Electrical, Computer & Energy Engineering
UNIVERSITY OF COLORADO BOULDER

ECEN 5053-003 Homework Assignment

Course Name: Embedding Sensors and Actuators
Corresponding Module: C3M2
Week Number: 10
Module Name: Force Sensors
Student Name: **Rushi James Macwan**

Note: Correct answer is in Blue Font

Homework is worth 100 points.

Part 1: Each question is worth 6.7 points.

A. Answer the following questions about strain gauge terminology:

A.1 What is the Gage Factor (GF)?

Sol.

A fundamental parameter of the strain gauge is its sensitivity to strain, expressed quantitatively as the gauge factor (GF). Gauge factor is defined as the ratio of fractional change in electrical resistance to the fractional change in length (strain):

$$GF = \frac{\Delta R/R}{\Delta L/L} = \frac{\Delta R/R}{\varepsilon}$$

Courtesy: Reference Link: [1]

A.2 What is the Temperature Coefficient of Resistance (TCR)?

Sol.

Temperature Coefficient of Resistance (TCR) shows the effect of temperature change on resistance change for the strain gauges.

$$\Delta R / R = TCR \times \Delta T$$

$$\Delta T = \text{ambient temperature change}$$

A.3 What is the Temperature Coefficient of Gauge Factor (TCGF)?

Sol.

Temperature Coefficient of Gauge Factor (TCGF) shows the effect of temperature change on the gauge factor (very low for metals, can be high for silicon gauges) for the strain gauges.

$$\Delta GF / GF = TCGF \times \Delta T$$

$$\Delta T = \text{ambient temperature change}$$

A.4 What is apparent strain?

Sol.

Apparent strain is any change in gage resistance that is not caused by the strain on the force element. Apparent strain is the result of the interaction of the thermal coefficient of the strain gage and the difference in expansion between the gage and the test specimen.

In addition to the temperature effects, apparent strain also can change because of aging and instability of the metal and the bonding agent.

Courtesy: Reference Link: [1]

A.5 What is bonded resistance?

Sol.

The resistance introduced in the strain gauge which is responsible for the apparent strain induced in the gauge by bonding the gauge with adhesive at a very high temperature and then using the same at room temperature is considered to be the bonded resistance.

A.6 What is null compensation?

Sol.

The four strain gauges are connected in a Wheatstone bridge network and this network is always assembled from four resistive elements. In the ideal case each of these resistive elements has the same resistance, which in turn leads to a bridge output of zero. This is almost never the real world case. These elements inherently have non--zero magnitude differences so the output of the unstressed Wheatstone bridge will rarely be zero, and this value is known as the **null output** for the bridge. A Wheatstone bridge consisting of gages having tolerance of just 0.05%, (most have much higher tolerances) can lead to an output zero at no load of 2.5% full scale output.

With the application of a single low TCR resistor, either a series resistance added to one of the bridge elements, or a parallel resistance shunting one of the elements, the unstressed bridge output voltage can be set to zero. **Thus, the null value of the bridge is compensated when the null output value of the bridge touches zero. This compensation process to nullify the null output value of the strain gauge bridge is called null compensation.**

Courtesy: Reference Link: [1]

A.7 High resistivity gauges exhibit nonlinearity due to a very large TCR, so that even if adjacent gauges in a four-wire bridge are strained equally and opposite, their resistance changes differently. What can you do differently in the excitation of the bridge to combat this effect?

Sol.

The TCR of a strain gauge can be negated by wiring 4 strain gauges into a 4-wire Wheatstone bridge where two gauges are in compression, two in tension, and the nominal resistances at the calibrated temperatures are identical. Empirically, it is proven that the strain in the gauge at the nominal and elevated temperatures is same.

The effect of thermal strains on the null output of Wheatstone bridge circuits formed from semiconductor gages can be minimized with the proper circuitry technique. Temperature induced resistance changes of adjacent arms of a bridge tend to cancel, but perfect compensation is rarely achieved due to inherent variations between matched gages themselves and the manner in which they are mounted (See Bonded Resistance Section). The effective TCR of a gage as a bridge element (see Equivalent Circuit Resistance and TCR) can be reduced to any desired value by choosing an appropriate shunt and/or series resistor of zero TCR. The use of “dummy--active” gage pairs as adjacent bridge elements is another common technique for temperature compensation. One simple trimming technique for improving the zero shift or no--load temperature compensation utilizes an experimentally determined shunt resistor across the gage with the higher TCR, and then a series resistor can then be found which corrects for the change in resistance introduced by the shunt.

With the application of a single low TCR resistor, either a series resistance added to one of the bridge elements, or a parallel resistance shunting one of the elements, the unstressed bridge output voltage can be set to zero.

Courtesy: Reference Link: [1]

B. A solid steel beam of width .06 meters and height .08 meters has an axial force placed on it of 250,000 Newtons. If the Modulus of Elasticity is 220 Gigapascals, what is the lateral strain ε_y? (Type in a five-decimal number)

Sol. $\boxed{\boldsymbol{\varepsilon_y = 67.47115 \times 10^{-6}}}$

Based on the class slide, we can use the below given formula to calculate the value of σ_x (stress normal to yz plane) in order to calculate the lateral strain value.

$\boldsymbol{\sigma_x = F / A}$ (here, F → axial force and A → the cross-sectional area)

Thus,

F = - 250,000 N (negative value because axial force is placed on the solid steel beam)

A = 0.06 x 0.08 = 0.0048 m^2

Therefore, σ_x = - 250,000 / 0.0048 Nm^{-2} = - 0.052083 GNm^{-2}

Furthermore, lateral strain ε_y can be calculated using the below given formula:

$$\varepsilon_y = -\nu\, \sigma_x / E$$

Here, E = modulus of elasticity = 220 GPa
v = poison's ratio of steel = 0.285 (average of limits) {reference}

Thus,

Lateral strain
= ε_y
= - 0.285 x (- 0.052083 G) / (220 G) = $\boxed{\boldsymbol{67.47115 \times 10^{-6}}}$

C. A cantilever beam has stress created by a downward point load F. The beam is instrumented with 4 strain gauges arranged in a Wheatstone bridge, where the two tensile gauges on the top have equal resistance to each other, and the two compressive gauges on the bottom have equal resistance to each other.

Other dimensions and parameters are given by:

Downward Force, F =	7,500	N
Length, L =	0.15	m
Location of strain gauges, x =	0.110	m
width, b =	0.032	m
height, h =	0.008	m
Poisson's Ratio, Υ =	0.3	
Modulus of Elasticity, E =	220	GPa
Guage Factor, GF =	2	
Output Voltage, V_o =	0.0230	volts
Input Voltage V_i =	10	volts

What is the absolute value of the axial strain ε_x based on beam theory?
What is the absolute value of the axial strain ε_{xg} based on the voltage output of the strain gauge bridge?

Sol. $\boxed{\varepsilon_x = \mathbf{0.00119850} \text{ and } \varepsilon_{xg} = \mathbf{0.00115}}$

Based on the class slides, the absolute value of the axial strain ε_x based on the beam theory can be found as follows:

$$\varepsilon_x = \varepsilon_y * \Upsilon = (\sigma_y / E) * \Upsilon$$

$$\sigma = Mc / I = F (L\text{-}x) (h/2) (bh^3/12) = 6F (L\text{-}x)/bh^2$$

Where

c = distance of load from neutral axis = h/2

I = cross-section modulus

σ = normal stress in beam at point x

E = modulus of elasticity

F = applied force at end of beam

M = bending moment in beam at x

Thus, based on the same above formula, we first calculate the formula for the normal stress in beam at point x.

Therefore, $\sigma_y = 6 \times 7500\ (0.15 - 0.110) / (0.032 \times 0.008^2)$
$= 0.878906\ GNm^{-2}$

Furthermore, **according to beam theory**,

$\varepsilon_X = \sigma_y \times v / E$ (where, v → poison's ratio and E → mod of elasticity)

Thus, $\varepsilon_X = (0.878906\ G \times 0.3) / 220\ G =$ $\boxed{\mathbf{0.00119850}}$

Similarly, to calculate the value of the lateral strain **based on the voltage output of the strain gauge**, we can use the below given formula:

$$(V_o/V_i)/GF = \varepsilon_x$$

Thus, $\varepsilon_{xg} = (0.0230 / 10) / 2 =$ $\boxed{\mathbf{0.00115}}$

D. A strain gauge is used at a temperature far above where it is calibrated. The parameters of operation are given by:

Nominal Temperature, T_0 =	25	°C
Elevated Temperature, T_e =	100	°C
TCGF =	1.150E-04	1 / °C
Nominal Gauge Factor, GF =	2	
Input Voltage, V_i =	10	volts
Output Voltage, V_o =	0.09	volts

Ignoring the effects of TCR, what strains are measured at the nominal and elevated temperatures? What is the percentage error in your measurement?

Sol. **0.0045, 0.0044615 and ERROR → 0.85512455%**

Based on the class slides, the TCGF equation is given as below:

$\Delta GF / GF = TCGF \times \Delta T$
Where, ΔT = ambient temperature change

Thus,

$\Delta T = 100 - 25 = 75\ °C$
$TCGF = 0.0001150\ °C^{-1}$
$GF = 2$

Therefore, the change in the GF value is given as under:

$\Delta GF = TCGF \times \Delta T \times GF = 0.0001150 \times 75 \times 2 = 0.01725$

Based on the calculated value of the change in the GF value, we will calculate the strain values at the nominal and elevated temperatures as under:

Strain at nominal temperature → $(V_0 / V_i) / GF = (0.09/10) / 2 = \boxed{\mathbf{0.0045}}$

Now, if the temperature rises, the gauge factor will also potentially rise. So, the new gauge factor at the elevated temperature will be $GF + \Delta GF$.

Strain at elevated temperature → $(V_0 / V_i) / (GF + \Delta GF)$

$= (0.09/10) / 2.01725 = \boxed{\mathbf{0.0044615}}$

Thus, the percentage error in the measurement of the strain values is:

$= ((0.0045 – 0.0044615) / 0.0045) \times 100\ \%$
$= \boxed{\textbf{0.85512455\% ERROR}}$

E. The TCR of a strain gauge can be negated by wiring 4 strain gauges into a 4-wire Wheatstone bridge where two gauges are in compression, two in tension, and the nominal resistances at the calibrated temperatures are identical. Perform a calculation to prove this empirically, using the information below. Calculate the strain in the gauge at the nominal and elevated temperatures, and show that you get the same number.

Excitation Voltage, V_i =	10	volts
R_1 =	994	ohms
R_2 =	1006	ohms
R_3 =	1006	ohms
R_4 =	994	ohms
Nominal Temperature, T_0 =	25	°C
Elevated Temperature, T_e =	125	°C
TCR	2.00E-05	ohm/ohm/ °C
Gauge Factor, GF =	2	

Sol. $\boxed{\textbf{0.003 in both the case}}$

Based on the class slides, the formula for the TCR of a strain gauge can be given as under:

$\Delta R / R = TCR \times \Delta T$

ΔT = ambient temperature change

Given this equation, we can find the change in the individual strain gauge resistance values as under:

Here,
$\Delta T = 125 - 25 = 100°C$
$TCR = 2 \times 10^{-5}$ Ohm / Ohm/°C

Also, $R_1 = R_4 = 994$ Ohms and $R_2 = R_3 = 1006$ Ohms

Therefore, the changes in the resistance values at the elevated temperature of 125 °C will be as under using the formula for TCR:

Thus,
$\Delta R_1 = \Delta R_4 = (2 \times 10^{-5}) \times 100 \times 994 = 1.988$ Ohms
$\Delta R_2 = \Delta R_3 = (2 \times 10^{-5}) \times 100 \times 1006 = 2.012$ Ohms

Now, the formula for the output voltage to input voltage ratio can be given as under:

$$V_o / V_i = (R_2/(R_1+R_2) - R_4/(R_3+R_4))$$

But, since $R_1 = R_4$ and $R_2 = R_3$, we can reduce the above equation to the form:

$$\boxed{V_o / V_i = (R_2/(R_1+R_2) - R_1/(R_2+R_1)) = (R_2 - R_1) / (R_1 + R_2) \ldots [1]}$$

Now, to calculate the strain value for the strain gauge at the nominal temperature, we can use the above equation as follows:

Strain = $(V_0 / V_i) / GF = [(1006 - 994) / (1006 + 994)] / 2 =$ **0.003**

To calculate the strain value for the strain gauge at the elevated temperature, we can modify equation as follows:

$$\boxed{V_0^* / V_i = ((R_2 + \Delta R_2) - (R_1 + \Delta R_1)) / ((R_1 + \Delta R_1) + (R_2 + \Delta R_2)) \ldots [2]}$$

Thus, the Strain value at the elevated temperature is given as below:

$\text{Strain}_{(\text{elevated temperature})}$
$= [((1006+2.012) - (994+1.988)) / ((994+1.988) + (1006+2.012))] / 2$

= [(1008.012 – 995.988) / (1008.012 + 995.988)] / 2
= [12.024/2004] / 2
= **0.003**

Thus, it can be empirically seen that both the strain values at the nominal and elevated temperature are same.

F. A strain gauge with Constantan wires is bonded to a ferrite steel structure at the nominal temperature. When the structure heats up an apparent strain is output by the gauge. What is the value of the apparent strain for this situation below?

Nominal Temperature, T_0 =	25	°C
Elevated Temperature, T_e =	150	°C
Constantan resistance change, TCR	2.00E-05	ohm/ohm/ °C
Gauge Factor, GF =	2	

Sol. $\boxed{\mathbf{0.0007}}$

To solve this problem, we can use the equation for calculating apparent strain as provided in the class slide:

$$\varepsilon_A = \left[(C_m - C_s) + \frac{TCR}{GF}\right](T - T_0)$$

GF = Gage factor (unitless)

C_m = Coefficient of thermal expansion of material to which gage is bonded (inches/inch/°F)

C_s = Coefficient of thermal expansion of gage (inches/inch/°F)

TCR = Temperature coefficient of resistance (ohms/ohm/°F)

$T - T_0$ = Temperature difference from reference temperature T_0 (°F)

Here,

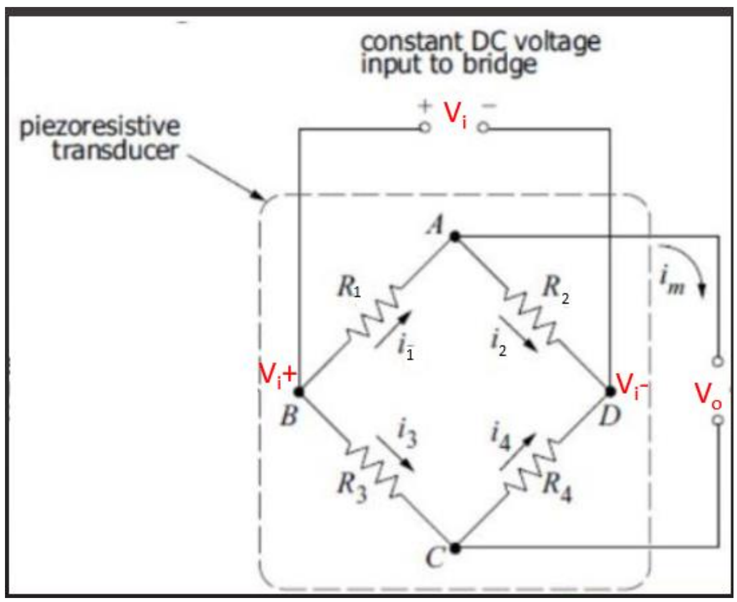
C_m = 10.5 x 10^{-6} in/in/°C {reference}
C_s = 14.9 x 10^{-6} in/in/°C {reference}

Thus,

ε_A (apparent strain) = [((10.5 – 14.9) x 10^{-6}) + ((20 / 2) x 10^{-6})] (150 – 25)
= [(-4.4 + 10) x 10^{-6}] (125)
= **0.0007**

G. Four identical strain gauges with Constantan wires are bonded to a titanium aircraft component at an elevated temperature, and then allowed to cool back to the nominal temperature. The gauges are arranged in the Wheatstone bridge circuit below. R2 and R3 are in tension, and R1 and R4 are in compression. This topology negates any effect due to the TCR of the gauge, and you get the given change of resistance under strain. However, there will still be an error in your strain measurement due to bonded resistance.

When the gauge is used at the nominal temperature, what is the value of the bonded resistance for this situation below? What strain does the gauge measure, ignoring the effects of bonded resistance? What strain does the gauge measure, ***including*** the effects of bonded resistance? What is the resulting % error in your measurement?



Nominal Temperature, T_0 =	25	°C
Elevated Temperature, T_e =	250	°C
Nominal Gauge resistance, R_G =	1000	ohms
Gauge Factor, GF =	2	
Excitation Voltage, V_i =	10	volts
Change in resistance purely due to strain, ΔR_s =	20	ohms

Sol. **1002.835 Ohms, 0.01, 0.009971730145, Error → 0.282698549 %**

To solve this problem, first we calculate the value of the bonded resistance for the given situation when the gauge is used at nominal temperature.

The formula for calculating the bonded resistance value can be given as under:

$$\Delta R = R_G GF(C_m - C_s)(T - T_0)$$

$$R_B = R_G - \Delta R$$

Where,

R_B = Strain gage resistance value after bonding (Bonded Resistance Value) (ohms)

R_G = Strain gage resistance value before bonding (ohms)

ΔR = Change in gage resistance due to bonding (ohms)

GF = Gage factor (unitless)

C_m = Coefficient of thermal expansion of material to which gage is bonded (inches/inch/°F)

C_s = Coefficient of thermal expansion of gage (inches/inch/°F)

$T - T_0$ = Temperature difference from reference temperature T_0 (°F)

In our case,

C_m = 8.6×10^{-6} in/in/°C {reference}
C_s = 14.9×10^{-6} in/in/°C {reference}
R_G = 1000 Ohms
GF = 2
$T-T_0$ = 225°C

Thus, the change in the gage resistance value due to bonding can be given as under:

$\Delta R = 1000 \times 2 \times ((8.6 – 14.9) \times 10^{-6}) (225) = - 2.835$ Ohms

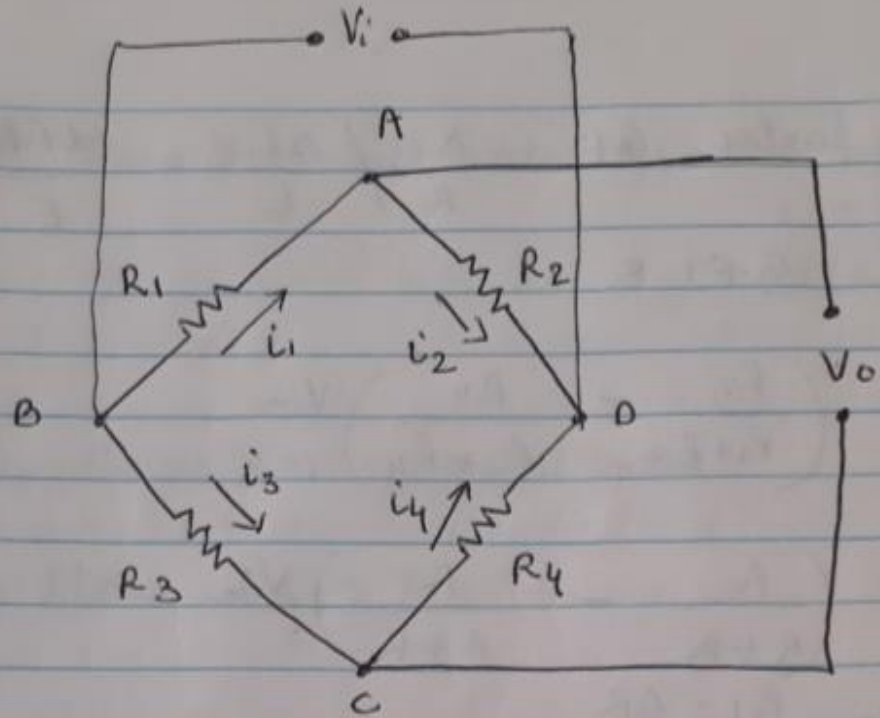
Therefore, the bonded resistance value is:

R_B = 1000 – (- 2.835) = $\boxed{\textbf{1002.835 Ohms}}$

Now, while the effects of bonded resistance are ignored, the strain measured by the gauge can be calculated as follows:

$V_o / V_i = \Delta R_S / R_G = 20 / 1000 = 0.02$

The above formula is obtained based on the attached reference document and the image given below:



Based on this diagram,

let $R_1 = R_2 = R_3 = R_4 = R$

Where, R = Nominal resistance of strain guage.

Let, ΔR = strain-induced change in resistance

$$\therefore R_1 = R - \Delta R \quad \text{——} (1)$$

$$R_2 = R + \Delta R \quad \text{——} (2)$$

$$R_3 = R + \Delta R \quad \text{——} (3)$$

$$R_4 = R - \Delta R \quad \text{——} (4)$$

Putting Eq (1), (2), (3) and (4) in the below eqn:-

$$V_o = \left[\frac{R_2}{R_1 + R_2} - \frac{R_4}{R_3 + R_4} \right] \times V_i$$

$$\therefore \frac{V_o}{V_i} = \left[\frac{R + \Delta R}{2R} - \frac{(R - \Delta R)}{2R} \right]$$

$$\therefore \frac{V_o}{V_i} = \frac{R + \Delta R - R + \Delta R}{2R}$$

$$\therefore \frac{V_o}{V_i} = \frac{2\Delta R}{2R}$$

$$\therefore \boxed{\frac{V_o}{V_i} = \frac{\Delta R}{R}} \quad \text{——} \quad (5)$$

Thus, the strain value can be given as follows:

$\varepsilon_{x\ \text{(ignoring bonded resistance)}}$ = (V_o / V_i) / GF = 0.02 / 2 = $\boxed{\mathbf{0.01}}$

Next, the strain value measured by the gauge while the effects of bonded resistance is considered can be calculated as follows:

V_o / V_i = ΔR_S / R_B = 20 / 1002.835 = 0.01994346

Thus, the strain value can be given as follows:

$\varepsilon_{x\ \text{(with bonded resistance)}}$

= (V_o / V_i) / GF = 0.01994346 / 2 = $\boxed{\mathbf{0.009971730145}}$

Finally, the error in the strain measurement can be given as follows:

% ERROR

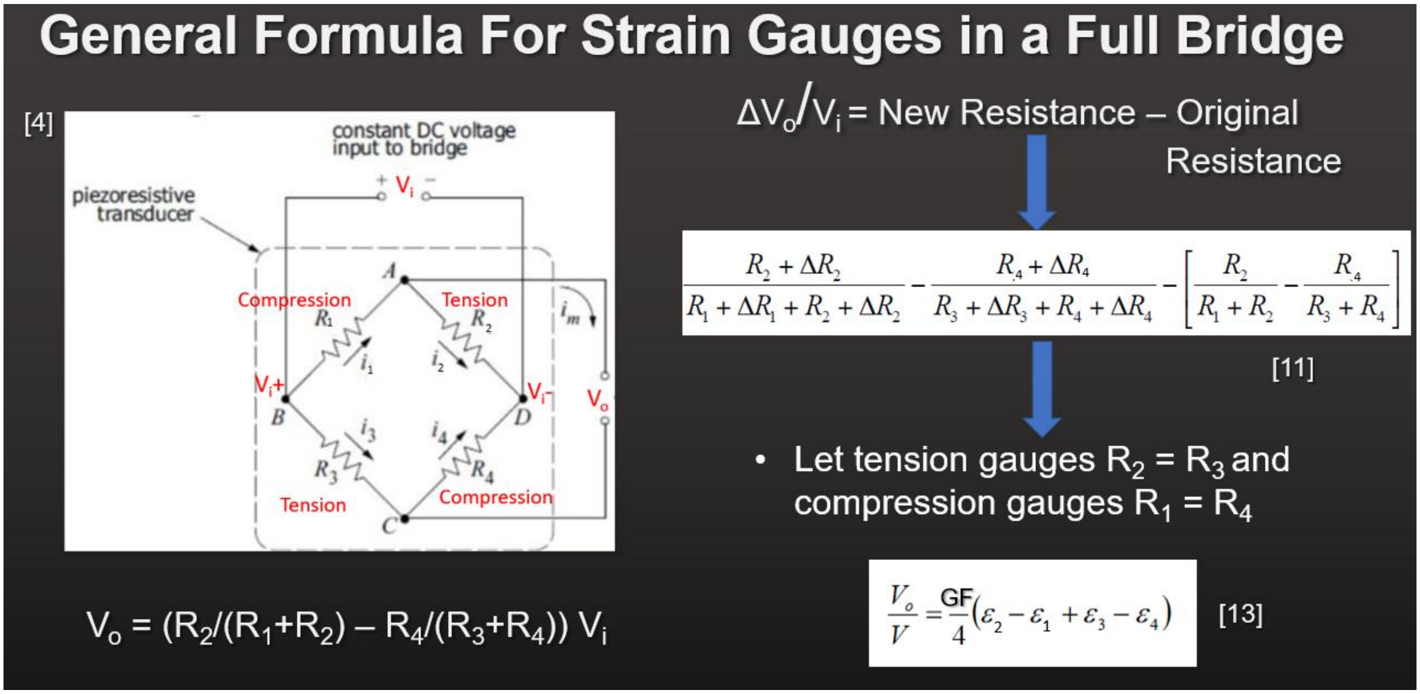
= [($\varepsilon_{x\ \text{(ignoring bonded resistance)}}$ - $\varepsilon_{x\ \text{(with bonded resistance)}}$) / $\varepsilon_{x\ \text{(ignoring bonded resistance)}}$] x 100 %

= [(0.01 - 0.009971730145) / 0.01] x 100 %

= $\boxed{\mathbf{0.282698549\ \%}}$

Courtesy: Reference Link: [1]

H. Derive the General Formula for Strain Gauges in a Full Bridge, shown in the lower right corner of this slide below.



Sol.

Based on the class slides and the information provided in the attached documents, for the given full-bridge circuit consisting of four identical strain gauges, the general formula for strain gauges can be derived as under:

Based on the information provided :-

$R_2 = R_3$ and $R_1 = R_4$ ——— (1)

But, given the construction of the strain gauge bridge (wheatstone) :-

$R_2 = \frac{R_1 \times R_4}{R_3}$ and $R_1 = \frac{R_2 \times R_3}{R_4}$ ——— (2)

Comparing Equations (1) and (2) :

$R_2 = \frac{R_1 \times R_1}{R_2}$ and $R_1 = \frac{R_3 \times R_3}{R_1}$

$\therefore R_2^2 = R_1^2$ $\therefore R_1^2 = R_3^2$

$\therefore R_2 = R_1$ $\therefore R_1 = R_3$ ——— (3)

From Equations (1), (2) and (3), we get :-

$R_1 = R_2 = R_3 = R_4 = R$

Now, let the change in resistance value of any strain gauge be ΔR.

Thus, the new configuration will be :-

$$\left.\begin{array}{l} R_1 \Rightarrow R_1 + \Delta R_1 \Rightarrow R \pm \Delta R \\ R_2 \Rightarrow R_2 + \Delta R_2 \Rightarrow R \mp \Delta R \\ R_3 \Rightarrow R_3 + \Delta R_3 \Rightarrow R \mp \Delta R \\ R_4 \Rightarrow R_4 + \Delta R_4 \Rightarrow R \pm \Delta R \end{array}\right\} \begin{array}{l} \text{since} \\ R_1 = R_2 = R_3 = R_4 \end{array}$$

Now, $\frac{V_o}{V_i} = \frac{R_2}{R_1 + R_2} - \frac{R_4}{R_3 + R_4}$

$$\therefore \frac{V_o}{V_i} = \frac{(R - \Delta R)}{(R + \Delta R) + (R - \Delta R)} - \frac{(R + \Delta R)}{(R - \Delta R) + (R + \Delta R)}$$

$$= \frac{(R - \Delta R)}{2R} - \frac{(R + \Delta R)}{2R}$$

$$= \frac{R - \Delta R - R - \Delta R}{2R}$$

$$= \frac{-2\Delta R}{2R}$$

$$= -\frac{\Delta R}{R} \quad \left(\text{or } \frac{\Delta R}{R}\right)$$

However, ΔR can be both positive & negative so, the sign depends on the resistance value.

$$\therefore \frac{V_o}{V_i} = -\frac{\Delta R_1}{R_1} = \frac{\Delta R_2}{R_2} = \frac{\Delta R_3}{R_3} = -\frac{\Delta R_4}{R_4} = \pm\frac{\Delta R}{R}$$

$$\therefore 4 \cdot \frac{V_o}{V_i} = \pm 4 \frac{\Delta R}{R}$$

$$\therefore 4 \cdot \frac{V_o}{V_i} = \frac{\Delta R_2}{R_2} + \left(-\frac{\Delta R_1}{R_1}\right) + \frac{\Delta R_3}{R_3} + \left(-\frac{\Delta R_4}{R_4}\right)$$

$$\therefore \frac{V_o}{V_i} = \frac{1}{4}\left[\frac{\Delta R_2}{R_2} - \frac{\Delta R_1}{R_1} + \frac{\Delta R_3}{R_3} - \frac{\Delta R_4}{R_4}\right]$$

Now, $(V_o/V_i) = \varepsilon \cdot (GF)$ and $\varepsilon = \frac{(\Delta R / R)}{GF}$ — ④

$$\therefore \frac{V_o}{V_i} = \frac{GF}{4}\left[\frac{(\Delta R_2/R_2) - (\Delta R_1/R_1) + (\Delta R_3/R_3) - \left(\frac{\Delta R_4}{R_4}\right)}{GF}\right]$$

Using Eq ④ :-

$$\therefore \boxed{\frac{V_o}{V_i} = \frac{GF}{4}\left[\varepsilon_2 - \varepsilon_1 + \varepsilon_3 - \varepsilon_4\right]} \quad \text{— ⑤}$$

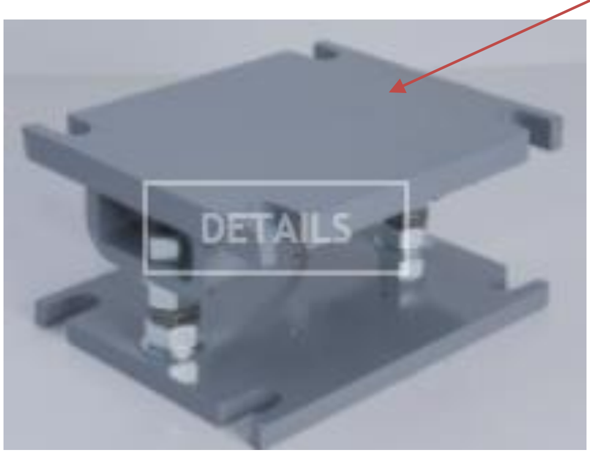
↓

Formula for strain gauges in a full-bridge

Courtesy: Reference Links: [1] [2]

I. Which type of load cell would you specify in each of these situations, how would you use this load cell, and why did you choose it?

I1. You have a minimum amount of clearance in your machine to fit the load cell and you plan on mounting the load cell with a series of bolts directly to the frame of your machine. You put the object to be weighed on a steel platform, which sits on top of the load cell.



Sol. **PANCAKE LOAD-CELL**

Based on the information provided in the class slides on the load cells and the problem specifications, the pancake load cell would be best suited for this problem.

Reasons:

1. The pancake load cell can be used for this problem since it is possible to mount the load cell with a series of bolts directly to the frame of the machine shown in the above image.

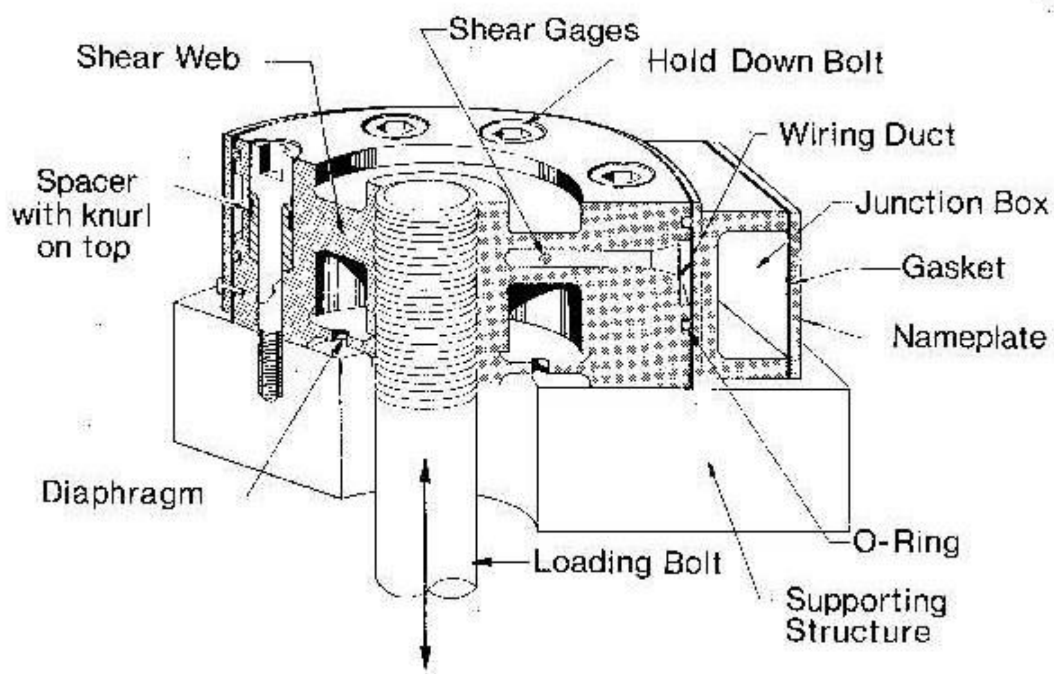
2. Furthermore, pancake load cells have a dual diaphragm design that provides excellent off-axis performance. Pancake load cells can be designed and constructed with welded stainless steel or aluminium and have a threaded thru-hole with standard setting and configurations that allow for the use of standard bolt diameters. Depending on the requirements of the buyer and the manufacturer from which the pancake load cell is to be produced, special threads or metric threads and mounting holes can be added to the pancake load cells being ordered.

3. Pancake load cells can come in several designs like high-accuracy, tension-compression, fatigue-rated, precision, and coiled tubing

injector pancake load cells, and they are built for endurance, off-center loading, and inline applications.

4. This type of load cell also usually comes with two installation options—with a base and without a base. To perform in a linear and predictable manner, the outside diameter of the load cell must be bolted to a flat rigid surface. To ensure that they can also be properly mounted in applications where there is no flat rigid surface, they include a factory installed base that provides a convenient threaded attachment point for easy installation and use in tension and compression. For best performance, repeatability and accuracy, it's a good idea to select one with the base included and permanently installed.

5. Their greatest attribute is that they are industry-standard, so you can generally use one manufacturer's load cell interchangeably with another's. These load cells are also known for their overall capability to handle extraneous loads due to misalignment without mechanical failures.



Based on the above points and as can be seen from the above image, the pan cake load cell can be thus fitted with the frame of the machine using a series of bolts through the holes provided in its construction. The steel platform can thus sit on the top of the load cell which can weigh the object.

Courtesy: Reference Links: [1] [2] [3]

I2. You are picking up a very heavy container with a helicopter using a set of thick steel cables. The load cell measures the tension in one cable section. If the tension gets too high, you know that the cable may break. You send a signal to the helicopter to immediately place the container back on the ground.

Sol. $\boxed{\textbf{S-BEAM LOAD-CELL}}$

Based on the information provided in the class slides on the load cells and the problem specifications, the S-Beam load cell would be best suited for this problem.

Reasons:

1. The S-Beam load cell uses 4 strain gauges that are bonded to the central section of the cell, where strains are high. The strain gauges are bonded at 45 degrees to horizontal or vertical planes in order to measure the shear strains.

2. Furthermore, the threaded holes on the top and bottom of the S-Beam load cell allows mounting of hooks to put beam in tension or blocks to put in compression. Thus, it can measure the tension in heavy cables that are attached to it using the hooks to put the beam under tension. It can be thus used for both measuring the tension in the cables and to measure the weight of a very heavy object like the container in the problem.

3. S beam load cells measure both tension and compression, and they are a good choice for performance and versatility as they are made of alloy steel with nickel plating and have a broad assortment of hardware which comes with them. S beam load cells are suitable for smaller silo vessel and tank weighing uses or to build on OEM projects or testing machines.

4. Each S beam load cell has a threaded hole at each end, making it a very versatile and flexible solution. Different accessories can be added to the S beam load cell to add an extra degree of utility or accuracy, depending on your requirements. The threaded hole can also be fitted to a threaded bar, eyebolt, or rod end bearing depending on the manufacturer and the purpose for which the load cell is to be used.

Based on the above mentioned points, the S-Beam load cell can accurately measure the tension in the cable that is connecting the container and the helicopter. The load cell will also be able to measure the off-centre load (since the helicopter is trying to lift the container which can be off-centre with respect to the container) and thus the load-cell will be accurately able to measure the tension in the cables irrespective of the movements of the helicopter with respect to the container.

Courtesy: Reference Links: [1] [2] [3] [4]

I3. You are designing a cheap electronic bathroom scale, for sale at retail of between $20 and $40. It would look like the one below.

Sol. **HALF-BRIDGE LOAD-CELL (also called STRAIN GAUGE LOAD-CELL)**

Based on the information provided in the class slides on the load cells and the problem specifications, the Half-bridge load cell (also called Strain-gauge load cell) would be best suited for this problem.

Reasons:

1. As found in the reference documents, the half-bridge load cell is simply the strain gauge that is the one that is found in the bathroom scales. This load cell is very simple in construction and is inexpensive.

2. It can measure a load of up to 50 Kgs and when four such load cells are used inside the bathroom scale on the four corners as shown in the image in this problem, the corresponding compression and tension is captured by the individual load cells and provides the weight of the object even though it can be a little off-center.

3. Moreover, the half-bridge load cell is very cheap and therefore it is possible to construct the said bathroom scale within the given price range of 20 – 40 $.

4. We can combine the four single strain gauge load cells (sometimes referred to as Load sensors) using the same wheatstone bridge principle and we can use a combinator to combine the single strain gauge load cells into a wheatstone bridge configuration where the force applied to all four single strain gauge load cells is added to give us a higher maximum load, and better accuracy than just one, and then the combinator can be hooked up to the same amplifier for easier measuring. This is the same layout that would work in a bathroom scale. There would be four strain gauge load cells hooked up to a combinator and an amplifier to give us the weight reading.

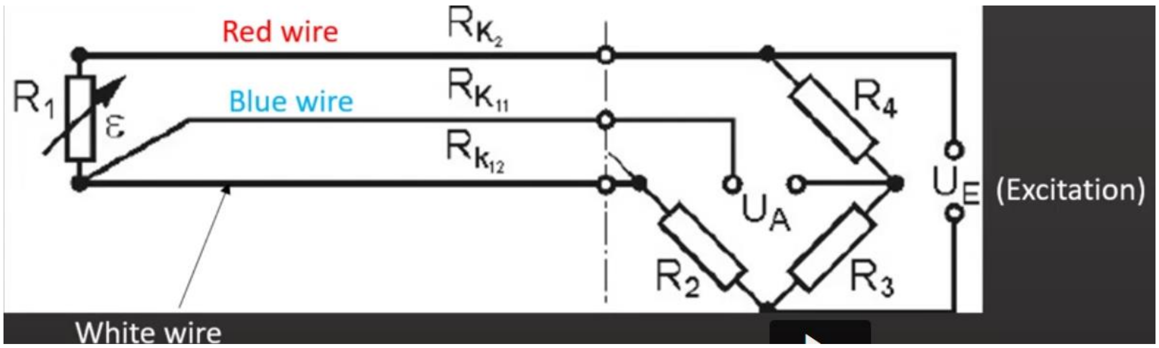
For higher reading values, the disc/button type load cell would suit our purpose. But, in our case of the bathroom scale, the half-bridge load cell is both appropriate enough for our use and is also inexpensive.

Courtesy: Reference Links: [1] [2] [3] [4] [5]

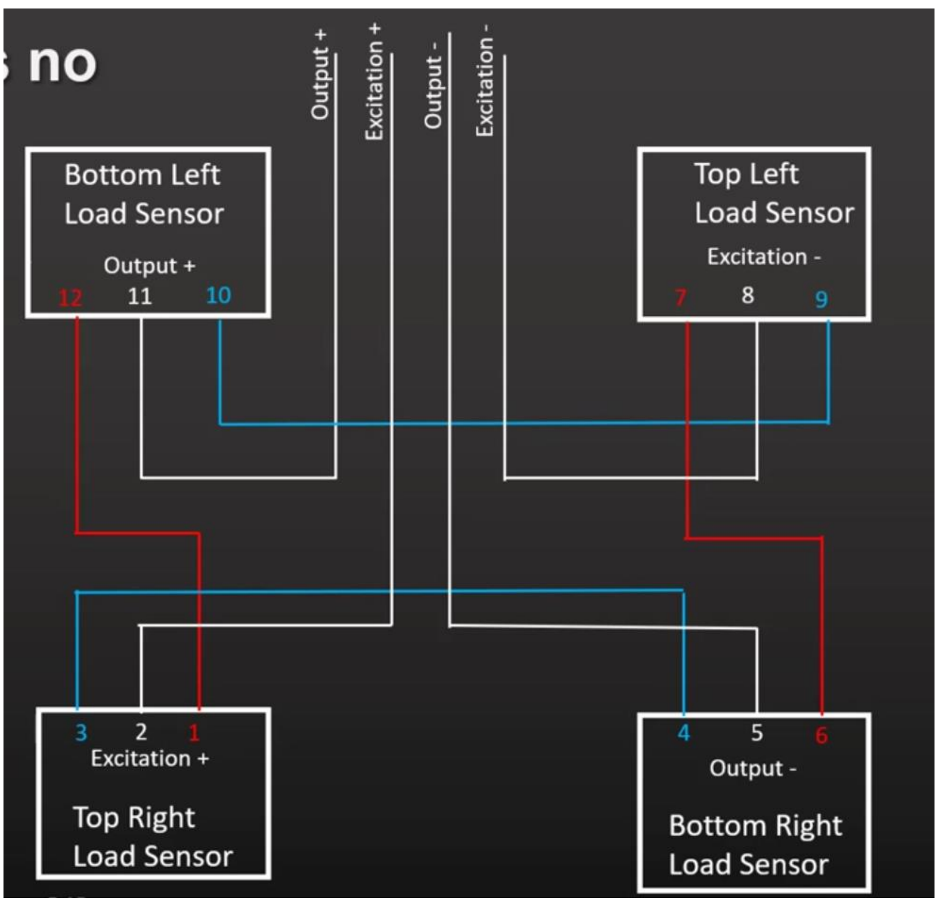
J. How are the load cells wired in our video for tearing down a bathroom scale (C3M2V4.mp4)? What compensation method is left out of this wiring method, and why could the manufacturer still get an accurate reading anyway?

Sol.

Normally, a three wire strain gauge is wired into a quarter bridge circuit as shown below:



The load cells in the bathroom scale shown in the video are wired as shown below:



Here, the four load cells are connected in such a way that every other load cell (or otherwise called load sensor) uses the other load cell as its fixed resistor, fulfilling the conditions that meet the operation in the quarter bridge circuit.

Using this method of wiring the load cells, the temperature compensation method is left out of this wiring method.

Still, the manufacture can get an accurate reading because of two reasons:

1. The bathroom scale is mostly used indoors where the temperature is fixed/stable. This means that the basic quarter bridge requirement that it should be used in a fixed temperature environment is fulfilled and therefore the connections of the load cells as shown above where every load cell treats the other load cell as its fixed resistor (as in a quarter bridge circuit) works the same way as the quarter bridge circuit works. This gives accurate readings.

2. Secondly, once the measurement stabilizes after a couple of seconds, since the person stands still/stable, the readings obtained are accurate because the value stabilizes to the actual value as the quarter bridge theory is applicable to our case.

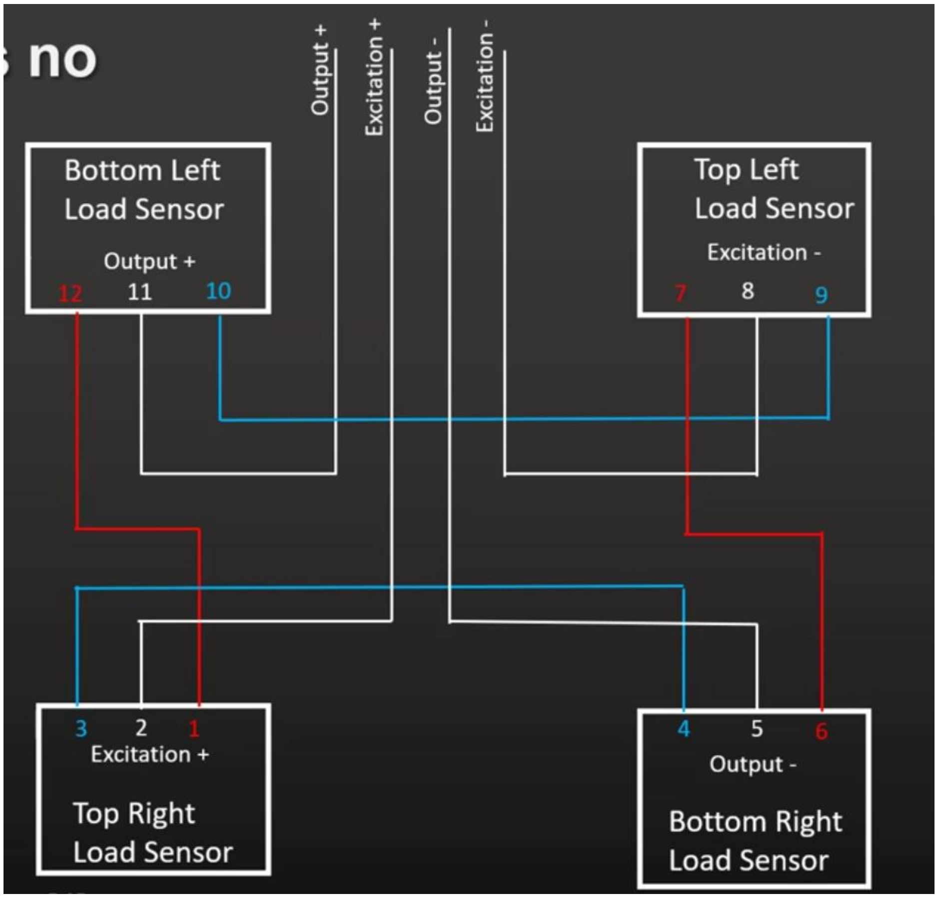
K. Why do the load cells used in our video for tearing down a bathroom scale (C3M2V4.mp4) have no fixed resistors, even though you normally need them to get a reading for the type of bridge circuit used?

Sol.

The load cells used in our video for a bathroom scale do not actually have any fixed resistors even though they are normally required to get a reading for the type of bridge circuit used in our case.

However, **the reason behind this is that the load cells, in our case, themselves are interconnected in a fashion that every load cell treats the other load cell as a fixed resistor** as if they are connected in a quarter bridge circuit. Thus, the load cells have no fixed resistors in our case.

This can be seen from the image below.



Secondly, since the bathroom scale is mostly used indoors in a fixed temperature environment, the concept of the quarter bridge circuit applies to the load cell connection in our case because the quarter bridge circuit can be used only when the temperature is constant. Therefore, the connections shown above for the load cells can provide accurate readings for the bathroom scale.

L. These questions refer to the Cypress Inc. reference AN64846 Getting Started with CapSense.pdf.

L1. Why does your capacitive touch circuit need a hatched ground plane?

Sol.

In general, a proper ground plane on the PCB reduces both RF emissions and interference. However, solid grounds near CapSense sensors, or traces connecting these sensors to the PSoC pins, increase the parasitic capacitance of the sensors.
The increase in parasitic capacitance is unwanted as it reduces the sensitivity. It is thus recommended that you use hatched ground planes surrounding the sensor and on the bottom layer of the PCB, below the sensors.

Thus, the capacitive touch circuit needs a hatched ground plane to reduce the increase in the parasitic capacitance which in turn provides decent sensitivity.

L2. How does Self-Capacitance work?

Sol.

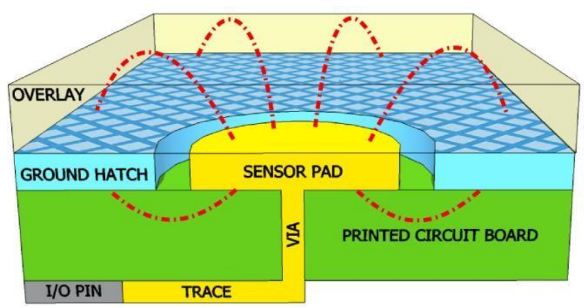
Self-capacitance uses a single pin and measures the capacitance between that pin and ground. A self-capacitance sensing system operates by driving current on a pin connected to a sensor and measuring the voltage. **When a finger is placed on the sensor, it increases the measured capacitance.** Self-capacitance sensing is best suited for single-touch sensors, such as buttons and sliders.

Cypress's CapSense solutions use self-capacitance sensing because it enables efficient use of pins for single-touch sensors and sliders.

WORKING:

In a CapSense self-capacitance system, the sensor capacitance measured by the controller is called C_S. When a finger is not on the sensor, C_S equals the parasitic capacitance (C_P) of the system. This parasitic capacitance is a simplification of the distributed capacitance that includes the effects of the sensor pad, the overlay, the trace between the CapSense controller pin and the sensor pad, the vias through the circuit board, and the pin capacitance of the CapSense controller. C_P is related to the electric field around the sensor pad. Although the following diagram shows field lines only around the sensor pad, the actual electric field is more complicated.

Figure 2-7. C_P and Electric Field



When a finger touches the sensor surface, it forms a simple parallel plate capacitor with the sensor pad through the overlay. The result is called finger capacitance, C_F, and is defined by Equation 1. *C_F is a simplification of a distributed capacitance that includes the effects of the human body and the return path to the circuit board ground.*

$$C_F = \frac{\varepsilon_0 \varepsilon_r A}{D}$$

Where:

ε_0 = Free space permittivity

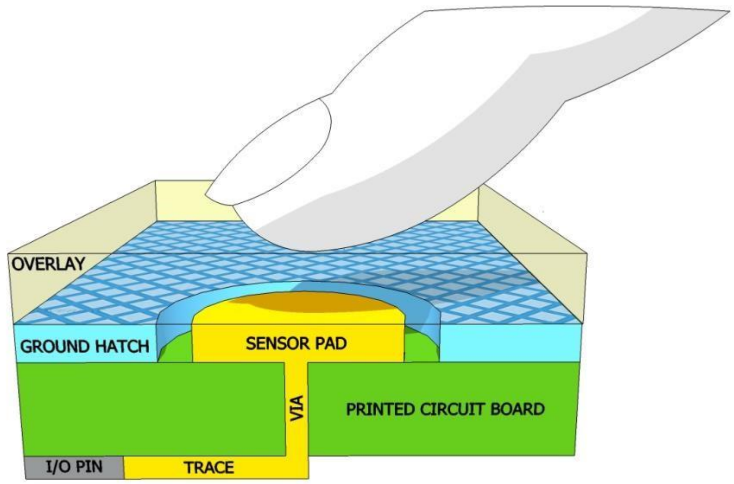
ε_r = Dielectric constant of overlay

A = Area of finger and sensor pad overlap

D = Overlay thickness

Equation – 1

Figure 2-8. CapSense System Equivalent Model



With a finger on the sensor surface, C_S equals the sum of C_P and C_F.

$$C_S = C_P + C_F$$

Equation – 2

L3. How does Mutual Capacitance work?

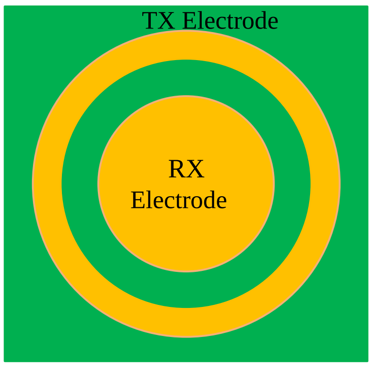
Sol.

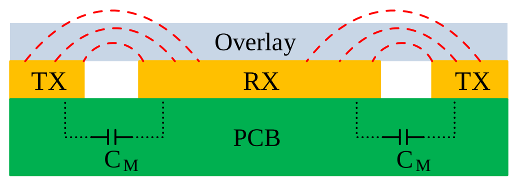
The given figure shows the button sensor layout for mutual-capacitance sensing. Mutual-capacitance sensing measures the capacitance between two electrodes. One of the electrodes is called the “transmit” (TX) electrode and the other electrode is called the “receive” (RX) electrode.

In a mutual-capacitance measurement system, a digital voltage (signal switching between V_{DDD} and GND) is applied to the TX pin and the amount of charge received on the RX pin is measured. The amount of charge received on the RX electrode is directly proportional to the mutual capacitance (C_M) between the two electrodes.

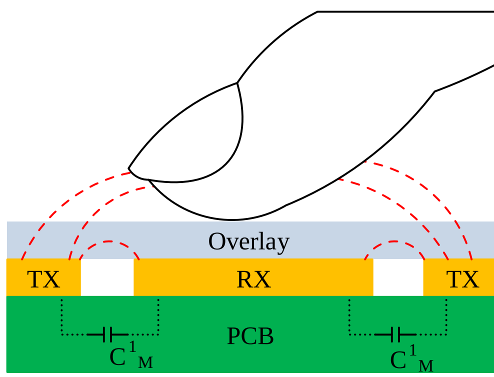
<u>When a finger is placed between the TX and RX electrodes, the mutual-capacitance decreases to C^1_M as shown in the figure.</u> <u>Because of the reduction in mutual-capacitance, the charge received on the RX electrode also decreases.</u> <u>The CapSense system measures the amount of charge received on the RX electrode to detect the touch/no touch condition.</u>

Mutual Capacitance Sensing Working





Mutual Capacitance with Finger Touch



The mutual-capacitance effect is best suited to multi-touch systems such as touchscreens and trackpads.

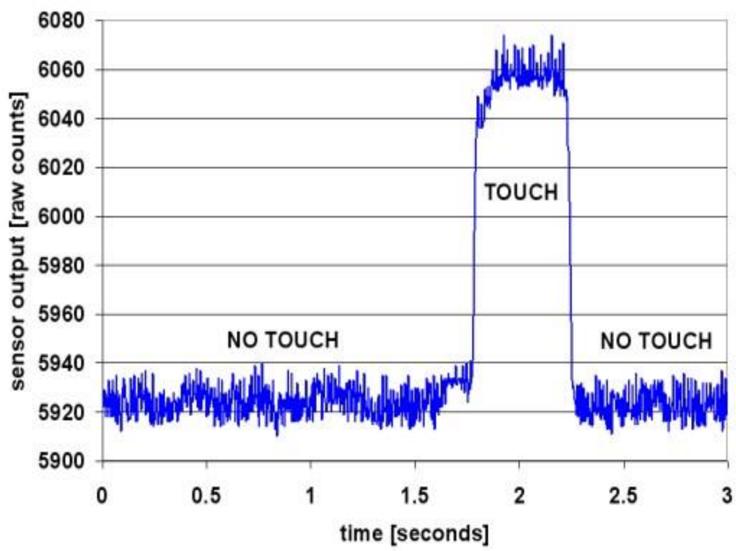
L4. What is the raw count of the CapSense hardware?

Sol.

Sensor capacitance is converted into a count value by the CapSense algorithm. The unprocessed count value is referred to as raw count. Processing of the raw count results in ON/OFF states for the sensor.

The CapSense hardware converts the sensor capacitance into a digital count, called raw count. The raw count is interpreted as either a TOUCH or NO TOUCH state for the sensor, as shown in the figure. The numerical value of the raw count is the digital representation of the sensor capacitance, and increases as the capacitance increases. The raw count is directly proportional to the average current drawn out by the sensor capacitance, which increases when the sensor capacitance increases. Sensitivity is a measure of how much the output will change for a given change on the input. The sensitivity of the CapSense sensor has units of counts-per-pF.

Sensing Algorithm Output



<u>Thus, the unprocessed digital count output of the CapSense hardware block that represents the physical capacitance of the sensor is called raw count of the CapSense hardware.</u>

L5. How is proximity sensing done using CapSense hardware? How is such a system constructed?

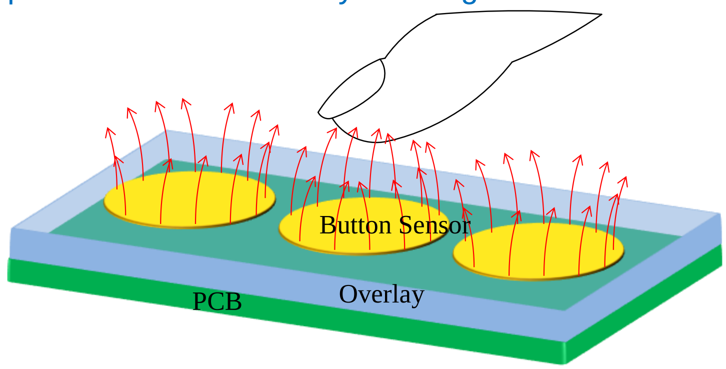
Sol.

Proximity sensing is the process of detecting a nearby object without any physical contact. **<u>Proximity sensors use an electromagnetic field, beam of electromagnetic radiation, or changes in ambient conditions to detect the proximity of a nearby object.</u>** There are various types of proximity sensors such as capacitive, inductive, and magnetic, Hall Effect, optical, and ultrasonic sensors; each has its own advantages and disadvantages. **<u>Capacitive proximity sensing has gained huge popularity because of its low cost, high reliability, low power, sleek aesthetics, and seamless integration with existing user interfaces.</u>**

A capacitive proximity sensor can be constructed using one of the following methods:

Button: A button sensor, when tuned for high sensitivity, can be used as a proximity sensor, as shown in Figure 3-35 The proximity-sensing distance is directly proportional to the sensor area. Because the diameter of a button sensor typically ranges from 5 mm to 15 mm, the proximity-sensing distance achieved with a button sensor is very less when compared to other sensor implementation methods.

Figure 3-35. CapSense-based Proximity Sensing with Button Sensor

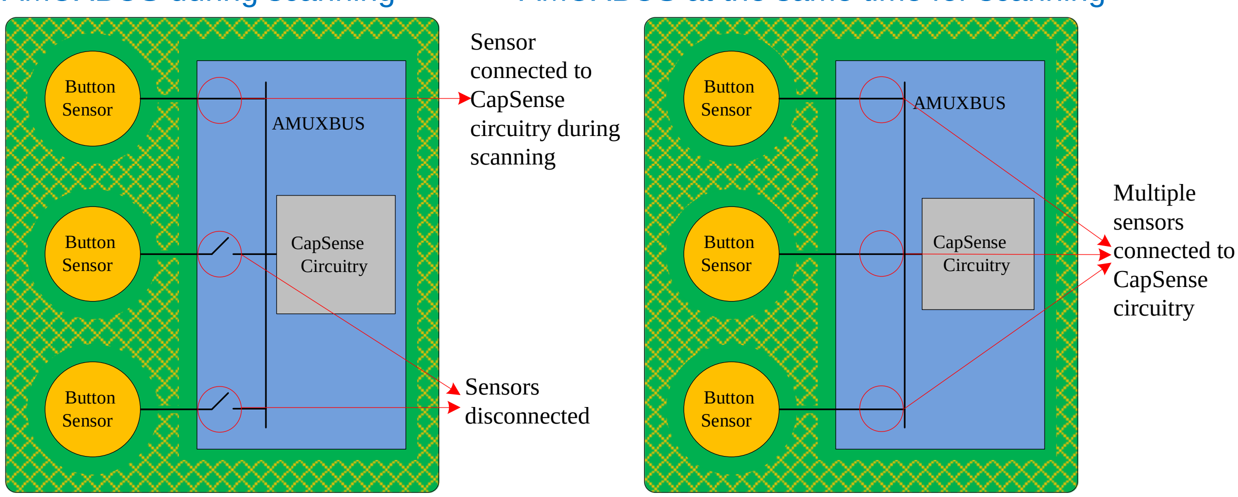


Sensor Ganging: Sensor ganging refers to connecting multiple sensors (Buttons, Proximity trace, Proximity loop) to the CapSense circuitry and scanning them as a single sensor as shown in Figure 3-36 (b). Ganging multiple sensors increases the effective sensor area and results in higher proximity-sensing distance, but ensure that the C_P of the ganged sensor does not cross the maximum C_P limit of 45 pF. Refer to AN92239 for details on how to implement proximity sensing using sensor ganging.

Figure 3-36. CapSense-based Proximity Sensing with Sensor Ganging

(a) Only one sensor is connected to AMUXBUS during scanning

(b) Multiple sensors connected to AMUXBUS at the same time for scanning

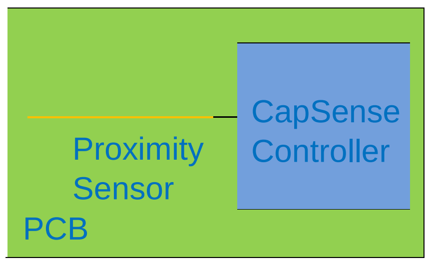


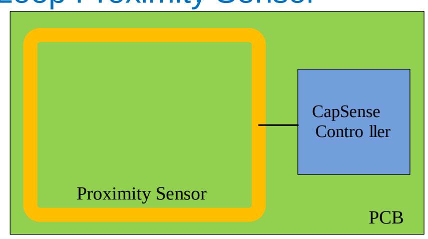
PCB Trace: A long PCB trace on a FR4 or a Flexible Printed Circuit (FPC) board can form a proximity sensor. The trace can be a straight line (Figure 3-37 (a)), or it can surround the perimeter of a system's user interface, as shown in Figure 3-37 (b). Implementing a proximity sensor with a PCB trace has the following advantages when compared to other sensor implementation methods:

- Proximity-sensor C_P is less
- Proximity-sensing distance is higher because more electric field lines couples to the hand
- More appropriate for mass production

Figure 3-37. CapSense-Based Proximity Sensing with PCB Trace

(a) Bar Proximity Sensor 1. (b) Loop Proximity Sensor





Wire: A single length of wire works well as a proximity sensor. The proximity distance achieved with a wire loop sensor is higher compared to a **PCB** trace. But using a wire sensor is not an optimal solution for mass production because of manufacturing cost and complexity.

M. Go to Google Patents (www.patents.google.com) and download US patent 8,040,142. Read the patent and answer the following questions:

M.1 How does the prior art circuit of Figure 1B work to sense a change in capacitance? What are the limitations of this method if you try to use it to sense a change in capacitance due to a finger touch?

Sol.

FIG. 1B is a block diagram illustrating a conventional capacitance sensor. It **illustrates another capacitance sensing technique using a charge transfer mechanism**.

Working:

FIG. 1B illustrates a conventional capacitance measurement circuit **101** including three switches **105** with control terminals **φ0**, **φ1**, and **φ2**, and summing capacitor **110** having a capacitance C_{SUM}, and an analog to digital ("ADC") converter **115**. Capacitance measurement circuit **101** may be used to sense changes in a DUT capacitor **120** having a changing capacitance C_{DUT}.

During operation, capacitance measurement circuit **101** operates as follows to sense capacitance changes on DUT capacitor **120**. First, summing capacitor **110** is discharged to a ground potential by asserting control terminal φ**0** to open circuit switch SW**0** and by asserting control terminal φ**1** to close circuit switch SW**1**. Once discharged to ground, integrating capacitor **110** is disconnected from ground by asserting φ**1** to open switch SW**1**. Then, DUT capacitor **120** is charged to the supply voltage VS by asserting φ**0** to open circuit switch SW**0** and asserting φ**2** to close circuit switch SW**2**. Once DUT capacitor **120** charges to the supply voltage VS, the charge on DUT capacitor **120** is transferred onto summing capacitor **110** and distributed between the two capacitors. Charge transfer occurs by asserting φ**1** and φ**2** to open circuit switches SW**1** and SW**2**, respectively, and asserting φ**0** to close circuit switch SW**0**.

Operation:

The above stages of charging DUT capacitor **120** and transferring the charge onto summing capacitor **110** are repeated a fixed number times causing the voltages of nodes N**1** and N**2** to ramp with time as illustrated in line graphs **130** and **135**, respectively. After a fixed number of consecutive charging stages and charge transferring stages, ADC converter **115** samples the final voltage on node N**2**. **The capacitance C_{DUT} is determined based on the output of ADC converter 115 and is proportional to the voltage at node N2 after the final charge transfer stage.**

Limitations:

Because the capacitance deviation of a capacitance sense switch due to a finger press is small compared to the underlying capacitance of the switch itself, the above two capacitance sensing techniques can be susceptible to external noise, interference, or other environmental factors.

For example, parasitic capacitances may couple to the user interface, electromagnetic interference ("EMI") may disrupt capacitance measurements and control signals, deviations in operating temperature can cause thermal expansions and dielectric variations that affect capacitance measurements, user error can result in malfunctions, and so forth.

These environmental factors can often result in disruptive capacitance deviations that are larger than the capacitance changes induced by a finger interaction with the capacitance sense interface.

<u>M.2</u> What can you tell from figure 4 capacitance values about the capacitances of the five CAP sensors represented by the curves S0, S1, S2, S3, and S4?

FIG. 4 is a chart illustrating a scanning technique of use with a multi-sensor capacitance sense interface for user activations, in accordance with an embodiment of the invention.

Chart 400 plots capacitance values for CAP sensors 325 versus time. The capacitance values plotted along the y-axis of chart 400 are values determined by capacitance measurement circuitry 305 that are representative of a measured capacitance of CAP sensors 325.

In one embodiment, the capacitance values are clock cycle counts gated by relaxation oscillator circuitry included within capacitance measurement circuitry **305**. In one embodiment, the capacitance values are the output of an analog-to-digital ("ADC") converter (e.g., ADC **115**).

Each trace S0, S1, S2, S3, S4 represents a measured capacitance value associated with a corresponding one of CAP sensors 325 (only five traces are illustrated so as not to clutter the drawing).

Each CAP sensor **325** is sampled in sequence by capacitance measurement circuitry **305** one time during each scan cycle (e.g., scan cycles **1**, **2**, **3**, **4**, and **5** are illustrated). For example, CAP sensor **325**A (represented by trace S**0**) is sampled at times T**0**, T**6**, T**8**, T**10**, T**12** . . . , CAP sensor **325**B (represented by trace S**1**) is sampled at time T**1**, CAP sensor **325**C (represented by trace S**2**) is sampled at time T**2**, and so on.

Traces S0 to S4 have been staggered vertically for clarity; however, if CAP sensors 325 are physically identical in size and orientation, traces S0 to S4 may in fact overlap with minor deviations due to localized variations in the capacitances of each CAP sensor 325.

During operation, the baseline capacitance of CAP sensors **325** may drift due to a variety of environmental factors, such as temperature changes, dynamically changing parasitic capacitances, localized disturbances, electromagnetic interference ("EMI"), or otherwise. **This drift is illustrated by the wandering traces S1 to S4.**

Furthermore, chart 400 illustrates a user activation of CAP sensor 325A sometime between the samplings of CAP sensor 325A at time T6 and time T8. In one embodiment, when the measured capacitance value of CAP sensor **325**A crosses the activation threshold, the user activation of CAP sensor **325**A is registered or acknowledged by software or hardware logic coupled to capacitance measurement circuitry **305**.

<u>M.3</u> What is the purpose of the Touch Sense Logic in figure 5?

FIG. 5 is a functional block diagram illustrating a system for improved capacitive touch sensing, in accordance with an embodiment of the invention.

FIG. 5 is a functional block diagram illustrating a system **500** for improved capacitive touch sensing, in accordance with an embodiment of the invention. The illustrated embodiment of system **500** includes capacitance sensor circuit **300**, touch sense logic **505**, application logic **510**, and I/O device **515**.

Touch sense logic **505** may be implemented entirely in software or firmware (e.g., machine readable instructions), entirely in hardware (e.g., application specific integrated circuit, field programmable gate array, etc.), or some combination of both.

Touch sense logic 505 analyzes signal 550 to compensate for various environmental factors (e.g., temperature drift), filter noise, reject false activation events (e.g., reject ESD events), interpolate higher resolution from capacitance sense interface 315, and compensate for various other user interactions with the capacitance sense interface 315.

Touch sense logic 505 analyzes signal 550 to determine whether an actuations of CAP sensors 325 should be registered (acknowledged) as valid touch events or rejected (masked) as false touch events.

M.4 How does the Baseline Drift Compensation Technique work? How does this system guard against long-term drift of capacitance measurements?

Baseline Drift Compensation Technique

FIG. 6 is a chart **600** illustrating a baseline drift compensation technique for improved capacitive touch sense operation, in accordance with an embodiment of the invention. In general, capacitance sensor circuit **300** measures a change in capacitance from a deactivated state to an activated state of a selected CAP sensor **325**. **In one embodiment, baseline logic 525 monitors the difference between the current value of signal 550 and a historical or baseline value. Thresholds for determining whether an activation event has in fact occurred are set related to these historical or baseline values. Measured capacitance values that pass over the activation threshold are considered to be touch events. Accordingly, it is important to accurately track the baseline capacitance values associated with CAP sensors 325 should they drift over time.**

Background (or "parasitic") capacitance may change slowly as a result of environmental factors (temperature drift, electrostatic charge build up, etc.). The activation threshold values should be adjusted to compensate for this background capacitance and other factors. This may be done by monitoring signal 550 in real-time, and updating the baseline capacitance value on a regular basis based on the actual capacitance values measured during each sampling cycle.

In one embodiment, the baseline capacitance value for each CAP sensor **325** is tracked and updated using a weighted moving average. As illustrated in FIG. 6, when an activation is sensed, the baseline capacitance value is held steady so that the elevated values due to the user interaction do not skew the baseline capacitance value calculation. Accordingly, during activations the baseline update algorithm is disabled and the measured capacitance values masked so as to hold the baseline capacitance value steady until the CAP sensor is deactivated.

The rate at which the baseline capacitance values are updated can be set as part of the system design and even updated by users at a later date. The automatic update rate or interval may also be set by the user to compensate for expected environmental variance. In the event that environmental changes are too rapid, the automatic update rate can be adjusted at run-time to compensate.

The baseline capacitance values for each CAP sensor 325 may be adjusted individually or as part of a sensor group (e.g., sensor groups 340 or 350). **Tracking and updating baseline capacitance values for each CAP sensor 325 allows different environmental effects to be compensated in each CAP sensor 325 independently. Alternatively, a group baseline capacitance value compensation enables variations to be averaged over a group of CAP sensors 325.** Accordingly, an embodiment of the invention enables convenient group compensation of baseline capacitance values and the update rate to be applied to a group of CAP sensors 325. **If the baseline capacitance values drift, then a weighted moving average may be applied to update the baseline average and track the baseline drift.**

M.5 What is the purpose of the Finger on Startup Detection Technique? Under what condition is this technique implemented in the code?

Finger on Startup Detection Technique

FIG. 7 is a chart **700** illustrating a finger on startup detection technique **for improved capacitive touch sense operation**, in accordance with an embodiment of the invention.

In one embodiment, the baseline capacitance values discussed above are initialized upon booting or starting system 500. **If during the initialization procedures (e.g., time T0 to T1 in FIG. 7), one or more CAP sensors 325 are actuated by a user (e.g., the user has his finger on one or more CAP sense buttons), it may be necessary to detect this condition and quickly update the startup baseline capacitance values.** After the user removes the activation (time T**2** in FIG. 7), the measured capacitance value of signal **550** will change to a negative value below the startup baseline capacitance value. Simply relying on the baseline logic **525** to slowly track down the startup baseline capacitance value using the weighted moving average can take too long, during which time user interaction with capacitance sense interface **315** will not be recognized. **Accordingly this finger on startup condition should be quickly recognized and compensated.**

In one embodiment, the finger on startup condition is determined by startup logic **520**, if the measured capacitance values of signal **550** cross a negative finger threshold below the startup baseline capacitance value and remains below it for a predetermined period of time (Tp). If this condition is found to be valid, then the startup baseline capacitance value is immediately updated by averaging the capacitance values measured after signal **550** dropped below the negative finger threshold.

M.6 Why would you use the Activation with Hysteresis Technique work? How does it work?

Activation with Hysteresis Technique

FIG. 8 is a chart **800** illustrating activation of CAP sensors **325** using hysteresis **for improved capacitive touch sense operation**, in accordance with an embodiment of the invention. Depending on the scan rate of a capacitance sense interface **315**, rapid repeat activation of the CAP sensors **325** may result in measured capacitance values that do not return all the way to the baseline capacitance value between activations. Unless compensated for, rapid repeat activations may not be detected.

To compensate for rapid repeat activations, hysteresis logic 530 may add hysteresis to the detection algorithm by applying two separate thresholds for determining when a selected CAP sensor 325 is activated and when the selected CAP sensor 325 is deactivated. As illustrated in FIG. 8, hysteresis logic 530 may add an activation threshold and a lower deactivation threshold.

When a measured capacitance value cross the activation threshold, the corresponding CAP sensor **325** is registered as “activated.” When the measured capacitance value falls below the deactivation threshold, the corresponding CAP sensor **325** is deemed “deactivated.”

In this manner, the measured capacitance value need not return all the way to the baseline capacitance value before an activation is deemed deactivated, nor does the measured capacitance value need to return to the baseline capacitance value to register a subsequent activation.

M.7 How does the ESD Compensation Technique determine whether a change in capacitance is from a finger touch or through ESD?

ESD Compensation Technique

FIG. 9A illustrates typical measured profiles for a finger event and an ESD event on CAP sensors **325**, in accordance with an embodiment of the invention. **As can be seen, the measured capacitance values rise and fall gradually as a user finger (or other conductive device) approaches and departs capacitance sense interface 315. In contrast, an ESD event is typified by a rapid spike above the baseline capacitance value, followed by a rapid drop below the baseline capacitance value followed by a ringing or transients with rapidly declining envelope.**

In order to prevent false activation, signal **550** is evaluated by ESD logic **535**. The nature of ESD events is to inject large fast transients into the measured capacitance values, as illustrated in FIG. 9A. ESD logic **535** can reject these

ESD events by quickly recognizing these transients and masking the false event for a period of time.

ESD logic 535 can recognize an ESD event by monitoring the slope between consecutive samplings of the measured capacitance values (e.g., calculating the derivative of traces S0 to S4 is real-time) and determining whether the measured capacitance values cross an ESD threshold below the baseline capacitance value. When the derivative value is significantly faster than typical of human activation, an activation may be rejected as an ESD event.

In one embodiment, ESD logic **535** determines that an ESD event has occurred if: (1) the slope of the measured capacitance value turns positive and has a magnitude greater than a positive slope threshold (POS_SLOPE_TH), (2) the slope of the measured capacitance value then turns negative and has a magnitude greater than a negative slope threshold (NEG_SLOPE_TH), and (3) the measured capacitance value cross an ESD threshold (ESD_TH) below the baseline capacitance value. ESD logic **535** may apply a fourth requirement that conditions (1), (2), and (3) occur within a predetermined time threshold. **If ESD logic 535 determines these conditions are valid, then the activation event is rejected as a false activation or ESD event and all activation on the particular CAP sensor 325 will be rejected or masked for a period of time.**

<u>M.8</u> Why would you use Variable Resolution via Interpolation? How does this system work?

Variable Resolution Via Interpolation

Some application logic **510** may require the use of a sliding switch, such as radial slider interface **320** or linear slider interface **330**. In many cases the resolution desired is much finer than is physically possible by simply using a greater number of smaller CAP sensors **325**. There may also be a desire for detecting greater resolution or granularity than there are physical CAP sensors in the physical array of CAP sensors. **Thus, the variable resolution via interpolation is used.**

<u>Example of such working system:</u>

In one embodiment, interpolation logic **540** includes detection algorithms to assess the signal strength on each CAP sensor in the array and map the measured values onto a user selected number of CAP sensors (e.g., interpolation).

Typical examples include mapping eight linearly spaced CAP sensors **325** onto a 0 to 100 scale, mapping twenty CAP sensors **325** onto a 0 to 256 scale, or mapping eight CAP sensors **325** onto six separate capacitance sense buttons. The calculations may be done with fractional fixed point for efficiency in a limited capability microcontroller.