This presentation is closely related to the following 2005 Masters presentation:

• DAQ 972, "Cool Data Acquisition Applications (or how to interface the PIC16F68X to the Real World)"

This presentation put more emphasis on the firmware (F/W) and software (S/W) issues. Its goal was to show how to integrate several analog and PIC solutions together in one application: a data logger that collects temperature, humidity, and light data. It also includes a real time clock using a watch crystal and two SPI memory chips. It included code, description of subroutines, and coding practices information.

This presentation places the emphasis on the analog signal conditioning aspects of the same applications. It also pulls together information in AN990 (a bird's eye view of sensor applications), FilterLab® V2.0's User's Guide, and adds essential material.

The lab exercises will help show how the theory applies to real applications, and demonstrates the power of our modular demo board approach.

Class Objectives

In this class you will:

- Review fundamentals
  - Four Common Sensor Types
  - Conditioning Circuits
  - Analog Filters
- Work with sensor conditioning circuits
- Obtain hands-on experience
- Receive references to additional design resources

 306 ASC Slide 2

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Class Abstract

Most sensor circuits require some analog signal conditioning before conversion to digital. This class gives background information on the many types of sensors and sensor conditioning circuits, including active filters. Three common sensors and their conditioning circuits are then covered in some detail. Hands-on experiments will help illustrate these sensor circuits and the filter design theory.

The three common sensors covered are: thermistor (temperature), photodiode (light), and capacitance (humidity). The filter designs will be generated by Microchip's FilterLab® Active Filter Design Tool software.

 306 ASC Slide 3

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Agenda

- **Analog Filters**
  - (FilterLab® Active Filter Design Tool 2.0 Lab)
- **Programmable Gain Amplifier (PGA)**
- **Sensor Circuits**
  - Voltage Sensors
  - Current Sensors (Photodiode Lab)
  - Resistive Sensors (Thermistor Lab)
  - Capacitive Sensors (Humidity Sensor Lab)

 306 ASC Slide 4

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Agenda

- **Combined Designs**
- **Summary**
- **Appendix A**
  - References

 306 ASC Slide 5

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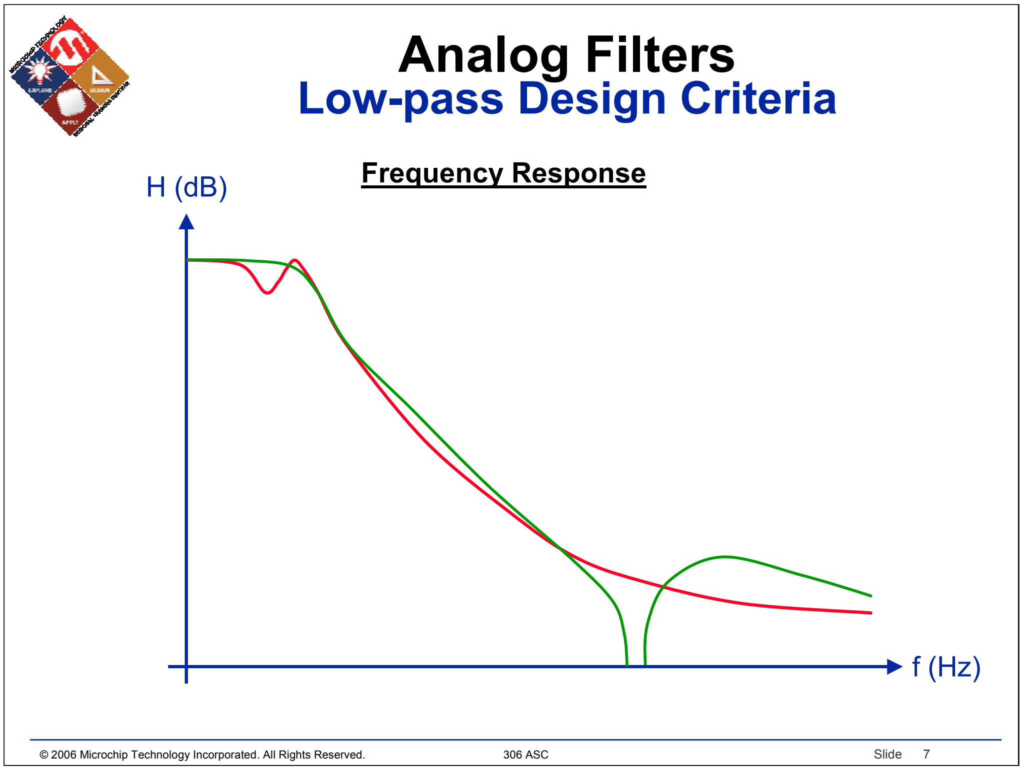
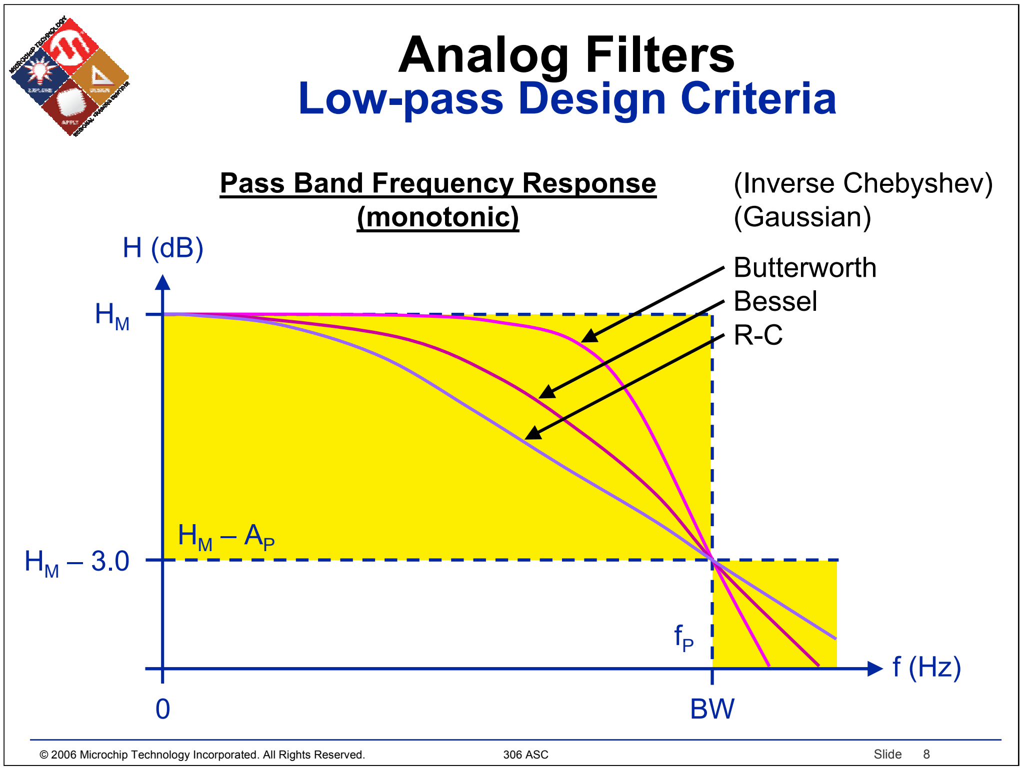
See application notes AN699 and AN737 (cited in Appendix A: References, Analog Filters).

Download FilterLab® V2.0 from Microchip Technology Inc.'s website (www.microchip.com). The User's Guide for FilterLab can be found under the Help menu button.



The two approximation functions shown display features in the more commonly used filters:

- Monotonic or Equal Ripple pass band
- Monotonic transition band
- Monotonic or Equal Ripple stop band
- This presentation follows the conventions found in FilterLab's User's Guide (see the following slides):
- The maximum gain is H_M (in units of dB)
- The attenuations
    - Are all ≥ 0 dB
    - A_P = Pass Band ripple (same as cut-off)
    - A_S = Stop Band ripple
- The frequencies
    - f_P = Pass Band frequency
    - BW = –3.0 dB bandwidth (half-power bandwidth) ≥ f_P
    - f_S = Stop Band frequency



Many filter approximation functions have a monotonic roll-off in the pass band. The three that we emphasize here are Butterworth, Bessel, and R-C. FilterLab includes Butterworth and Bessel, while the R-C filter designs will be covered in this presentation. The Inverse Chebyshev and Gaussian filters are only mentioned in passing for those that are familiar with them.

Butterworth filters are described as being maximally flat; this means that as many derivatives as possible are zero at the point f = 0 Hz. As a consequence, the attenuation in the pass band is near to 0 dB for as long as possible.

Bessel filters give up some of the pass-band performance shown by the Butterworth filters in exchange for better time domain behavior.

R-C filters have the worst pass band response of any of these filters, but have no step response overshoot.

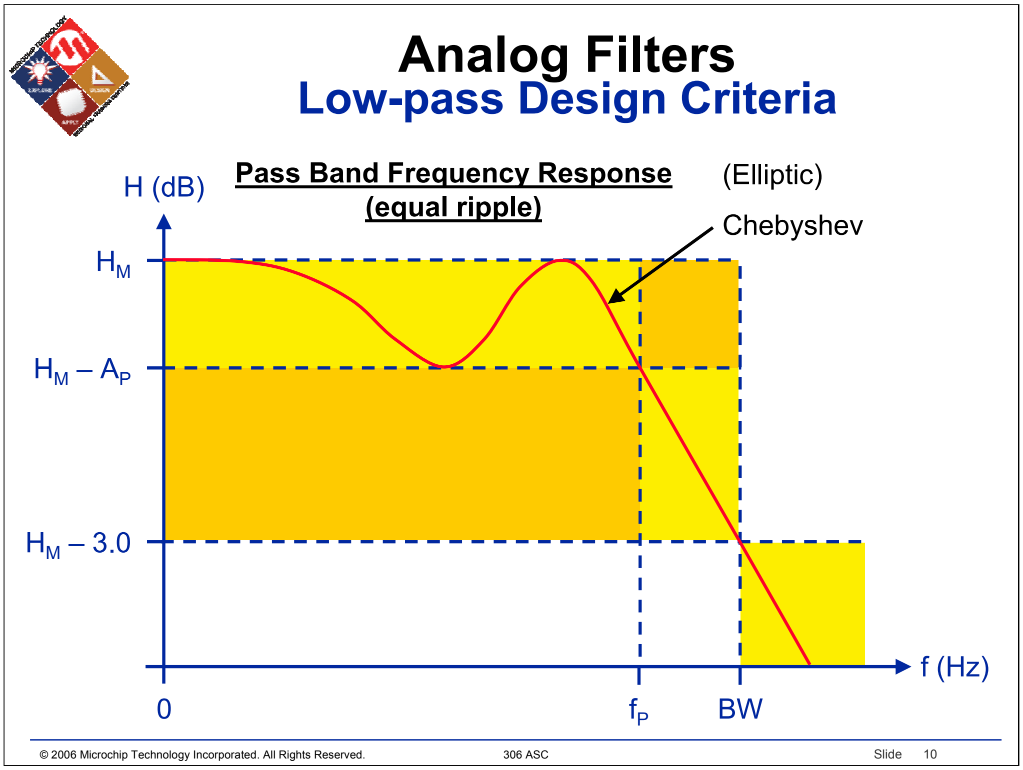
Analog Filters
Low-pass Design Criteria

- **Frequency Response**
  - Pass Band for R-C, Bessel, and Butterworth
    - A_C (pass band attenuation)
      - –3.0 dB
    - BW (–3.0 dB bandwidth)
      - Greater than signal bandwidth
    - f_P (pass band frequency)
      - f_P = BW

 306 ASC Slide 9

For these filters, f_P and BW are usually considered to be equal (A_P = -3.0 dB). In some industries, the -3 dB bandwidth is called the half-power bandwidth.

The filter bandwidth needs to be greater than the signal bandwidth so that the signal is faithfully reproduced.



Some filter approximation functions have equal ripple in the pass band below f_P. These filters achieve a sharper cut-off at the pass-band edge (BW) than those with a monotonic roll-off, at the cost of more step response overshoot, ringing, and longer settling time. We will emphasize the Chebyshev function here, which is included in FilterLab. The Elliptic approximation function is only mentioned in passing for those that are familiar with it.

Analog Filters
Low-pass Design Criteria

– Pass Band for Chebyshev
- A_P (pass band attenuation)
  – As small as possible ($A_P \le 3.0$ dB)
- BW (–3.0 dB bandwidth)
  – Greater than signal bandwidth
- f_P (pass band frequency)
  – Set for desired BW ($f_P \le$ BW)

 306 ASC Slide 11

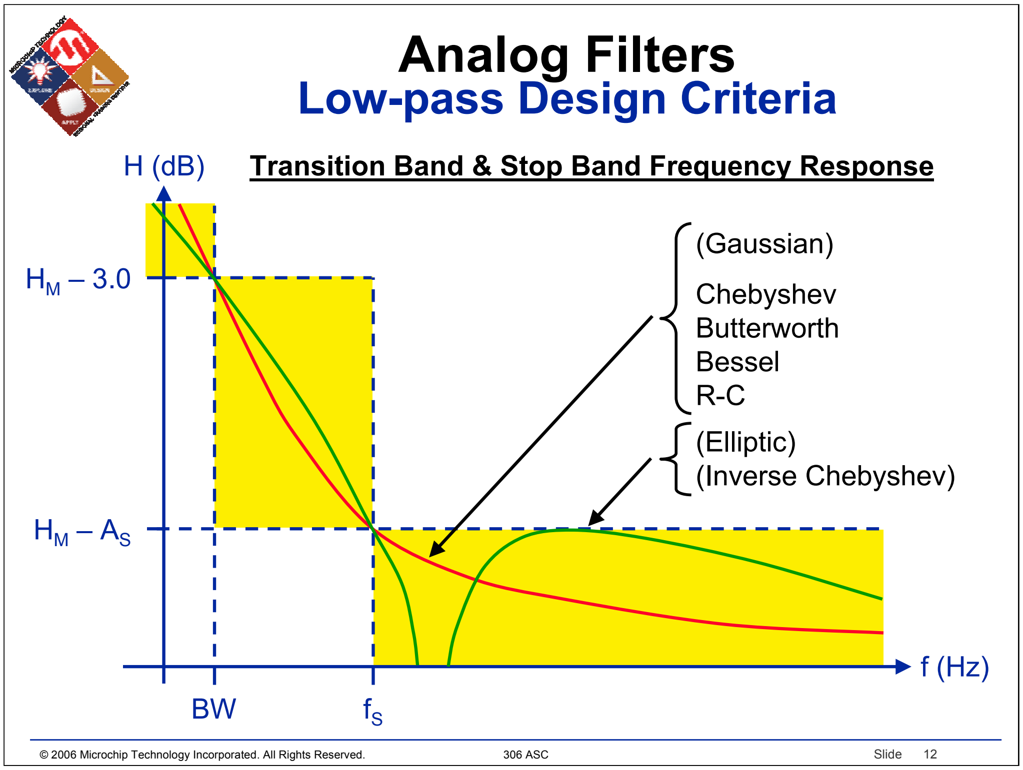
For these filters, f_P is usually less than BW ($A_P < 3.0$ dB). It is best to make A_P as small as possible to minimize:

• Frequency response ripple

• Time domain echoes

• Step response overshoot, ringing, and settling time

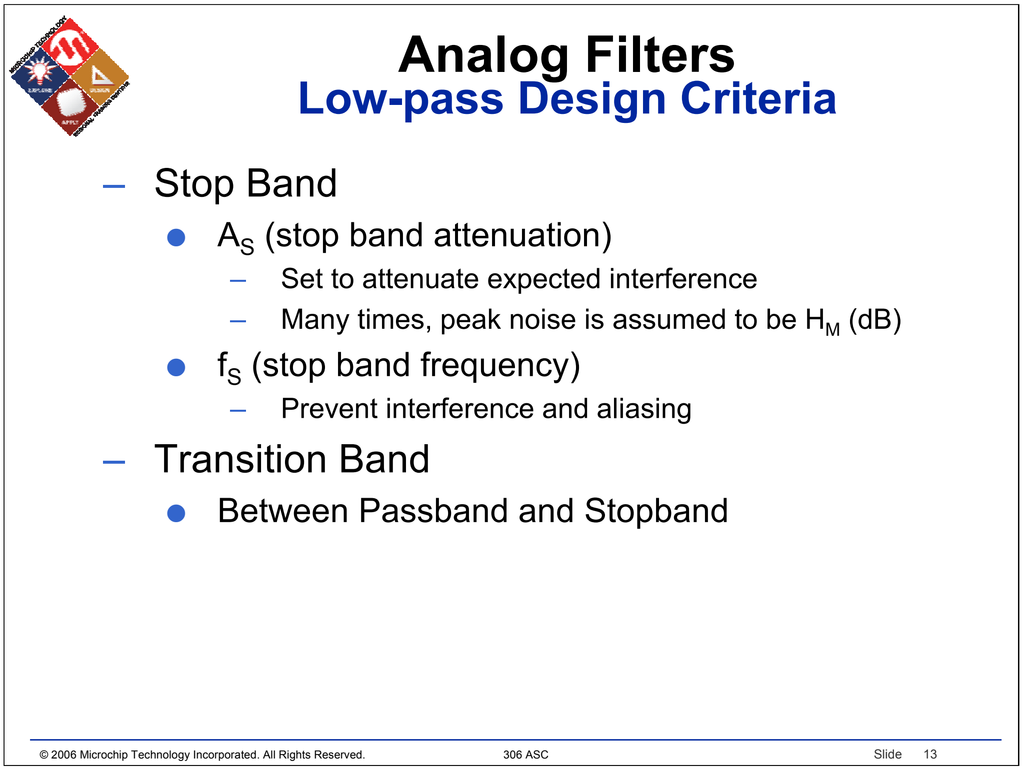
The filter bandwidth needs to be greater than the signal bandwidth so that the signal is faithfully reproduced.



The transition band is the region where the attenuation is between 3.0 dB and A_S. The vast majority of useful filters have a monotonic response in this region.

The stop band has an attenuation greater than A_S. Most filters have either a monotonic response or an equal ripple response in the stop band. The ripple is generated by transmission zeros on the $j\omega$ axis; this type of zero does not disturb the group delay response, and has little effect on the step response, and helps make the transition region more narrow. For these reasons, the Inverse Chebyshev filter has a better step response than a Chebyshev filter with the same frequency domain requirements.

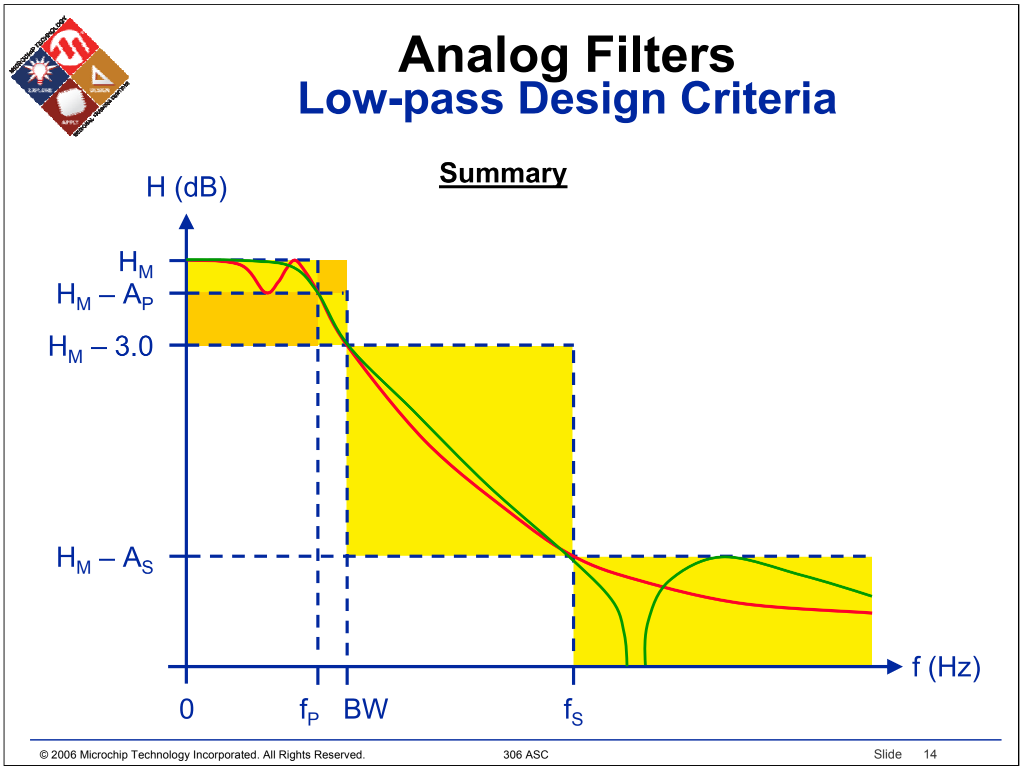
The Gaussian, Inverse Chebyshev, and Elliptic approximation functions are only mentioned in passing for those that are familiar with them.



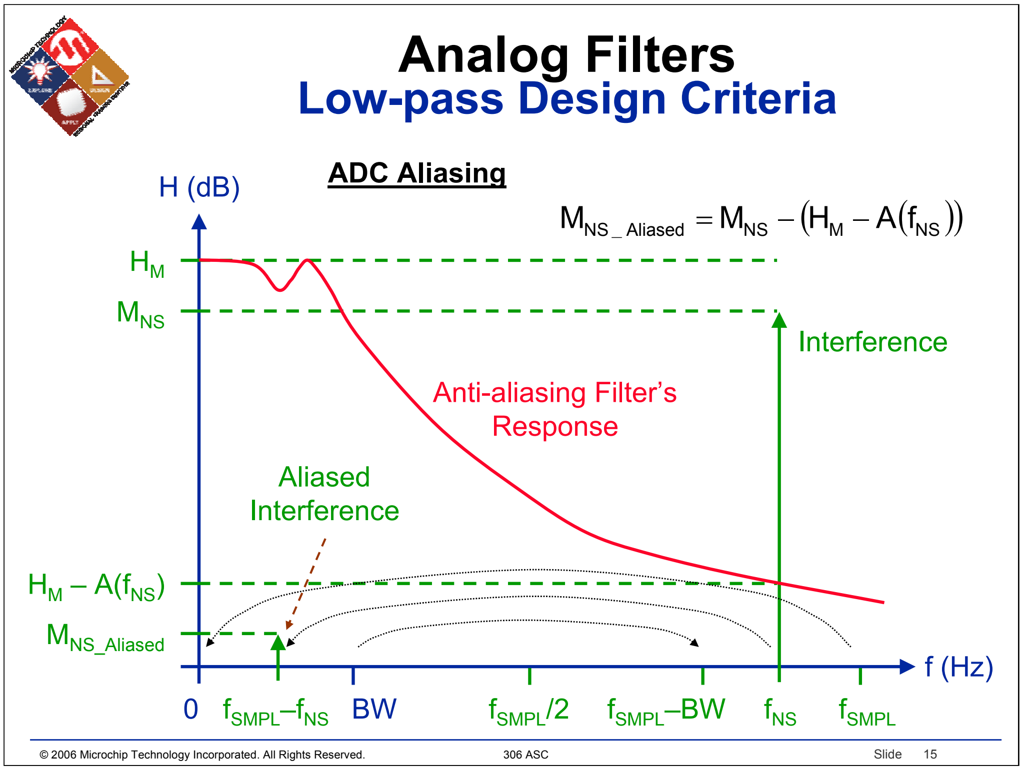
The stop band attenuation and frequency (A_S and f_S) are set based on the worst case interference or noise that needs to be rejected. Increasing A_S and/or decreasing f_S has the following effects:

- Better frequency response magnitude
- Worse time domain response
    - More peaking in group delay
    - More step response overshoot, peaking, and settling time

If possible, the transition region should be above the signal bandwidth and below any interference.



This slide combines all of the previous information; this is the big picture



The ADC's sample rate is f_{SMPL}. This picture only shows aliasing from the first band above Nyquist ($f_{SMPL}/2 \le f \le f_{SMPL}$). All tones above the Nyquist rate ($f_{SMPL}/2$) are aliased into the base band ($0 \le f_{TONE} \le f_{SMPL}/2$). It is the filter's attenuation at the original interfering tone's frequency that determines how large the aliased tone will be.

Once a tone is aliased into the sampled signal's bandwidth, it is not possible to simply filter it out using a digital filter. If the aliased tone is above the digital filter's bandwidth, then it can be attenuated further.

Analog Filters
Low-pass Design Criteria

– ADC Aliasing
- BW < $f_{SMPL}/2$ = the Nyquist Rate
- A_S (stop-band attenuation)
  – Attenuate expected interference to acceptable level
  – Set to attenuate expected interference

    A_S > (# ADC bits) (6.02 dB) – (H_{FSR} – M_{NS})

    where H_{FSR} = the ADC's full scale range
- f_S (stop-band frequency)
  – $f_S < f_{SMPL} - BW$
  – When interference may be in the transition band
    - Increase f_{SMPL} or reduce f_S

There are two common assumptions about aliasing that have the potential to cause problems with a design:

• Assumption # 1 – The anti-aliasing filter's BW should be as close to the Nyquist rate ($f_{SMPL}/2$) as possible

• This is true for systems that are not concerned about step response (e.g., settling time) and have a strongly constrained sample rate

• e.g., music CD-ROM's when they first came out

• This increases the filter order and cost

• It can be easier and cheaper to increase the sampling rate

• Assumption # 2 – This design doesn't need an anti-aliasing filter; it has been proven on the bench

• Some designs may not need an explicit filter;

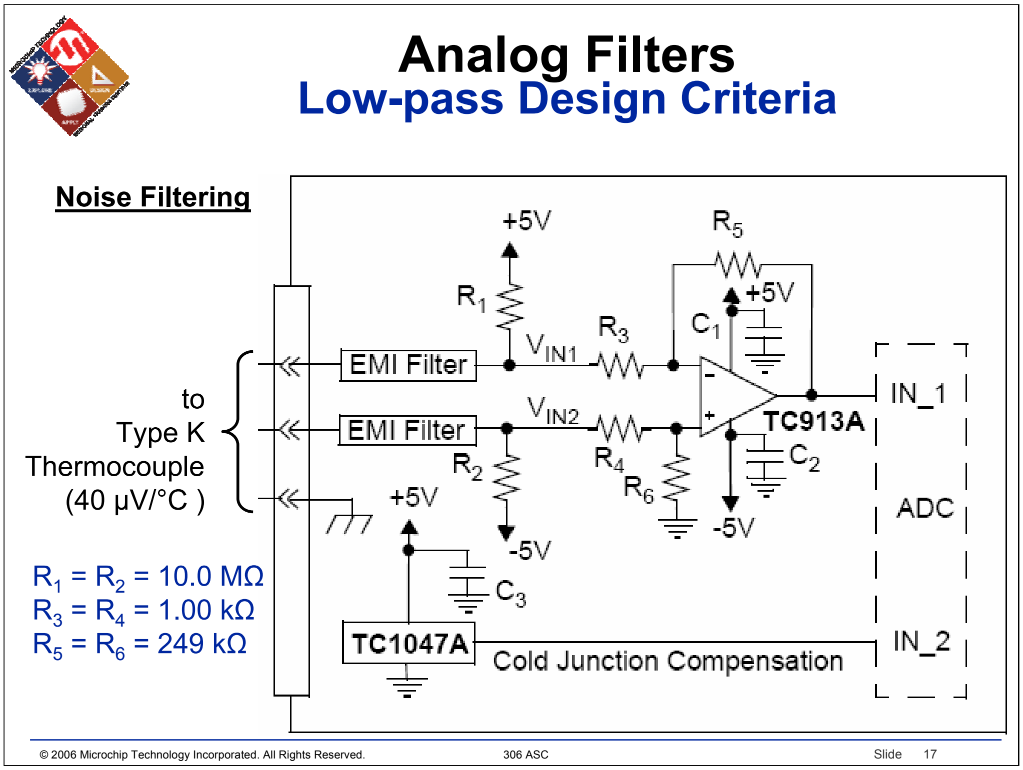
• The parasitic capacitances will attenuate the interference

• There may be almost no interference

• In today's environment, however

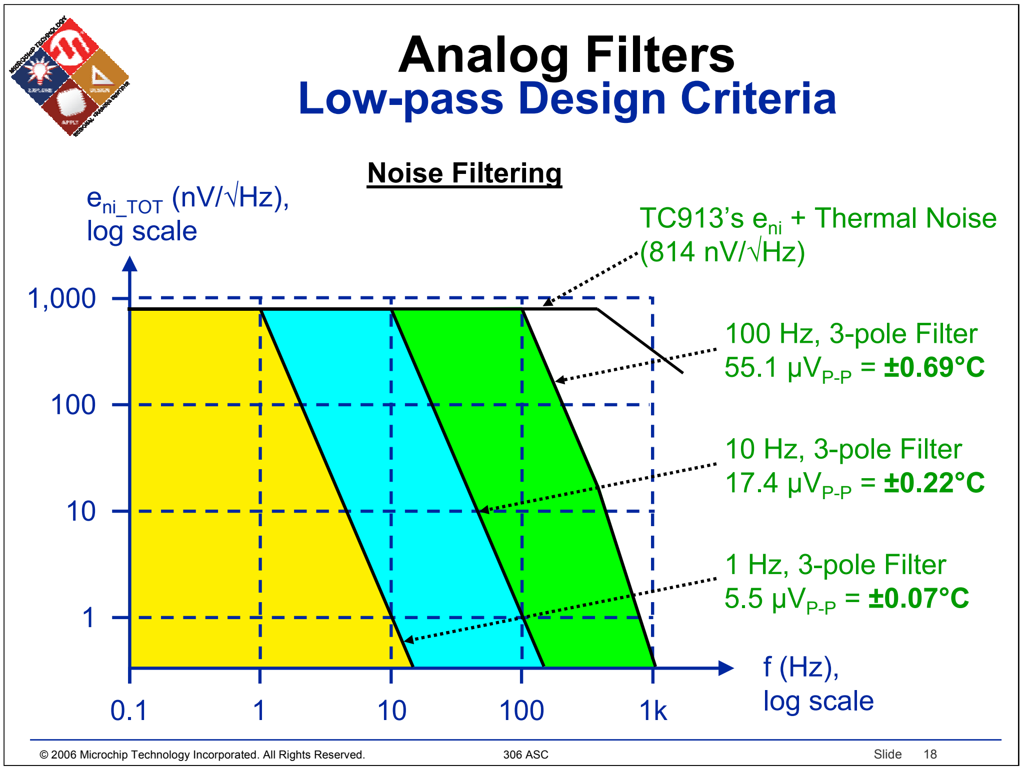
• There is a lot more noise (cell phones, PDA's, IPOD's, …)

• Simple filters can be quite low cost



This design produces device noise that is an appreciable fraction of the desired measurement accuracy. The op amp and the resistor are the sources of the device noise.

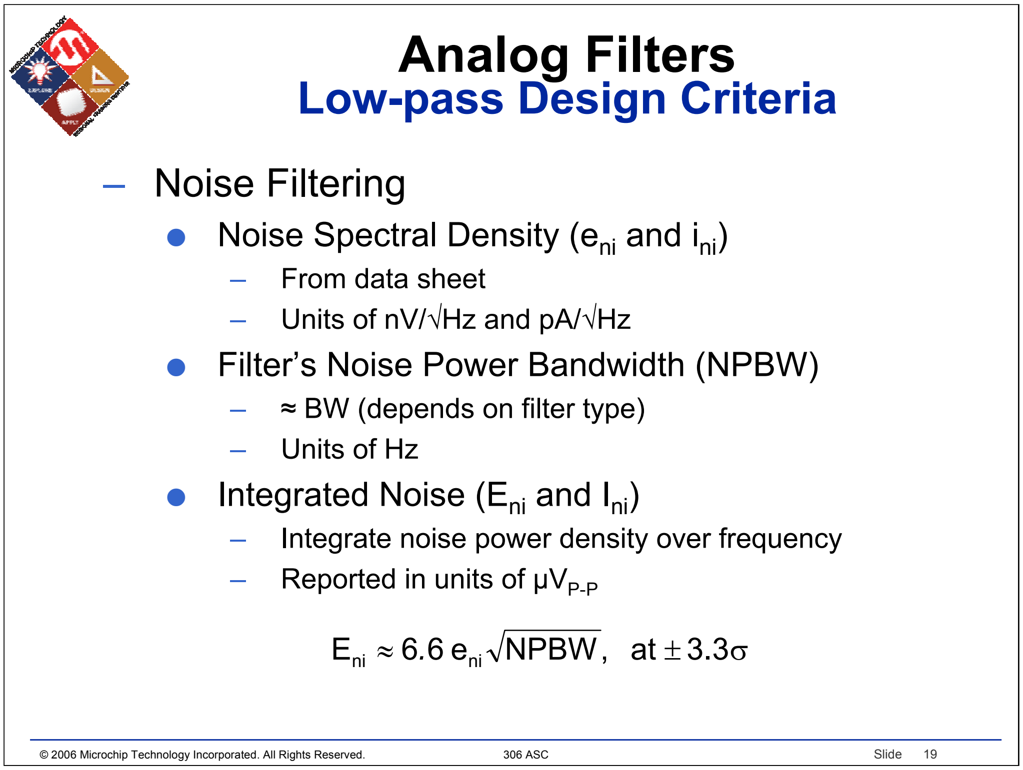
This is Figure 18 of AN929 (cited in Appendix A, References, Thermistor / Temperature).



The easiest way to analyze the device noise's contribution to the peak-to-peak output noise is based on the frequency domain. The op amp and resistors together produce a spectral density of about 814 nV/√Hz. The three colored regions show how much noise is included by the three different filters:

- The TC913 at a gain of 249 V/V has
    - BW of 6.0 kHz
    - Single pole roll-off
    - Noise Power Bandwidth (NPBW) of 9.4 kHz
- The 10 Hz, 3-pole Filter
    - Has a 3rd order Butterworth response
    - NPBW = 10.5 Hz
- The 1 Hz, 3-pole Filter
    - Has a 3rd order Butterworth response
    - NPBW = 1.05 Hz

For many applications using a thermocouple, any one of the three filters would keep the device and thermal noise low enough. The choice would depend more on interference (e.g., mains or line power) and the step response rise time.

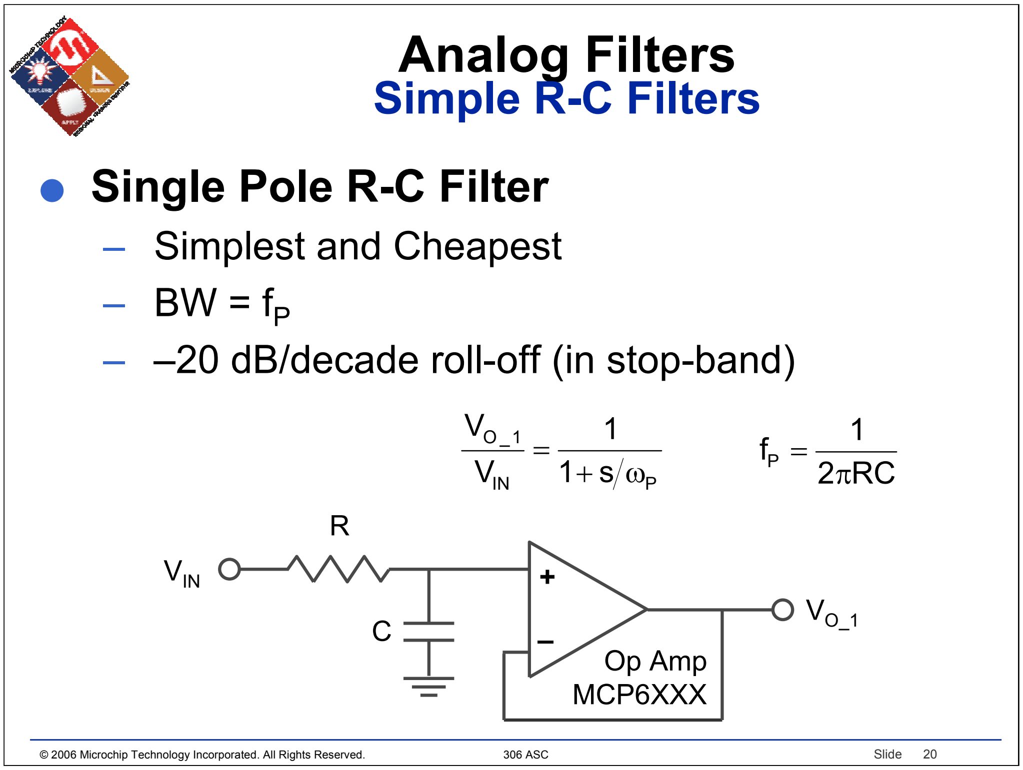


Some typical NPBW / BW ratios follow (assuming constant input noise spectral density):

NPBW / BW	Description
1.57	Single pole, low-pass filter
1.11	n = 2, Butterworth filter
1.05	n = 3, Butterworth filter
1.03	n = 4, Butterworth filter
1.02	n = 5, Butterworth filter

We use a crest factor of ±3.3σ for our integrated noise specs. It is not necessary to change the crest factor (CF) much to achieve significantly different failure rates (fraction of time CF is exceeded), assuming a stationary, Gaussian random distribution:

CF	Failure Rate (fraction of time CF is exceeded)
±2.58σ	1% = 10,000 ppm
±3.29σ	**0.1% = 1,000 ppm**
±3.89σ	0.01% = 100 ppm
±4.42σ	0.001% = 10 ppm
±4.89σ	0.0001% = 1 ppm
±5.33σ	0.00001% = 0.1 ppm

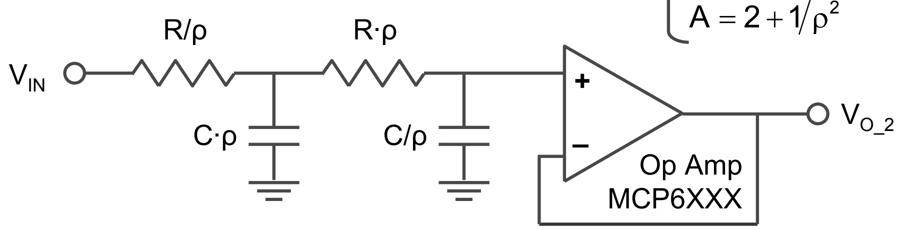


This low-pass filter is widely used; it effectively reduces interference and noise in many DC and base band applications. Adding an op amp buffer can be very helpful when V_{O_1} is connected to a significant load that could affect the filter response.

Analog Filters
Simple R-C Filters

- **Double Pole R-C Filter**
  - BW < f_P
  - –40 dB/decade roll-off (in stop-band)
  - Resistor ratio = Capacitor ratio = ρ^2

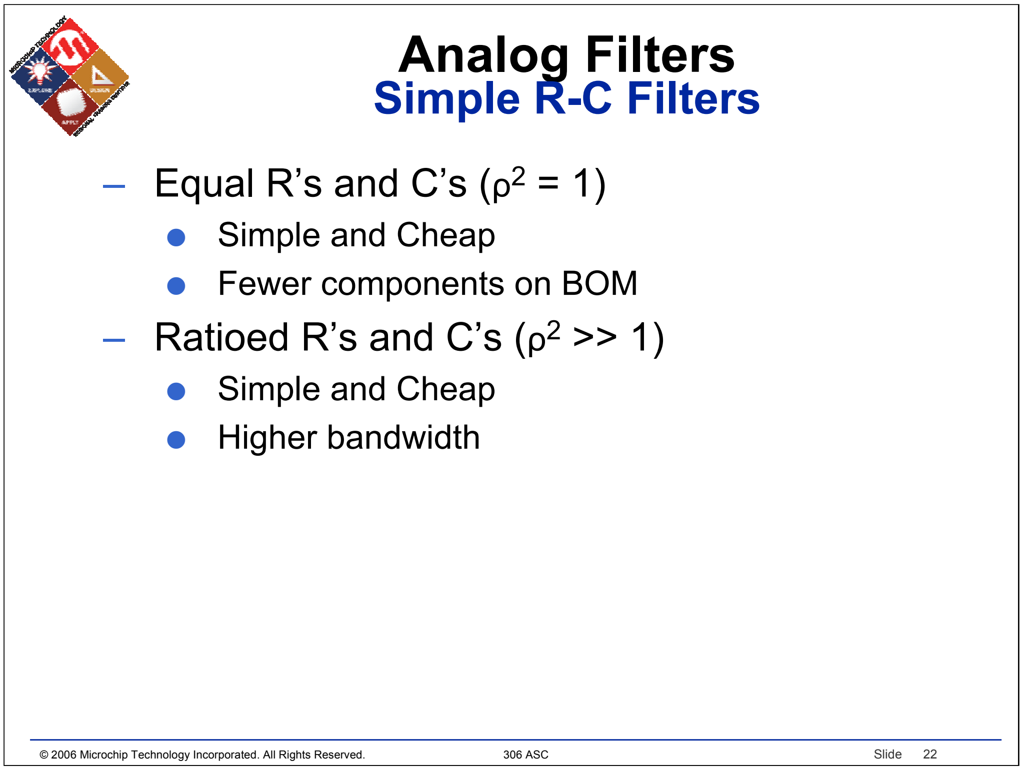
$$\frac{V_{O\_2}}{V_{IN}} = \frac{1}{1 + A(s/\omega_P) + (s/\omega_P)^2} \qquad \begin{cases} f_P = \dfrac{1}{2\pi RC} \\ A = 2 + 1/\rho^2 \end{cases}$$



 306 ASC Slide 21

This filter is less commonly used, but is still effective at reducing interference and noise in many DC and base band applications. Adding an op amp buffer can be very helpful when V_{O_2} is connected to a significant load that could affect the filter response.

The impedance ratio (ρ^2) has a significant impact on the transfer function; the smaller it is, the higher the BW is (see the next two slides).



Most designers choose equal R’s and C’s because it makes their life easier.

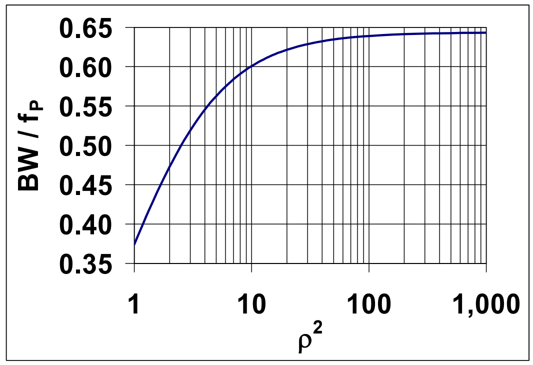
Making $\rho^2 >> 1$ produces better performance when the –3dB bandwidth is a critical spec. This happens because the first R-C pair looks more like a low impedance (voltage) source for the second R-C pair. The downside is having more parts on the BOM (Bill of Materials).

Note that making $\rho^2 < 1$ is counterproductive.

Analog Filters
Simple R-C Filters

$$A = 2 + \frac{1}{\rho^2}$$

$$BW = \frac{f_P}{\sqrt{\left(\frac{A^2}{2} - 1\right) + \sqrt{1 + \left(\frac{A^2}{2} - 1\right)^2}}}$$



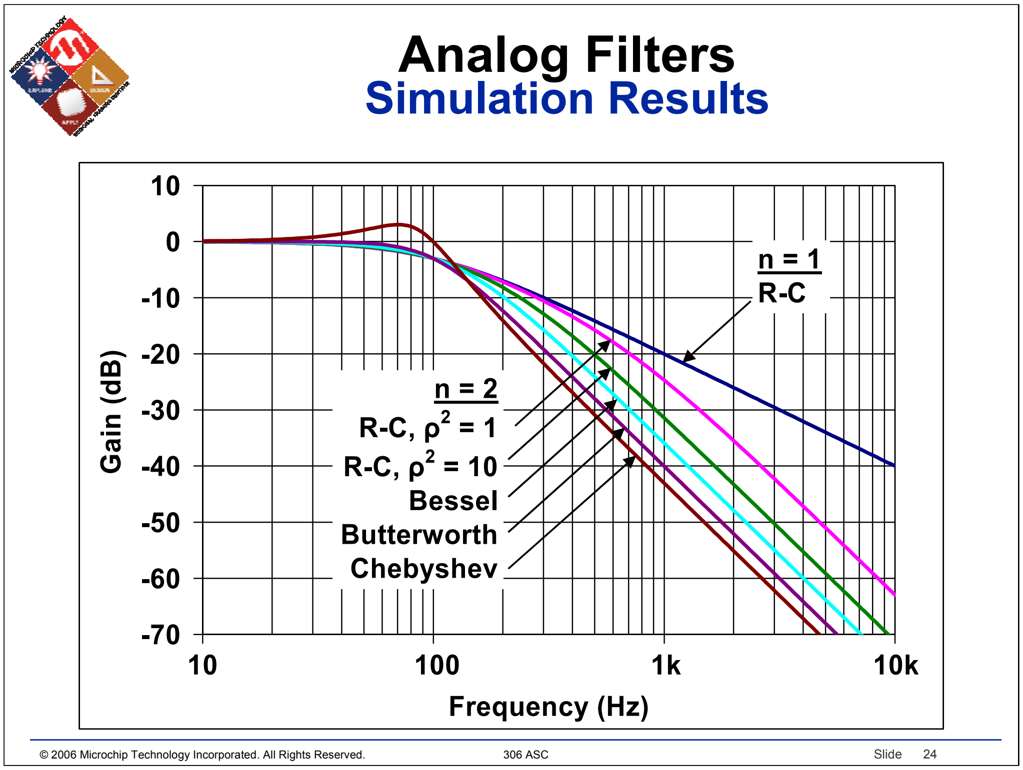
 306 ASC Slide 23

Increasing ρ^2 to 10 gives most of the benefit, with a reasonable component ratio:

ρ^2	BW / f_P	$BW / f_P / 0.644$	
1	**0.374**	**0.581**	**(–41.9%)**
2	0.473	0.735	(–26.5%)
5	0.563	0.874	(–12.6%)
10	**0.601**	**0.933**	**(–6.7%)**
20	0.622	0.966	(–3.4%)
50	0.635	0.986	(–1.4%)
100	0.639	0.993	(–0.7%)
infinity	0.644	1.000	(0%)

As was mentioned before, making $\rho^2 < 1$ is counterproductive:

ρ^2	BW / f_P
0.1	0.00980
$\to 0$	ρ^2



These filters were designed to have f_P = 100 Hz and A_P = –3.0 dB. The attenuation well into the stop band is:

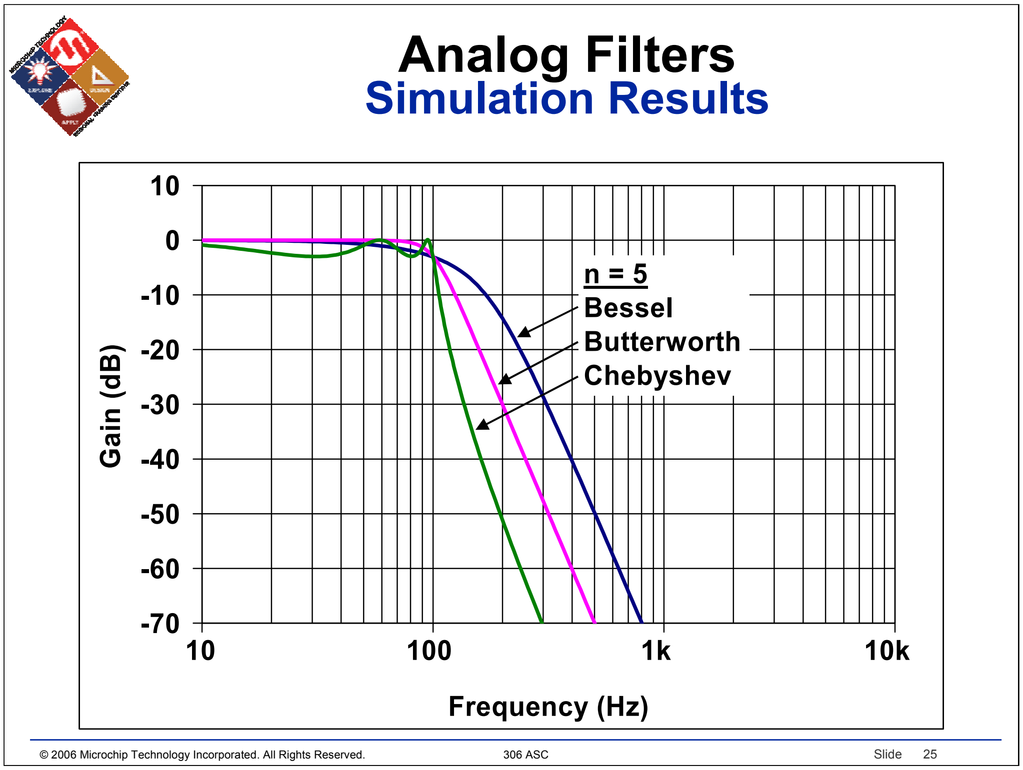
$$n\,[\,20\log(f/f_P) + A_{EX}\,]$$

where:

n = # Poles	Filter	A_{EX} = "Excess" Attenuation
1	R-C	0.0 dB
2	R-C, $\rho^2 = 1$	–7.6 dB
2	R-C, $\rho^2 = 10$	–4.3 dB
2	Bessel	–2.0 dB
2	Butterworth	0.0 dB
2	Chebyshev	1.5 dB

Notes:

(1) The Chebyshev was designed with the Sallen-Key topology, with a gain of 1 V/V (0 dB). FilterLab, in this case, forces the DC gain to be 0 dB, so H_M is 3.0 dB. This choice was made because a unity gain buffer configuration has significant implementation advantages.



These filters were designed to be 5th order, and to have f_P = 100 Hz. The Chebyshev filter has a pass band ripple of A_P = 3.0 dB to make the comparisons easier. Chebyshev filters with lower A_P, and the same BW, will be closer to the Butterworth in frequency domain performance.

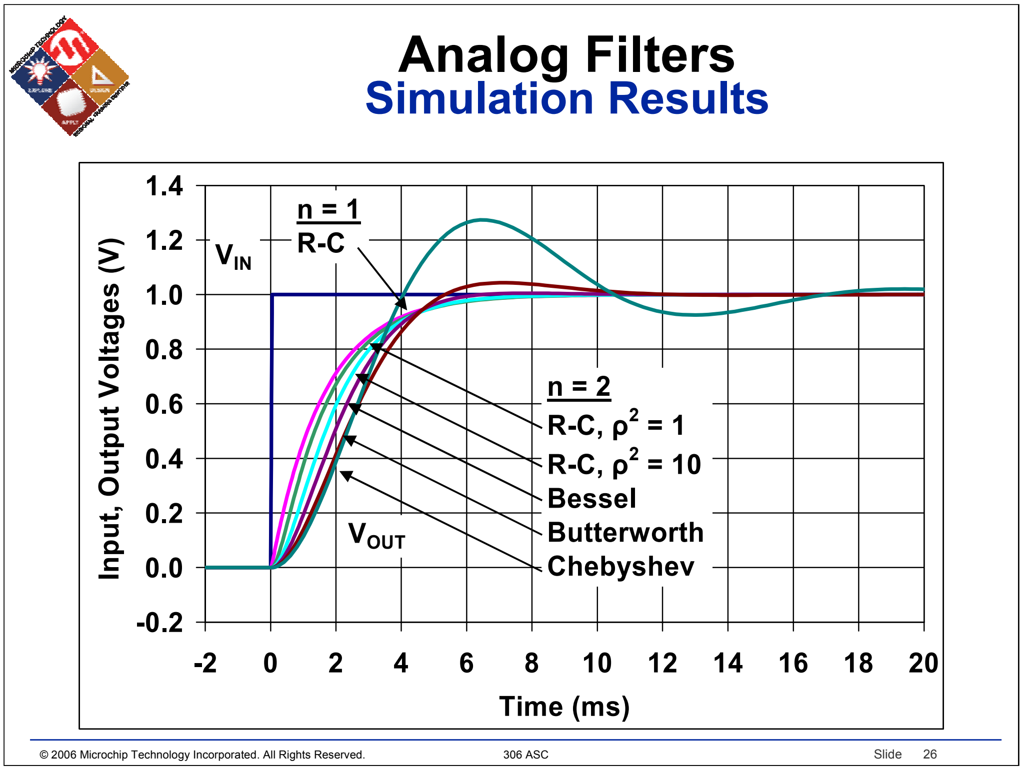
The trade-off is in the time domain, as will be seen on the next two slides.

The attenuation well into the stop band is:

$$n\,[\,20 \log(f / f_P) + A_{EX}\,]$$

where:

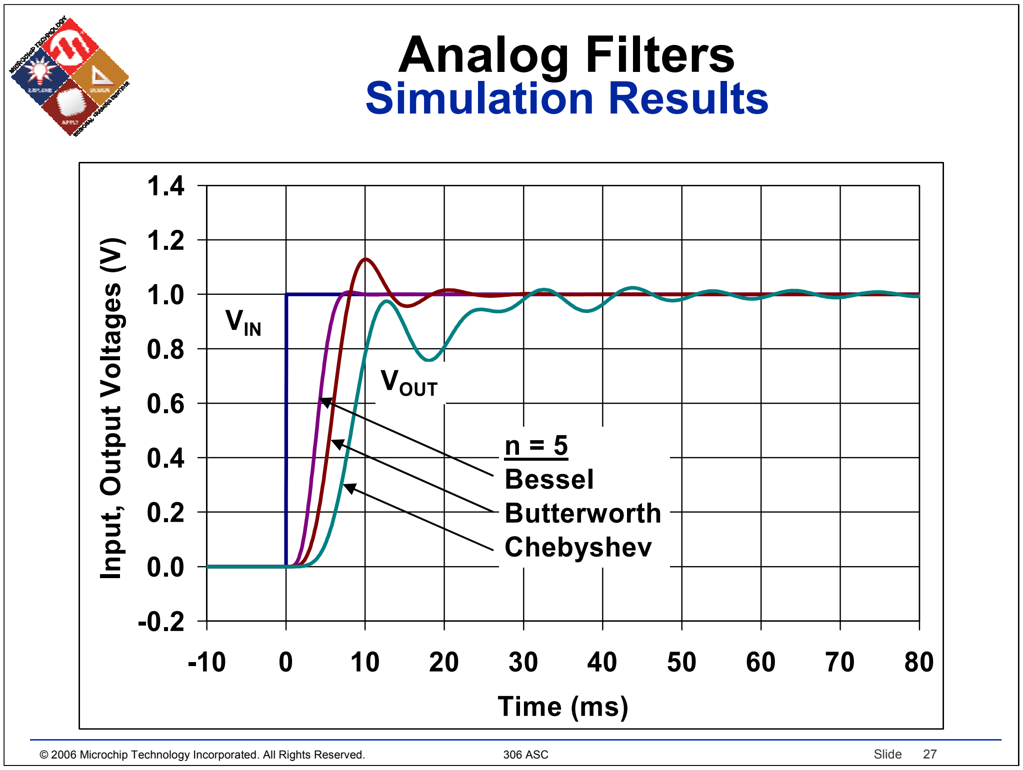
n = # Poles	Filter	A_{EX} = "Excess" Attenuation
5	Bessel	–4.1 dB
5	Butterworth	0.0 dB
5	Chebyshev	4.8 dB



All of these filters show about the same rise time, but the propagation delays and overshoot vary considerably.

Filter	Prop. Delay	Rise Time	Overshoot
n = 1, R-C	1.13 ms	3.50 ms	0%
n = 2, R-C, $\rho^2 = 1$	1.35 ms	3.49 ms	0%
n = 2, R-C, $\rho^2 = 10$	1.68 ms	3.45 ms	0%
n = 2, Bessel	1.98 ms	3.41 ms	0.5%
n = 2, Butterworth	2.31 ms	3.43 ms	4.3%
n = 2, Chebyshev	2.36 ms	2.74 ms	27.3%

Filter	Undershoot	1% Settling	0.1% Settling
n = 1, R-C	0%	7.3 ms	11.0 ms
n = 2, R-C, $\rho^2 = 1$	0%	7.4 ms	11.0 ms
n = 2, R-C, $\rho^2 = 10$	0%	7.0 ms	10.1 ms
n = 2, Bessel	0.002%	5.8 ms	11.0 ms
n = 2, Butterworth	0.2%	10.5 ms	16.4 ms
n = 2, Chebyshev	7.5%	21.8 ms	34.2 ms



All of these filters show greater variation in the rise time, and the propagation delays and overshoot/undershoot vary considerably.

Filter	Prop. Delay	Rise Time	Overshoot
n = 5, Bessel	3.88 ms	3.54 ms	0.8%
n = 5, Butterworth	5.59 ms	4.08 ms	12.8%
n = 5, Chebyshev	8.25 ms	5.86 ms	2.4%

Filter	Undershoot	1% Settling	0.1% Settling
n = 5, Bessel	0.2%	6.9 ms	11.7 ms
n = 5, Butterworth	4.4%	22.1 ms	32.9 ms
n = 5, Chebyshev	24.3%	32.9 ms	137.4 ms

Analog Filters
Low-pass Design Criteria

- **Choosing the Approximation Function**

Approximation	Frequency Response	Step Response	
	Transition Ratio (f_S/f_C)	Overshoot	Settling Time
R-C	Higher	None	Lower
Bessel	High	Low	Low
Butterworth	Reasonable	Reasonable	Reasonable
Chebyshev	Low	High	High

 306 ASC Slide 28

This table gives a rough summary of what the previous slide have shown.

It can be argued that a particular design might compare a 3rd order Chebyshev against a 5th order Butterworth, for instance, and so the tradeoffs shown above would be different. This argument has been ignored here for the simple reason that most designs cannot afford the cost of one more filter section (1 op amp, several R's and C's); most designs are constrained more by the cost than by the performance.

Analog Filters
FilterLab® Software V2.0

- **Introduction to FilterLab Software V2.0**
  - Active Filters
  - Design Program
    - Download at www.microchip.com
    - Free
  - Supports most common filters
  - Practical implementations

This free tool can be a significant help in designing the most common filters. It has been set up to simplify the task, and to give design values that should work in actual circuits

Many textbook circuits and/or design procedures overlook the practical issues; they are more focused on the math and network theory (no author can cover everything). For this reason, some of their solutions have a much narrower application than is stated. One of FilterLab's goals is to avoid many of the common pitfalls.

FilterLab also saves the designer the tedious process of working through many complex mathematical equations before reaching a possible implementation.

Analog Filters
Exercise 1

- **Exercise 1**
  - FilterLab® Software
  - See Handout

 306 ASC Slide 30

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See application notes AN248 and AN867, and two Web Seminars (cited in Appendix A: References, PGA).

Two evaluation boards (MCP6S22 PGA PICtail™ Demo Board and MCP6S2X PGA Evaluation Board , cited in Appendix A: References, Demo Boards) are available for evaluating these PGA’s. Their User’s Guides (DS51481 and DS51327, cited in Appendix A: References, Demo Boards) help obtain the most from these boards.

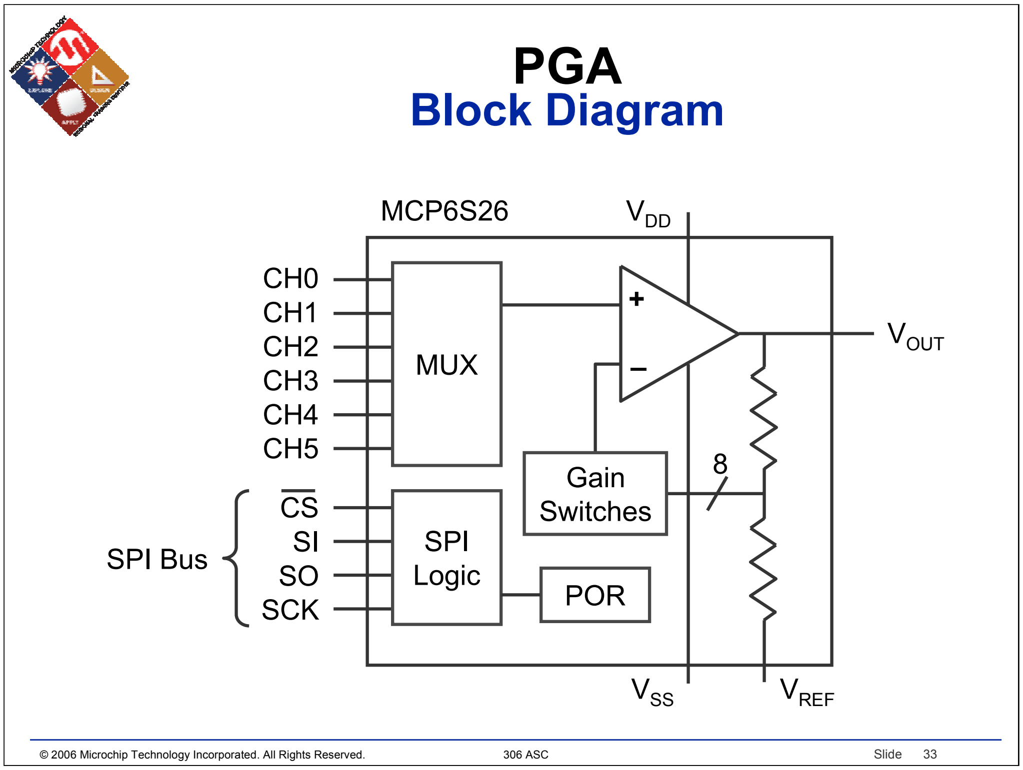
PGA
Description

- **MCP6S26**
  - 6 Multiplexed Channels
  - 8 Gain Settings:
    - +1, 2, 4, 5, 8, 10, 16, and 32 V/V
  - Serial Peripheral Interface (SPI)

 306 ASC Slide 32

This family of PGA’s benefits greatly from the combination of MUX’ed inputs and digitally selected gains; each input can be individually selected along with the most appropriate gain.

The SPI interface simplifies the task of communicating with these devices.



The only comments on this diagram that need to be made are:

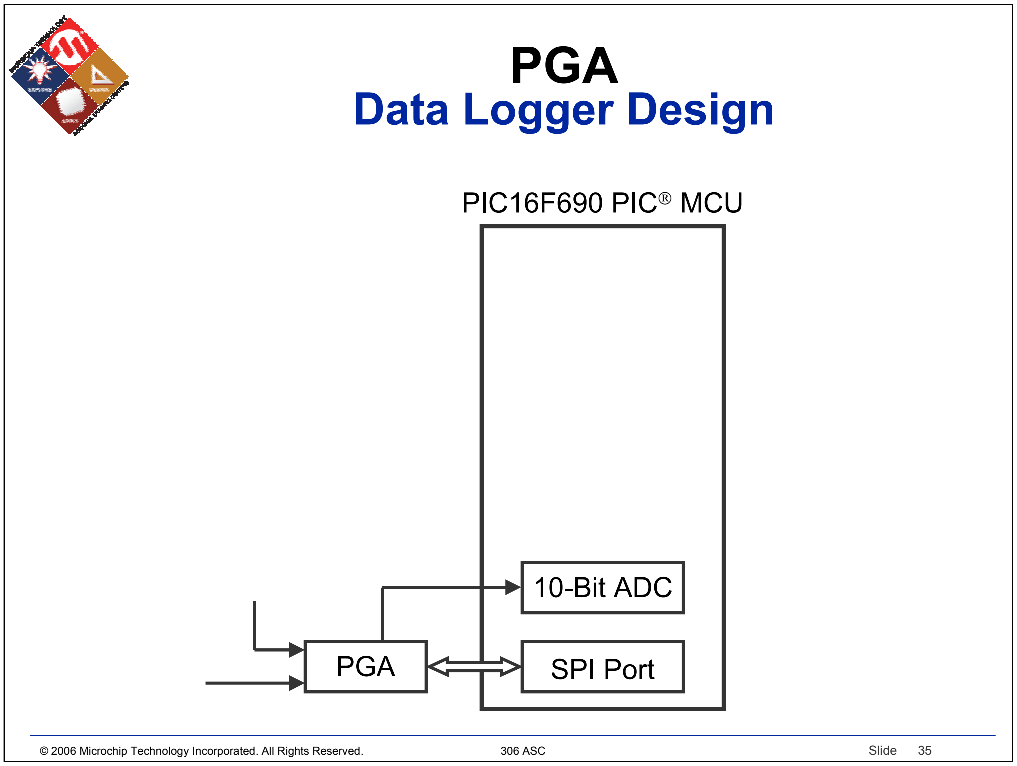
- The gain switches are outside the resistor ladder's current path
  - They can be quite small
  - They have very small parasitic capacitance
  - They have little impact on the bandwidth or gain peaking
- The compensation capacitor has three values, depending on gain, to optimize the bandwidth
- The internal POR ensures the PGA powers up in a known state

PGA
Data Logger Design

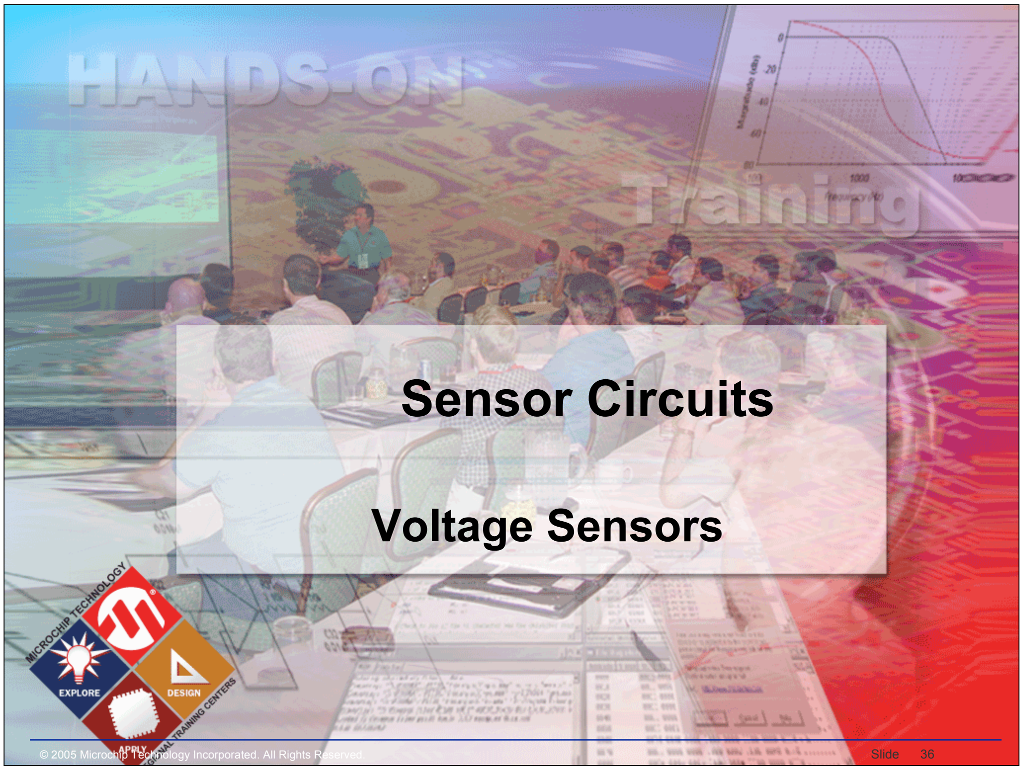
- **Introduction**
  – PGA as an analog MUX (many sensors)
  – PIC® MCU for digital smarts
  – Memory for data logging sensor outputs
- **External Components:**
  – PGA
- **PIC MCU Resources:**
  – ADC
  – 4 GPIO pins
  – Firmware routines

 306 ASC Slide 34

This presentation discusses several pieces that are part of our “Data Logger Design.” This design includes three sensors: a thermistor (temperature), a photodiode, and a humidity (capacitive) sensor. The PGA allows multiplexing of multiple inputs, and the selection of appropriate gains for each sensor. The signal is acquired (usually by using the PIC’s internal ADC), processed, and the results are stored on two SPI memory chips. The clock is driven by a watch crystal so that the PIC conserves power.



This is the block diagram of what we have been discussing.



See application note AN990 (cited in Appendix A: References, General Topics); it is the source for most of the circuits shown here.

Many sensors have been left off, such as:

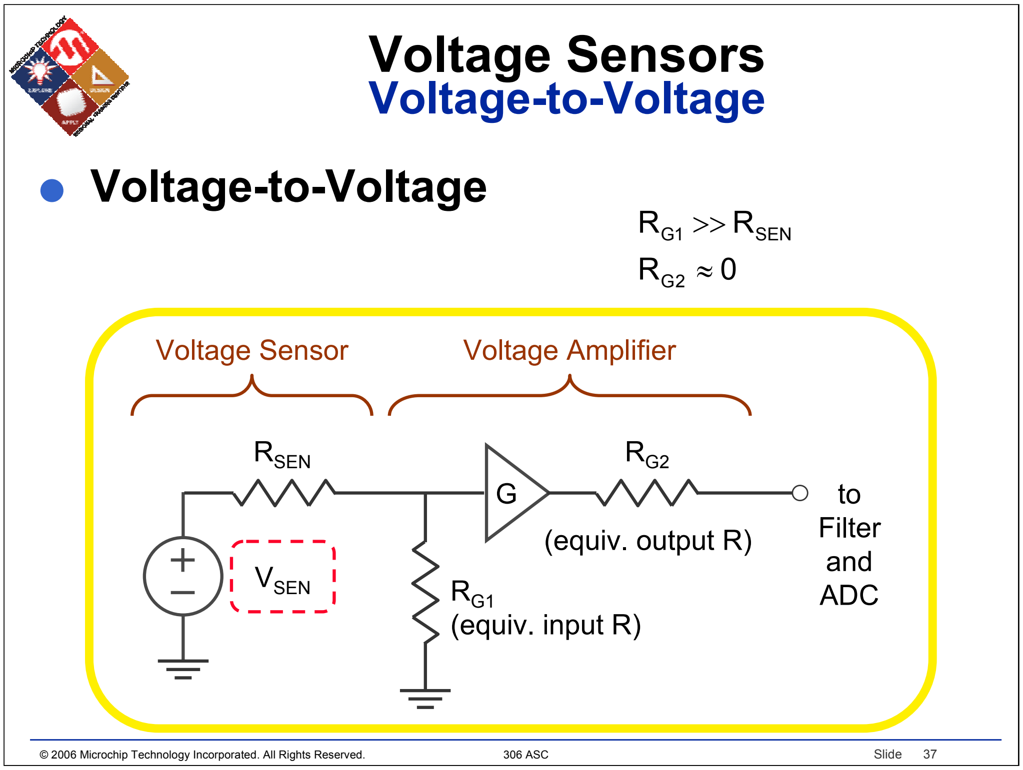
• LVDT and Piezo-electric motion sensors

• Ultrasonic sensors

• IC temperature sensors

• Mass flow meter

This presentation is intended as an introduction and overview, not as an exhaustive reference.



An ideal voltage sensor (source) has zero output impedance. For practical purposes, R_{SEN} only needs to be much less than the input impedance of the signal conditioning circuitry.

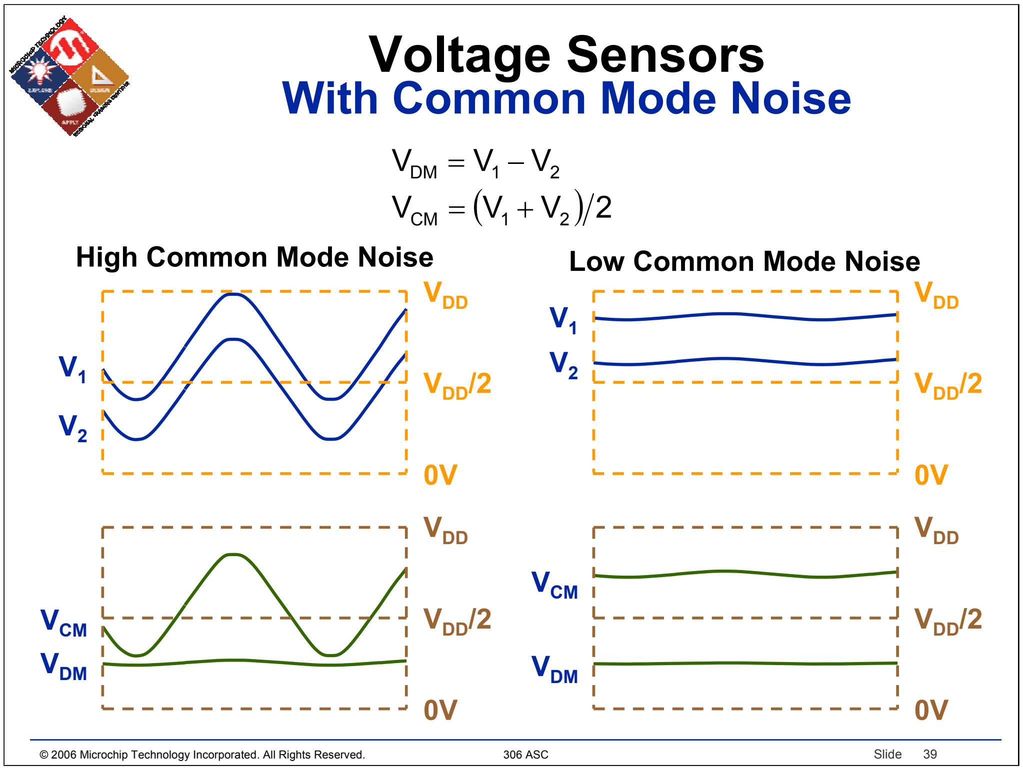
Most voltage amplifiers, including op amps, will have very low output impedance.

Voltage Sensors
With Common Mode Noise

- **High Common Mode Noise**
  - Remote sensors, typically
  - Thermocouple
  - RTD
  - Wheatstone Bridge
    - Strain Gage
    - Pressure Sensor
  - High side current sensing

 306 ASC Slide 38

"Remote sensors" are not on the PCB with the conditioning circuit. An example would be a temperature sensor (i.e., thermocouple) for an engine.



The plots are arranged as follows:

- Left Column – For high common mode noise / remote sensors
- Right Column – For low common mode noise / local sensors
- Top Row – For original two signals
- Bottom Row – For derived common mode and differential mode signals

Since difference and instrumentation amplifiers reject the DC common mode component shown above, it is not usually considered a part of the common mode noise.

Any sensor with differential output is subject to common mode noise. Examples include Wheatstone bridges and Hall effect sensors. Common mode noise is reduced by shielding, PCB layout, and using a difference or instrumentation amplifier.

Power supply noise also affects the results through an analog part's PSRR. This phenomenon is reduced by better power supply filtering and bypassing, and by using better analog parts.

Filtering can also reduce the common mode, or power supply, noise. It is best to remove as much as possible before the filter.

Voltage Sensors
With Common Mode Noise

$$V_{OUT} \approx \frac{R_2}{R_1} \cdot (V_{SEN+} - V_{SEN-})$$

$$CMRR \le \frac{1}{\dfrac{1}{CMRR_{OA}} + 2 \cdot TOL_R}$$

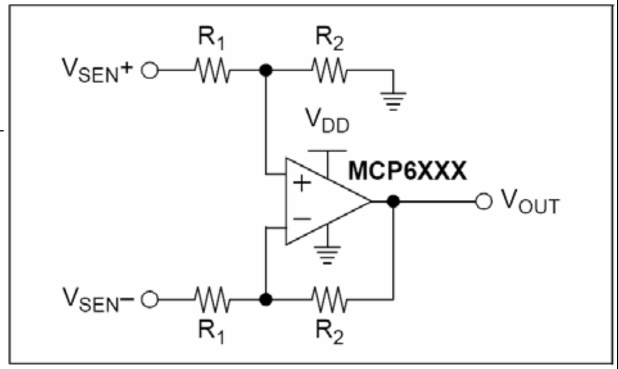


FIGURE 4: *Difference Amp.*

This difference amplifier rejects common mode signals. Imagine the case where $V_{SEN}+ = V_{SEN}-$. Because the voltage dividers top and bottom are the same, and the op amp’s inputs are at the same voltage (“virtual short”), we must have V_{OUT} = 0V.

To a first order approximation, the output is a function of the input difference voltage, but not the common mode input voltage, as can be seen by the top equation.

The second equation give a more practical result. If the op amp’s CMRR ($CMRR_{OA}$) is given in V/V (e.g., 80 dB is converted to 10,000 V/V), and the resistor tolerance (TOL_R) is given in absolute terms (e.g., 0.1% becomes 0.001), then the total effective CMRR will be in V/V (for the examples already given, 476 V/V = 54 dB).

Voltage Sensors
With Common Mode Noise

- Pros:
  - Simple
  - Allows high input voltages
- Cons:
  - Loads sensor
  - Unbalanced loading of input
- PIC® MCU Resources Required:
  - 1 GPIO pin
  - ADC

 306 ASC Slide 41

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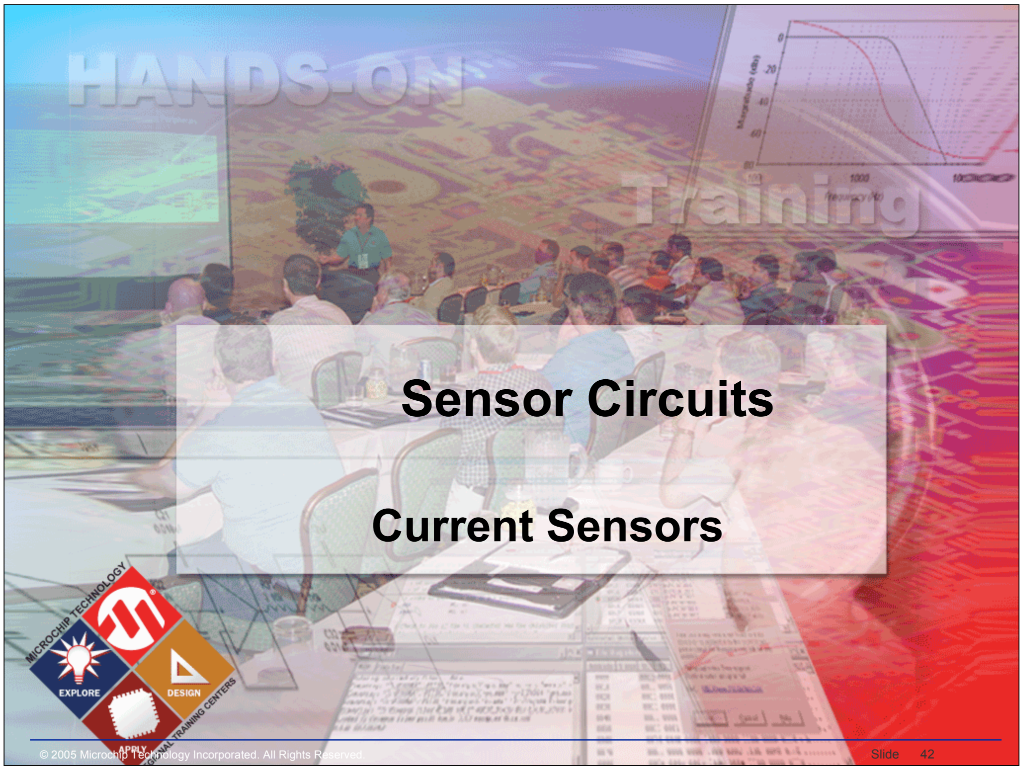
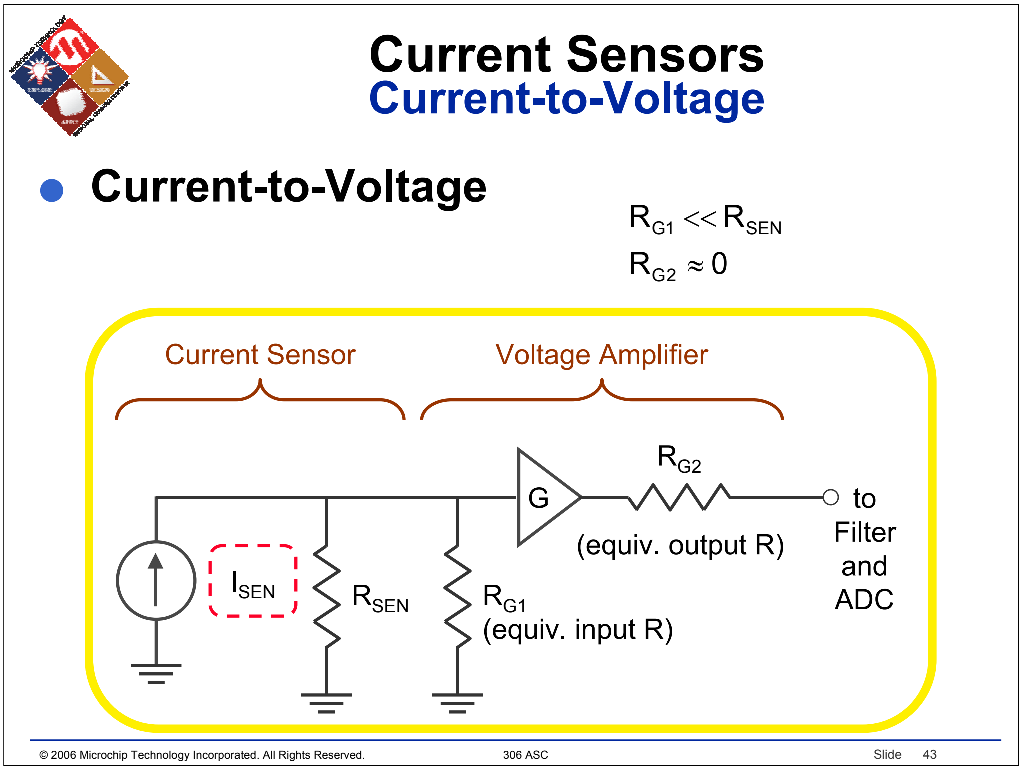


See application note AN990 (cited in Appendix A: References, General Topics); it is the source for most of the circuits shown here.



R_{G1} needs to be much less than R_{SEN}. The resistor R_{G1} describes one of two situations:

- R_{G1} converts I_{SEN} to a voltage which is gained up by the amplifier
- R_{SEN} represents a very low equivalent input impedance to a transimpedance amplifier (converts current to voltage); this will be described shortly

Current Sensors
Transimpedance Amplifier

- **Transimpedance Amplifier**
  - Rain Detector for car wipers
  - IR Smoke Detector
  - Motion Detector
  - Photodiode

The op amp that forms the transimpedance amplifier forces the current sensor's output to be at a fixed voltage. This bootstraps the sensor's output impedance so that it does not affect the measurement.

Current Sensors
Transimpedance Amplifier

$$V_{OUT} = -I_{SEN}R_1 + \frac{V_{DD}}{2}$$

$$BW = \frac{1}{2\pi R_1 C_1}$$

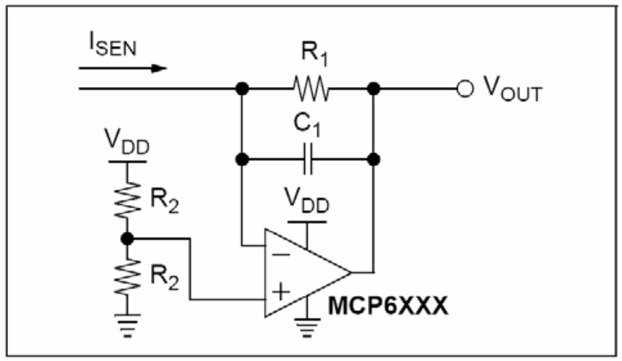


FIGURE 8: *Transimpedance Amplifier.*

 306 ASC Slide 45

The non-inverting input of the op amp is at mid-supply ($V_{DD}/2$). The op amp's "virtual short" forces the inverting input to also be at $V_{DD}/2$. Thus, I_{SEN} is forced to enter a circuit node with zero resistance; this is the ideal case for conditioning current sources. R_1 convert I_{SEN} to a voltage. C_1 and R_1 cause a low-pass pole, and C_1 stabilizes the amplifier (see AN951 cited in Appendix A, References, Photodiode).

Current Sensors
Transimpedance Amplifier

- Pros:
  - Simple
  - Flexible
  - Bootstraps sensor’s output impedance
- Cons:
  - May need stabilization
- PIC® MCU Resources Required:
  - 1 GPIO pin
  - ADC

 306 ASC Slide 46

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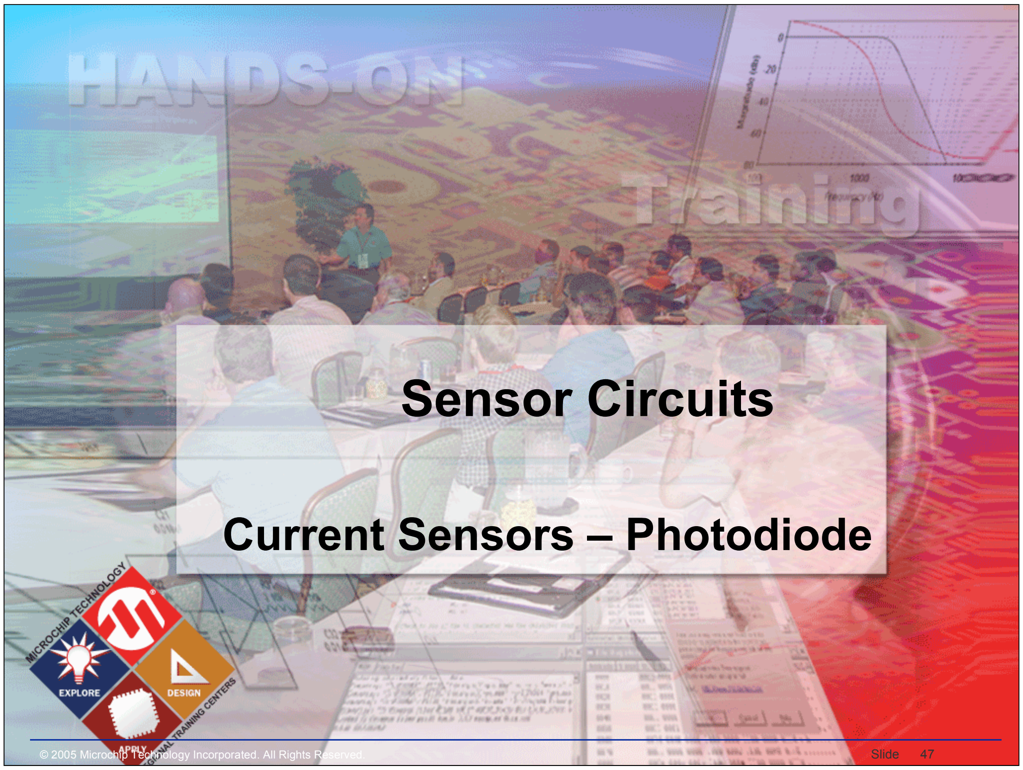
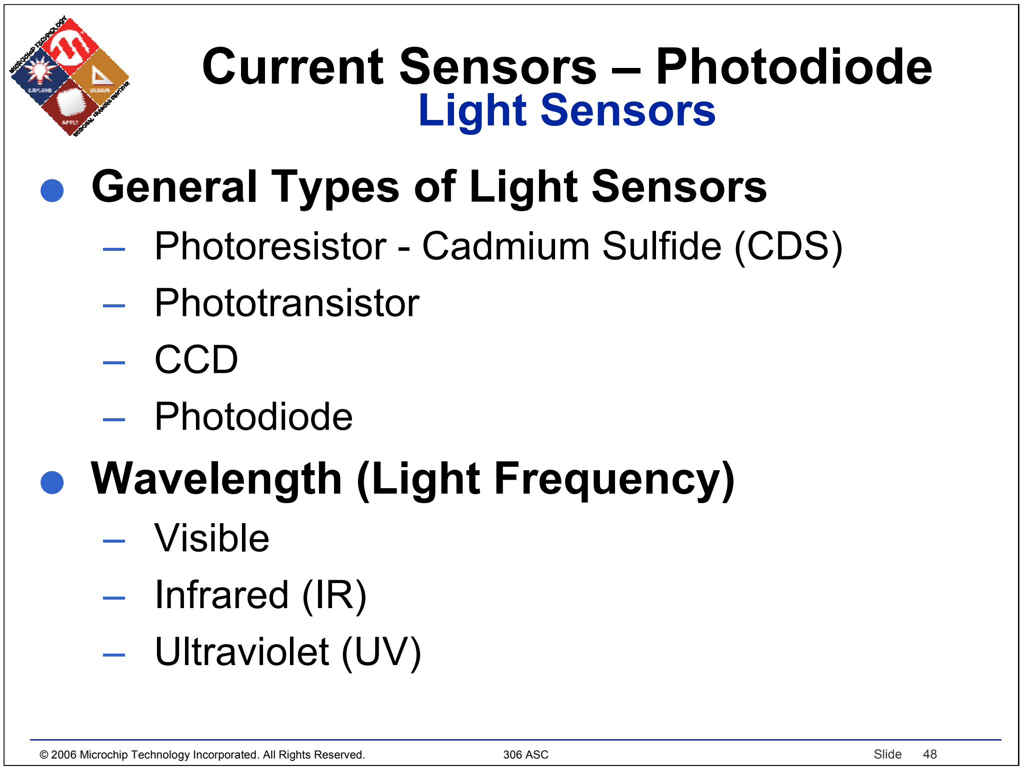


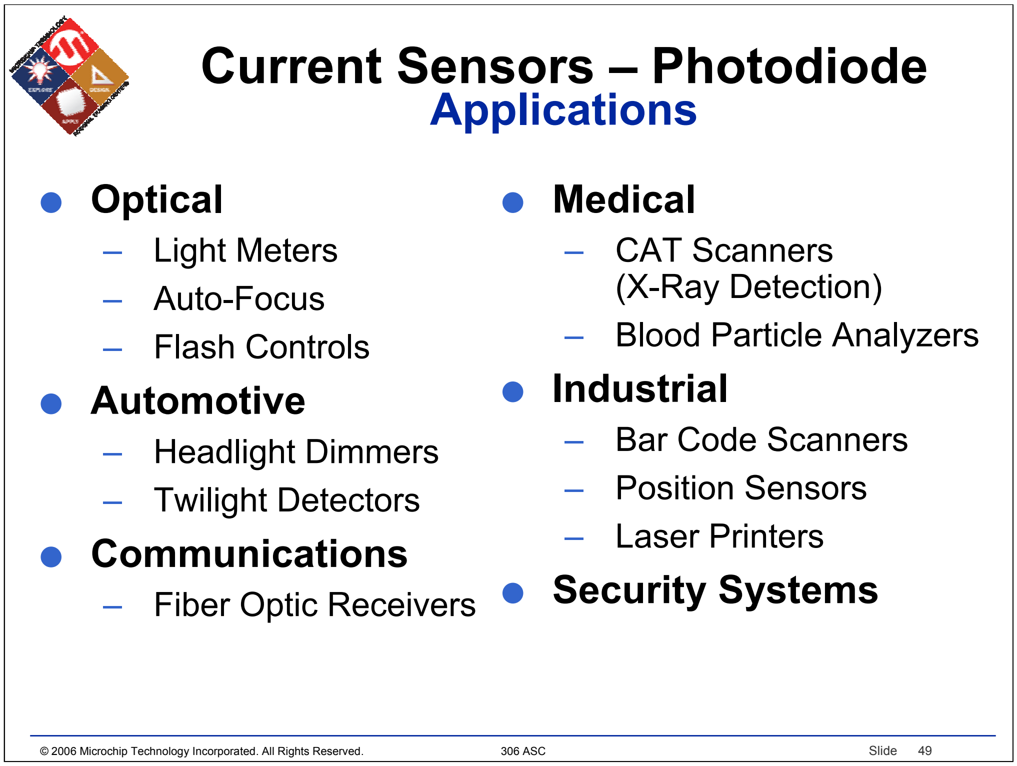
See application notes AN682, AN692, and AN951 (cited in Appendix A: References, Photodiode).

The companion demo board (MCP6SX2 PGA Photodiode PICtail™ Demo Board , cited in Appendix A: References, Demo Boards) is used to demonstrate this application in this class. Its User's Guide (DS51514, cited in Appendix A: References, Demo Boards) gives more information on the board used for the measurements in AN951.

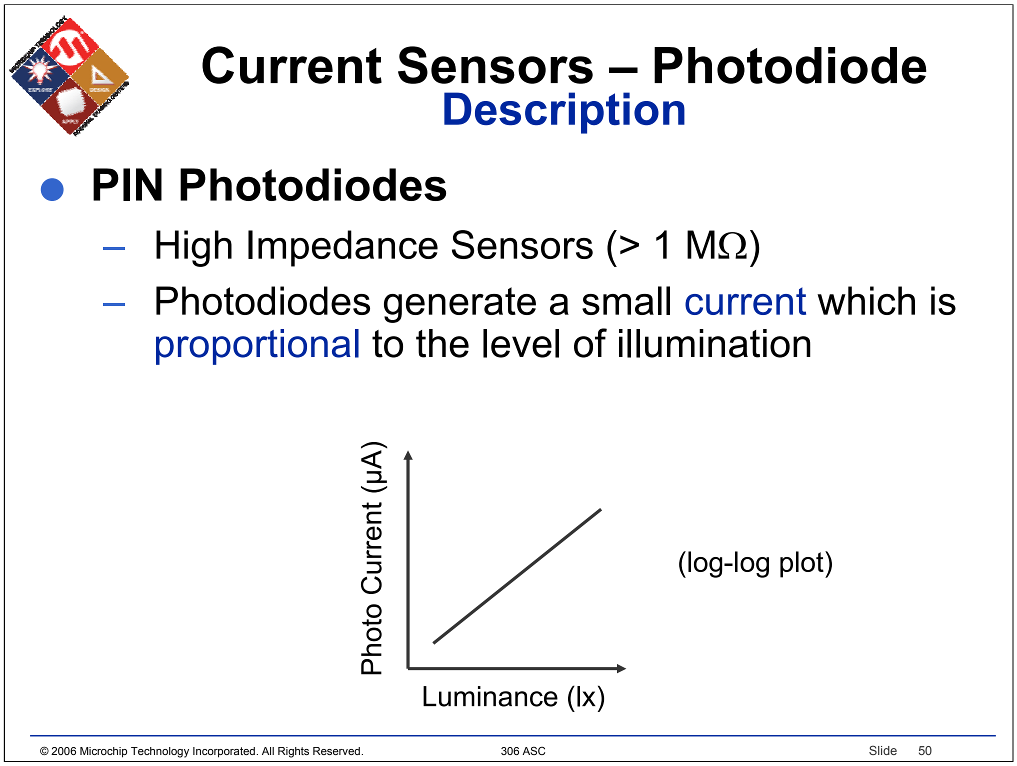


Here we give an overview of light sensors.

CCD = Charge Coupled Devices



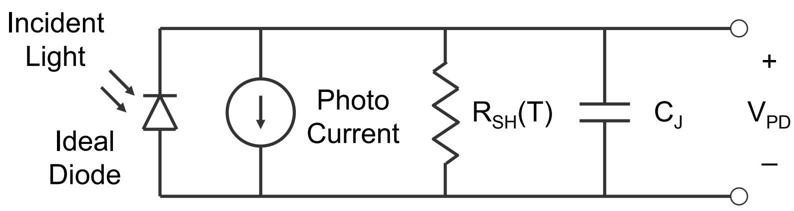
These are just a few of the many applications in use today.



PIN stands for P-type silicon, Insulator, N-type silicon; the very thin insulator layer is used to convert light to electrical energy.

This plot could just as well be a linear-linear plot; it is shown as log-log in order to emphasize that it operates over many orders of magnitude of light (luminance).

Current Sensors – Photodiode
Equivalent Circuit



– Short Circuit Current
- Linear over 6 to 9 decades of light intensity
- Often measured for absolute light levels

 306 ASC Slide 51

This is a simplified model of the PIN photodiode. The ideal diode models its current vs. voltage and temperature behavior. The photo current source models the charges that are freed by incident light energy. The shunt resistor (R_{SH}) models the dynamic resistance ($\Delta V/\Delta I$) of the diode junction; it is strongly dependent on temperature. The parallel capacitance (C_J) models the diodes junction capacitance; this will be the depletion capacitance only when being used as a photodiode.

Current Sensors – Photodiode
Equivalent Circuit

- Open Circuit Forward Voltage Drop
  - Varies logarithmically with light level
  - Large temperature coefficient
  - Seldom used
- Shunt Resistance – $R_{SH}(T)$
  - Usually about 1000 MΩ at room temperature
  - Decreases by a factor of 2 for every 10°C rise in temperature
- Diode Junction Capacitance – C_J
  - Function of junction area and diode bias voltage
  - Typical small area diodes at zero bias ≈ 50 pF

 306 ASC Slide 52

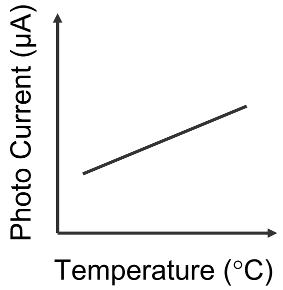
The forward voltage drop depends on the diode's saturation current, area, and temperature. It is difficult to obtain reasonable accuracy this way.

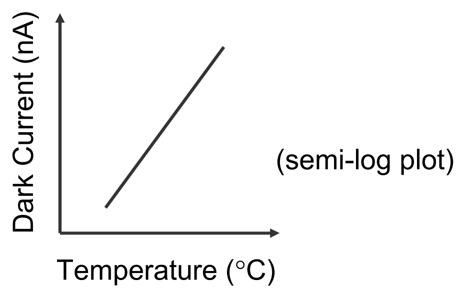
R_{SH} can have a significant impact on the results of a poorly designed circuit; especially at high temperatures (e.g., 125°C).

C_J limits the bandwidth of the circuit. As will be discussed later on, its interaction with the op amp (in the transimpedance amplifier) causes a greater reduction in bandwidth than may be apparent at first sight.

Current Sensors – Photodiode
Temperature Coefficient

- Photo Current (I_L)
  - Increases with increasing temperature
- Dark Current (I_D)
  - Current that exists without incident light
  - Increases with increasing temperature



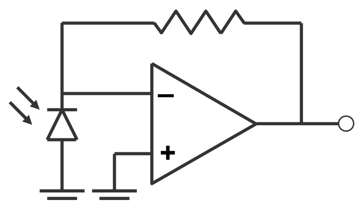


 306 ASC Slide 53

The dark current causes an error in the reading. As will be seen, it is mostly eliminated in precision applications by forcing a 0V bias across the photodiode, but is important in high speed applications because a significant reverse bias is imposed on the photodiode.

Current Sensors – Photodiode
Transimpedance Amplifier

- **Transimpedance Amplifier**
  - Converts current to voltage
  - Photodiode bias is maintained at a constant voltage by the op amp's "virtual short"
  - Short circuit current is converted to a voltage



 306 ASC Slide 54

The op amp's feedback forces the inverting input to be at 0V (because the non-inverting input is at 0V). This implies that all of the photodiode's current goes through the resistor; the op amp and resistor together convert the photocurrent to the output voltage.

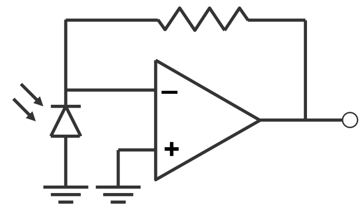
The most important points to be made are:

• There is no dark current because the photodiode is at 0V bias

• The photodiode's shunt resistance is bootstrapped by the op amp (forced to stay at 0V), so it has no effect on the output voltage

Current Sensors – Photodiode
Modes of Operation

- **Photovoltaic Mode**
  - Operated with zero bias
  - Pros:
    - Most precise
    - Linear operation
    - Virtually no influence by dark current
  - Cons:
    - Lower speeds
    - Not for switching (digital)

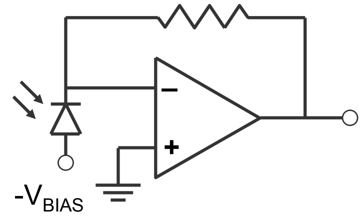


 306 ASC Slide 55

This configuration is slower because the photodiode's junction capacitance is large. Being large, it has a greater impact on the op amp; bandwidth has to be traded for stability.

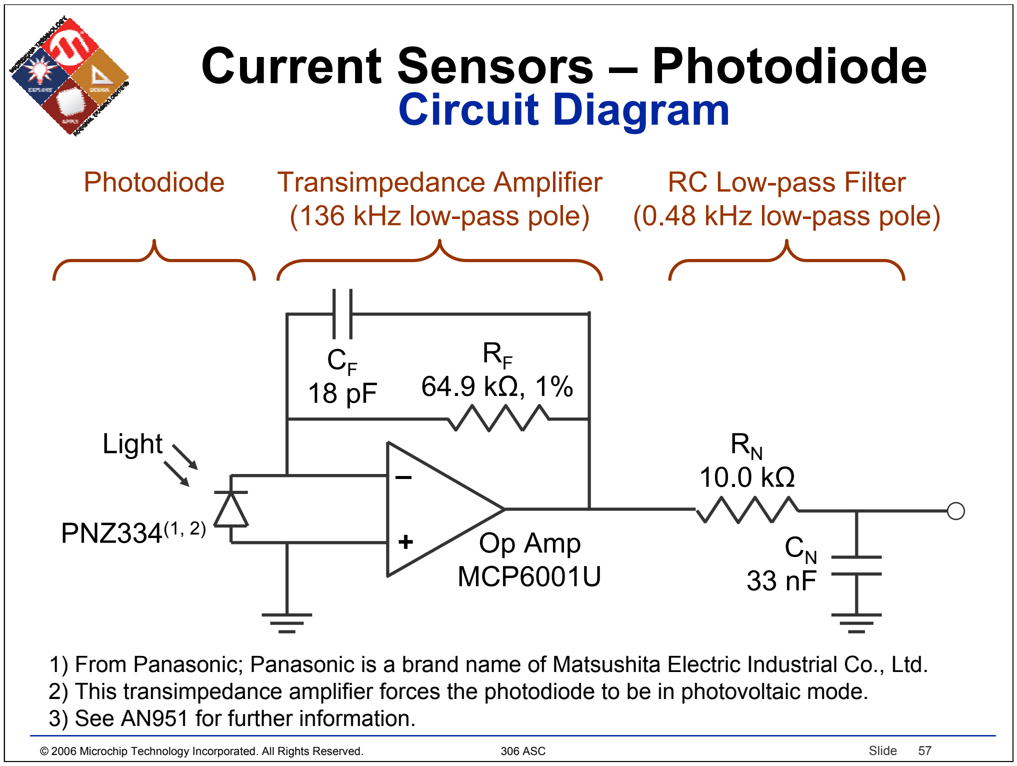
Current Sensors – Photodiode
Modes of Operation

- **Photoconductive Mode**
  - Operated in reverse bias
  - Pros:
    - Higher speeds
    - Good for switching (digital)
  - Cons:
    - Poor linearity
    - A small amount of dark current flows (i.e., current with no illumination)

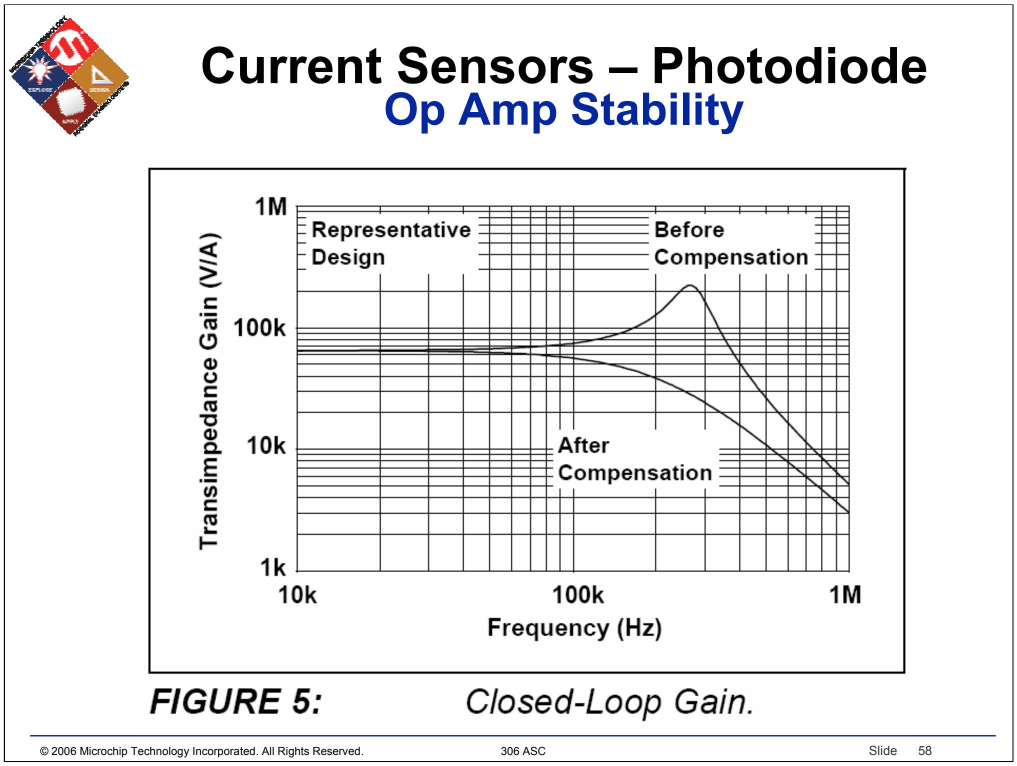


 306 ASC Slide 56

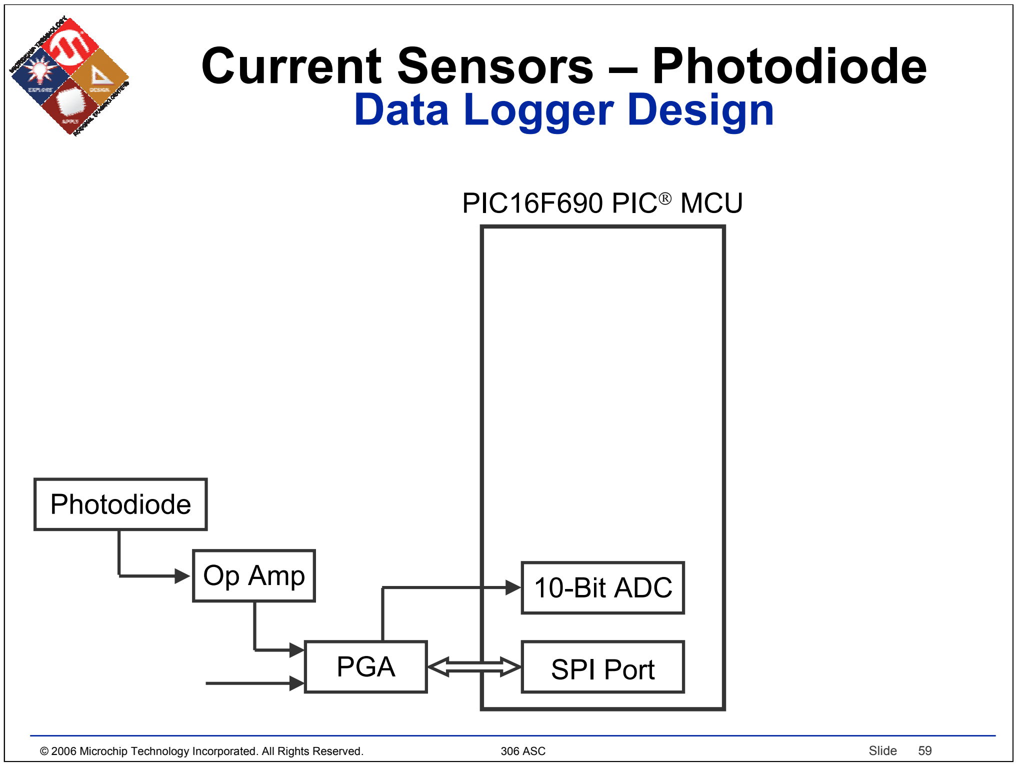
The photodiode’s junction capacitance is much smaller because it is reverse biased. The dark current (and current noise) are greater, so the error is greater.



This is a complete design, including a compensation capacitor (C_F) and noise filter.



This plot from AN951 (cited in Appendix A: References, Photodiode) shows the need for compensating the transimpedance amplifier. That application note gives a straightforward algorithms for designing the transimpedance amplifier, including the selection of C_F for stability and low peaking.



This is the portion of the "Data Logger Design" that pertains to the photodiode application.

Current Sensors – Photodiode
Data Logger Design

- **External Components:**
  - Photodiode
  - Op Amp
  - PGA (as MUX)
  - Filter
- **PIC® MCU Resources:**
  - ADC
  - 4 GPIO pins
  - Firmware routines

 306 ASC Slide 60

This is a high level resources analysis based on the block diagram.

Current Sensors – Photodiode
Exercise 2

- **Exercise 2**
  - MCP6SX2 PGA Photodiode PICtail™ Demo Board
  - See handout

 306 ASC Slide 61

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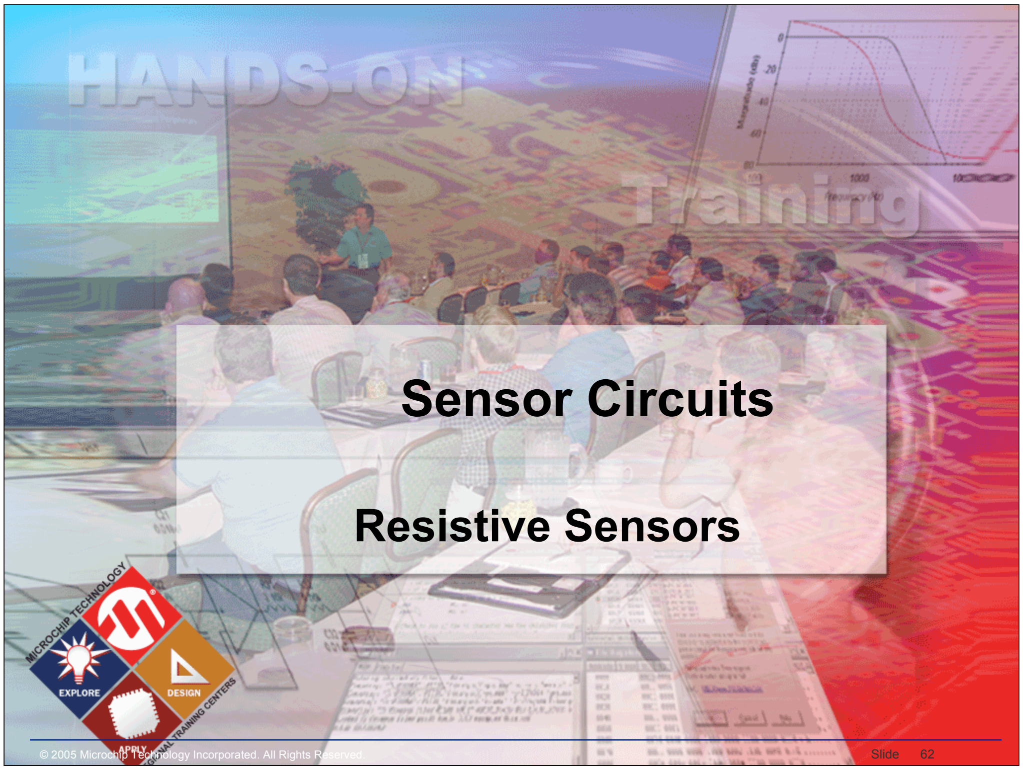
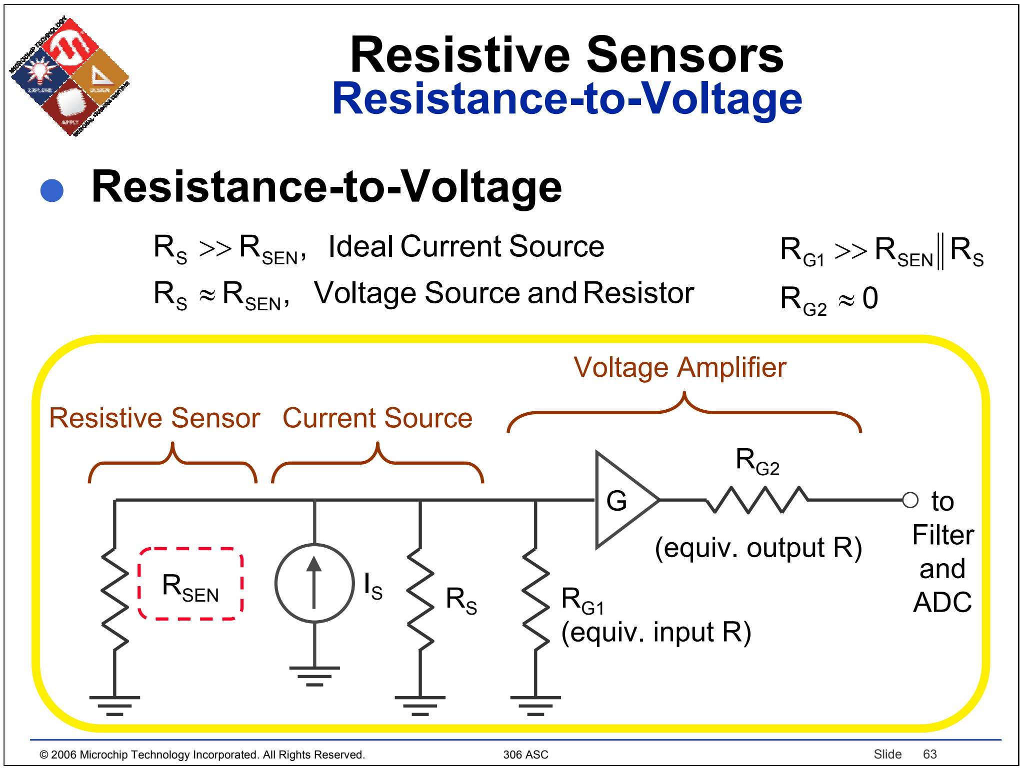


See application note AN990 (cited in Appendix A: References, General Topics); it is the source for most of the circuits shown here.



The Current Source can be

- Ideal (very large output impedance), so a voltage $R_{SEN}I_S$ is generated
- As simple as a resistor tied to a known voltage, creating a voltage divider; see the following slides

Resistive Sensors
Resistor Dividers

- **R-to-V Conversion**
  - Thermistor
  - RTD
  - Magneto-resistive compass
  - Wheatstone Bridge
    - Strain Gage
    - Pressure Gage

 306 ASC Slide 64

These are a few of many common applications based on this approach (using a voltage divider). As a matter of fact, the Wheatstone Bridge is nothing more than two voltage dividers; the important thing is that the difference between the two divider voltages can be a much more sensitive measurement than a single ended one.

Resistive Sensors
Resistor Dividers

$$V_{OUT} = V_{DD} \cdot \frac{R_1}{R_1 + R_{SEN}}$$

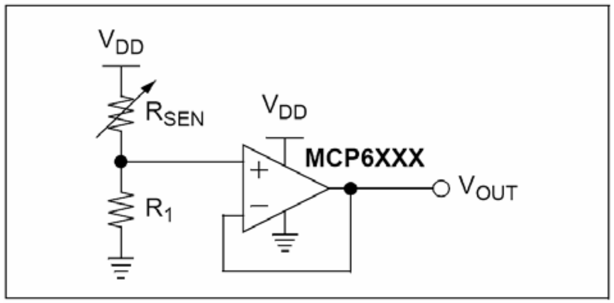


FIGURE 10: *Voltage Divider with Op Amp.*

 306 ASC Slide 65

The sensor R_{SEN} is placed in series with R_1 to convert resistance to voltage. It can be placed on top or bottom, depending on a particular design's needs. The op amp buffers the voltage divider.

Resistive Sensors
Resistor Dividers

- Pros:
  - Simple
  - Easy to filter
    - (just add a capacitor to ground in parallel with R_1)
  - Ratiometric
- Cons:
  - Non-linear transformation from resistance to voltage
  - Self-heating
  - DC drift vs. time due to constant DC bias
- PIC® MCU Resources Required:
  - 1 GPIO pin
  - ADC

 306 ASC Slide 66

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Resistive Sensors
Resistor Dividers

$$V_{OUT} = V_{DD} \cdot \left(\frac{R_1}{R_1 + R_{SEN}} + \frac{R_2}{R_{SEN}} - \frac{R_2}{R_1} \right)$$

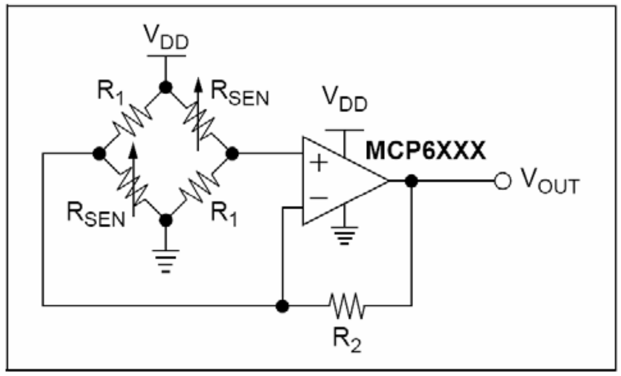

FIGURE 12: *Wheatstone Bridge – Single Op Amp Circuit.*

The sensors R_{SEN} are placed in series with the two resistors R_1 to convert resistance to a differential voltage (note that a Wheatstone bridge is nothing more than two resistor dividers in parallel). The op amp and resistor R_2 convert that differential voltage to V_{OUT}.

Usually, this circuit is designed so that $R_1 \approx R_{SEN}$ and $R_2 >> R_1$. In that case, V_{OUT} is a strong function of R_{SEN}.

Other amplifier arrangements can also be used: a difference amplifier or an instrumentation amplifier.

Resistive Sensors
Resistor Dividers

- Pros:
  - High CMRR
  - Ratiometric
- Cons:
  - Non-linear transformation from resistance to voltage
  - Self-heating
  - DC drift vs. time due to constant DC bias
- PIC® MCU Resources Required:
  - 1 GPIO pin
  - ADC

 306 ASC Slide 68

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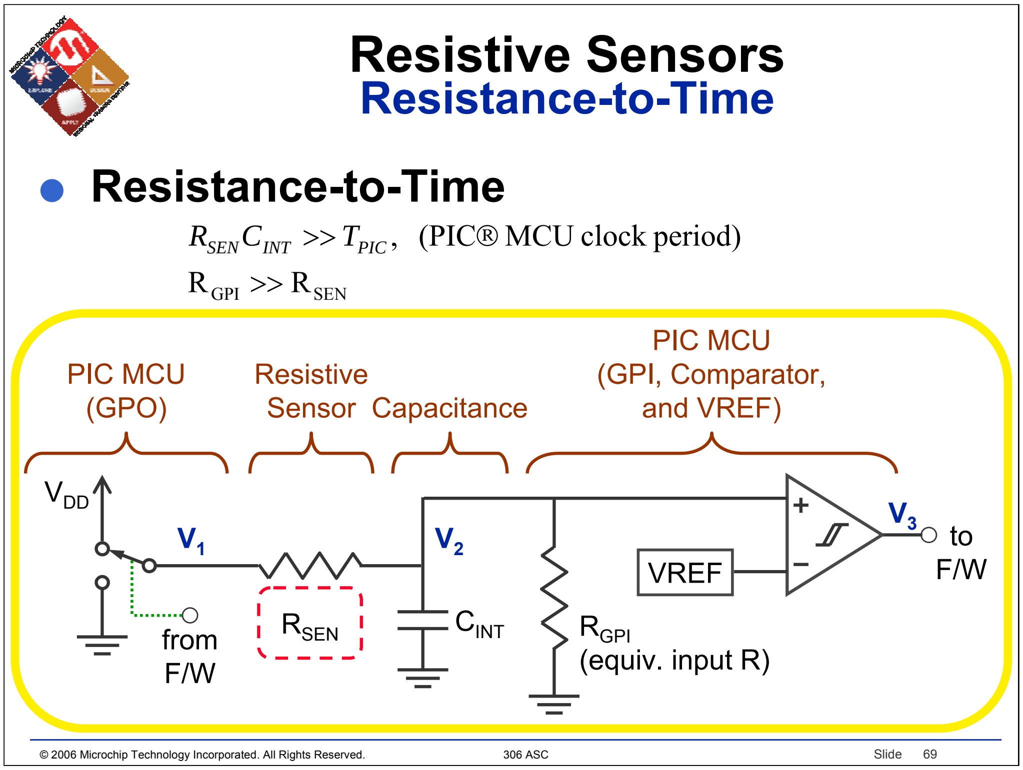


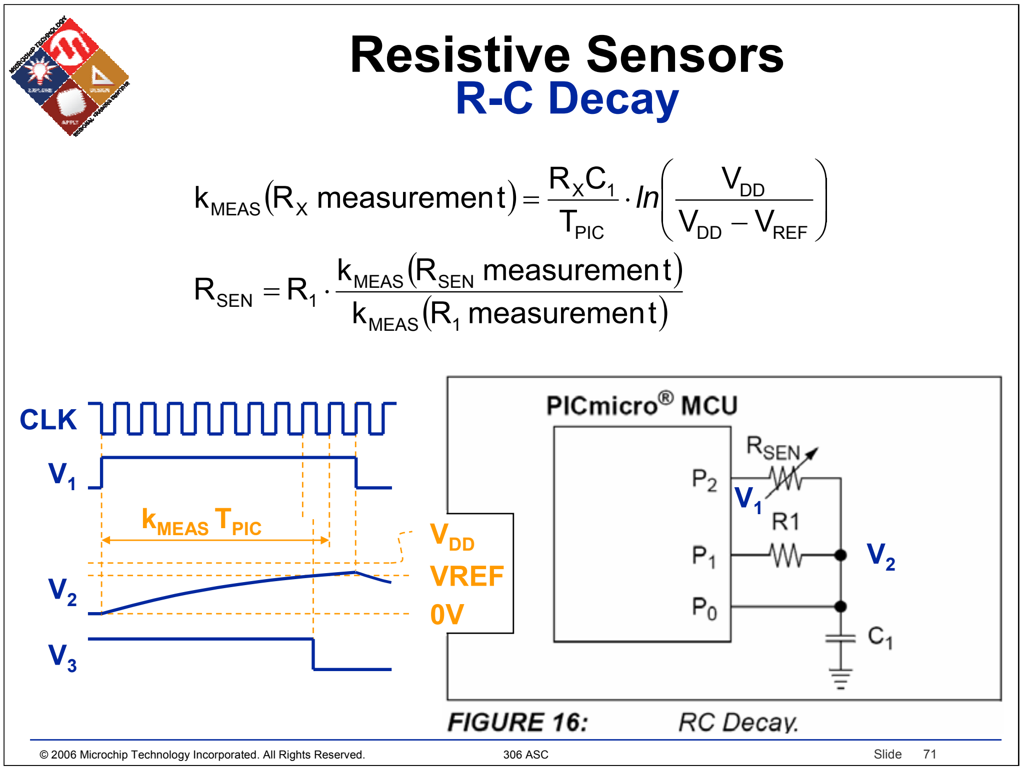
This circuit uses minimal circuitry (including a PIC microcontroller) to do the measurement. It is appropriate for larger R_{SEN} and C_{INT} values (e.g., higher than 10 kΩ and 1 nF). Obviously, at least one of R_{SEN} and C_{INT} would need to be significantly larger for the time constant to be much larger than T_{PIC}; this requirement ensures enough resolution on the elapsed time measurement.

Resistive Sensors
R-C Decay

- **R-C Decay**
  - Thermistor
    - Low temperatures when R_{SEN} is high

 306 ASC Slide 70

This circuit uses the PIC's ability to measure time (i.e., the CCP module).



Measuring the time for the $R_{SEN}C_1$ step response gives a measure of R_{SEN}. Since C_1 is apt to change significantly over temperature and process, the R_1C_1 step response is also measured; the ratio of the times is proportional to the ratio of resistances (the other terms, including C_1, cancel out).

Measuring a thermistor is one practical application of this circuit.

It is possible to eliminate R_1 in low precision applications; C_1's variation over temperature and process will tend to be the strongest limit on the accuracy.

Resistive Sensors
R-C Decay

- Pros:
  - Simple
- Cons:
  - Parasitic capacitance (on the order of 10 pF)
  - Clock period (needs to be << RC time constant)
  - Requires division (could use lookup table)
  - DC drift vs. time due to average DC bias
- PIC® MCU Resources Required:
  - 3 GPIO pins
  - CCP Module
  - Comparator

 306 ASC Slide 72

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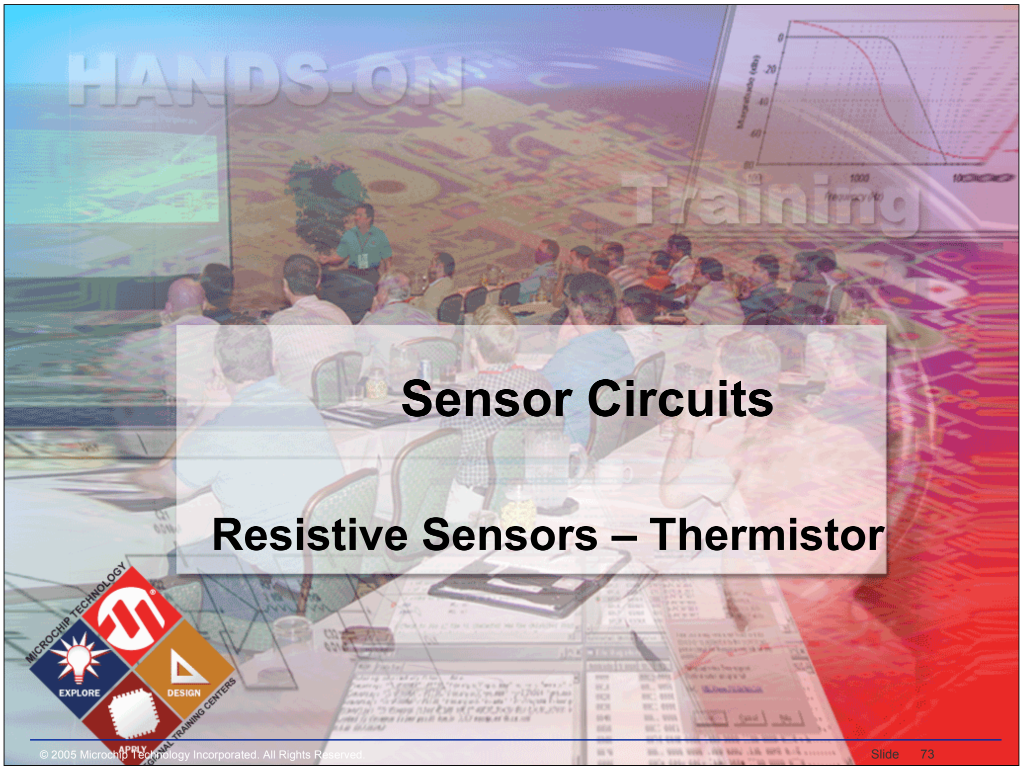


See application note AN897 (cited in Appendix A: References, Thermistor / Temperature). Additional references are included in Appendix A on thermistors and temperature sensing.

The companion demo board (MCP6SX2 PGA Thermistor PICtail™ Demo Board , cited in Appendix A: References, Demo Boards) is used to demonstrate this application in this class. Its User's Guide (DS51517, cited in Appendix A: References, Demo Boards) gives information on this board which was used for the measurements in AN897.

Resistive Sensors – Thermistor
Applications

- **THERMally sensitive resISTOR**
  - Negative Temperature Coefficient (NTC)
    - Precision temperature measurement
    - Exponential temperature to resistance relationship
  - Positive Temperature Coefficient (PTC)
    - For setting temperature trip points
    - Circuit protection (current limiter)

 306 ASC Slide 74

We will emphasize the NTC thermistors in this application.

Resistive Sensors – Thermistor
Description

- **Vishay BCcomponents[1] 2381 640 55103**
  - NTC
  - Resistance at 25°C = R_{25} = 10 kΩ
  - Tolerance on R_{25} = ±1%
  - Response time ≈ 1.7 s, in oil
  - Operating Temperature ranges
    - Zero dissipation, short periods: -40 to +150 °C
    - Zero dissipation, continuous: -40 to +125 °C
    - Maximum dissipation (100 mW): 0 to +55 °C

1) Vishay BCcomponents is a brand name of Vishay Intertechnology, Inc.

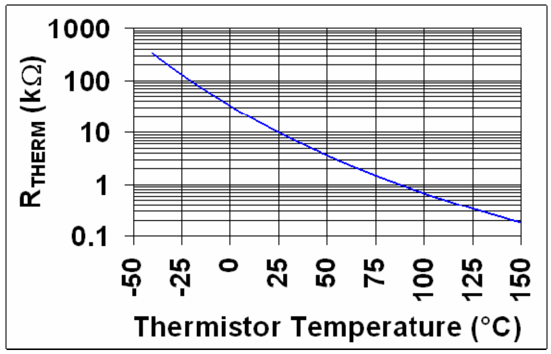
This was the thermistor chosen for AN897, and its accompanying demo board (MCP6SX2 PGA Photodiode PICtail™ Demo Board; see Appendix A: References, Demo Boards).

This thermistor is relatively accurate for its cost, plus it has a large temperature range and quick response time.

Resistive Sensors – Thermistor
Resistance vs. Temperature

T_{THERM} (°C)	T_{THERM} (°F)	R_{THERM} (kΩ)
-40	-40	332.1
0	32	32.56
25	77	10.00
70	158	1.753
125	257	0.3387
150	302	0.1826

BCcomponents P/N: 2381 640 55103

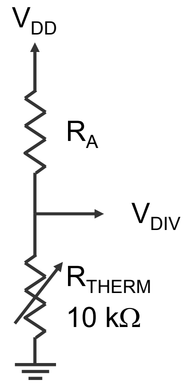


 306 ASC Slide 76

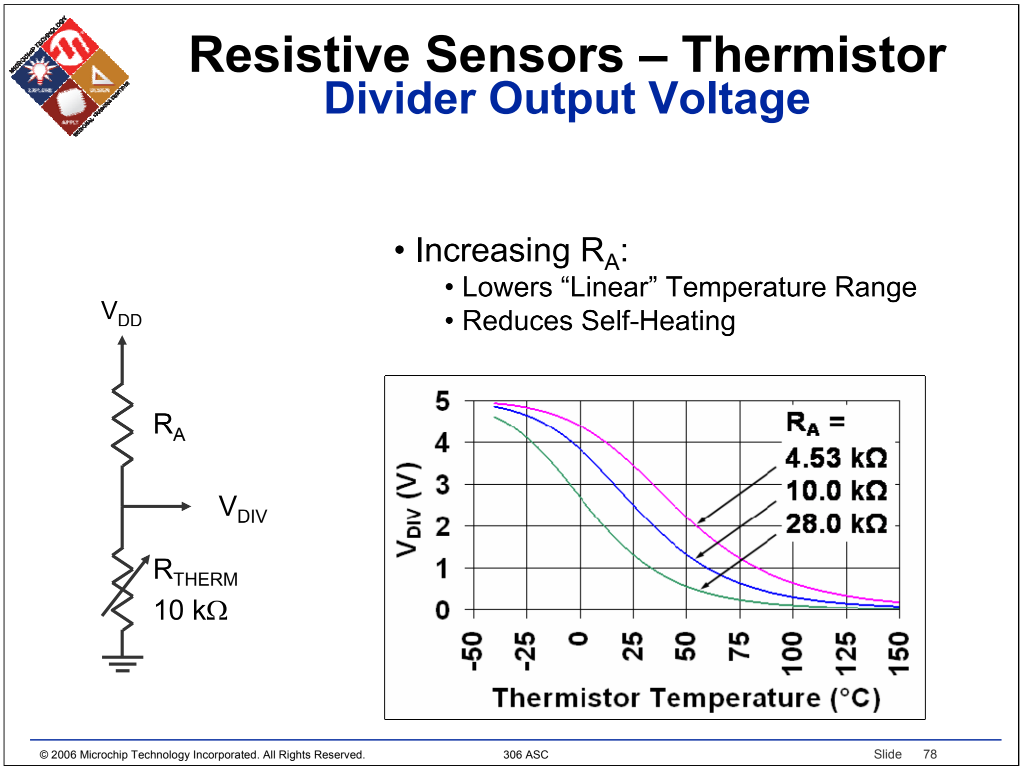
This thermistor changes by more than 3 orders of magnitude across its temperature range.

Resistive Sensors – Thermistor
Description

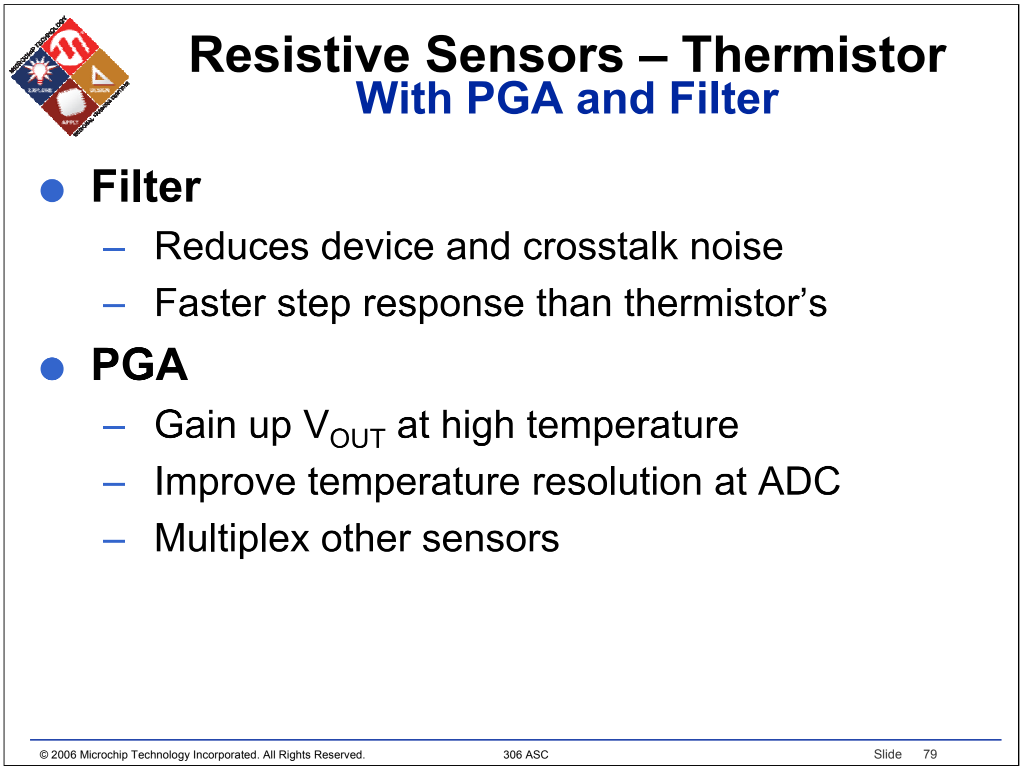
- **Resistance to Voltage Conversion**
  - Requires current excitation
    - Voltage Reference and Resistor
    - Current Reference
  - R_{THERM} vs. temperature
    - Is very non-linear
    - Is approximately exponential
  - Voltage divider
    - Ratiometric (V_{DD} = ADC's VREF)



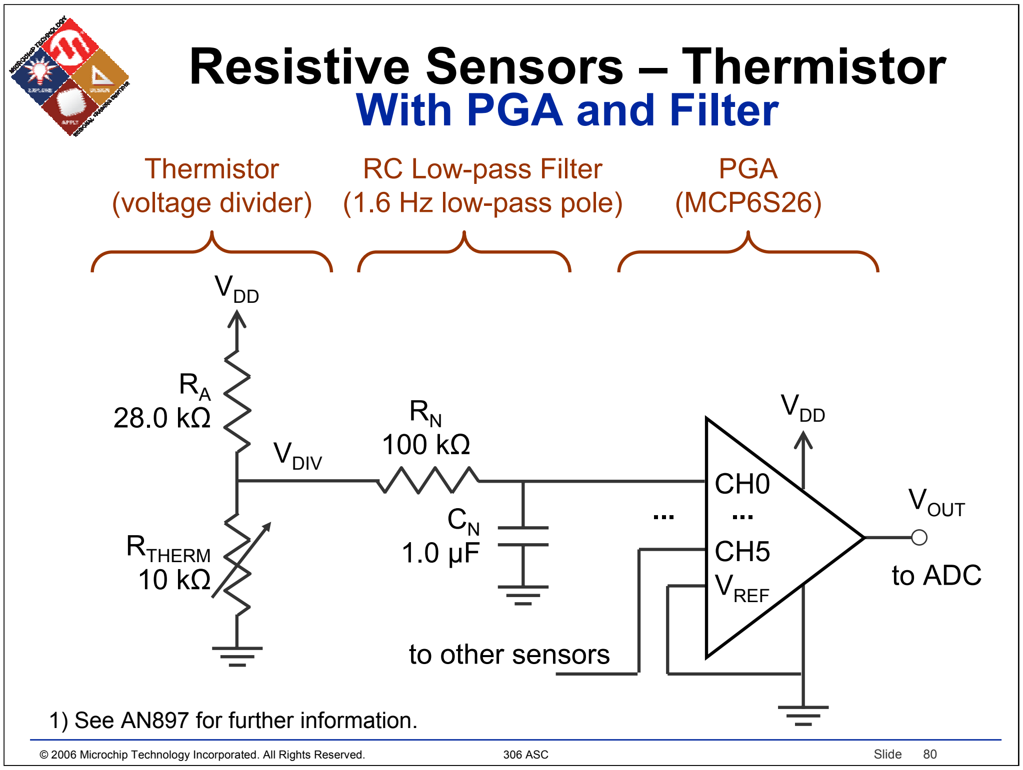
The previously discussed circuits for converting resistance to voltage could be applied here. This design uses the simplest circuit. As will be discussed later, the F/W includes a piecewise linear interpolation table and algorithm to overcome the non-linearity.



This slide shows the fundamental tradeoffs in the chosen circuit. The temperature to voltage conversions is reasonably linear over a 40°C to 80°C wide temperature range. All three of these curves show a greatly reduced sensitivity to temperature (i.e., $\Delta V_{DIV}/\Delta T_A$) at high temperatures. This is acceptable for applications needing a small temperature range, but is unacceptable for applications over a larger temperature range.



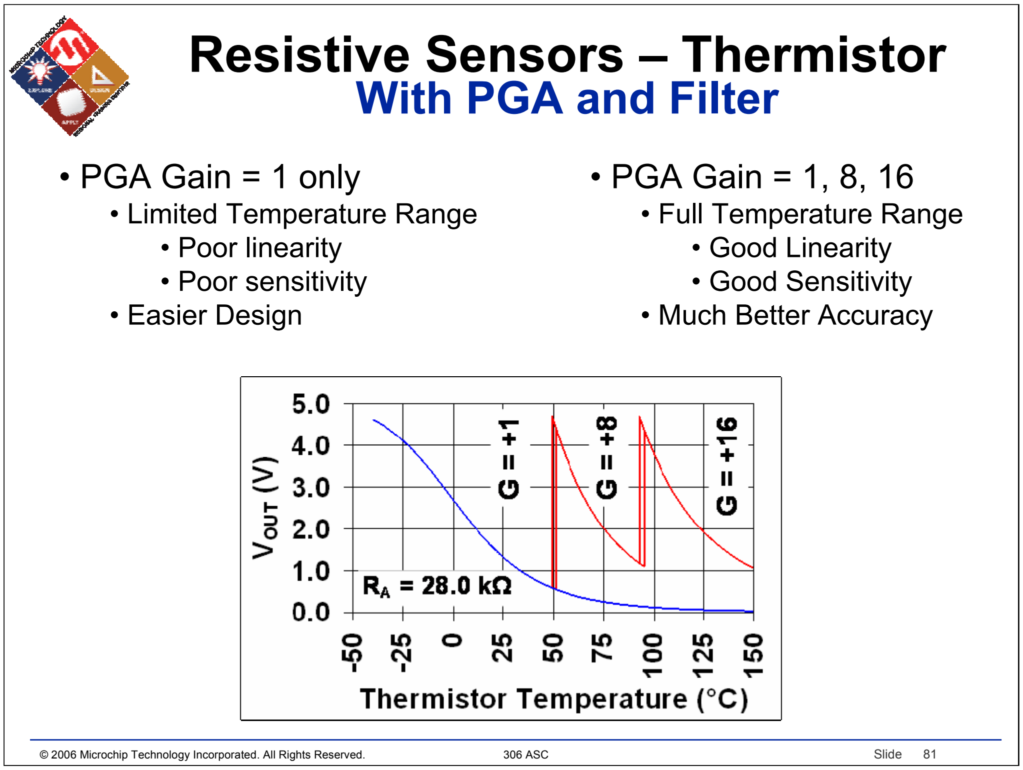
If the filter is twice as fast as the thermistor, the combined filter and thermistor rise time is only 12% slower. Choosing an R-C filter ensures no overshoot and a relatively quick settling time. A more complex filter would be needed, however, if the design proves to have significant mains (line power) noise.



The PGA is set at higher gains when the thermistor temperature is high, which improves the temperature resolution at the ADC. The PGA also multiplexes other sensors.

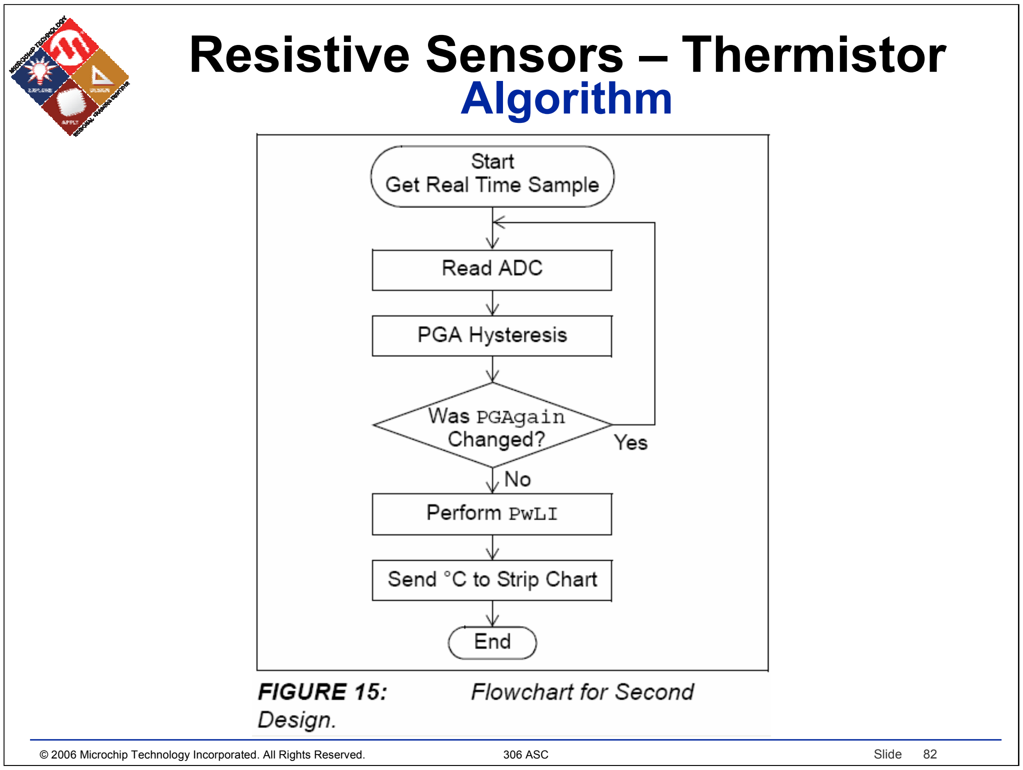
The filter reduces device and crosstalk noise. It needs to have a faster step response than thermistor so that the overall time constant is not significantly slower.

This circuit diagram is deceptively simple. The next slide will show the importance of setting the PGA's digitally selectable gain according to the thermistor temperature.



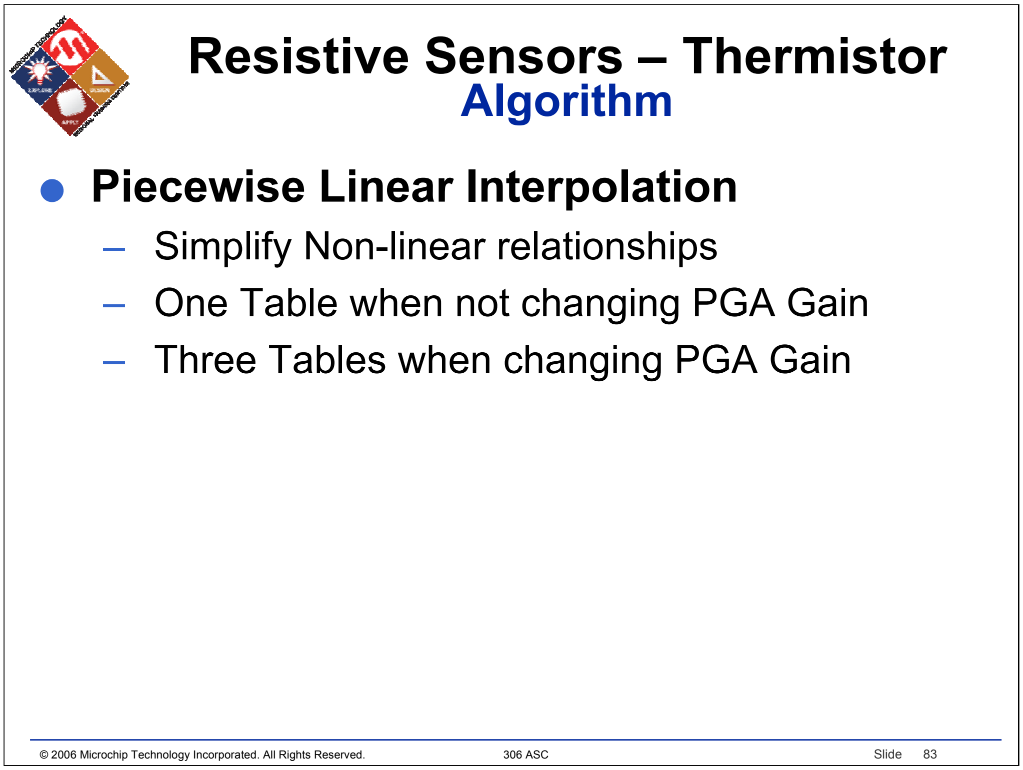
The PGA Gain = 1 curve includes both the blue curve on the bottom. At temperatures above 50°C, the non-linear behavior causes V_{OUT}'s sensitivity ($\Delta V_{OUT}/\Delta T_A$) to T_A to be greatly reduced; this application is good for the limited range of –40°C to +50°C.

The PGA Gain = 1, 8, 16 curves are in red to the right of +50°C, and blue to the left of +50°C. These curves show three different gains, which were chosen to optimize the overall error performance (see AN897). Also shown is a significant amount of hysteresis at each gain selection point (1.7°C and 2.0°C), which will greatly reduce the number of times the gain will need to be changed. V_{OUT}'s sensitivity ($\Delta V_{OUT}/\Delta T_A$) is greatly improved (11×); it only varies between 0.03°C/LSb and 0.27°C/LSb over the full temperature range (instead of between 0.1°C/LSb and 2.4°C/LSb).



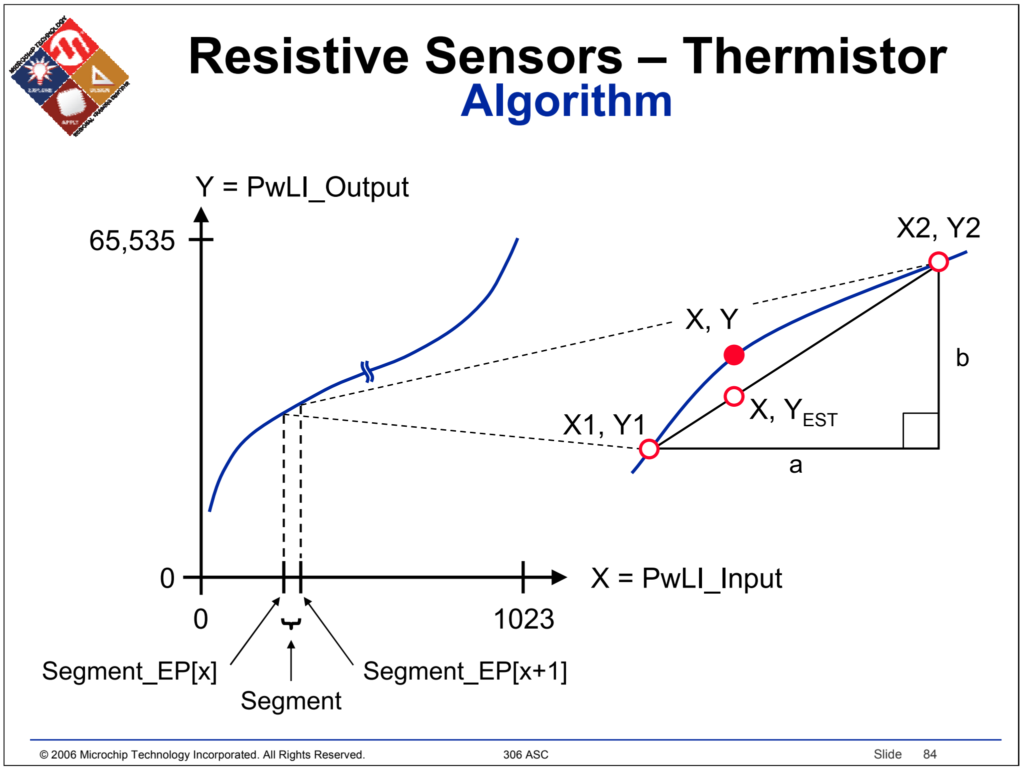
This algorithm was kept simple by avoiding complex decision structures. It operates as follows:

- Read/Set Gain
  - “Read ADC” obtains the ADC output value
  - “PGA Hysteresis” selects the appropriate gain for the PGA, while including the appropriate hysteresis behavior
  - If the gain changes, this process is restarted (extra decisions could have been made to speed up the execution in this case, but the structure would have become too complex)
  - If the gain doesn’t change, then the current ADC output is a valid one
- “Perform PwLI” executes the piecewise linearization subroutine, which will be described soon, to obtain the correct T_A for a given V_{OUT}
- “Send °C to Strip Chart” sends T_A to a PC GUI interface



This approach is simple enough to be easy to implement in a 8-bit MCU (see AN942 cited in Appendix A: References, General topics).

One table is provided for each gain; only 1 table is needed when G = 1 only, and three tables are needed when G = 1, 8, and 16. Each table is filled with valid data across the entire temperature range (–40°C to +150°C). ADC code values that outside that range are set to the nearest valid value. This increases the likelihood that a processing error would not produce an erroneous output code.



The top 10 bits of X are used to point to the correct Y1 value and slope (b/a) in the lookup table. The bottom 6 bits (X – X1) are multiplied by the slope, and then added to Y1:

$$Y \approx Y1 + (b/a)(X - X1)$$

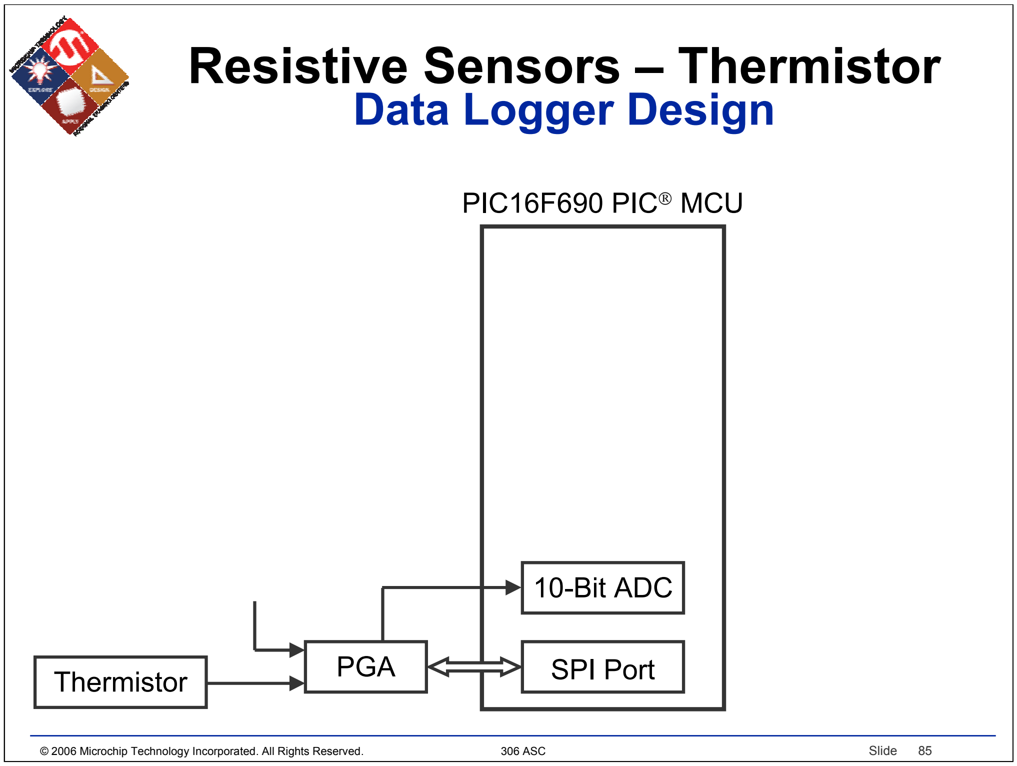
where:

$$a = X2 - X1$$

$$b = Y2 - Y1$$

This greatly reduces the time spent in interpolating, and makes the multiplications much easier to perform (shift and add).

See application note AN942 (cited in Appendix A: References, General Topics) for a more complete description.



This is the block diagram level design for the thermistor.

Resistive Sensors – Thermistor
Data Logger Design

- **External Components:**
  - Thermistor
  - PGA
    - (as MUX)
    - Provides Gain
  - Filter
- **PIC® MCU Resources:**
  - ADC
  - 4 GPIO pins
  - Firmware routines

 306 ASC Slide 86

This is a simplified high level resources analysis.

Resistive Sensors – Thermistor
Exercise 3

- **Exercise 3**
  - MCP6SX2 PGA Thermistor PICtail™ Demo Board
  - See handout

 306 ASC Slide 87

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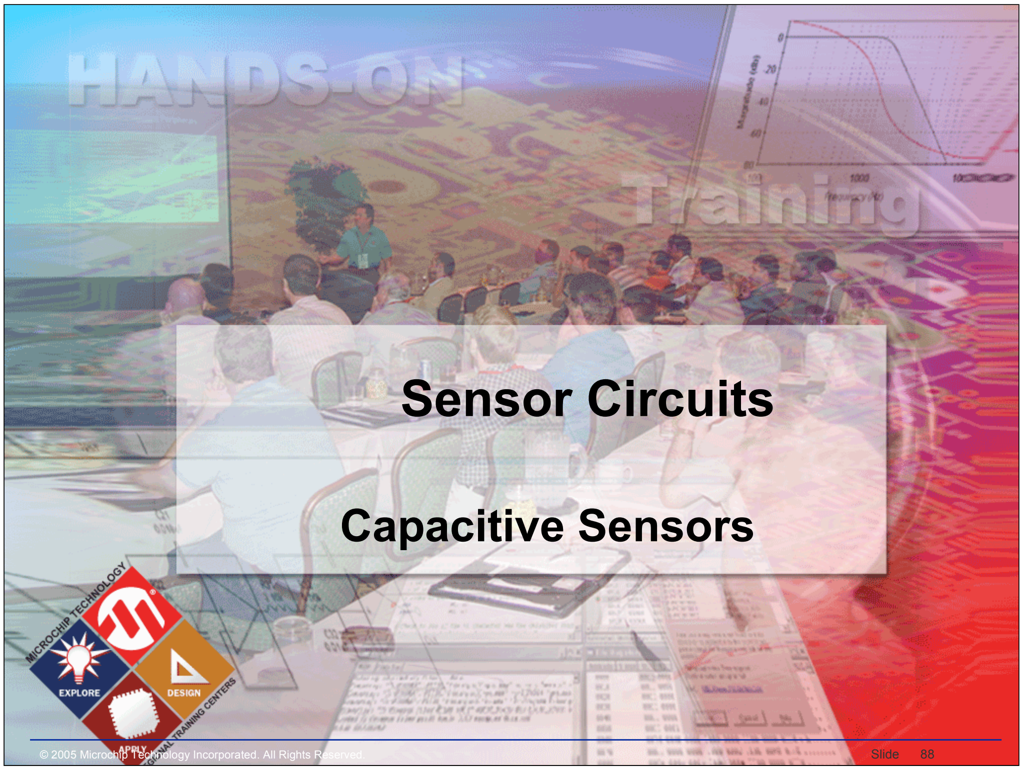
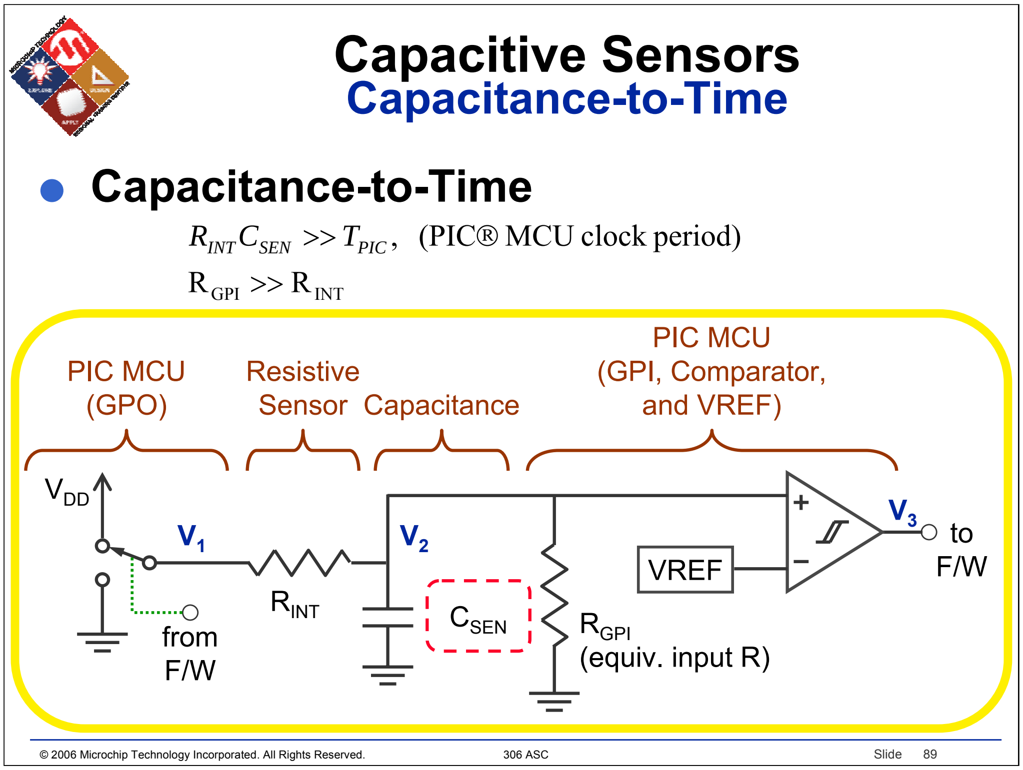


See application note AN990 (cited in Appendix A: References, General Topics); it is the source for most of the circuits shown here.



This circuit uses minimal circuitry (including a PIC microcontroller) to do the measurement. It is appropriate for larger R_{INT} and C_{SEN} values (e.g., higher than 10 kΩ and 1 nF). Obviously, at least one of R_{INT} and C_{SEN} would need to be significantly larger for the time constant to be much larger than T_{PIC}; this requirement ensures enough resolution on the elapsed time measurement.

The application circuit that follows modifies this basic design by adding an op amp. With the op amp, the circuit achieves a much higher level of accuracy, and the ability to measure much smaller capacitances.

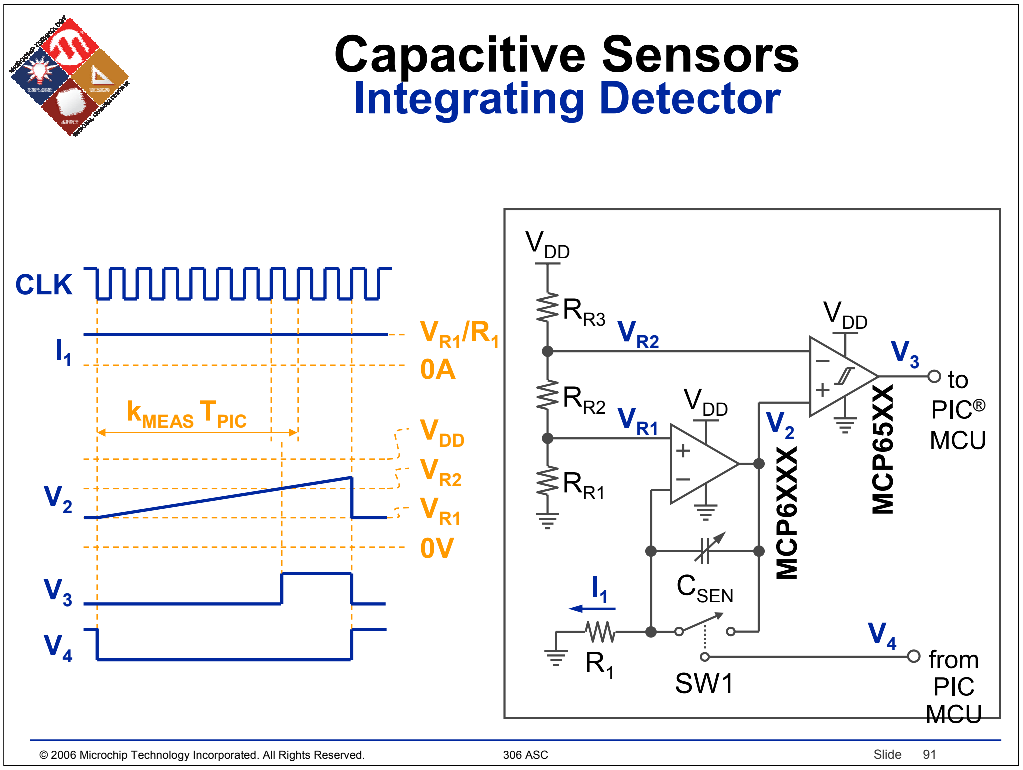
Capacitive Sensors
Integrating Detector

- **Single Slope Integrating Detector**
  - Tank Level Sensor
  - Humidity Sensor
  - Touch Sensor

 306 ASC Slide 90

This application bootstraps the sensor's parasitic impedances (e.g., case-to-ground capacitance). The conversion from capacitance to time is a linear one.

Greater accuracy is possible by using a dual slope approach; see the write-up in AN1016 (cited in Appendix A: References, Capacitive / Humidity Sensor).



The op amp forces the voltage V_{REF} to be across R_1, which causes a constant current to be dumped on C_{SEN} and SW1. When SW1 (usually a FET whose gate is controlled by the PIC) is closed, the voltage across C_{SEN} is forced to zero. Once it releases, C_{SEN} starts to integrate the current V_{REF}/R_1 with constant slope. Once the op amp output reaches V_{REF}, the output comparator signals the MCU. The MCU calculates the elapsed time, which is converted to the capacitance C_{SEN}, then starts the cycle over again by closing SW1.

Capacitive Sensors
Integrating Detector

$$I_1 = \frac{V_{R1}}{R_1}$$

$$\frac{\Delta V_2}{\Delta t} = \frac{I_1}{C_{SEN}}, \quad \text{SW1 is open}$$

$$k_{MEAS} = \frac{R_1 C_{SEN}}{T_{PIC}} \cdot \left(\frac{V_{R2} - V_{R1}}{V_{R1}} \right)$$

$$C_{SEN} = k_{MEAS} \left(\frac{T_{PIC}}{R_1} \cdot \frac{V_{R1}}{V_{R2} - V_{R1}} \right)$$

 306 ASC Slide 92

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Capacitive Sensors
Integrating Detector

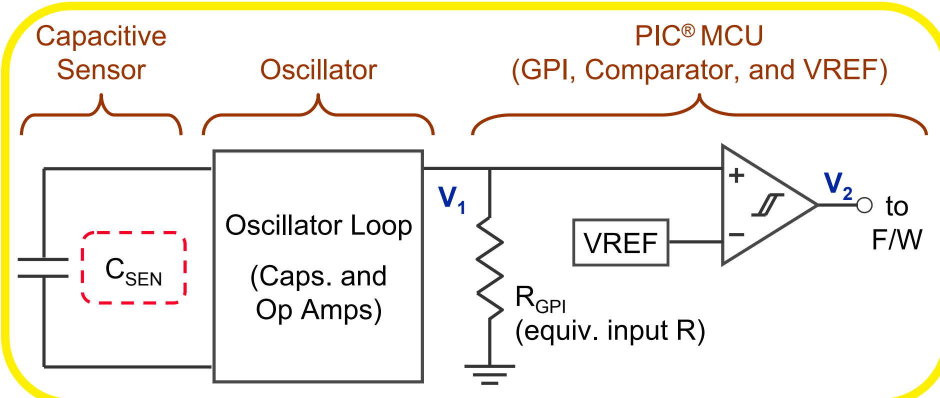
- Pros:
  - Linear conversion from capacitance to time
  - No DC drift vs. time due to zero average DC bias
- Cons:
  - Moderately complex circuitry
  - Calibration may be required
- PIC® MCU Resources Required:
  - Number of GPIO pins
    - 4 for external V_{REF} and Comparator
    - 3 for internal V_{REF} and Comparator
  - CCP Module

 306 ASC Slide 93

Dielectric absorption is much slower, and is averaged out. Offset voltages are not corrected as in the dual slope approach previously mentioned.

Capacitive Sensors
Capacitance-to-Frequency

- **Capacitance-to-Frequency**



 306 ASC Slide 94

This approach works well with the PIC’s CCP module.

Capacitive Sensors
Oscillator Frequency

- **Oscillator Frequency**
  - Tank Level Sensor
  - Humidity Sensor
  - Touch Sensor
  - Proximity Sensor

 306 ASC Slide 95

Many more applications are possible; those listed are very common ones.

See application notes AN866 and AN895 (cited in Appendix A: References, General Topics).

Capacitive Sensors
Oscillator Frequency

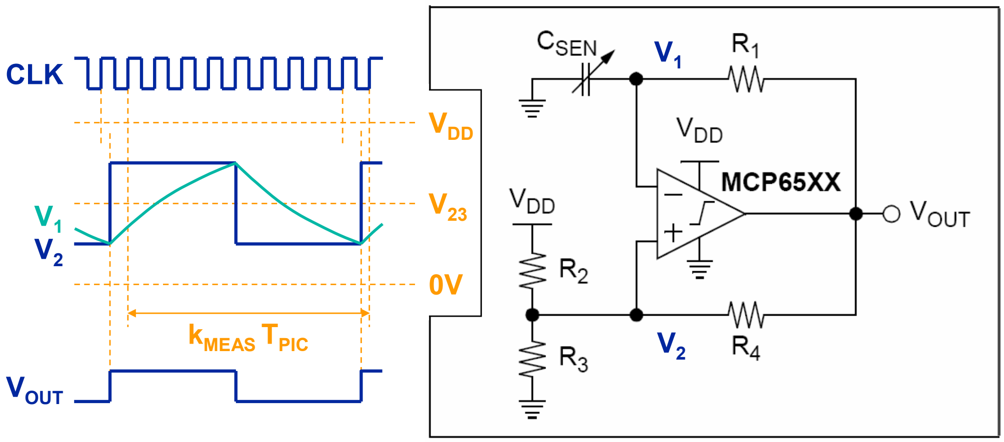


FIGURE 19: *Multi-vibrator (oscillator).*

 306 ASC Slide 96

This is a simple multi-vibrator, or relaxation oscillator. V_{OUT} toggles from rail to rail as V_1 moves past V_2. V_2 moves up and down with V_{OUT}, and sets the points where V_{OUT} toggles. R_1 and C_{SEN} produce a single time constant step response whose length is proportional to C_{SEN}.

See application notes AN866 and AN895 (cited in Appendix A: References, General Topics) for a simple design solution for this circuit. Basically, the design is set so that V_1 is centered on mid-supply ($V_{DD}/2$).

Capacitive Sensors
Oscillator Frequency

$$R_{23} = R_2 \| R_3$$

$$V_{23} = V_{DD} \cdot \frac{R_3}{R_2 + R_3}$$

$$\Delta V_2 = V_{DD} \cdot \frac{R_{23}}{R_{23} + R_4}$$

$$V_{2\_AVE} = \frac{V_{23} R_4 + (V_{DD}/2) R_{23}}{R_{23} + R_4}$$

$$V_2 = V_{2\_AVE} - \frac{\Delta V_2}{2}, \quad V_{OUT} = 0V$$

$$V_2 = V_{2\_AVE} + \frac{\Delta V_2}{2}, \quad V_{OUT} = V_{DD}$$

 306 ASC Slide 97

This sequence of equations has been included for completeness. They will need to be double checked; the author has not had time to verify them on the bench.

Capacitive Sensors
Oscillator Frequency

$$k_{MEAS}(\text{High to Low}) = \frac{R_1 C_{SEN}}{T_{PIC}} \cdot \ln\left(\frac{V_{2\_AVE} + \Delta V_2/2}{V_{2\_AVE} - \Delta V_2/2}\right)$$

$$k_{MEAS}(\text{Low to High}) = \frac{R_1 C_{SEN}}{T_{PIC}} \cdot \ln\left(\frac{V_{DD} - (V_{2\_AVE} - \Delta V_2/2)}{V_{DD} - (V_{2\_AVE} + \Delta V_2/2)}\right)$$

$$C_{SEN} = k_{MEAS}(\text{High to Low}) \cdot T_{PIC} \Big/ R_1 \ln\left(\frac{V_{2\_AVE} + \Delta V_2/2}{V_{2\_AVE} - \Delta V_2/2}\right)$$

$$C_{SEN} = k_{MEAS}(\text{High to Low}) \cdot T_{PIC} \Big/ R_1 \ln\left(\frac{V_{DD} - (V_{2\_AVE} - \Delta V_2/2)}{V_{DD} - (V_{2\_AVE} + \Delta V_2/2)}\right)$$

 306 ASC Slide 98

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Capacitive Sensors
Oscillator Frequency

- Pros:
  - Simple
  - Start up is robust
- Cons:
  - Comparator inside the loop adds timing errors
  - No DC drift vs. time due to zero average DC bias
- PIC® MCU Resources Required:
  - 1 GPIO pin
  - CCP Module

 306 ASC Slide 99

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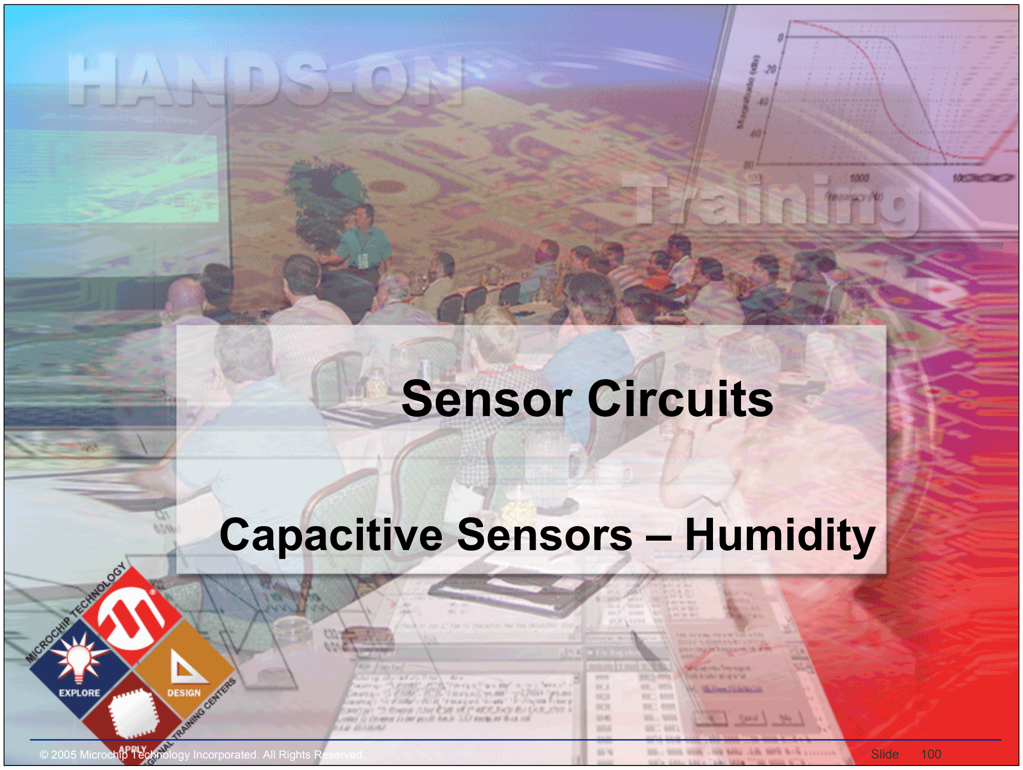


See application note AN1016 (cited in Appendix A: References, Capacitive / Humidity Sensor). Appendix A includes a reference to AN1014 that discusses an alternative method to measuring a small capacitive sensor; it is not as accurate, but is simpler to build.

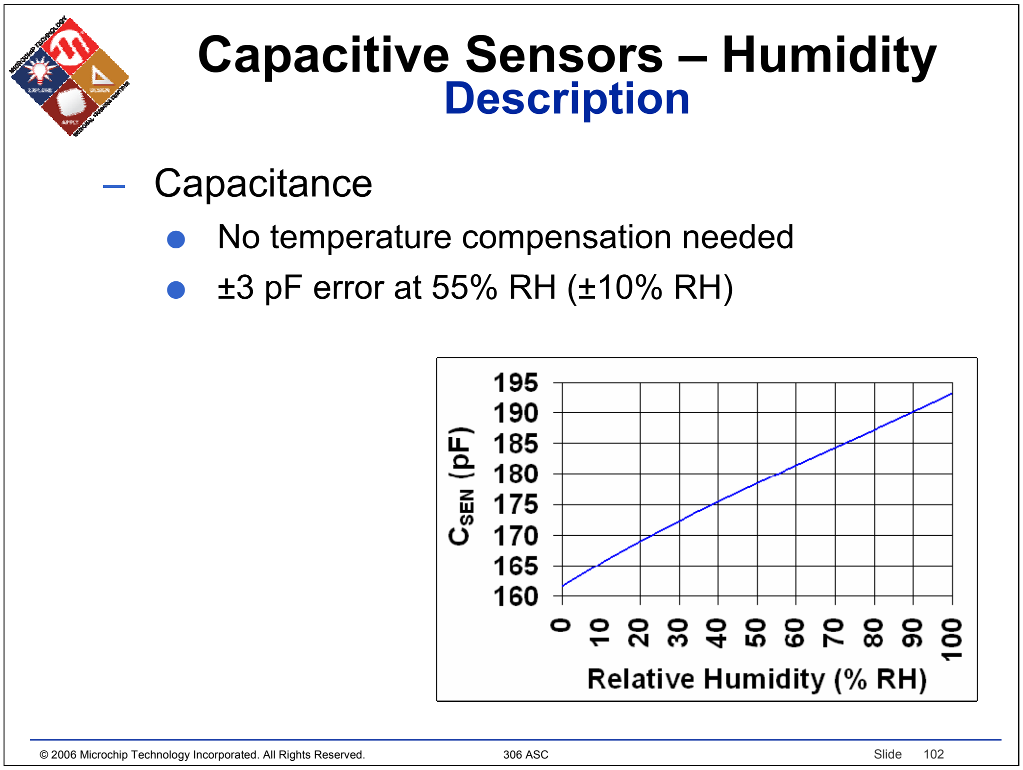
The companion demo board (Humidity Sensor PICtail™ Demo Board, cited in Appendix A: References, Demo Boards) is used to demonstrate this application in this class. Its User's Guide (DS51594, cited in Appendix A: References, Demo Boards) gives useful information on the board used for the measurements in AN1016.

Capacitive Sensors – Humidity
Description

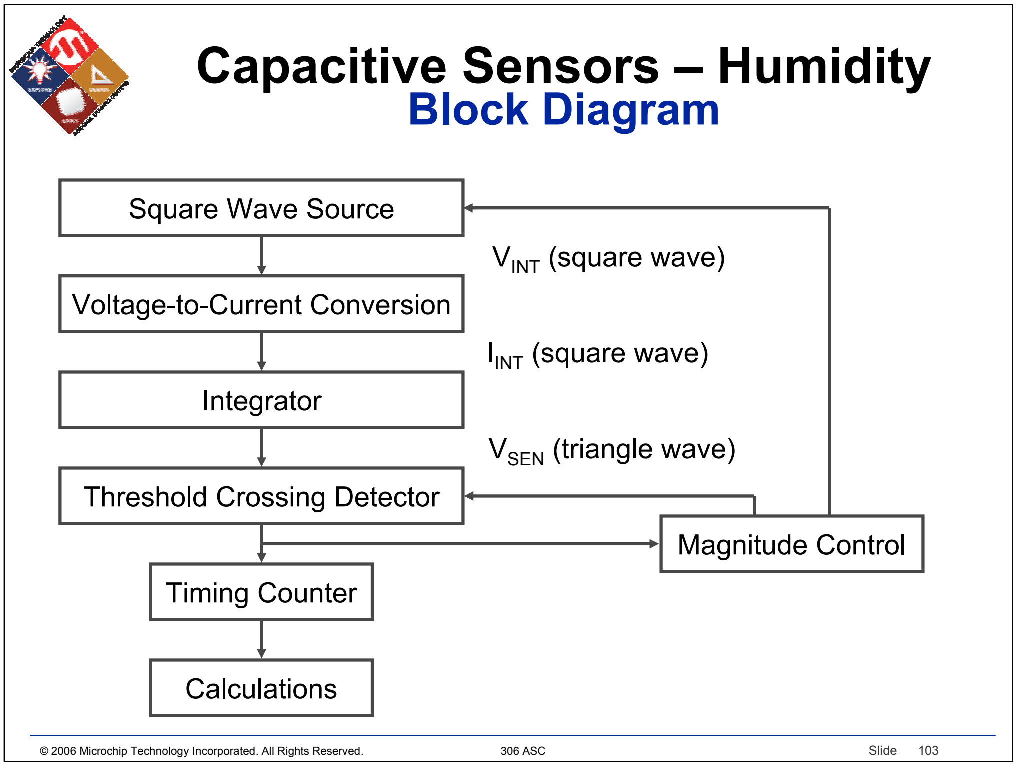
- **HS1101LF humidity sensor from Humirel**
  - Two terminal package (case tied to one pin)
  - Temperature Range
    - Operating = -60 to 140°C
    - Storage = -60 to 140°C
  - Humidity Range
    - 1 to 99 % RH, measurement range
  - Time Response
    - Time Constant ≤ 5 s
    - Recovery Time = 10 s, after 150 hrs of condensation

 306 ASC Slide 101

This is a relatively low cost sensor with good performance.

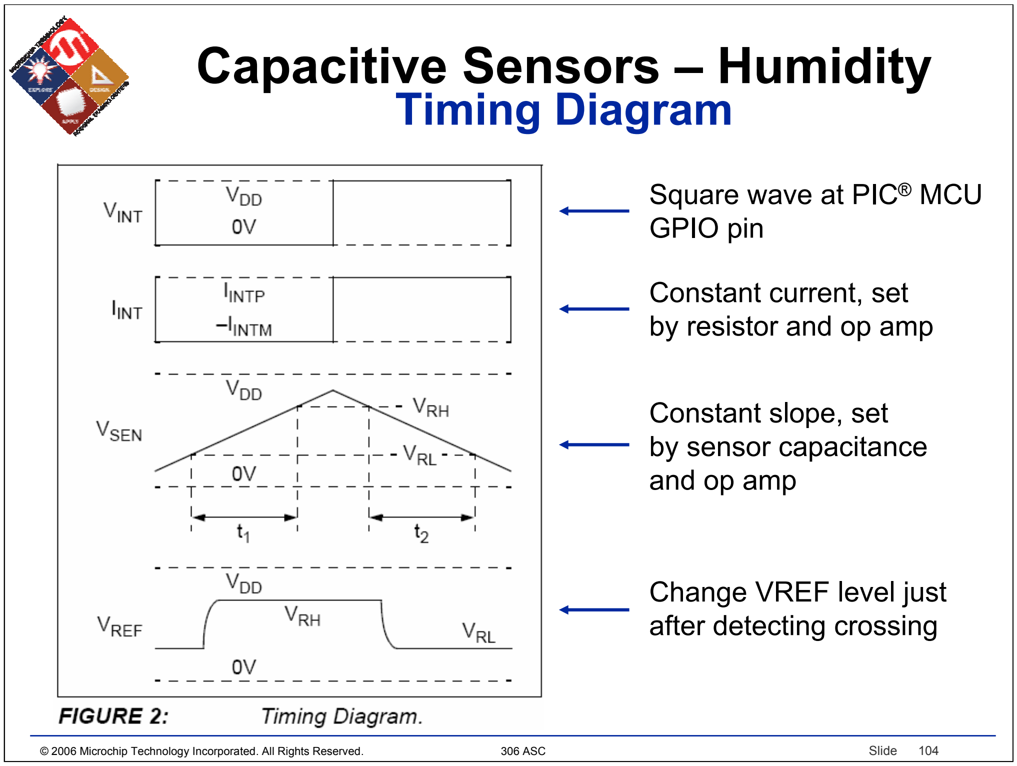


The response curve is not quite linear; it will be corrected later on with a piecewise linear interpolation subroutine. The sensor's error can be high, but it is possible to correct it in production with a calibration step.

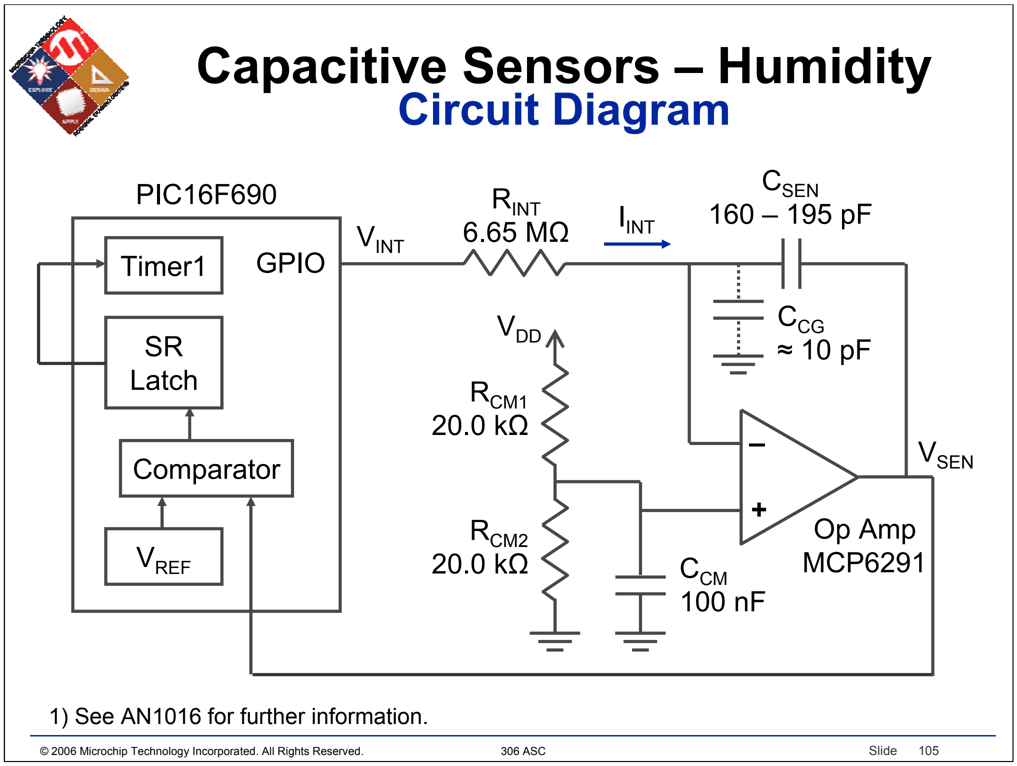


This humidity sensor cannot be measured using the R-C decay circuit because the capacitance is too low; it is too close in value to the parasitic capacitances, and the time constant is too low.

The block diagram shows a mixed signal control loop. The analog portion (top left) generates a triangle wave whose slope is a function of the sensor capacitance (C_{SEN}). A comparator detects when the triangle wave crosses either the top or bottom threshold value. The magnitude control (F/W) changes the polarity of the square wave each time this happens. The timing counter outputs the accumulated count each time a threshold is crossed, and resets the CCP Module. The elapsed times are used in a F/W routine to calculate capacitance, and then Relative Humidity.



This is a timing diagram that goes with the block diagram on the previous slide.



This is the analog circuit diagram; most of the analog details are outside the PIC. Note that this implementation is simpler and cleaner than the single slope integrating circuit covered in the "Sensor Overview: Sensor Conditioning Circuits" section of this presentation.

The op amp's input is biased at mid-supply ($V_{DD}/2$) so that the triangle wave has equal slope magnitudes going up and down. The sensor's case to ground capacitance (C_{CG}) and the op amp's common mode capacitance are in parallel from the inverting input to ground. The op amp's feedback action bootstraps this capacitance so that it does not create any dynamic (AC) currents; most other circuits do not have this feature. The op amp's output voltage swings between the limits set by V_{REF}, the comparator, and the control loop's F/W.

The Humidity Sensor PICtail™ Demo Board that implements this circuit uses the PIC16F690's internal oscillator as its clock source, which is set to run at 8 MHz. With the divide by four, the PIC's instruction cycle is then $T_{PIC} = 0.5\ \mu s$.

Capacitive Sensors – Humidity
Comments

- **Like Dual Slope Integrating ADC**
  - Similarities:
    - Measures triangle wave's rate of change
    - Resistance is constant
  - Dissimilarities
    - Input Voltage is constant
    - Capacitance changes

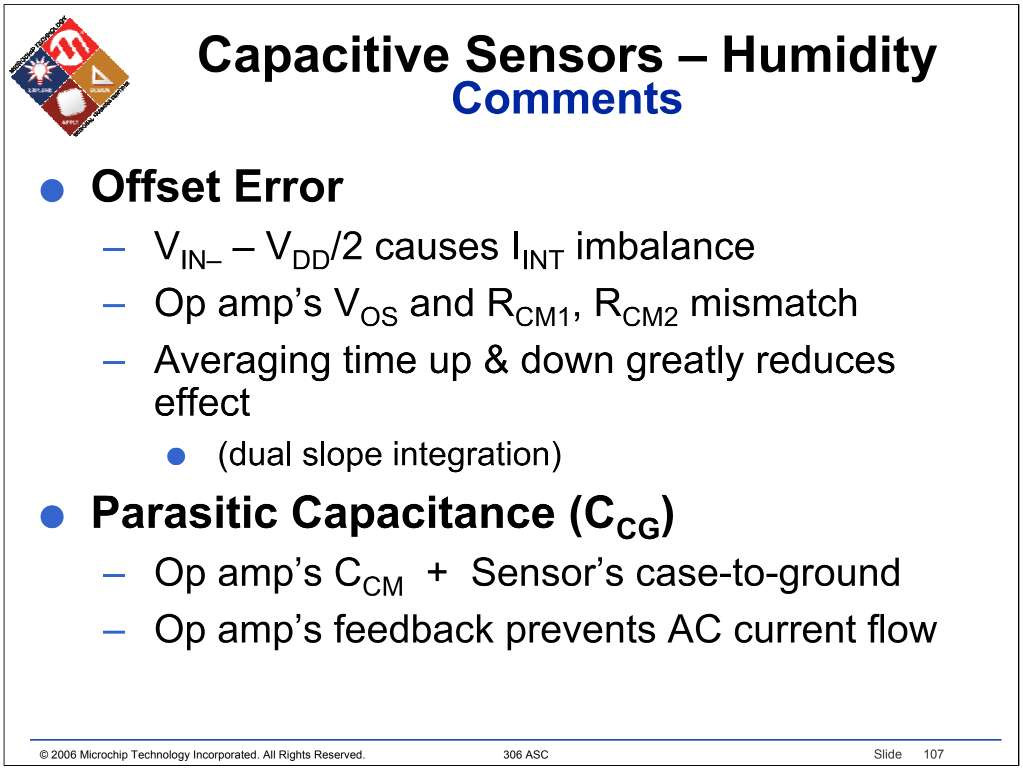
 306 ASC Slide 106

Many low frequency, high resolution ADC's have been of the "integrating ADC's" type. This application uses the same basic architecture, but modifies it in these ways:

• Uses two constant (ratiometric) voltages, with equal magnitude and opposite polarity, as the references

• The capacitance changes

• The triangle wave's slopes are approximately equal in magnitude; they are both used to measure the same quantity (C_{SEN})



AN1016 goes into significant detail about how using the dual slope approach reduces the impact of offset and mismatch errors. Suffice it to say that averaging the two elapsed times (up and down) makes the relative error go as the square of the offset and mismatch errors (I_{INT} imbalance):

$$V_{DD}/2 = 5.0V / 2 = 2.5V$$

$$V_{OS} = 5\ mV = 0.2\%\ of\ 2.5V$$

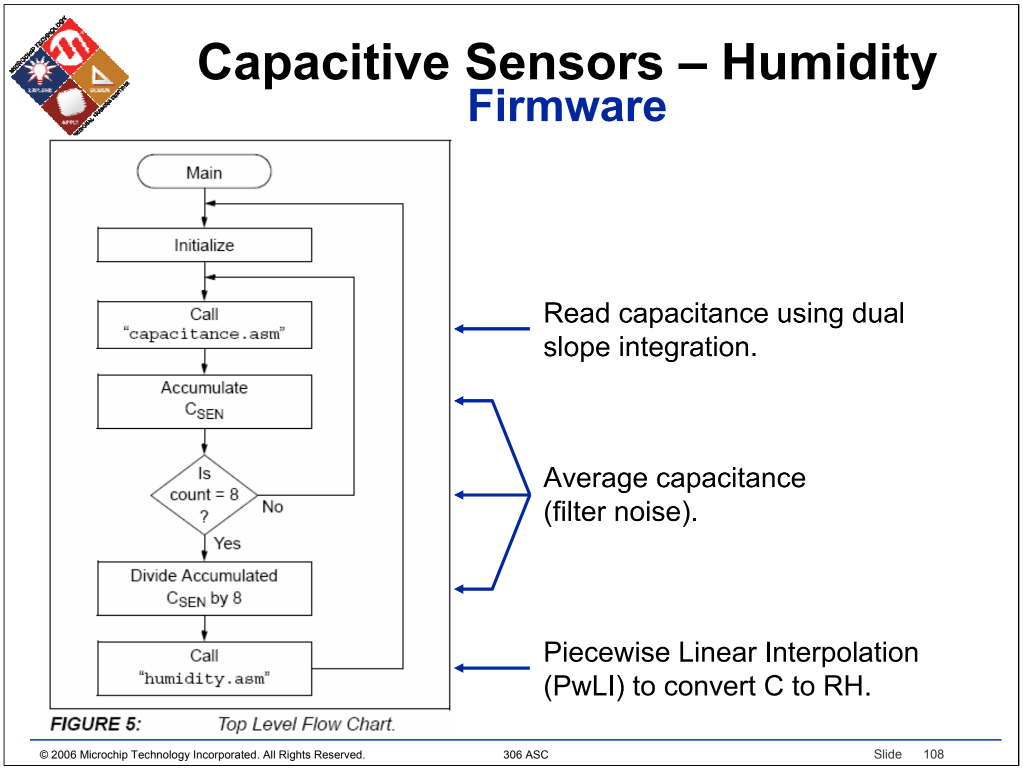
$$TOL_R = 1\%$$

gives:

$$\text{Offset and Mismatch Relative Errors} = 0.2\% + 1\% = 1.2\%$$

$$C_{SEN}\ \text{Relative Error} = (0.012)^2 = 0.0144\% = 144\ ppm$$

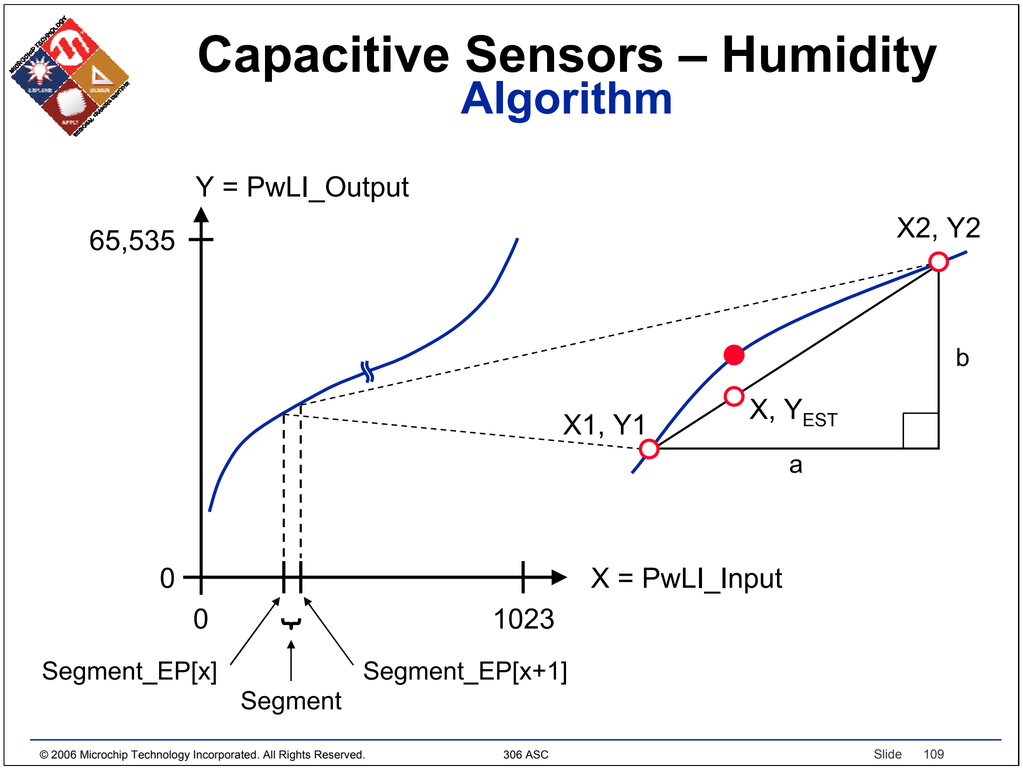
As was discussed before, the sensor's parasitic case to ground capacitance (C_{CG}) is bootstrapped by the op amp. This prevents any AC current to flow in C_{CG}.



The "Initialize" subroutine sets several important variables, including the oscillator. The selections it makes are the internal, calibrated oscillator, and a clock frequency of 8 MHz (giving T_{PIC} = 0.5 µs, as explained before).

The "capacitance.asm" subroutine also converts the average elapsed time into capacitance. R_{INT} (converts voltage square wave to a current square wave) was chosen (assuming T_{PIC} = 0.5 µs) so that the output increments at 0.1 pF / count; no actual calculations are needed.

The PwLI subroutine takes care of the sensor's offset and non-linearity.



The top 10 bits of X are used to point to the correct Y1 value and slope (b/a) in the lookup table. The bottom 6 bits (X – X1) are multiplied by the slope, and then added to Y1:

$$Y \approx Y1 + (b/a)\,(X - X1)$$

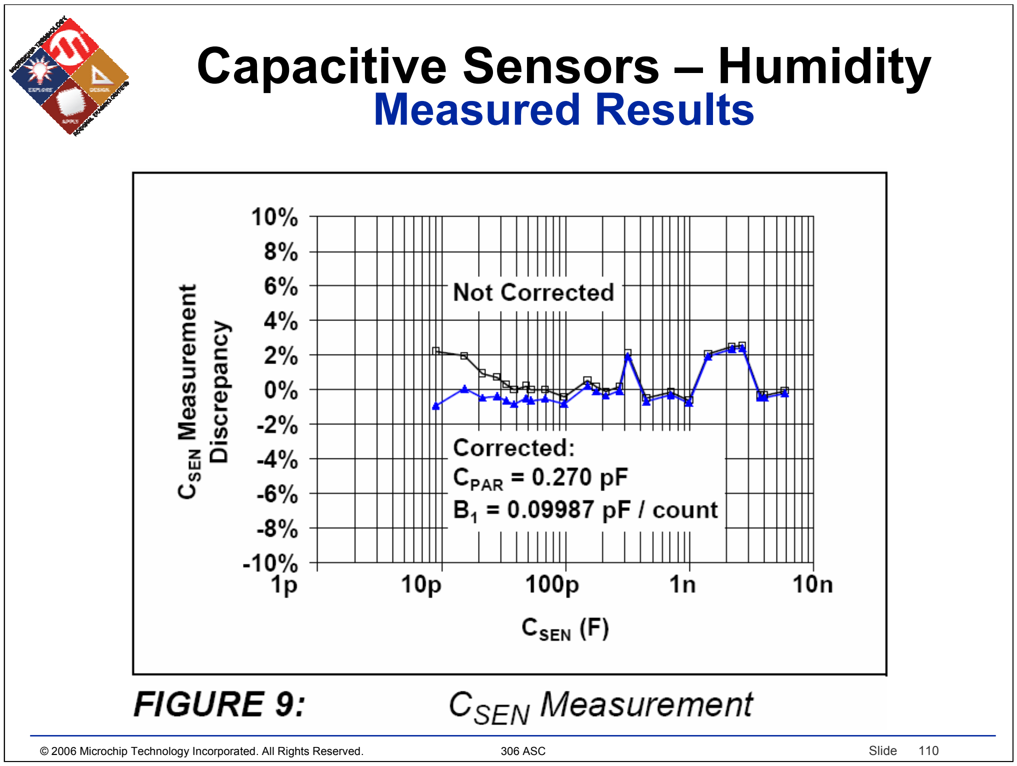
where:

$$a = X2 - X1$$

$$b = Y2 - Y1$$

This greatly reduces the time spent in interpolating, and makes the multiplications much easier to perform (shift and add).

See application note AN942 (cited in Appendix A: References, General Topics) for a more complete description.



The top curve shows the measurement results when using the circuit and F/W as designed. It is possible to do further correction to improve the accuracy:

• Subtract the parasitic capacitance in parallel with C_{SEN} (mainly the op amp's pin-to-pin package capacitance)

• Correct for the circuit's gain error (e.g., R_{INT} is too high, or the PIC's clock is too slow)

As can be seen in the plot, these corrections may, or may not, be of use in a particular application.

Capacitive Sensors – Humidity
Measured Results

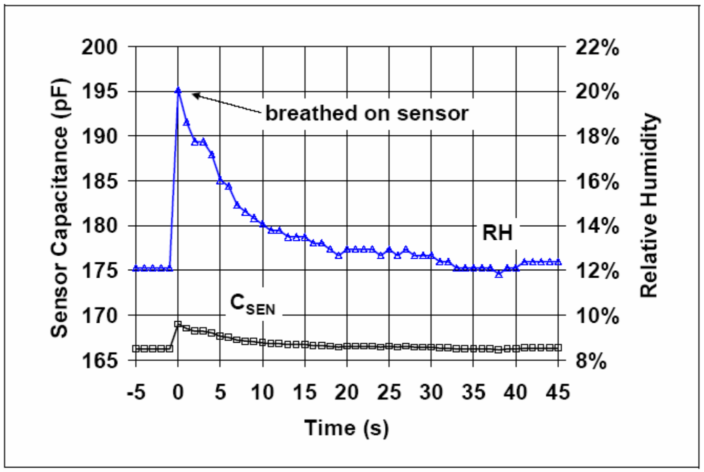
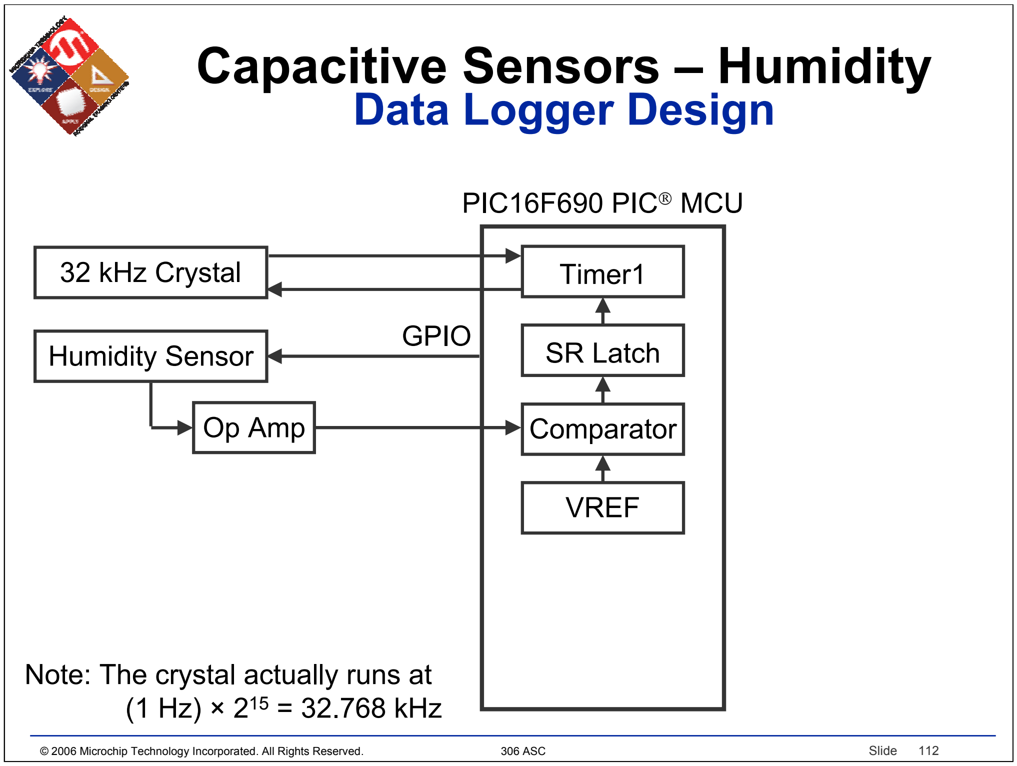


FIGURE 10: *HS1101LF Impulse Response.*

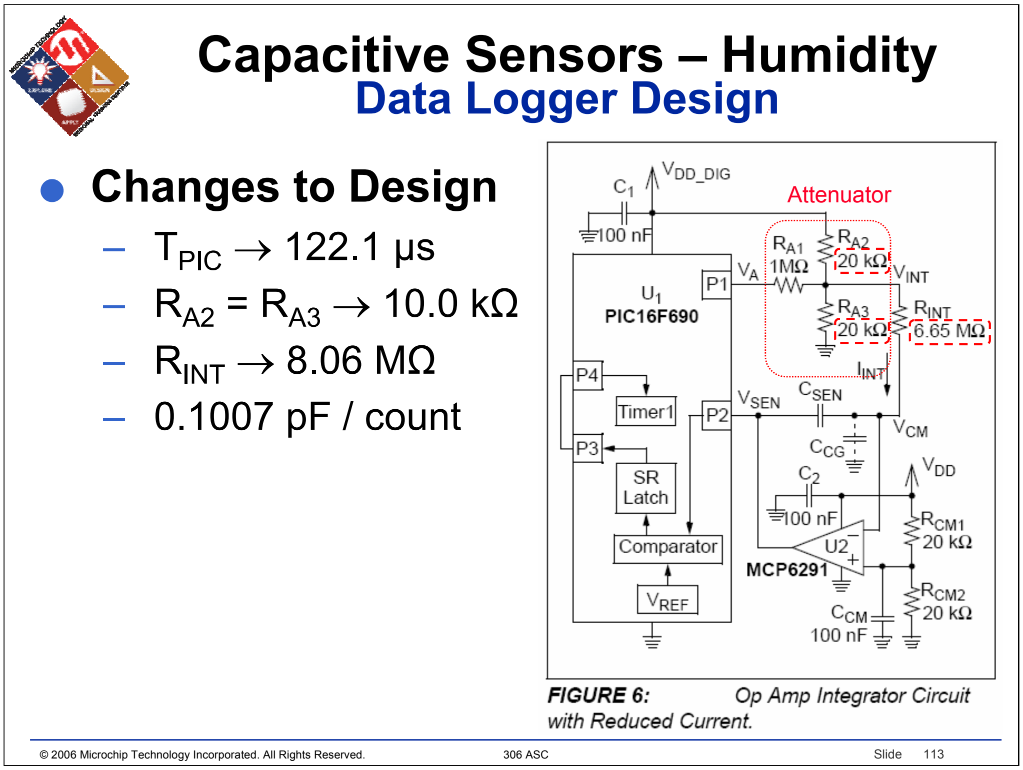
Since the measurement resolution is 0.1 pF / count, the resolution in relative humidity (RH) is 0.6% RH. The application was set to take one reading every second.

Breathing on the humidity sensor is a quick way to verify its time constant, and that the circuit is operating properly.

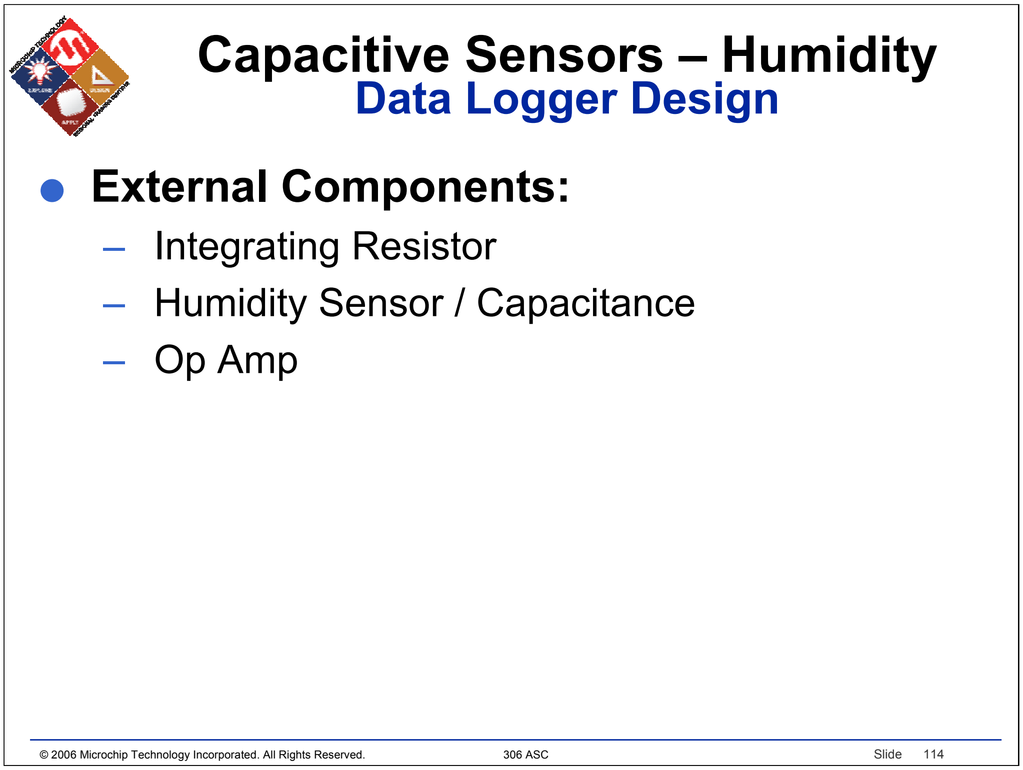


This is the block diagram of the "Data Logger Design" implementation of this circuit. The "32 kHz Crystal" is a watch crystal (it actually runs at at 1 Hz times 2^{15} = 32.768 kHz); it is used to make the PIC operate as slow as possible, conserving energy. The "SR Latch" is a relatively new feature which made it possible to do a hardware test instead of a F/W test; this cut the F/W test loop overhead very significantly, which increased the available resolution.

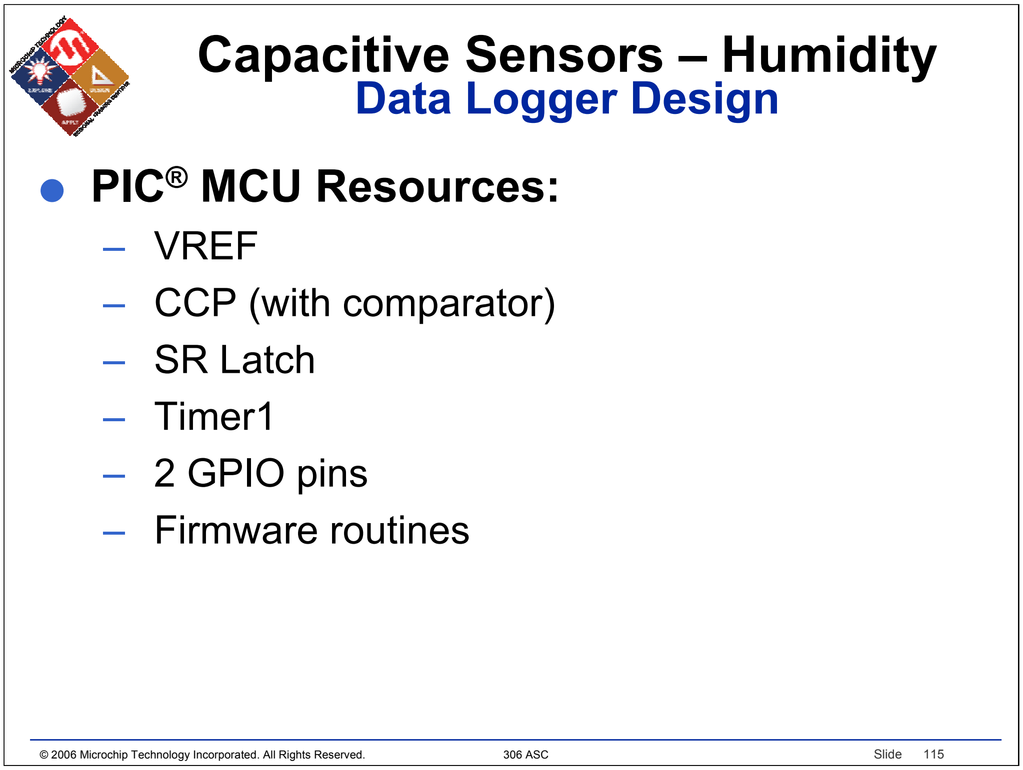
Changing the clock rate to 32.768 kHz will change T_{PIC} to approximately 122.1 µs. With R_{INT} = 6.65 MΩ, as shown in the circuit diagram, the measurement resolution would be approximately 24.41 pF / count. The next slide shows a solution, based on a discussion in AN1016, that would bring the resolution back to 0.1 pF / count.



As discussed on the last slide, we need to change the measurement resolution from 24.41 pF / count to 0.1 pF / count, because T_{PIC} changed from 0.5 µs to 122.1 µs. The attenuator formed by R_{A1}, R_{A2}, and R_{A3} has a modified gain of 0.005 V/V, so the measurement resolution would be 0.1221 pF / count if R_{INT} were left at 6.65 MΩ. Changing R_{INT} to 8.06 MΩ finalizes the change in measurement, giving a value of 0.1007 pF / count.



The op amp contributes greatly to the performance of this circuit.



This application requires much more from the PIC than the others we have covered.

Capacitive Sensors – Humidity
Exercise 4

- **Exercise 4**
  - Humidity Sensor PICtail™ Demo Board
  - See handout

 306 ASC Slide 116

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This is a paper design that integrates the three applications we covered in detail: thermistor, photodiode, and humidity sensor.

Data Logger Design
Comments

- **Comments**
  - Previously shown in this presentation
  - Possible future board
  - Need PIC® MCU that can accommodate Real Time Clock and Humidity Sensor application simultaneously

 306 ASC Slide 118

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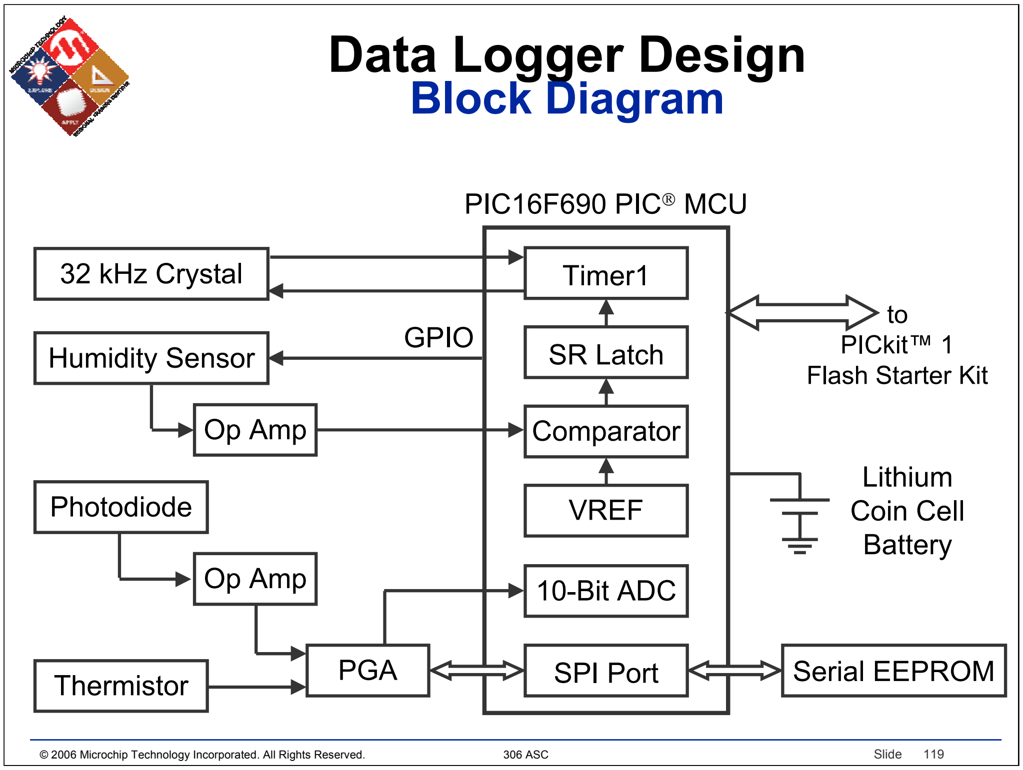


All of the building blocks on the left and inside the PIC have already been covered. The serial bus shown on the top right is a way for the PIC to communicate easily with a PC (via the PICkit 1 Flash Starter Kit board and GUI). The Lithium Coin Cell Battery makes this board portable, and is big enough to allow data collection over a significant period of time. The Serial EEPROM block represents SPI memory IC's that store the collected data.

This board collects data when in its stand alone mode, then it sends the data to the PC when it is connected via the PICkit 1.

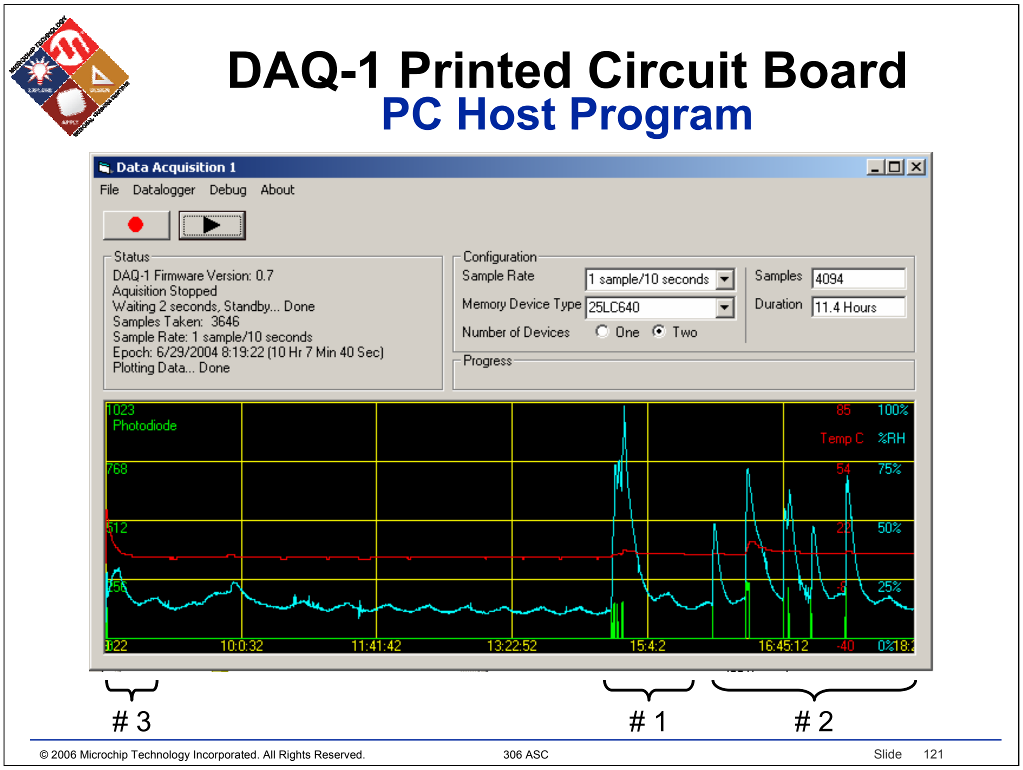
DAQ-1 Printed Circuit Board
Comments

- **Designed for previous MASTERs class**
  - Not released
- **Used alternate Humidity Sensor**
  - Integrated Solution
  - Powered

 306 ASC Slide 120

This board was part of the 2005 Masters presentation:

• DAQ 972, “Cool Data Acquisition Applications (or how to interface the PIC16F68X to the Real World)”



This screen shot was the result of leaving the DAQ-1 board in a fridge one day.

The curves are:

• The luminescence trace is green, and is shown as ADC code numbers (0 to 1023). It is the bottom curve.

• The temperature trace is red, with units of °C (-40°C to +85°C). It is the top curve on the left side.

• The relative humidity trace is blue, with units of % RH (0% to 100%). It is the middle curve on the left side, and shows the most variability.

The events are:

• # 1 – At 3 PM, the daughter arrives home from school and opens the fridge multiple times, over a short period of time, for a snack

• # 2 – Between 4 PM and 5 PM, the wife prepares dinner; she opens the fridge multiple times at well spaced intervals; she has planned her preparations well

• # 3 – At about 8 AM, the fridges auto-defrost cycle starts


HANDS-ON
Training
Summary
MICROCHIP TECHNOLOGY
EXPLORE
DESIGN
APPLY
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Slide 122

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Summary

- **We have covered sensor design issues**
  - Voltage, current, capacitance sensors
  - Filtering interference
  - PGA's
    - Non-linearity
    - MUXing

 306 ASC Slide 123

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Summary

- **We have focused on three applications**
  - Photodiode
  - Thermistor
  - Humidity Sensor

 306 ASC Slide 124

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Summary

- **We Have Met Our Objectives**
  - Reviewed fundamentals of analog conditioning
  - Learned how to design sensor conditioning circuits
  - Obtained hands-on experience
  - Worked on real applications
  - Received references to additional design resources (to follow)

 306 ASC Slide 125

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Summary

- **We Have Additional Resources**
  - See Appendix A
    - (immediately after this slide)

 306 ASC Slide 126

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All of the material in this appendix can be obtained at www.microchip.com.

References

- **General Topics**
  - Application Notes
    - AN866, "Designing Operational Amplifier Oscillator Circuits For Sensor Applications"
    - AN884, "Driving Capacitive Loads With Op Amps"
    - AN895, "Oscillator Circuits for RTD Temperature Sensors"
    - AN942, "Piecewise Linear Interpolation of PIC12/14/16 Series Microcontrollers"
    - AN990, "Analog Sensor Conditioning Circuits – An Overview"

 306 ASC Slide 128

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References

- MAPS
  - Combined cross reference and product selector guide
  - Compares technical specs
  - Simple GUI interface
  - Search for “MAPS” at www.microchip.com

 306 ASC Slide 129

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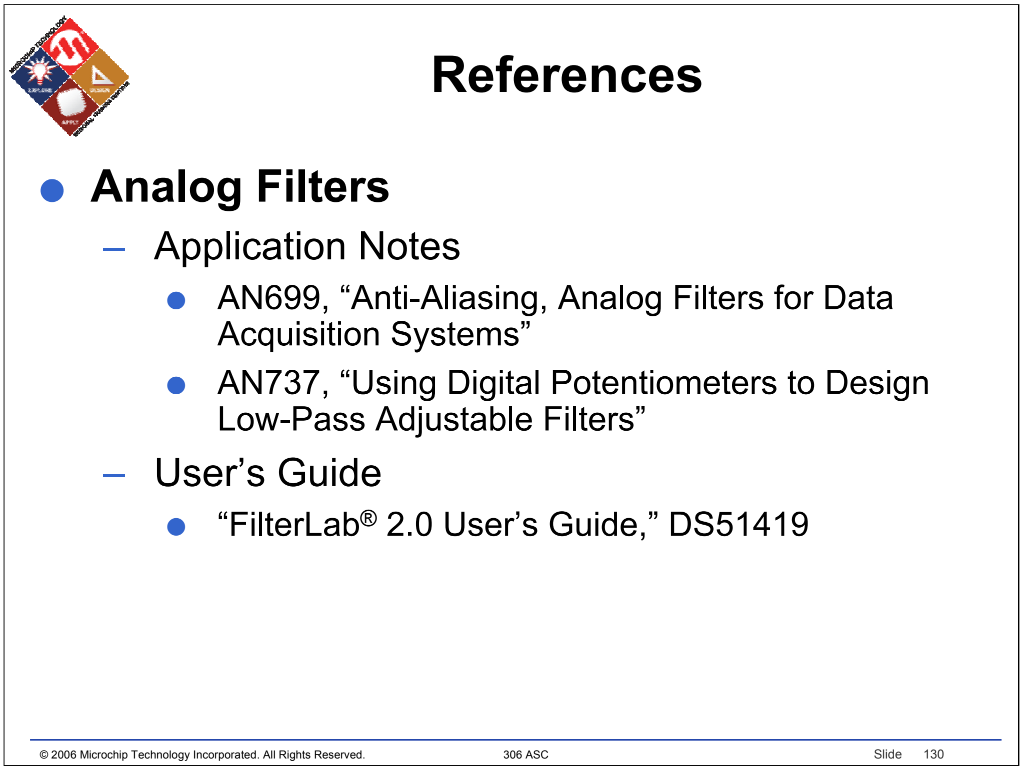

Download FilterLab® V2.0 from Microchip Technology Inc.'s website (www.microchip.com). The User's Guide for FilterLab can be found under the Help menu button.

References

- **PGA**
  - Application Notes:
    - AN248, “Interfacing MCP6S2X PGAs to PICmicro® Microcontroller”
    - AN867, “Temperature Sensing with a Programmable Gain Amplifier”
  - Web Seminars:
    - Thermistor Application for the New MCP6S9X PGA
      - Recorded 17 NOV 2004
    - Amplify sensor signals using the PGA
      - Recorded 24 SEP 2003

 306 ASC Slide 131

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References

- **Photodiode**
  - Application Notes:
    - AN682, “Using Single Supply Operational Amplifiers in Embedded Systems”
    - AN692, “Using a Digital Potentiometer to Optimize a Precision Single Supply Photo Detection Circuit”
    - AN951, “Amplifying High Impedance Sensors - Photodiode Example”

 306 ASC Slide 132

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References

- **Thermistor / Temperature**
  - Application Notes:
    - AN679, "Temperature Sensing Technologies"
    - AN685, "Thermistors in Single Supply Temperature Sensing Circuits"
    - AN679, "Temperature Sensing Technologies"
    - AN867, "Temperature Sensing with a Programmable Gain Amplifier"
    - AN897, "Thermistor Temperature Sensing with MCP6S2X PGA"
    - AN929, "Temperature Measurement Circuits for Embedded Applications"

 306 ASC Slide 133

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References

- Web Seminars:
  - Thermistor Application for the New MCP6S9X PGA
    - Recorded 17 NOV 2004

- **Capacitive / Humidity Sensor**
  - Application Notes:
    - AN1014, "Measuring Small Changes in Capacitive Sensors"
    - AN1016, "Detecting Small Capacitive Sensors Using the MCP6291 and PIC16F690 Devices"

 306 ASC Slide 134

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References

- **Demo Boards – For Exercises 1- 4**
  - PICkit™ 1 Flash Starter Kit
    - Order # DV164101
    - PICkit™ 1 Flash Starter Kit User's Guide (DS40051)
  - Signal Analysis PICtail™ Daughter Board
    - Order # AC164120
    - Signal Analysis PICtail™ Daughter Board User's Guide (DS51476)

 306 ASC Slide 135

It is possible to search our website using the Order Numbers.

References

- MCP6SX2 PGA Photodiode PICtail™ Demo Board
  - Order # MCP6SX2DM-PCTLPD
  - Photodiode PGA PICtail™ Daughter Board User's Guide (DS51514)
- MCP6SX2 PGA Thermistor PICtail™ Demo Board
  - Order # MCP6SX2DM-PCTLTH
  - MCP6SX2 PGA Thermistor PICtail™ Demo Board User's Guide (DS51517)

 306 ASC Slide 136

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References

- Humidity Sensor PICtail™ Demo Board
  - Order # PIC16F690DM-PCTLHS
  - Humidity Sensor PICtail™ Demo Board User's Guide (DS51594)
- Microchip Website Location
  - www.microchip.com
    - → Design
    - → Development Tools
    - → Demo Boards
    - → Analog
    - → Linear
    - → …

 306 ASC Slide 137

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References

- **Demo Boards – For Information Only**
  - MCP6S22 PGA PICtail™ Demo Board
    - Order # MCP6S22DM-PICTL
    - MCP6S22 PGA PICtail™ Demo Board User's Guide (DS51481)
  - MCP6S2X PGA Evaluation Board
    - Order # MCP6S2XEV
    - MCP6S2X Evaluation Board (Rev. 4) User's Guide (DS51327)

 306 ASC Slide 138

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References

- PICkit™ 2 Starter Kit
  - Order # DV164120
  - PICkit™ 2 Microcontroller Programmer USER'S GUIDE (DS51553)
- DAQ-1 Printed Circuit Board
  - DAQ 972, "Cool Data Acquisition Applications (or how to interface the PIC16F68X to the Real World)," Masters 2005 Class
  - Unavailable (a prototype board only)

 306 ASC Slide 139

Many of these demo boards are companions to application notes highlighted in this presentation.

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