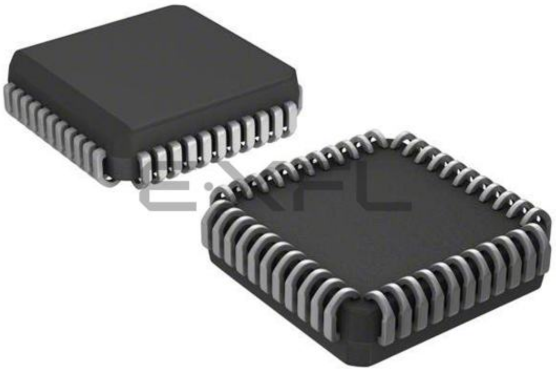

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What is "Embedded - Microcontrollers"?

"Embedded - Microcontrollers" refer to small, integrated circuits designed to perform specific tasks within larger systems. These microcontrollers are essentially compact computers on a single chip, containing a processor core, memory, and programmable input/output peripherals. They are called "embedded" because they are embedded within electronic devices to control various functions, rather than serving as standalone computers. Microcontrollers are crucial in modern electronics, providing the intelligence and control needed for a wide range of applications.

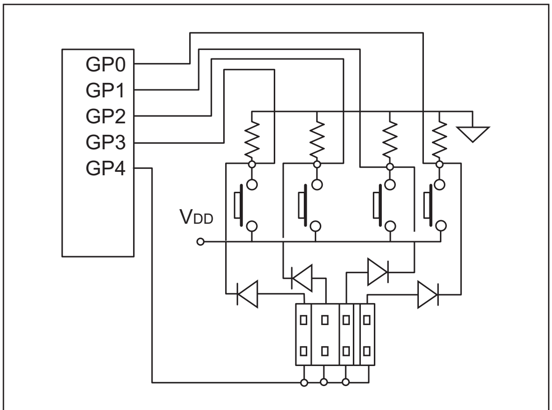
Applications of "Embedded - Microcontrollers"

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-LCC (J-Lead)
Supplier Device Package	44-PLCC (16.59x16.59)
Purchase URL	https://www.e-xfl.com/product-detail/microchip-technology/pic16c65bt-20-l

TIP #9 Decode Keys and ID Settings

Buttons and jumpers can share I/O's by using another I/O to select which one is read. Both buttons and jumpers are tied to a shared pull-down resistor. Therefore, they will read as '0' unless a button is pressed or a jumper is connected. Each input (GP3/2/1/0) shares a jumper and a button. To read the jumper settings, set GP4 to output high and each connected jumper will read as '1' on its assigned I/O or '0' if it's not connected. With GP4 output low, a pressed button will be read as '1' on its assigned I/O and '0' otherwise.

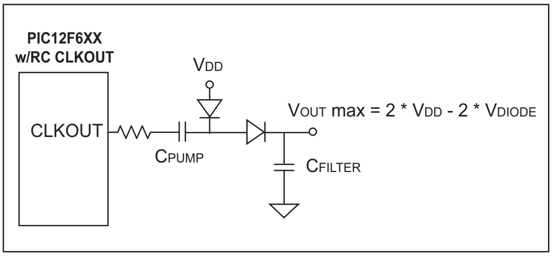
Figure 9-1



- When GP4 = 1 and no keys are pressed, read ID setting
- When GP4 = 0, read the switch buttons

TIP #10 Generating High Voltages

Figure 10-1



Voltages greater than V_{DD} can be generated using a toggling I/O. PIC MCUs CLKOUT/OSC2 pin toggles at one quarter the frequency of OSC1 when in external RC oscillator mode. When OSC2 is low, the V_{DD} diode is forward biased and conducts current, thereby charging C_{PUMP}. After OSC2 is high, the other diode is forward biased, moving the charge to C_{FILTER}. The result is a charge equal to twice the V_{DD} minus two diode drops. This can be used with a PWM, a toggling I/O or other toggling pin.

Application Notes:

AN512, “I*mplementing Ohmmeter/Temperature Sensor*”

AN611, “*Resistance and Capacitance Meter Using a PIC16C622*”

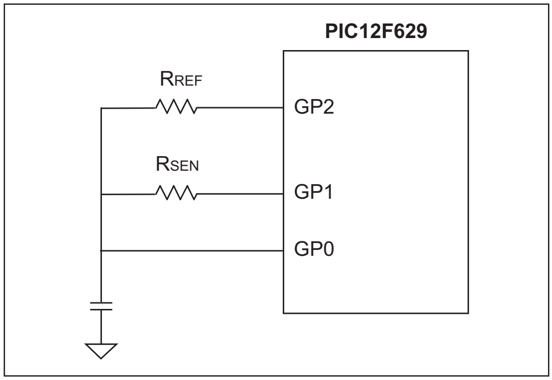
Here is the schematic and software flow for using a reference resistor to improve the accuracy of an analog sensor reading. The reference resistor (R_{REF}) and sensor (R_{SEN}) are assigned an I/O and share a common capacitor. GP0 is used to discharge the capacitor and represents the capacitor voltage.

Through software, a timer is used to measure when GP0 switches from a ‘0’ to a ‘1’ for the sensor and reference measurements. Any difference measured between the reference measurement and its calibrated measurement is used to adjust the sensor reading, resulting in a more accurate measurement.

The comparator and comparator reference on the PIC12F629/675 can be used instead of a port pin for a more accurate measurement. Polypropylene capacitors are very stable and beneficial in this type of application.

1. Set GP1 and GP2 to inputs, and GP0 to a low output to discharge C
2. Set GP0 to an input and GP1 to a high output
3. Measure tR_{SEN} (GP0 changes to 1)
4. Repeat step 1
5. Set GP0 to an input and GP2 to a high output
6. Measure tR_{REF} (GP0 changes to 1)
7. Use film polypropylene capacitor
8. $R_{TH} = x\ R_{REF}\ \dfrac{tR_{SEN}}{tR_{REF}}$

Figure 13-2



Other alternatives: voltage comparator in the PIC12F6XX to measure capacitor voltage on GP0.

TIP #2 Power Budgeting

Power budgeting is a technique that is critical to predicting current consumption and battery life. Power budgeting is performed by calculating the total charge for each mode of operation of an application by multiplying that mode’s current consumption by the time in the mode for a single application loop. The charge for each mode is added, then averaged over the total loop time to get average current. Table 1 calculates a power budget using the application from Figure 2 in Tip #1 using a typical nanoWatt XLP device.

Mode	Time in Mode (mS)	Current (mA)		Charge Current * Time (mA * Sec)
		By Device	Mode Total	
Sleep MCU Sleep Sensor Off EEPROM Off	1989	0.00005 0 0	5.00E-05	9.95E-05
Initialize MCU Sleep Sensor On EEPROM Off	1	0.00005 0.0165 0	1.66E-02	1.66E-05
Sample Sensor MCU Run Sensor On EEPROM Off	1	0.048 0.0165 0	6.45E-02	6.45E-05
Scaling MCU Run Sensor Off EEPROM Off	1	0.048 0 0	4.80E-02	4.80E-05
Storing MCU Run Sensor Off EEPROM On	8	0.048 0 1	1.05E+00	8.38E-03
Total	2000	—	—	8.61E-03

Average Current

$$= \frac{8.61e\text{-}3}{2000e\text{-}3} \quad \frac{\text{mA*Sec}}{\text{Sec}}$$

$= 0.0043$ mA

Peak Current 1.05 mA

Computing Battery Life

Using the average current from the calculated power budget, it is possible to determine how long a battery will be able to power the application. Table 2 shows lifetimes for typical battery types using the average power from Table 1.

Battery	Capacity (mAh)	Life			
		Hours	Days	Months	Years
CR1212	18	4180	174	5.8	.48
CR1620	75	17417	726	24.2	1.99
CR2032	220	51089	2129	71.0	5.83
Alkaline AAA	1250	290276	12095	403.2	33.14
Alkaline AA	2890	671118	27963	932.1	76.61
Li-ion*	850	197388	8224	274.1	22.53

NOTE: Calculations are based on average current draw only and do not include battery self-discharge.

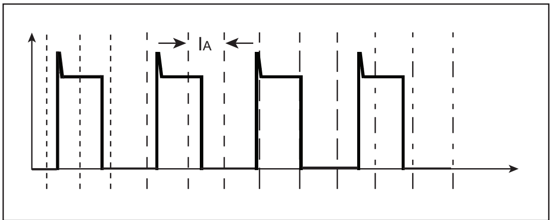
*Varies by size; value used is typical.

After completing a power budget, it is very easy to determine the battery size required to meet the application requirements. If too much power is consumed, it is simple to determine where additional effort needs to be placed to reduce the power consumption.

TIP #12 Repetitive Phase Shifted Sampling

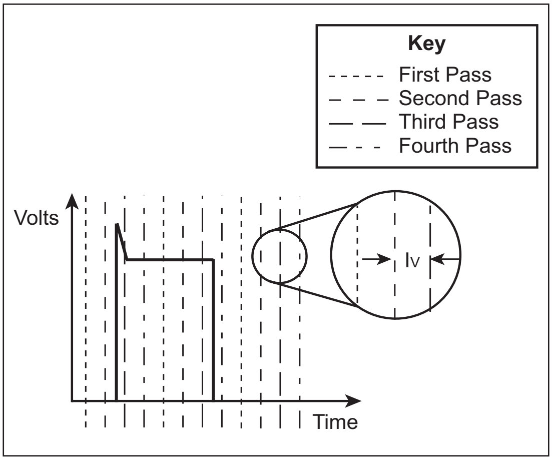
Repetitive phase shifted sampling is a technique to artificially increase the sampling rate of an A/D converter when sampling waveforms that are both periodic and constant from period to period. The technique works by capturing regularly spaced samples of the waveform from the start to finish of the 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



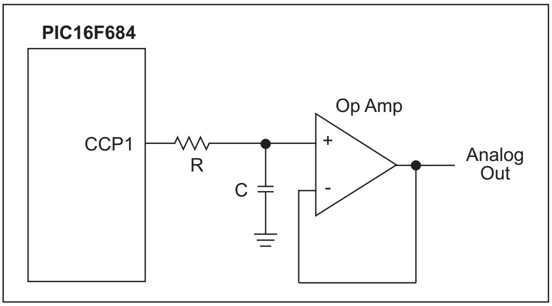
The CCP module is configured in Compare Special Event Trigger mode to accomplish this task. The phase shift is implemented by picking values of CCPRxL and CCPRxH that are not synchronous with the period of the sampling waveform. For instance, if the period of a waveform is 100 µs, then sampling at a 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 #16 Generating an Analog Output

Figure 16-1: Low-Pass Filter



Pulse-width modulated signals can be used to create Digital-to-Analog (D/A) converters with only a few external components. Conversion of PWM waveforms to analog signals involves the use of an analog low-pass filter. In order to eliminate unwanted harmonics caused by a PWM signal to the greatest degree possible, the frequency of the PWM signal (F_{PWM}) should be significantly higher than the bandwidth (F_{BW}) of the desired analog signal. Equation 16-1 shows this relation.

Equation 16-1

$$F_{PWM} = K{*}F_{BW}$$

Where harmonics decrease as K increases

R and C are chosen based on the following equation:

Equation 16-2

$$RC = 1/(2\pi F_{BW})$$

Pick a value of C arbitrarily and then calculate R. The attenuation of the PWM frequency for a given RC filter is:

Equation 16-3

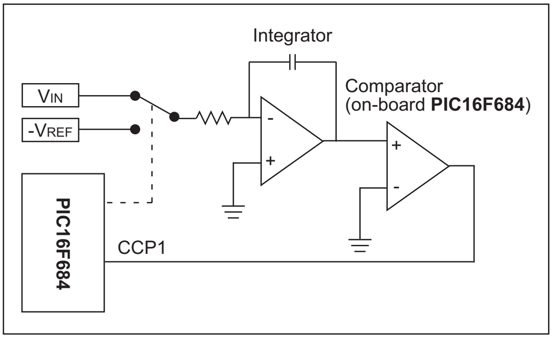
$$Att(dB = -\ 10{*}\log[1+(2\pi F_{PWM}RC)2]$$

If the attenuation calculated in Equation 16-3 is not sufficient, then K must be increased in Equation 16-1. See Application Note AN538 "*Using PWM to Generate Analog Output in PIC17C42*" for more details on using PWM to generate an analog output.

TIP #21 Dual-Slope Analog-to-Digital Converter

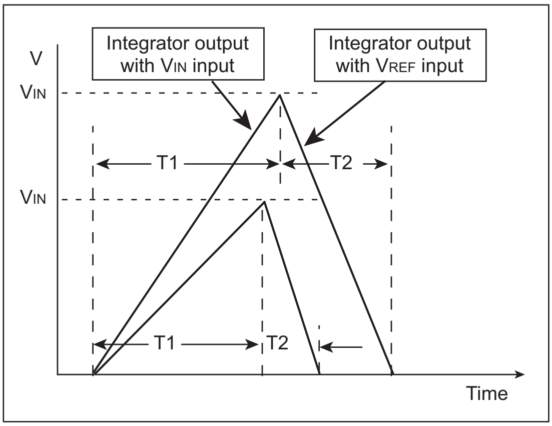
A circuit for performing dual-slope A/D conversion utilizing the CCP module is shown in Figure 21-1.

Figure 21-1: Dual-Slope Analog-to-Digital Converter



Dual-slope A/D conversion works by integrating the input signal (V_{IN}) for a fixed time (T1). The input is then switched to a negative reference (-V_{REF}) and integrated until the integrator output is zero (T2). V_{IN} is a function of V_{REF} and the ratio of T2 to T1.

Figure 21-2: V vs. Time



The components of this conversion type are the fixed time and the timing of the falling edge. The CCP module can accomplish both of these components via Compare mode and Capture mode respectively. Here’s how:

1. Configure the CCP module in Compare mode, Special Event Trigger.
2. Switch the analog input into the integrator from V_{REF} to V_{IN}.
3. Use the CCP module to wait T1 (T1 chosen based on capacitor value).
4. When the CCP interrupt occurs, switch the analog input into the regulator from V_{IN} to V_{REF} and reconfigure the module in Capture mode; wait for falling edge.
5. When the next CCP interrupt occurs, the time captured by the module is T2.
6. Calculate V_{IN} using Equation 21-1.

Equation 21-1

$$V_{IN} = V_{REF} \frac{T2}{T1}$$

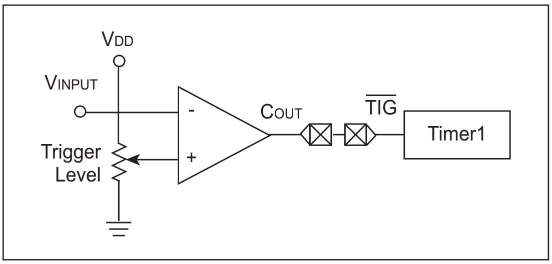
TIP #4 Pulse Width Measurement

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 $\overline{\text{T1G}}$ input is high, Timer1 does not count. Combining $\overline{\text{T1G}}$ 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.

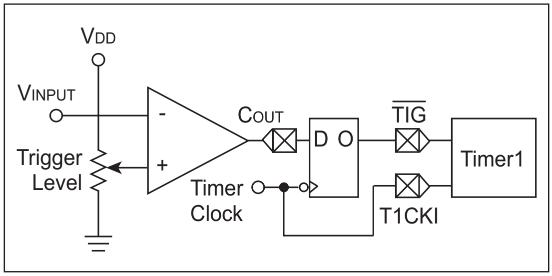
Figure 4-1: Comparator with Timer1 and $\overline{\text{T1G}}$



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



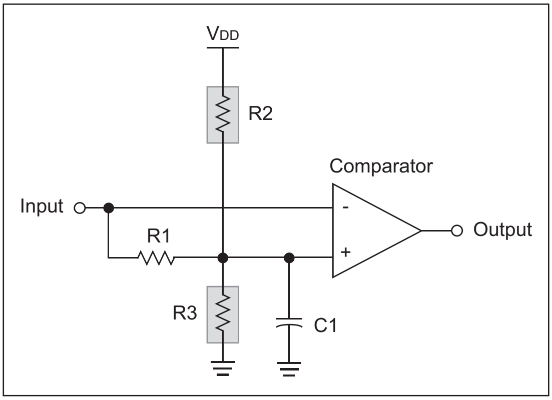
TIP #6 Data Slicer

In both wired and wireless data transmission, the data signal may be subject to DC offset shifts due to temperature shifts, ground currents or other factors in the system. When this happens, using a simple level comparison to recover the data is not possible because the DC offset may exceed the peak-to-peak amplitude of the signal. The circuit typically used to recover the signal in this situation is a data slicer.

The data slicer shown in Figure 6-1 operates by comparing the incoming signal with a sliding reference derived from the average DC value of the incoming signal. The DC average value is found using a simple RC low-pass filter (R1 and C1). The corner frequency of the RC filter should be high enough to ignore the shifts in the DC level while low enough to pass the data being transferred.

Resistors R2 and R3 are optional. They provide a slight bias to the reference, either high or low, to give a preference to the state of the output when no data is being received. R2 will bias the output low and R3 will bias the output high. Only one resistor should be used at a time, and its value should be at least 50 to 100 times larger than R1.

Figure 6-1: Data Slicer



Example:

Data rate of 10 kbits/second. A low pass filter frequency of 500 Hz: R1 = 10k, C1 = 33 μF. R2 or R3 should be 500k to 1 MB.

TIP #20 Logic: Set/Reset Flip Flop

This tip shows the use of the comparator to implement a Set/Reset Flip Flop.

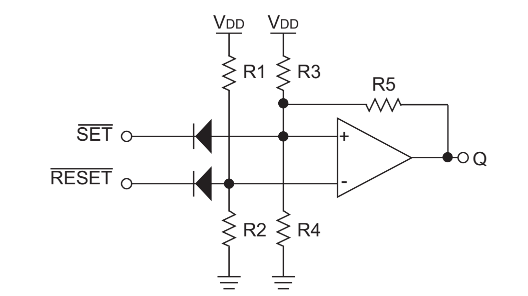
The inverting and non-inverting inputs are biases at VDD/2 by resistors R1 through R4. The non-inverting input also receives positive feedback from the output through R5. The common bias voltages and the positive feedback configure the comparator as a bistable latch. If the output Q is high, the non-inverting input is also pulled high, which reinforces the high output. If Q is low, the non-inverting input is also pulled low, which reinforces the low output. To change state, the appropriate input must be pulled low to overcome the positive feedback. The diodes prevent a positive state on either input from pulling the bias of either input above VDD/2.

> **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 Set and Reset must be driven near ground for the circuit to operate properly.
2. The combination of R1/R2 and R3/R4 will draw current constantly, so they must be kept large to minimize current draw.
3. R1 through R4 must be equal for a VDD/2 trip level.
4. R5 must be greater or equal to R3.
5. R1 through R4 will react with the input capacitance of the comparator, so larger values will limit the minimum input pulse width.

Figure 20-1: Set/Reset Flip Flop



Example:

- Diodes = 1N4148
- R1 = R2 = R3 = R4 = 10k
- R5 = 10k

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.

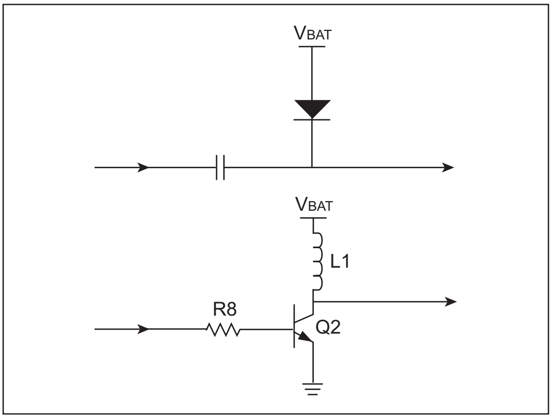
TIP #2 LCD PIC® MCU Segment/ Pixel Table

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).

Figure 5-2: Two Types of Boost Converter

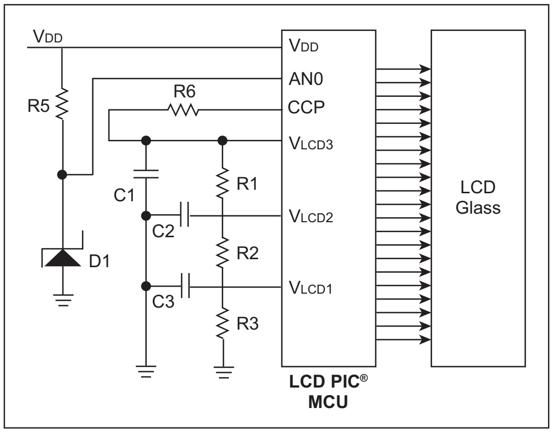


The two methods of producing a boost converter are shown above. The first circuit is simply a switched capacitor type circuit. The second circuit is a standard inductor boost circuit. These circuits work by raising VDD. This allows the voltage at VLCD to exceed VDD.

TIP #6: Software Controlled Contrast with PWM for LCD Contrast Control

In the previous contrast control circuits, the voltage output was set by a fixed reference. In some cases, the contrast must be variable to account for different operating conditions. The CCP module, available in the LCD controller devices, allows a PWM signal to be used for contrast control. In Figure 6-1, you see the buck contrast circuit modified by connecting the input to RA6 to a CCP pin. The resistor divider created by R4 and R5 in the previous design are no longer required. An input to the ADC is used to provide feedback but this can be considered optional. If the ADC feedback is used, notice that it is used to monitor the VDD supply. The PWM will then be used to compensate for variations in the supply voltage.

Figure 6-1: Software Controlled Voltage Generator



Application Note References

- AN220, “*Watt-Hour Meter Using PIC16C923 and CS5460*” (DS00220)
- AN582, “*Low-Power Real-Time Clock*” (DS00582)
- AN587, “*Interfacing PIC® MCUs to an LCD Module*” (DS00587)
- AN649, “*Yet Another Clock Featuring the PIC16C924*” (DS00649)
- AN658, “*LCD Fundamentals Using PIC16C92X Microcontrollers*” (DS00658)
- TB084, “*Contrast Control Circuits for the PIC16F91X*” (DS91084)

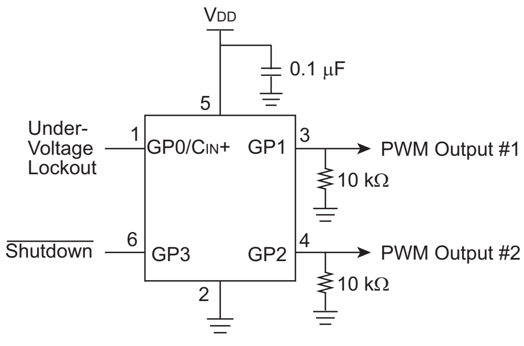
Application notes can be found on the Microchip web site at www.microchip.com.

TIP #2 A Start-Up Sequencer

Some new devices have multiple voltage requirements (e.g., core voltages, I/O voltages, etc.). The sequence in which these voltages rise and fall may be important.

By expanding on the previous tip, a start-up sequencer can be created to control two output voltages. Two PWM outputs are generated to control the shutdown pins of two SMPS controllers. Again, this type of control only works on controllers that respond quickly to changes on the shutdown pin (such as those that do cycle-by-cycle limiting).

Figure 2-1: Multiple PWM Output Soft-Start Controller



Note: Assumes SOT-23 packaging.

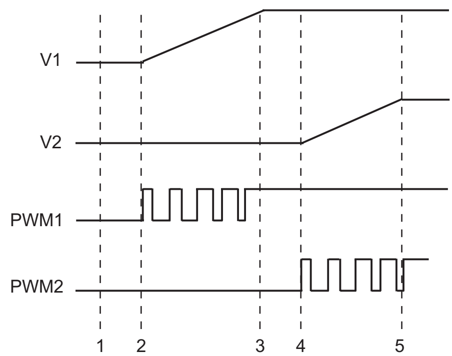
This design uses the PIC MCU comparator to implement an under-voltage lockout. The input on the GP0/C_{IN}+ pin must be above the internal 0.6V reference for soft-start to begin, as shown in Figure 2-2.

Two conditions must be met in order for the soft-start sequence to begin:

1. The shutdown pin must be held at V_{DD} (logic high).
2. The voltage on GP0 must be above 0.6V.

Once both start-up conditions are met, the sequences will delay and PWM #1 will ramp from 0% to 100%. A second delay allows the first voltage to stabilize before the sequencer ramps PWM #2 from 0% to 100%. All delays and ramp times are under software control and can be customized for specific applications. If either soft-start condition becomes invalid, the circuit will shutdown the SMPS controllers.

Figure 2-2: Timing Diagram



1. Start-up conditions met
2. Initial delay
3. PWM Ramp #1 complete
4. Between PWM delay
5. PWM #2 complete

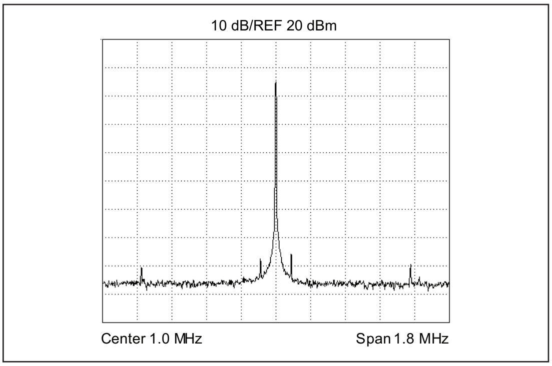
Example software is provided for the PIC10F200 which was taken from TB093, “*Multiple PWM Output Soft-Start Controller for Switching Power Supplies*” (DS91093).

TIP #4 Creating a Dithered PWM Clock

In order to meet emissions requirements as mandated by the FCC and other regulatory organizations, the switching frequency of a power supply can be varied. Switching at a fixed frequency produces energy at that frequency. By varying the switching frequency, the energy is spread out over a wider range and the resulting magnitude of the emitted energy at each individual frequency is lower.

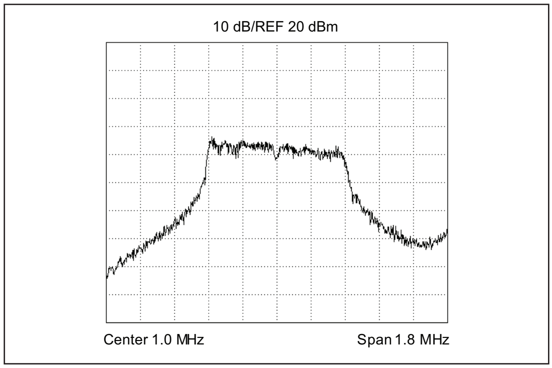
The PIC10F200 has an internal 4 MHz oscillator. A scaled version of oscillator can be output on a pin (Fosc/4). The scaled output is 1/4 of the oscillator frequency (1 MHz) and will always have a 50% duty cycle. Figure 4-1 shows a spectrum analyzer shot of the output of the Fosc/4 output.

Figure 4-1: Spectrum of Clock Output Before Dithering



The PIC10F200 provides an Oscillator Calibration (OSCCAL) register that is used to calibrate the frequency of the oscillator. By varying the value of the OSCCAL setting, the frequency of the clock output can be varied. A pseudo-random sequence was used to vary the OSCCAL setting, allowing frequencies from approximately 600 kHz to 1.2 MHz. The resulting spectrum is shown in Figure 4-2.

Figure 4-2: Spectrum of Clock Output After Dithering



By spreading the energy over a wider range of frequencies, a drop of more than 20 dB is achieved.

Example software is provided for the PIC10F200 that performs the pseudo-random sequence generation and loads the OSCCAL register.

TIP #8 Transformerless Power Supplies

When using a microcontroller in a line-powered application, such as the IR remote control actuated AC switch described in Tip #9, the cost of building a transformer-based AC/DC converter can be significant. However, there are transformerless alternatives which are described below.

Capacitive Transformerless Power Supply

Figure 8-1: Capacitive Power Supply

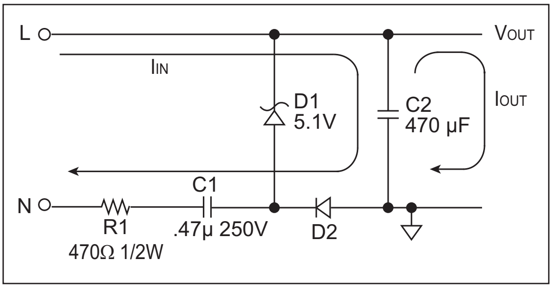


Figure 8-1 shows the basics for a capacitive power supply. The Zener diode is reverse-biased to create the desired voltage. The current drawn by the Zener is limited by R1 and the impedance of C1.

Advantages:

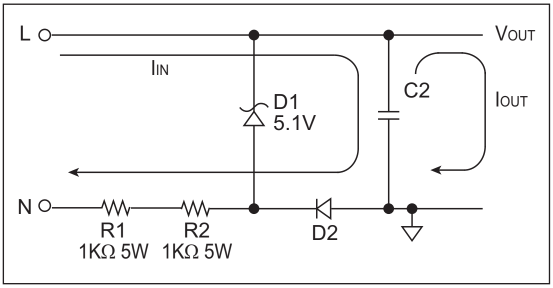
- Significantly smaller than a transformer-based power supply
- Lower cost than a transformer-based or switcher-based power supply
- Power supply is more efficient than a resistive transformerless power supply

Disadvantages:

- Not isolated from the AC line voltage which introduces safety issues
- Higher cost than a resistive power supply because X2 rated capacitors are required

Resistive Power Supply

Figure 8-2: Resistive Power Supply



The resistive power supply works in a similar manner to the capacitive power supply by using a reversed-biased Zener diode to produce the desired voltage. However, R1 is much larger and is the only current limiting element.

Advantages:

- Significantly smaller than a transformer-based power supply
- Lower cost than a transformer-based power supply
- Lower cost than a capacitive power supply

Disadvantages:

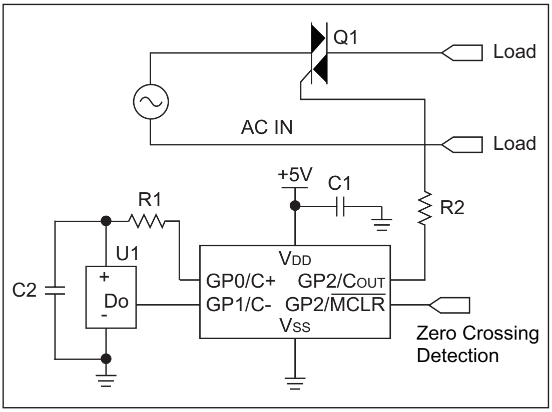
- Not isolated from the AC line voltage which introduces safety issues
- Power supply is less energy efficient than a capacitive power supply
- More energy is dissipated as heat in R1

More information on either of these solutions, including equations used for calculating circuit parameters, can be found in AN954, "*Transformerless Power Supplies: Resistive and Capacitive*" (DS00954) or in TB008, "*Transformerless Power Supply*" (DS91008).

TIP #9 An IR Remote Control Actuated AC Switch for Linear Power Supply Designs

Many line-powered applications (audio amplifiers, televisions, etc.) can be turned on and off using an infrared remote control. This requires that some components be energized to receive the remote signals even when the device is off. Low current PIC microcontrollers are best in this application. Figure 9-1 shows an example circuit layout.

Figure 9-1: PIC MCU Infrared Receiver Schematic



The PIC10F200 has several features that make it ideally suited for this type of application:

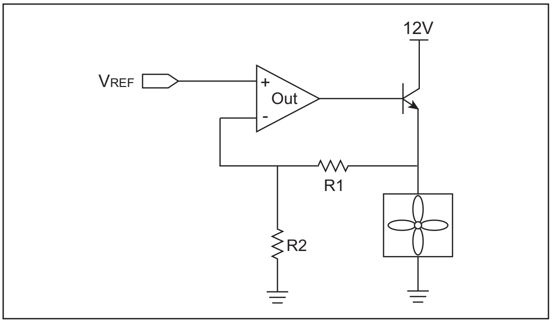
- Extremely low operating and standby current (350 μA operating, 0.1 μA when asleep)
- Input/Output pins with configurable pull-ups and reset-on-change capability
- High sink/source ability (±25 mA) allows driving external devices, such as the IR receiver, directly from the I/O pin
- Ability to use a low-cost resistive power supply
- Small form factor (SOT-23 packaging)

TB094, “*Dimming AC Incandescent Lamps Using A PIC10F200*” (DS91094) provides both software and hardware examples of an infrared controller.

Method 2 – Linear Control

When using PWM, the voltage will vary between a maximum and a minimum, however, is it also possible to use a linear method to control fan speed, as shown in Figure 14-4.

Figure 14-4: Linear Control Drive



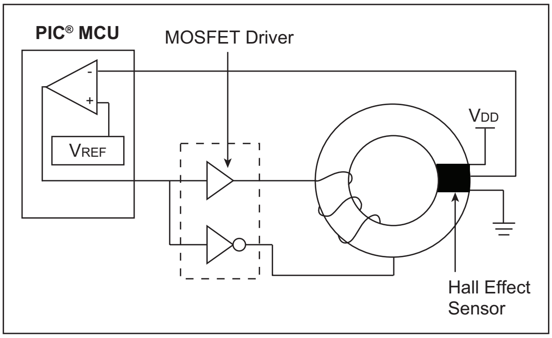
The voltage applied at the non-inverting terminal of the op amp is used to vary the voltage across the op amp. The non-inverting terminal voltage can be produced by a Digital-to-Analog Converter (DAC) or by the method shown in Tip #11.

When using this method, care must be taken to ensure that the fan voltage is not too low or the fan will stop spinning. One advantage this method has over PWM is that the tachometer output will function properly on 3-wire fans. The disadvantage, however, is that it often offers less speed control. For example, a 12V fan will not spin below 8V, so a range of only 4V is available for speed control. A 5V fan will not spin below 4V and so the control range is only 1V, which is often unacceptable. Another disadvantage is the power consumption of the circuit. The transistor will dissipate more power than the PWM method.

TIP #15 High Current Delta-Sigma Based Current Measurement Using a Slotted Ferrite and Hall Effect Device

Many current sensors rely on ferrite cores. Non-linearity in the ferrite can lead to inaccurate results, especially at high currents. One way to avoid the non-linearities is to keep the net flux in the ferrite near zero. Consider the circuit in Figure 15-1.

Figure 15-1: Hall Effect Current Measurement Schematic



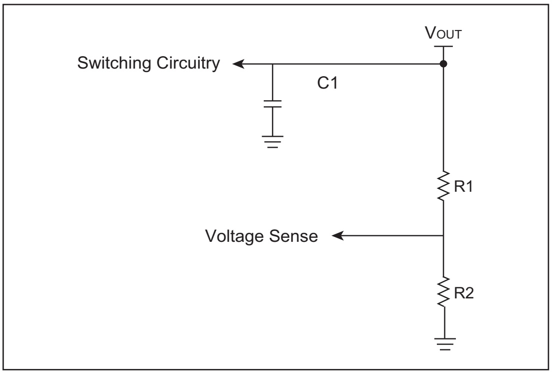
The Hall effect sensor output is proportional to the current being measured. When I_{IN} = 0 amps, the output of the sensor will be $V_{DD}/2$. A current passing through the sensor in one direction will increase the output of the sensor, and a current in the other direction will decrease the output of the sensor.

The output of the comparator is used to drive a coil of wire wound around the ferrite core. This coil of wire will be used to create flux in the opposite direction as the flux imposed in the core.

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 #5 3.3V → 5V Direct Connect

The simplest and most desired way to connect a 3.3V output to a 5V input is by a direct connection. This can be done only if the following 2 requirements are met:

- The V_{OH} of the 3.3V output is greater than the V_{IH} of the 5V input
- The V_{OL} of the 3.3V output is less than the V_{IL} of the 5V input

An example of when this technique can be used is interfacing a 3.3V LVCMOS output to a 5V TTL input. From the values given in Table 4-1, it can clearly be seen that both of these requirements are met.

3.3V LVCMOS V_{OH} of 3.0 volts is greater than 5V TTL V_{IH} of 2.0 volts, and

3.3V LVCMOS V_{OL} of 0.5 volts is less than 5V TTL V_{IL} of 0.8 volts.

When both of these requirements are not met, some additional circuitry will be needed to interface the two parts. See Tips 6, 7, 8 and 13 for possible solutions.

TIP #6 3.3V → 5V Using a MOSFET Translator

In order to drive any 5V input that has a higher V_{IH} than the V_{OH} of a 3.3V CMOS part, some additional circuitry is needed. A low-cost two component solution is shown in Figure 6-1.

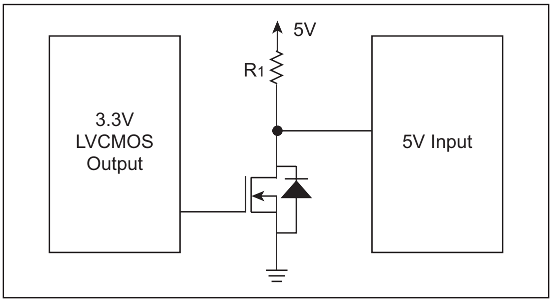
When selecting the value for R1, there are two parameters that need to be considered; the switching speed of the input and the current consumption through R1. When switching the input from a ‘0’ to a ‘1’, you will have to account for the time the input takes to rise because of the RC time constant formed by R1, and the input capacitance of the 5V input plus any stray capacitance on the board. The speed at which you can switch the input is given by the following equation:

Equation 6-1

$$T_{SW} = 3 \times R_1 \times (C_{IN} + C_S)$$

Since the input and stray capacitance of the board are fixed, the only way to speed up the switching of the input is to lower the resistance of R1. The trade-off of lowering the resistance of R1 to get faster switching times is the increase in current draw when the 5V input remains low. The switching to a ‘0’ will typically be much faster than switching to a ‘1’ because the ON resistance of the N-channel MOSFET will be much smaller than R1. Also, when selecting the N-channel FET, select a FET that has a lower V_{GS} threshold voltage than the V_{OH} of 3.3V output.

Figure 6-1: MOSFET Translator



TIP #16 5V → 3.3V Active Analog Attenuator

Reducing a signal’s amplitude from a 5V to 3.3V system using an op amp.

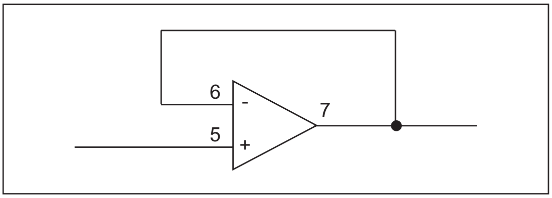
The simplest method of converting a 5V analog signal to a 3.3V analog signal is to use a resistor divider with a ratio R1:R2 of 1.7:3.3. However, there are a few problems with this.

1. The attenuator may be feeding a capacitive load, creating an unintentional low pass filter.
2. The attenuator circuit may need to drive a low-impedance load from a high-impedance source.

Under either of these conditions, an op amp becomes necessary to buffer the signals.

The op amp circuit necessary is a unity gain follower (see Figure 16-1).

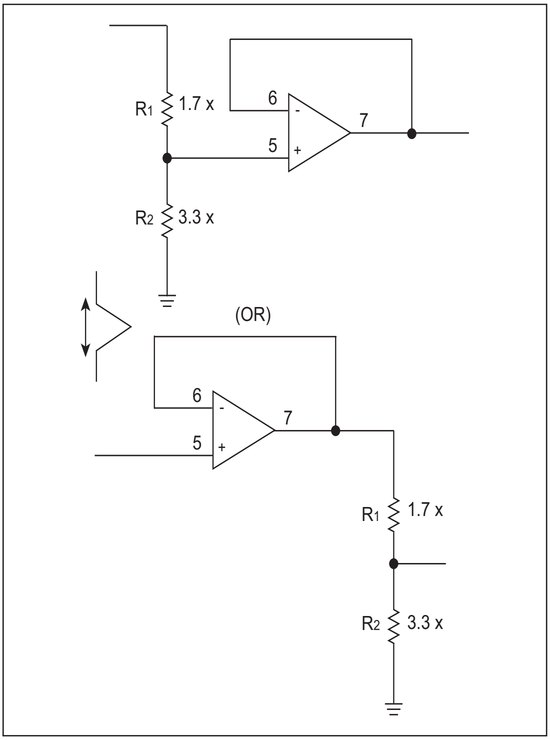
Figure 16-1: Unity Gain



This circuit will output the same voltage that is applied to the input.

To convert the 5V signal down to a 3V signal, we simply add the resistor attenuator.

Figure 16-2: Op Amp Attenuators



If the resistor divider is before the unity gain follower, then the lowest possible impedance is provided for the 3.3V circuits. Also, the op amp can be powered from 3.3V, saving some power. If the X is made very large, then power consumed by the 5V side can be minimized.

If the attenuator is added after the unity gain follower, then the highest possible impedance is presented to the 5V source. The op amp must be powered from 5V and the impedance at the 3V side will depend upon the value of R1||R2.

