MICROCHIP

AN1237

Inductive Touch Hardware Design

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Note:	Microchip inductive mTouch™ sensing solution is proprietary technology. It is available to customers free-of-charge under a license agreement permitting use and implementation of the technology on any PIC® microcontroller or dsPIC® digital signal controller.

INTRODUCTION

Microchip's inductive touch user interface system uses the electromagnetic interaction between a conductive target and a sensing coil to detect the pressure applied by the user on the surface of a touch panel. The pressure created by the user causes a minute flex in the target, changing the spacing between the coil and target, which results in a change in the impedance of the sensing coil. A simple impedance measurement circuit is then multiplexed between the various sensor coils to detect the impedance changes indicating a user's press on a specific coil.

THEORY OF OPERATION

To measure the impedance of a sensor coil, the measurement system must first excite the coil with a pulsed current. This produces a pulsed voltage across the coil that is proportional to both the current and the impedance of the coil. The impedance system then converts the pulsed voltage into a DC voltage proportional to the amplitude of the pulsed voltage. The resulting DC value is converted to a binary number using an ADC, and software in the system then decides if the shift in impedance is indicative of a user's touch.

To assist in maintaining consistency in the readings, a reference coil has been added to remove variations in the readings due to long term variations in the current and due to temperature shifts. The reference coil is in series with the sensor coil, and is measured as part of the conversion process. The resulting value for the reference coil is divided into the sensor coil, resulting in a "normalized value" for the coil. By using the impedance ratio of the two coils, any variation in the impedances due to temperature or long term drive current variation fall out and we are left with a temperature and voltage compensated value.

DRIVER

At the start of the measurement process is the pulsed current driver. This circuit must produce a current of sufficient amplitude and frequency to obtain measurable voltage across the coil. The relationship between voltage, current, inductance and frequency is given in Equation 1.

EQUATION 1: RELATIONSHIP BETWEEN DRIVE AND COIL INDUCTION

$$\Delta V_{OUT} = \frac{(\Delta I_{DRV} \bullet L_{COIL} \bullet F_{DRV})}{2}$$

V_{OUT} = Pulsed Output Voltage

ΔI_{DRV} = AC Drive Current Amplitude

F_{DRV} = AC Drive Current Frequency

L_{COIL} = Inductance of the Sensor Coil

NOTE: These equations assume a 50% duty cycle.

Note:	The design of the coils is covered in application note *AN1239, "Inductive Touch Sensor Design"*.

For a typical design, a drive current of 2 mA at 2 MHz will produce a pulsed output voltage of 10 mV across a 5 uH coil.

Figure 2 shows a typical circuit for generating the pulse current drive. R1 and C1 convert the square wave output of the PWM into a triangle drive. The transistor then converts the triangle waveform into a drive current, with the maximum current level determined by R3.

FIGURE 1: INDUCTIVE TOUCH BLOCK DIAGRAM

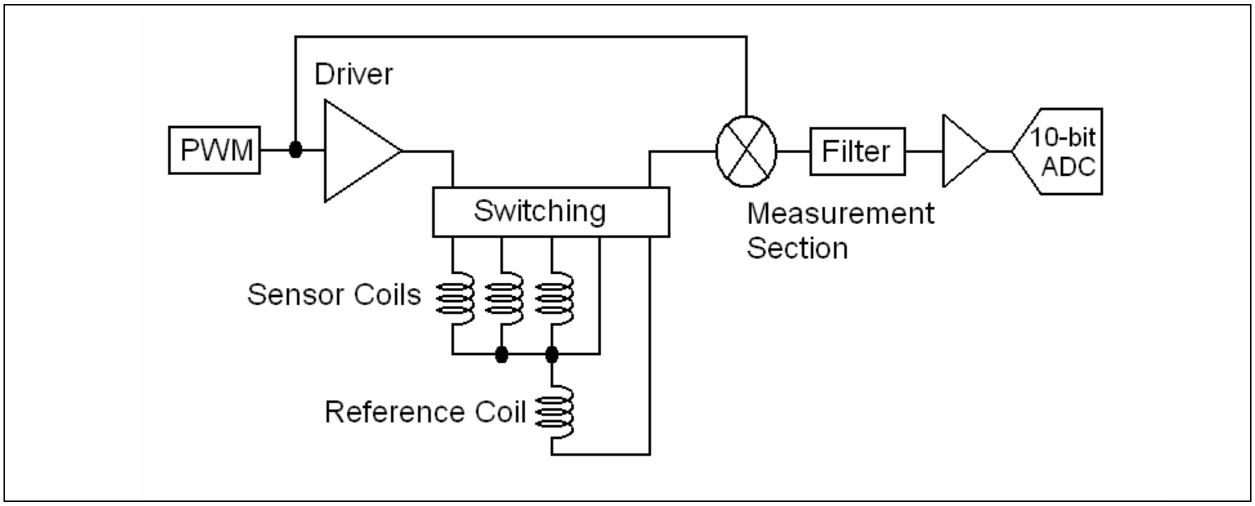
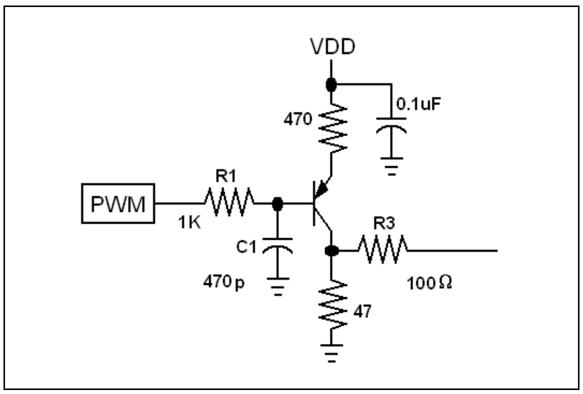


FIGURE 2: TYPICAL PULSED CURRENT DRIVER SCHEMATIC



Equation 3 provides the predicted drive current for specific values of R1, R3 and C1. The predicted drive current value, however, will be reduced by the parallel input resistance of the measurement module. The equivalent impedance at 2 MHz of this circuit is about 700 Ohm. Because the equivalent impedance of the reference and sensor coils is about 120 Ohm, only a part of driver current will pass trough inductors, as shown in Equation 2.

EQUATION 2: REDUCTION IN DRIVE CURRENT DUE TO DETECTOR INPUT IMPEDANCE

$I_{actual} = (700/(120 + 700)) \cdot \Delta I$

$I_{actual} = 0.85 \cdot \Delta I$

Iactual = actual drive current through the inductors

ΔI = Calculated drive current

85% = Typical percentage of drive current passing through the coils

While a more accurate drive current calculation can be made, the variance between sensor coils, the variance between different switching topologies, and the variance in the input impedance of the measurement section is neither consistent nor constant with supply voltage and temperature, making an exact calculation problematic at best. Given the measurement system used by the system is ratio metric in nature; these variances are eliminated from the measurement as part of the measurement process, so an exact calculation is of no value in the final design. This makes the approximation of 85% sufficient for a reliable design.

A typical design at 2 MHZ, with a supply of 3V, will typically use the values shown in Figure 2 to produce a 1.8 mA drive current.

The driver is also shown using a PNP transistor. While this may seem unusual, the polarity of the transistor was chosen to provide the maximum drive head room for the inductors. Because the digital drive into the R1C1 filter has a 50% duty cycle, the base voltage on the transistor will be centered at $V_{DD}/2$. If an NPN transistor were used in the driver, the output of the diver would be approximately ($V_{DD}/2$) – 0.7V. Given a 3V supply voltage, that would put the average output voltage at approximately 0.8V. By using a PNP, the base-emitter diode drop is on the high side of the transistor, putting the emitter voltage at ($V_{DD}/2$) + 0.7, or 2.2V. The saturation voltage for the transistor would then put the collector voltage at 1.5V – 1.7V maximum, giving the circuit approximately 0.7V more drive than the NPN version.

EQUATION 3: OUTPUT DRIVE CURRENT

$$\Delta I = (V_{DD}/R3) \bullet \frac{[e^{(-t/R1C1)} - 1]}{1 + e^{(-t/R1C1)}}$$

ΔI = Change in drive current

V_{DD} = Supply voltage

t = Time

R1 = Input Filter Resistor

C1 = Input Filter Capacitor

> **Note:** The actual drive current delivered to the coils will be reduced due to the input resistance of the measurement circuit. Typically this results in approximately 85% of the current drive passing through the coils. See **"Appendix A: Drive Current Calculations"** for further information.

SWITCHING

There are two methods for switching the driver and measurement circuitry between the different sensor coils: analog multiplexers and GPIO grounding. Refer to Figure 3 and Figure 4.

The analog multiplexer-based system uses paired switches to connect the driver output and the measurement input to each of the coils to be measured. Separate drive and measurement analog switches are used to eliminate the on resistance of the driver multiplexer from creating an offset in the measured coil voltage.

FIGURE 3: ANALOG MULTIPLEXER

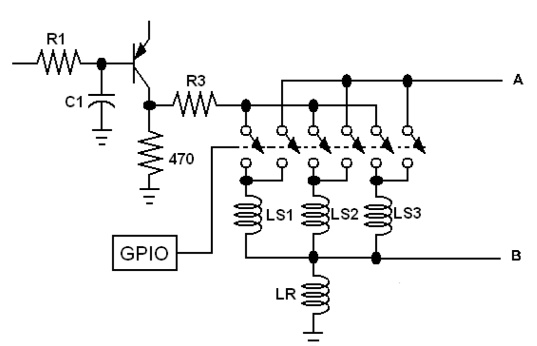
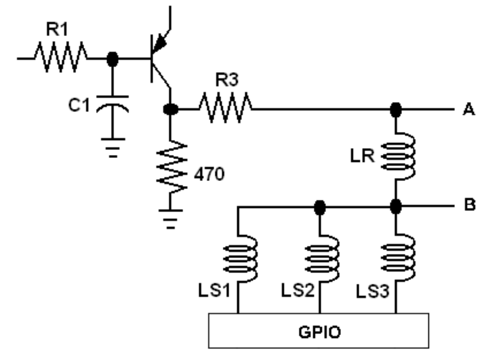


FIGURE 4: GPIO GROUNDING SYSTEM



To reduce cost and complexity, the GPIO grounding system uses IO pins on the microcontroller as switches which ground the coil to be measured. The on resistance of the IO pin will introduce a 100 ohm resistance in series with the coils being measured, and must be taken into account when designing the measurement system to prevent clipping in the output filter.

Fortunately, the output resistance of the IO pins is reasonably stable with temperature/voltage, which allows the normalization and averaging routines in the software to compensate for the offset.

The main configuration difference between the two systems is the placement of the reference coil LR. In the analog multiplexer system, LR is grounded, while the GPIO system has LR between the driver and the sensor coils.

Both systems work well, although the multiplexer-based system is slightly more robust due to the direct connection of the reference inductor to ground. The GPIO grounding system typically produces a smaller deviation due to a press. This is because gain of the output filter must be reduced due to the larger voltage produced across the coils plus the output resistance of the IO pins. Refer to *AN1239, "Inductive Touch Sensor Design"* for more information.

MEASUREMENT CIRCUIT

The Analog-to-Digital Converter (ADC) peripheral, included in most small microcontrollers, does not have the speed required to measure voltages pulsed at a rate of 1-2 MHz. Therefore, the measurement system must convert the pulse DC voltage into a stable DC voltage proportional to the amplitude of the pulsed DC voltage. It must also provide sufficient gain to allow the ADC sufficient resolution of the pulsed DC amplitude to allow the detection of the 5-10% deviation created by the user's press on the sensor target.

While the measurement system, with sufficient gain, can be created using any one of several RF power measurement chips, the cost of the devices can be prohibitive in cost sensitive applications. Therefore, Microchip's inductive touch system uses a simpler synchronous detector, based on an analog multiplexer and an op amp. Refer to Figure .

SENSOR/REFERENCE SELECT

The SPDT analog multiplexer at the left selects from one of two inputs. These inputs are from the multiplexing section and are connected to the reference and sensor coils. The two .01 uF capacitors and 1K resistors level shift the pulsed signals up to the virtual ground used by the detector.

The phase of the incoming signal is then shifted back in line with the PWM signal by the two 330 pF capacitors in series. The phase of the incoming signal is out of phase due to the shift introduced by the output impedance of the driver circuit and the inductance of the coils. Restoring the phase improves the performance of the converter due to the frequency mixing employed in the detector.

FREQUENCY MIXER

The Double Pole, Double Throw (DPDT) multiplexer in the center of the circuit is driven by the same PWM signal that was used to generate the pulse current drive. Together they form a balance mixer, which produces two output frequencies, one at 2x the PWM frequency and the other at 0 Hz. It is the 0 Hz signal that is amplified by the amplifier/filter and passed to the ADC. The 2x frequency is removed by the filter network around the detector amplifier/filter, and the Gain Bandwidth Product (GBWP) limitation of the op amp itself.

VIRTUAL GROUND

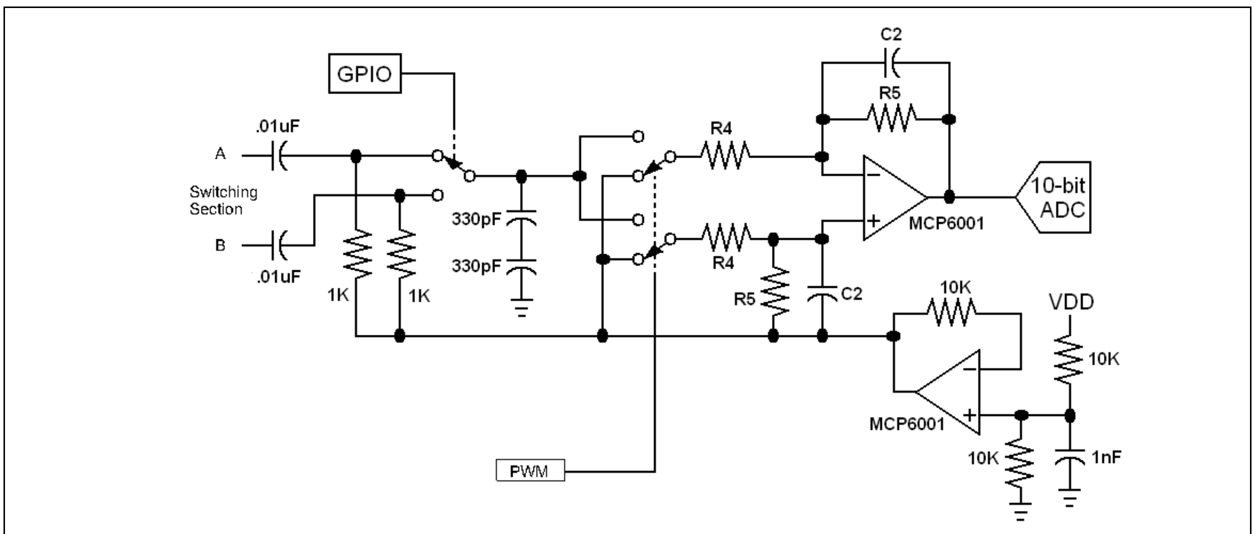
The second op amp, with the 10K resistor divider network on its input, produces a virtual ground for the system at VDD/2. This allows the amplifier/filter in the detector to operate as if it was powered by a split supply.

AMPLIFIER FILTER

The gain of the first op amp is set by R4 and R5. When the inverting input is connected to the coils through the multiplexer, the gain of the amplifier is found using Equation 4, and produces an output that is inverted relative to the virtual ground. When the non-inverting input is connected to the coils, the gain of the amplifier is found using Equation 5 and the output is not inverted, though it is still referenced to the virtual ground.

Because the inverting/non-inverting configuration of the amplifier is switched at the drive frequency, the output is essentially a composite of the positive going outputs of both the inverting and non-inverting configurations. While the gain in the inverting configuration is slightly different from the non-inverting configuration, the low pass nature of the amplifier will average the difference and produce a DC voltage proportional to the input signal, multiplied by the gain.

FIGURE 5: SYNCHRONOUS DETECTOR MEASUREMENT CIRCUIT



EQUATION 4: DETECTOR GAIN IN THE INVERTING CONFIGURATION

$$Gain = -R5/R4$$

EQUATION 5: DETECTOR GAIN IN THE NON-INVERTING CONFIGURATION

$$Gain = R5/R4$$

Capacitor C2 will reduce the gain of the detector at higher frequencies. This is done both to attenuate the 2X frequency component, and also to limit the amount of noise that is passed through the detector to the ADC. The corner frequency of the detector will also determine the response time of the detector, or how fast the detector will settle to a new value in response to a change in the input selection. While it is tempting to set the corner frequency as high as possible, the bandwidth of the detector should also be kept as low as possible to limit the amount of noise passed through the detector into the ADC.

A good rule of thumb is to set the detector corner frequency to 8x to 10x the sensor scan rate of the system. This will provide sufficient settling time for the three measurements that must be made for each sensor. The corner frequency of the detector is determined by Equation 6.

EQUATION 6: CORNER FREQUENCY FOR THE DETECTOR

$$F_{3dB} = 1/(2\pi \bullet R5 \bullet C2)$$

F_{3dB} = Corner frequency of low pass filter

Note:	The polarity of the PWM signal, used to drive the control input of the DPDT multiplexer is important. If the polarity is reversed, the detector filter/amplifier will produce an inverted output signal. This signal would produce an output voltage near $V_{DD}/2$ for low level signals, and a voltage near ground for higher level signals. Care should be taken in the design to ensure that the correct control signal polarity is used, and that the inputs to the multiplexer are also correct.

DETECTOR GAIN

The gain of the filter detector should be chosen to produce an output voltage that provides the maximum resolution in the ADC, while also maintaining a reasonable safety margin for component tolerances and drift. If the GPIO multiplexed configuration is used, then the output resistance of the IO pins should also be considered when determining the detector gain.

From Equation 1, the voltage generated across the coils can be calculated from the drive current. For the example sited in the **"Driver"** section, this is approximately 10 mV per coil. The total voltage across both the reference and sensor coils would then be 20 mV. Given a supply voltage of 3V, this means that a range of 1.5V, relative to the virtual ground, is available. Assuming a 20% margin, then desired gain for the detector should be:

EQUATION 7: DETECTOR GAIN CALCULATION (ANALOG MULTIPLEXER)

$$Gain = (((V_{DD}/2) \bullet 80\%)/(2 \bullet V_{COIL})$$

V_{COIL} = coil voltage

EQUATION 8: DETECTOR GAIN CALCULATION (GPIO)

$$Gain = ((V_{DD}/2) \bullet 80\%)/((2 \bullet V_{COIL}) + (R_{GPIO} \bullet I_{actual}))$$

V_{COIL} = Coil Voltage

R_{GPIO} = Output resistance of GPIO

I_{actual} = Output drive current

Given a voltage gain of 50 to 60, a 10-bit ADC, with a minimum step size of 3 mV, then the system should be able to theoretically resolve a change in impedance on the order of approximately 0.5%, assuming two coils in series. Because the system oversamples the values and averages the results, a resolution of closer to 11-12 bits is actually achieved to give the system the ability to resolve down to .1%-.2%. From *AN1239, "Inductive Touch Sensor Design"*, we know that the typical shift in sensor impedance is typically 3-4%, so the actual number of counts per press is typically between 15 to 40 counts.

DESIGN EXAMPLE FOR AN ANALOG MULTIPLEXER SYSTEM

$V_{DD} = 3V$
10 mv per coil
$F_{3dB} = 1$ kHz

Gain = (1.5V•.8)/(2•10 mV)
Gain = 60

Gain = R5/R4
60 = R5/10K
R5 = 560K (closest value)

$F_{3dB} = 1/(2\pi \bullet R5 \bullet C2)$
1.0 kHz = 1/(6.28•560K•C2)
C2 = 270 pF (closest value)

DESIGN EXAMPLE FOR A GPIO GROUNDED SYSTEM

$V_{DD} = 3V$
10 mV per coil
$F_{3dB} = 1$ kHz
Iactual = 1.8 mA
$R_{GPIO} = 100$ ohms

Gain = R5/R4
66 = R5/10K
R5 = 56K (closest value)

$F_{3dB} = 1/(2\pi \bullet R5 \bullet C2)$
2.0 kHz = 1/(6.28•56K•C2)
C2 = 2700 pF (closest value)

NOISE

One of the first questions that is usually asked about a user interface is how resistant is the system to radiated and conducted noise? In *AN1239, "Inductive Touch Sensor Design"* it is recommended that the target above the sensor coils be grounded. If a ground plane is also put under the coils, then the sensor coils are essentially locked in a shielded box, making the system more resistant to both radiated and conducted noise. The only stipulation is that the power supply should be adequately filtered with proper bypass capacitors, and that good layout practices be used in the design of the PCBs. It should be noted that even with all the shielding on the sensor coils, a small amount of noise will still be observable in the raw data collected on each of the sensor coils.

Note: Any ground plane placed under the sensor coil must be a minimum of 2x-3x further away from the coil, than the spacing between the coil and the target material to prevent loss of sensitivity in the coil.

CONCLUSION

As we have seen in this application note, inductive touch is a relatively simple interface to implement, requiring only a handful of discrete semiconductors and analog blocks. Further, the resulting system is largely resistant to most forms of noise, and, given its requirement for an actuation force to activate a sensor, it constitutes a major step forward in user interface technology.

FIGURE 6: SCHEMATIC OF A COMPLETE GPIO MULTIPLEXED THREE BUTTON INDUCTIVE TOUCH INTERFACE

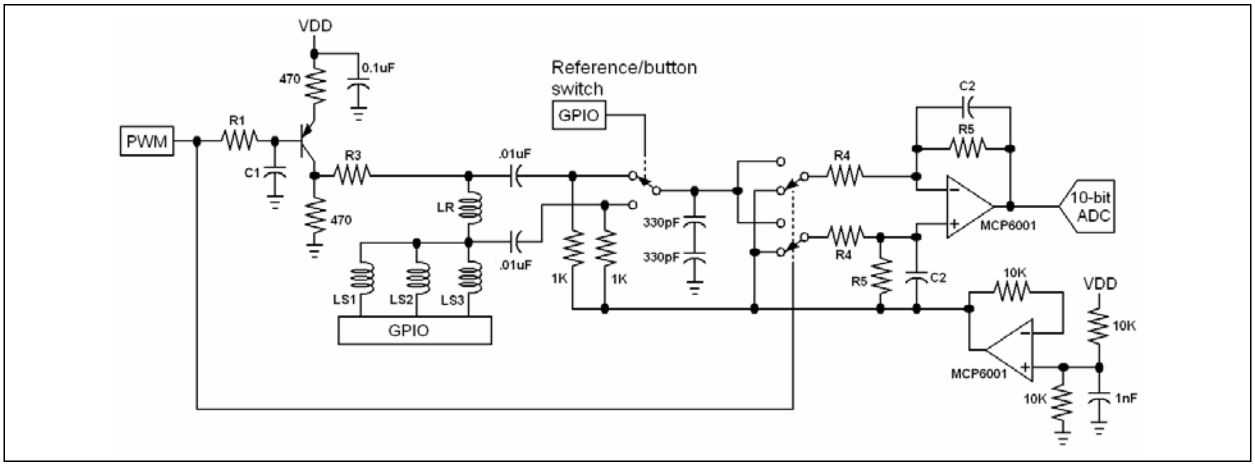
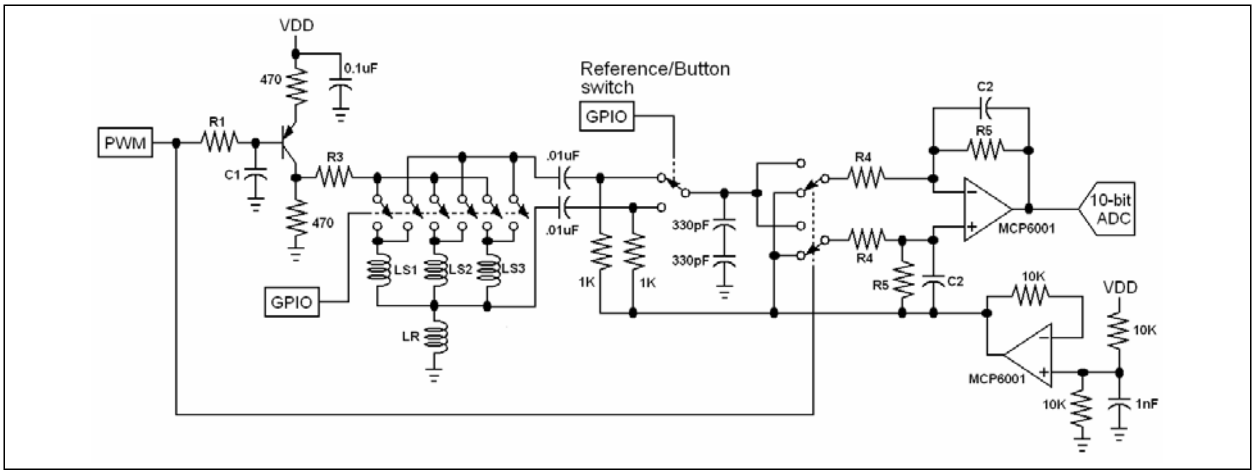


FIGURE 7: SCHEMATIC OF A COMPLETE MULTIPLEXER-BASED THREE BUTTON INDUCTIVE TOUCH INTERFACE



APPENDIX A: DRIVE CURRENT CALCULATIONS

The equation for the drive current is derived by starting with the standard time constant equation for an RC network.

$$V(t) = Vstart \bullet \exp(-t/RC)$$

For the first half of the square wave, capacitor C1 is charged through R1, for the second half, it is discharged through R1, so we can consider;

$$Vstart = V_{DD}/2 - \Delta V$$
$$Vstop = V_{DD}/2 + \Delta V$$

Substituting in the equation for an RC network:

$$V_{DD}/2 = \Delta V = (V_{DD}/2 - \Delta V) \bullet \exp(-t/RC)\Delta V(1 + \exp(-t/RC)) = (V_{DD}/2) \bullet [\exp(-t/RC) - 1]$$

$$\Delta V = (V_{DD}/2) \bullet \frac{[e(-t/RC) - 1]}{1 + e(-t/RC)}$$

The peak to peak amplitude of the resulting triangular waveform, at coil driver input will then be:

$$V_{PK-PK} = 2\Delta V$$

The peak to peak current trough coil driver will therefore be:

$$I_{PK-PK} = V_{PK-PK}/R3$$

The resulting equation is then:

$$\Delta I = V_{DD}/R3 \bullet \frac{\lfloor e^{(-t/R1C1)}(-1) \rfloor}{1 + e^{(-t/R1C1)}}$$

The actual drive current will be reduced due to the input resistance of the detector circuit. Typically this is 2K on both the coil and reference inputs. The equivalent impedance at 2 MHz of this circuit is approximately 700 Ohm. If the equivalent impedance of reference and active coil is approximately 120 Ohms, the portion of the current that actually passes through the inductors is given by:

$$I_{ACTIVE} \sim (700/(120+700)) \bullet I_{PK-PK} \rightarrow .085\ I_{PK-PK}$$

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