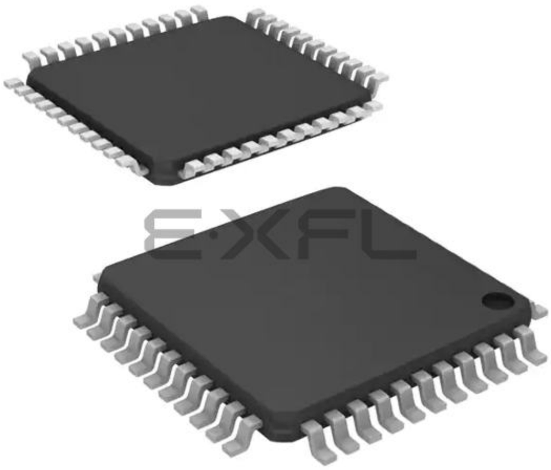

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Applications of "Embedded - Microcontrollers"

Details	
Product Status	Active
Core Processor	PIC
Core Size	8-Bit
Speed	20MHz
Connectivity	I²C, SPI, UART/USART
Peripherals	Brown-out Detect/Reset, POR, PWM, WDT
Number of I/O	33
Program Memory Size	7KB (4K x 14)
Program Memory Type	OTP
EEPROM Size	-
RAM Size	192 x 8
Voltage - Supply (Vcc/Vdd)	4V ~ 5.5V
Data Converters	-
Oscillator Type	External
Operating Temperature	0°C ~ 70°C (TA)
Mounting Type	Surface Mount
Package / Case	44-TQFP
Supplier Device Package	44-TQFP (10x10)
Purchase URL	https://www.e-xfl.com/product-detail/microchip-technology/pic16c65b-20-pt

CHAPTER 2
PIC® Microcontroller Low Power Tips ‘n Tricks

Table Of Contents

TIPS ‘N TRICKS INTRODUCTION

Microchip continues to provide innovative products that are smaller, faster, easier to use and more reliable. The Flash-based PIC® microcontrollers (MCUs) are used in an wide range of everyday products, from smoke detectors, hospital ID tags and pet containment systems, to industrial, automotive and medical products.

PIC MCUs featuring nanoWatt technology implement a variety of important features which have become standard in PIC microcontrollers. Since the release of nanoWatt technology, changes in MCU process technology and improvements in performance have resulted in new requirements for lower power. PIC MCUs with nanoWatt eXtreme Low Power (nanoWatt XLP™) improve upon the original nanoWatt technology by dramatically reducing static power consumption and providing new flexibility for dynamic power management.

The following series of Tips n’ Tricks can be applied to many applications to make the most of PIC MCU nanoWatt and nanoWatt XLP devices.

GENERAL LOW POWER TIPS ‘N TRICKS

The following tips can be used with all PIC MCUs to reduce the power consumption of almost any application.

TIP #5 Measuring RPM Using an Encoder

Revolutions Per Minute (RPM), or how fast something turns, can be sensed in a variety of ways. Two of the most common sensors used to determine RPM are optical encoders and Hall effect sensors. Optical encoders detect the presence of light shining through a slotted wheel mounted to a turning shaft (see Figure 5-1.) As the shaft turns, the slots in the wheel pass by the eye of the optical encoder. Typically, an infrared source on the other side of the wheel emits light that is seen by the optical encoder through slots in the wheel. Hall effect sensors work by sensing the position of the magnets in an electric motor, or by sensing a permanent magnet mounted to a rotating object (see Figure 5-2). These sensors output one or more pulses per revolution (depending on the sensor).

Figure 5-1: Optical Encoder

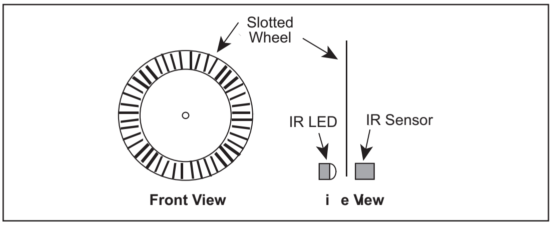
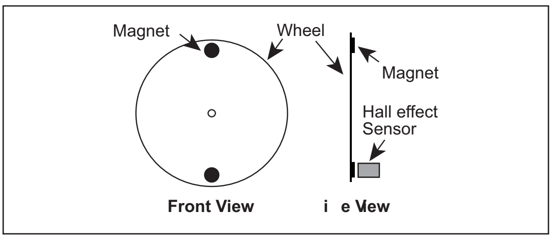


Figure 5-2: Hall Effect Sensor



In Figure 5-3 and Figure 5-4, the waveform is high when light is passing through a slot in the encoder wheel and shining on the optical sensor. In the case of a Hall effect sensor, the high corresponds to the time that the magnet is in front of the sensor. These figures show the difference in the waveforms for varying RPMs. Notice that as RPM increases, the period (T) and pulse width (W) becomes smaller. Both period and pulse width are proportional to RPM. However, since the period is the greater of the two intervals, it is good practice to measure the period so that the RPM reading from the sensor will have the best resolution. See Tip #1 for measuring period. The technique for measuring period with averaging described in Tip #2 is useful for measuring high RPMs.

Figure 5-3: Low RPM

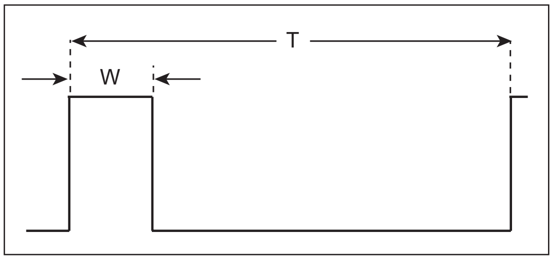
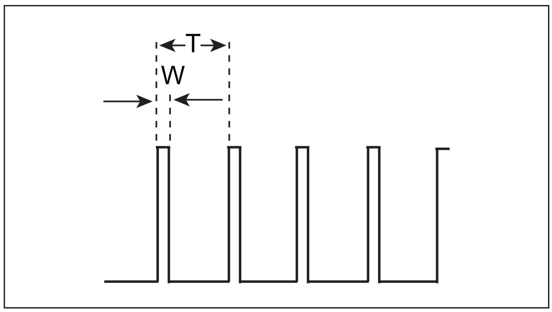


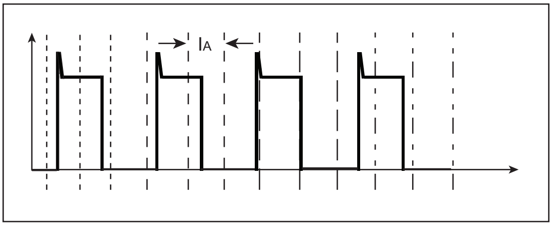
Figure 5-4: High RPM



TIP #12 Repetitive Phase Shifted Sampling

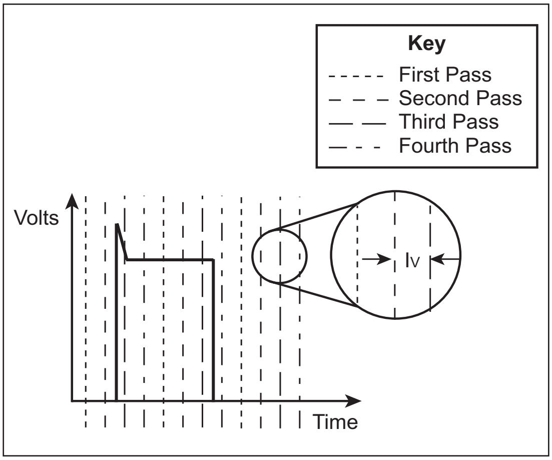
Repetitive phase shifted sampling is a technique to artificially increase the sampling rate of an A/D converter when sampling waveforms that are both periodic and constant from period to period. The technique works by capturing regularly spaced samples of the waveform from the start to finish of the waveform's period. Sampling of the next waveform is then performed in the same manner, except that the start of the sample sequence is delayed a percentage of the sampling period. Subsequent waveforms are also sampled, with each sample sequence slightly delayed from the last, until the delayed start of the sample sequence is equal to one sample period. Interleaving the sample sets then produces a sample set of the waveform at a higher sample rate. Figure 12-1 shows an example of a high frequency waveform.

Figure 12-1: High Frequency Periodic Waveform



As indicated in the key, the finely dotted lines show where the A/D readings are taken during the first period of the waveform. The medium sized dashed lines show when the A/D readings are taken during the second period, and so on. Figure 12-2 shows these readings transposed onto one period.

Figure 12-2: Transposed Waveform



The CCP module is configured in Compare Special Event Trigger mode to accomplish this task. The phase shift is implemented by picking values of CCPRxL and CCPRxH that are not synchronous with the period of the sampling waveform. For instance, if the period of a waveform is 100 μs, then sampling at a rate of once every 22 μs will give the following set of sample times over 11 periods (all values in μs).

1st	2nd	3rd	4th	5th	6th	7th	8th	9th	10th	11th
0	10	20	8	18	6	16	4	14	2	12
22	32	42	30	40	28	38	26	36	24	34
44	54	64	52	62	50	60	48	58	46	56
66	76	86	74	84	72	82	70	80	68	78
88	98		96		94		92		90	

When these numbers are placed in sequential order, they reveal a virtual sampling interval (I_V) of 2 μs from 0 μs to 100 μs, although the actual sampling interval (I_A) is 22 μs.

TIP #2 Faster Code for Detecting Change

When using a comparator to monitor a sensor, it is often just as important to know when a change occurs as it is to know what the change is. To detect a change in the output of a comparator, the traditional method has been to store a copy of the output and periodically compare the held value to the actual output to determine the change. An example of this type of routine is shown below.

Example 2-1

```
Test
   MOVF  hold,w     ;get old Cout
   XORWF CMCON,w    ;compare to new Cout
   ANDLW COUTMASK
   BTFSC STATUS,Z
   RETLW 0          ;if = return "no change"
   MOVF  CMCON,w    ;if not =, get new Cout
   ANDLW COUTMASK   ;remove all other bits
   MOVWF hold       ;store in holding var.
   IORLW CHNGBIT    ;add change flag
   RETURN
```

This routine requires 5 instructions for each test, 9 instructions if a change occurs, and 1 RAM location for storage of the old output state.

A faster method for microcontrollers with a single comparator is to use the comparator interrupt flag to determine when a change has occurred.

Example 2-2

```
Test
   BTFSS PIR1,CMIF  ;test comparator flag
   RETLW 0          ;if clear, return a 0
   BTFSS CMCON,COUT ;test Cout
   RETLW CHNGBIT    ;if clear return
                    ;CHNGFLAG
   RETLW COUTMASK + CHNGBIT;if set,
                    ;return both
```

This routine requires 2 instructions for each test, 3 instructions if a change occurs, and no RAM storage.

If the interrupt flag can not be used, or if two comparators share an interrupt flag, an alternate method that uses the comparator output polarity bit can be used.

Example 2-3

```
Test
   BTFSS CMCON,COUT ;test Cout
   RETLW 0          ;if clear, return 0
   MOVLW CINVBIT    ;if set, invert Cout
   XORWF CMCON,f    ;forces Cout to 0
   BTFSS CMCON,CINV ;test Cout polarity
   RETLW CHNGFLAG   ;if clear, return
                    ;CHNGFLAG
   RETLW COUTMASK + CHNGFLAG;if set,
                    ;return both
```

This routine requires 2 instructions for each test, 5 instructions if a change occurs, and no GPR storage.

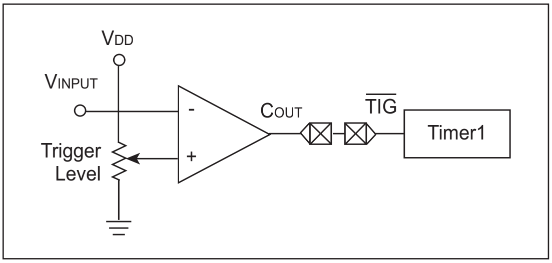
TIP #4 Pulse Width Measurement

To measure the high or low pulse width of an incoming analog signal, the comparator can be combined with Timer1 and the Timer1 Gate input option (see Figure 4-1). Timer1 Gate acts as a count enable for Timer1. If the input is low, Timer1 will count. If the $\overline{T1G}$ input is high, Timer1 does not count. Combining $\overline{T1G}$ with the comparator allows the designer to measure the time between a high-to-low output change and a low-to-high output change.

To make a measurement between a low-to-high and a high-to-low transition, the only change required is to set the CINV bit in the comparator CMCON register which inverts the comparator output.

Because the output of the comparator can change asynchronously with the Timer1 clock, only comparators with the ability to synchronize their output with the Timer1 clock should be used and their C2SYNC bits should be set.

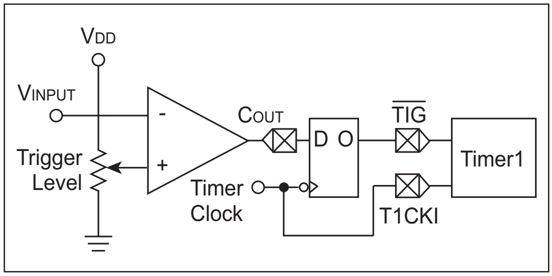
Figure 4-1: Comparator with Timer1 and $\overline{T1G}$



If the on-chip comparator does not have the ability to synchronize its output to the Timer1 clock, the output can be synchronized externally using a discrete D flip-flop (see Figure 4-2).

> **Note:** The flip-flop must be falling edge triggered to prevent a race condition.

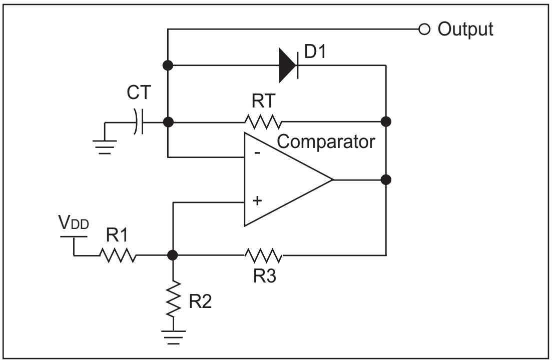
Figure 4-2: Externally Synchronized Comparator



TIP #9 Multi-Vibrator (Ramp Wave Output)

A multi-vibrator (ramp wave output) is an oscillator designed around a voltage comparator or operational amplifier that produces an asymmetrical output waveform (see Figure 9-1). Resistors R1 through R3 form a hysteresis feedback path from the output to the non-inverting input. Resistor RT, diode D1 and capacitor CT form a time delay network between the output and the inverting input. At the start of the cycle, CT is discharged holding the non-inverting input at ground, forcing the output high. A high output forces the non-inverting input to the high threshold voltage (see Tip #3) and charges CT through RT. When the voltage across CT reaches the high threshold voltage, the output is forced low. A low output drops the non-inverting input to the low threshold voltage and discharges CT through D1. Because the dynamic on resistance of the diode is significantly lower than RT, the discharge of CT is small when compared to the charge time, and the resulting waveform across CT is a pseudo ramp function with a ramping charge phase and a short-sharp discharge phase.

Figure 9-1: Ramp Waveform Multi-Vibrator



To design this multi-vibrator, first design the hysteresis feedback path using the procedure in Tip #3. Remember that the peak-to-peak amplitude of the ramp wave will be determined by the hysteresis limits. Also, be careful to choose threshold voltages (V_{TH} and V_{TL}) that are evenly spaced within the common mode range of the comparator.

Then use V_{TH} and V_{TL} to calculate values for RT and CT that will result in the desired oscillation frequency F_{OSC}. Equation 9-1 defines the relationship between RT, CT, V_{TH}, V_{TL} and F_{OSC}.

Equation 9-1

$$F_{OSC} = \frac{1}{RT * CT * \ln(V_{TH}/V_{TL})}$$

This assumes that the dynamic on resistance of D1 is much less than RT.

Example:

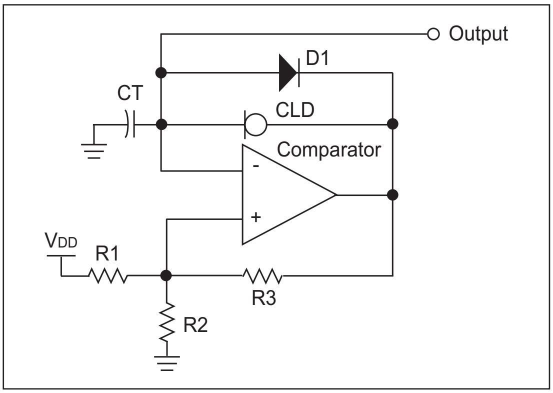
- V_{DD} = 5V, V_{TH} = 1.666V and V_{TH} = 3.333V
- R1, R2 and R3 = 10k
- R_T = 15k, C_T = .1 μF for a F_{OSC} = 906 Hz

> **Note:** Replacing R_T with a current limiting diode will significantly improve the linearity of the ramp wave form. Using the example shown above, a CCL1000 (1 mA Central Semiconductor CLD), will produce a very linear 6 kHz output (see Equation 9-2).

Equation 9-2

$$F_{OSC} = \frac{I_{CLD}}{C\ (V_{TH} - V_{TL})}$$

Figure 9-2: Alternate Ramp Waveform Multi-Vibrator Using a CLD



TIP #11 PWM Generator

This tip shows how the multi-vibrator (ramp wave) can be used to generate a voltage controlled PWM signal. The ramp wave multi-vibrator operates as described in Tip #9, generating a positive going ramp wave. A second comparator compares the instantaneous voltage of the ramp wave with the incoming voltage to generate the PWM output (see Figure 11-2).

When the ramp starts, it is below the input voltage, and the output of the second comparator is pulled high starting the PWM pulse. The output remains high until the ramp wave voltage exceeds the input, then the output of the second comparator goes low ending the PWM pulse. The output of the second comparator remains low for the remainder of the ramp waveform. When the ramp waveform returns to zero at the start of the next cycle, the second comparator output goes high again and the cycle starts over.

Figure 11-1: PWM Wave Forms

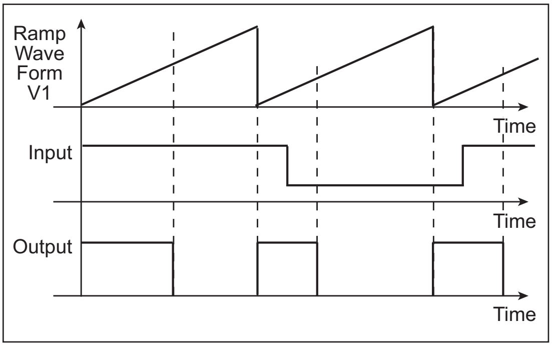
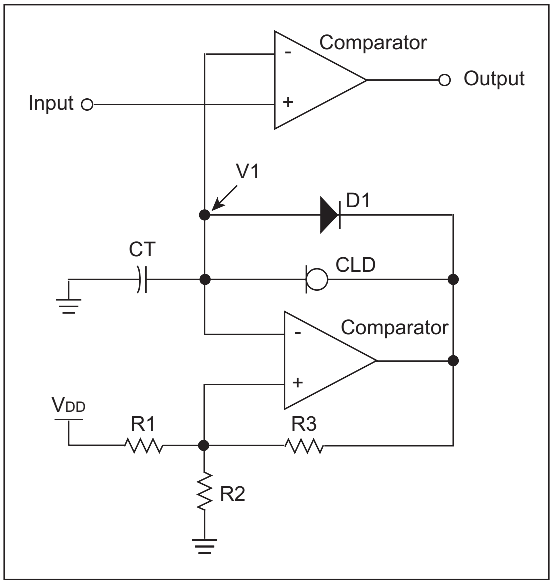


Figure 11-2: PWM Circuit



To design a PWM generator, start with the design of a ramp wave multi-vibrator using the design procedure from Tip #9. Choose high and low threshold voltages for the multi-vibrators hysteresis feedback that are slightly above and below the desired PWM control voltages.

> **Note:** The PWM control voltage will produce a 0% duty cycle for inputs below the low threshold of the multi-vibrator. A control voltage greater than the high threshold voltage will produce a 100% duty cycle output.

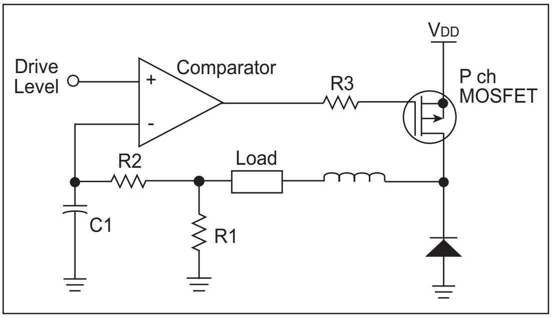
Using the example values from Tip #9 will result in a minimum pulse width at an input voltage of 1.7V and a maximum at an input of 3.2V.

TIP #13 PWM High-Current Driver

This tip combines a comparator with a MOSFET transistor and an inductor to create a switch mode high-current driver circuit. (See Figure 13-1).

The operation of the circuit begins with the MOSFET off and no current flowing in the inductor and load. With the sense voltage across R1 equal to zero and a DC voltage present at the drive level input, the output of the comparator goes low. The low output turns on the MOSFET and a ramping current builds through the MOSFET, inductor, load and R1.

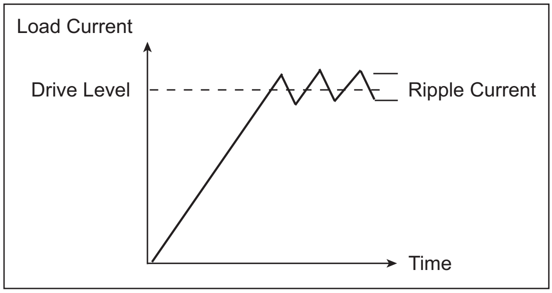
Figure 13-1: High Current Driver



When the current ramps high enough to generate a voltage across R1 equal to the drive level, the comparator output goes high turning off the MOSFET. The voltage at the junction of the MOSFET and the inductor then drops until D1 forward biases. The current continues ramping down from its peak level toward zero. When the voltage across the sense resistor R1 drops below the drive level, the comparator output goes low, the MOSFET turns on, and the cycle starts over.

R2 and C1 form a time delay network that limits the switching speed of the driver and causes it to slightly overshoot and undershoot the drive level when operating. The limit is necessary to keep the switching speed low, so the MOSFET switches efficiently. If R2 and C1 were not present, the system would run at a speed set by the comparator propagation delay and the switching speed of the MOSFET. At that speed, the switching time of the MOSFET would be a significant portion of the switching time and the switching efficiency of the MOSFET would be too low.

Figure 13-1: Current Through the Load



To design a PWM high current driver, first determine a switching speed (F_{SWX}) that is appropriate for the system. Next, choose a MOSFET and D1 capable of handling the load current requirements. Then choose values for R2 and C1 using Equation 13-1.

Equation 13-1

$$F_{SWX} = \frac{2}{R2 * C1}$$

Next determine the maximum ripple current that the load will tolerate, and calculate the required inductance value for L1 using Equation 13-2.

Equation 13-2

$$L = \frac{V_{DD} - V_{LOAD}}{I_{RIPPLE} * F_{SWX} * 2}$$

Finally, choose a value for R1 that will produce a feedback ripple voltage of 100 mV for the maximum ripple current I_{RIPPLE}.

Example:

- F_{SWX} = 10 kHz, R2 = 22k, C1 = .01 μF
- I_{RIPPLE} = 100 mA, V_{DD} = 12V, V_L = 3.5V
- L = 4.25 mH

TIP #19 Logic: XOR/XNOR Gate

This tip shows the use of the comparator to implement an XOR gate and its complement the XNOR gate.

The operation is best described in three sections:

1. Both A and B inputs are low
   With both inputs low, the inverting input is held at .7V and the non-inverting is held at ground. This combination results in a low output.
2. Both A and B inputs are high
   With both inputs high, the inverting input is pulled up to VDD and the non-inverting input is equal to 2/3 VDD (the average of VDD inputs and GND). This combination also results in a low output.
3. Input A or B is high
   With one input high and one low, The inverting input is held at .7V and the non-inverting input is equal to 1/3 VDD (the average of a VDD input and GND). This combination results in a high output.

> **Note:** Typical propagation delay for the circuit is 250-350 ns using the typical on-chip comparator peripheral of a microcontroller. Delay measurements were made with 10k resistance values.

While the circuit is fairly simple, there are a few requirements for correct operation:

1. The inputs A and B must drive from ground to VDD for the circuit to operate properly.
2. All resistances on the both inputs react with the input capacitance of the comparator, so the speed of the gate will be affected by the source resistance of A and B, as well as, the size of resistors R1, R2, R3 and R4.
3. Resistor R1, R2 and R3 must be equal.
4. Resistor R4 must be small enough to produce a 1.0V, or lower, voltage drop across D1 and D2.

Figure 19-1: XOR Gate

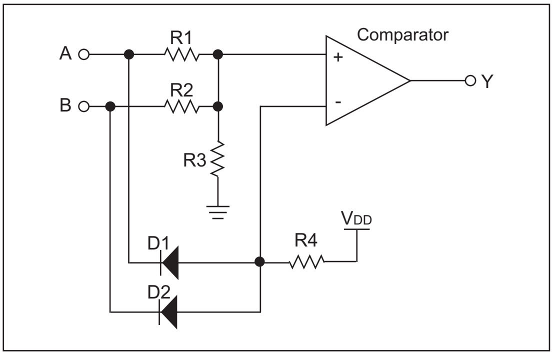
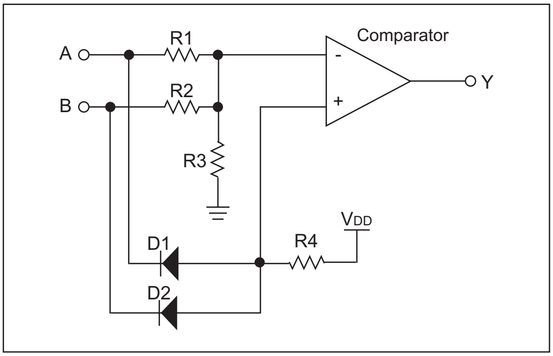


Figure 19-2: XNOR Gate



Example:

- D1 = D2, = 1N4148
- R4 = 10k, R1 = R2 = R3 = 5.1k

Figure 3-1: 4 and 5 Wire Stepper Motors

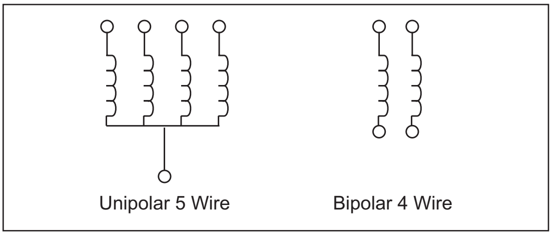


Figure 3-2: 6 and 8 Wire Stepper Motors

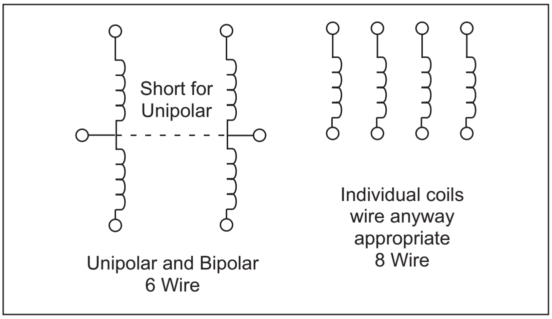


Figure 3-3: Unipolar Motor (4 Low Side Switches)

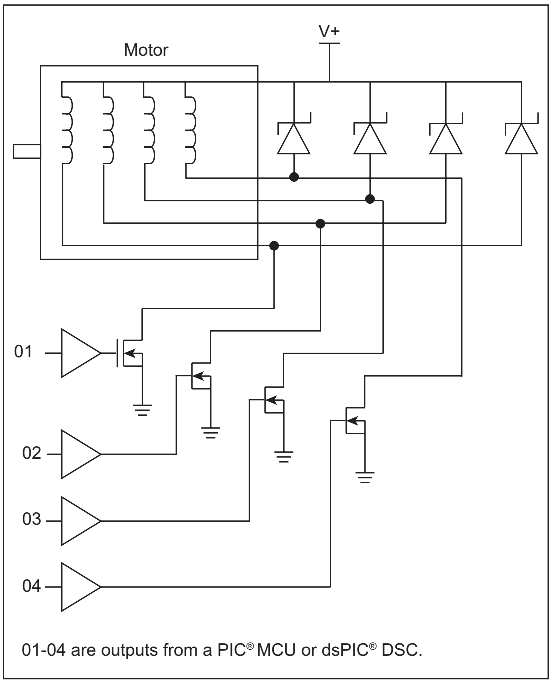
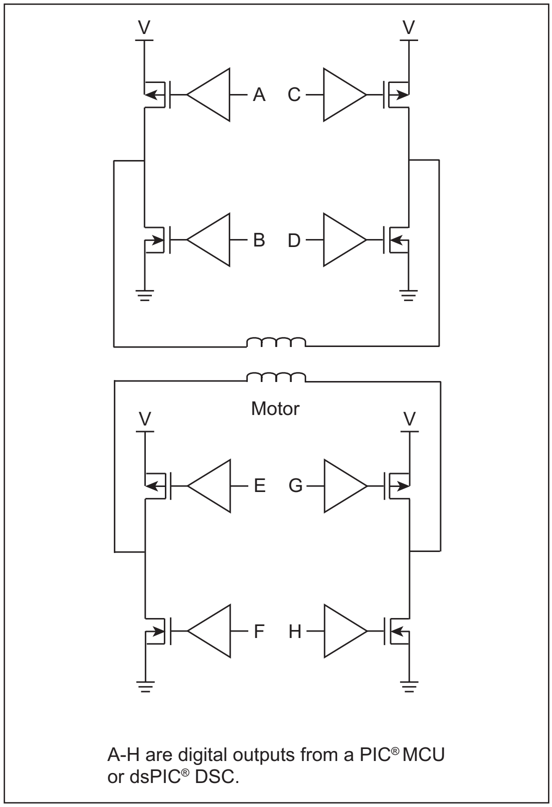


Figure 3-4: Bipolar Motor (4 Half-Bridges)



TIP #5 Writing a PWM Value to the the CCP Registers With a Mid-Range PIC® Microcontroller

The two PWM LSb's are located in the CCPCON register of the CCP. This can make changing the PWM period frustrating for a developer. Example 5-1 through Example 5-3 show three macros written for the mid-range product family that can be used to set the PWM period. The first macro takes a 16-bit value and uses the 10 MSb's to set the PWM period. The second macro takes a 16-bit value and uses the 10 LSb's to set the PWM period. The last macro takes 8 bits and sets the PWM period. This assumes that the CCP is configured for no more than 8 bits.

Example 5-1: Left Justified 16-Bit Macro

```
pwm_tmp    equ xxx   ;this variable must be
                     ;allocated someplace
setPeriod  macro a   ;a is 2 SFR's in
                     ;Low:High arrangement
                     ;the 10 MSb's are the
                     ;desired PWM value
 RRF       a,w       ;This macro will
                     ;change w
 MOVWF     pwm_tmp
 RRF       pwm_tmp,w
 ANDLW     0x30
 IORLW     0x0F
 MOVWF     CCP1CON
 MOVF      a+1,w
 MOVWF     CCPR1L
```

Example 5-2: Right Justified 16-Bit Macro

```
pwm_tmp    equ xxx   ;this variable must be
                     ;allocated someplace
setPeriod  macro a   ;a is 2 bytes in
                     ;Low:High arrangement
                     ;the 10 LSb's are the
                     ;desired PWM value
 SWAPF     a,w       ;This macro will
                     ;change w
 ANDLW     0x30
 IORLW     0x0F
 MOVWF     CCP1CON
 RLF       a,w
 IORLW     0x0F
 MOVWF     pwm_tmp
 RRF       pwm_tmp,f
 RRF       pwm_tmp,w
 MOVWF     CCPR1L
```

Example 5-3: 8-Bit Macro

```
pwm_tmp    equ xxx   ;this variable must be
                     ;allocated someplace
setPeriod  macro a   ;a is 1 SFR
 SWAPF     a,w       ;This macro will
                     ;change w
 ANDLW     0x30
 IORLW     0x0F
 MOVWF     CCP1CON
 RRF       a,w
 MOVWF     pwm_tmp
 RRF       pwm_tmp,w
 MOVWF     CCPR1L
```


TIP #6 Current Sensing

The torque of an electric motor can be monitored and controlled by keeping track of the current flowing through the motor. Torque is directly proportional to the current. Current can be sensed by measuring the voltage drop through a known value resistor or by measuring the magnetic field strength of a known value inductor. Current is generally sensed at one of two places, the supply side of the drive circuit (high side current sense) or the sink side of the drive circuit (low side current sense). Low side sensing is much simpler but the motor will no longer be grounded, causing a safety issue in some applications. High side current sensing generally requires a differential amplifier with a common mode voltage range within the voltage of the supply.

Figure 6-1: Resistive High Side Current Sensing

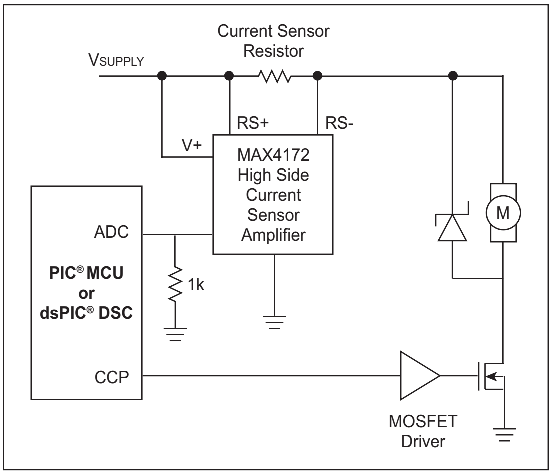
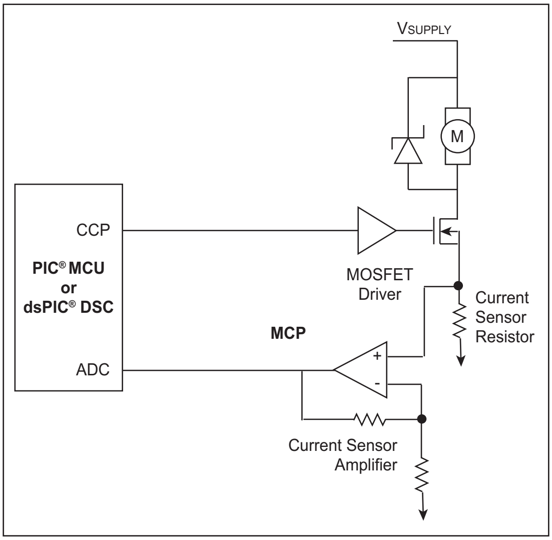


Figure 6-2: Resistive Low Side Current Sensing



Current measurement can also be accomplished using a Hall effect sensor to measure the magnetic field surrounding a current carrying wire. Naturally, this Hall effect sensor can be located on the high side or the low side of the load. The actual location of the sensor does not matter because the sensor does not rely upon the voltage on the wire. This is a non-intrusive method that can be used to measure motor current.

Figure 6-3: Magnetic Current Sensing

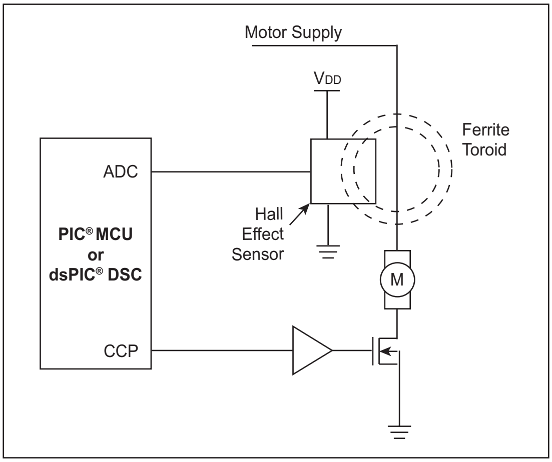
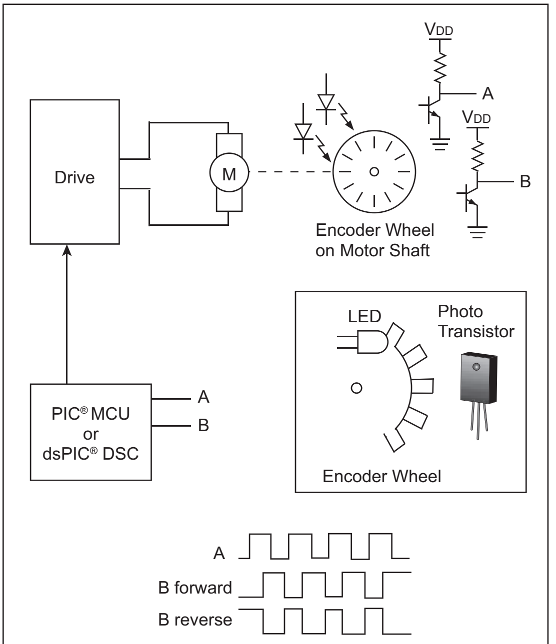


Figure 7-2: Optical Speed/Direction/Position Sensing



Note: Frequency of one signal provides RPM of motor. Pulse count provides motor position. A-B phase provides motor direction.

Quadrature sensing can easily be accomplished in software, but there is generally an upper limit to the RPM. By using a few gates, the sensing can be done partially in hardware and partially in software. The new PIC18FXX31 and dsPIC 16-bit Digital Signal Controller families include an encoder interface that allows MUCH higher RPM motors to be measured with an excellent degree of accuracy.

Older Methods of Motor Sensing

Resolvers and analog tachometers are two older technologies for motor position/velocity sensing. An analog tachometer is simply an electric generator with a linear output over a specified range of RPM's. By knowing the output characteristics, the RPM can be known by simply measuring the voltage across the tachometer terminals.

A resolver is a pair of coils that are excited by an external AC signal. The two coils are at 90° to each other so they pick up the AC signal at different strengths, depending on their orientation. The result is a sine or cosine output related to the angle of the resolver in reference to the AC signal. Inverse cosine/sine will produce the angle of the sensor. This type of sensor can be very accurate and is still used where absolute position must be known.

NOTES:

TIP #1 Typical Ordering Considerations and Procedures for Custom Liquid Displays

1. Consider what useful information needs to be displayed on the custom LCD and the combination of alphanumeric and custom icons that will be necessary.
2. Understand the environment in which the LCD will be required to operate. Operating voltage and temperature can heavily influence the contrast of the LCD and potentially limit the type of LCD that can be used.
3. Determine the number of segments necessary to achieve the desired display on the LCD and reference the PIC Microcontroller LCD matrix for the appropriate LCD PIC microcontroller.
4. Create a sketch/mechanical print and written description of the custom LCD and understand the pinout of the LCD. (Pinout definition is best left to the glass manufacturer due to the constraints of routing the common and segment electrodes in two dimensions.)
5. Send the proposed LCD sketch and description for a written quotation to at least 3 vendors to determine pricing, scheduling and quality concerns.
   a) Take into account total NRE cost, price per unit, as well as any setup fees.
   b) Allow a minimum of two weeks for formal mechanical drawings and pin assignments and revised counter drawings.
6. Request a minimal initial prototype LCD build to ensure proper LCD development and ensure proper functionality within the target application.
   a) Allow typically 4-6 weeks for initial LCD prototype delivery upon final approval of mechanical drawings and pin assignments.
7. Upon receipt of prototype LCD, confirm functionality before giving final approval and beginning production of LCD.

Note: Be sure to maintain good records by keeping copies of all materials transferred between both parties, such as initial sketches, drawings, pinouts, etc.

TIP #2 LCD PIC® MCU Segment/ Pixel Table

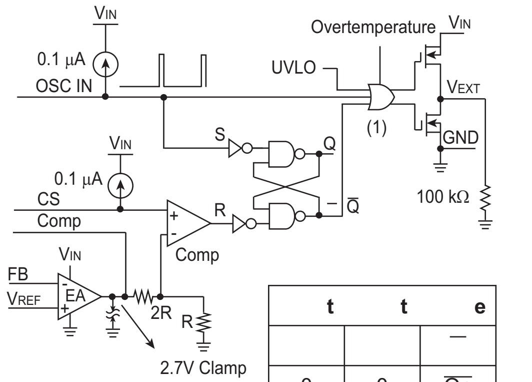
Table 2-1: Segment Matrix Table

Multiplex Commons	Maximum Number of Segments/Pixels					Bias
	PIC16F913/ 916	PIC16F914/ 917	PIC16F946	PIC18F6X90 (PIC18F6XJ90)	PIC18F8X90 (PIC18F8XJ90)	
Static (COM0)	15	24	42	32/ (33)	48	Static
1/2 (COM1: COM0)	30	48	84	64/ (66)	96	1/2 or 1/3
1/3 (COM2: COM0)	45	72	126	96/ (99)	144	1/2 or 1/3
1/4 (COM3: COM0)	60	96	168	128/ (132)	192	1/3

This Segment Matrix table shows that Microchip's 80-pin LCD devices can drive up to 4 commons and 48 segments (192 pixels), 64-pin devices can drive up to 33 segments (132 pixels), 40/44 pin devices can drive up to 24 segments (96 pixels) and 28-pin devices can drive 15 segments (60 segments).

TIP #6 Current Limiting Using the MCP1630

Figure 6-1: MCP1630 High-Speed PWM



t	t	e
		—
0	0	$\overline{Qn}$
0	1	1
1	0	0
1	1	1

Note 1: During overtemperature, V_{EXT} driver is high-impedance.

The block diagram for the MCP1630 high-speed PWM driver is shown in Figure 6-1. One of the features of the MCP1630 is the ability to perform current limiting. As shown in the bottom left corner of the diagram, the output of the Error Amplifier (EA) is limited by a 2.7V clamp. Therefore, regardless of the actual error, the input to the negative terminal of the comparator (labeled Comp) is limited to 2.7V ÷ 3 or 0.9V.

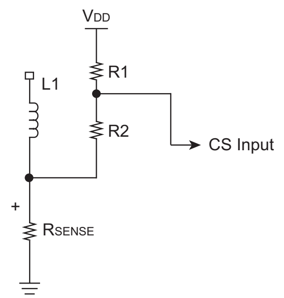
It is possible to implement the current limiting by using a single sense resistor. In this case, the maximum current would be given by Equation 6-1.

Equation 6-1

$$I_{MAX} = (0.9V) / R_{SENSE}$$

For high current applications, this method may be acceptable. When lower current limits are required, the size of the sense resistor, R_{SENSE}, must be increased. This will cause additional power dissipation. An alternative method for lower current limits is shown in Figure 6-2.

Figure 6-2: Low Current Limits



In this case, the Current Sense (CS) input of the MCP1630 is biased upward using the R1/R2 resistor divider. The equations for the new current limit are shown in Equation 6-2.

Equation 6-2

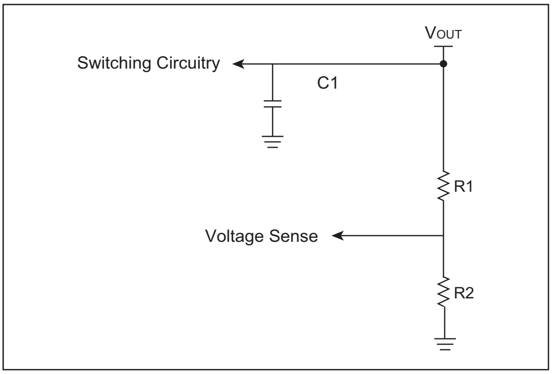
$$0.9V = \frac{(V_{DD} - I_{MAX} \cdot R_{SENSE}) \cdot R2}{R1 + R2}$$

Equation 6-2 can be solved to determine the values of R1 and R2 that provide the desired current limit.

TIP #21 Using Output Voltage Monitoring to Create a Self-Calibration Function

A PIC microcontroller can be used to create a switching power supply controlled by a PID loop (as described in Tip #16). This type of power supply senses its output voltage digitally, compares that voltage to the desired reference voltage and makes duty cycle changes accordingly. Without calibration, it is sensitive to component tolerances.

Figure 21-1: Typical Power Supply Output Stage



The output stage of many power supplies is similar to Figure 21-1. R1 and R2 are used to set the ratio of the voltage that is sensed and compared to the reference.

A simple means of calibrating this type of power supply is as follows:

1. Supply a known reference voltage to the output of the supply.
2. Place the supply in Calibration mode and allow it to sense that reference voltage.

By providing the supply with the output voltage that it is to produce, it can then sense the voltage across the resistor divider and store the sensed value. Regardless of resistor tolerances, the sensed value will always correspond to the proper output value for that particular supply.

Futhermore, this setup could be combined with Tip #20 to calibrate at several temperatures.

This setup could also be used to create a programmable power supply by changing the supplied reference and the resistor divider for voltage feedback.

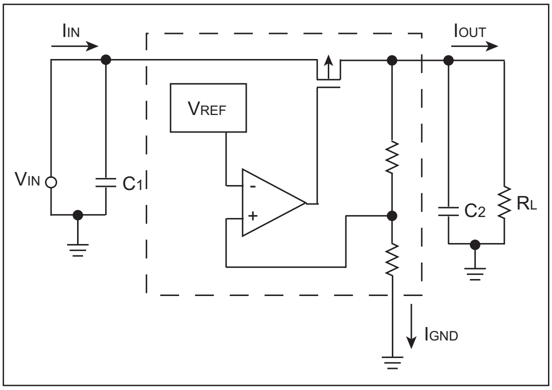
TIP #1 Powering 3.3V Systems From 5V Using an LDO Regulator

The dropout voltage of standard three-terminal linear regulators is typically 2.0-3.0V. This precludes them from being used to convert 5V to 3.3V reliably. Low Dropout (LDO) regulators, with a dropout voltage in the few hundred milli-volt range, are perfectly suited for this type of application. Figure 1-1 contains a block diagram of a basic LDO system with appropriate current elements labeled. From this figure it can be seen that an LDO consists of four main elements:

1. Pass transistor
2. Bandgap reference
3. Operational amplifier
4. Feedback resistor divider

When selecting an LDO, it is important to know what distinguishes one LDO from another. Device quiescent current, package size and type are important device parameters. Evaluating for each parameter for the specific application yields an optimal design.

Figure 1-1: LDO Voltage Regulator



An LDOs quiescent current, I_Q, is the device ground current, I_{GND}, while the device is operating at no load. I_{GND} is the current used by the LDO to perform the regulating operation. The efficiency of an LDO can be approximated as the output voltage divided by the input voltage when $I_{OUT} >> I_Q$. However, at light loads, the I_Q must be taken into account when calculating the efficiency. An LDO with lower I_Q will have a higher light load efficiency. This increase in light load efficiency has a negative effect on the LDO performance. Higher quiescent current LDOs are able to respond quicker to sudden line and load transitions.

TIP #4 Powering 3.3V Systems From 5V Using Switching Regulators

A buck switching regulator, shown in Figure 4-1, is an inductor-based converter used to step-down an input voltage source to a lower magnitude output voltage. The regulation of the output is achieved by controlling the ON time of MOSFET Q1. Since the MOSFET is either in a lower or high resistive state (ON or OFF, respectively), a high source voltage can be converted to a lower output voltage very efficiently.

The relationship between the input and output voltage can be established by balancing the volt-time of the inductor during both states of Q1.

Equation 4-1

$$(V_S - V_O) * t_{on} = V_o * (T - t_{on})$$
$$\text{Where: } T \equiv t_{on}/\text{Duty\_Cycle}$$

It therefore follows that for MOSFET Q1:

Equation 4-2

$$\text{Duty\_Cycle}_{Q1} = V_O/V_S$$

When choosing an inductor value, a good starting point is to select a value to produce a maximum peak-to-peak ripple current in the inductor equal to ten percent of the maximum load current.

Equation 4-3

$$V = L * (di/dt)$$
$$L = (V_S - V_O) * (t_{on}/I_o * 0.10)$$

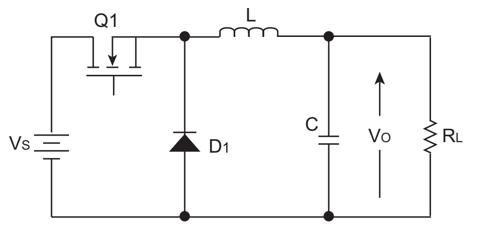
When choosing an output capacitor value, a good starting point is to set the LC filter characteristic impedance equal to the load resistance. This produces an acceptable voltage overshoot when operating at full load and having the load abruptly removed.

Equation 4-4

$$Z_o \equiv \sqrt{L/C}$$
$$C = L/R^2 = (I_O^2 * L)/V_O^2$$

When choosing a diode for D1, choose a device with a sufficient current rating to handle the inductor current during the discharge part of the pulse cycle (I_L).

Figure 4-1: Buck Regulator



Digital Interfacing

When interfacing two devices that operate at different voltages, it is imperative to know the output and input thresholds of both devices. Once these values are known, a technique can be selected for interfacing the devices based on the other requirements of your application. Table 4-1 contains the output and input thresholds that will be used throughout this document. When designing an interface, make sure to reference your manufacturers data sheet for the actual threshold levels.

Table 4-1: Input/Output Thresholds

	V_{OH} min	V_{OL} max	V_{IH} min	V_{IL} max
5V TTL	2.4V	0.5V	2.0V	0.8V
3.3V LVTTL	2.4V	0.4V	2.0V	0.8V
5V CMOS	4.7V (Vcc-0.3V)	0.5V	3.5V (0.7xVcc)	1.5V (0.3xVcc)
3.3V LVCMOS	3.0V (Vcc-0.3V)	0.5V	2.3V (0.7xVcc)	1.0V (0.3xVcc)

Table 18-1: Bipolar Transistor DC Specifications

Characteristic	Sym	Min	Max	Unit	Test Condition
OFF CHARACTERISTICS					
Collector-Base Breakdown Voltage	$V_{(BR)CBO}$	60	–	V	I_C = 50 µA, I_E = 0
Collector-Emitter Breakdown Voltage	$V_{(BR)CEO}$	50	–	V	I_C = 1.0 mA, I_B = 0
Emitter-Base Breakdown Voltage	$V_{(BR)EBO}$	7.0	–	V	I_E = 50 µA, I_C = 0
Collector Cutoff Current	I_{CBO}	–	100	nA	V_{CB} = 60V
Emitter Cutoff Current	I_{EBO}	–	100	nA	V_{EB} = 7.0V
ON CHARACTERISTICS					
DC Current Gain	h_{FE}	120 180 270	270 390 560	–	V_{CE} = 6.0V, I_C = 1.0 mA
Collector-Emitter Saturation Voltage	$V_{CE(SAT)}$	–	0.4	V	I_C = 50 mA, I_B = 5.0 mA

When using bipolar transistors as switches to turn on and off loads controlled by the microcontroller I/O port pin, use the minimum h_{FE} specification and margin to ensure complete device saturation.

Equation 18-1: Calculating the Base Resistor Value

$$R_{BASE} = \frac{(V_{DD} - V_{BE}) \times h_{FE} \times R_{LOAD}}{V_{LOAD}}$$

3V Technology Example

V_{DD} = +3V, V_{LOAD} = +40V, R_{LOAD} = 400Ω, h_{FE} min. = 180, V_{BE} = 0.7V

<u>R_{BASE} = 4.14 kΩ, I/O port current = 556 µA</u>

5V Technology Example

V_{DD} = +5V, V_{LOAD} = +40V, R_{LOAD} = 400Ω, h_{FE} min. = 180, V_{BE} = 0.7V

<u>R_{BASE} = 7.74 kΩ, I/O port current = 556 µA</u>

For both examples, it is good practice to increase base current for margin. Driving the base with 1 mA to 2 mA would ensure saturation at the expense of increasing the input power consumption.