output (differential voltage) over the resistance deviation in a full-bridge (four active resistors). A bridge with constant voltage supply exposes more non-linear than a bridge with constant current supply. It should be noted that Figure 9 shows the resistance deviation far beyond the 0.2% deviation used with typical load cells.

If so desired, the output function can be expressed as a polynomial, and the output signal of the scale can be linearized using polynomial correction. The CE in the 71M6511H/71M6511 ICs can be used to implement linearization of almost any type, and it helps that the signal is very slow. This allows for computation-intensive algorithms to be performed.

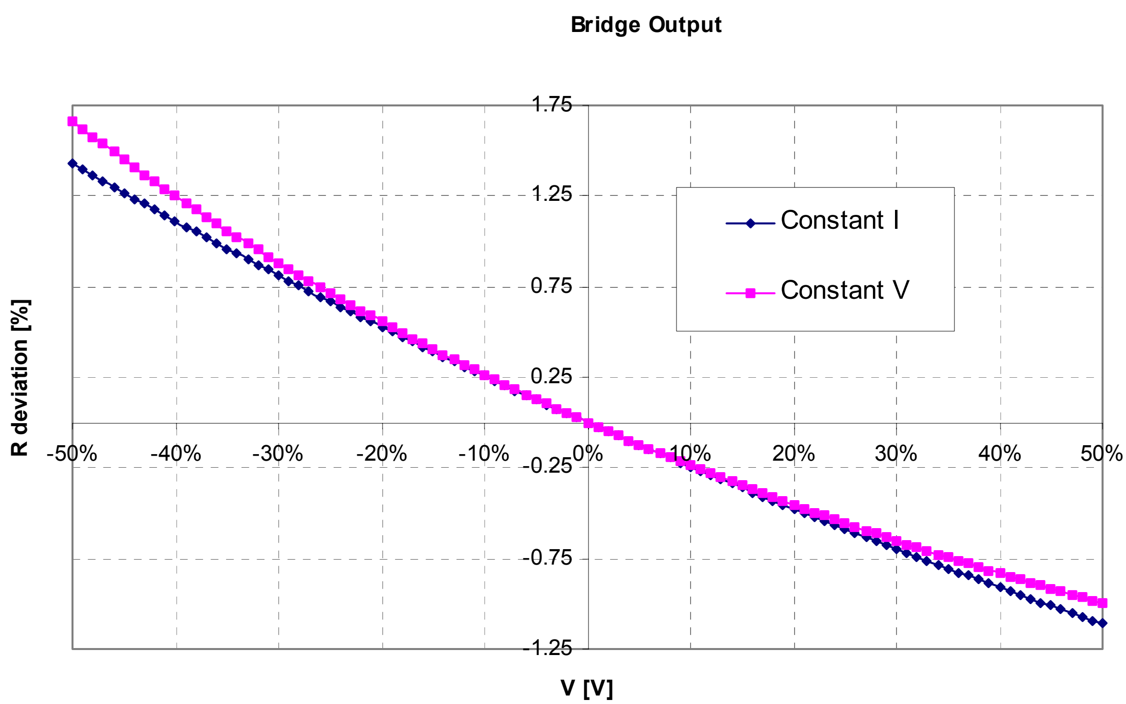

Figure 9: Bridge Non-Linearity

Similar considerations apply to non-linearity caused by the gain stage.

Temperature compensation may be employed to improve the performance of the scale over temperature. The factors affecting performance over temperature are:

- Gain drift of the amplifier
- Offset drift of the amplifier
- Gain and offset drift of the load cell

All of the above factors can be compensated for if they can be characterized and if they are repeatable. The 71M6511H/71M6511 ICs have an on-chip temperature sensor that can be used to implement almost any form of temperature compensation. Obviously, if the TC of the load cell has to be compensated, the load cell has to be co-located with the 71M6511H/71M6511 IC or it has to be guaranteed that the load cell and the 71M6511H/71M6511 are exposed to the same thermal environment.

TDK Semiconductor application note AN_651X_009 may help in designing mechanisms for temperature compensation.

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