SECTION 6
POSITION AND MOTION SENSORS
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Modern linear and digital integrated circuit technology is used throughout the field of position and motion sensing. Fully integrated solutions which combine linear and digital functions have resulted in cost effective solutions to problems which in the past have been solved using expensive electro-mechanical techniques. These systems are used in many applications including robotics, computer-aided manufacturing, factory automation, avionics, and automotive.

This section is an overview of linear and rotary position sensors and their associated conditioning circuits. An interesting application of mixed-signal IC integration is illustrated in the field of AC motor control. A discussion of micromachined accelerometers ends the section.

Figure 6.1

LINEAR VARIABLE DIFFERENTIAL TRANSFORMERS (LVDTS)

The linear variable differential transformer (LVDT) is an accurate and reliable method for measuring linear distance. LVDTs find uses in modern machine-tool, robotics, avionics, and computerized manufacturing. By the end of World War II, the LVDT had gained acceptance as a sensor element in the process control industry largely as a result of its use in aircraft, torpedo, and weapons systems. The publication of *The Linear Variable Differential Transformer* by Herman Schaevitz in 1946 (Proceedings of the SASE, Volume IV, No. 2) made the user community at large aware of the applications and features of the LVDT.

The LVDT (see Figure 6.2) is a position-to-electrical sensor whose output is proportional to the position of a movable magnetic core. The core moves linearly inside a transformer consisting of a center primary coil and two outer secondary coils wound on a cylindrical form. The primary winding is excited with an AC voltage source (typically several kHz), inducing secondary voltages which vary with the position of the magnetic core within the assembly. The core is usually threaded in order to facilitate attachment to a nonferromagnetic rod which in turn in attached to the object whose movement or displacement is being measured.

Figure 6.2

The secondary windings are wound out of phase with each other, and when the core is centered the voltages in the two secondary windings oppose each other, and the net output voltage is zero. When the core is moved off center, the voltage in the secondary toward which the core is moved increases, while the opposite voltage decreases. The result is a differential voltage output which varies linearly with the core's position. Linearity is excellent over the design range of movement, typically

0.5% or better. The LVDT offers good accuracy, linearity, sensitivity, infinite resolution, as well as frictionless operation and ruggedness.

A wide variety of measurement ranges are available in different LVDTs, typically from ±100µm to ±25cm. Typical excitation voltages range from 1V to 24V RMS, with frequencies from 50Hz to 20kHz. Key specifications for the Schaevitz E100 LVDT are given in Figure 6.3.

Figure 6.3

Note that a true null does not occur when the core is in center position because of mismatches between the two secondary windings and leakage inductance. Also, simply measuring the output voltage V_{OUT} will not tell on which side of the null position the core resides.

A signal conditioning circuit which removes these difficulties is shown in Figure 6.4 where the absolute values of the two output voltages are subtracted. Using this technique, both positive and negative variations about the center position can be measured. While a diode/capacitor-type rectifier could be used as the absolute value circuit, the precision rectifier shown in Figure 6.5 is more accurate and linear. The input is applied to a V/I converter which in turn drives an analog multiplier. The sign of the differential input is detected by the comparator whose output switches the sign of the V/I output via the analog multiplier. The final output is a precision replica of the absolute value of the input. These circuits are well understood by IC designers and are easy to implement on modern bipolar processes.

Figure 6.4

Figure 6.5

The industry-standard AD598 LVDT signal conditioner shown in Figure 6.6 (simplified form) performs all required LVDT signal processing. The on-chip excitation frequency oscillator can be set from 20Hz to 20kHz with a single external capacitor. Two absolute value circuits followed by two filters are used to detect the amplitude of the A and B channel inputs. Analog circuits are then used to generate the ratiometric function [A–B]/[A+B]. Note that this function is independent of the amplitude of the primary winding excitation voltage, assuming the sum of the LVDT output voltage amplitudes remains constant over the operating range. This is usually the case for most LVDTs, but the user should always check with the manufacturer if it is not specified on the LVDT data sheet. Note also that this approach requires the use of a 5-wire LVDT.

Figure 6.6

A single external resistor sets the AD598 excitation voltage from approximately 1V RMS to 24V RMS. Drive capability is 30mA RMS. The AD598 can drive an LVDT at the end of 300 feet of cable, since the circuit is not affected by phase shifts or absolute signal magnitudes. The position output range of V_{OUT} is ±11V for a 6mA load and it can drive up to 1000 feet of cable. The V_A and V_B inputs can be as low as 100mV RMS.

The AD698 LVDT signal conditioner (see Figure 6.7) has similar specifications as the AD598 but processes the signals slightly differently. Note that the AD698 operates from a 4-wire LVDT and uses synchronous demodulation. The A and B signal processors each consist of an absolute value function and a filter. The A output is then divided by the B output to produce a final output which is ratiometric and independent of the excitation voltage amplitude. Note that the sum of the LVDT secondary voltages does not have to remain constant in the AD698.

Figure 6.7

The AD698 can also be used with a half-bridge (similar to an auto-transformer) LVDT as shown in Figure 6.8. In this arrangement, the entire secondary voltage is applied to the B processor, while the center-tap voltage is applied to the A processor. The half-bridge LVDT does not produce a null voltage, and the A/B ratio represents the range-of-travel of the core.

Figure 6.8

It should be noted that the LVDT concept can be implemented in rotary form, in which case the device is called a *rotary variable differential transformer* (RVDT). The shaft is equivalent to the core in an LVDT, and the transformer windings are wound on the stationary part of the assembly. However, the RVDT is linear over a relatively narrow range of rotation and is not capable of measuring a full 360º rotation. Although capable of continuous rotation, typical RVDTs are linear over a range of about ±40º about the null position (0º). Typical sensitivity is 2 to 3mV per volt per degree of rotation, with input voltages in the range of 3V RMS at frequencies between 400Hz and 20kHz. The 0º position is marked on the shaft and the body.

HALL EFFECT MAGNETIC SENSORS

If a current flows in a conductor (or semiconductor) and there is a magnetic field present which is perpendicular to the current flow, then the combination of current and magnetic field will generate a voltage perpendicular to both (see Figure 6.9). This phenomenon is called the *Hall Effect*, was discovered by E. H. Hall in 1879. The voltage, V_H, is known as the *Hall Voltage*. V_H is a function of the current density, the magnetic field, and the charge density and carrier mobility of the conductor.

Figure 6.9

The Hall effect may be used to measure magnetic fields (and hence in contact-free current measurement), but its commonest application is in motion sensors where a fixed Hall sensor and a small magnet attached to a moving part can replace a cam and contacts with a great improvement in reliability. (Cams wear and contacts arc or become fouled, but magnets and Hall sensors are contact free and do neither.) Since V_H is proportional to magnetic field and not to rate of change of magnetic field like an inductive sensor, the Hall Effect provides a more reliable low speed sensor than an inductive pickup.

Although several materials can be used for Hall effect sensors, silicon has the advantage that signal conditioning circuits can be integrated on the same chip as the sensor. CMOS processes are common for this application. A simple rotational speed detector can be made with a Hall sensor, a gain stage, and a comparator as shown in Figure 6.10. The circuit is designed to detect rotation speed as in automotive applications. It responds to small changes in field, and the comparator has built-in hysteresis to prevent oscillation. Several companies manufacture such Hall switches, and their usage is widespread.

Figure 6.10

There are many other applications, particularly in automotive throttle, pedal, suspension, and valve position sensing, where a linear representation of the magnetic field is desired. The AD22151 is a linear magnetic field sensor whose output voltage is proportional to a magnetic field applied perpendicularly to the package top surface (see Figure 6.11). The AD22151 combines integrated bulk Hall cell technology and conditioning circuitry to minimize temperature related drifts associated with silicon Hall cell characteristics.

The architecture maximizes the advantages of a monolithic implementation while allowing sufficient versatility to meet varied application requirements with a minimum number of external components. Principle features include dynamic offset drift cancellation using a chopper-type op amp and a built-in temperature sensor. Designed for single +5V supply operation, low offset and gain drift allows operation over a –40ºC to +150ºC range. Temperature compensation (set externally with a resistor R1) can accommodate a number of magnetic materials commonly utilized in position sensors. Output voltage range and gain can be easily set with external resistors. Typical gain range is usually set from 2mV/Gauss to 6mV/Gauss. Output voltage can be adjusted from fully bipolar (reversible) field operation to fully unipolar field sensing. The voltage output achieves near rail-to-rail dynamic range (+0.5V to +4.5V), capable of supplying 1mA into large capacitive loads. The output signal is ratiometric to the positive supply rail in all configurations.

Figure 6.11

OPTICAL ENCODERS

Among the most popular position measuring sensors, optical encoders find use in relatively low reliability and low resolution applications. An *incremental* optical encoder (left-hand diagram in Figure 6.12) is a disc divided into sectors that are alternately transparent and opaque. A light source is positioned on one side of the disc, and a light sensor on the other side. As the disc rotates, the output from the detector switches alternately on and off, depending on whether the sector appearing between the light source and the detector is transparent or opaque. Thus, the encoder produces a stream of square wave pulses which, when counted, indicate the angular position of the shaft. Available encoder resolutions (the number of opaque and transparent sectors per disc) range from 100 to 65,000, with absolute accuracies approaching 30 arc-seconds (1/43,200 per rotation). Most incremental encoders feature a second light source and sensor at an angle to the main source and sensor, to indicate the direction of rotation. Many encoders also have a third light source and detector to sense a once-per-revolution marker. Without some form of revolution marker, absolute angles are difficult to determine. A potentially serious disadvantage is that incremental encoders require external counters to determine absolute angles within a given rotation. If the power is momentarily shut off, or if the encoder misses a pulse due to noise or a dirty disc, the resulting angular information will be in error.

Figure 6.12

The *absolute* optical encoder (right-hand diagram in Figure 6.12) overcomes these disadvantages but is more expensive. An absolute optical encoder's disc is divided up into N sectors (N = 5 for example shown), and each sector is further divided radially along its length into opaque and transparent sections, forming a unique N-bit digital word with a maximum count of $2^N - 1$. The digital word formed radially by each sector increments in value from one sector to the next, usually employing Gray code. Binary coding could be used, but can produce large errors if a single bit is incorrectly interpreted by the sensors. Gray code overcomes this defect: the maximum error produced by an error in any single bit of the Gray code is only 1 LSB after the Gray code is converted into binary code. A set of N light sensors responds to the N-bit digital word which corresponds to the disc's absolute angular position. Industrial optical encoders achieve up to 16-bit resolution, with absolute accuracies that approach the resolution (20 arc seconds). Both absolute and incremental optical encoders, however, may suffer damage in harsh industrial environments.

RESOLVERS AND SYNCHROS

Machine-tool and robotics manufacturers have increasingly turned to resolvers and synchros to provide accurate angular and rotational information. These devices excel in demanding factory applications requiring small size, long-term reliability, absolute position measurement, high accuracy, and low-noise operation.

A diagram of a typical synchro and resolver is shown in Figure 6.13. Both sycnchros and resolvers employ single-winding rotors that revolve inside fixed stators. In the case of a simple synchro, the stator has three windings oriented 120º apart and electrically connected in a Y-connection. Resolvers differ from synchros in that their stators have only two windings oriented at 90º.

Figure 6.13

Because synchros have three stator coils in a 120º orientation, they are more difficult than resolvers to manufacture and are therefore more costly. Today, synchros find decreasing use, except in certain military and avionic retrofit applications.

Modern resolvers, in contrast, are available in a brushless form that employ a transformer to couple the rotor signals from the stator to the rotor. The primary winding of this transformer resides on the stator, and the secondary on the rotor. Other resolvers use more traditional brushes or slip rings to couple the signal into the rotor winding. Brushless resolvers are more rugged than synchros because there are no brushes to break or dislodge, and the life of a brushless resolver is limited only by its bearings. Most resolvers are specified to work over 2V to 40V RMS and at frequencies from 400Hz to 10kHz. Angular accuracies range from 5 arc-minutes to 0.5 arc-minutes. (There are 60 arc-minutes in one degree, and 60 arc-seconds in one arc-minute. Hence, one arc-minute is equal to 0.0167 degrees).

In operation, synchros and resolvers resemble rotating transformers. The rotor winding is excited by an AC reference voltage, at frequencies up to a few kHz. The magnitude of the voltage induced in any stator winding is proportional to the sine of the angle, θ, between the rotor coil axis and the stator coil axis. In the case of a synchro, the voltage induced across any pair of stator terminals will be the vector sum of the voltages across the two connected coils.

For example, if the rotor of a synchro is excited with a reference voltage, $V\sin\omega t$, across its terminals R1 and R2, then the stator's terminal will see voltages in the form:

$$\begin{aligned}
&\text{S1 to S3} = V \sin\omega t \sin\theta \\
&\text{S3 to S2} = V \sin\omega t \sin(\theta + 120^\circ) \\
&\text{S2 to S1} = V \sin\omega t \sin(\theta + 240^\circ),
\end{aligned}$$

where θ is the shaft angle.

In the case of a resolver, with a rotor AC reference voltage of $V\sin\omega t$, the stator's terminal voltages will be:

$$\begin{aligned}
&\text{S1 to S3} = V \sin\omega t \sin\theta \\
&\text{S4 to S2} = V \sin\omega t \sin(\theta + 90^\circ) = V \sin\omega t \cos\theta.
\end{aligned}$$

It should be noted that the 3-wire synchro output can be easily converted into the resolver-equivalent format using a Scott-T transformer. Therefore, the following signal processing example describes only the resolver configuration.

A typical resolver-to-digital converter (RDC) is shown functionally in Figure 6.14. The two outputs of the resolver are applied to cosine and sine multipliers. These multipliers incorporate sine and cosine lookup tables and function as multiplying digital-to-analog converters. Begin by assuming that the current state of the up/down counter is a digital number representing a trial angle, φ. The converter seeks to adjust the digital angle, φ, continuously to become equal to, and to track θ, the analog angle being measured. The resolver's stator output voltages are written as:

$$\begin{aligned}
V_1 &= V \sin\omega t \sin\theta \\
V_2 &= V \sin\omega t \cos\theta
\end{aligned}$$

where θ is the angle of the resolver's rotor. The digital angle φ is applied to the cosine multiplier, and its cosine is multiplied by V_1 to produce the term:

$$V \sin\omega t \sin\theta \cos\varphi.$$

The digital angle φ is also applied to the sine multiplier and multiplied by V_2 to product the term:

$$V \sin\omega t \cos\theta \sin\varphi.$$

These two signals are subtracted from each other by the error amplifier to yield an AC error signal of the form:

$$V \sin\omega t\,[\sin\theta \cos\varphi - \cos\theta \sin\varphi].$$

Using a simple trigonometric identity, this reduces to:

$$V \sin\omega t\,[\sin(\theta - \varphi)].$$

The detector synchronously demodulates this AC error signal, using the resolver's rotor voltage as a reference. This results in a DC error signal proportional to $\sin(\theta-\varphi)$.

The DC error signal feeds an integrator, the output of which drives a voltage-controlled-oscillator (VCO). The VCO, in turn, causes the up/down counter to count in the proper direction to cause:

$$\sin(\theta - \varphi) \to 0.$$

When this is achieved,

$$\theta - \varphi \to 0,$$

and therefore

$$\varphi = \theta$$

to within one count. Hence, the counter's digital output, φ, represents the angle θ. The latches enable this data to be transferred externally without interrupting the loop's tracking.

Figure 6.14

This circuit is equivalent to a so-called type-2 servo loop, because it has, in effect, two integrators. One is the counter, which accumulates pulses; the other is the integrator at the output of the detector. In a type-2 servo loop with a constant rotational velocity input, the output digital word continuously follows, or tracks the input, without needing externally derived convert commands, and with no steady state phase lag between the digital output word and actual shaft angle. An error signal appears only during periods of acceleration or deceleration.

As an added bonus, the tracking RDC provides an analog DC output voltage directly proportional to the shaft's rotational velocity. This is a useful feature if velocity is to be measured or used as a stabilization term in a servo system, and it makes tachometers unnecessary.

Since the operation of an RDC depends only on the ratio between input signal amplitudes, attenuation in the lines connecting them to resolvers doesn't substantially affect performance. For similar reasons, these converters are not greatly susceptible to waveform distortion. In fact, they can operate with as much as 10% harmonic distortion on the input signals; some applications actually use square-wave references with little additional error.

Tracking ADCs are therefore ideally suited to RDCs. While other ADC architectures, such as successive approximation, could be used, the tracking converter is the most accurate and efficient for this application.

Because the tracking converter doubly integrates its error signal, the device offers a high degree of noise immunity (12 dB-per-octave rolloff). The net area under any given noise spike produces an error. However, typical inductively coupled noise spikes have equal positive and negative going waveforms. When integrated, this results in a zero net error signal. The resulting noise immunity, combined with the converter's insensitivity to voltage drops, lets the user locate the converter at a considerable distance from the resolver. Noise rejection is further enhanced by the detector's rejection of any signal not at the reference frequency, such as wideband noise.

The ADS290 is one of a number of integrated RDCs offered by Analog Devices. Key specifications are shown in Figure 6.15. The general architecture is similar to that of Figure 6.14. The input signal level should be 2V RMS ± 10% in the frequency range from 3kHz to 20kHz.

Figure 6.15

INDUCTOSYNS

Synchros and resolvers inherently measure rotary position, but they can make linear position measurements when used with lead screws. An alternative, the Inductosyn™ (registered trademark of Farrand Controls, Inc.) measures linear position directly. In addition, Inductosyns are accurate and rugged, well-suited to severe industrial environments, and do not require ohmic contact.

The linear Inductosyn consists of two magnetically coupled parts; it resembles a multipole resolver in its operation (see Figure 6.16). One part, the scale, is fixed (e.g. with epoxy) to one axis, such as a machine tool bed. The other part, the slider, moves along the scale in conjunction with the device to be positioned (for example, the machine tool carrier).

Figure 6.16

The scale is constructed of a base material such as steel, stainless steel, aluminum, or a tape of spring steel, covered by an insulating layer. Bonded to this is a printed-circuit trace, in the form of a continuous rectangular waveform pattern. The pattern typically has a cyclic pitch of 0.1 inch, 0.2 inch, or 2 millimeters. The slider, about 4 inches long, has two separate but identical printed circuit traces bonded to the surface that faces the scale. These two traces have a waveform pattern with exactly the same cyclic pitch as the waveform on the scale, but one trace is shifted one-quarter of a cycle relative to the other. The slider and the scale remain separated by a small air gap of about 0.007 inch.

Inductosyn operation resembles that of a resolver. When the scale is energized with a sine wave, this voltage couples to the two slider windings, inducing voltages proportional to the sine and cosine of the slider's spacing within the cyclic pitch of the scale. If S is the distance between pitches, and X is the slider displacement within a pitch, and the scale is energized with a voltage V $\sin\omega t$, then the slider windings will see terminal voltages of:

$$V \text{ (sine output)} = V \sin\omega t \sin[2\pi X/S]$$

$$V \text{ (cosine output)} = V \sin\omega t \cos[2\pi X/S].$$

As the slider moves the distance of the scale pitch, the voltages produced by the two slider windings are similar to those produced by a resolver rotating through 360º. The absolute orientation of the Inductosyn is determined by counting successive pitches in either direction from an established starting point. Because the Inductosyn consists of a large number of cycles, some form of coarse control is necessary in order to avoid ambiguity. The usual method of providing this is to use a resolver or synchro operated through a rack and pinion or a lead screw.

In contrast to a resolver's highly efficient transformation of 1:1 or 2:1, typical Inductosyns operate with transformation ratios of 100:1. This results in a pair of sinusoidal output signals in the millivolt range which generally require amplification.

Since the slider output signals are derived from an average of several spatial cycles, small errors in conductor spacing have minimal effects. This is an important reason for the Inductosyn's very high accuracy. In combination with 12-bit RDCs, linear Inductosyns readily achieve 25 microinch resolutions.

Rotary inductosyns can be created by printing the scale on a circular rotor and the slider's track pattern on a circular stator. Such rotary devices can achieve very high resolutions. For instance, a typical rotary Inductosyn may have 360 cyclic pitches per rotation, and might use a 12-bit RDC. The converter effectively divides each pitch into 4096 sectors. Multiplying by 360 pitches, the rotary Inductosyn divides the circle into a total of 1,474,560 sectors. This corresponds to an angular resolution of less than 0.9 arc seconds. As in the case of the linear Inductosyn, a means must be provided for counting the individual pitches as the shaft rotates. This may be done with an additional resolver acting as the coarse measurement.

VECTOR AC INDUCTION MOTOR CONTROL

Long known for its simplicity of construction, low-cost, high efficiency and long-term dependability, the AC induction motor has been limited by the inability to control its dynamic performance in all but the crudest fashion. This has severely restricted the application of AC induction motors where dynamic control of speed, torque and response to changing load is required. However, recent advances in digital signal processing (DSP) and mixed-signal integrated circuit technology are providing the AC induction motor with performance never before thought possible. Manufacturers anxious to harness the power and economy of Vector Control can reduce R&D costs and time to market for applications ranging from industrial drives to electric automobiles and locomotives with a standard chipset/development system.

It is unlikely that Nikola Tesla (1856-1943), the inventor of the induction motor, could have envisaged that this workhorse of industry could be rejuvenated into a new class of motor that is competitive in most industrial applications.

Before discussing the advantages of Vector Control it is necessary to have a basic understanding of the fundamental operation of the different types of electric motors in common use.

Until recently, motor applications requiring servo-control tasks such as tuned response to dynamic loads, constant torque and speed control over a wide range were almost exclusively the domain of DC brush and DC permanent magnet synchronous motors. The fundamental reason for this preference was the availability of well understood and proven control schemes. Although easily controlled, DC brush motors suffer from several disadvantages; brushes wear and must be replaced at regular intervals, commutators wear and can be permanently damaged by inadequate brush maintenance, brush/commutator assemblies are a source of particulate contaminants, and the arcing of mechanical commutation can be a serious fire hazard is some environments.

The availability of power inverters capable of controlling high-horsepower motors allowed practical implementation of alternate motor architectures such as the DC permanent magnet synchronous motor (PMSM) in servo control applications. Although eliminating many of the mechanical problems associated with DC brush motors, these motors required more complex control schemes and suffered from several drawbacks of their own. Aside from being costly, DC PMSMs in larger, high-horsepower configurations suffer from high rotor moment-of-inertia as well as limited use in high speed applications due to mechanical constraints of rotor construction and the need to implement field weakening to exceed baseplate speed.

In the 1960's, advances in control theory, in particular the development of *indirect field-oriented control*, provided the theoretical basis for dynamic control of AC induction motors. Because of the intensive mathematical computations required by indirect field-oriented control, now commonly referred to as *vector control*, practical implementation was not possible for many years. Available hardware could not perform the high-speed precision sensing of rotor position and near real-time computation of dynamic flux vectors. The current availability of precision optical encoders, isolated gate bipolar transistors (IGBTs), high-speed resolver-to-digital converters and high-speed digital signal processors (DSPs) has pushed vector control to the forefront of motor development due to the advantages inherent in the AC induction motor.

A simplified block diagram of an AC induction motor control system is shown in Figure 6.17. In this example, a single-chip IC (ADMC300, ADMC330, or ADMC331) performs the control functions. The inputs to the controller chip are the motor currents (normally three-phase) and the motor rotor position and velocity. Hall-effect sensors are often used to monitor the currents, and a resolver and an RDC monitor the rotor position and velocity. The DSP is used to perform the real time vector-type calculations necessary to generate the control outputs to the inverter processors. The transformations required for vector control are also accomplished with the DSP.

The ADMC300 comprises a high performance, 5 channel 16-bit ADC system, a 12-bit 3-phase PWM generation unit, and a flexible encoder interface for position sensor feedback. The ADMC330 includes a 7 channel 12-bit ADC system and a 12-bit 3-phase PWM generator. The ADMC331 includes a 7 channel 12-bit ADC system, and a programmable 16-bit 3-phase PWM generator. It also has additional power factor correction control capabilities. All devices have on-chip DSPs (approximately 20MHz) based on Analog Device's Modified Harvard Architechure 16-bit DSP core. Third-party DSP software and reference designs are available to facilitate motor control system development using these chips.

Figure 6.17

ACCELEROMETERS

Accelerometers are widely used to measure tilt, inertial forces, shock, and vibration. They find wide usage in automotive, medical, industrial control, and other applications. Modern micromachining techniques allow these accelerometers to be manufactured on CMOS processes at low cost with high reliability. Analog Devices iMEMS® (Integrated Micro Electro Mechanical Systems) accelerometers represent a breakthrough in this technology. A significant advantage of this type of accelerometer over piezoelectric-type charge-output accelerometers is that DC acceleration can be measured (e.g. they can be used in tilt measurements where the acceleration is a constant 1g).

Figure 6.18

The basic unit cell sensor building block for these accelerometers is shown in Figure 6.19. The surface micromachined sensor element is made by depositing polysilicon on a sacrificial oxide layer that is then etched away leaving the suspended sensor element. The actual sensor has tens of unit cells for sensing acceleration, but the diagram shows only one cell for clarity. The electrical basis of the sensor is the differential capacitor (CS1 and CS2) which is formed by a center plate which is part of the moving beam and two fixed outer plates. The two capacitors are equal at rest (no applied acceleration). When acceleration is applied, the mass of the beam causes it to move closer to one of the fixed plates while moving further from the other. This change in differential capacitance forms the electrical basis for the conditioning electronics shown in Figure 6.20.

Figure 6.19

Figure 6.20

The sensor's fixed capacitor plates are driven differentially by a 1MHz square wave: the two square wave amplitudes are equal but are 180° out of phase. When at rest, the values of the two capacitors are the same, and therefore the voltage output at their electrical center (i.e., at the center plate attached to the movable beam) is zero. When the beam begins to move, a mismatch in the capacitance produces an output signal at the center plate. The output amplitude will increase with the acceleration experienced by the sensor. The center plate is buffered by A1 and applied to a synchronous demodulator. The direction of beam motion affects the phase of the signal, and synchronous demodulation is therefore used to extract the amplitude information. The synchronous demodulator output is amplified by A2 which supplies the acceleration output voltage, V_{OUT}.

An interesting application of low-g accelerometers is measuring tilt. Figure 6.21 shows the response of an accelerometer to tilt. The accelerometer output on the diagram has been normalized to 1g fullscale. The accelerometer output is proportional to the sine of the tilt angle with respect to the horizon. Note that maximum sensitivity occurs when the accelerometer axis is perpendicular to the acceleration. This scheme allows tilt angles from –90º to +90º (180º of rotation) to be measured. However, in order to measure a full 360º rotation, a dual-axis accelerometer must be used.

Figure 6.21

Figure 6.22 shows a simplified block diagram of the ADXL202 dual axis ±2g accelerometer. The output is a pulse whose duty cycle contains the acceleration information. This type of output is extremely useful because of its high noise immunity, and the data is transmitted over a single wire. Standard low cost microcontrollers have timers which can be easily used to measure the T1 and T2 intervals. The acceleration in g is then calculated using the formula:

$$A(g) = 8\,[T1/T2 - 0.5]\,.$$

Note that a duty cycle of 50% (T1 = T2) yields a 0g output. T2 does not have to be measured for every measurement cycle. It need only be updated to account for changes due to temperature. Since the T2 time period is shared by both X and Y channels, it is necessary to only measure it on one channel. The T2 period can be set from 0.5ms to10ms with an external resistor.

Analog voltages representing acceleration can be obtained by buffering the signal from the X_{FILT} and Y_{FILT} outputs or by passing the duty cycle signal through an RC filter to reconstruct its DC value.

Figure 6.22

A single accelerometer cannot work in all applications. Specifically, there is a need for both low-g and high-g accelerometers. Low-g devices are useful in such applications as tilt measurements, but higher-g accelerometers are needed in applications such as airbag crash sensors. Figure 6.23 summarizes Analog Devices family of ADXL accelerometers to date. Note that dual-axis versions as well as duty-cycle output versions are also available for some of the devices.

Figure 6.23

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