8 Front-End Electronics for Embedded Systems

This chapter is primarily dedicated to front-end analog electronics interfacing microcontrollers. It brings together most of the topics covered in the previous seven chapters. The chapter starts with basic electromechanical device control and sensor circuits, then moves on to implementing full scale electronic platforms that contain sensor(s), amplifier(s), logic blocks, ADC and DAC.

8.1 Electromechanical Device Control

Relay switches are excellent choices to isolate digital circuits from noisy electrical environments or from circuits that require high voltages or currents to operate electromechanical devices. Figure 8.1 shows the operation of a relay switch. The parallel diode placed across the relay prevents transient currents to form and protects rest of the circuit from current surges emanating from the relay. When a high logic level is applied to the V_{IN} terminal, the NPN transistor saturates. The resulting current transforms the inductor into an electromagnet, and closes the mechanical switch inside the relay enclosure. This action starts the current conduction in the secondary circuit to operate an electromechanical device. When there is no current present, the switch pulls back to its no contact position and stops the current in the secondary circuit.

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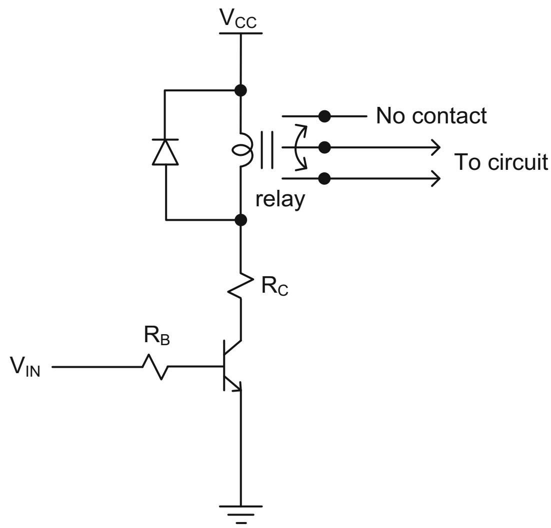


Fig. 8.1 Bipolar circuit controlling a relay

Figure 8.2 uses the relay to activate a DC motor. When closed, the 12 V supply runs the DC motor. In this circuit V_{CC} can safely be set to a voltage lot less than 12 V. Similarly, the supply voltage for the DC motor can be set much higher than 12 V without causing noise or harming the digital circuit.

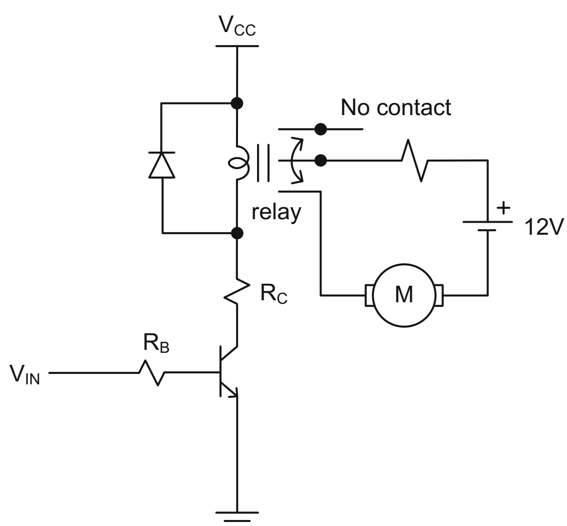


Fig. 8.2 A dc motor activation circuit using a relay for electrical isolation

Opto-isolator is another device that uses optical means to isolate a digital circuit from an analog circuit that includes an electromechanical device and often requires high voltages or currents. The box defined by the dashed lines in Fig. 8.3 represents an opto-isolator. This device is composed of an infrared LED and a photo-transistor whose base is optically coupled with the light emanating from the LED. When the LED emits infrared light, it produces high voltage at the base of the photo-transistor to saturate it. Since opto-isolator only comes with an open-collector configuration, the collector of the photo-transistor has to be externally connected to a supply voltage with a current limiting resistor to function.

Figure 8.3 shows an application of the opto-isolator to operate a piezo-electric buzzer. When V_{IN} terminal receives logic 1, the output of the inverter goes low. The supply current that flows through RD turns on the "internal" LED. Infrared light that emanates from the LED saturates the photo-transistor provided that the current through R_O meets the manufacturer's requirements for the transistor to saturate. As a result, the collector of the photo-transistor pulls low and turns off the output NPN transistor. Since there will be no current through the piezo-electric buzzer, it remains silent.

However, when V_{IN} is lowered to 0 V, the inverter output goes to logic 1, and this produces 0 V across the infrared LED. The LED does not emit light, and the photo-transistor stops operating. Consequently, the base terminal of the photo-transistor goes to 12 V, and saturates the output NPN transistor. The buzzer starts operating.

It is essential to separate the ground terminals of the 5 V digital supply from the 12 V analog supply to prevent transmitting any electrical noise through ground connections that may adversely affect the operation of the digital circuit.

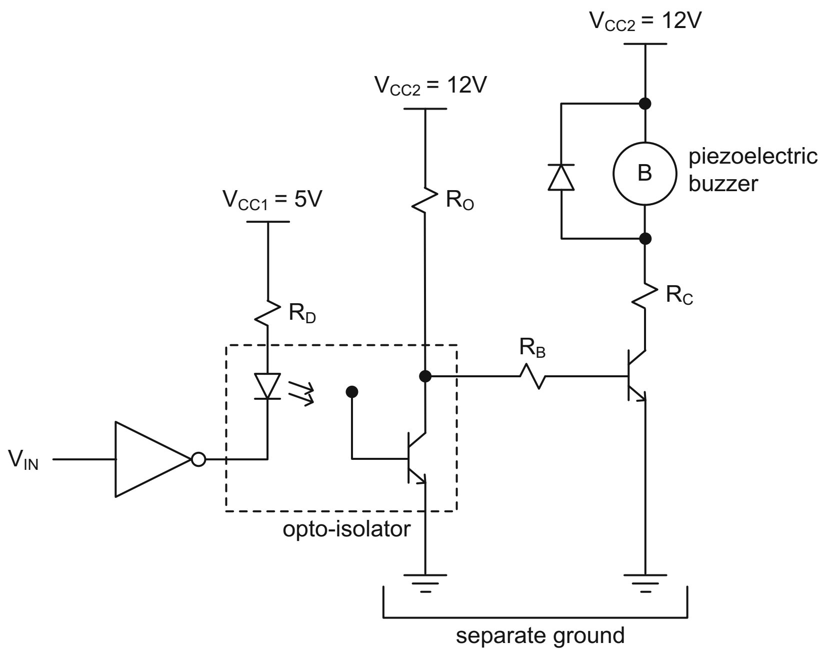


Fig. 8.3 Piezoelectric buzzer control using an opto-isolator

Figure 8.4 shows another application of the opto-isolator device if the DC motor requires even higher voltage than 12 V to operate. As discussed above with the piezo-electric buzzer, a low logic applied to V_{IN} disables the opto-isolator and operates the DC motor. Again, it is a vital importance to separate the ground connections of the 5 and 24 V supplies from each other to prevent electrical noise produced in the analog circuitry from propagating through the ground terminal.

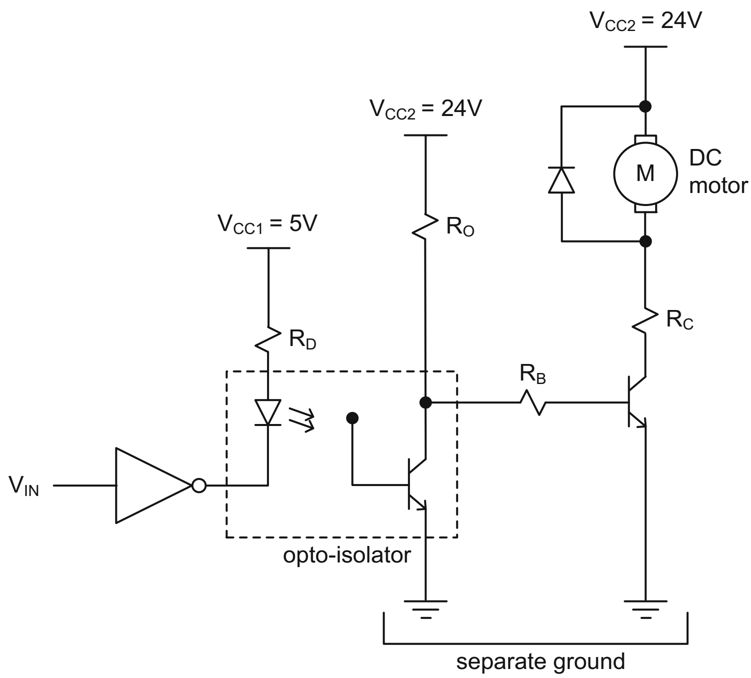


Fig. 8.4 DC motor control using an opto-isolator

8.2 Pulse Width Modulation Circuits

A simple Pulse Width Modulation (PWM) circuit is shown in Fig. 8.5. This circuit is an enhanced form of the square wave generator discussed earlier in Chapter 7.

In this circuit, there are two feedback loops. The primary feedback loop connects the output to the positive input of the operation amplifier with R_F, setting up the reference signal, V_{REF}, for the input. The secondary feedback loop connects the output to the negative input of the operational amplifier with the combination of a variable resistor, a diode and a capacitor. This RC circuit basically adjusts the pulse width at the output.

Assume $V_{SIG} = -V_{CC}$ and $V_{OUT} = +V_{CC}$ initially. $V_{OUT} = +V_{CC}$ produces:

$$V_{REF} = V_{CC}\frac{R_F}{R_F + R_O} \tag{8.1}$$

The output node starts charging the capacitor through the resistor, αR (α represents the portion of the potentiometer), and the diode, D_1, as shown in the top portion of Fig. 8.6. This produces:

$$V_{SIG} = -V_{CC} + 2V_{CC}\left[1 - \mathbf{exp}\left(-\frac{t}{\alpha RC}\right)\right] \tag{8.2}$$

As soon as V_{SIG} reaches slightly above $V_{REF} = V_{CC}\frac{R_F}{R_F + R_O}$, V_{OUT} changes its polarity, and becomes equal to $-V_{CC}$. As a result, V_{REF} changes its value and becomes:

$$V_{REF} = -V_{CC}\frac{R_F}{R_F + R_O} \tag{8.3}$$

With $V_{OUT} = -V_{CC}$, this time the capacitor starts discharging through the diode, D_2, and the unused portion of R, $(1 - \alpha)R$, from its initial value of $\frac{R_F}{R_F + R_O}V_{CC}$ towards $-V_{CC}$ as shown in the bottom portion of Fig. 8.6. Thus,

$$V_{SIG} = V_{CC}\left(1 + \frac{R_F}{R_F + R_O}\right)\mathbf{exp}\left(-\frac{t}{(1-\alpha)RC}\right) - V_{CC} \tag{8.4}$$

When V_{SIG} reaches $-\frac{R_F}{R_F + R_O}V_{CC}$, however, the output changes its polarity to $+V_{CC}$ and forms a positive reference voltage, $V_{REF} = V_{CC}\frac{R_F}{R_F + R_O}$. As a result, the capacitor starts charging through αR and D_1 again, and V_{SIG} begins to increase from $-\frac{R_F}{R_F + R_O}V_{CC}$ to $\frac{R_F}{R_F + R_O}V_{CC}$.

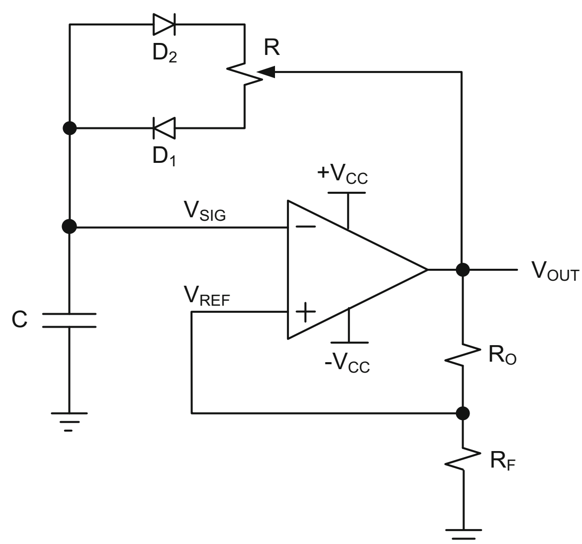


Fig. 8.5 A simple pulse width modulation (PWM) circuit

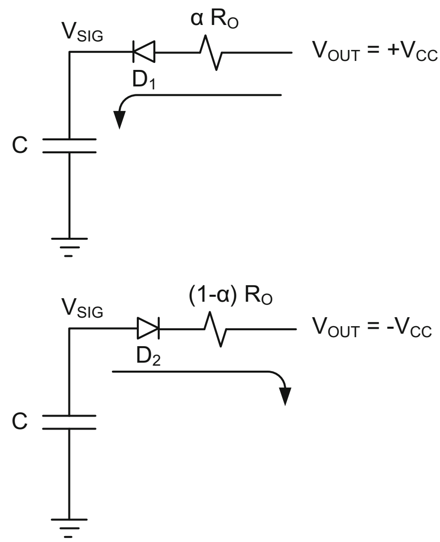


Fig. 8.6 Charge (*top*) and discharge (*bottom*) paths of the PWM circuit in Fig. 8.6

The waveforms of V_{SIG} and V_{OUT} are very similar to the ones produced by the square wave generator in Chapter 7, and shown here again in Fig. 8.7. In this figure, the time constants during the rise and the fall of V_{SIG} are different from each other due to α. This property enables fine-tuning the pulse width, and therefore the duty cycle of the output waveform.

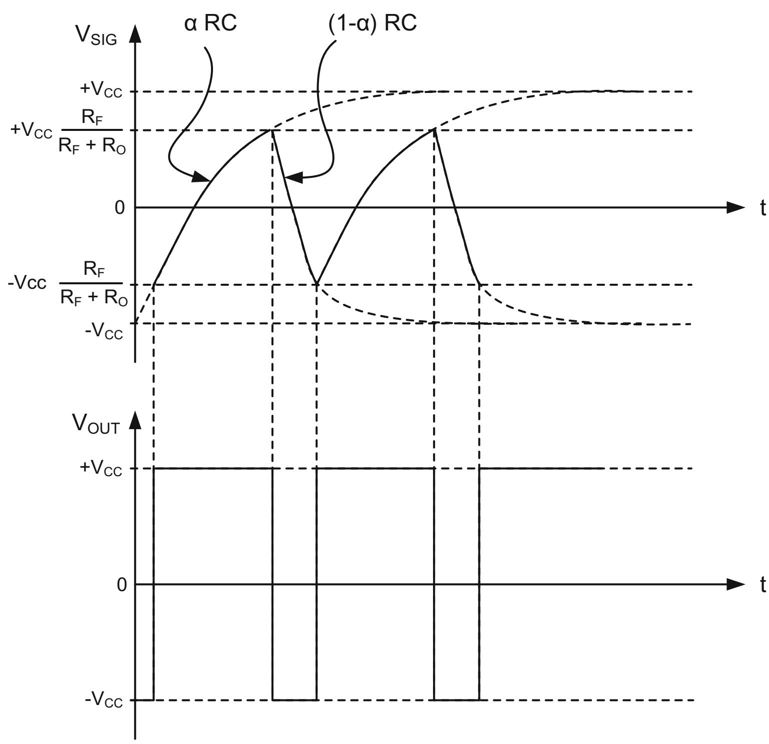


Fig. 8.7 The waveforms at V_{SIG} and V_{OUT} of the PWM circuit in Fig. 8.6

8.3 DC Motor Control

Operating a DC motor in different speeds can be achieved by the combination of a PWM circuit cascaded with an opto-isolator circuit as shown in Fig. 8.8. The opto-isolator separates the ground terminals of the noise-receptive digital circuit from the circuit that operates the motor which usually requires high voltages.

The PWM can generate voltages between 0 V and $+V_{CC}$ to turn on and off the optical transistor in high speeds not too suitable for a relay.

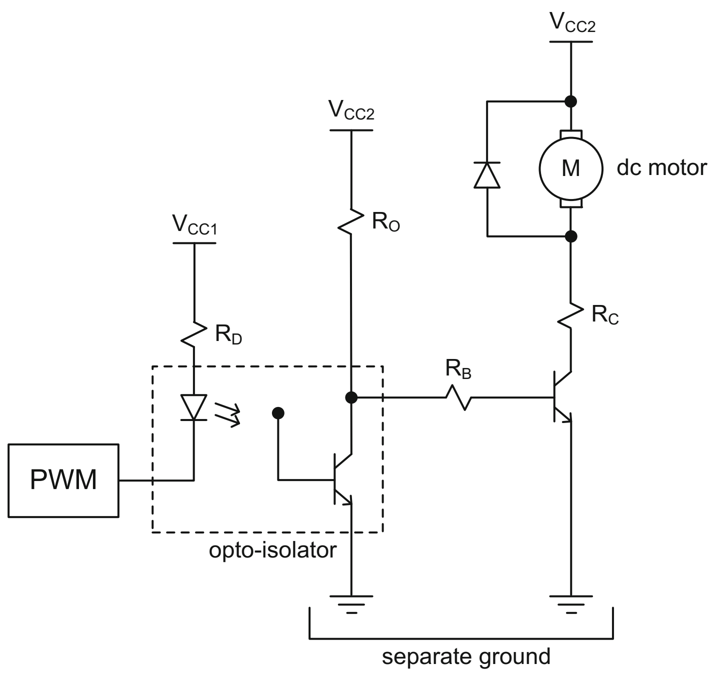


Fig. 8.8 DC motor control

The pulse duration of the PWM output at $+V_{CC}$ lasts only micro seconds, during which the output bipolar transistor supplies current to the DC motor. The only way to observe this is through measuring the motor's rpm. Motor's speed can be increased by increasing the time constant of the secondary feedback circuit or αRC. The motor can practically be turned off if the value of the variable resistance is reduced to 0 Ω.

8.4 Servo Control

It is possible to design a pulse generator to control a servo motor. The signal input to an analog servo calls for a pulse with a period of 20 ms as shown in Fig. 8.9.

If the pulse width is maintained at T = 1.5 ms, the servo arm stays at neutral position. Reducing T to 1.25 ms turns the servo arm to −90° from the neutral axis. Similarly increasing T to 1.75 ms rotates the servo arm to +90°. However, some servo motor manufacturers specify T = 1 ms and T = 2 ms pulse widths for −90 and +90° arm rotations in their datasheets, respectively, and this requires calibrating servo arm movement for each servo manufacturer.

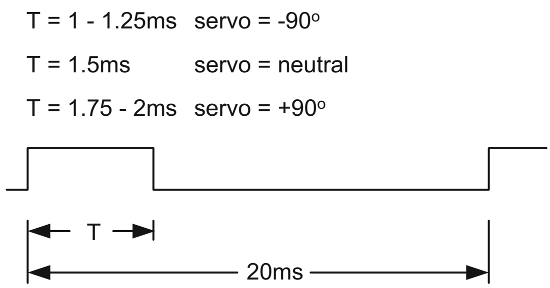


Fig. 8.9 Servo control waveforms

8.5 Hall-Effect Sensor Control

The Hall-effect device studied in Chapter 5 operates in the presence of a magnetic field, and forms a very useful device for systems that require controlling mechanical rotation.

The magnetic properties of Hall-effect devices have enabled manufacturers to devise a simple device resistant to heat, cold, pressure and many other external effects. The basic usage of this device is car industry. However, many other industries that employ magnetic field in the operation of their systems also use this device. Figure 8.10 shows a unipolar Hall-effect device manufactured by Allegro Microsystems.

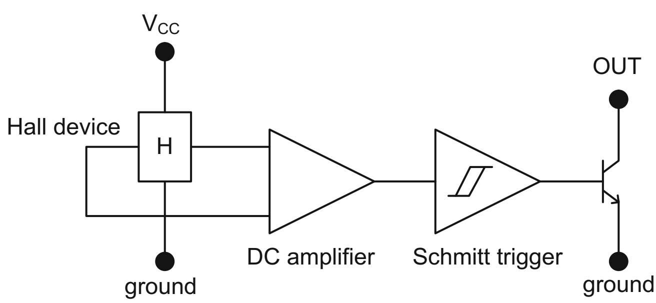


Fig. 8.10 Unipolar Hall-effect device manufactured by Allegro Microsystems

The package contains a Hall-effect device that produces milli-volt range Hall voltage when exposed to a magnetic field, a DC amplifier that amplifies the Hall voltage to full logic levels at its output, and finally an NPN bipolar transistor with an open collector.

When this device is connected to activate an LED as shown in Fig. 8.11, the exposure to magnetic field enables the Schmitt trigger to produce a high logic output to saturate the NPN transistor. The resulting current, I_L, turns on the LED and becomes:

$$I_L = \frac{(V_{CC} - V_{LED} - V_{CESAT})}{R_L} \tag{8.5}$$

Here, V_{LED} is the voltage drop across the LED. I_L in this equation must be adjusted to be able to saturate the bipolar transistor and turn on the LED.

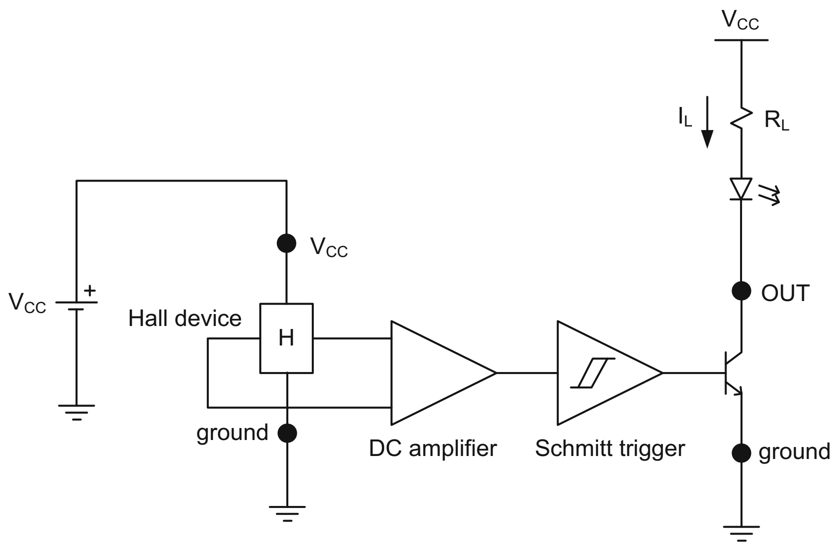


Fig. 8.11 Usage of the unipolar Hall-effect device to activate LED

The remainder of this chapter is dedicated to numerous design projects that prepare the sensor signal to interface with the microcontroller. Every project uses a different type of sensor and follows a unique data-path to condition the analog signal prior to analog-to-digital conversion. The last two projects integrate digital logic blocks with analog circuitry, therefore they require logic design skills outlined in Chapter 9.

8.6 Design Project 1: Designing Front-End Electronics for an Analog Microphone

The first design project simply prepares the analog signal from a microphone for the ADC. The microphone used in Fig. 8.12 is analog microphone ADMP504 produced by Analog Devices. From the manufacturer's datasheet, the microphone produces a maximum voltage of 0.25 V, which needs to be amplified to a maximum value of 5 V for the analog-to-digital converter, AD7819, also produced by Analog Devices. The microphone produces an output DC offset voltage of 0.8 V that needs to be cleaned from the analog signal before the signal is amplified. Therefore, V_{REF} in Fig. 8.12 should be adjusted to a certain value that offsets the 0.8 V DC offset in the microphone output signal.

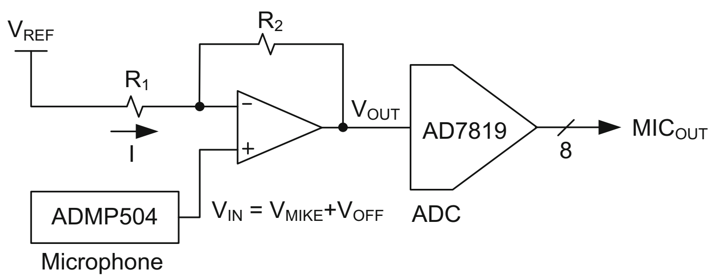


Fig. 8.12 Microphone front end circuit

According to the datasheet we have:

$$V_{IN} = V_{MIKE} + V_{OFF} \tag{8.6}$$

where, V_{MIKE} is the analog voltage produced by the microphone without DC offset, and $V_{OFF} = 0.8$ V. We also have:

$$V_{REF} = (R_1 + R_2)I + V_{OUT} \tag{8.7}$$

$$V_{REF} = R_1 I + V_{IN} \tag{8.8}$$

From Eq. 8.8,

$$I = \frac{V_{REF} - V_{IN}}{R_1} \tag{8.9}$$

Substituting Eq. 8.9 into Eq. 8.7 yields:

$$V_{REF} = (R_1 + R_2)\frac{V_{REF} - V_{IN}}{R_1} + V_{OUT} \tag{8.10}$$

Rearranging the terms in Eq. 8.10 yields:

$$V_{OUT} = \frac{(R_1 + R_2)}{R_1}V_{IN} - \frac{R_2}{R_1}V_{REF} \tag{8.11}$$

Substituting Eq. 8.6 into Eq. 8.11 yields:

$$
\begin{aligned}
V_{OUT} &= \frac{(R_1 + R_2)}{R_1}(V_{MIKE} + V_{OFF}) - \frac{R_2}{R_1}V_{REF} \\
&= \frac{(R_1 + R_2)}{R_1}V_{MIKE} + \left[\frac{(R_1 + R_2)}{R_1}V_{OFF} - \frac{R_2}{R_1}V_{REF}\right]
\end{aligned}
\tag{8.12}
$$

In order to isolate the term with V_{MIKE}, we need to eliminate the terms with V_{OFF} and V_{REF}. Thus,

$$\frac{(R_1 + R_2)}{R_1}V_{OFF} = \frac{R_2}{R_1}V_{REF}$$

or

$$V_{REF} = \frac{(R_1 + R_2)}{R_2}V_{OFF} \tag{8.13}$$

Then,

$$V_{OUT} = \frac{(R_1 + R_2)}{R_1}V_{MIKE} \tag{8.14}$$

From the datasheets, the maximum value produced by the microphone is $V_{MIKE} = 0.25$ V and the maximum value at the input of the ADC is $V_{OUT} = 5$ V. Therefore, from Eq. 8.14:

$$\frac{V_{OUT}}{V_{MIKE}} = \frac{5}{0.25} = 20 = \frac{(R_1 + R_2)}{R_1}$$

$$R_2 = 19\,R_1 \tag{8.15}$$

We use $R_1 = 1$ KΩ and $R_2 = 19$ KΩ

Using Eqs. 8.15 in 8.13 yields $V_{REF} = 1.05 \times 0.8 = 0.85$ V.

Therefore, we must find a Zener diode at 0.85 V or a potentiometer that can produce 0.85 V. However, the latter choice is prone to temperature fluctuations.

8.7 Design Project 2: Designing Front-End Electronics for a Temperature Measurement System

The second design project amplifies the signal from a type-K thermocouple, and prepares it for the ADC. The thermocouple signal changes linearly with temperature as shown in Fig. 8.13. From this figure, the thermocouple signal can be approximated as $V_T \approx 0.0407$ T in mV. However, at low temperatures the deviation from linearity introduces error as large as 2 mV.

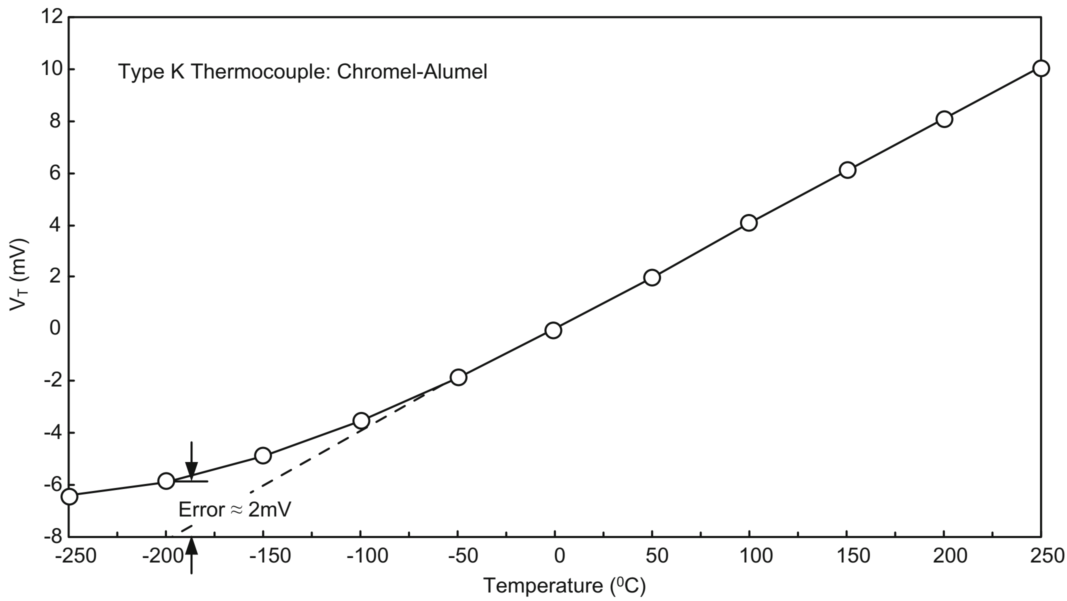


Fig. 8.13 Voltage-temperature characteristics of type-K thermocouple

In the first attempt, let us build a circuit that measures ambient temperatures between 0 and 200 °C. This is a very straightforward design that requires only signal amplification as shown in Fig. 8.14.

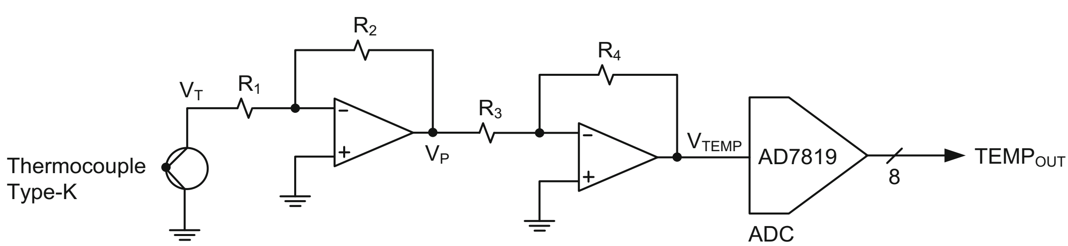


Fig. 8.14 A circuit that measures temperatures between 0 and 200 °C

The amplification is performed by two cascaded inverting stages. In the first stage, we have:

$$\frac{V_P}{V_T} = -\frac{R_2}{R_1} \tag{8.16}$$

In the second stage,

$$\frac{V_{TEMP}}{V_P} = -\frac{R_4}{R_3} \tag{8.17}$$

We know the maximum voltages at V_T = 8.14 mV (at T = 200 °C) and V_{TEMP} = 5 V. This requires an overall gain of approximately 614. If the first stage amplifies the thermocouple signal by 20 times, then the second stage will need an amplification of 30.7. Thus,

$$R_2 = 20\,R_1 \tag{8.18}$$

Similarly,

$$R_4 = 30.7\,R_3 \tag{8.19}$$

Selecting R_1 = 1 KΩ, R_2 = 20 KΩ, R_3 = 2 KΩ, and $R_4 \approx$ 62 KΩ according to Eqs. 8.18 and 8.19 will proportionally change the sensor readings at V_{TEMP} between 0 and 5 V.

The second circuit in Fig. 8.15 conditions the temperature readings at V_T between −200 and 200 °C.

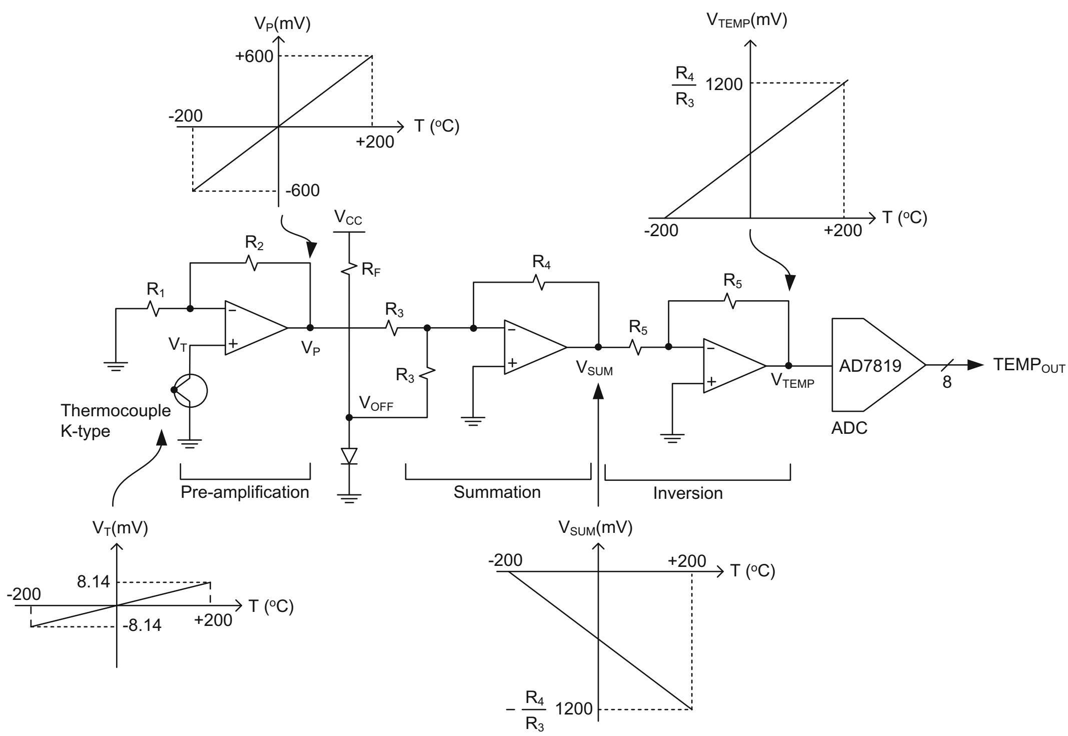


Fig. 8.15 A circuit that measures temperatures between −200 and 200 °C

Since the thermocouple voltage readings become negative for temperatures less than 0 °C, the best method is to add a DC offset to the actual thermocouple signal to eliminate the negative values of V_T. The maximum value of the "extrapolated" V_T at −200 °C is −8.14 mV according to Fig. 8.13. This is a very small value to offset. Therefore, a better option is to amplify the thermocouple signal without any DC offset in the first stage, and then employ an offset voltage for the amplified signal in the next stage. The amplification in the first stage produces the result in Eq. 8.20.

$$\frac{V_P}{V_T} = \frac{R_1 + R_2}{R_1} \tag{8.20}$$

In the second stage, a steady DC offset voltage can be generated by a rectifying diode. Assuming the diode has $V_F = 600$ mV, the amplification at the first stage must yield:

$$\frac{V_P}{V_T} = \frac{R_1 + R_2}{R_1} = \frac{600\,\text{mV}}{8.14\,\text{mV}} \approx 73.7 \tag{8.21}$$

In this equation, if $R_1 = 1$ KΩ then $R_2 \approx 72$ KΩ.

This amplification factor makes the value of V_P to be approximately −600 mV at −200 °C, and +600 mV at +200 °C.

For the second stage, when $V_{OFF} = 600$ mV, then:

$$V_{SUM} = -\frac{R_4}{R_3}(V_P + V_{OFF}) = -\frac{R_4}{R_3}(V_P + 600\,\text{mV}) \tag{8.22}$$

According to Eq. 8.22, when temperature drops to −200 °C and V_P becomes −600 mV, then:

$$V_{SUM} = -\frac{R_4}{R_3}(-600\text{ mV} + 600\text{ mV}) = 0\text{ V} \tag{8.23}$$

Similarly, when temperature rises to 200 °C and V_P becomes +600 mV, then:

$$V_{SUM} = -\frac{R_4}{R_3}(600\,\text{mV} + 600\,\text{mV}) = -\frac{R_4}{R_3}1200\,\text{mV} \tag{8.24}$$

A Zener diode can also be used to create a DC offset in the second stage. The smallest reference voltage from a Zener diode is approximately 1.8 V, and using this value in Eq. 8.21 requires a gain over 200 in the first stage. However, producing a high voltage gain from a single stage amplifier should be avoided because small deviations in resistor values introduce large errors in amplifier gain.

To calculate the biasing resistor value, R_F, in the second stage requires determining the forward current, I_F, of the rectifying diode. Assuming that $V_{CC} = 5$ V, and this particular diode calls for $I_F = 1$ mA in order to produce $V_F = 600$ mV across its terminals, R_F becomes approximately equal to 4.4 KΩ to set up the correct DC offset, V_{OFF}.

The third stage simply provides a voltage inversion. Thus:

$$V_{TEMP} = -\frac{R_5}{R_5}V_{SUM} = -V_{SUM} = \frac{R_4}{R_3}(V_P + V_{OFF}) \tag{8.25}$$

The maximum value of V_{TEMP} should be around 5 V before the analog-to-digital converter stage. Thus:

$$V_{TEMP} = 5000\,\mathrm{mV} = \frac{R_4}{R_3}(V_P + V_{OFF}) = \frac{R_4}{R_3}1200\,\mathrm{mV} \tag{8.26}$$

$$\frac{R_4}{R_3} = \frac{5000\,\mathrm{mV}}{1200\,\mathrm{mV}} \approx 4.2 \tag{8.27}$$

According to Eq. 8.27, we can use $R_3 = 1.1\ \mathrm{K\Omega}$ and $R_4 = 4.7\ \mathrm{K\Omega}$. In Eq. 8.25, R_5 can be taken as 10 KΩ.

All three cascaded stages produce the following output voltage at V_{TEMP}:

$$V_{TEMP} = \frac{R_4}{R_3}(V_P + V_{OFF}) = \frac{R_4}{R_3}\left(\frac{R_1 + R_2}{R_1}V_T + V_{OFF}\right) \tag{8.28}$$

In this equation, V_{TEMP} becomes 0 V at T = −200 °C (and $V_T = -8.14$ mV). Similarly, at T = 200 °C (and $V_T = 8.14$ mV) V_{TEMP} becomes approximately 5 V.

8.8 Project 3: Designing Front-End Electronics for a Light Level Measurement System

Another interesting project with operational amplifiers is to measure the intensity of a light source with a photodiode. Once reverse biased, photodiodes produce high levels of reverse saturation current when exposed to light. An OSRAM SFH-213 is a popular photodiode with a spectral range of 400–1100 nm, peaking at 900 nm. It has a relatively small dark (thermal) current, I_{DARK}, of 1 nA and area of 1 mm^2. Its light sensitivity of 550 mA/W makes this diode generate 5.5 µA of photo current (short circuit), I_{PH}, when exposed to 10 W/m^2 intensity light source within its spectral range.

The reverse breakdown voltage of SFH-213 is 50 V per manufacturer's datasheet. Therefore, reverse biasing this diode with $-V_{CC} = -15$ V as shown in Fig. 8.16 is safe although this voltage level may also increase the dark current characteristics beyond what is specified in the datasheet. The pre-amplification stage in Fig. 8.16 produces a gain proportional to the resistor R_1 as shown in Eq. 8.29.

$$V_{PH} = R_1(I_{PH} + I_{DARK}) \tag{8.29}$$

The resistor, R_P, limits the current through the photodiode and also provides a reverse biased voltage less than V_{CC} across its terminals in case the power supply voltage is greater than the breakdown voltage.

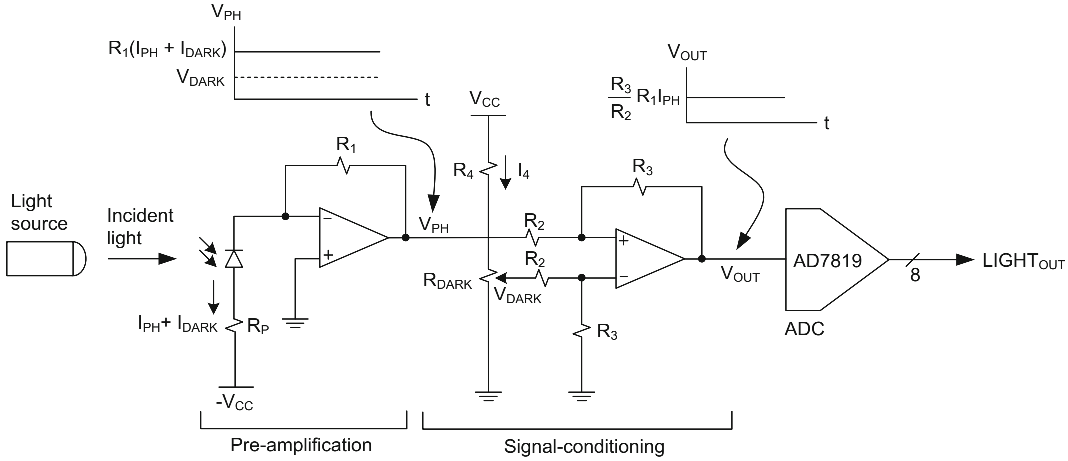


Fig. 8.16 A circuit that measures light level ($-V_{CC}$ is used to bias the photodiode)

The second stage in Fig. 8.16 is primarily used for signal conditioning such as elimination of the dark current component of the photodiode and possibly noise cancellation with some voltage amplification. This stage produces the following voltage gain:

$$V_{OUT} = \frac{R_3}{R_2}(V_{PH} - V_{DARK}) \tag{8.30}$$

Substituting Eq. 8.29 into Eq. 8.30 yields:

$$V_{OUT} = \frac{R_3}{R_2}[R_1(I_{PH} + I_{DARK}) - V_{DARK}] \tag{8.31}$$

In Eq. 8.31, if we tune the potentiometer such that:

$$V_{DARK} = R_1 I_{DARK} \tag{8.32}$$

Then Eq. 8.31 can be reformed as:

$$V_{OUT} = \frac{R_3}{R_2} R_1 I_{PH} \tag{8.33}$$

Using $I_{PH} = 5.5\ \mu A$ and the maximum value of $V_{OUT} = 5\ V$, Eq. 8.33 can be rewritten as:

$$\frac{R_3}{R_2} R_1 = \frac{5V}{5.5 \times 10^{-6} A} \approx 909\ K\Omega \tag{8.34}$$

According to Eq. 8.34, selecting $R_2 = 1\ K\Omega$, $R_1 = 27\ K\Omega$ produces $R_3 = 33\ K\Omega$. From Eq. 8.32, the dark voltage becomes:

$$V_{DARK} = 27 \times 10^3 \times 1 \times 10^{-9} = 27\ \mu V \tag{8.35}$$

Also from Eq. 8.29, V_{PH} becomes approximately equal to 0.15 V.

If we allow $I_4 = 1$ mA at the input of the second stage, R_4 can be approximately equal to:

$$R_4 = \frac{15\text{V}}{1 \times 10^{-3}\text{A}} \approx 15\ \text{K}\Omega \tag{8.36}$$

The potentiometer, R_{DARK}, should be a small value compared to R_4 such as 100 Ω. This way, values up to 100 μV can be produced at the V_{DARK} input.

The second schematic in Fig. 8.17 is used if the reverse bias voltage for the photodiode is $+V_{CC} = 15$ V for the same project. This configuration requires both amplification stages in Fig. 8.16 to be inverting type.

Both circuits use Analog Devices eight-bit ADC, AD7819.

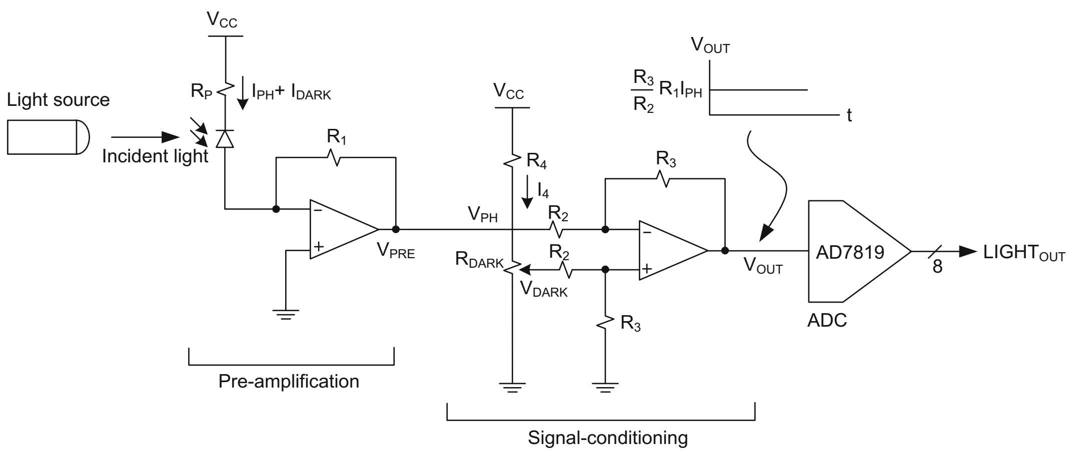


Fig. 8.17 A circuit that measures light level ($+V_{CC}$ is used to bias the photodiode)

8.9 Project 4: Designing Photo Detector Circuits

The circuits in Figs. 8.18 and 8.19 are simple on/off circuits sensitive to light, and do not require any ADC to interface with the microcontroller. The sole purpose of these circuits is to create a hardware interrupt for the microcontroller or engage an electro-mechanical device based on light detection.

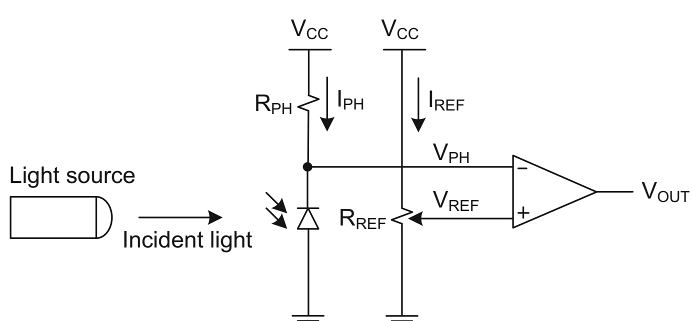


Fig. 8.18 Photo detector circuit with an operational amplifier

When the light is exposed to the photodiode in Fig. 8.18, the voltage level at V_{PH} drops with respect to V_{REF} because the photo current through the diode, I_{PH}, becomes much larger than the thermally-generated reverse saturation current, I_{DARK}. Therefore,

$$V_{PH} = V_{CC} - R_{PH} I_{PH} \tag{8.37}$$

Suppose $V_{CC} = 5$ V. If we use the same photodiode, SFH-213, from the earlier design with $I_{PH} = 5.5\ \mu A$, and assume $V_{PH} = 3$ V (although values less that 3 V or closer to 5 V are also acceptable design entries for V_{PH}) when the light is on, then R_{PH} becomes:

$$R_{PH} = \frac{(V_{CC} - V_{PH})}{I_{PH}} = \frac{5 - 3}{5.5 \times 10^{-6}} \approx 363\,\text{K}\Omega \tag{8.38}$$

When the light is off, $I_{PH} \approx 0$ A and $V_{PH} = V_{CC} = 5$ V.

This means that V_{REF} should be somewhere between $V_{PH} = 3$ and 5 V. Suppose $V_{REF} = 4$ V and $I_{REF} = 1$ mA. Then the maximum potentiometer value, R_{REF}, becomes:

$$R_{REF} = \frac{V_{CC}}{I_{REF}} = \frac{5}{1 \times 10^{-3}} = 5\,\text{K}\Omega \tag{8.39}$$

If we adjust the tap on the potentiometer, R_{REF}, such that V_{REF} becomes 4 V, then $\Delta v = V_{PH} - V_{REF}$ becomes $\Delta v = 3 - 4 = -1$ V when the light turns on and $V_{OUT} = 5$ V. When the light turns off, V_{REF} is still at 4 V, but V_{PH} becomes 5 V. This produces $\Delta v = 5 - 4 = +1$ V, and V_{OUT} becomes 0 V.

The circuit in Fig. 8.19 is an extension of the photo detector circuit in Fig. 8.18, and is used to operate a small DC motor. In this figure, the DC motor spins as long as the light is on, but it turns off when the light is switched off. A Zener diode is used to set a reference voltage for the difference amplifier.

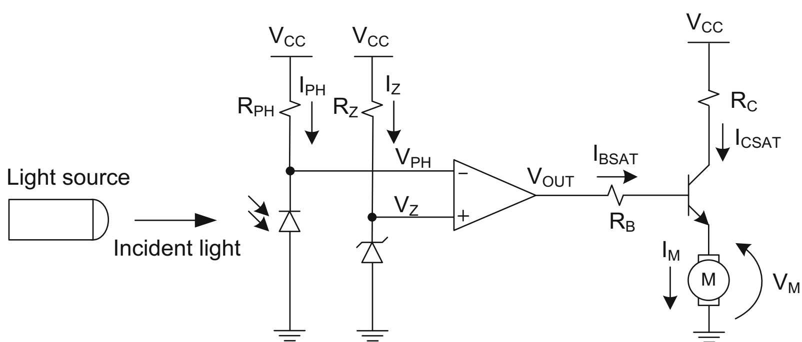


Fig. 8.19 Photo detector circuit operating a small DC motor

Assume that the power supply voltages are $+V_{CC} = 15$ V and $-V_{CC} = 0$ V. $I_{PH} = 5\ \mu A$ when the light turns on, and the DC motor operates at a voltage of $V_M = 5$ V at $I_M = 100$ mA.

When the light is on, assume $V_{PH} = 13$ V. This produces:

$$R_{PH} = \frac{(V_{CC} - V_{PH})}{I_{PH}} = \frac{15 - 13}{5 \times 10^{-6}} = 400\,\mathrm{K\Omega} \tag{8.40}$$

When the light is off, the photodiode only generates negligible reverse biased current (I_{DARK}), and V_{PH} becomes $V_{CC} = 15$ V.

Therefore, V_Z must be selected somewhere between $V_{PH} = 13$ and 15 V although voltages much lower then $V_{PH} = 13$ V are also acceptable design entries. Thus, assume $V_Z = 14$ V with $I_Z = 1$ mA from the Zener diode datasheet.

When the light is on, $\Delta v = V_{PH} - V_Z$, becomes $\Delta v = 13 - 14 = -1$ V, and the output of the operational amplifier, V_{OUT}, becomes 15 V. This voltage should drive the NPN transistor into saturation. Thus:

$$V_{OUT} = R_B I_{BSAT} + V_{BESAT} + V_M \tag{8.41}$$

$$V_{CC} = R_C I_{CSAT} + V_{CESAT} + V_M \tag{8.42}$$

Substituting $V_{OUT} = 15$ V, $V_{CC} = 15$ V, and the typical values of V_{BESAT} and V_{CESAT} into Eqs. 8.41 and 8.42 yields:

$$R_B I_{BSAT} = 15 - 0.8 - 5 = 9.2\,\mathrm{V} \tag{8.43}$$

$$R_C I_{CSAT} = 15 - 0.2 - 5 = 9.8\,\mathrm{V} \tag{8.44}$$

If 2N3904 is used in the circuit, then we can ignore the value of I_{BSAT} compared to I_{CSAT}. This produces $I_M = 100$ mA $\approx I_{CSAT}$. Then from Eq. 8.44:

$$R_C = \frac{9.8\,\mathrm{V}}{100\,\mathrm{mA}} \approx 100\,\Omega \tag{8.45}$$

From the I-V characteristics of 2N3904, the NPN transistor goes into saturation at $I_{CSAT} = 100$ mA when $I_{BSAT} \geq 100$ μA. Thus:

$$R_B = \frac{9.2\,\mathrm{V}}{100\,\mu\mathrm{A}} \approx 92\,\mathrm{K\Omega} \tag{8.46}$$

When the light turns off, $\Delta v = V_{PH} - V_Z = 15 - 14 = +1$ V, and V_{OUT} becomes 0 V. This voltage turns off the NPN transistor and therefore the DC motor.

8.10 Project 5: Designing Front-End Electronics for an Optoelectronic Tachometer

Among all the other projects the optoelectronic tachometer design is the most demanding because it contains both analog and digital building blocks to complete the design. The analog part shown in Fig. 8.20 is similar to Fig. 8.17 except the incident light beam to the photodiode is chopped by a four-bladed propeller. The data path in Fig. 8.20 amplifies and cleans the pulsed photo current, I_{PH}, generated by the photodiode. The output of this circuit, $clock_{VAR}$, indicates a variable clock signal, and is used to compute the prop rpm which varies in time. This variable but low frequency clock signal is tracked by a known reference clock, $clock_{REF}$, oscillating at 1 MHz, in order to compute the unknown frequency at $clock_{VAR}$.

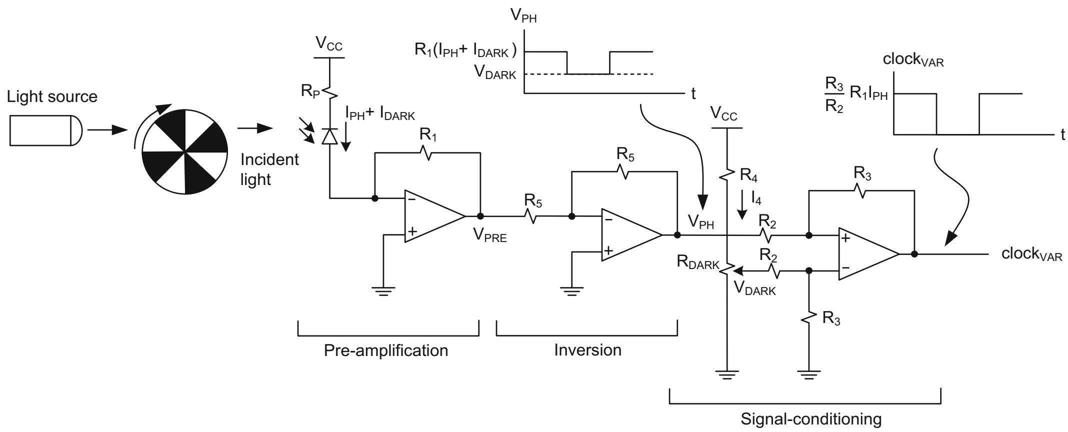


Fig. 8.20 Clock generation circuit for optoelectronic tachometer

The circuit in Fig. 8.21 shows two counters. The first counter, Pulse Counter, is a 16-bit wide counter and operates with the unknown clock frequency at $clock_{VAR}$. The second reference counter, Ref Counter, operates with 1 MHz clock at $clock_{REF}$, and it is 26 bits wide. The output of Ref Counter, $CountOut_{REF}$, is connected to a decoder whose sole purpose is to detect the end of a minute long time period. This translates to generating logic 1 at OneMinute node for a binary value of 11-1001-0011-1000-0111-0000-0000 at $CountOut_{REF}$ or a waiting for a period of 6×10^7 µs before generating a Stop signal with the 1 MHz reference clock. When the end detection occurs, the pulse generated at OneMinute port stops both the Pulse and Ref Counters. The total number of counts at the output of Pulse Counter, $CountOut_{PULSE}$, signifies the number of optical pulses received within 1 min time interval. However, depending on the number of blades on the propeller, this number still needs to be divided by the number of blades on the prop. Therefore, a four bladed prop requires $CountOut_{PULSE}$ to be shifted to the right by two. The output of the shifter, RPM_{OUT}, thus reads the number of revolutions of the four-bladed prop per minute.

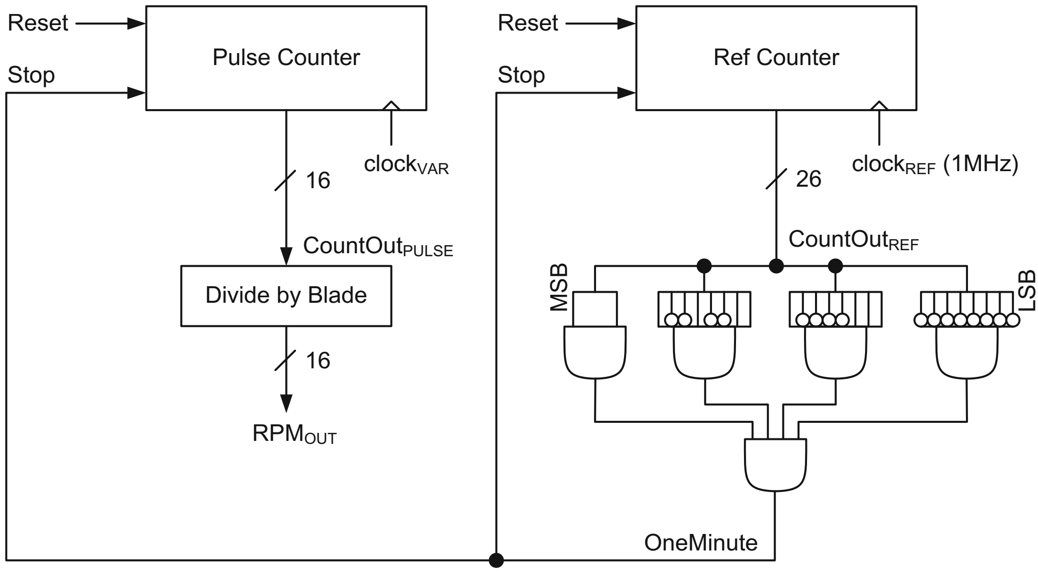


Fig. 8.21 An optoelectronic tachometer architecture

Figure 8.22 shows the circuit schematic of the counters. After an external reset, the pulse and reference counter outputs become $CountOut_{PULSE} = 0$ and $CountOut_{REF} = 0$, respectively, and both counters immediately start counting upwards using different clocks until they reach the 1 min mark. At the 1 min mark, the pulse received from the Stop input switches the port configuration of the 3-1 MUX from C-port to S-port, halting the count mechanism and putting both counters in idle mode. The counters stay in this state until they simultaneously receive an external reset.

The digital circuits in Fig. 8.21 can be implemented with discrete ICs from Texas instruments. However, it is best to consider a Field-Programmable-Gate-Array (FPGA) platform for implementations that require higher clock frequencies, especially when building timing-critical units such as a reference counter in a similar design.

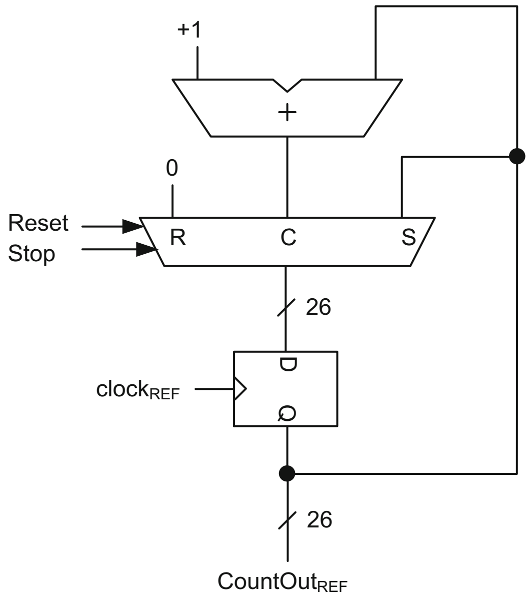


Fig. 8.22 Pulse and reference counter circuit for the optoelectronic tachometer

8.11 Project 6: Designing Front-End Electronics for a Hall-Effect-Based Tachometer

The circuit in Fig. 8.23 operates similar to the optoelectronic tachometer in the previous project but with a Hall-effect device in place. This project uses Allegro A1120 Hall-effect sensor shown in Fig. 8.10. Every time the sensor is exposed to magnetic field, as small as 35 Gauss, on a rotating disc in Fig. 8.23, the NPN bipolar transistor in the device turns on and pulls down the output node, clock$_{VAR}$, to 0 V.

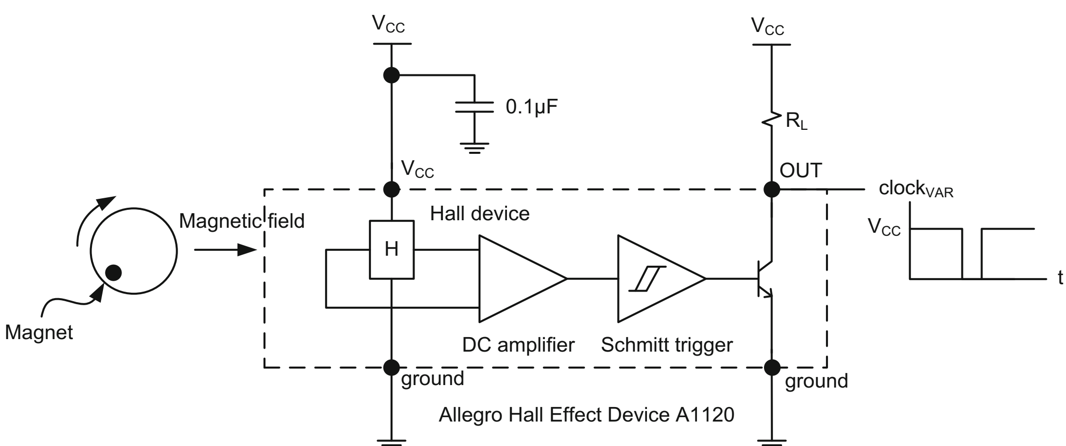


Fig. 8.23 The clock generation circuit for a Hall-effect-based tachometer

When the magnetic field is removed from the sensor or drops below 25 Gauss, on the other hand, the NPN transistor turns off, and the resistor, R_L, pulls up the $clock_{VAR}$ node to V_{CC}. This produces a clock waveform with a very brief duration at 0 V. A capacitor is added between V and ground to eliminate noise components in the power supply.

Applying this clock signal to the Pulse Counter, and a 1 MHz reference clock to the Ref Counter in Fig. 8.21 produces the same scenario described in the previous project. When 1 min elapses, $CountOut_{REF}$ produces a pulse at OneMinute port, and disables the counting mechanism in both the Pulse and Ref counters. The number of pulses at $CountOut_{PULSE}$ determines the rpm of the rotating disc. There is no need for a Divide by Blade unit, representing the number of blades in a prop, unless there is more than one permanent magnet on the rotating disc.

Review Questions

1. Design an analog servo controller that can rotate a servo arm to the left and right by producing pulse widths between 1 and 2 ms in a 20 ms period. How can this servo controller be designed using only digital blocks from Chapter 9?

2. Design a controller that generates pulses for a brushless motor. The design should include fine tuning the pulse width as well as the period of the pulse. Implement the same design using only digital blocks.

3. Design an analog interface to a microcontroller for reading the changing capacitance values of an accelerometer. Refer to Chapter 5 to be acquainted with the physical principles of accelerometers. Use digital logic design concepts from Chapter 9 to support the front-end electronics.

4. Wenner four pin method measures the resistivity of soil. A constant current is applied to the outer pins while the inner pins measure the voltage. The resistivity is calculated based on $\rho = 2\pi LR$, where L is the distance between the pins that measures voltage, and R is the resistance obtained from dividing the measured voltage by the constant current. When the resistance value rises above $R = 30\ \Omega$, irrigation becomes necessary. $R = 3\ \Omega$, on the other hand, indicates well-moist soil. The irrigation system for soil turns on with an active-high pulse when $R = 30\ \Omega$ at 6 am every morning and stays on as long as the pulse. A 10 min irrigation drops the resistance back to $3\ \Omega$. Design a system composed of analog and digital components that operate the irrigation system. Refer to Chapter 9 to design the logic blocks required for this system.

5. Design a traffic light module composed of red, yellow and green lights, which controls the flow of traffic on a major street. An optoelectronic device, such as a photodiode, monitors the rate of traffic (cars per minute) on this street. When the rate is 10 cars/min or less, the controller stops the incoming traffic for 1 min and turn on the green light for the cross traffic. The yellow light stays on for 1 s before the module transitions from green to red. Use the digital design principles from Chapter 9 to implement this system.

6. Design the front end electronics of a wind meter. Assume that wind rotates a four bladed propeller connected to the wind meter. Draw the circuit schematic between the prop and the microcontroller. Indicate the appropriate sensor(s) to be used in the system. Refer to Chapter 9 in order to design the digital portion of this system.

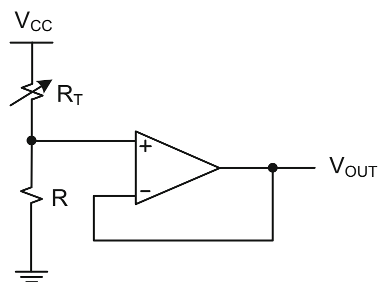
7. Distance between two points needs to be measured. A pulsed laser beam is placed at the point of origin; a reflective target is located at the destination. When the laser beam hits the target and reflected back to the unit, the unit effectively measures the time difference between the rising edge of the original pulsed laser beam and the rising edge of the

reflected beam to calculate the distance. Design the front-end electronics between the laser beam and the microcontroller. Use the digital logic design concepts given in Chapter 9 to support the analog electronics.

8. A resistive transducer measures the applied pressure in grams. When there is no force, the sensor shows in excess of 1 MΩ resistance. However, when pressure is applied, the resistance decreases with increasing pressure. Some of the resistance readings are given below:

$$\begin{array}{ll} \text{Force} = 10\,\text{grams} & \text{Resistance} = 100\,\text{K}\Omega \\ \text{Force} = 100\,\text{grams} & \text{Resistance} = 10\,\text{K}\Omega \\ \text{Force} = 1000\,\text{grams} & \text{Resistance} = 1\,\text{K}\Omega \\ \text{Force} = 10000\,\text{grams} & \text{Resistance} = 0.1\,\text{K}\Omega \end{array}$$

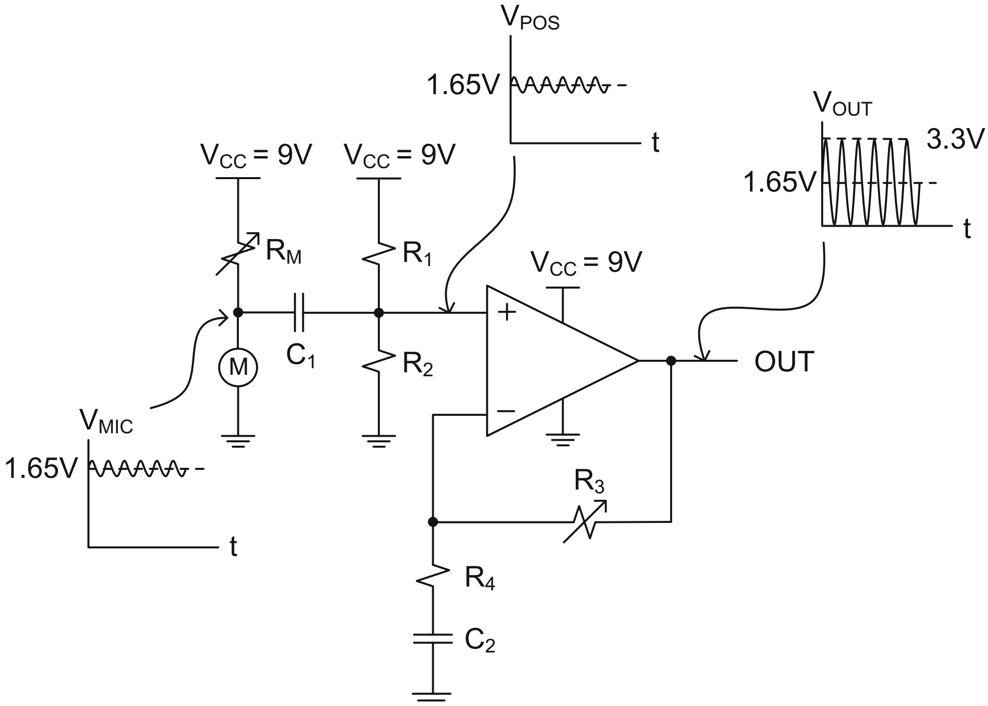
An operational amplifier circuit given below can be used as the initial stage for this system with R_T, corresponding to the changing resistance of the transducer.



If it becomes necessary, use the digital logic design principles given in Chapter 9 to implement the system.

9. A triple light optoelectronic detection system needs to be designed such that it can distinguish red, yellow and green lights from each other. Conceptualize the type of sensor array to implement such a system, and design its analog front-end electronics for the microcontroller. Refer to Chapter 9 in order to use the necessary digital logic blocks to complete this system.

10. Use the operational amplifier circuit below to amplify human voice with an analog microphone. Use $V_{CC} = 9$ V, $C_1 = C_2 = 10$ µF and the potentiometers, $R_M = R_3 = 100$ KΩ. Here, C_1 and C_2 become short circuits when an Alternating Current (AC) signal (i.e. human voice) is applied to the circuit. However, these capacitors become open circuits in Direct Current (DC) when there is no AC signal. Obtain the values of R_1, R_2 and R_4, in order to obtain the waveforms at V_{MIC}, V_{POS} and V_{OUT} as shown.



V_{POS}
1.65V
t
V_{OUT}
3.3V
1.65V
t
V_{CC} = 9V
V_{CC} = 9V
R_M
R_1
V_{CC} = 9V
M
C_1
R_2
OUT
V_{MIC}
1.65V
t
R_3
R_4
C_2