Component Design for Load Cell Electronics (PSAS Wax Hybrid Tests)

Version 1.01, Last Modified 7/15/2004

Introduction

The load cell electronics measure the millivolt level signals produced by strain gage type load cells and amplify these signals into a 0−5V range.

Goals

Amplify low level signal from strain gage load cell and transmit it over several feet of cable to PC−baseddata acquisition card.
Accuracy sufficient to avoid degrading overall signal quality .
Small overall package to facilitate mounting near load cell
Minimize construction time, while achieving moderate cost.

Specs

Primarily through−holecomponents to facilitate vector board construction
Four wire interface (Power, Gnd, and Signal out, Signal Gnd)
Wide supply range, at least 12−24V (Try for 10V allowing discharged car batteries.)
Current mode output for easy interfacing over long cable runs
Overall accuracy sufficient for 12 bits of precision

Power Supply Section (PS)

PS Goals

Simplicity / Reliability
Wide input voltage range
Accuracy sufficient to maintain overall accuracy
Adequate heat dissipation
Low cost

PS Design notes

The basic idea is to use an ordinary 3 terminal regulator, possibly a low drop−outtype, and servo it with an opamp. The regulator buys a thermally protected, power capable device. The opamp circuit improves the accuracy of the regulator, since 3 terminal types are not sufficiently accurate.

PS Issues

Bandwidth could be rolled off with a capacitor at the adj. Terminal if required for stability (This was done).

To allow operation at 12 Volts, the bridge voltage could be lowered to 8.75V. The bridge noise is thus increased, but 10V operation is relatively easy to achieve. This appears to be a simple jumper change, so some design flexibility is retained.
(We decided to use 8.75V excitation to allow operation down to 10V. The as−builtvoltage was actually 8.66V.)

■ Instrumentation Amplifier (IA)

■ IA Goals

Single opamp package
Low parts count
12 bit accuracy over moderate range
Low cost

■ IA Design notes

Our load cell is a very standard unit. The nominal bridge resistance is 350 Ohms. The sensitivity is 3mV/V full scale, and the recommended maximum excitation is 10V. Typical errors are on order 0.1%, which is consistent with a 12 bit accuracy specification.

Rejected amplifiers:
LT1168_0500_Mag.pdf — Middle thirds rule
an87.pdf — p.69 has nice ECG application
LT1167_0598_Mag.pdf — Noise analysis, repetition, also interesting on ESD protection (supersedes dn182f.pdf ?)

LTC1100 (12$) chopper, single supply, 18V
LT1167 (6$) not single supply
LT1789 (6$) no DIPs

INA122

■ IA Issues

The main issue is with single supply operation. Swinging the output near ground introduces distortion. To avoid this an LM10 or similar is added to provide an approximately 1/2V output offset to move the zero point away from ground. This is a compromise between dynamic range and accuracy near zero load.

Connector(s)

■ CM1 Amphenol PT02A10−6P

Amphenol circular bayonet lock receptacle with solder pins

10	shell size number
6	number of pins
???A	current rating / pin
???V AC/DC	voltage rating
???VAC	withstanding voltage, 1 minute
???mΩ	contact resistance Max (initial, after environmental testing)
???MΩ	insulation resistance
23.825mm	flange size (square)
21mm	depth

Aluminum shell, olive drab, cadmium plated.

This is the connector installed on the load cell we have.

■ CF1 Amphenol PT06E10−6S(SR)(DigiKey PT06E10−6S(SR)−ND 29.97$ea)

Amphenol circular bayonet lock plug with solder sockets

10	shell size number
6	number of sockets
???A	current rating / socket
???V AC/DC	voltage rating
???VAC	withstanding voltage, 1 minute
???mΩ	contact resistance Max (initial, after environmental testing)
???MΩ	insulation resistance

Aluminum shell, olive drab, cadmium plated.

This is the mating connector for CM1. Boy, it sure is expensive.

■ CN2 Solder to board

6	number of contacts

No idea what to do here. Maybe just solder to the board without a connector?

■ CN3 Solder to board

2	number of contacts

Not clear what to do here. If we use 12V, i suggest Anderson "powerpole" connectors. Possibly hang them out on a short pig−tail?

■ CN4 Solder to board

2	number of contacts (possibly with added shield)

2 signal connector for twisted pair or shielded wire. Maybe some audio connector???

Integrated Circuit(s)

■ U1 Burr−Brown(Now TI) INA125P, Instrumentation Amp w/Precision Reference (DigiKey INA125P−ND5.15$ea)

Texas Inst. INA125P
DIP−16 package
(2.7,, 36)V supply voltage
(, 460, 525)μA supply current @ 5V (Vs dependent specs below taken at 5V)
(−40, 85)C T_Ambient range

OpAmp
±(, 75, 500)μV initial input offset voltage
±(, 0.25, 2)μV/C tempco of input offset voltage
(, 3, 20)μV/V supply voltage coefficient for input offset voltage
(100, 114,)dB CMRR @ G=100, Vs=5V
(−1.2,−0.8,)V positive output swing from the positive rail
(0.3, 0.15,)V negative output swing from the negative rail
(, 10, 25)nA input bias current
±(, 60,)pA/°C tempco of input bias current
±(, 0.5, 2.5)nA input offset current
±(, 0.5,)pA/°C tempco of offset current
$38\,\mathrm{nV}\big/\sqrt{\mathrm{Hz}}$ voltage noise
0.8μVp−p 1/f voltage noise 0.1−10Hz
$56\,\mathrm{fA}\big/\sqrt{\mathrm{Hz}}$ current noise @ 1kHz
$170\,\mathrm{fA}\big/\sqrt{\mathrm{Hz}}$ current noise @ 10Hz
5pAp−p 1/f current noise 0.1−10Hz
±(, 0.05, 0.5)% Gain error @ G=100
±(25, 100)ppm/°C tempco of Gain
±(, 0.001, 0.01)%FS non−linearity@ G=100
4.5kHz BW @ G=100
375μs 0.01% settling time for 10V output step (±15V supplies)
5μs overload recovery

Reference
±(, 0.15, 0.5)% initial accuracy
±(, 18, 35)ppm/°C tempco of reference
±(, 20, 50)ppm/V supply voltage coefficient for reference voltage
(, 3, 75)ppm/mA load current coefficient for reference voltage
(1,25, 1,)V drop out voltage
1.25V nominal bandgap voltage
±0.5% bandgap initial accuracy
±18ppm/°C tempco of bandgap voltage

```
(* V_RMS *) √(4 k T R B) /. {B → Hz, k → 1.380662*^−23, T → 25 + 273.15} /. R → 350
```

$$2.40063\times10^{-9}\,\sqrt{\mathrm{Hz}}$$

For the load cell (see SG100 below) it was determined that the initial off set is less than 350μV, the DC error over the FS range is less than 45μV, and the DC error tempco is less than 1.5μV/°C . Additionally, since the bridge resistance is about 350Ω the Johnson noise in the bridge can be calculated at 25C as around $2.4\,\mathrm{nV}\big/\sqrt{\mathrm{Hz}}$.

For a 350Ω source impedance, the femto−Amp current noise is negligible. The amplifier noise dominates the Johnson noise from the source. The bandwidth can be limited by 2.2kΩ resistors to, say, a break frequency of 200Hz (reconsider this) with only a couple percent increase in overall noise (prove this (done below)).

```
(* V_RMS total noise *) Sqrt[4 k T R B + ((38*^-9)^2 + (170*^-15 * R)^2) B] /.
   {B → Hz, k → 1.380662*^-23, T → 25 + 273.15} /. R → 350 + 2 * 2.2*^3 // N
```

```
3.90239 × 10^-8 Sqrt[Hz]
```

The total noise assuming two 2.2kΩ resistors (metal film of course) is $39\,\mathrm{nV}/\sqrt{\mathrm{Hz}}$ which is a noise increase of 2.6%, which for our purposes is probably negligible.

```
250 * 39*^-9 * Sqrt[100] / 30*^-3 // N
```

```
0.00325
```

Assuming a bandwidth around 100Hz, the rms noise expressed in pounds is about 0.003 Lbs. This is negligible for our purposes. In fact the bandwidth could be raised to 1kHz with no extreme penalty in extra noise (and possibly should be, what is the bandwidth of the sensor?). Care should be taken not to create frequency aliasing due to a low sample rate.

The 1/f noise is about 0.8μVp−pfrom specs, this is around 0.007 Lbs. Still respectably low for us. The 2.2kΩ metal film resistors should not add unacceptable to the 1/f noise. (Worst case about 0.4μV, better resistors in this class are often 10 times better.)

It appears the AC noise of the INA125P is acceptable, now to move on to the DC noise.

The typical offset is 75μV, but the max is 500μV. I don’t know if i trust the statistics from the load cell people, i’m willing to call it 250μV, making the total max input referred offset 600μV. The output offset is then about 100mV. Since the output swing is only down to 300mV (worst case), a very reasonable (and convenient) offset point is 0.4V. This should give enough headroom to bring the real output zero point well into the linear range.

All the DC error sources due to the instrumentation amp appear to be comparable or less than the sensor itself, so that should work.

The reference also looks ok assuming we run it off the regulated 10V supply.

Looks like the INA125P will work.

■ U2 Linear LT1635CN8, Micropower RR OpAmp w/Reference (improved LM10) (DigiKey LT1635CN8−ND 3.50$ea)

Linear Tech. LT1635CN8
DIP−8 package
(1.2,, 10)V supply voltage
(, 150, 260)μA supply current @ 5V (Vs dependent specs below taken at 5V)
(−40, 85)C T_Ambient range

OpAmp
±(, 0.5, 1.8)mV initial input offset voltage
±(, 3, 7)μV/C tempco of input offset voltage
(85, 97,)dB CMRR @ G=100, Vs=5V
(90, 97,)dB PSRR
(−0.35,−0.2,)V positive output swing from the positive rail sourcing 5mA
(, 125, 250)mV negative output swing from the negative rail sinking 5mA
(, 2.5, 5.5)nA input bias current
±(, 0.2, 0.6)nA input offset current
$50\,\mathrm{nV}/\sqrt{\mathrm{Hz}}$ voltage noise
1μVp−p 1/f voltage noise 0.1−10Hz
$50\,\mathrm{fA}/\sqrt{\mathrm{Hz}}$ current noise @ 1kHz
(45, 200,)V/mV large signal Gain w/RL=1kΩ
(20, 40,)mA short circuit current (either supply)
175kHz GBW product
45V/ms slew rate

Reference
(189, 200, 211)mV reference voltage
±(, 30, 100)ppm/°C tempco of reference
(, 5, 15)nA bias current into pin 8
±(, 20, 100)ppm/V supply voltage coefficient for reference voltage
(, 200, 500)ppm/mA load current coefficient for reference voltage [0−1]mA

All that is done with U2 is the generation of a ~1/2V reference. This reference is applied to the output offset of U1, thus raising the zero point into the guaranteed linear range of U1's output.

The DC and AC noise of this amplifier is not of great concern because its signal is added after the 1st gain stage (G~150) so the 1st stage amplifier dominates the overall noise.

The LT1635 is an improved LM10, probably a genuine LM10 would work here as well.

■ U3 National LP2954AIT, 5.0V Micro Power Low Drop Out Voltage Regulator (DigiKey LP2954AIT−ND 4.30$ea)

National Semiconductor LP2954AIT
TO−220−3 package
(5.5,, 30)V supply voltage
(, 1.1, 2.5)mA ground pin current @ 50mA output
(−40, 125)C T_Junction range
(4.975,, 5.025)V initial output voltage
(, 20, 100)ppm tempco of output voltage
(, 0.03, 0.20)% supply voltage coefficient for output voltage
(, 0.04, 0.16)% load current coefficient for output voltage [1, 250]mA (pulsed)
(, 240, 420)mV drop out voltage
(, 380, 530)mA short circuit current limit
(, 80,)μVRMS output voltage noise 10Hz −100kHz (0.1μF feedback, 33μF output)

U3 is used here as a super power transistor with thermal and short circuit protection. Since the minimum regulated voltage is 5V, the circuit will always start.

Capacitors(s)

■ C1 2.2μF, 50V, tantalum, radial, Kemet T350E225K050AS (DigiKey 399−1447−ND 0.93$ea)

Kemet	T350E225K050AS
2.2 μF	nominal capacitance
±10%	tolerance
50V	WV DC
3.5Ω	ESR @ 100kHz, 25°C
0.9μA	leakage @ 25°C
5%	dissipation factor @ 120Hz, 25°C
tantalum	dielectric
(5.5x8.9)mm	case size (DxH)
2.54mm	lead spacing
???C	operating temperature range

For long lead lengths, a 1μF input capacitor is recommended for U3. Larger values within reason are "better". Probably a 35V rating is adequate.

■ C2 33μF, 16V, tantalum, radial, Kemet T350H336K016AS (DigiKey 399−1406−ND 1.81$ea)

Kemet	T350H336K016AS
33 μF	nominal capacitance
±10%	tolerance
16V	WV DC
1.6Ω	ESR @ 100kHz, 25°C
4μA	leakage @ 25°C
6%	dissipation factor @ 120Hz, 25°C
tantalum	dielectric
(7.6x10.2)mm	case size (DxH)
2.54mm	lead spacing
???C	operating temperature range

Required output capacitor for U3. 10μF is adequate for stability, but noise voltage is slightly decreased by using a larger value.

■ C3 1oμF, Tantalum

C3 was modified in version 1.01. Formerly C3 connected the output and "ground" pin of U3, and was a 0.1μF cap. In v1.01 this was changed to an enormous 10μF unit to ground. The reason for this was instability in the power output loop. Seemingly the lag through U3 and it’s output cap was too long, so the output of the opamp was slowed down tremendously to keep the loop stable. This kills the transient response of the opamp, but this is just not important in this application. (The transient response of U3 is unaffected.)

■ C4, C5 0.1μF, 50V, X7R, axial, BC A104K15X7RF5TAA (DigiKey 1109PHCT−ND3.44$/10)

BC	A104K15X7RF5TAA
0.1 μF	nominal capacitance
±10%	tolerance
50V	WV DC
X7R	dielectric
(2.54x3.81)mm	case size (DxL)
(−55,,125)C	operating temperature range

Bypass capacitors. C3 reduces output noise from U3.

■ C6 2x0.1μF, see C3

Changed to 0.2μF, just use two bypass capacitors in parallel.

Kemet	C430154K5R5CA7200
0.15 μF	nominal capacitance
±10%	tolerance
50V	WV DC
X7R	dielectric
(3.81x7.371)mm	case size (DxL)
(−55,,125)C	operating temperature range

C6 sets the 3dB input bandwidth of the instrumentation amplifier. Since the bandwidth of the transducer remains unknown, we guesstimate 100Hz is a good choice. Due to the nature of cascaded filters, arbitrarily set 200Hz as the bandwidth to the initial filter. The series input resistance includes R5 and R6

```
Solve[f == 1 / (2 π R C), C][[1, 1]] /. {f → 200, R → 2 * 2.2*^3 + 350} //
 EngineeringForm
```

```
C → 167.532 × 10^-9
```

0.15μF is close enough

```
f == 1 / (2 π R C) /. {C → 0.15*^-6, R → 2 * 2.2*^3 + 350}
```

```
f == 223.375
```

Really 0.2μF is probably fine, and saves buying another cap.

```
f == 1 / (2 π R C) /. {C → 0.2*^-6, R → 2 * 2.2*^3 + 350}
```

```
f == 167.532
```

Let's just do that.

Resistors(s)

■ R1, R2 any 5% axial lead resistor 1/10W or more

R1 and R2 combine with C1 to filter input noise from the power supply. To allow operation at 10V the voltage drop across R1+R2 must be kept small. U3's drop out voltage is under 1/2V so

```
R1 + R2 < (Vs - (Vexcite + Vdrop)) / Is
  /. Is → Vexcite / 350 + 20*^-3 /. {Vs → 10, Vexcite → 8.75, Vdrop → 0.5}
```

```
R1 + R2 < 16.6667
```

For convenience, set R1=R2=7.5Ω .

Note that there can be up to a 1/2V drop across R1. This means if the acquisition board is referred to the negative supply there would be of order 1/2V error. This is way too much. R2 could be eliminated, but that is equivalent to assuming that there is no drop throughout the whole ground loop. The only accurate way to deal with this problem is to make a differential measurement of the signal at the acquisitor board. In that case R2 is helpful in that it allows a single supply opamp to make the measurement by raising the common mode voltage above the ground rail.

The power dissipation assuming 50mA is

```
P → I^2 R /. {I → 0.05, R → 7.5} // EngineeringForm (* Watts *)
```

```
P → 18.75 × 10^-3
```

■ R3 1kΩ , Any 5% or better film resistor in axial package (DigiKey 1.00KXBK−ND 0.54$/5)

Yageo 1.00KXBK-ND
1.00kΩ nominal resistance
250V WV DC
(2.3x6.5)mm case size (DxL)
±1% tolerance
±100ppm temp.co.
(−65,+150)C operating temperature range
metal film

R3 isolates the DC voltage setting opamp in U1 from the "ground" pin on U3. The pins could be connected directly, but R3 is beneficial in that it isolates the AC feedback from C3 to the ground pin. It also makes the designer 'feel' better.

```
Solve[f == 1 / (2 π R C), R][[1, 1]] /. {f → 1000, C → 0.1*^-6} // EngineeringForm
```

```
R → 1.59155 × 10^3
```

The maximum ground current from U3 is 2.5mA, the voltage drop across R3 is

```
R * 2.5*^-3 /. % (* Volts *)
```

```
3.97887
```

This is a bit much (10V −5V −4V = 1V output from U1). Cut R3 to 1k , U1 rises to 2.5V, which should be fine even at 8.75V.

■ R4 348Ω , 1% or better metal film resistor in axial package (DigiKey 348XBK−ND 0.54$/5)

Yageo 348XBK-ND
348Ω nominal resistance
250V WV DC
(2.3x6.5)mm case size (DxL)
±1% tolerance
±100ppm temp.co.
(−65,+150)C operating temperature range
metal film

(Repeat the calculations below for a reduced excitation voltage.)

Nominal excitation voltage

```
1.24 * 7
```

```
8.68
```

Target gain (using nominal 3mV/V output spec)

$$\frac{5 - 0.454}{3*^{\wedge}-3 * \%}$$

```
174.578
```

```
Solve[% == 4 + 60*^3 / R4, R4]⟦1, 1⟧
```

```
R4 → 351.746
```

348Ω is an available value. In theory a 348Ω 1% and a 3.0Ω 5% could be used in series, but that seems unnecessary.

This has changed, see above.

This is the gain setting resistor for the instrumentation amplifier. The nominal input range is 30mV. The intentional offset is 0.454V and the maximum output voltage target is 5V, therefore the gain target is

$$\frac{5 - 0.454}{30*^{\wedge}-3}$$

```
151.533
```

The gain formula for U1 is G = 4 + 60k / R4 , R4 is then

```
Solve[% == 4 + 60*^3 / R4, R4]⟦1, 1⟧
```

```
R4 → 406.688
```

DigiKey has 100ppm/°C 1% resistors, which should be adequate. A 402Ω unit is available, the resulting gain is

```
G → 4 + 60*^3 / R4 /. R4 → 402 // N
```

```
G → 153.254
```

This should be plenty close enough.

■ R5, R6 2.2kΩ , Any 5% or better metal film resistor in axial package (DigiKey 2.21KXBK−ND 0.54$/5)

Yageo 2.21KXBK−ND
2.21kΩ nominal resistance
250V WV DC
(2.3x6.5)mm case size (DxL)
±1% tolerance
±100ppm temp.co.
(−65,+150)C operating temperature range
metal film

Mostly these resistors are part of a bandwidth limiting circuit for the input amplifier. They also increase the robustness of the inputs during static discharge. The highest usable value is set either by the amplifier bias currents or the Johnson noise in the resistor. The bias currents are small, so Johnson noise dominates. 2.2kΩ is ok, see the calculations for U1.

■ R7 14.0kΩ , Any 1% or better metal film resistor in axial package (DigiKey 14.0KXBK−ND0.54$/5)

Yageo 14.0KXBK−ND
14.0kΩ nominal resistance
250V WV DC
(2.3x6.5)mm case size (DxL)
±1% tolerance
±100ppm temp.co.
(−65,+150)C operating temperature range
metal film

R7 and R8 together determine the intentional offset voltage applied to the output signal. The value should be about 1/2V to assure that the worst case input offset error does not drive the output below the negative range of the output amplifier (about 0.3V without pull down). Somewhat arbitrarily, limit the ±250 Lbs. range of the load cell to [+250, −25]Lbs. and map this to 0−5V.The target offset Voltage is

```
5 * 25 / (250 + 25) // N (* Volts *)
```

```
0.454545
```

The nominal reference Voltage is 0.2V, the desired ratio of resistances is

```
Solve[% == 0.2 (1 + r), r][[1, 1]]
```

```
r → 1.27273
```

In standard values this ratio is closely approached by 14.0k / 11.0k . Using slightly less useless resistors, it could be 12.7k / 10.0k .

■ R8 11.0kΩ , Any 1% or better metal film resistor in axial package (DigiKey 11.0KXBK−ND0.54$/5)

Yageo 11.0KXBK−ND
11.0kΩ nominal resistance
250V WV DC
(2.3x6.5)mm case size (DxL)
±1% tolerance
±100ppm temp.co.
(−65,+150)C operating temperature range
metal film

See R7

Miscellaneous

■ SG100 Transducer Techniques, LPU−250Tension or Compression Load Cell

Transducer Techniques LPU−250(See http://www.transducertechniques.com/LPU−Load−Cell.cfm)

±1112N	Full Scale (FS) capacity
150%	Allowable overload
30mV/V	output @ FS (compression or tension???)
350Ω	nominal bridge resistance
10V	maximum excitation voltage
±0.1%FS	non−linearity
±0.1%FS	hysteresis
±0.05%FS	non−repeatability
±90ppm/°C	Temperature coefficient of output compared to load
±1%FS	output zero position variation (±correct???)
±18ppm/°C	Temperature coefficient of zero point compared to FS
(+15, 70)C	operating temperature range
(−50,90)C	storage temperature range
76.2mm	puck (case) diameter
25.4mm	puck thickness
???g	device mass

What is the maximum variation in zero point? Assuming that the specifications refer to the maximum value, it is 1% of full scale. Assuming semi−normal scaling this is about 1% of 5V or 50mV. If the INA125 is used, the output limit is around +300mV to ground, so the effective zero point should be around 350mV to compensate for variations in the bridge output. More offset will need to be added to allow for instrumentation amp errors.

What is the estimated accuracy of the sensor? Assume that the figures given (such as ±0.1%FS) refer to 95% confidence intervals and Gaussian distributions, then the estimated standard deviation is 1/2 the given error (±0.1%FS implies σ = 0.05%). Therefore, further assuming independent errors, the total error standard deviation is approximately the RMS sum of the non−linearity, hysteresis, and non−repeatability deviations.

$$\sqrt{2\left(\frac{0.05}{100}\right)^2+\left(\frac{0.025}{100}\right)^2}$$

```
0.00075
```

This works out to about 0.075% standard deviation, or 0.15%FS in terms of the assumed 95% confidence interval. In practice there are probably regions of the input to output relation that are less accurate than this, and many that are more accurate.

The temperature specification for zero point drift indicates ≤18ppm/°C compared to FS. For example, if the FS output is 30mV, the drift per°C should be less than 1/2μV/°C .

The output tempco is given as 90ppm relative to load. As stated this sounds like a gain error. For example the FS output could change up to 2.7μV/°C .

Since the bridge resistance is about 350Ω , the power dissipated in the bridge at 10V is about 285mW.

```
P → V^2 / R /. {V → 10, R → 350} // N
```

```
P → 0.285714
```

The thermal resistance of the bridge element is unspecified, but might be 10's of degrees per Watt. Heating in the bridge will shift the output parameters as outlined. If we assume on order 10°C rise the induced voltage errors are around 30μV.

The DC errors are likewise around 0.15%FS or $45\mu V$ at 10V plus the initial zero offset, which is less than $300\mu V$. This puts the total (maximum) initial error at about $350\mu V$, and the error coefficient is about $1.5\mu V/°C$ (bare knuckle estimate). It is therefore hoped that the instrumentation amplifier for this system has an initial error not much worse than $100\mu V$, and a tempco comparable to $1\mu V/°C$.