TECHNICAL NOTE

ONE TECHNOLOGY WAY • P.O. BOX 9106 • NORWOOD, MASSACHUSETTS 02062-9106 • 781/329-4700

MINIMIZING POWER CONSUMPTION OF iMEMS ACCELEROMETERS

by Harvey Weinberg

Introduction

Portable battery powered devices are perhaps the largest growth market segment today. In the effort to reduce the size and weight of these devices, battery capacity is often minimized. To maintain good performance, designers are forced to carefully examine their circuits for ways to decrease power consumption.

This technical note will outline methods to reduce the power consumption of iMEMS accelerometers using both hardware and software techniques. A special section will cover some techniques that are specific to certain parts.

Basic Methods

There are three basic methods to reduce power consumption.

- Lower the supply voltage.
- Turn off the accelerometer when measurements are not taking place.
- Use clever software.

All of the techniques outlined in this application note are no more than extensions of these methods.

Lower the Supply Voltage

Often the most straightforward, and lowest cost, way to reduce power consumption is to simply reduce the supply voltage. Figure 1 shows the typical power consumption versus supply voltage for several iMEMS accelerometers.

While there are great power savings to be had by simply lowering the supply voltage, there is a price to be paid as well. As all these accelerometers are ratiometric, lowering their supply voltage will lower the sensitivity by roughly the same ratio. The exception to this is the PWM outputs of the ADXL202/210. These outputs remain fairly constant as the supply voltage changes.

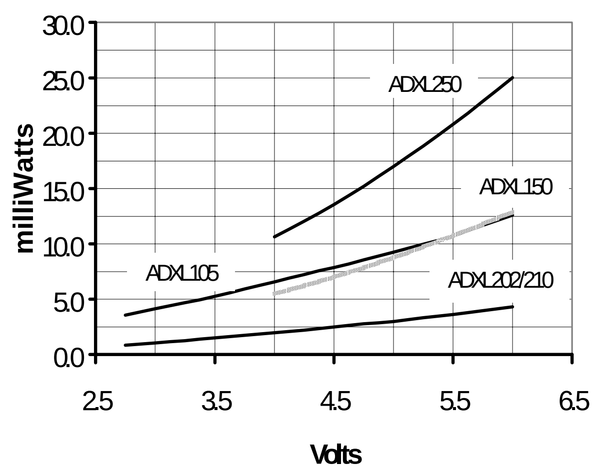

Figure 1. Power consumption vs Supply Voltage for Several Accelerometers

If the entire system is ratiometric (i.e. the A to D reference voltage is proportional to Vdd) this reduction of sensitivity is not a problem.

A potentially more serious problem is that the accelerometer's noise performance is generally degraded as the supply voltage is reduced. This cannot be mitigated by using a ratiometric system, and must be kept in mind during system design.

Turn Off the Accelerometer
Turning off the accelerometer when measurement is not occurring can result in great power savings. This is particularly true in applications where the sampling rate is low. Figure 2 illustrates the average power consumed by an ADXL150 versus the power cycling (sampling) frequency (for Vdd=5V and a 25µs A to D conversion time). Note that the Nyquist criteria must be satisfied in any case and the sampling frequency must be at least twice the input frequency.

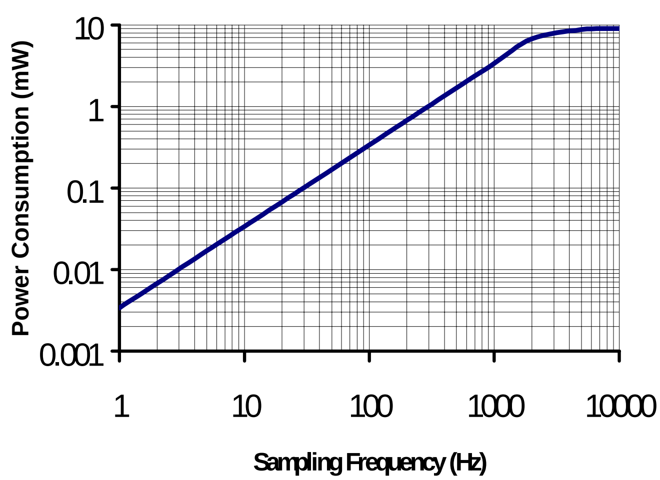

Figure 2. ADXL150 Average Power Consumption vs. Power Cycling Frequency

In iMEMS accelerometers the turn-on time is mainly a function of the bandwidth. While higher bandwidths will allow lower power operation (due to faster power cycling rates), it will also generally result in more noise. Many systems include an anti-aliasing filter between the accelerometer and the A to D converter. The time constant of this anti-aliasing filter must also be considered when power cycling.

Table 1 shows the approximate turn-on time (including the internal low pass filter) for several accelerometers. Faster turn-on times allow the user to very quickly turn on the accelerometer, measure the acceleration, and then turn off the accelerometer.

Table 1. Accelerometer Turn-On Time

MODEL	BANDWIDTH	TURN-ON
ADXL202/210	5000Hz	460µs
ADXL105	12KHz	700µS
ADXL150/250	1KHz	360µS
ADXL190	400Hz	750µS

Adding a single-pole low pass filter (anti-aliasing) to the accelerometer output will lengthen the turn on time by $5/(2\pi f)$ seconds, where f is the corner frequency of the filter.

For example, restricting the bandwidth of an accelerometer to 50Hz by adding a single-pole low pass filter would add 15.9ms to the settling time.

Low Power Design Example
Most low power designs are devices that measure movements that take place infrequently. These applications are ideal for power cycling. A good example is an automatic shutoff gas valve. In the event of an earthquake, the valve shuts down the natural gas supply to prevent ruptured gas pipes from leaking. Minor tremors (as can be created by large trucks passing by), and single impulse shocks (as would be generated by bumping into the valve) should be ignored.

Earthquakes are low frequency (below 20Hz), low *g*, multi-axis events, so we will use an ADXL202 and sample at 40Hz (25ms) to avoid aliasing.

In this design we can assume that vibrations of less than 200m*g* can be ignored (an earthquake is defined as having sufficient energy to cause accelerations greater than 200m*g*). So our peak-to-peak noise floor must be less than 200m*g*. Using peak-to-peak to RMS ratio of 6.6 we find:

RMS noise = 200m*g* / 6.6 = 30.3m*g*

So we will select a bandwidth that will result in an RMS noise floor that is less than 30.3m*g*.

$$\text{Noise} = \text{Noise density} * \sqrt{\text{bandwidth} * 1.6}$$

For the ADXL202 with a typical noise density of 500µ*g*/√Hz:

$$\text{Noise} = 500\mu g/\sqrt{\text{Hz}} * \sqrt{\text{bandwidth} * 1.6}$$

Re arranging the equation;

$$\text{Bandwidth} = (\text{noise}^2)/(1.6 * \text{noise density}^2)$$

In this example the maximum bandwidth is approximately 2.3KHz. Using the closest standard value we can set the bandwidth to 2Khz. Therefore Cx and Cy are 0.0022µF, and the noise floor is approximately 28m*g* RMS (185m*g* peak to peak).

For the ADXL202 the turn on time is approximately;

Ton = 160 * Cx + 0.3 ms

Where Cx is in µF. The 0.3ms is the turn on time of the accelerometer itself, while the 160 * Cx (or Cy) term is the settling time of the bandwidth limiting filter. Using a 0.0022µF capacitor the turn on time to steady state is approximately 650µs. We will be using the analog outputs and an A to D converter, so we must add 25µS for conversion time. So the total on time is 675µs.

Therefore the average power consumed is:

600µA * (0.675ms / 25ms) = 16.2µA

with a 5V supply. Lowering the supply voltage to 3V will reduce the average current consumption to 10.8µA.

Obviously 200m*g* peak-to-peak noise is too high to make a good measurement, therefore if a measurement of greater than 200m*g* is made the sampling speed can be increased for a few seconds to 800Hz and groups of 20 samples can be averaged. This will bring the noise floor down to approximately 6m*g* RMS (42m*g* peak to peak) allowing more precise measurement. More power will be utilized at this time, but as this happens infrequently, the average power consumed will still be under 20µA.

Software can then determine if the higher acceleration (>200m*g*) was actually due to an earthquake or just a single impulse event due to jostling and take appropriate action.

The ADXL202/210

The ADXL202/210 with it's PWM outputs is a special case and merits special attention when power cycling.

Using the PWM Outputs

The duty cycle modulator of the ADXL202/210 runs asynchronously to the rest of the accelerometer. Since we have no way of knowing what state the PWM output will be when the accelerometer analog output data is valid, we must wait at least one T2 period after the specified turn-on time (to ensure that the data is valid).

This leads us to two conclusions:

1. Using the analog outputs and an A to D converter rather than the PWM outputs will allow faster power cycling and therefore, lowest power operation.

2. If we choose to use the PWM outputs with power cycling, we should use the shortest T2 time possible. However using a short T2 implies using a fast counter, which is normally inconsistent with low power operation.

Although not explicitly specified in the data sheet, the maximum PWM frequency is typically 5KHz (Rset = 25KΩ). Therefore the additional overhead needed to use the PWM outputs will be at least 400µs (two T2 periods – 1 period wait for valid data, and another period for measurement (to calculate the counter time - the ADXL202 has a total dynamic range of ±4*g*, or │8000m*g*│. At T2=5KHz, 1mg per count = [(1/5KHz) / 8000m*g*] total range, or 250ηs per m*g*). In addition, if we want to use such a fast T2 period and maintain 10m*g* resolution (for the ADXL202) we would need a counter that counts in 250ηs increments

Even if only 32m*g* resolution is sufficient (roughly equivalent to 2° of tilt) the counter increment would rise to 800ηs. Still too fast for most low power 8-bit microcontrollers.

Using counter that runs at 1MHz (1µs per count) and maintaining 10m*g* resolution, we would have to run T2 at approximately 1.25KHz, and the T2 overhead would rise to 1.6ms compared to the 25µs or so that an A to D converter would take to complete a conversion. Clearly the overall system design becomes more involved when we are looking to minimize power consumption and use the PWM outputs of the ADXL202/210. Often the best choice when looking to minimize both component cost and power consumption is to use and A to D converter along with the ADXL202/210.

The exception to this is when the sampling rate is very low. In a system where a measurement is only made from time to time, the additional time required to use the PWM output is not a great handicap. The high resolution and low cost (i.e. no A to D converter required) may be more important than the additional power consumption.

Reducing Turn-On Time with Charge Conservation

In most applications most of the turn-on time of the ADXL202/210 is attributable to the time constant of the bandwidth limiting filter (formed by the internal 32KΩ resistors and Cx or Cy). The Cx and Cy values are normally dictated by the resolution required in the application. Many tilt sensing applications are low speed in nature and therefore good candidates for power cycling, however these applications often also require high resolution. So large Cx and Cy values are mandated resulting in long turn on times. In many cases the turn-on time can be greatly reduced by conserving the charge on Cx and Cy.

Figure 3 shows a typical circuit used for charge conservation. Cx and Cy are switched out of the circuit (via S2 and S3) just prior to the removal of Vdd to the ADXL202/210 (via S1). They are switched back in just after (a few µs) Vdd is applied to the ADXL202/210. This removes any path for the filter capacitors to discharge (other than leakage through Cx, Cy and the switches themselves). In an actual system a CMOS analog switch would be used for S1, S2, and S3.

By keeping the filter capacitors (Cx and Cy) from discharging we can speed up the settling time as shown in figure 4. Allowing us to turn off the accelerometer quickly and conserve power. Note that it takes the same time to arrive at steady state with or without charge conservation.

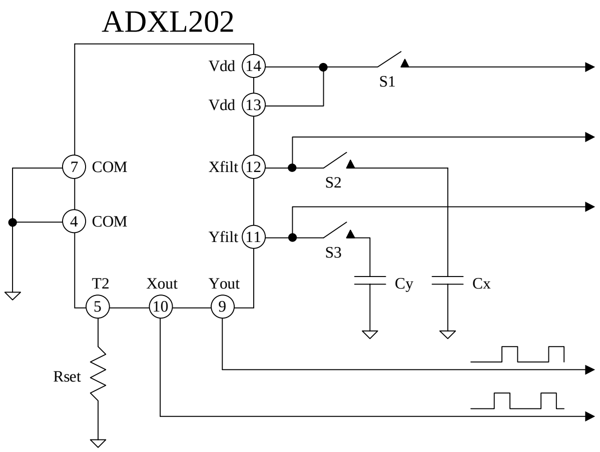

Figure 3. Charge Conservation Circuit

But if the goal is to arrive at measurements that are accurate within a certain tolerance of the steady state output (100m*g* as an example), this can be done more quickly using charge conservation. Figure 4 shows the Cy (pin 11) output of a system that is being power cycled 10ms ON/17ms OFF, for an effective bandwidth of about 18Hz. With Vdd = 5V and using a charge conservation architecture with Cx and Cy being 0.1µF, the average current used is:

Iavg = 600µA * (10ms/27ms) = 222µA

After 17ms the Cx and Cy voltage droops down to 1.81V. 10ms after Vdd is applied to the ADXL202, Cx and Cy are 2.471V. That is within 28mV (or 92m*g*) of the steady state output in this case.

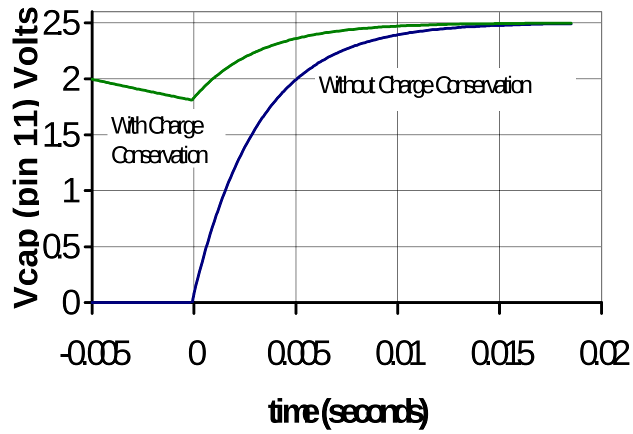

Figure 4. Cy Voltage With, and Without Charge Conservation

Had charge conservation not been used, we would have had an error of 107mV (or 342m*g*) versus the steady state output after 10ms. To get to within 92m*g* of steady state output without charge conservation we would have had to wait 14.1ms After turn-on. Resulting in an average current consumption of:

Iavg = 600µa * (14.1ms/27ms) = 313µA

A difference of 41%.

Had we lowered the supply voltage to 3.3V, the average current consumed would be 146µA using charge conservation and 207µA without.

Note that the time required to arrive within 100m*g* of the steady state response depends a great deal on the amount of droop, which is chiefly determined by the switch leakage.

Switch Selection

The performance of a charge conservation system depends greatly on the switch selected. The 74HC4016, and others like it, perform fairly well. However certain switches, like the metal gate CMOS 4016 type, should be avoided because of their high “ON” resistance at supply voltages of 5V or less. The 4066 and 74HC4066 should also be avoided because their internal architecture, tends to discharge the filter capacitors when switched. Precisely the opposite of what we wish to do!

There are many CMOS analog switches available that will work satisfactorily in this application. It is important to try out the switch you choose to see if it's performance is adequate.

Using I/O Ports Instead of Switches

In minimal systems the CMOS analog switch may be eliminated totally at the expense of some power consumption.

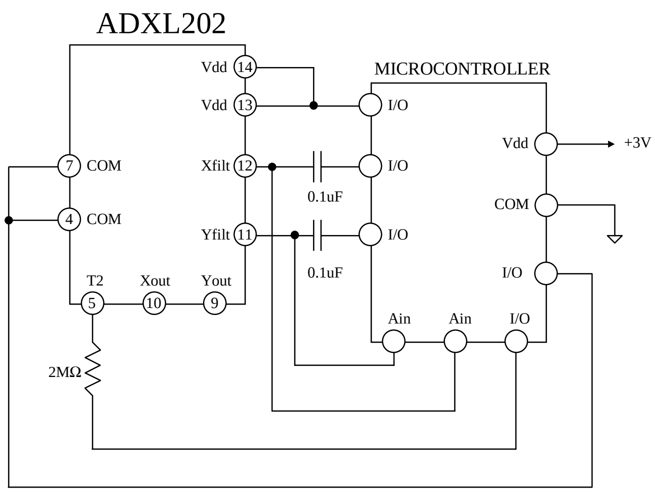

Figure 5. Using Microcontroller I/O

Rather than using CMOS analog switches to high-side switch the capacitors (as shown in Figure 3), one can use the output ports of a microcontroller to low side switch the capacitors as shown in Figure 5. Here the bandwidth limiting capacitors, Rset, and the ADXL202/210 are tri-stated when no measurements are being made so as to remove any discharge path for Cx and Cy. As the duty cycle outputs are not being used, we are free to choose a high value for Rset allowing us to save a few microamps.

Using I/O ports is usually a very low cost and compact solution as no additional components are required. However it is generally lower performance than analog CMOS switches as microcontroller I/O ports are not optimized for low leakage. Figure 6 outlines a model of the nodes at the filter capacitors when both sides are connected to I/O ports that are in tri-state mode. The value of the "leakage" resistors depends on the construction of the given microcontroller. Obviously the charged capacitor will discharge through the leakage resistors.

The performance of the system shown in Figure 5 is described in Table 2. A Microchip Technology PIC16C73 was used. Different microcontrollers may yield different results because of differences in the I/O structure.

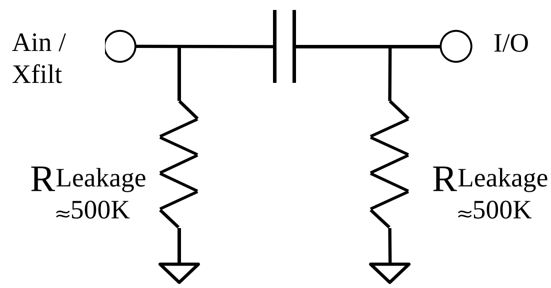

Figure 6. Model of Microcontroller Ports When in Tri-State Mode

Table 2. Peak to Peak Output Error vs. On Time for Figure 5. Vdd = 3.3V, Sampling Rate =36Hz, Bandwidth = 18Hz

Ton	Toff	Iavg	Error
infinity	0	400µA	0m*g*
11ms	17ms	157µA	32m*g*
9ms	19ms	128µa	80m*g*
7ms	21ms	100µA	160m*g*
5ms	23ms	87µA	320m*g*

If a slower (than 36Hz) sampling rate could be tolerated, additional power savings can be realized by lengthening the Toff time. For example, a 10Hz bandwidth (20Hz sampling rate) system with a Ton of 11ms and a Toff of 39ms would consume 87µA while still settling to within 32mg using the circuit shown in Figure 5.

Automatic Shut Off Gas Valve Revisited

Looking back at the earlier low power design example, we can further reduce the power consumption by using the charge conservation techniques discussed here.

Using a circuit similar to Figure 5, but with Cx and Cy 0.0022µF the settling time to 200m*g* is approximately 430µs plus 25µs of A to D acquisition time. So the average current consumed is;

600µA * 455µs/25ms = 10.9µA

with a 5V supply, and only 7.3µA with a 3V supply. A 33% reduction in current consumption compared to the previous example.

Conclusion

While iMEMS accelerometers are inherently low power devices, several techniques are available to designers to lower their power consumption even further. Generally, lower power operation results in degraded noise performance. So it is important for the designer to understand the compromises they will be making in pursuing very low power operation. Current consumption of well under 100µA is feasible in many low speed applications, and even under 10µA in some cases.