Chapter 5

Interfacing to Integrated Hall-Effect Devices

Integrated Hall-effect sensors almost always incorporate enough signal-processing and support circuitry to provide an immediately useful output signal, be it a proportional voltage or a digital switched output. There are situations, however, when the electrical interfaces provided by the sensor do not meet the requirements of a particular application. In these cases, the designer must provide some additional circuitry to bridge the gap between the sensor's output and the inputs of the system to which it is being interfaced.

5.1 Interface Issues—Linear Output Sensors

Because the output of a linear Hall-effect sensor is a voltage proportional to magnetic flux density, it offers the highest degree of flexibility in interfacing to an outside system. By providing the ability to measure magnetic field, a linear sensor is a building block with which one can begin to implement nearly any kind of magnetic sensing function required. Potentially an enormous number of types of interfaces can be implemented; four of the more common types are:

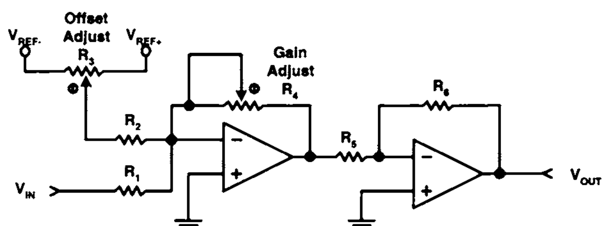
1) Offset and sensitivity adjustment
2) Line-driver circuits
3) Output thresholding
4) Analog-to-digital converter (ADC) interface

The following sections describe circuits to perform these interface functions.

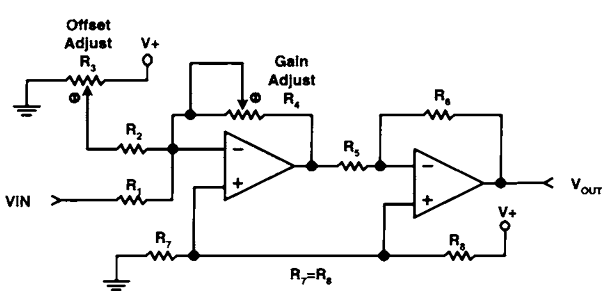
5.2 Offset and Gain Adjustment

A common interfacing situation occurs when the output offset and span of a sensor do not match the input offset and span of some other electronic circuit that needs that sensor's input. While some of the newer linear Hall-effect sensors offer user-adjustable gain and offset controls on-chip, circumstances will still arise in which what is available will not satisfy a given set of requirements. One example would be where a signal swing of ±10V is required. No common integrated linear Hall-effect sensor provides an output that swings below 0V, or has an output range that spans 20V (–10V to +10V). If such an output is needed, it requires the addition of some custom circuitry.

In systems where one has the luxury of operating from ± split power supplies, such as +5/–5V or +15/–15V, the circuit of Figure 5-1a provides trim for both sensitivity and offset. The sensitivity gain is given by $(R_4 \times R_6)/(R_5/R_1)$. The range of offset adjustment as referred to the output of op-amp A_2 is $\pm (V_{ref} \times R_4 \times R_6)/(R_2 \times R_5)$ volts (within the limitations of the op-amps' range of output voltage swing). With some types of op-amps, it may be necessary to keep $R_4 \geq R_1$ and $R_6 \geq R_4$ to keep the circuit from oscillating. In many cases, where a large amount of gain is not needed, you can set $R_5 = R_6$, providing the output stage with a gain of –1, and simplifying the total gain to R_4/R_1 and the offset adjust range to $V_{ref} \times R_4/R_2$. When implementing this circuit with "typical"op-amps, suitable values for the resistors will range from roughly 1000Ω to 1 MΩ.



(a)



(b)

Figure 5-1: Offset and gain adjustment circuits for split voltage supply systems (a) and single voltage supply systems (b).

Note that this circuit has two separate amplification stages and requires two op-amps. This is because the first-stage amplifier configuration is inverting; inputting a positive signal results in a negative output signal. Adding another inverting gain stage flips the polarity back to positive. In this circuit, the offset adjustment is applied "upstream" in the signal-processing chain before the gain adjustment. The proper order in which to adjust this circuit is to first adjust the gain to the desired level, through resistor R_4 and, once a satisfactory gain has been achieved, then to adjust the output offset voltage, using R_3.

In many cases, one doesn't have the luxury of operating from a split voltage supply, and has to make the system work from whatever is available, often from a single +5V or +12V supply. The circuit of Figure 5-1a can be modified for single-supply operation by providing the op-amps with a "virtual ground" somewhere between the positive supply and ground. For a 5V system, 2.5V is often a good choice for the virtual ground value. The simplest way to make a virtual ground is by connecting a resistive divider between the positive supply and ground. In this circuit, this function is performed by resistors R_7 and R_8. Placing a small bypass capacitor at this virtual ground point is often helpful in improving the circuit's stability and reducing the output noise.

In cases where cost or space is an issue, it is also possible to perform gain and offset adjustments with fewer components. An example of a single-supply gain and offset adjustment circuit that uses only a single op-amp is shown in Figure 5-2. In this circuit, gain and offset adjustment are very much interdependent. Adjusting the offset will also change the gain. This is because the gain of the circuit is partially dependent on the impedance looking into the wiper of potentiometer R_3, which ranges from zero when it is at either of the ends (minimum or maximum adjustment) to $R_3/4$ when it is in the middle (where R_3 is the end-to-end resistance of the potentiometer). Needless to say, having the gain and offset interdependent can make this circuit truly a joy to properly adjust. If you choose a value for R_3 that is much smaller than that of R_2, the gain of the circuit can be approximated by $R_2/(R_1 + R_2)$, and the output offset adjustment range by $5V \times (R_1/R_2)$ (for 5V operation). Note that when you adjust the wiper on R_3 toward the positive supply, the output voltage will go down.

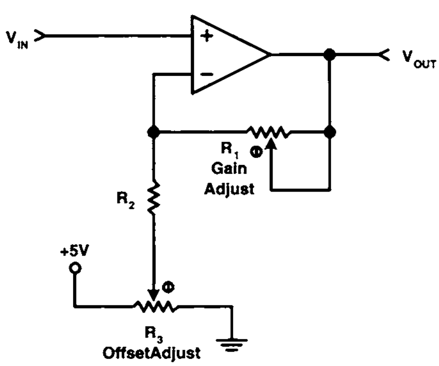


Figure 5-2: Offset and gain adjustment with a single op-amp.

Depending on the requirements of your application, it is also possible to "hard-wire" the gain and offset adjustments on the previous circuits by replacing the potentiometers and variable resistors with fixed resistors. If fixed offset and gain adjustments are all that are needed, this is often a desirable thing to do. Variable resistors are among the least reliable electronic components available; fixed-precision resistors offer orders of magnitude higher reliability, for orders of magnitude lower cost. Additionally, manually adjusting a variable component is an expensive and error-prone operation to perform in even a medium-volume production environment. If you don't absolutely need the adjustability, don't use variable resistors in your circuits.

5.3 Output Thresholding

Although there is an enormous variety of switch and latch-type Hall-effect sensors available, providing a multitude of available B_{OP} and B_{RP} points, there may be occasions where you need a device with specific B_{OP} and B_{RP} values. One option is to use a programmable switch-device such as the Allegro Microsystems A3250. In cases where your requirements fall outside the range of available fixed and programmable devices, it is possible to build your own switches and latches around a linear output Hall sensor. This approach allows you nearly complete freedom in setting B_{OP}, B_{RP} and hysteresis. The primary drawbacks are that your discrete version of a switched device will be considerably more expensive than a comparable integrated device and will take up much more space. If your application allows for these increased cost and space requirements, however, then the do-it-yourself approach can be viable.

One method of building a digital sensor out of a linear one is through the circuit of Figure 5-3. This is similar to the conceptual model of a switch presented in Chapter 3, except with a few more implementation details included. When the output of the linear Hall-effect sensor exceeds the B_{OP} threshold, it causes a HIGH (5V) condition at the output of comparator U_{2A}. This HIGH condition causes the flip-flop (U_3) to latch into the HIGH state. When the output of the linear device drops below the B_{RP} threshold, it causes a HIGH at the output of comparator U_{2B}, latching the flip-flop into a LOW state. The hysteresis is the difference between the B_{OP} and B_{RP} points.

The circuit in Figure 5-3 allows for B_{OP} and B_{RP} to be set completely independently, with B_{OP} normally set higher than B_{RP}. For many applications, however, one will be interested in being able to set a single threshold, with only a very small amount of hysteresis. In the presence of a low level of hysteresis, the B_{OP} and B_{RP} thresholds may be nearly indistinguishable. A simple circuit that provides a single threshold with a small amount of hysteresis is shown in Figure 5-4.

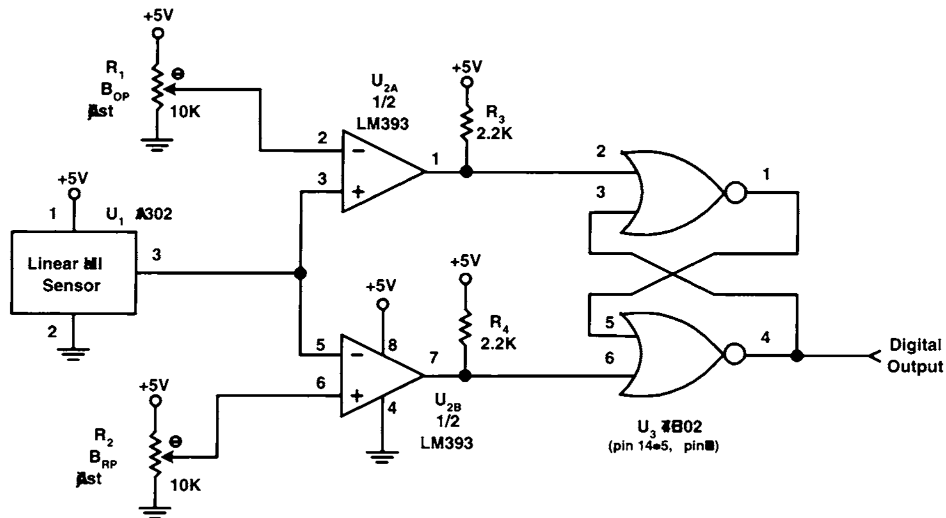


Figure 5-3: Discrete implementation of switch.

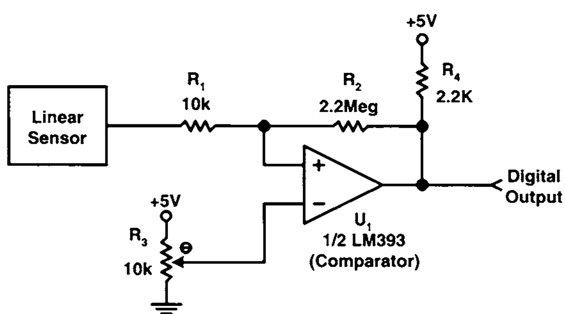


Figure 5-4: Threshold switch with a small amount of hysteresis.

The comparator in this circuit works in the same way as those of the previous circuit, outputting a HIGH condition when the voltage at its positive input is greater than the voltage at its negative input. The network comprising R_1 and R_2 provides positive feedback in this circuit. This means that when the output switches HIGH, it forces the comparator's positive input to also move to a slightly higher voltage. Conversely, when the output switches low, it pulls the positive input down with it. The effect is analogous to the snap in a light switch on an electric lamp—when you push it past the halfway point in either direction, a spring in the switch tends to pull it the rest of the way. The amount of this shift, which is the threshold detector's hysteresis, is approximately given by $V_S * R_1/(R_1 + R_2)$, where $V_S = 5V$ in this circuit. For the values of resistors and V_S

shown, the hysteresis will be about 24 mV. For a linear sensor such as the Allegro A3515, with a sensitivity of 5 mV/gauss, 24 mV of electrical hysteresis would translate into just under 5 gauss of magnetic hysteresis.

Despite its apparent simplicity, this circuit has several drawbacks. Because the amount of hysteresis is dependent on the swing of the comparator's output, anything you do to load down that point will reduce the effective hysteresis. Another issue is that the resistance of R_2 must be significantly greater than that of R_4; otherwise, the hysteresis feedback will load down the output all by itself. If one takes precautions to ensure that the loading on the circuit's output is controlled, this circuit can provide reasonably well-behaved and predictable performance.

5.4 Interfacing to Switches and Latches

One might think that, because the output of a switch or a latch is digital, the whole job of interfacing is already complete. While it is a straightforward matter to interface the output of a typical Hall-effect switch to a digital input, there are also many other, not so obvious ways that digital Hall-effect sensors can be used.

5.5 The Pull-Up Resistor

As mentioned earlier, most digital Hall-effect switches and latches provide an open-collector or open-drain type of output. The output therefore behaves like a switch to ground, and does not output any voltage on its own. In order to derive a useful voltage-level signal, one must add a pull-up resistor, as shown in Figure 5-5. One simple but common question that arises from the necessity of the output pull-up resistor is how to select an appropriate resistor.

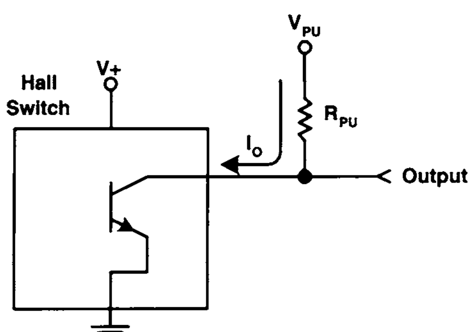


Figure 5-5: Output pull-up resistor.

Only a few pieces of information are required to select and size a pull-up resistor. The first is V_{PU}, or the voltage to which the pull-up resistor is to be tied, which is not necessarily the same as the supply voltage for the sensor. The second piece of information is the current (I_{ON}) that should be flowing through the Hall sensor's output when it is in the ON state. This current is not necessarily the maximum current the device can handle (I_{Omax}), but is determined by the application (limited, of course, to less than I_{Omax} !). Just because a Hall-effect sensor can handle 20 mA through its output doesn't mean that the output current must be 20 mA. Sometimes, indeed most of the time, sizing a pull-up resistor for a lower current level makes sense. In any case, for a given V_{PU} and a desired output current I_O, the resistance value of a suitable pull-up resistor is given by:

$$R_{PU} \approx \frac{V_{PU}}{I_O}$$ **(Equation 5-1)**

For the case of V_{PU} =12 V, and I_O=10 mA, an appropriate pull-up resistor value would be 1.2KΩ.

The second piece of information is the power rating of the resistor. If you do not pick a resistor with a power rating greater than its actual power dissipation, the circuit may experience reliability problems in the field. Again, this is a simple calculation, with power dissipated given by:

$$P = \frac{(V_{PU})^2}{R_{PU}}$$ **(Equation 5-2)**

To continue the example above, a 1200Ω pull-up resistor attached to a 12V pull-up supply would dissipate ($12^2/1200$) or 0.12 watts. While this would not pose a problem for a through-hole ¼W resistor, it would be an overload condition for an 0805-type surface-mount device, which is typically rated for 1/10W of power dissipation.

The above analysis for power dissipation assumes that the Hall sensor is constantly in the ON state, and that the resistor is constantly dissipating power. In many applications, the Hall sensor will be turning on and off, and the average power dissipation may be considerably lower than the case in which the sensor is constantly on. While one could size the resistor based on a presumed "duty cycle," it is much better practice to design the circuit to be able to deal with the worst-case power dissipation—i.e., that the sensor is always on and dissipating power in the resistor. Additionally, it is also generally good design practice not to operate electronic components at their maximum power ratings. For example, one would probably not want to select a 1/8W (0.125W) resistor to serve in a place where it could dissipate 0.12W on a continuous basis; a larger resistor, say 1/4W, might be a better choice. In addition to considering a device's power dissipation and ratings, one must also consider the effects of factors such as ambient operating temperature, airflow, and secondary packaging. Selecting component power ratings can be a complex process, well beyond the scope of this text.

Because electronic components tend to be particularly rugged devices, however, a device being driven at a significant thermal overload may work for a while. This means that, instead of failing immediately in the factory, where it merely becomes scrap, an assembly with overloaded devices may make its way out to a customer, where it can become a field-return or a field-service call, or worse. With the introduction of small surface-mount components with limited power dissipation (1/16W for a 0603-sized resistor), component power ratings have become an issue that must be taken very seriously, even in circuitry not normally perceived as being "high-power."

5.6 Interfacing to Standard Logic Devices

Interfacing digital Hall-effect output devices to most modern digital logic devices is extremely straightforward. Here are some tips for interfacing to some of the more common logic families in present use.

Bipolar TTL – (74xx, 74Sxx, 74LSxx, 74Fxx) – connect the pull-up resistor to the logic power supply (Vcc, +5V). Because very little current (< 1 mA) is required to pull a TTL input to logic HIGH (>2.4V), you can use a large pull-up (2. 2kΩ – 4.7 kΩ) for lower system power consumption. While a TTL logic input driven from an open-collector output may sometimes seem to work fine with NO pull-up resistor, this is a bad situation from a reliability standpoint, and should be avoided.

CMOS – (74HCxx, 74HCTxx, CD4xxx) – The pull-up should be connected to the positive voltage supply (VDD) for the logic. For 74H and 74HC families, VDD can range from about 3V–6V, and from 3V–15V for CD4xxx family devices. Because CMOS digital inputs look like a capacitor, very high pull-up resistors (100 kΩ) can be used in situations where one does not especially care about the rise time of the signal. If the resistor is made too large, however, the rise time on the sensor output signal may become unacceptably long. In the case of CMOS logic, if you forget to put in the pull-up resistor, the circuit will work fine as long as the Hall-effect switch's output is ON (pulling the output low to ground) but when it turns off, the input to the CMOS logic gate will float and assume a highly unpredictable value. This type of design error can be very difficult to troubleshoot. One symptom of this problem is that the output stays low when a DVM or an oscilloscope is hooked to the offending logic input, but becomes spurious when the test instrument is removed.

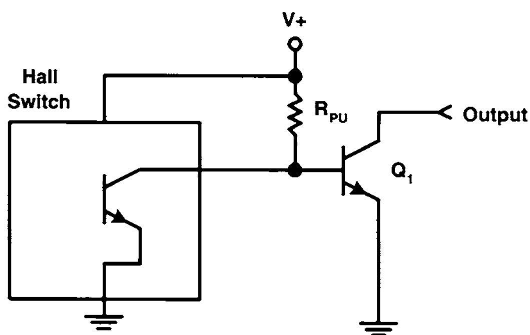
Microprocessor inputs – While most microprocessor input ports look like CMOS inputs (and usually are CMOS inputs nowadays), several variations are occasionally encountered. Built-in pull-up resistors are fairly common on many microprocessor input ports, and can save you from having to add external pull-up resistors. One problem, however, that comes up when trying to interface digital Hall-effect sensors to a microprocessor or microcontroller is that many of the I/O lines are bi-directional, meaning that they can be either inputs or outputs, depending on how they are configured by the software running on the microprocessor. It is important to ensure that there are no tug-of-war situations in which the output of the Hall sensor and an output of the microcontroller both try to control the logic level of a single line. When such a contentious condition exists, one device will lose, your circuit will not work reliably, and in extreme cases hardware damage may result.

5.7 Discrete Logic

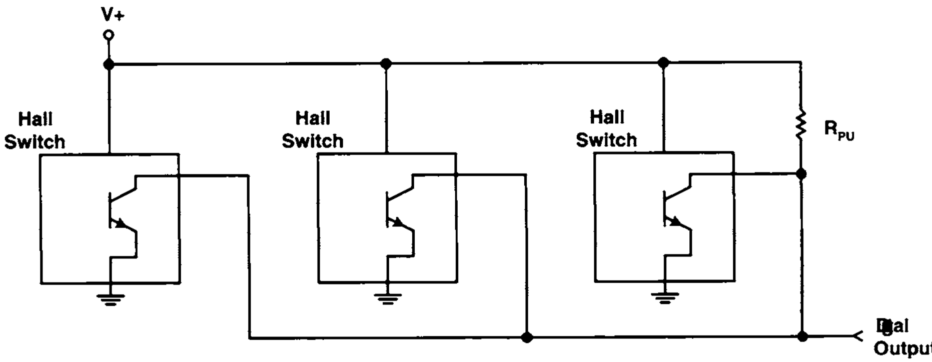
It is not uncommon for a sensor assembly to incorporate some amount of logical decision-making capability. In the future this will become increasingly common. In most cases, to implement simple logic functions one would use small-scale integrated logic circuits, such as 74HCxx family devices or programmable logic devices (PLDs). For more sophisticated decision-making capabilities, a small microcontroller might be used.

To perform a few very simple logical operations, such as inverting the polarity of an output, it sometimes makes more sense in a design to use logic functions made from discrete components, such as resistors, transistors, and diodes. There are two reasons for taking this approach. The first is cost. If a logic function can be implemented with a few discrete components, it can actually be more cost effective than one implemented with an integrated circuit. The second reason is that most integrated logic families have a limited operating voltage range. For a sensor assembly that must operate over a 4–24V supply voltage range, using off-the-shelf logic circuits usually requires a voltage regulator to provide a suitable stable power supply (often 5V or 3.3V) for the logic circuits. For both of these reasons there are situations where it often makes sense to design one's own logic functions from discrete components.

The design of robust discrete-transistor logic is a nontrivial task. Many factors, such as operating voltage range, power consumption, immunity to electrical noise, and switching speed must be considered in the course of designing the circuits. When you use a standard logic IC, these issues have already been addressed by the logic manufacturer. When you design your own logic from discrete components, you are on your own. The example circuits that follow should be considered a starting point and not final designs; as simple as they are, they may require significant modification to meet the exact requirements of a given application.



(a)



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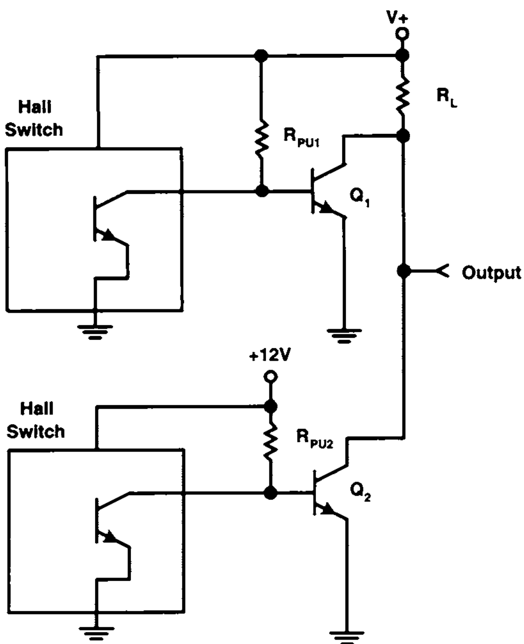

(c)

Figure 5-6: Discrete logic circuits. Logic NOT (a), Logic OR (b), Logic AND (c) polarity inverter.

Logic NOT Output

This circuit (Figure 5-6a) converts a "normally open" (OFF) open collector output into a "normally closed" (ON) open collector output. When the Hall-effect sensor (U_1) is

OFF, current flows through the pull-up resistor (R_{PU}) into the base terminal of transistor Q_1, turning it ON. Conversely, when the Hall-effect sensor is ON, it diverts current from Q_1, leaving it OFF. This function is called *logic NOT* or *logic inversion.*

Wired OR Output

Open collector outputs can be tied together using a single pull-up resistor, as shown in Figure 5-6b. The resulting circuit is called a *wired-OR* configuration (also sometimes called a *wired AND* in logic design). When one or more of the Hall-effect sensors turns ON, the output is pulled low (ON). This function is known as *logic OR.*

AND function

Sometimes you need to know when *all* of a group of sensors are on. This function is called *logic AND.* The circuit of Figure 5-6c shows one way to do this. It works by combining logic NOT functions with the wired OR function. If a given sensor (U_1) is ON, it causes the associated transistor at its output (Q_1) to be OFF. If both Q_1 and Q_2 were OFF, as would result from both sensors (U_1 and U_2) being ON, then the output will be pulled high to +12V through R_L. If one needs an output that is switched to ground when all of the sensors are on, one can use the output signal shown to drive the base of an additional transistor (not shown) to obtain this switching function.

This circuit also suggests the upper limit of complexity for which it is probably worth designing logic from discrete transistors. This circuit contains three resistors and three transistors. When the costs of the components and the costs of stuffing them onto a circuit board are considered, using an integrated AND gate is likely to be a more economical option, despite possible requirements for additional support circuitry.

5.8 Driving Loads

Hall-effect sensors are sometimes used to control various types of lamps and electromechanical devices that require a significant amount of power to operate. This section will describe some simple circuits for interfacing with these types of devices.

5.9 LED Interfaces

Most digital Hall-effect sensors are capable of sinking up to 20 or 25 mA of output current safely. This also happens to be a more-than-adequate amount of current for driving many small light-emitting diodes (LEDs). This makes it easy in most cases to directly drive an LED from a digital Hall sensor. Figure 5-7a shows a circuit in which the LED lights up when the Hall-effect sensor is ON. Figure 5-7b, on the other hand, shows a circuit in which the LED illuminates when the Hall-effect sensor is OFF; when the sensor is ON in this case, it shunts current away from the LED. In either case, the major design parameters are the desired operating current for the LED and the voltage at which one needs to operate. Because a red LED typically has a forward voltage drop of about 1.8V (the voltage drop depends on the variety of LED—for example, blue LEDs usually have voltage drops of around 3.5 to 4.0V) you need to take this into account to get the right operating current, especially when operating at lower (5V) supply voltages. An appropriate resistor value (when using our typical red LED) can be estimated by

$$R_1 \approx \frac{V_S - 1.8}{I_{LED}} \qquad \textbf{(Equation 5-3)}$$

where V_S is the supply voltage and I_{LED} is the desired operating current. Another issue is selecting the appropriate power dissipation for the resistor. In the case of the first circuit (Figure 5-7a), the resistor only dissipates power when the LED is ON, and the approximate dissipation is given by $(V_S - 1.8V)^2/R_1$, (again, where 1.8V is the LED's voltage drop). In the case of the circuit of Figure 5-7b, the resistor dissipates power all the time, with a maximum dissipation of V_S^2/R when the LED is OFF. One odd "feature" of this circuit is that it actually draws more current when the LED is not illuminated.

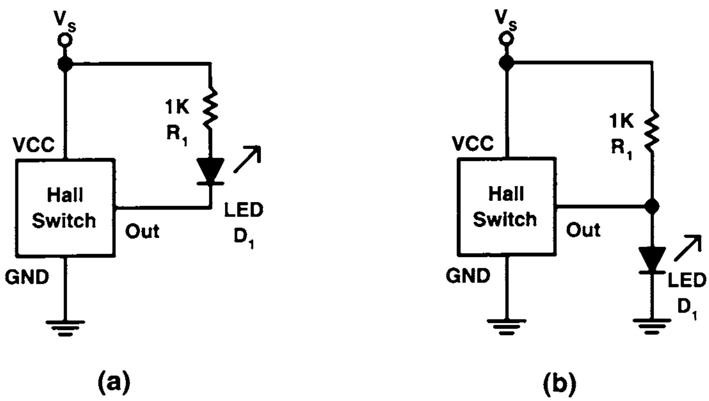

Figure 5-7: Normally OFF LED driver (a), normally ON LED Driver (b).

5.10 Incandescent Lamps

A low-voltage (3–12V) incandescent lamp can provide much more light output than an LED, but also requires much more current to produce that light. From a circuit standpoint, an incandescent lamp is a resistor. Unlike a typical resistor, however, an incandescent lamp will draw a brief surge of up to several times its normal operating current when turning on because the filament resistance is lower when the lamp is cold. This surge is called the cold-filament current or inrush current, and can be more than

10 times the lamp's normal operating current. When designing a lamp driver, one must use transistors that can handle these short high-current inrush spikes without self-destructing. Two simple circuits for controlling small lamps (12V, < 1A current) from the outputs of digital Hall-effect sensors are shown in Figure 5-8.

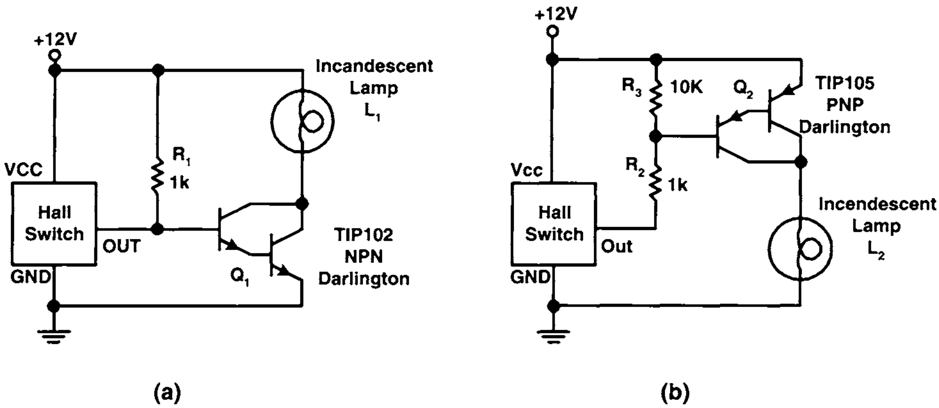

Figure 5-8: Incandescent lamp drivers using bipolar Darlington transistors. Inverting (a), and noninverting (b).

These lamp drivers use what is known as Darlington transistors because they provide a much greater current gain (typically >1000) than that of a typical bipolar transistor (typical gains from 50 to 200). This ensures that the ≈10 mA of available base current will completely switch the transistor ON while carrying a load current of up to several amperes. Figure 5-8a shows an inverting driver (Lamp ON when sensor OFF), while Figure 5-8b shows a noninverting driver (Lamp ON when sensor ON). R_3 is used in this circuit to ensure that the transistor turns completely OFF when the output of the sensor is OFF (and not sinking current).

The circuits shown above do show a few nonideal behaviors. The price paid by using a Darlington transistor is high output saturation voltage. The Darlington's collector-emitter saturation voltage, even when switched ON completely, will never drop below about 1V. Aside from the reduction in voltage to operate the lamp (resulting in less light output), this also results in significant power dissipation in the transistor. For example, a 1Ampere current in either of the circuits of Figure 5-8 will dissipate approximately 1W in Q_1, possibly requiring the transistor to be heat-sinked.

Another, and typically better, approach is to use a power MOSFET as the lamp driver. Figure 5-9 shows an inverting lamp driver using a power N-channel MOSFET (lamp is ON when sensor is OFF). When switched completely ON, the particular MOSFET shown in Figure 5-9 has an ON resistance of about 0.2Ω, and will only dissipate 200 mW with a 1A load current. MOSFETs with lower ON resistances are also available, down to values in the range of a few milliohms. Since no current flows into the gate terminal of a MOSFET, it is possible to use MOSFET drivers to switch large load currents, while being controlled directly from the output of a digital Hall-effect sensor.

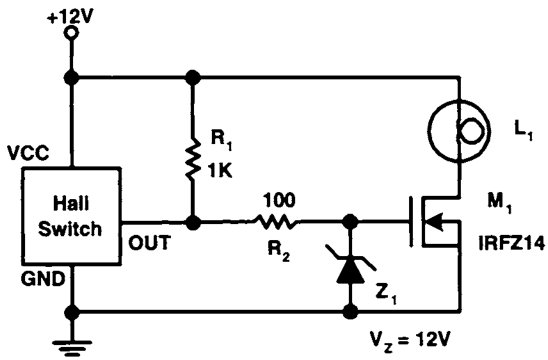

Figure 5-9: Intverting lamp driver using power MOSFET.

MOSFETs, however, have their own set of idiosyncrasies. The first is that since they are voltage controlled devices, you need a significant voltage swing at their gates terminal to completely turn them ON. Depending on the device and the operating conditions, this usually ranges anywhere from 3 - 12 volts. If R_1 were tied to a 5 volt supply, this circuit would either work marginally (the lamp would be dim and the MOSFET would get hot in the ON state), or not at all. One remedy to this problem is to select a MOSFET that is designed for operation at low-gate drive voltages.

Another characteristic of MOSFETS is that their gates are sensitive to damage from even momentary over-voltage conditions. For this reason, it is common to put a zener diode or other device to clamp the maximum gate-source voltage excursions. In this example a 12V zener diode was selected to clamp the gate-to-source voltage excursion. Some modern power MOSFETs, however, have maximum gate-to-source voltage ratings of as little as ±6V, and must be protected accordingly.

A final characteristic of MOSFETs to be discussed here is gate capacitance. The gate of a power MOSFET looks like a large capacitor, typically in the range of 100–10000 pF. It is often advisable to put a small resistor (R2) in series with the gate to both limit the turn-on and turn-off current spikes, and to help prevent oscillation. Despite a few additional and nonobvious design issues to consider, for many power switching applications MOSFET transistors are a better solution than bipolar transistors (Darlington or otherwise).

5.11 Relays, Solenoids, and Inductive Loads

While resistive loads, such as incandescent lamps are relatively benign (once one accounts for the issue of cold-filament inrush current), inductive loads such as relays and solenoids are another story. If you were to drive even a small relay (e.g., 12V, 50-mA coil drive) with one of the lamp drivers described above, the circuit would probably work for a few ON-OFF cycles, then cease operating. This is because an inductive load such as the coil of a relay can produce very large voltage spikes (hundreds or even thousands of volts) if it is abruptly turned off. Consider the circuit of Figure 5-10a: when the relay coil is energized, current will flow through it. When the coil is switched off, because the relay coil is an inductor, the current will keep trying to flow. The coil will raise the voltage at the collector of the transistor value to try to keep the current flowing. If the voltage at this point happens to be greater than the breakdown voltage rating of the transistor, this tends to damage the transistor. In many cases, this circuit may operate successfully a few times and then fail after the transistor becomes sufficiently damaged by multiple voltage spikes.

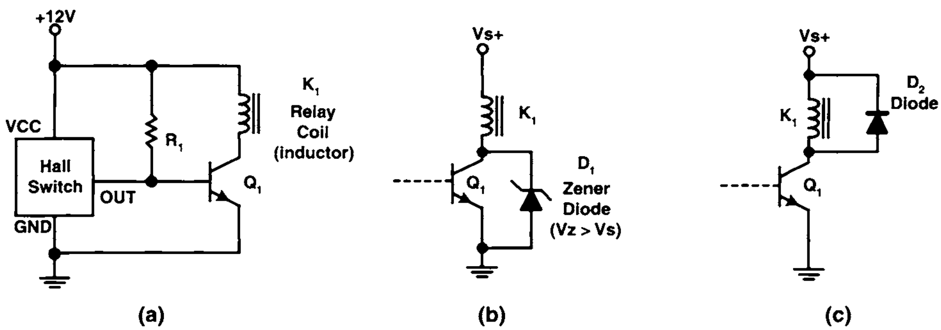

Figure 5-10: Driving a relay load. In a way that may damage the transistor (a). Protecting the output transistor with a zener diode (b) and a rectifier diode (c).

Having a circuit fail after a few or even a single operating cycle is clearly unacceptable in the overwhelming majority of applications one could imagine. There are, however, several simple techniques that can be used to protect the output transistor from an inductive load. One way is to connect a zener diode (D_1) at the transistor's collector, as shown in Figure 5-10b. While this works, it requires that the zener diode be able to dissipate the bulk of the energy stored in the inductor. As an example, for a relay operating at 12V and 250 mA, you would want to select a diode with a zener voltage of >12V, let us say 16V for this example. If you pick a zener diode with a zener voltage less than the supply voltage, it will be turned on all the time, and continuously conduct current. When the relay turnes off, the diode will have to handle a current spike of 250 mA (the relay ON current), with a drop of 16V (the zener voltage), resulting in a 4W peak power dissipation. While this power dissipation may only need be sustained for a few microseconds, it is important that the diode can handle it (see the manufacturer's datasheets!). When taking this approach, it is also important to select a transistor that has a collector breakdown voltage that is higher than the zener diode's breakdown voltage, or the transistor will simply break down before the zener begins conducting. In cases where an inductive load requires high-voltage or high current, a suitable zener diode may be either difficult to find or excessively expensive.

Another protection circuit is shown in Figure 5-10c. When the coil is turned off and the collector voltage rises above the supply voltage, diode D_2 will turn on, effectively shorting out the coil. This circuit provides the advantage of dissipating most of the inductor's energy through the inductor's winding resistance. This circuit configuration is commonly called a *fly-back diode*. An additional benefit of using a fly-back diode protection circuit is that you can use ordinary diodes as opposed to zener diodes. Suitable rectifier diodes tend to be less expensive than zener diodes of comparable power and current ratings. There are three major requirements for the diode. The first is that it can handle the coil current; diodes are readily available that can handle several amperes continuously. The second requirement is that the diode's reverse breakdown voltage be greater than the supply voltage; again diodes with reverse breakdown voltages of up to several hundred volts are easy to find and inexpensive. The final requirement is that the diode be able to switch on quickly enough to keep the voltage at the collector of the output transistor from rising too high. Ideally, the fly-back diode will limit the transistor collector voltage to the supply voltage plus 0.6V (one diode drop). In reality, the actual peak voltage may be somewhat higher; how much higher is dependent on the characteristics of the coil, the driver circuit, and the fly-back diode used.

5.12 Wiring-Reduction Schemes

If you have a large number of digital Hall-effect sensors in a system, wiring and interconnection can become a serious issue from size, manufacturability, and cost standpoints. This section presents several approaches to reducing the number of wires needed to read sensor status. The trade-off, of course, is that you will need some additional electronics.

5.13 Encoding and Serialization

If only one digital Hall-effect sensor in a group is activated at a time, it is possible to have it impress an identifying code on several wires. Figure 5-11 shows one method for encoding up to 2^N-1 digital sensors on N data lines. Diodes are used to isolate the data lines from each other. If the outputs of each sensor were tied directly to several wires, they would all be shorted together, and the scheme would not work. While the output voltage on the data lines will rise to the V_{CC} voltage when HIGH, they will only go down to about 0.7V when LOW, because of the voltage drop caused by the diodes. Because the LOW voltage will be about 0.7V, some additional interface circuitry may be needed before interfacing the outputs to a microcontroller or other logic circuitry.

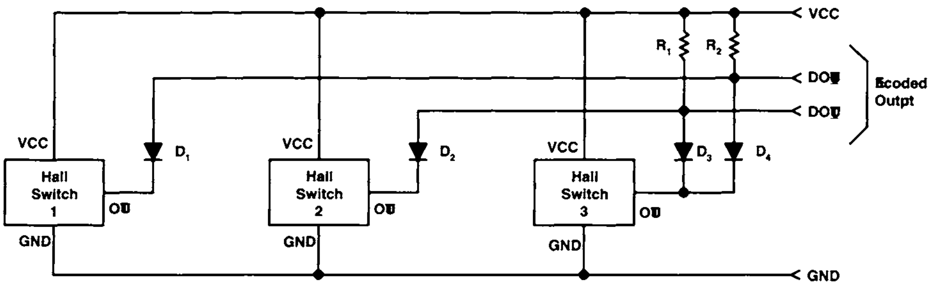

Figure 5-11: Encoding multiple digital Hall-effect sensors.

In the previous circuit, if more than one Hall-effect sensor were activated, it resulted in an ambiguous output. If the possibility exists that multiple devices can be activated simultaneously, this circuit cannot be used reliably. In cases like these, it is possible to add more electronics to be able to read devices independently of each other. An example of a system that allows independent sensor polling is shown in the block diagram of Figure 5-12.

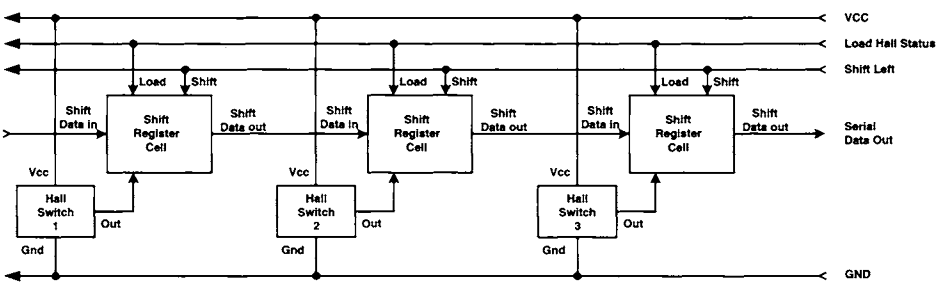

Figure 5-12: Serial shift-register multiplexing.

In this scheme, the status of all the Hall-effect sensors is first read into a string of shift register cells when the LOAD line is pulsed. Then, each time the shift line is pulsed, the status of each shift register cell is passed along to the one to its right. By loading the status of all of the sensors once, and then by repeatedly shifting the data out one bit at a time, it is possible to sequentially read the status of all of the sensors. The advantage of this technique is that it only requires five wires (VCC, GND, Shift, Load, and Serial Data Out), regardless of the number of sensors.

5.14 Digital-to-Analog Encoding

If one has a spare analog-to-digital input line with which to read sensor status, it is also possible to encode the status of multiple sensors as an analog signal. Figure 5-13 shows how the outputs of a few digital sensors can be encoded into a single binary-weighted analog signal with the addition of a few resistors.

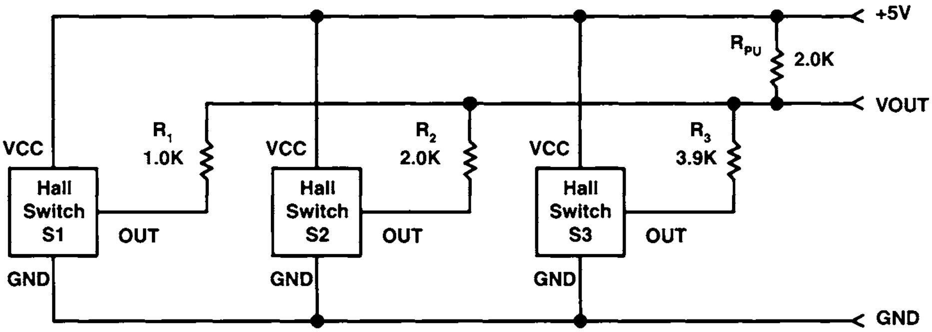

Figure 5-13: Binary-weighted encoding of digital sensors into an analog signal.

Table 5-1 shows the relationship between the output states of each sensor, and the resulting voltage at V_O. In this example the resistors are chosen to increase by a factor of two (binary weighting) between adjacent devices, ensuring that each combination is unique. One feature of this technique is that it is possible to independently monitor the status of any sensor in the circuit.

The voltages seen in any given implementation will be slightly different, varying as a function of resistor tolerance and because of the effects of the output saturation voltage of the Hall sensors used. Because of these resistor tolerance and V_{SAT} error issues, the number of sensors which can be encoded together in this manner will be restricted, with four or five likely representing a practical upper limit.

Table 5-1: Output voltage vs. sensor state for circuit of Figure 5-13.

Sensor Status			Output Voltage (V_O)
S1	S2	S3	
OFF	OFF	OFF	5.00
OFF	OFF	ON	3.31
OFF	ON	OFF	2.50
OFF	ON	ON	1.99
ON	OFF	OFF	1.67
ON	OFF	ON	1.42
ON	ON	OFF	1.25
ON	ON	ON	1.11

Another digital-to-analog encoding scheme can be used when one only needs to know the identity of the first device activated in a string of digital-output sensors. This circuit is shown in Figure 5-14. One application in which this circuit could be used is in liquid level measurement. A float containing a magnet is set up so that it moves past a line of sensors, its exact position dependent on the level of some fluid, such as water. The rightmost sensor actuated represents the level of the float, and consequently the level of the liquid. Because the rightmost sensor actuated shorts the resistive chain to ground, it is irrelevant if any sensors to the left are also activated; the output voltage is determined solely by the position of the rightmost sensor.

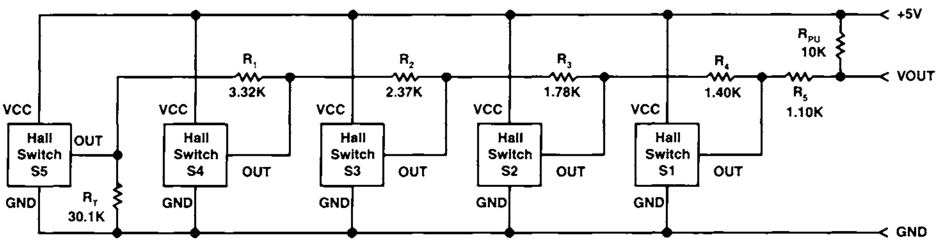

Figure 5-14: Encoding the rightmost sensor of a string as a voltage.

Note that in this circuit, the values of the resistors are not equal; they were chosen so that the output voltage would decrease in roughly equal steps. Equal-valued resistors would also result in a monotonically decreasing output voltage, but with different step sizes between output voltage levels. Table 5-2 shows the output voltage as a function of rightmost activated sensor. As in the prior analog-to-digital encoding example, we are ignoring the effects of resistor tolerance and sensor output saturation voltage. Nevertheless, with proper resistor selection it is possible to chain a significant number of sensors into this type of a "thermometer" sensing arrangement before it becomes difficult to reliably identify which sensor in the chain is activated.

Table 5-2: Output voltage vs. sensor activated in Figure 5-15.

Rightmost Sensor Activated	Output Voltage V_O
S1	0.50
S2	1.00
S3	1.50
S4	2.00
S5	2.50
None	4.00

5.15 Mini-Networks

Another way to reduce wiring is to network the sensors together by pairing each sensor with a microcontroller. In the past few years, both the cost and package sizes of microcontrollers have fallen dramatically. Microcontrollers are now available for less than $1 US in moderate quantities and are available in SOIC-8 and smaller packages. As an example, Microchip's PIC10F200 provides an 8-bit processor core, 256 words of user-programmable ROM, and 16 bytes of RAM in a 6-pin SOT-23 package, making it actually smaller than many contemporary Hall-effect sensors. The continuing downward trends in both size and cost make microcontrollers increasingly attractive options for adding inexpensive intelligence to one's designs.

Figure 5-15 shows the general organization of a simple sensor network. Each Hall-effect sensor is associated with a microcontroller that can monitor the sensor's output and provide an interface to a common data bus. Each sensor-microcontroller combination is often referred to as a network node. The network may be controlled by a central "bus master," which coordinates data transfers or, alternatively, each sensor node may have the ability to initiate and coordinate data transfers on its own—this latter arrangement is often called a *multi-master* system.

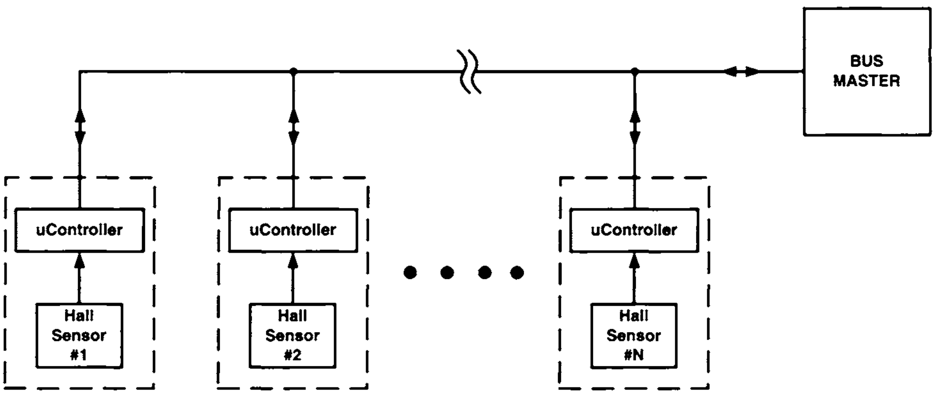

Figure 5-15: Simple sensor network.

When one thinks of a sensor network, the first options that may come to mind are Ethernet or industrial sensor busses such as Profibus or Fieldbus. While these networks have their advantages, they can require a substantial amount of processing power and software to be effectively realized. For the purposes of wiring reduction, much simpler protocols can often be used, especially if data transmission rates are relatively low and not over long physical distances. Examples of where such a network might be used are in a high-performance copying machine or automobile. Both systems must monitor numerous sensors and do so economically.

There are many options available for inexpensive network protocols. Some of the major factors in selecting a network protocol are the number of sensors that need to be monitored, the physical separation of sensors, and the speed (bandwidth). Some options available to the designer are:

SPI – Developed by Motorola, this simple-to-implement synchronous serial bus offers bandwidths up to several megabits/second over short distances, and is primarily intended to facilitate communications between chips on the same circuit card. Many low-end microcontrollers have built-in hardware to support this standard, but at lower speeds (a few tens of kilobits/second) SPI functionality can often be implemented in software.

I^2C – Developed by Philips as a means of allowing chips on the same board to communicate, I^2C provides bandwidths up to around 400 kbits/sec. I^2C provides more functionality than SPI, and many low-end microcontrollers also provide support hardware on-chip.

CAN bus – A relatively sophisticated bus developed by Bosch for automotive applications. It offers high bandwidths and many advanced features that make it suitable for a wide range of applications. A few microcontrollers are now available that offer CAN bus support, but at this time they still tend to be relatively expensive when compared to Hall-effect sensor ICs.

Linbus – This bus was recently introduced as a low-cost, low-bandwidth (20 kilobits/second) way of monitoring sensors and controlling actuators in automobiles. A few microcontrollers are starting to come on the market with dedicated hardware support for Linbus. It is also possible to implement Linbus protocols completely in software with a small amount of external interfacing circuitry. This is definitely a standard to watch, as its increasing acceptance in the automotive industry will result in lots of inexpensive compatible chips and support.

Note that using a microcontroller as the heart of a sensor network node can provide many capabilities. First, if an analog-to-digital converter is available on the microcontroller, the network node can monitor the output of a linear-output sensor in addition to digital output devices. Because microcontrollers typically provide several I/O pins, it may also be possible for a single node to monitor several sensors—for example, a linear Hall-effect sensor may be used as a position sensor while a thermistor provides a temperature reading. Finally, the microcontroller's processing power can be used for local data reduction, reducing the amount of bandwidth needed for data transmission as well as the amount of processing power needed by a central monitor.

5.16 Voltage Regulation and Power Management

While most digital-output Hall-effect sensors contain internal voltage regulators that allow them to operate over an extended range of power supply voltages, most linear-output sensors do not, and require a regulated power supply for proper operation. Additionally, if you plan on using microcontrollers, discrete logic, or many other kinds of support circuitry, you may find that you need a source of regulated power internal to your sensor assembly, to power these other devices.

There are two fundamental approaches to providing a regulated power supply; design it from discrete components, or buy an integrated regulator in the form of an off-the-shelf integrated circuit. Unless you have a really unusual set of design requirements, the do-it-yourself approach is not recommended, as it takes a lot of discrete circuitry and design acumen to provide performance comparable to that of even the least expensive integrated regulator. For this reason we will focus on the characteristics and use of integrated devices.

Among the most popular voltage regulators are the members of the LM78xx family of linear regulators. "Linear" in the context of a voltage regulator means that the device does not contain any switching elements, and operates in a continuous manner, with no oscillators or clocks. One especially useful type of linear regulators is the 78L05 and its related variants, which provide a regulated 5V output voltage with a peak current output of 100 mA (for many versions). This device is available in both TO-92 3-leaded transistor and SOIC-8 surface-mountable packages. To use one of these parts, you provide an input voltage ranging from 8–24V to one pin, tie another pin to system ground, and 5V (± a specified tolerance) appears on the third pin. To reduce levels of noise and provide some power-supply decoupling, it is also advisable to connect small capacitors between the input from both the inputs and outputs to ground, as shown in Figure 5-16.

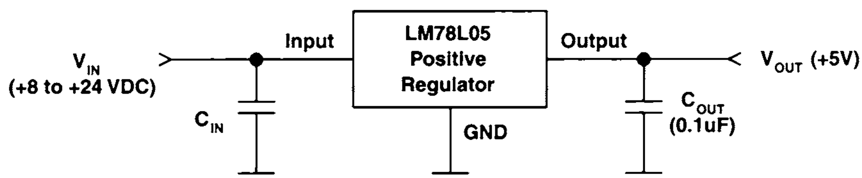

Figure 5-16: LM78L05 5V linear voltage regulator circuit.

Although these devices are about as easy to use as was just described, they have certain characteristics that one must understand in order to employ them effectively. The first is that they have definite limits on the amount of power they can safely dissipate. If one draws 100 mA from the 5V output of one of these devices while running from a 24V input, the regulator will dissipate nearly 2W of power, in the form of heat. The power dissipation for a linear regulator is approximately equal to $(V_{IN} - V_{OUT}) \times I_{OUT}$, neglecting a small amount of power consumed for internal "housekeeping" functions. This means that large differences in the input and output voltages will severely limit the amount of current one can draw from the device. For regulators in smaller packages (TO-92, SOIC-8) it can be easy to exceed the device's maximum power ratings. Maximum power ratings for integrated regulators also decrease with increasing ambient temperature. Fortunately, the manufacturers of these devices will usually provide information on derating the component for operation at higher temperatures.

The second important characteristic of these devices is that there is a "drop-out" voltage for the input below which regulation is lost, and the output is not guaranteed to be 5V (for a 5V regulator). For 78Lxx devices, this dropout voltage is usually specified as 2–3V above the output voltage; a 5V regulator would begin to "drop out" at around 7–8V of input voltage. When operation from lower power supplies is required, one can often use what is known as a low-drop-out regulator often referred to simply as an LDO. One family of such devices is the TL750 series produced by Texas Instruments. These devices can maintain regulation when operating with dropout voltages of less than a volt. One important consideration when using many types of LDOs is that the capacitive load on the device's output is a critical factor in maintaining stability. Most LDOs have very specific requirements for the amount of capacitance you bypass their outputs with, and even the type of capacitor you use! If you put either too much or too little, or even the wrong kind of capacitor on a given device's output, it may begin oscillating. Fortunately, manufacturers of LDOs tend to provide very specific information about acceptable values and types of bypassing capacitors in their devices' datasheets.

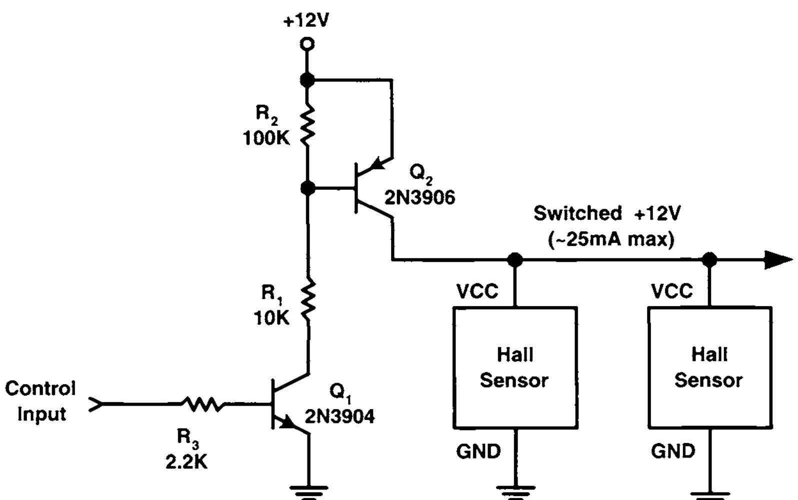

Figure 5-17: Switching power to Hall-effect devices.

Another power-related issue is that of power management. Because most Hall-effect sensors consume several milliamperes of operating current, they can be difficult to use in battery-operated systems that must operate for an extended length of time. One solution to this problem is to shut them off when not in use. The circuit of Figure 5-17 provides one way to do this. When the control input is at 0V, transistor Q_1 and, consequently, transistor Q_2 are both off. When 5V is applied to the control input, both transistors are turned on. Because Q_2 is a PNP device, it can go into saturation, and introduce a relatively small (<100 mV) voltage drop when sourcing moderate amounts of current (a few tens of milliamperes in this case). While this small voltage drop should pose little problem when power-switching digital Hall-effect sensors in this manner, it should be considered when attempting to use this technique to power-switch a ratiometric linear-output device, where the small drop in supply voltage will be reflected as changes in zero-flux offset and sensitivity.