

approach can be particularly effective in combination with some of the aforementioned techniques for reducing stiction; however, it does require the introduction of several additional fabrication steps into the process flow.

4.4 MECHANICAL CHARACTERIZATION OF POLYSILICON

With the advent of surface micromachining technology, numerous nontraditional methods for the characterization of thin-film mechanical properties on the micrometer and submicrometer scales have been developed. These techniques involve the fabrication of micromechanical test structures that exhibit a controlled response to an internally or externally applied load that can be analytically related to various mechanical properties of the thin film. The micromechanical test structures can be placed at different locations across the wafer surface, thus providing an accurate picture of the on-wafer variation in these properties. Many of these devices can also be fabricated in parallel with actual device structures to serve as in-situ process control monitors.

Thin films used in the fabrication of surface-micromachined devices tend to have large residual stress fields that are extremely sensitive to both the deposition conditions of the film and its postdeposition process history. These residual stresses can affect the load-response behavior of the device, the frequency dependence of its output, and many other critical operating parameters; therefore they must be fully understood and carefully controlled if one is to develop a manufacturable surface-micromachining process. In the following section we will review some of the micromechanical test structures and discuss their application to the mechanical characterization of thin films.

4.4.1 Test Structures for In-Situ Characterization

One class of test structures used for evaluation of internal strain fields is based on the measurement of dimensional changes that occur upon the release of a stressed film from the underlying substrate. After release from the substrate, a stressed film is free to relax to its unstressed state, and the resultant contraction or elongation of the film is directly proportional to the magnitude of the internal strain field. These test structures have been devised to provide both direct and indirect measurements of the stress relaxation [80–83].

4.4.1.1 Cantilevers

Direct measurement of the strain relaxation can be achieved by measuring the contraction or elongation of a thin cantilever beam after release from the substrate [81].

The residual strain in the film is related to the dimensional change in the beam by the following expression:

$$\varepsilon = \frac{L_0 - L}{L_0} \quad (4.3)$$

Where ε is the residual strain, L_0 is the initial length of the cantilever beam, and L is the length of the beam after release from the substrate. The difference between L_0 and L is typically determined by comparing the end of the cantilever beam to fiducial marks patterned on the wafer surface [81].

This measurement technique can prove problematic when applied to films with relatively high elastic moduli, as is often the case in surface-micromachined devices. That is, the dimensional changes in the film tend to be quite small and difficult to measure, even when large aspect ratio structures are utilized. Senturia and coworkers have developed [80,82,83] several structures for the characterization of tensile films that effectively amplify the stress relaxation. However, for practical device dimensions the amplification factors are on the order of one-and-a-half to two; therefore devices of this type are still best suited to the measurement of films that exhibit large internal stress, such as polyimide or other polymeric materials.

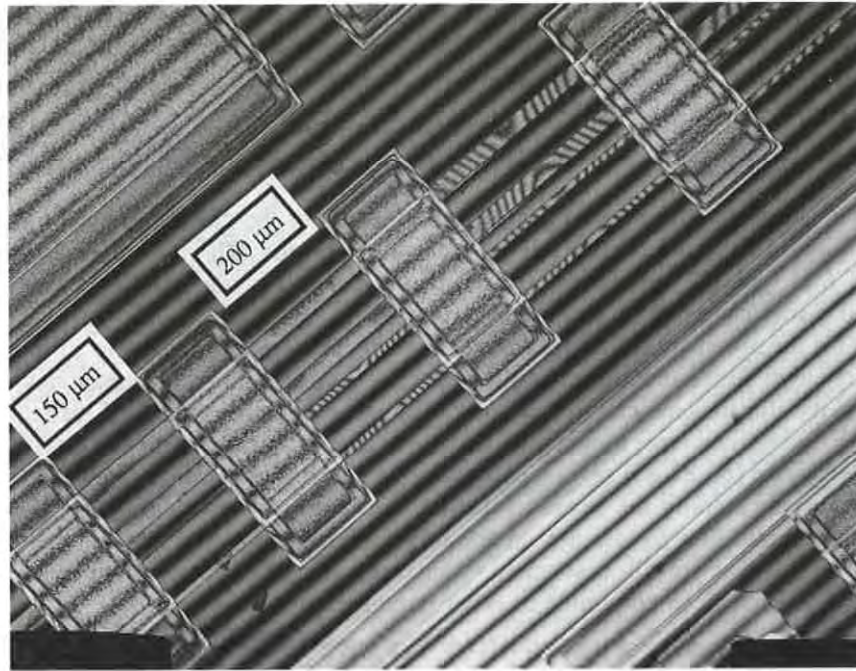
4.4.1.2 Bridges

Guckel et al. [24,25,47,84] have developed structures for the measurement of localized strain fields based on the tendency of slender members to buckle when a compressive load acting on its ends exceeds a critical value ε_{CR} . The Euler buckling strain for a thin beam under an axial load is given by [84–86]:

$$\varepsilon_{CR} = \frac{\pi^2 t^2}{KL^2} \quad (4.4)$$

where K is a constant related to the geometrical boundary conditions, t is the thickness of the thin film, and L is the length of the beam. For surface-micromachined bridges, the values for K lie between those of a simply supported beam ($K = 12$) and those of a doubly clamped beam ($K = 3$).

The upper and lower bounds for the residual strain in a compressively stressed thin film can be obtained by fabricating an array of bridge structures of varying lengths and observing the point at which the critical buckling strain is exceeded. An example of this technique is given in Figure 4.17, where an optical interference micrograph of a polysilicon bridge array is shown. Interference fringes on the unbuckled beams are not straight while those on the buckled beams are straight. In this array all bridges that exceed 150 μm in length have bowed, whereas all those that



(a)

Figure 4.17 Surface-micromachined bridge array for the evaluation of residual compressive strains in thin films: (a) optical interference micrograph of an array of polysilicon bridges; (b) theoretical buckling strains for the bridges in this array.

are shorter than $150\ \mu\text{m}$ remain straight. One can see from Figure 4.17(b) that this indicates that the magnitude of the internal strain lies between -3.2×10^{-4} and -5.6×10^{-4} . This is consistent with substrate curvature measurements taken on similarly processed films.

4.4.1.3 Ring and Beam Structures

A similar technique can be applied to the characterization of tensile films if one utilizes the ring and beam structure [24,25,47] shown schematically in Figure 4.18. When this structure is released from the substrate, the relaxation of the internal stress in the film will result in the application of a tensile load to the ring at the two opposing anchor points. As a result of this tensile load, the ring will want to contract

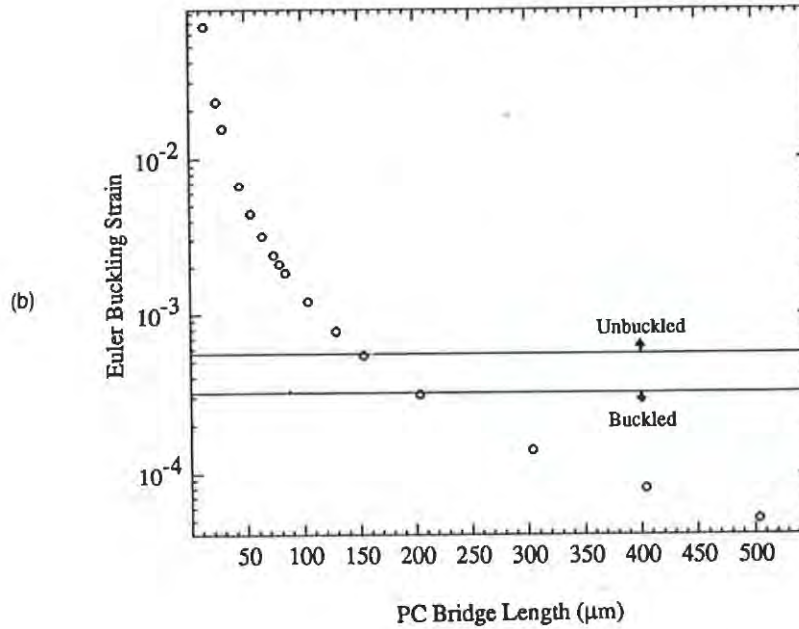


Figure 4.17 continued.

radially at the point where the ring joins the internal crossbar [87]. This radial contraction will introduce compressive stress in the crossbar that is related to the average residual strain in the film by the following expression:

$$\epsilon_{\text{crossbar}} = G\epsilon_{\text{film}} \quad (4.5)$$

where G is determined by the specific geometry of the ring and beam structure [24,25,47]. Combining (4.4) and (4.5) one obtains the following expression for the critical value of the residual strain required to buckle the crossbar:

$$\epsilon_{\text{film}}^{\text{CR}} = \left(\frac{\pi^2 t^2}{3(2R_{\text{CR}})^2} \right) \left(\frac{1}{G} \right) \quad (4.6)$$

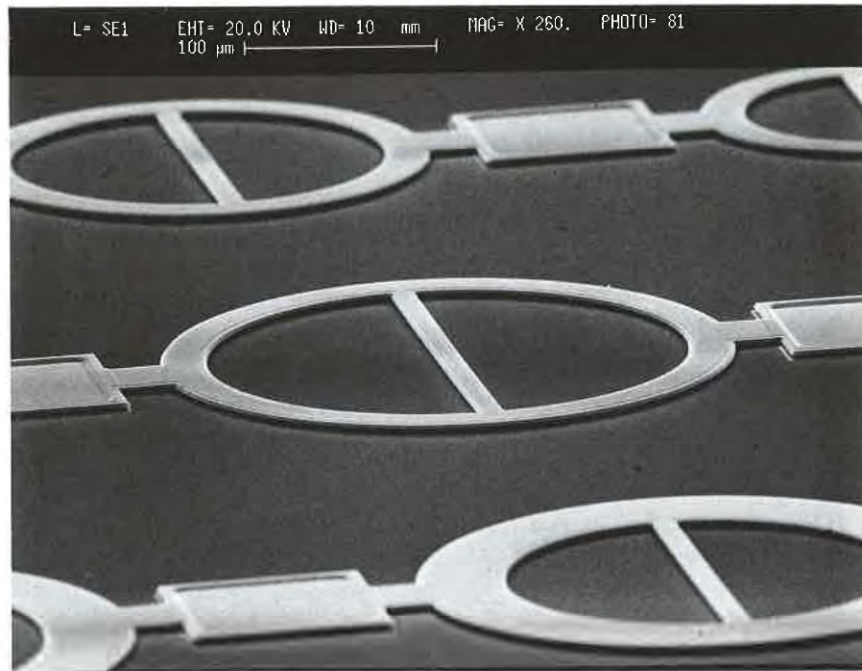


Figure 4.18 Schematic drawing of a ring-and-beam micromechanical test structure for the measurement of residual tensile stress in thin films (after [47]).

As in the bridge structures, upper and lower bounds for the residual strain in the film can be obtained by fabricating and analyzing an appropriate array of these test structures.

4.4.2 Gradient of Residual Stress

It should be noted that the expressions for the critical buckling stresses are truly valid only for films in which the residual stresses do not vary through the thickness of the film, as a residual strain gradient introduces an eccentricity to the axial load that will tend to alter the elastic stability of the beam [86]. In-plane stresses typically vary through the thickness of a thin film, particularly if the film has been subjected to substantial postdeposition processing [88]. This variation in the film stress results in an internal bending moment M whose magnitude is given by [88]:

$$M = \int_{-t/2}^{t/2} \sigma_x(y) y dy \quad (4.7)$$

Where t is the thickness of the thin film and $\sigma_x(y)$ is the magnitude of the in-plane stress at a distance y from the center of the film. This effective bending moment will cause the film to curl when it is released from the substrate, resulting in a nonplanar device structure. Several test structures have been devised to evaluate the magnitude of this internal stress gradient [88–90], the simplest being the surface-micromachined cantilever beam.

Relaxation of the internal stress gradients in the film will cause a surface-micromachined cantilever to bend out of the plane of the substrate upon release from the sacrificial oxide layer, as illustrated in Figure 4.6. The magnitude of this deflection can be measured using either a SEM or a phase-contrast microscope [64]. Such data can then be used to estimate the magnitude of the internal stress field.

For a residual stress gradient that remains constant along the length of the beam, the vertical deflection $\delta(x)$ at a point x can be described by [85]:

$$\frac{\delta(x)}{x} = K + \frac{(1 - \nu^2)}{2EI} Mx \quad (4.8)$$

Where x is the position along the length of the cantilever, K is a constant that depends upon the boundary conditions at the support point, $E/(1 - \nu^2)$ is the biaxial modulus of the film, and I is the moment of inertia of the beam about the z -axis. The first term in (4.8) is related to the geometrical boundary conditions at the constraint end; the second term describes the warping of the cantilever beam due to the residual strain gradient in the film. Figure 4.19 shows a plot of $\delta(x)/x$ versus x whose slope is proportional to $M(1 - \nu^2)/E$, which is process sensitive.

Fan et al. [90] have developed a spiral microstructure for the characterization of residual strain gradients in thin films. When fabricated from a thin film that exhibits a residual strain gradient, spiral structures tend to contract or expand upon release from the substrate. The resultant deformation provides three different response variables, which can be directly related to the residual strain gradient: endpoint rotation, endpoint height, and lateral contraction. The advantage of the Archimedean spiral structure is that the strain gradient can be estimated from a single data point, and two of the response variables (end rotation and lateral contraction) do not require the use of a SEM or phase contrast microscope to obtain the measurement. This substantially reduces the time required to characterize the thin film. However, the relatively large area required to fabricate Archimedean spirals with the requisite sensitivity is somewhat prohibitive. As an illustration, a SEM micrograph of an Archimedean spiral structure is shown in Figure 4.20. This particular structure is fabricated from a polysilicon film with a small residual stress gradient.

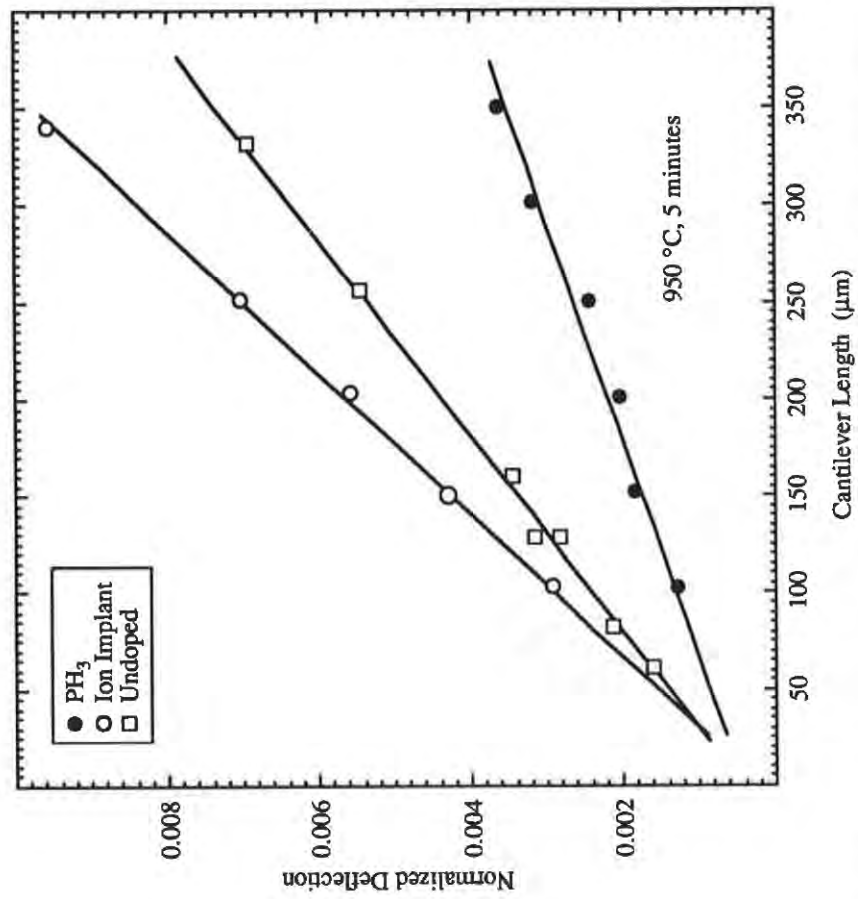


Figure 4.19 Plot of normalized deflection versus cantilever length for polysilicon cantilever beams that have been subjected to different process conditions.

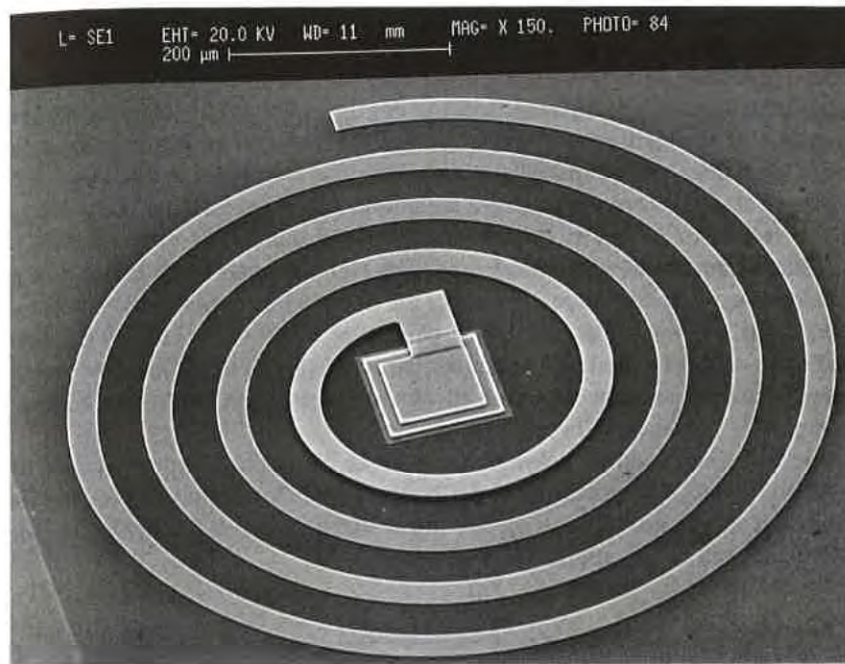


Figure 4.20 SEM micrograph of a polysilicon Archimedean spiral structure used for the characterization of internal stress gradients.

4.4.3 Load-Response Characterization

The preceding sections have discussed methods of measuring the magnitude and homogeneity of residual stresses in thin films. Numerous other thin film mechanical properties, such as Young's modulus, Poisson's ratio, and the yield strength of the material, play an equal if not greater role in determining the characteristics of a micromechanical device. These properties can also be quite sensitive to the nature of the film and therefore must be carefully monitored. Measurements designed to evaluate these properties typically involve characterizing the response of a micromechanical structure to an externally applied load. The variation in the response of the structure with the magnitude of the applied load can then be directly related to the mechanical properties of the materials from which it is fabricated. The remainder of this section will review some of these methods and their application to the evaluation of thin film mechanical properties.

Several authors have developed characterization techniques based on the load-response behavior of cantilever beams. The mechanics of a thin cantilever beam are well defined and closed-form solutions to the load-response behavior of this structure are available for a variety of different loading configurations [85], therefore the measured response of the cantilever beam to an applied load can easily be related to the mechanical properties of the material from which the cantilever is fabricated.

Weihs et al. [91] have developed a technique for characterizing the mechanical response of thin film cantilever beams using a load controlled submicrometer indentation instrument called a *Nanoindenter*. The Nanoindenter has a diamond tipped stylus, which is used to mechanically deflect the beam while continuously monitoring the applied load and the deflection of the cantilever beam. This is shown schematically in Figure 4.21(a). The load-response characteristics of the cantilever beam can be extracted from these data and used to obtain information about the elastic and plastic properties of the cantilever material.

Brittle materials such as polysilicon will exhibit linear-elastic behavior up to their breaking points. A typical load-deflection curve for a polysilicon cantilever beam is shown in Figure 4.21(b). In the linear-elastic regime the deflection of the beam $\delta(x)$ is related to the applied load P by:

$$\delta(x) = \frac{4Px^3(1 - \nu^2)}{Ewh^3} \quad (4.9)$$

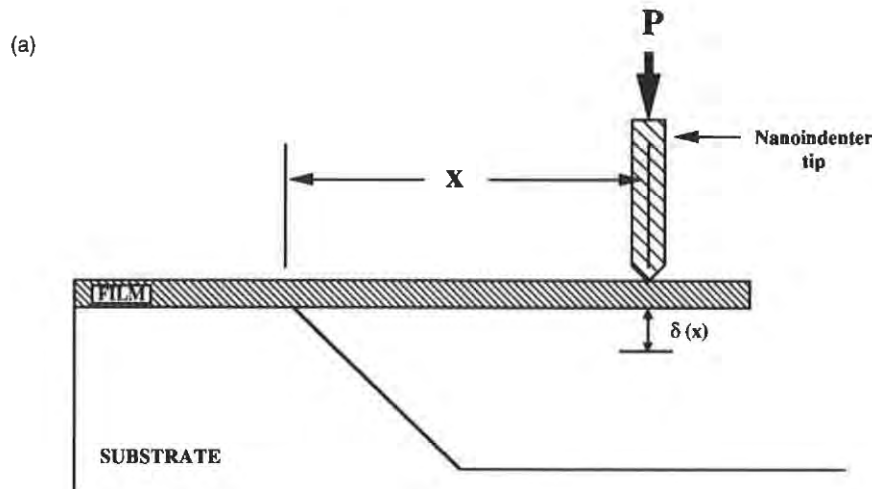


Figure 4.21 Mechanical characterization of thin films using the Nanoindenter technique: (a) schematic drawing of the measurement technique (after [92]); (b) load-deflection data for a surface-micromachined polysilicon cantilever beam.

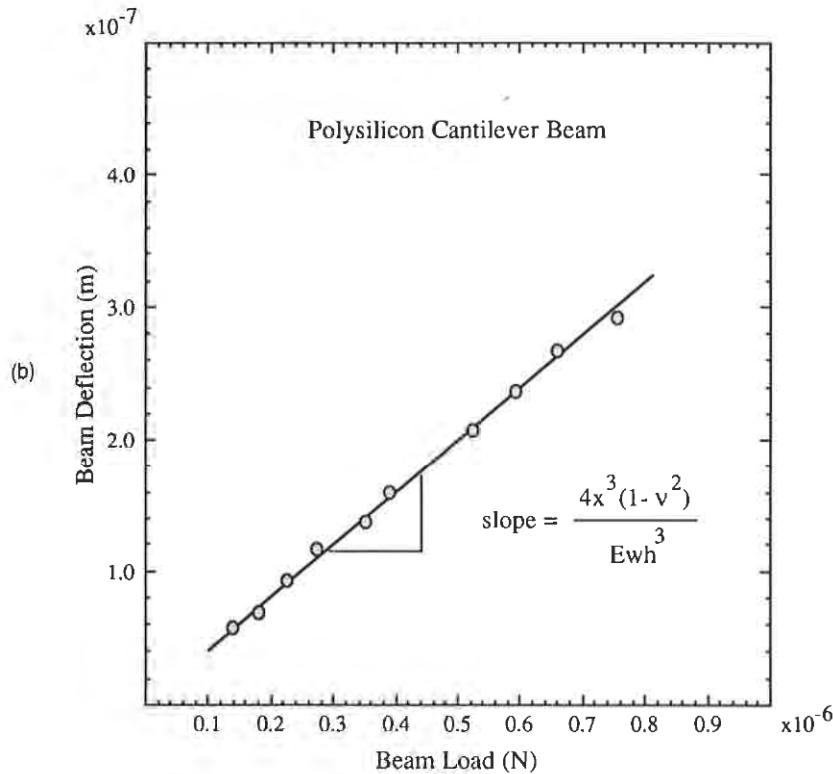


Figure 4.21 continued.

where x is the point at which the load is being applied and w is the width of the cantilever beam. $E/(1 - \nu^2)$ is the biaxial modulus of the material, which can be calculated from the slope of this line to the accuracy that the dimensions of the cantilever beam can be determined.

Under small loads, beams fabricated from metals or other nonbrittle materials will also exhibit linear-elastic behavior. However, the onset of plastic flow will cause the shape of the load-response curve to become nonlinear at higher loads. The load at which the response of the cantilever deviates from linear elastic behavior is defined as the yield load P_y . The yield strength σ_y of the material can be estimated from P_y by calculating the maximum stress in the beam under these conditions [91].

Dynamic properties of cantilever beams have also been used to evaluate the elastic properties of thin films [93–95], although this technique cannot be extended to the characterization of plastic deformation. The beam is electrostatically vibrated by the application of a sinusoidally varying voltage between the cantilever beam and

the substrate while the amplitude of oscillation of the cantilever beam is measured optically using the reflected laser light, as illustrated schematically in Figure 4.22(a). Variations in the vibrational amplitude of the cantilever beam are then monitored as a function of the frequency of applied voltage.

As shown in Figure 4.22(b), the vibrational amplitude of an underdamped cantilever beam goes through a maxima at some critical frequency f_{cr} [93]. This frequency is related to the mechanical resonance frequency of the cantilever beam by:

$$f_{cr} = f_r \left(1 - \frac{1}{4Q^2} \right)^{1/2} \quad (4.10)$$

where Q reflects the energy loss due to damping of the mechanical vibrations of the cantilever beam. Q can be extracted from the width of the resonance frequency or from the ratio of the peak vibrational amplitude to the amplitude of vibration at very low frequencies. The resonant frequency of a cantilever beam is a function of both the geometry of the device and the properties of the material or materials from which it is fabricated. It is related to these parameters by:

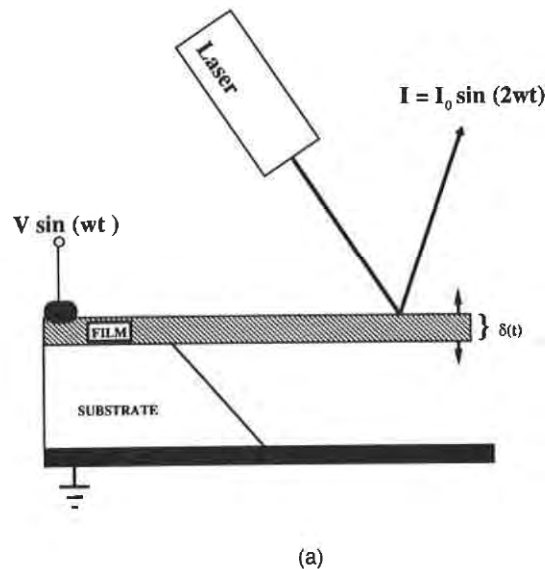


Figure 4.22 Mechanical characterization of thin films using resonant frequency measurements: (a) schematic drawing of the experimental setup; (b) typical frequency response of an SiO₂ cantilever beam; (c) resonant frequency versus beam length for an array of SiO₂ beams.

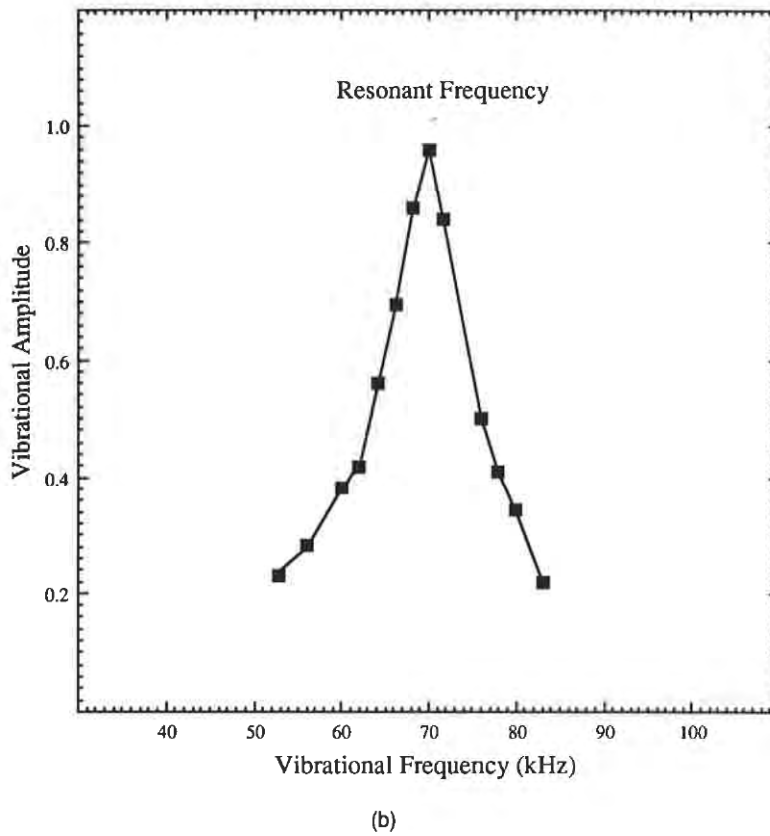
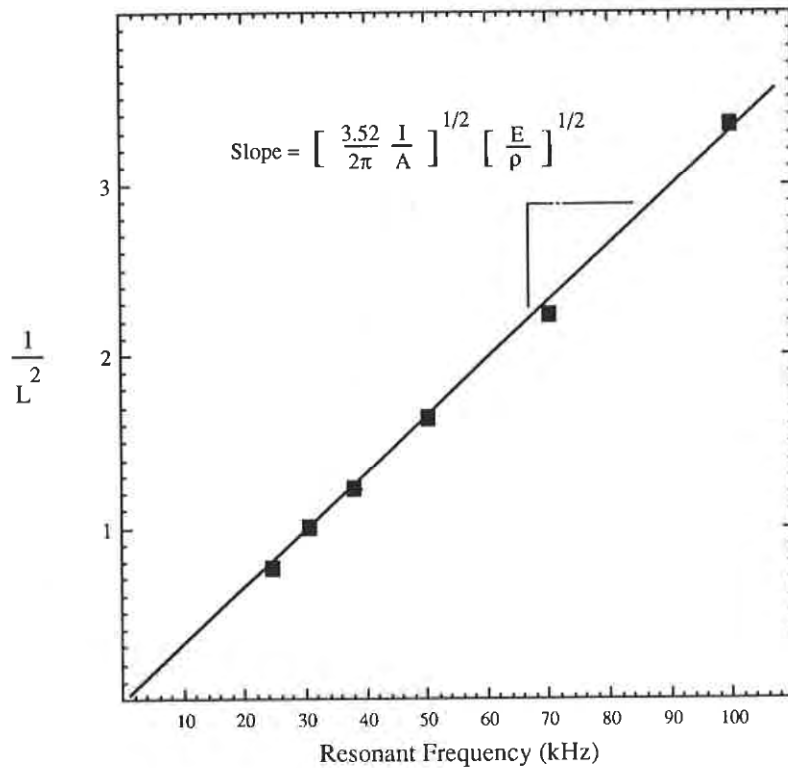


Figure 4.22 continued.

$$f_r^2 = \frac{3.52}{2\pi} \left(\frac{I}{AL^4} \frac{E}{\rho} \right) \quad (4.11)$$

where L is the length of the cantilever beam, A is its cross-sectional area, and I is its moment of inertia about the z -axis. E is the elastic modulus of the material from which the beam is fabricated, and ρ is the material density. A plot of $1/L^2$ versus f_r for an array of cantilever beams will yield a straight line whose inverse slope is proportional to E/ρ [93]. E/ρ can be calculated from the slope of this line using (4.11), and E can then be estimated from this value by using published values of the material density. Modulus values obtained in this manner are typically in reasonable agreement with bulk values of the elastic modulus.



(c)

Figure 4.22 continued.

Guckel and coworkers have used a similar technique to evaluate both the average residual stress field and the elastic properties of polysilicon thin films [24]. In this case, the resonating structure was a micromachined beam that was clamped at both ends rather than a cantilever beam. Constraining the beam at both ends causes the beam to bear an axial load proportional to the magnitude of internal stress field in the film. The resonant frequency of the structure varies with the magnitude of an applied load much like the pitch of a guitar string changes as the tension on the string is varied [24,96].

The response of micromachined diaphragms to an applied pressure can also be used to evaluate the mechanical properties of thin films [80,82,83,97,98]. To obtain

information on the mechanical properties of the material, a pressure differential is created across the diaphragm, causing the diaphragm to bow outward. The vertical displacement at the center of the diaphragm is then measured using an optical interferometer or other optical measurement technique. The center displacement, d , can be related to the applied pressure differential ($P_1 - P_0$) by the following expression:

$$P_1 - P_0 = A \frac{t}{a^2} \sigma_0 d + B f(\nu) \left(\frac{t}{a^4} \right) \left(\frac{E d^3}{1 - \nu} \right) \quad (4.12)$$

where $2a$ is the width of the diaphragm and t is the thickness of the thin film [97]. The term σ_0 is the average residual stress in the film, and $E/(1 - \nu)$ is the biaxial modulus. A and B are constants that depend upon the geometry of the device, and $f(\nu)$ is a function of Poisson's ratio for the diaphragm material [80,82,83,97,98]. With the appropriate geometrical constants, the load-deflection data can be fit to yield values for the average residual film stress as well as the biaxial modulus of the thin film, as illustrated in Figure 4.23, [80].

As micromachining technology matures, the level of understanding of the mechanical behavior of thin-film materials that is required increases tremendously. To address these issues, micromechanical versions of a wide range of bulk mechanical characterization techniques have been developed. A survey of the recent literature reveals discussions of devices for the characterization of friction and wear, fracture strain and fracture toughness [79,99,100], and many other thin-film mechanical properties. Because of space limitations we refer the reader to the available literature on these topics.

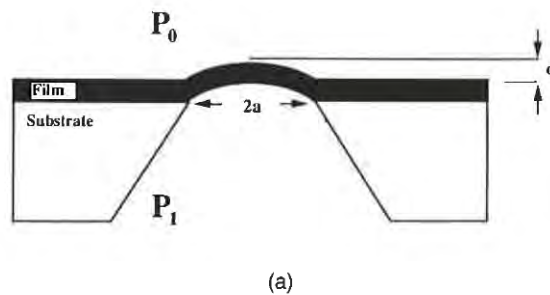
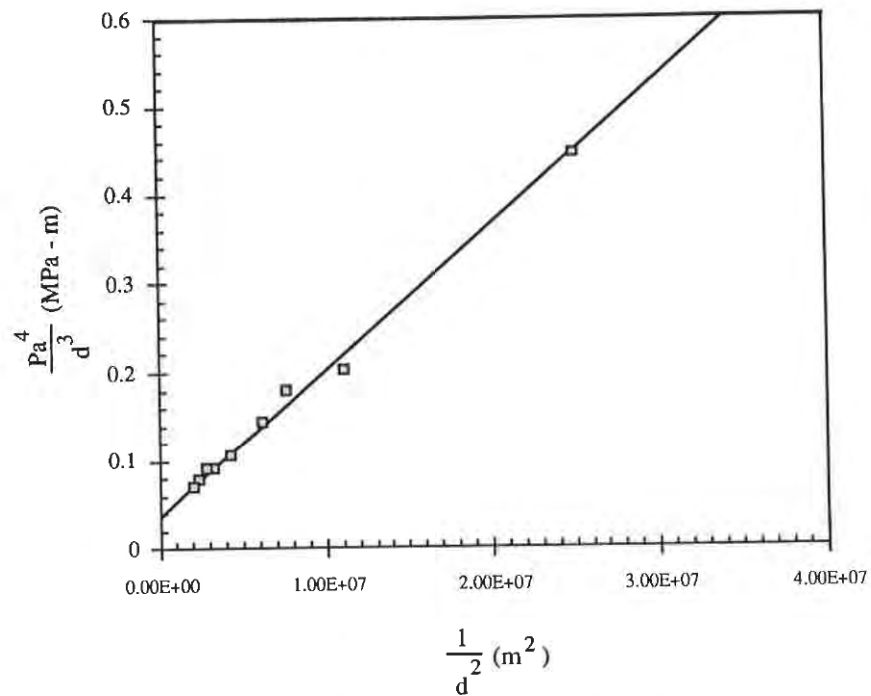


Figure 4.23 Mechanical characterization of thin films with the membrane-deflection technique: (a) schematic drawing of the test structure; (b) pressure-deflection data for polyimide membranes.



(b)

Figure 4.23 continued.

4.5 APPLICATION OF SURFACE MICROMACHINING

A wide variety of different sensors have already been produced using surface micromachining technology. A complete description would require a chapter in itself. Therefore, we have limited our choice of examples to two specific applications: the accelerometer and the electrostatically driven actuator. We feel that these examples illustrate the basic technology concepts quite adequately. In addition, this section will include discussion concerning integrated surface micromachined systems. This issue is becoming increasingly important with the current trend toward integration of sensor components with signal processing circuits.

4.5.1 Surface Micromachined Accelerometer

The development of a reliable accelerometer for automotive applications currently represents one of the most challenging tasks in the field of sensor technology. To

meet this challenge the designers must weigh the pros and cons of each different approach, such as piezoresistive, capacitive, or piezoelectric. Each approach has its inherent limitations.

The technology chosen for description here is a capacitive accelerometer manufactured by surface micromachining [101,102]. The basic structure is a double-pinned differential capacitor utilizing three polysilicon layers. The generic term *double-pinned* refers to a structure that is anchored to the substrate at locations on opposite sides of the suspended mass. Similar devices can be constructed using a cantilever-type design [101]. There are potentially significant differences in processing between the cantilever-type and the double-pinned structures. For example, for double-pinned designs it is desirable that the resultant residual stress in the polysilicon film be tensile to produce a flat undistorted structure. However, with the cantilever design the stress gradient through the thickness of the polysilicon is the critical constraint that must be minimized to avoid generation of a bending moment. An example of a double-pinned device is shown in Figure 4.24. The movable part (polysilicon layer 2) is composed of a suspended mass and four tethers that are anchored to the substrate. The differential capacitor in Figure 4.24 is made of three polysilicon layers:

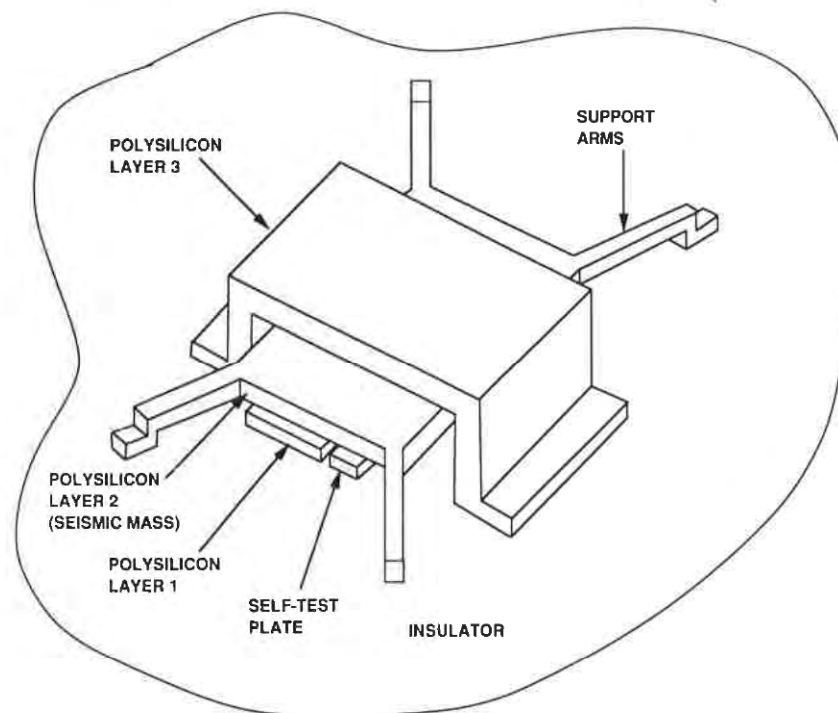


Figure 4.24 Polysilicon surface micromachined accelerometer.

the bottom capacitor is formed between polysilicon layer 2 and polysilicon layer 1, whereas the top capacitor is formed between the polysilicon layer 3 and polysilicon layer 2. The poly-1 and poly-3 layers are fixed. The basic principle of operation involves movement of the polysilicon-2 layer (suspended mass) under acceleration. This movement occurs in a direction perpendicular to the surface of the chip, and it is translated into a change of capacitance.

An additional feature of this device is a built-in self-test capability. This is achieved relatively easily by partitioning the poly-1 layer. That is, in addition to the poly-1 plate that makes up the bottom capacitance there is a self-test plate beneath poly-2. The self-test plate is electrically isolated from the bottom plate of the sense capacitor. When a voltage is applied to the self-test plate with respect to poly-2, electrostatic force will act on poly-2, causing a deflection that results in a change of capacitance, which in turn causes a change in the output voltage. The applied self-test voltage can be adjusted to correspond to a desired level of acceleration.

The differential capacitors of the micromachined structure can be coupled to either open- or closed-loop detector circuitry. An example of an open-loop configuration is shown in Figure 4.25. C_t and C_b are the top and bottom sensing capacitances of the micromechanical structure. The oscillator and reference blocks determine the internal clock frequency and reference voltages, respectively. The control logic contains digital circuitry that provides the pulses for proper operation of the switched-

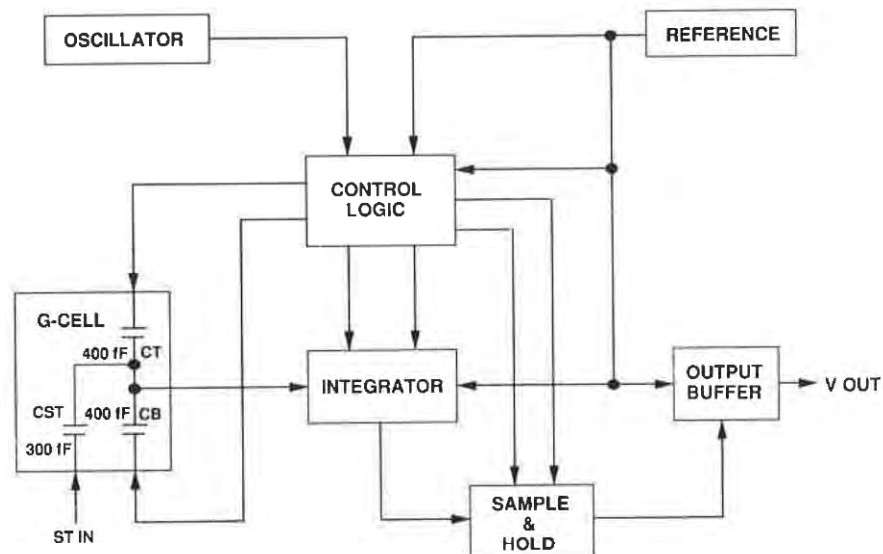


Figure 4.25 Block diagram of an open-loop accelerometer. C_t and C_b are sensing capacitors, and C_{st} is a self-test capacitor.

capacitor networks. These networks include an integrator and a sample-and-hold block. The switched-capacitor networks provide the output signal that is converted into analog form in the output buffer.

The response of a typical open loop system is shown in Figure 4.26. As can be seen, the output voltage is linear with a sensitivity of about 100 mV/g.

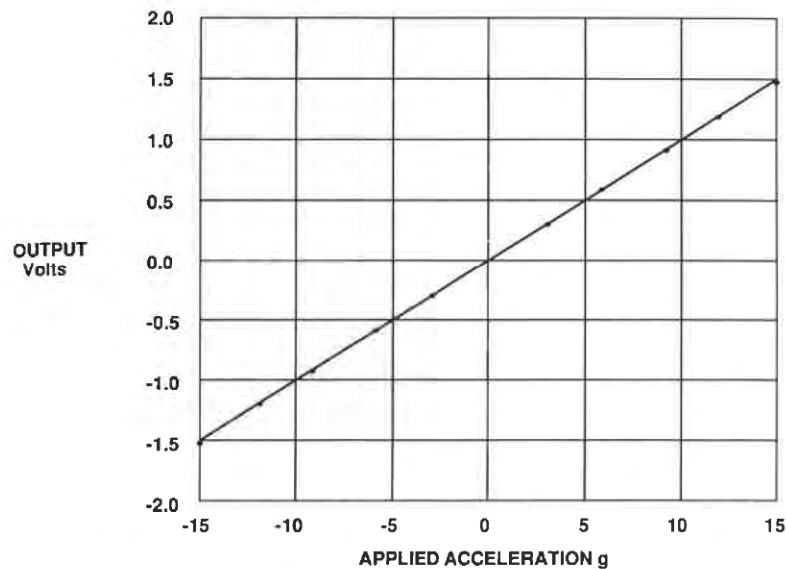


Figure 4.26 Output voltage as a function of acceleration for the open-loop system. The supply voltage is 5V. The sensitivity is 100 mV/g.

4.5.2 Electrostatically Driven Resonators

A particularly interesting group of surface micromachined devices is electrostatically driven actuators. They can be realized as micromotors [103,104] or comb-drive actuators [105]. For this class of devices the driving force is the electrostatic force created by a high electric field developed between the electrodes. The spacing between the electrodes is on the order of a few micrometers and the voltage between the electrodes is small, on the order of 10V. The spacing between the electrodes can also be adjusted into the submicrometer region [106].

In this section we focus on the electrostatic-comb-drive actuators that can be excited to work as resonators. The importance of resonators comes from their potential use for measurement of such quantities as pressure, acceleration, or vapor

concentration. In all of these applications physical or chemical parameters alter the potential or kinetic energy of the resonator, thus shifting the resonance frequency.

An example of a polysilicon resonator is shown in Figure 4.27. The resonator is made of three electrodes: a resonant plate (*B*-electrode), and two electrostatic comb drivers (*A* and *C* electrodes). The resonant structure is suspended by folded beams attached to the anchors. Fingers of the *B*-electrode are placed between the fingers of comb drivers *A* and *C*. The *A* and *C* electrodes are stationary and used for driving or sensing. To excite the resonant structure both a dc and an ac drive voltage are needed. Usually one stationary electrode is used for driving (*A*, for example) and the other is used for sensing (*C*-electrode). When the *B*-electrode is excited, it moves in a lateral direction, causing the change in comb capacitance. The change of capacitance is a function of finger dimensions as well as of the comb gap (space between the fingers). The current at the sense comb caused by the motion of the *B*-electrode is proportional to the change of capacitance and thus to the velocity of the *B*-electrode. This current can be measured by an electromechanical amplitude modulation scheme [105]. The resonance frequency of the structures is found to be in the range between 20 and 75 KHz, depending on structure dimensions. The quality factor depends on damping and is in the range between 20 and 100. Typical dc driving voltage is about 50V, and the amplitude of the ac driving voltage around 10V. It should be pointed out that resonators are also suitable for characterization of thin films. For example, for structures like the one shown in Figure 4.27, Young's modulus of polysilicon can be estimated [107]. Further, the resonant structures can be designed with straight beams instead of folded beams [107], in which case the intrinsic stress in the film remains in the structure causing axial forces. The existence of axial forces is the source of nonlinearity in the deflection motion of the structure and can be used to estimate the intrinsic stress.

4.5.3 Integration of Surface Micromachined Structures—Processing Issues

Integration of sensors and actuators with signal processing circuitry on the same chip offers numerous advantages, including higher sensitivity and accuracy, compensation of parasitic effects, faster data acquisition, improved reliability, and lower cost. However, integration is not an easy task, particularly when surface micromachined devices and the control circuitry have conflicting processing requirements. Therefore, in this section, we will discuss specific processing requirements of the surface micromachined structures and integrated circuits and outline possible solutions to problems associated with integration.

Analysis of polysilicon surface micromachined structures reveals several potential areas of incompatibility. First, the thickness of the polysilicon film used in surface micromachining is almost an order of magnitude greater than those used for the gate in IC technology. The thicker polysilicon is required to achieve structurally

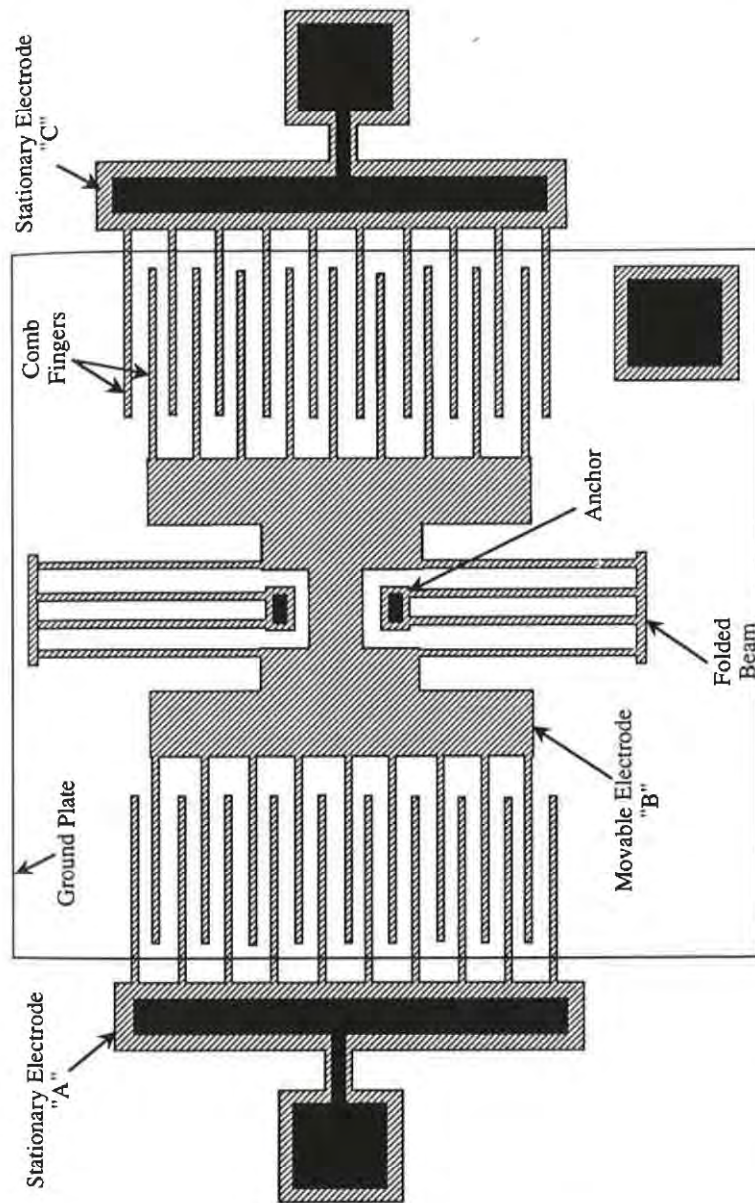


Figure 4.27 Electrostatic comb-drive actuator.

sound designs [35]. To distinguish this structural polysilicon from the gate polysilicon we will call it *thick polysilicon* in further discussion in this section. Second, the mechanical properties of polysilicon are extremely dependent on the deposition and annealing processing scheme. Finally, the thick polysilicon films have to be conductive because they are used simultaneously as mechanical and electrical elements.

The logical question, when it comes to integration, is how polysilicon surface micromachined structures should be integrated together with the circuitry using a full IC process, and at the same time, how can the specific process requirements of both technologies be met? Should one complete the structural layer processing followed by IC processing or vice versa?

Actually separating structural layer and IC processing completely is a difficult proposition. Full IC processing typically includes several high-temperature steps, such as initial oxidation, *p*-well drive-in (diffusion), *n*-well drive-in, field oxidation, gate oxidation, *n*+ source-drain drive-in, and *p*+ source-drain drive-in, or corresponding predeposition steps. All of these processes require temperatures in excess of 900°C. The mechanical properties of polysilicon are extremely sensitive to post-deposition thermal processing and required design specifications may not be achievable if the structural layers are deposited prior to high-temperature IC steps. If the thick polysilicon is deposited after full IC processing, three elements must be considered: (1) influence of polysilicon stress relaxation annealing on IC devices, (2) protection of IC devices during sacrificial etching, and (3) effects of doping intended for polysilicon on IC devices.

Annealing temperatures, if above 900°C for extended time periods, can modify circuit parameters resulting in catastrophic failure. One possible method to circumvent this problem is to achieve a high temperature quickly and maintain this temperature for only a very short time, using rapid thermal annealing (RTA) procedures [34,104]. The IC designs can also be more readily compensated to tolerate short (on the order of seconds) RTA cycles. Protection of the IC devices during sacrificial etching may be accomplished with a passivating layer of silicon nitride [34]. Finally, appropriate dopant levels in the thick polysilicon may be achieved either by ion implantation [34] or in-situ doping [104]. Both are relatively low temperature process steps and should not change parameters of already fabricated IC devices. In-situ doping offers advantages over ion implantation for certain surface micromachining applications because ion implantation can induce structural alterations in the polysilicon films.

4.6 LIGA PROCESS

The LIGA process is a technique for fabrication of three-dimensional micromechanical structures with high aspect ratios (height/width) having heights of several hundred micrometers. LIGA, an acronym for, lithography, electroforming, and micromolding

(in German, lithographie, galvanoförmung, abförmung) was first demonstrated in Germany [109]. The process combines X-ray lithography with thick resist layers and electroplated metal layers to form three-dimensional structures. It also can be combined with sacrificial layers [110,111], in which case it is called *sacrificial LIGA* (SLIGA) [111]. The basic steps in SLIGA are presented in Figure 4.28.

Processing starts with the deposition and patterning of the sacrificial layer. Requirements for the sacrificial layer include good adhesion to the substrate, good coverage of the sacrificial layer by a plating base, resistance to X-ray damage, and ease of removal during etching. Because the LIGA process involves temperatures no greater than 200°C, polyimide films meet all of the sacrificial layer requirements. Following deposition, the polyimide is patterned and then covered by a plating base that will be used later as a seed for electroplating of core material. The plating base (seed) for the example shown in Figure 4.28 consists of 150Å of titanium (adhesive metal) and 150Å of nickel. Both of these films are sputtered. The next step involves application of a thick photoresist layer (the photoresist can be as thick as several hundred micrometers, depending on the type of structure desired). The photoresist should have high selectivity between the exposed and unexposed areas, which, in turn, should produce vertical walls. This requirement can be satisfied with polymethylmethacrylate (PMMA) combined with aqueous developing system. After the application of PMMA, it is exposed to X-ray photons from a synchrotron. The synchrotron generates high-energy X-ray collimated photons needed to achieve complete exposure of the thick photoresist. The X-ray mask is also specially designed and constructed of a basic material (membrane) that is X-ray transparent. Silicon nitride or tensile polysilicon films can be used for this purpose. In addition, a patterned layer of gold on the membrane serves as an absorber. The combination of the membrane and absorber allows for locally exposed patterns that produce vertical photoresist walls after the development of PMMA.

Development of PMMA is followed by electroplating of a core material (Ni in this case) and subsequent removal of PMMA and the plating base in selective areas. The final step in SLIGA is etching of the sacrificial layer thus producing a suspended structure (bridge in Figure 4.28).

Note that further evolution of the LIGA process has led to polyimide-based processing [11], which does not require use of the synchrotron. This simplification of the technology will play an important role in future applications.

It should be emphasized that the LIGA process greatly expands micromachining capabilities, making possible vertical cantilevers, coils, microoptical devices, microconnectors, actuators, and so forth. One possible application of the SLIGA process is for constructing actuators. Guckle et al. provide an example of such an application [112]. Magnetic or thermal actuation mechanisms are accomplished by passing a current through the actuator. The actuator geometry determines the magnitude of sensitivity to one mechanism or the other, and the geometries can be tailored to emphasize one effect and suppress the other. The example in Figure 4.29

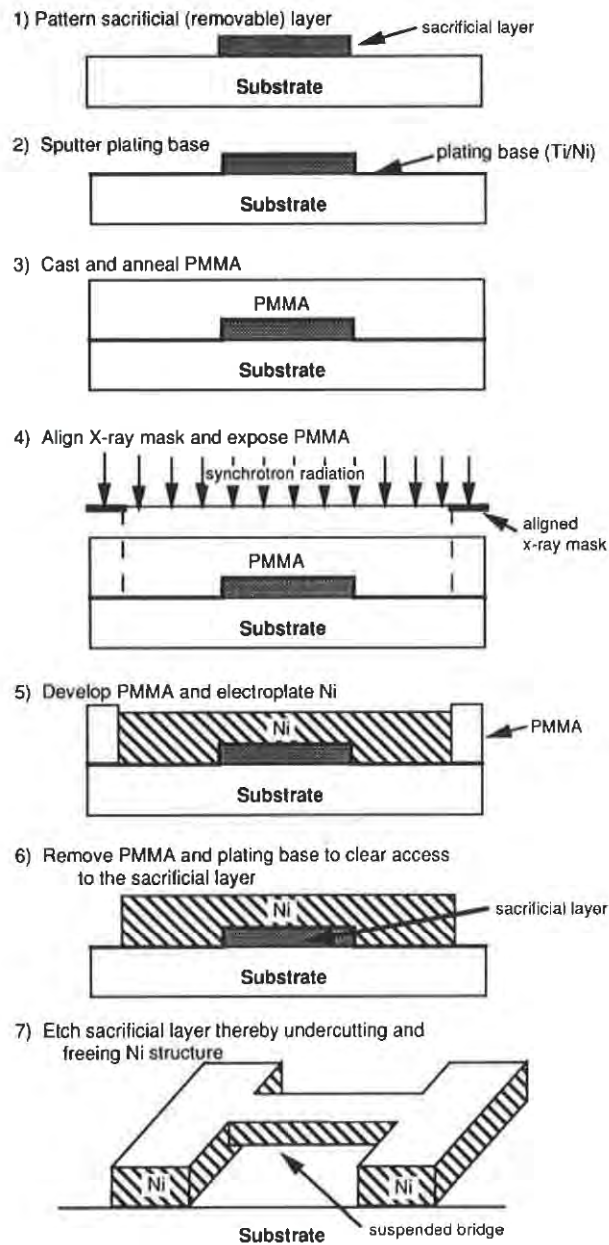


Figure 4.28 The SLIGA process.

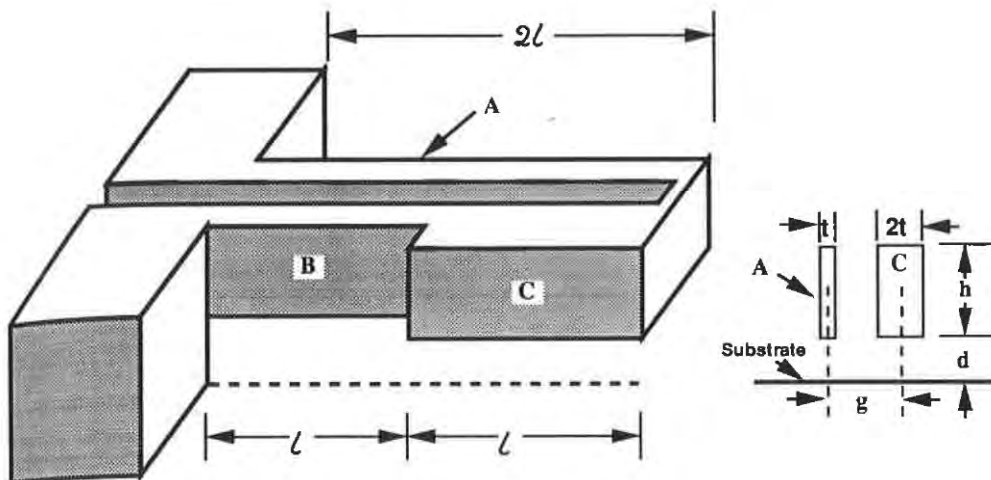


Figure 4.29 Thermally driven actuator: $l = 1000 \mu\text{m}$, $g = 10 \mu\text{m}$, $t = 4 \mu\text{m}$, $r = 50 \mu\text{m}$.

shows a cantilever designed to explore the thermal actuation mechanism. Sections A and B are identical in thickness but differ in length. Section C is thicker than A and B, resulting in a lower resistance. The length of A is equal to the length of B and C together. By passing a current through the device, dissipated power will raise its temperature. The lower resistance of section C will cause lower power dissipation in that area, which in turn will cause a smaller temperature increase in C relative to sections A and B. As a consequence, section A will expand more than sections B and C combined. The final result of the thermal actuation mechanism is a deflection of the beam. This is a good example of how geometric differences (asymmetry) can be used as a driver for the thermal actuation mechanism. It is interesting to note that the measured deflection of the tip of the beam, made of nickel with overall length of 1 mm and height of $50 \mu\text{m}$, exceeds $100 \mu\text{m}$. This type of actuator can be used either as a microswitch or for position sensing.

4.7 SUMMARY

The focus of this chapter was on polysilicon surface micromachining technology. The basic concept of surface micromachining was presented as were the key processing issues pertinent to polysilicon. The discussion included deposition conditions, doping, annealing, and stress of the films. Furthermore, attention was paid to sacrificial layers and sacrificial etching including stiction. Also, a section was devoted to test structures for mechanical characterization of polysilicon. Finally, several examples of practical applications of surface micromachining were cited. In addition, this chapter contains a brief section describing the LIGA process.

REFERENCES

- [1] H. C. Nathanson, W. E. Newell, R. A. Wickstrom, and J. R. Davis, Jr., "The Resonant Gate Transistor," *IEEE Trans. Electron Dev.*, Vol. ED-14, 1967, p. 117.
- [2] L. S. Fan, Y. C. Tai, and R. S. Muller, "Pin Joints, Gears, Springs, Cranks, and Other Novel Micromechanical Structures," *Tech. Digest, Transducers '87*, 1987, p. 849.
- [3] K. J. Gabriel, W. S. N. Trimmer, and M. Mehregany, "Microgears and Turbines Etched From Silicon," *Tech. Digest, Transducers '87*, 1987, p. 853.
- [4] S. Sugiyama, T. Suzuki, K. Kawahata, K. Shimaoka, M. Takigawa, and I. Igarashi, "Micro-Diaphragm Pressure Sensor," *Tech. Digest, IEDM*, Los Angeles, 1986, p. 184.
- [5] R. Gutteridge, J. Schmiesing, and Lj. Ristic, "Characterization of Poly-SiO₂: Application for Dynamic Micromechanical Structures," unpublished results.
- [6] G. Blackburn and J. Janata, "The Suspended Mass Ion Selective Field Effect Transistor," *Electrochem. Soc. Spring Mtg*, Vol. 82-1, Montreal, 1982, p. 196.
- [7] W. Yun, "CMOS Metallization For Integration With Micromachining Processes," M.S. thesis, University of California, Berkeley, 1989.
- [8] R. B. Brown, M. L. Ger, and T. Nguyen, "Characterization of Molybdenum Thin Films for Micromechanical Structures," *Proc. IEEE MEMS '90*, Napa Valley, CA, 1990, p. 77.
- [9] L. Tong and M. Mehregany, "Silicon Carbide as a New Micromechanics Material," *Tech. Digest, Solid-St. Sensor and Actuator Workshop*, Hilton Head Island, SC, 1992, p. 198.
- [10] J. A. Walker, K. J. Gabriel, and M. Mehregany, "Thin-Film Processing of TiNi Shape Memory Alloy," *Sensors and Actuators*, Vols. A21-23, 1990, p. 243.
- [11] C. H. Ahn and M. G. Allen, "A Fully Integrated Micromagnetic Actuator With a Multilevel Meander Magnetic Core," *Tech. Digest, IEEE Solid-St. Sensor and Actuator Workshop*, Hilton Head Island, SC, 1992, p. 14.
- [12] T. Tawagawa, P. Schiller, H. Yoon, and D. L. Polla, "Micromachined Zinc Oxide Thin Film Sensors," Abs. no. 762, ECS Fall Mtg, Seattle, 1990, p. 1084.
- [13] F. Shemansky, Lj. Ristic, D. Koury, and H. Hughes, "Mechanical Properties of Polysilicon Based Composite Films," unpublished results.
- [14] H. Guckel and D. W. Burns, "Planar Processed Polysilicon Sealed Cavities for Pressure Transducer Arrays," *Tech. Digest, IEDM*, 1984, p. 223.
- [15] C. H. Mastrangelo and R. S. Muller, "Vacuum-Sealed Silicon Micromachined Incandescent Light Source," *Tech. Digest, IEDM*, 1989, p. 503.
- [16] R. S. Rosler, "Low Pressure CVD Production Processes for Poly, Nitride, and Oxide," *Solid State Technology*, 1977, p. 63.
- [17] T. I. Kamins, M. M. Mandurah, and K. C. Saraswat, "Structure and Stability of Low Pressure Chemically Vapor Deposited Silicon Films," *J. Electrochem. Soc.*, Vol. 125, No. 6, 1978, p. 927.
- [18] C. Linder and N. F. De Rooij, "Investigations on Free Standing Polysilicon Beams in View of Their Application as Transducers," *Sensors and Actuators*, Vol. A21-A23, 1990, p. 1053.
- [19] W. A. P. Claassen, J. Bloem, W. G. J. N. Valkenburg, and C. H. J. Van Den Brekel, "The Deposition of Silicon from Silane in a Low Pressure Hot Wall System," *J. Crystal Growth*, Vol. 57, 1982, p. 259.
- [20] A. M. Beers and J. Bloem, "Temperature Dependence of the Growth Rate of Silicon Prepared Through Chemical Vapor Deposition from Silane," *Appl. Phys. Lett.*, Vol. 41, No. 2, 1982, p. 153.
- [21] G. Harbeke, L. Krausbauer, E. F. Steigmeier, A. E. Widmer, H. F. Kappert, and G. Neugebauer, "High Quality Polysilicon by Amorphous Low Pressure Chemical Vapor Deposition," *Appl. Phys. Lett.*, Vol. 42, No. 3, 1983, p. 249.
- [22] G. Harbeke, L. Krausbauer, E. F. Steigmeier, A. E. Widmer, H. F. Kappert, and G. Neugebauer,

- "Growth and Physical Properties of LPCVD Polycrystalline Silicon Films," *J. Electrochem. Soc.*, Vol. 131, No. 3, 1984, p. 675.
- [23] P. Joubert, B. Loisel, Y. Chouan, and L. Haji, "The Effect of Low Pressure on the Structure of LPCVD Polycrystalline Silicon Films," *J. Electrochem. Soc.*, Vol. 134, No. 10, 1987, p. 2541.
 - [24] H. Guckel, D. W. Burns, H. A. C. Tilmans, D. W. DeRoo, and C. R. Rutigliano, "Mechanical Properties of Fine Grained Polysilicon the Repeatability Issue," *Tech. Digest*, IEEE Solid State Sensor and Actuator Workshop, Hilton Head Island, SC, 1988, p. 96.
 - [25] H. Guckel and D. W. Burns, "Polysilicon Thin Film Process," U.S. Patent # 4,897,360, 1990.
 - [26] P. Krulevitch, R. T. Howe, G. C. Johnson, and J. Huang, "Stress in Undoped LPCVD Polycrystalline Silicon," *Tech. Digest*, Transducers '91, 1991, p. 949.
 - [27] T. I. Kamins, "Design Properties of Polycrystalline Silicon," *Sensors and Actuators*, Vols. A21-A23, 1990, p. 817.
 - [28] L. S. Fan and R. S. Muller, "As Deposited Low Strain LPCVD Polysilicon," *Tech. Digest*, IEEE Solid State Sensor and Actuator Workshop, Hilton Head Island, SC, 1988, p. 55.
 - [29] E. Kinsbron, M. Sternheim, and R. Knoell, "Crystallization of Amorphous Silicon Films During Low Pressure Chemical Vapor Deposition," *Appl. Phys. Lett.*, Vol. 42, No. 9, 1983, p. 835.
 - [30] H. Guckel, T. Randazszo, and D. W. Burns, "A Simple Technique for the Determination of Mechanical Strain in Thin Films with Applications to Polysilicon," *J. Appl. Phys.*, Vol. 57, No. 5, 1985, p. 1671.
 - [31] H. Guckel, J. J. Sniogowski, T. R. Christenson, and F. Raissi, "The Application of Fine Grained, Tensile Polysilicon Mechanically Resonant Transducers," *Sensors and Actuators*, Vols. A21-A23, 1990, p. 346.
 - [32] H. Guckel, D. W. Burns, C. R. Rutigliano, D. K. Showers, and J. Uglow, "Fine Grained Polysilicon and its Application to Planar Pressure Transducers," *Tech. Digest*, Transducers '87, 1987, p. 277.
 - [33] H. Guckel and D. W. Burns, "Measurement and Control of Mechanical Properties of Thin Films," IV International Workshop on the Physics of Semiconductor Devices, Madras, India, 1987.
 - [34] M. W. Putty, S. Chang, R. T. Howe, A. L. Robinson, and K. D. Wise, "Process Integration for Active Polysilicon Resonant Microstructures," *Sensors and Actuators*, Vol. 20, 1989, p. 143.
 - [35] R. T. Howe and R. S. Muller, "Polycrystalline and Amorphous Silicon Micromechanical Beams: Annealing and Mechanical Properties," *Sensors and Actuators*, Vol. 4, 1983, p. 447.
 - [36] R. T. Howe and R. S. Muller, "Stress in Polycrystalline and Amorphous Silicon Thin Films," *J. Appl. Phys.*, Vol. 54, 1983, p. 4674.
 - [37] T. I. Kamins and T. R. Cass, "Structure of Chemically Deposited Polycrystalline Silicon Films," *Thin Solid Films*, Vol. 16, 1973, p. 147.
 - [38] J. Adamczewska and T. Budzynski, "Stress in Chemically Vapor Deposited Silicon Films," *Thin Solid Films*, Vol. 113, 1984, p. 271.
 - [39] W. A. Bryant, "The Kinetics of the Deposition of Silicon by Silane Pyrolysis at Low Temperatures and Atmospheric Pressure," *Thin Solid Films*, Vol. 60, 1979, p. 19.
 - [40] F. C. Eversteyn and B. H. Put, "Influence of AsH₃, PH₃, and B₂H₆ on the Growth Rate and Resistivity of Polycrystalline Silicon Films Deposited from a SiH₄-H₂ Mixture," *J. Electrochem. Soc.*, Vol. 120, No. 1, 1973, p. 106.
 - [41] P. Krulevitch, T. D. Nguyen, G. C. Johnson, R. T. Howe, H. R. Wenk, and R. Gronsky, "LPCVD Polycrystalline Silicon Thin Films: The Evolution of Structure, Texture and Stress," *Mat. Res. Soc. Symp. Proc.*, Vol. 202, 1991, p. 167.
 - [42] A. T. Voutsas and M. K. Hatalis, "Structure of As-Deposited LPCVD Silicon Films at Low Deposition Temperatures and Pressures," *J. Electrochem. Soc.*, Vol. 139, No. 9, 1992, p. 2659.
 - [43] M. K. Hatalis and D. W. Greve, "Pressure Chemical Vapor Deposited Amorphous Silicon Films," *J. Appl. Phys.*, Vol. 63, 1988, p. 2260.
 - [44] M. K. Hatalis and D. W. Greve, "High-Performance Thin-Film Transistors in Low-Temperature

- Crystallized LPCVD Amorphous Silicon Films," *IEEE Electron Dev. Lett.*, Vol. EDL-8, 1987, p. 361.
- [45] A. G. R. French and P. J. Evans, "Effect of Deposition Temperature on LPCVD Polysilicon," *Electron. Lett.*, Vol. 22, 1986, p. 716.
 - [46] P. Krulevitch, G. C. Johnson, and R. T. Howe, "Stress and Microstructure in Phosphorus Doped Polycrystalline Silicon," *Mat. Res. Soc. Symp. Proc.*, Vol. 276, 1992.
 - [47] H. Guckel, D. W. Burns, C. C. G. Visser, H. A. C. Tilmans, and D. Deroo, "Fine Grained Polysilicon Films with Built in Tensile Strain," *IEEE Trans. Electron Dev.*, Vol. ED-35, No. 6, 1988, p. 800.
 - [48] S. Chang, W. Eaton, J. Fulmer, C. Gonzalez, B. Underwood, J. Wong, and R. L. Smith, "Micromechanical Structures in Amorphous Silicon," *Tech. Digest, Transducers '91*, 1991, p. 751.
 - [49] H. Guckel, "Surface Micromachined Pressure Transducers," *Sensors and Actuators*, Vol. 28, 1991, p. 133.
 - [50] F. A. Shemansky, "Mechanical and Electrical Properties of In-situ Phosphorus Doped Polysilicon," unpublished results.
 - [51] J. Huang, P. Krulevitch, G. C. Johnson, R. T. Howe, and H. R. Wenk, "Investigation of Texture and Stress in Undoped Polysilicon Films," *Mat. Res. Soc. Symp. Proc.*, Vol. 182, 1990, p. 201.
 - [52] G. Harbeke, L. Krausbauer, E. F. Steigmeier, A. E. Widmer, H. F. Kappert, and G. Neugebauer, "LPCVD Polycrystalline Silicon: Growth and Physical Properties of In-Situ Phosphorus Doped and Undoped Films," *RCA Review*, Vol. 44, 1983, p. 287.
 - [53] A. J. Learn and D. W. Foster, "Deposition and Electrical Properties of In-Situ Phosphorus Doped Silicon Films Formed by Low Pressure Chemical Vapor Deposition," *J. Appl. Phys.*, Vol. 61, No. 5, 1987, p. 1898.
 - [54] B. S. Meyerson and W. Olbricht, "Phosphorus Doped Polycrystalline Silicon via LPCVD: I. Process Characterization," *J. Electrochem. Soc.*, Vol. 131, No. 10, 1984, p. 2361.
 - [55] B. S. Meyerson and M. L. Yu, "Phosphorus Doped Polycrystalline Silicon via LPCVD: II. Surface Interactions of the Silane/Phosphine/Silicon System," *J. Electrochem. Soc.*, Vol. 131, No. 10, 1984, p. 2366.
 - [56] A. Yeckel and S. Middleman, "A Model of Growth Rate Nonuniformity in the Simultaneous Deposition and Doping of a Polycrystalline Silicon Film by LPCVD," *J. Electrochem. Soc.*, Vol. 134, No. 5, 1987, p. 1275.
 - [57] A. Yeckel, S. Middleman, and A. K. Hochberg, "The Origin of Nonuniform Growth of LPCVD Films from Silane Gas Mixtures," *J. Electrochem. Soc.*, Vol. 136, No. 7, 1989, p. 2038.
 - [58] H. Kurokawa, "P-Doped Polysilicon Film Growth Technology," *J. Electrochem. Soc.*, Vol. 129, No. 11, 1982, p. 2620.
 - [59] S. Nakayama, H. Yonezawa, and J. Murota, "Deposition of Phosphorus Doped Silicon Films by Thermal Decomposition of Disilane," *Jap. J. Appl. Physics*, Vol. 23, No. 7, 1984, p. L493.
 - [60] T. Whidden, E. Lindow, and S. Stiles, "In-Situ Doped Polysilicon Thin Films Utilizing Disilane in a Vertical Reactor," *Microelectronics Manufacturing Technology*, 1991, p. 35.
 - [61] S. P. Murarka and T. F. Retajczk, Jr., "Effect of Phosphorus Doping on Stress in Silicon and Polycrystalline Silicon," *J. Appl. Physics*, Vol. 54, No. 4, 1983, p. 2069.
 - [62] D. Flowers, Lj. Ristic, and H. G. Hughes, "Mechanical and Structural Characterization of In-Situ Doped Plasma Enhanced Alpha Silicon Films," *Tech. Digest, Transducers '91*, San Francisco, 1991, p. 961.
 - [63] Lj. Ristic, M. L. Kniffin, R. Gutteridge, and H. G. Hughes, "Properties of Polysilicon Films Annealed by RTA Process," *Thin Solid Films*, Vol. 220, 1992, p. 106.
 - [64] T. Lober, J. Huang, M. Schmidt, and S. Senturia, "Characterization of the Mechanisms Producing Bending Moments in Polysilicon Micro-Mechanical Beams by Interferometric Deflection Measurements," *Tech. Digest, IEEE Solid St. Sensor and Actuator Workshop*, Hilton Head Island, SC, 1988, p. 92.

- [65] F. A. Shemansky and M. L. Kniffin, "Mechanical and Electrical Properties of Ex-situ Phosphorus Doped Polysilicon Films," unpublished results.
- [66] J. C. Bravman and R. Sinclair, "Transmission Electron Microscopy Studies of the Polycrystalline Silicon-SiO₂ Interface," *Thin Solid Films*, Vol. 104, 1983, p. 153.
- [67] T. I. Kamins, "Structure and Properties of LPCVD Silicon Films," *J. Electrochem. Soc.*, Vol. 127, 1980.
- [68] T. I. Kamins, J. Manoliu, and R. N. Tucker, "Diffusion of Impurities in Polycrystalline Silicon," *J. Appl. Phys.*, Vol. 43, 1972, p. 83.
- [69] Y. Shioya and M. Maeda, "Comparison of Phosphosilicate Glass Films Deposited by Three Different Chemical Vapor Deposition Methods," *J. Electrochem. Soc.*, Vol. 133, No. 9, 1986, p. 1943.
- [70] A. Takamatsu, S. Miyoko, H. Sakai, and T. Yoshimi, "Plasma-Activated Deposition and Properties of Phosphosilicate Glass Film," *J. Electrochem. Soc.*, Vol. 131, No. 8, 1984, p. 1865.
- [71] D. J. Monk, D. S. Soane, and R. T. Howe, "A Diffusion/Chemical Reaction Model for HF Etching of LPCVD Phosphosilicate Glass Sacrificial Layers," *Tech. Digest*, IEEE, Solid State Sensor and Actuator Workshop, Hilton Head Island, SC, 1992, p. 46.
- [72] R. B. Bird, W. E. Stewart, and E. N. Lightfoot, in *Transport Phenomena*, New York, John Wiley & Sons, 1960.
- [73] R. A. Levy and K. Nassau, "Viscous Behavior of Phosphosilicate and Borophosphosilicate Glasses in VLSI Processing," *Solid State Technology*, 1986, p. 123.
- [74] P. R. Scheeper, J. A. Voorthuyzen, W. Olthius, and P. Bergveld, "Investigation of Attractive Forces Between PECVD Silicon Nitride Microstructures and an Oxidized Silicon Substrate," *Sensors and Actuators*, Vol. A30, 1992, p. 231.
- [75] R. L. Alley, G. J. Cuan, R. T. Howe, and K. Komvopoulos, "The Effect of Release-Etch Processing on Surface Microstructure Stiction," *Tech. Digest*, IEEE Solid St. Sensor and Actuator Workshop, Hilton Head Island, SC, 1992, p. 202.
- [76] M. Orpana and A. O. Korhonen, "Control of Residual Stress of Polysilicon Thin Films by Heavy Doping in Surface Micromachining," *Tech. Digest*, Transducers '91, 1991, p. 957.
- [77] H. Guckel and J. J. Sniewowski, "Formation of Microstructures with Removal of Liquids by Freezing and Sublimation," U.S. Patent No. 5,013,693.
- [78] G. K. Fedder and F. T. How, "Thermal Assembly of Polysilicon Microstructures," Proceedings IEEE Microelectro-Mechanical Systems Workshop, Salt Lake City, 1989.
- [79] Y. C. Tai and R. S. Muller, "Fracture Strain of LPCVD Polysilicon," *Tech. Digest*, IEEE Solid-State Sensor and Actuator Workshop, Hilton Head Island, SC, 1988, p. 88.
- [80] M. G. Allen, M. Mehregany, R. T. Howe, and S. D. Senturia, "Microfabricated Structures for the In-Situ Measurement of Residual Stress, Young's Modulus, and Ultimate Strain of Thin Films," *Appl. Phys. Lett.*, Vol. 51, 1987, p. 241.
- [81] L. S. Fan, Y. C. Tai, and R. S. Muller, "Integrated Movable Micromechanical Structures for Sensors and Actuators," *IEEE Trans. Electron Dev.*, Vol. ED-35, 1988, p. 724.
- [82] M. Mehregany, R. T. Howe, and S. D. Senturia, "Novel Microstructures for the In-situ Measurement of Mechanical Properties of Thin Films," *J. Appl. Phys.*, Vol. 62, 1987, p. 3579.
- [83] Stephen D. Senturia, "Microfabricated Structures for the Measurement of Mechanical Properties and Adhesion of Thin Films," *Tech. Digest*, Transducers '87, Tokyo, 1987, p. 11.
- [84] H. Guckel, T. Randazzo, and D. W. Burns, "A Simple Technique for the Determination of Mechanical Strain in Thin Films with Applications to Polysilicon," *J. Appl. Phys.*, Vol. 57, 1985, p. 1671.
- [85] R. Roark and W. Young, in *Formulas for Stress and Strain*, New York, McGraw-Hill, 1989.
- [86] S. H. Crandall, N. C. Dahl, and T. J. Lardner, in *An Introduction to the Mechanics of Solids*, New York, McGraw-Hill, 1978.
- [87] A. Blake, in *Practical Stress Analysis in Engineering Design*, New York, Marcel Dekker, 1990.

- [88] D. W. Burns and H. Guckel, "Thin Films for Micromechanical Sensors," *Journal of Vac. Sci. Tech.*, Vol. A8, 1990, p. 3606.
- [89] R. Mahadevan, M. Mehregany, and K. J. Gabriel, "Application of Electric Microactuators to Silicon Micromechanics," *Sensors and Actuators*, Vols. A21-A23, 1990, p. 219.
- [90] L. S. Fan, R. S. Muller, W. Yun, R. T. Howe, and J. Huang, "Spiral Microstructures for the Measurement of Average Strain Gradients in Thin Films," *Proc. IEEE Conf. on Micro Electro Mechanical Systems*, 1990, p. 177.
- [91] T. P. Weihs, S. Hong, J. C. Bravman, and W. D. Nix, "Mechanical Deflection of Cantilever Microbeams: A New Technique for Testing the Mechanical Properties of Thin Films," *J. Mater. Res.*, Vol. 3, 1988, p. 931.
- [92] W. D. Nix, "Mechanical Properties of Thin Films," *Metallurgical Trans.*, Vol. 20A, 1989, p. 2217.
- [93] K. E. Petersen, "Dynamic Micromechanics on Silicon: Techniques and Devices," *IEEE Trans. Electron. Dev.*, Vol. ED-25, 1978, p. 1241.
- [94] K. E. Petersen and C. R. Guarnieri, "Young's Modulus Measurement of Thin Films Using Micromechanics," *J. Appl. Phys.*, Vol. 50, 1979, p. 6761.
- [95] L. Kiesewetter, J. M. Zhang, D. Houdeau, and A. Steckenborn, "Determination of Young's Moduli of Micromechanical Thin Films Using the Resonance Method," *Sensors and Actuators*, Vol. A35, 1992, p. 153.
- [96] J. D. Zook, D. W. Burns, H. Guckel, J. J. Sniegowski, R. L. Engelstad, and Z. Feng, "Characteristics of Polysilicon Resonant Microbeams," *Sensors and Actuators*, Vol. A35, 1992, p. 51.
- [97] J. Y. Pan, P. Lin, F. Maseeh, and S. D. Senturia, "Verification of FEM Analysis of Load-Deflection Methods for Measuring Mechanical Properties of Thin Films," *Tech. Digest, IEEE Solid State Sensor and Actuator Workshop*, Hilton Head Island, SC, 1990, p. 70.
- [98] F. Maseeh, M. A. Schmidt, M. G. Allen, and S. D. Senturia, "Calibrated Measurements of Elastic Limit, Modulus and the Residual Stress of Thin Films Using Micromachined Suspended Structures," *Tech. Digest, IEEE Solid State Sensor and Actuator Workshop*, Hilton Head Island, SC, 1988, p. 84.
- [99] K. Najafi and K. Suzuki, "Measurement of Fracture Stress, Young's Modulus, and Intrinsic Stress of Heavily Boron-Doped Silicon Microstructures," *Thin Solid Films*, Vol. 181, 1989, p. 251.
- [100] L. S. Fan, R. T. Howe, and R. S. Muller, "Fracture Toughness Characterization of Brittle Thin Films," *Sensors and Actuators*, Vol. A21-A23, 1990, p. 872.
- [101] Lj. Ristic, R. Gutteridge, B. Dunn, D. Mietus, and P. Bennett, "Surface Micromachined Polysilicon Accelerometer," *Tech. Digest, IEEE Solid-State Sensor and Actuator Workshop*, Hilton Head Island, SC, 1992, p. 118.
- [102] Lj. Ristic, R. Gutteridge, J. Kung, D. Koury, B. Dunn, and H. Zunino, "A Capacitive Type Accelerometer with Self-Test Features Based on a Double-Pinned Polysilicon Structure," *Tech. Digest, Transducers '93, Yokohama*, 1993, p. 810.
- [103] L. S. Fan, Y. C. Tai, and R. S. Muller, "IC-Processed Electrostatic Micro-Motors," *Tech. Digest, IEDM, San Francisco*, 1988, p. 666.
- [104] M. Mehregany, S. F. Bart, L. S. Tavrow, J. H. Lang, S. D. Senturia, and M. F. Schlecht, "A Study of Three Microfabricated Variable-Capacitance Motors," *Sensors and Actuators*, Vols. A21-A23, 1990, p. 173.
- [105] W. C. Tang, T.-C. H. Nguyen, and R. T. Howe, "Laterally Driven Polysilicon Resonant Microstructures," *Sensors and Actuators*, Vol. A20, 1989, p. 25.
- [106] T. Hirano, T. Furuhashi, K. J. Gabriel, and H. Fujita, "Operation of Sub-micron Gap Electrostatic Comb-Drive Actuators," *Tech. Digest, Transducers '91, San Francisco*, 1991, p. 873.
- [107] P. I. Pratt, G. C. Johnson, R. T. Howe, and J. C. Chang, "Micromechanical Structures for Thin Film Characterization," *Tech. Digest, Transducers '91, San Francisco*, 1991, p. 205.
- [108] Lj. Ristic, R. Gutteridge, and R. Roop, "Integration of Surface Micromachined Dynamic Structures," *Proc. IV International Forum on ASIC and Transducers, Leuven*, 1991, p. 129.

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- [109] E. W. Becker, W. Ehrfeld, P. Hagmann, A. Maner, and D. Munchmeyer, "Fabrication of Microstructures with High Aspect Ratios and Great Structural Heights by Synchrotron Radiation Lithography, Galvanoforming, and Plastic Molding (LIGA Process)," *Microelectronic Engineering*, Vol. 4, 1986, p. 35.
 - [110] K. Suzuki, "Single Crystal Silicon Micro-Actuators," *Tech. Digest*, IEDM, 1990, p. 26.
 - [111] H. Guckel, K. J. Skrobis, T. R. Christensen, J. Klein, S. Han, B. Choi, E. G. Lovel, and T. W. Chapman, "On the Application of Deep X-Ray Lithography with Sacrificial Layers to Sensor and Actuator Construction (The Magnetic Micromotor with Power Take Offs)," *Tech. Digest*, Transducers '91, 1991, p. 393.
 - [112] H. Guckel, J. Klein, T. Christenson, K. Skrobis, M. Laudon, and E. G. Lovell, "Thermomagnetic Metal Flexure Actuators," *Tech. Digest*, Solid-St. Sensor and Actuator Workshop, Hilton Head Island, SC, 1992, p. 73.

Chapter 11

Sensors for the Automotive Industry

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Motorola

11.1 INTRODUCTION

The automotive industry has been specifically chosen to describe sensor applications for two reasons: it represents an extremely harsh environment for electronic components, and it provides high-volume applications that in turn drive the development of new sensing technologies. New systems being developed for automobiles are the result of government legislation (emissions and safety regulations) and the need to have a competitive advantage (antilock brake systems, air bags, and entertainment systems). Many of these applications require a wide variety of sensing inputs and use sensor technologies that are also found in medical, industrial, and instrumentation applications.

A few of the factors influencing the increased use of electronic systems in automobiles are improved performance, product differentiation, system diagnostics for improved serviceability, systems integration for improved reliability, reduced emissions, improved safety and convenience, communications, and improved manufacturability. A simplified model of an automotive control system is shown in Figure 11.1. Input, either from vehicle sensors (or switches) or driver actuated switches, are converted to a form that can be used by the microprocessor. In certain cases the sensor is the feedback element for the closed loop control system. The interface requirements are minimal or nonexistent for simple switched inputs but can require amplification, calibration, buffering, temperature compensation, or analog-to-digital (A/D) conversion for many sensing inputs. The complexity and number of components can be considerably reduced if custom integrated circuits and integrated sensors are developed to provide the interface from the sensing element to the microprocessor.

Parameters that can and need to be sensed in the car vary from system to system. Certain common measurements such as wheel speed are required for several

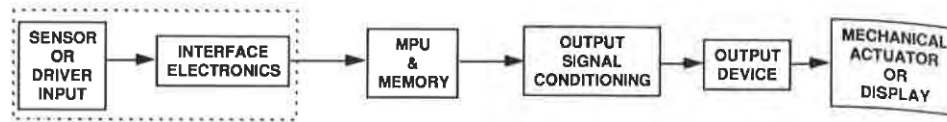


Figure 11.1 MCU-based automotive control system.

systems. Figure 11.2 [1] shows automotive sensor volumes in 1990 for three of the major measurement requirements. The numbers exclude commodity sensors such as temperature and standard vehicle speed sensors. (Temperature sensors account for as much as one-fifth the total volume in 1990.) The total number of motion-position, pressure, and acceleration sensors is projected to double by 1996. This growth is caused by the extension of control systems that are presently only on limited production vehicles to higher volume car lines and in some cases almost the entire fleet. In addition, new systems that are still in the development phase will achieve a level of high-volume production. Acceleration sensors are projected to experience the highest growth.

The automotive environment is recognized to be one of the more difficult applications for electronic systems and microelectronic sensors. The environment includes a wide temperature range, high-humidity conditions, and the need to withstand several chemicals, operate under high electromagnetic interference (EMI) conditions and yet cause relatively low levels of radio frequency interference (RFI), despite extremely wide voltage swings. At the same time, the acceptance of electronics is extremely customer driven and demands low cost and high reliability. Projected sales volumes can decrease (or increase) quickly, depending on customer acceptance of the system, the vehicle in which the system is offered, and external economic factors that affect purchasing decisions for high-cost items such as automobiles.

A major differentiating factor for automotive usage is whether the location of sensors is in the passenger compartment or under the hood. Table 11.1 (after SAE J1211) shows the variation that can occur to several variables depending on the location of the electronic component. High temperature and humidity can significantly reduce the useful life, especially in underhood applications of electronics. The SAE (formerly, Society of Automotive Engineers) has developed "Recommended Environmental Practices for Electronic Equipment Design—SAE J1211" to address the unique problems of automotive electronics.

The system voltage in passenger vehicles is nominally 12V. However, several normal and abnormal voltage variations occur, which must be taken into account at either the component or system level. Normal operating charging systems regulate the output of the alternator to provide sufficient voltage to keep the battery charged under various temperature and load conditions. This voltage can range from 16 V

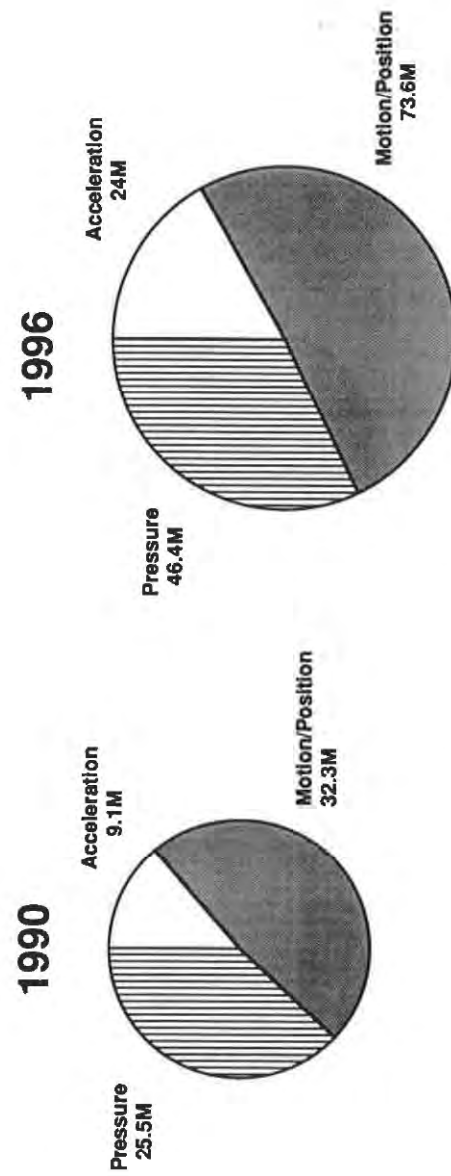


Figure 11.2 Sensor volumes in the North American automotive market. The total sensor market is expected to more than double by 1996.

Table 11.1
The Automotive Environment

<i>Variable</i>	<i>Underhood</i>	<i>Passenger Compartment</i>
Storage temperature	−40°C to +150°C	−40°C to +85°C
Operating temperature	−40°C to +125°C	−40°C to +85°C
Vibration	15g, 10 to 200 Hz	2g, 20 Hz
Humidity	Up to 100%	Up to 100%
Chemicals	Salt spray, water, fuel, oil, coolant and solvent immersion	—
Thermal cycling	>1,000 cycles from −40°C to +125°C	>1,000 cycles from −40°C to +85°C
EMI protection	Up to 200 V/m	Up to 50 V/m

when very cold (−40°C) to about 12V when maximum underhood temperatures are reached. Electronic assemblies on the car must be able to withstand reverse battery conditions (−12V), jump starts from tow trucks (24V), short-duration voltage transients that can easily exceed $\pm 100\text{V}$, and long-duration (>400 ms) alternator load dump transients, which can be as high as 120V. To prevent damage due to excessive voltage, microelectronic sensors with sufficiently high breakdown voltage (to withstand the variety of automotive voltages) must be used, or the systems they are used in must provide a regulated and protected power bus.

The increasing electronic complexity of modern vehicles has increased the possibility of EMI causing problems between various vehicle systems. In addition, electronics can be susceptible to and radiate RFI causing poor radio reception and malfunctioning vehicle systems with a resulting no-fault condition found during service diagnostic procedures. The potential problems in this area have been addressed by SAE with several information reports and a recommended practice as shown in [2].

Recommended Environmental Practices for Electronic Equipment Design, SAE J1211

Performance Levels and Methods of Measurement of Electromagnetic Radiation from Vehicles and Devices, SAE J551

Performance Levels and Methods of Measurement of EMR from Vehicles and Devices (Narrowband RF), SAE J1816

Electromagnetic Susceptibility Procedures for Vehicle Components (Except Aircraft), SAE J1113

Vehicle Electromagnetic Radiated Susceptibility Testing Using a Large TEM Cell, SAE J1407

Open Field Whole-Vehicle Radiated Susceptibility 10 kHz–18 GHz, Electric Field, SAE J1338

Class B Data Communication Network Interface, SAE J1850

Diagnostic Acronyms, Terms and Definitions for Electrical/Electronic Components, SAE J1930

Failure Mode Severity Classification, SAE J1812

Guide to Manifold Absolute Pressure Transducer Representative Test Method, SAE J1346

Guide to Manifold Absolute Pressure Transducer Representative Specification, SAE J1347

Some possible applications for sensors in existing and future automotive systems are indicated in Table 11.2. This shows a potential of over 100 sensors for 18 different automotive systems. Many of the same parameters are required for different systems, and this will change the way that sensors are used in future vehicle systems. In some cases, such as manifold absolute pressure and mass air flow in engine control systems, one measurement will be preferred over another possible measurement. The list does not include sensors that could be used in measuring current in brushless dc motors or sensors required in entertainment systems and systems with little or no sensor usage, such as electronic mufflers and heated seats.

At least 21 sensing technologies can and are being used to provide these sensors.

- | | |
|---|-----------------------------------|
| • Acoustical | • Potentiometer |
| • Capacitive | • Reed switch |
| • Doppler radar | • Silicon piezoresistive |
| • Electrochemical | • Silicon capacitive |
| • Hall effect | • Surface acoustic wave (SAW) |
| • Inclinator | • Temperature bimetallic switch |
| • Linear variable differential transformer (LVDT) | • Temperature—transistor V_{BE} |
| • Magnetostrictive element (MRE) | • Thermistor |
| • Optoelectronic | • Ultrasonic |
| • Piezoelectric | • Variable reluctance |
| | • Weigand effect |

Many of the sensors are digital (actually a switch) and not analog sensors. Certain technologies have also had much greater acceptance and more widespread use in automotive applications, such as thermistors, Hall-effect sensors, piezoresistive pressure sensors, and variable reluctance sensors.

The requirements for automotive sensors can be divided into three areas: the transducer element; interface electronics; and packaging. The basic transducer element is expected to meet the accuracy requirements of the system and be able to

Table 11.2
Sensor Requirements versus Vehicle Systems

<i>System</i>	<i>Parameter</i>	<i>System</i>	<i>Parameter</i>
Engine control	Manifold absolute pressure	NODS (near obstacle detection) collision avoidance	Relative speed vehicle #2
	Barometric pressure (altitude)		Distance vehicle #2
	Exhaust gas oxygen		(See nods)
	NO _x		(See navigation)
	Fuel composition	Intelligent highways	Traffic flow
	Mass air flow		Humidity in passenger compartment (PC)
	Engine knock		Temperature (PC)
	EGR valve position		Chemical composition of air (PC)
	EGR pressure	HVAC (climate control)	Air flow (PC)
	Throttle position		Outside air temperature
	Combustion pressure		Chemical composition of outside air
	Exhaust pressure		Motor phase angle sensors
	Fuel composition/flex fuel	Cruise control	Vehicle speed
	Fuel pressure		Brake switch
	Vehicle speed		Selector switch
	Air charge temperature		Multiple usage of sensors
	Engine speed	MUX/diagnostics	AC clutch sensor/switch
	Crank angle/position		Power steering pressure
	Engine coolant temperature		Shift lever (PRNDL) switch
	Humidity		Engine speed
	Torque	Electronic transmission (continuously variable transmission)	Throttle position
	Oil quality		Shift lever position (PRNDL)
	Fluid condition		Trans oil pressure
	Engine noise		Shaft speed
Anti-skid brakes	Ambient temperature	Driver information	Transmission oil temperature
	Wheel speed		Vacuum modulation
	Brake switch		Vehicle speed
	Brake pressure		Engine speed
	Deceleration		Oil pressure
Traction control/abs	Fluid level		Fuel level
	Fish tailing		Oil level
	Engine speed		Oil quality/condition
	Steering wheel angle		Coolant pressure
Air bags	Steering wheel rate of change		Coolant condition
	Deceleration		Coolant temperature
	Bag pressure		Ambient air temp
Suspension	Vehicle speed		Coolant level
	Vehicle height (wh-body displ)		

Table 11.2
(Continued)

<i>System</i>	<i>Parameter</i>	<i>System</i>	<i>Parameter</i>
Electronic power steering (also electronic assisted)	Steering wheel angle	Memory seat	Windshield washer level
	Steering wheel direction		Transmission oil level
	Acceleration		Tire pressure
	Vehicle weight		Tire surface temperature
	Road surface conditions		Battery fluid level
	Side slip angle		Rain sensor
	Speed of yaw angle		Sun sensor
	Lateral displacement		Driver selector switch
	Vehicle speed		Lumbar pressure
	Pneumatic spring pressure		Seat position
4-wheel steering	Ride selector switch	Navigation	Gyroscope
	Steering wheel torque		Steering wheel turns
	Hydraulic pressure		Geomagnetic (magnetic field) sensors
	Vehicle speed		Distance traveled
Security and keyless entry	Steering wheel position		
	Vehicle speed		
	Steering wheel angle		
	Front-rear wheel angle		
	Illegal access indication		
	Remote access indication		

tolerate the normal overrange occurrences that could be anticipated in certain failure modes of other components. The output should be stable over the operating temperature range and the lifetime of the vehicle. Electronics is frequently required to enhance the output signal, calibrate the zero and sensitivity, compensate for the effects of temperature and other variables, tolerate electromagnetic interference, provide protection against short circuits, reverse battery and overvoltage conditions, and is increasingly being asked to provide diagnostic and self-test features. The ease of interface for digital processing also requires the output of an analog sensor to be converted through an (integral) analog to a digital interface or a serial communication interface. The sensor package required for the automotive environment must be small, easy to handle and mount in the application, extremely robust and capable of withstanding thermal cycling, thermal shock, vibration, various chemicals, and high levels of humidity, and provide the EMI protection for the sensor and interface electronics.

Microelectronic sensors are extremely well suited to satisfy the requirements of increased reliability, easier electronic interface, and self-diagnostics and to reduce the system's cost. Therefore inherent limitations in these devices must be resolved by appropriate sensor design and signal conditioning. Below are a number of undesirable characteristics of semiconductor sensors that have to be addressed [3].

- Nonlinearity
- Drift
- Offset
- Temperature dependence of offset
- Temperature dependence of sensitivity
- Nonrepeatability
- Cross-sensitivity to temperature and strain
- Hysteresis
- Low resolution
- Low sensitivity
- Unsuitable output impedance
- Self-heating
- Unsuitable frequency response

11.2 SENSING TECHNOLOGY IN VEHICLE SYSTEMS

New technologies can provide a sensing solution for an automotive application only if they can meet performance criteria, form factor (packaging) requirements, program development timetables, and pass the required reliability qualification tests. The decision on the most appropriate technology can then be reduced to a purchasing decision—the lowest cost approach. Some examples regarding current sensing techniques will be discussed in this section.

Electronic engine controls were the first area of on-board computer usage in automobiles. Early systems provided controlled delivery of fuel, based on engine load, engine speed, and engine temperature. Federal and local (for example, California, Denver) standards for emission levels and corporate average fuel economy (CAFE) have necessitated that the precise control provided by electronic systems be incorporated in all passenger vehicles sold in the United States. The most sophisticated systems incorporate several input sensors such as an oxygen (O_2) sensor, coolant temperature sensor, manifold absolute pressure (MAP) sensor, vehicle speed sensor, throttle position sensor, engine speed reference or distributor reference (rpm), mass air flow sensor, air intake temperature sensor, crankshaft sensor, detonation (knock) sensor, throttle body temperature sensor, throttle idle switch, altitude (or barometric) pressure sensor, and a wide open throttle switch. The requirements for these sensors and the number of sensor inputs required for the control system are being increased to meet legislated demands for tighter emission controls and higher fuel economy.

The electronic control unit processes the input signals to provide proper fuel, spark, and engine speed-load control and actuates several output drivers such as the ignition coil, injector drivers, and emission-related solenoids and valves. Figure 11.3 [4] shows a typical engine control system with all of the inputs that are required for

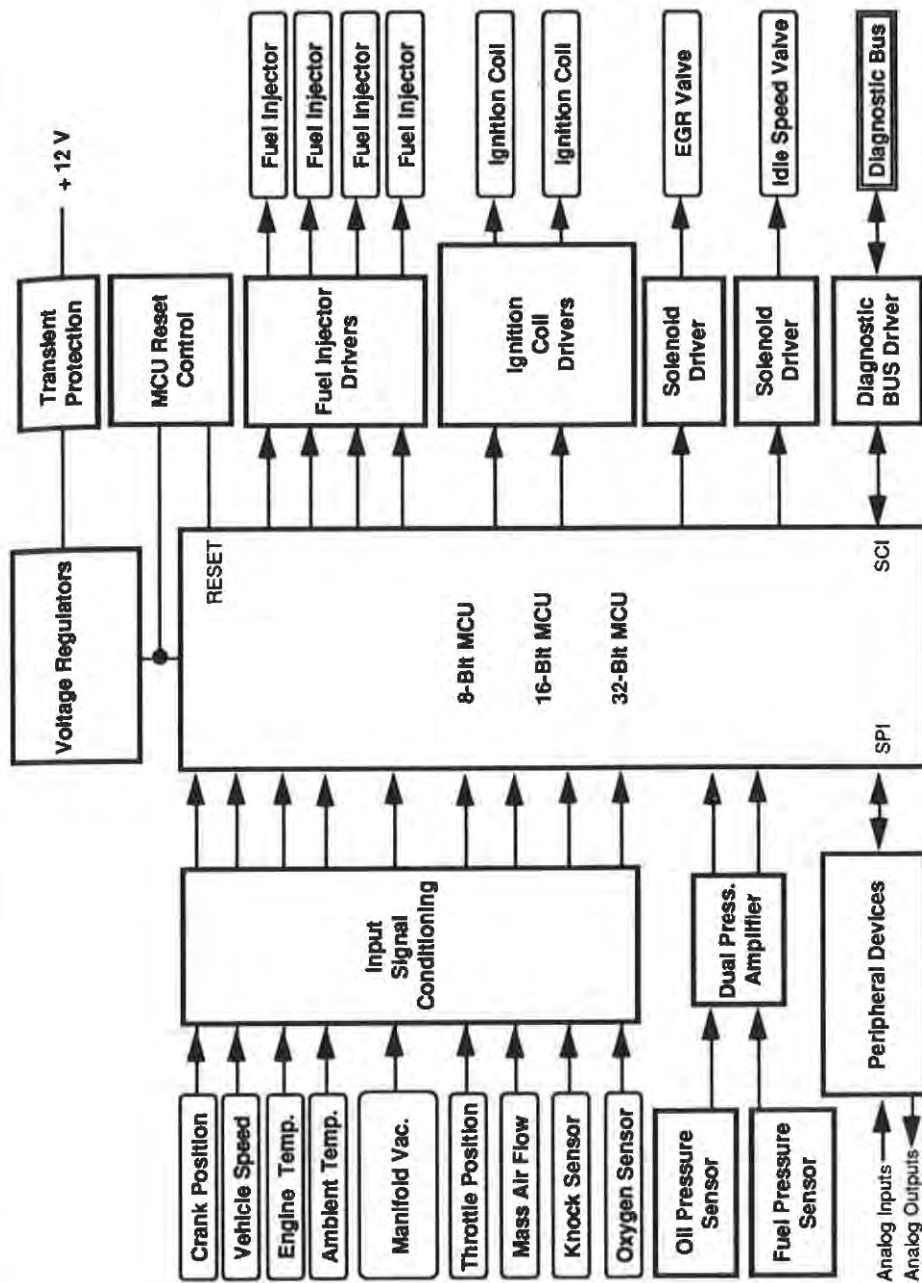


Figure 11.3 Engine control system.

meeting current emissions standards and providing improved fuel economy and drivability. All of the sensors in this system, except the oxygen sensor, are or possibly can be microelectronic sensors. The microelectronic sensors can also integrate the input signal conditioning functions.

11.2.1 Manifold Absolute Pressure

The first and currently largest volume automotive application for micromachined sensors is for sensing manifold absolute pressure and barometric absolute pressure (BAP) required in electronic engine control systems. The measurement of absolute pressure has been accomplished with LVDT (linear variable differential transformer) sensors, aneroid chambers, capacitive ceramic pressure sensors, and piezoresistive and capacitive silicon sensors. The last two will be discussed here as representatives of the new technology.

11.2.1.1 Piezoresistive Pressure Sensor

The most frequently used approach to semiconductor pressure sensors utilizes the piezoresistive effect in silicon. Sensing elements are either a four-element Wheatstone bridge or a single element positioned to maximize the sensitivity to shear stress.

The latest development in semiconductor pressure sensors is shown in Figure 11.4, where the circuitry necessary to adjust the offset and span and to compensate for temperature effects on both offset and span, as well as to provide an amplified signal, has been integrated on the same chip with the micromachined diaphragm. The sensor signal is provided from a single piezoresistive element located at a 45° angle in the center of the edge of the square diaphragm, and this allows for an extremely linear measurement. The offset voltage and full-scale span of the basic sensing element vary with temperature but in a highly predictable manner. This approach requires a minimum number of external components for amplification to provide a usable output signal.

Two silicon wafers are used to produce the piezoresistive silicon pressure sensor (Figure 11.5). The top wafer is etched until a thin, square diaphragm, approximately 25 μm in thickness, is achieved. The square area is extremely reproducible, as is the 54.7° angle of the cavity wall. The top wafer is attached to a support wafer by a glass frit to provide a structure that is isolated from mounting stresses. An absolute reference is obtained by bonding the top wafer to the solid bottom wafer in a vacuum. In addition to the basic sensing element, an interactively laser-trimmed four-stage network is also produced in a single monolithic structure.

The photomicrograph of the fully signal conditioned pressure sensor is shown in Figure 11.6. The additional circuitry is accomplished using the silicon area that is required to provide the mechanical support for the diaphragm. The die size for

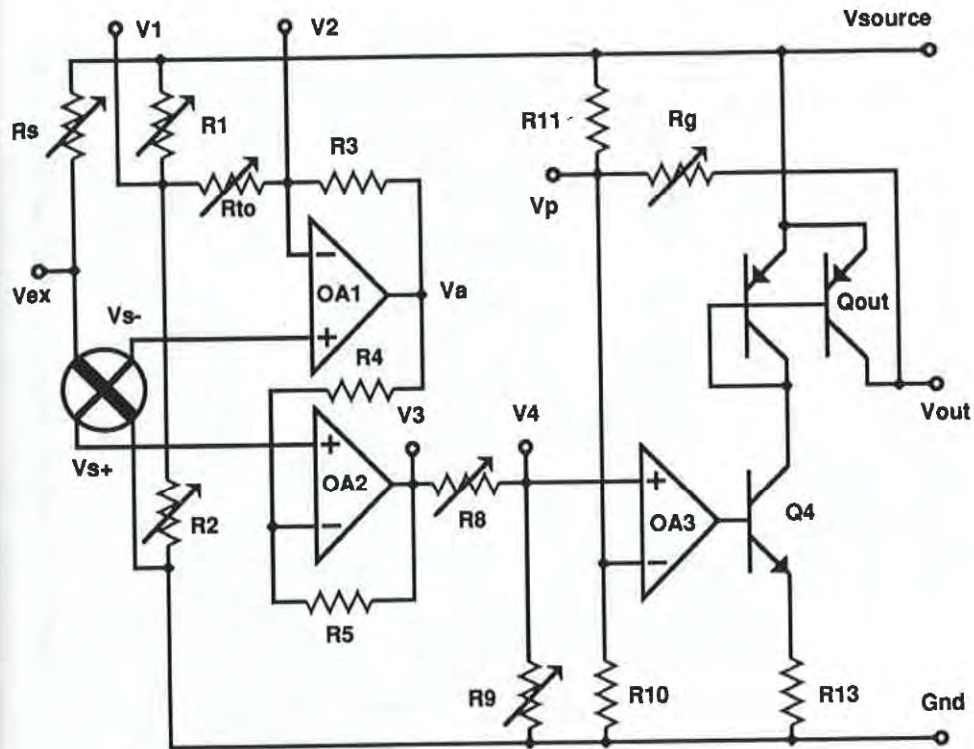


Figure 11.4 Integrated MAP sensor circuit.

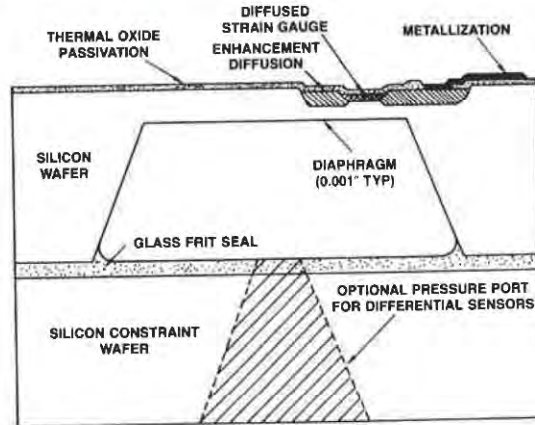


Figure 11.5 Cross section of piezoresistive pressure sensor.

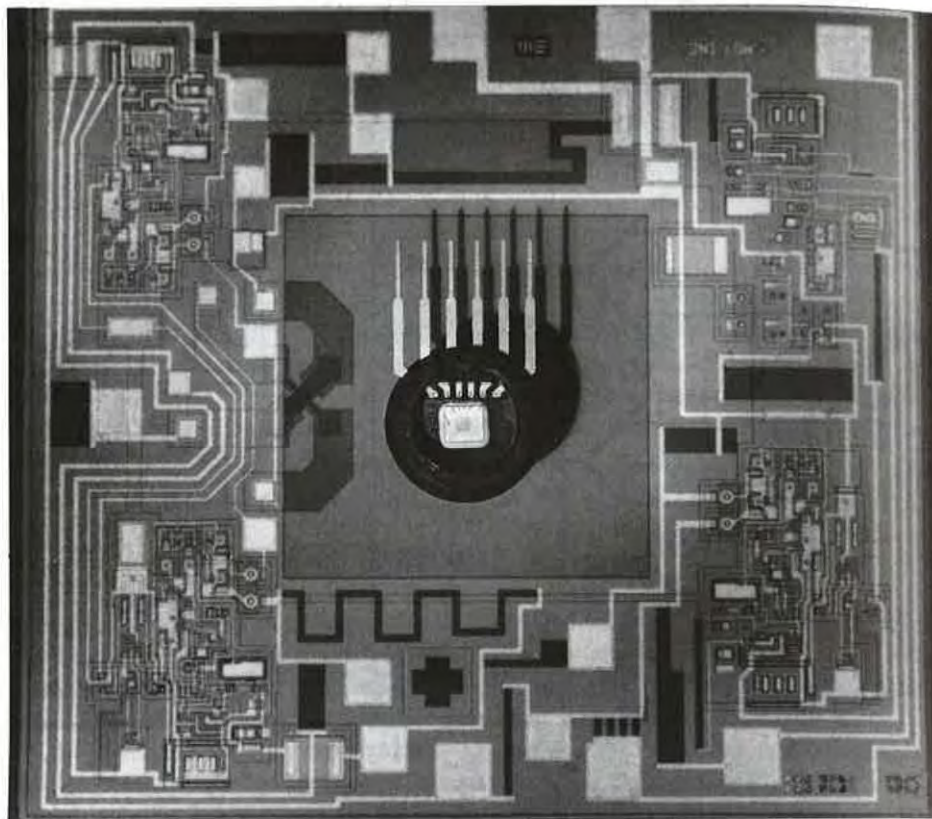


Figure 11.6 Integrated MAP sensor.

the uncompensated sensing element is $3048\ \mu\text{m} \times 3048\ \mu\text{m}$, and the die size for the fully signal conditioned unit is $3685\ \mu\text{m} \times 3302\ \mu\text{m}$.

The cost is minimized when the laser trimming is performed at the earliest stage of assembly. Automatic pattern recognition allows for high throughput in assembly. The assembly techniques, equipment, and statistical process controls are developed for high-volume production.

The laser trimming process can be used to provide a wide variety of transfer functions. Once the trim procedure is completed, it yields a device with a zero-pressure offset voltage of 0.5V nominal and a full-scale output voltage of 4.5V when connected to a 5.0-V supply. Therefore, the output dynamic range due to an input pressure swing of 0–15 psi is 4.0V. The output is ratiometric with the supplied voltage, and a maximum supply voltage of 10V is allowed.

The performance of an automotive version of the sensor for measuring manifold absolute pressure is shown in Figure 11.7. The performance at five temperatures is

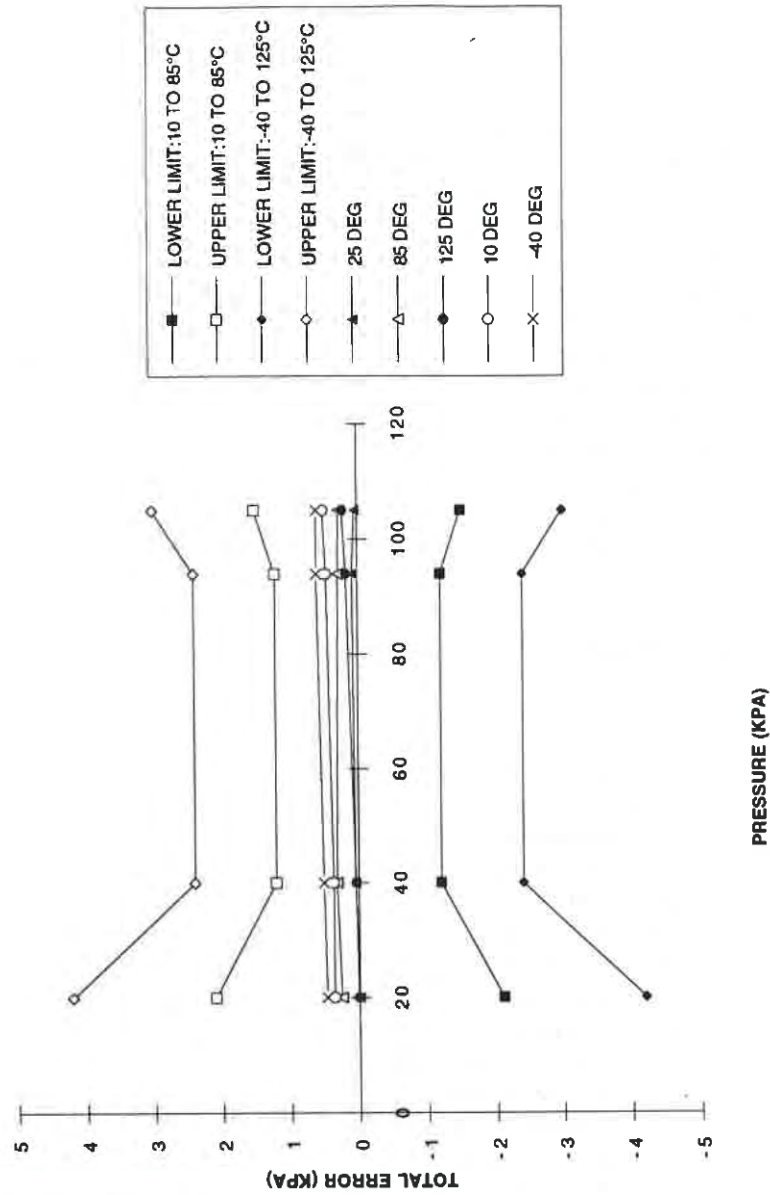


Figure 11.7 Performance of integrated MAP sensor versus temperature.

plotted versus the error band commonly specified for automotive applications from -40°C to 125°C . The performance at -40°C and 125°C is even within the limits required for $+10^{\circ}\text{C}$ to 85°C operation.

11.2.1.2 Capacitive Absolute Pressure Sensor

A silicon capacitive absolute pressure (SCAP) sensor has also been designed to measure barometric or manifold absolute pressure for an automotive engine control system [5]. The plates of the capacitor are formed by a silicon diaphragm and a metallized, glass substrate. The diaphragm in the SCAP sensor is a $5\text{ }\mu\text{m}$ cavity etched into a $127\text{-}\mu\text{m}$ thick silicon wafer. Holes are laser drilled through the glass substrate to obtain the electrical connection and allow the reference vacuum to be sealed inside the cavity. The glass substrate is anodically bonded to the silicon top wafer to achieve a hermetically sealed structure. Solder bumps seal in the reference vacuum and provide the external connections. Additional external circuitry is used to produce a variable frequency output that can have a direct interface to a microcontrol unit (MCU). The sensing element is 45 mm^2 and achieves a full-scale capacitance of approximately 39 pF . The capacitance change over the operating pressure range of 17 to 105 kPa is 7 pF , resulting in an output frequency change from 20 to 180 Hz . The variation of the capacitance with temperature is -30 to $800\text{ ppm}/^{\circ}\text{C}$ and the linearity is typically 1.4% .

11.2.2 Tire Pressure Sensor

One of the traditional measurements necessary for safe vehicle operation is tire pressure. Electronics has recently made it possible to measure tire pressure at each wheel and transmit the data to the passenger compartment for continuous driver awareness of the status of each tire. A system such as the one in Figure 11.8 [6] utilizes semiconductor pressure and temperature sensors mounted inside the tire.

Power for the system is supplied by a combination of rotating and fixed antennas. A high-frequency alternating current applied to the fixed antenna allows an induced voltage on the rotating antenna to provide sufficient power for the wheel's electronic components. Transmission is interrupted every 150 ms to allow the sensors to measure pressure and temperature and transmit digitally coded information back to the fixed antenna. A control unit decodes the sensor signals 6 times/second and compares this measurement to a reference pressure stored in memory to determine if a critical situation exists. The system can distinguish between an increase in tire temperature caused by increased flexing of an underinflated tire and the heat that results from braking or the tire's response to high outside ambient temperatures.

The key sensor characteristic for a piezoresistive pressure sensor such as the one described in Section 11.2.1.1 is an extremely low-current draw (high-impedance device). Also, tire pressure measurements are typically gauge measurements, but

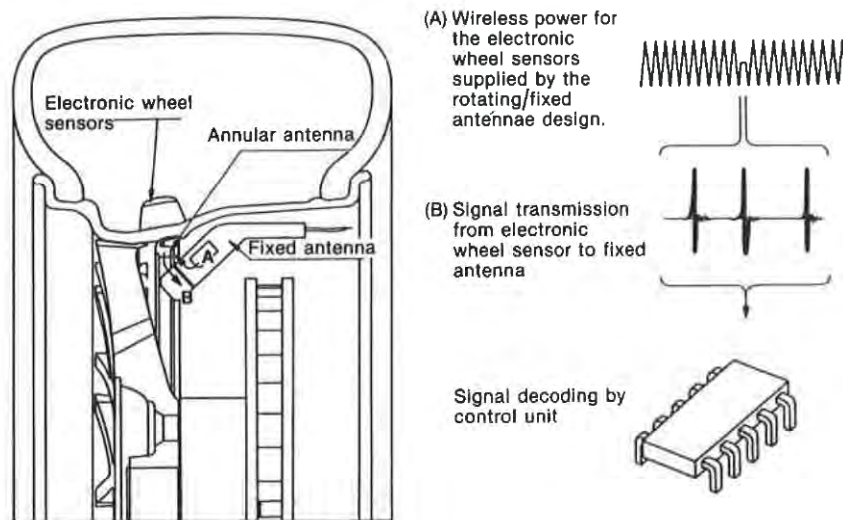


Figure 11.8 Tire pressure sensing.

since the electronic tire pressure is measured inside the tire, an absolute sensor is used.

Other applications for pressure sensing in the automobile that involve both higher and lower pressure ranges and gauge or absolute pressure measurements include the AC compressor, pneumatic lumbar support, air bag pressure, ABS (antilock braking system) brake pressure, engine oil and transmission pressure, fuel injection pressure, coolant system pressure, EGR (exhaust gas recirculation) pressure, pneumatic spring pressure, and combustion chamber pressure. Differential pressure can also be used to measure flow, which will be covered in Section 11.2.4. Pressure sensors can further be used to measure fluid level, as discussed in Section 11.3.2.

11.2.3 Position, Rotation, and Speed Sensing

As indicated in Figure 11.2, motion and position sensing is the highest volume application. The most popular speed sensor in automotive applications is the variable reluctance transducer, which consists of a magnetic pole with a coil of wire wrapped around it. An output voltage proportional to the rate of change of flux in the magnetic circuit produces a pulse when a tooth on a trigger wheel passes in front of the pole. The output voltage can vary from 500 mV to over 80V and requires additional electronic circuitry to eliminate unwanted noise, clamp the output voltage to a safe level, and provide a useful signal to the MCU. Several semiconductor approaches are also being used to detect wheel speed, crankshaft speed, angular rotation, and direction. A comparison of the various techniques is shown in Table 11.3 [7]. In all cases the

Table 11.3
A Comparison of Various Sensors and Their Features

<i>Sensor Type</i>	<i>Zero Speed</i>	<i>Maximum Frequency</i>	<i>Magnets</i>	<i>Nonferrous Wheel</i>	<i>Square Wave Output</i>	<i>S/N</i>	<i>Immunity to EMC</i>
Eddy current	Yes	500 kHz	No	Yes	Yes	High	Poor
Hall effect	Yes/No*	1 MHz	Yes	No	No	Moderate	Poor
Variable reluctance	No	10–100 MHz	Yes	No	No	Low	Moderate
Reed switch	Yes	600 Hz	Yes	No	Yes	High	High
Wiegand effect	Yes	20 kHz	Yes	No	No	High	—
Opto	Yes	10 MHz	No	Any	Yes	Very high	High

*Sensors with capacitor in feedback circuit cannot achieve zero speed.

output of the sensor is a pulse that is counted, or the time between pulses is calculated, to provide the desired speed or angular position. Some of these same techniques are also used to detect linear position. For linear position such as throttle and EGR position, potentiometers are commonly used.

11.2.3.1 Opto Sensing

Properly designed optical sensors have the smallest jitter angle, or variation from actual angular position, for a given diameter sense wheel when compared to other crankshaft position sensors. In addition opto sensors can sense static position and have a constant output with speed. They must be protected from oil and dirt to function properly. In addition, optical sensors have an upper temperature limit that must not be exceeded if the expected life is to be obtained.

Opto sensing techniques using infrared emitters and detectors have been successfully used for several years in rather benign environments to accurately sense rotational and linear motion. Their extension into the automobile has been delayed due to temperature sensitivity and stability problems associated with emitters. Recent developments have resulted in significant improvement, which has allowed their usage in higher (up to 150°C) temperature applications.

Detailed studies of LEDs revealed that under accelerated conditions it was not unusual to experience 15–20% degradation after 1,000 hours of testing. Processing revisions and improvements in process control implemented (by Motorola) to increase LED performance have resulted in extremely stable, long-term reliability. The current transfer ratio (CTR), the ratio of transistor collector current to forward current, was used to measure the LED light output. As shown in Figure 11.9 [8], after 7,500 hours, the average CTR was still almost 98% of the initial value at room temperature.

Improved performance and packaging concepts that take into account the mounting location and surrounding environment have resulted in several custom opto sensors being developed for automotive applications. For example, two pairs of opto emitters and detectors with integrated Schmitt triggers are used to provide steering-wheel rate of turn and direction of rotation in a ride control system (Figure 11.10). The sensor is mounted inside the passenger compartment. Multiple slots in a shutter interrupt the light path between the two pairs of emitters and detectors used to provide the quadrature signal. The opto sensor does not require a magnetic trigger wheel, like most speed sensors. A wheel of stamped steel or any material that will block the light path is acceptable. Schmitt trigger outputs that are integrated on the opto detector provide a signal easily connected to an MCU. The MCU in the ride control system uses an algorithm that determines the straight-ahead wheel position and then calculates the lateral acceleration that will result from the turn angle. If the vehicle acceleration is above a programmed level, the shock absorber damping is switched to firm. The optical steering sensor also provides information that allows the MCU

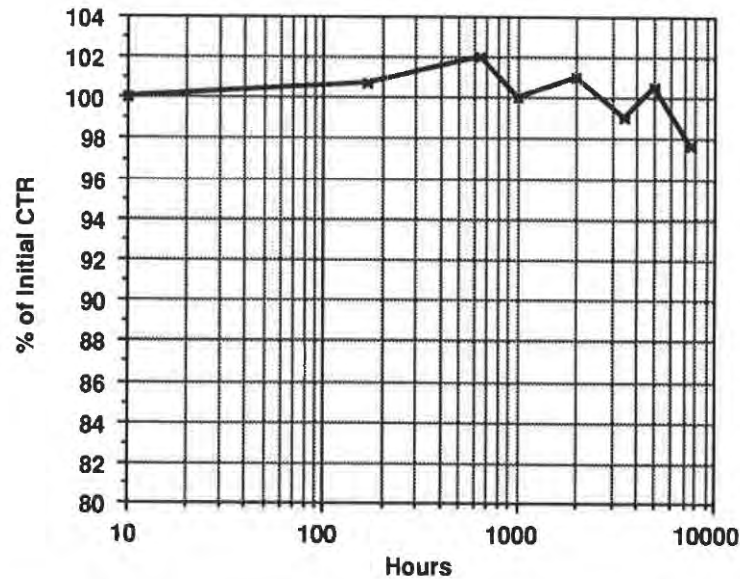


Figure 11.9 Improvement in opto stability.

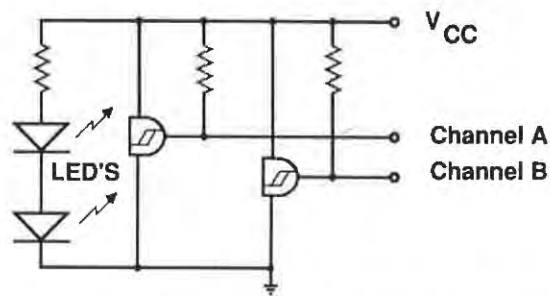


Figure 11.10 Opto sensing for steering-wheel position and direction.

to determine the rate at which the steering wheel is turning. If this rate is too high, then the damping is switched to firm to provide maneuverability for avoiding accidents [9].

Packaging techniques have also allowed opto devices to be used as a shock absorber-mounted vehicle height sensor (Figure 11.11). Reflective sensing from the stainless steel shaft provides information for the MCU in this case. Opto sensing is made possible by utilizing the clean environment that must exist inside the shock

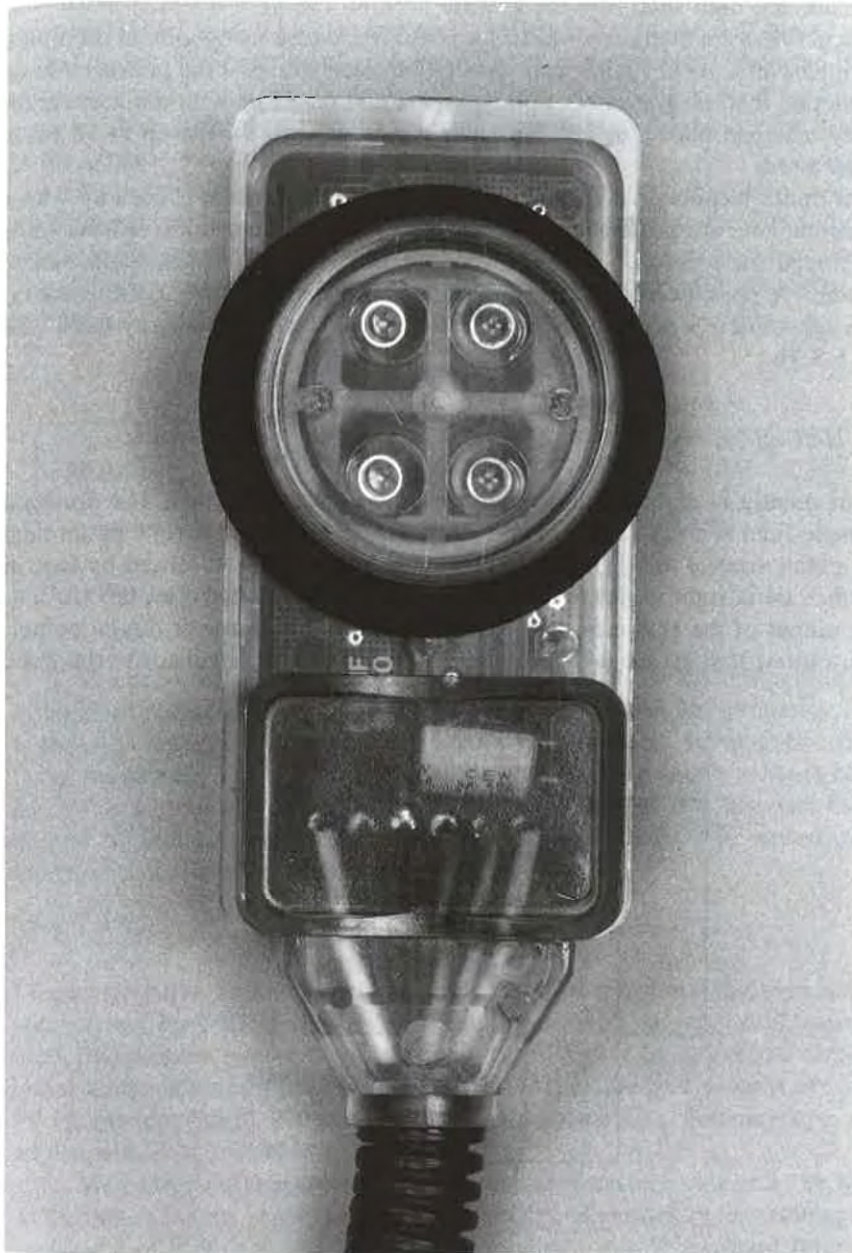


Figure 11.11 Opto vehicle height sensor.

absorber for the stainless steel shaft to move without galling. Dirt and grease are sealed outside the opto sensor by the gasket.

Optical fibers are being considered for noise-free data transmission of multiplex (MUX) signals in critical high-speed data requirements. Low-cost packages have been developed that integrate fiber optic emitters and detectors with the connection to properly trimmed plastic optical fiber. Applications with speeds up to 10 MHz can be addressed.

Fiber optics technology also has several sensing applications, especially where harsh environments are present and isolation is required. Fiber optic sensors have been developed for rotation, vibration, acceleration, acoustic, strain, temperature, and pressure. A possible automotive application is the use of fiber optic gyros to determine the vehicle's position in future intelligent vehicle highway systems (see Section 11.3.7).

11.2.3.2 Hall-Effect Sensing

Another frequently used principle to sense position is the Hall effect. The presence of a magnetic field at a right angle to an epitaxial layer with current flowing through it causes a Hall voltage to be generated [10]. The Hall voltage is sensed by taps in the layer that are at right angles to the current flow, as shown in Figure 11.12 [11].

The output of the Hall-effect device is converted to a linear or digital output. The linear output Hall sensor adds a dc amplifier and voltage regulator to the basic

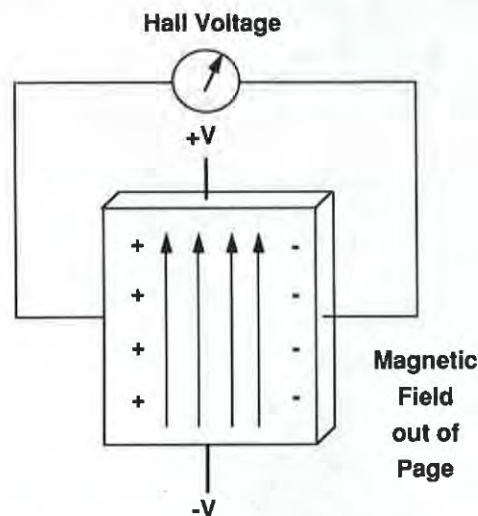


Figure 11.12 Hall-effect principle.

Hall element to provide an output that is linear, proportional to the applied magnetic flux density, and ratiometric (proportional) to the supply voltage. A Schmitt trigger threshold detector with built-in hysteresis added to the linear Hall sensor provides a digital output. The addition of an open collector *npn* transistor allows the Hall switch to be compatible with digital logic. Hall-effect devices have been used for several years as sensors for detecting the speed and position of the distributor trigger wheel in the ignition system.

Addressable Hall-effect sensors have been designed for automotive multiplex wiring systems [12]. The sensor has two pins that connect to the bus and a third pin available for sensing an open or closed switch without requiring a magnetic field. The two-wire bus can power up to 30 sensors that require 2.5 mA each, as shown in Figure 11.13 [10].

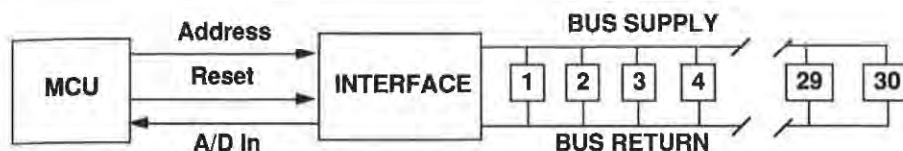


Figure 11.13 MCU addressable Hall-effect sensors.

The presence of the factory programmed address on the bus causes the sensor to transmit a diagnostic signal indicating that it is functioning properly. This signal is followed by a bit indicating the presence or absence of the magnetic field that is being monitored. The chip is addressed by switching the supply voltage between 6V and 9V in a serial pulse train. When the sensor is addressed it increases its quiescent current by 300 μ A when the rail is 9V. If a magnetic field is sensed, the current increase is continued when the rail is reduced to 6V.

11.2.3.3 Magnetoresistive Elements

Magnetoresistive sensors are being developed to overcome the limitations of inductive and Hall-effect sensors for linear position, rotational, and speed sensing. Inductive pickups require complex signal conditioning for the signal output and external temperature compensation networks [13]. Hall-effect devices are limited due to temperature drift, temperature range, repeatability, nonlinearity, and offset adjustment.

Magnetoresistive elements (also magnetic resonance elements, or MRE) such as the one in Figure 11.14 [14] utilize four strips of permalloy, a ferromagnetic alloy composed of 20% iron and 80% nickel, arranged in a Wheatstone bridge configuration [15]. The output signal, based on a change in resistance, is proportional to

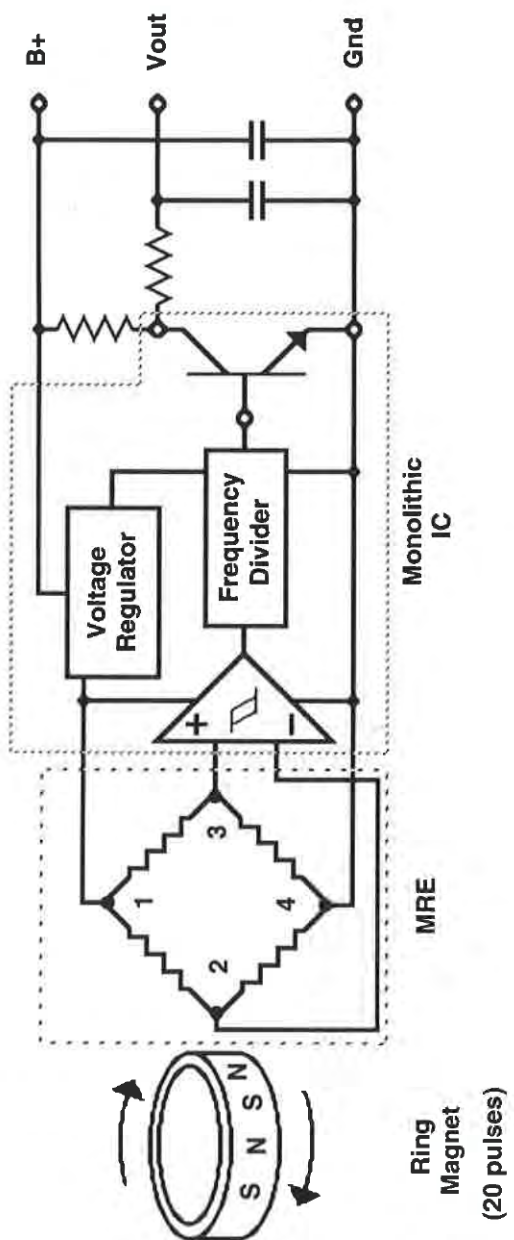


Figure 11.14 Magnetoresistive element (MRE).

the input signal magnitude modulated by the magnitude of the magnetic field. The rotation of a 20-pole magnet ring causes the magnetic field to change. Each rotation of the magnetic ring produces 20 cycles of alternating current output from the MRE. A comparator changes the ac signal to a digital signal, the signal is inverted and then divided into five equal parts by a combination meter. One part is used to drive the vehicle's speedometer and the other four signals are used for inputs to suspension, engine and transmission, cruise control, and power steering.

Linear characteristics up to 175°C have been demonstrated by MRE sensors. An additional application being pursued is replacing the potentiometer, with its associated wearout mechanism, by a noncontact MRE sensor for measuring the throttle position.

11.2.4 Flow

The measurement of flow is performed in automobiles for both liquid and air. One of the more important flow measurements is mass air flow. Karman vortex and hot wire anemometer approaches are used for sensing the mass air flow in engine control systems. A mass air flow measurement in an engine control system provides an improvement in the ability to adjust the fuel flow for improved combustion efficiency over speed-density engine control systems that utilize MAP measurements. However, the mass air flow sensor is more expensive than a MAP sensor. Stricter emission standards are increasing the use of mass air flow sensors. The mass of the air is measured instead of the volume to avoid the effect of temperature and density on the air. Other vehicle flow measurements include fuel flow and air flow in the passenger compartment. Fuel flow is frequently sensed by a turbine in the flow line whose blades are counted by magnetic sensor such as a Hall-effect sensor. Passenger compartment air flow could possibly utilize microsensor technology.

11.2.4.1 Mass Air Flow Sensing

A typical approach for mass air flow involves a hot wire anemometer with a platinum wire as the sense elements as shown in Figure 11.15 [16]. The air temperature probe detects the temperature of the air flowing into the combustion process. The hot wire probe is heated by current supplied from a power transistor (Figure 11.15(a)). The difference in temperature between the probes is detected as a difference in electrical resistance. Electronic circuitry regulates the heating current to the hot wire probe to keep the temperature difference constant at any flow rate. The current required is proportional to the air flow rate. The construction of one of the probes is shown in Figure 11.15(b). One of the main issues in this type of sensor is response time. Improvements that reduce the response time from 50 to 30 ms have been made in recently developed units.

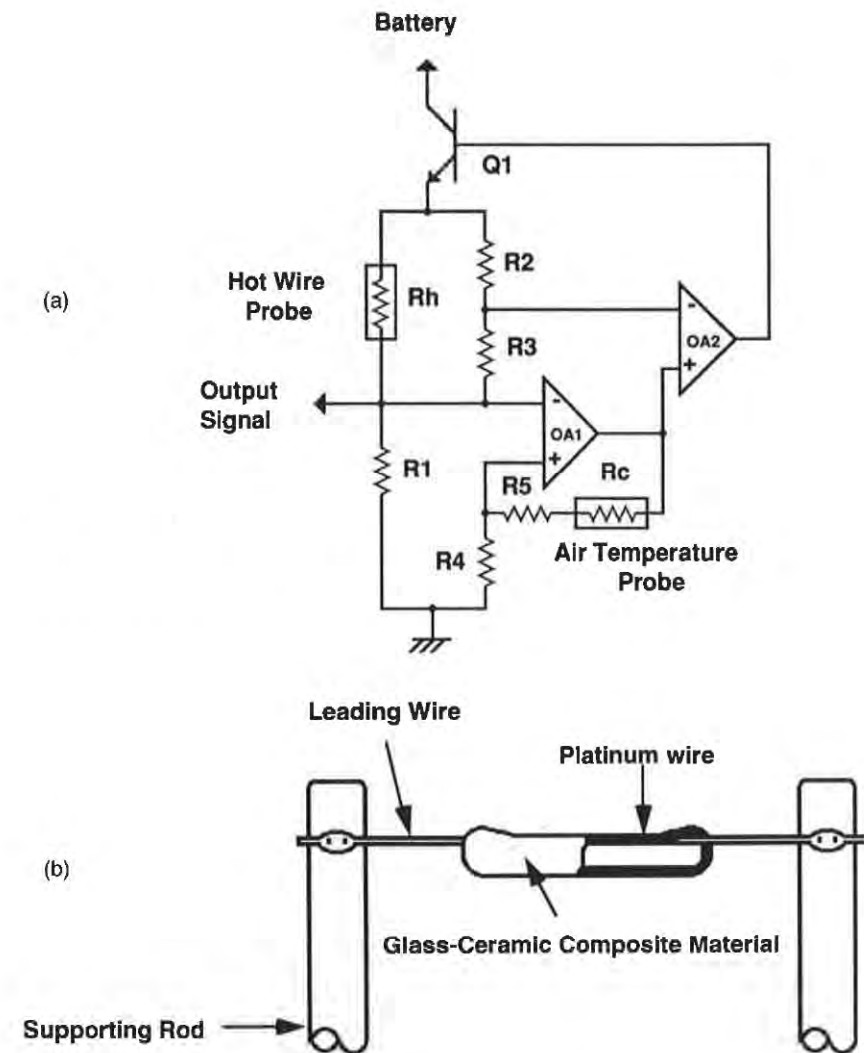


Figure 11.15 (a) Circuit for hot wire anemometer mass air flow sensor and (b) construction of probe.

Microbridge silicon mass air flow sensors have also been developed that utilize temperature resistive films laminated within a thin film (2 to 3 μm) of dielectric material and micromachined cavity as shown in Figure 11.16 [17]. The resistors are suspended over the etched cavity, which also provides thermal isolation for the heater and sensing resistor. Heat is transferred from one resistor to the other as the result

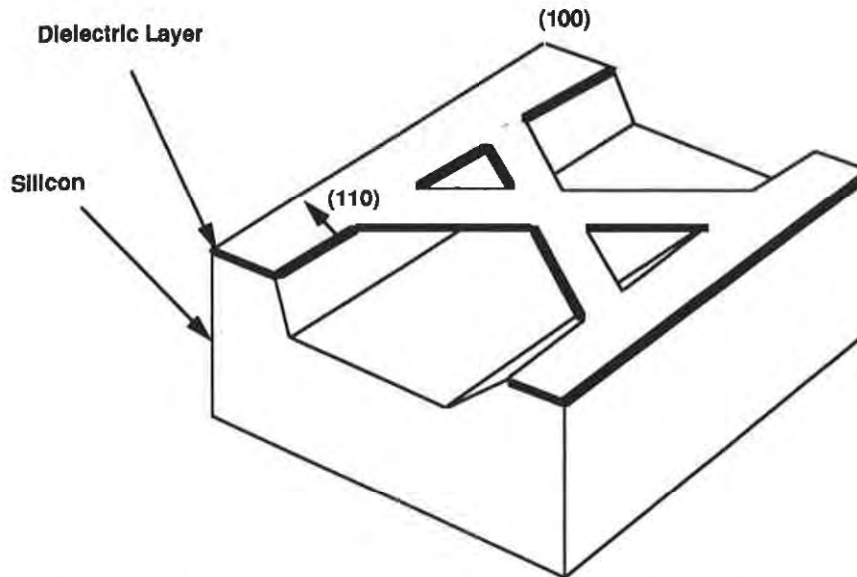


Figure 11.16 Microbridge mass air flow sensor.

of flow. The imbalance in the resistance caused by the heat transfer is directly proportional to the flow. The small size of the resistors allows fast response time (≤ 3 ms) and high sensitivity to flow. The device is also useful as a differential or absolute temperature sensor. Silicon microbridges are being pursued for sensing mass air flow in future vehicles.

11.2.5 Accelerometers

Sensing acceleration is one of the most intense development areas. Accelerometers are required for air bag and active suspension systems but can also be used in antilock brake and ride control systems if the cost is low enough. The most critical and highest volume application is currently in the area of crash sensors for air bags.

An inflatable restraint (air bag) system is one of the more important new vehicle systems designed to address the need for improved vehicle safety and satisfy the legislated requirements of FMVSS208 (Federal Motor Vehicle Safety Standard). Air bag systems utilize as many as five accelerometers (crash sensors) to provide information to an electronic control unit. The crash sensors are positioned at strategic locations in the front of the vehicle, and a low arming sensor is located in the passenger compartment. When a crash occurs, the (series) arming sensor is closed and

the rapid deceleration produces a signal from the crash sensor. If this signal is above the calibrated level of a crash event (equivalent to running into a wall at 12 mph) it is allowed to pass the trigger current necessary to ignite the propellant that inflates the bag.

Present systems utilize electromechanical and piezoelectric sensors to provide an air bag deployment signal. These systems have an excellent record. However, the system costs about \$400 to \$500 for the driver-side-only system and is estimated to cost from \$800 to \$1,000 when passenger-side units are added. A significant portion of the cost is attributed to the sensors. When light trucks with shorter crash zones, especially vans, are added to the list of vehicles that require air bags, a faster input from the crash sensor will be required. In addition to meeting the system performance levels, reliability requirements, and cost objectives of this application, semiconductor accelerometers are being pursued for the capability to provide a faster trigger, self-test (diagnostic) feature, reduce the number of sensors in the system, and their potential for increased integration [18].

11.2.5.1 Mechanical Crash Sensors

One version of the crash sensor used in current vehicles is shown in Figure 11.17 [19]. The sensor consists of a roller positioned against a calibration backstop by a

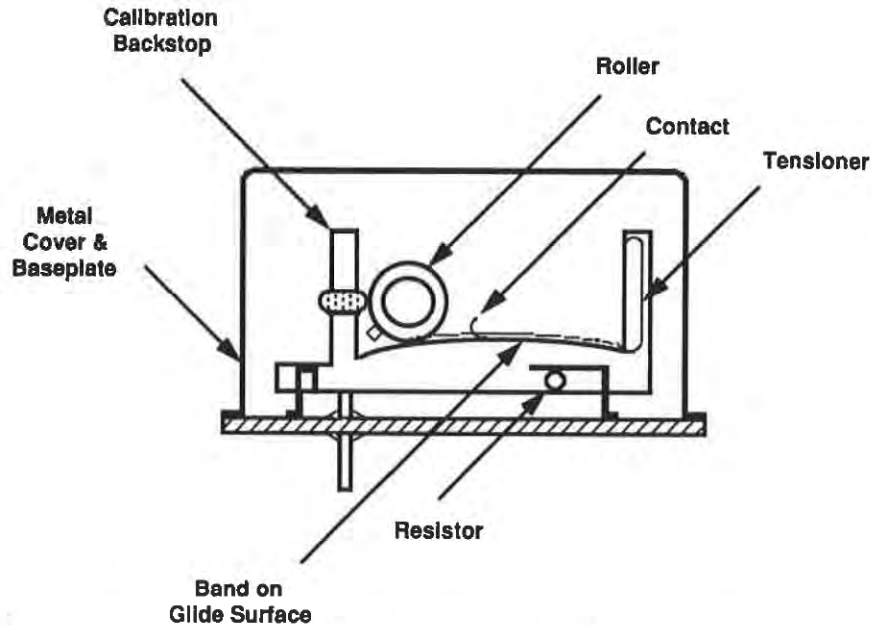


Figure 11.17 Mechanical crash sensor.

flat spring band. Sufficient deceleration causes the roller to move forward and close the contact. Calibration allows the level of deceleration necessary to close the switch to be set at different levels for different vehicles. Other mechanical approaches include a ball in tube that is restrained by either a spring or magnetic force. Deceleration overcomes this force and allows a crash signal to be indicated.

11.2.5.2 Piezoelectric Accelerometers

Ceramic and thin-film piezoelectric sensors solve some of the cost and operational problems associated with mechanical crash sensors. They are analog sensors and not switches like the mechanical units. Figure 11.18 shows a piezoelectric sensor currently used in some vehicles [20]. The sensing element consists of two reversely polarized piezo-oxide bars with two electrodes. The bars are cemented together to form a biomorph element. The cantilever structure is positioned in the module in the direction of vehicle motion. When deceleration occurs, the upper bar is stretched and the lower bar is compressed. Because the bars are reversely polarized a signal doubling occurs. An external stage with an impedance converter, low-pass filter, and amplification are necessary to provide a usable output.

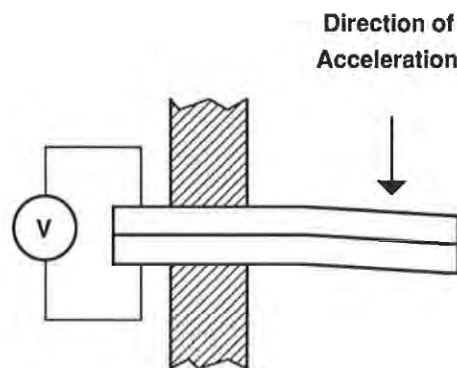


Figure 11.18 Piezoelectric accelerometer (crash sensor).

11.2.5.3 Silicon Accelerometers

Micromachining techniques utilized in the manufacturing of piezoresistive pressure sensors such as the MAP sensor have led to the development of silicon accelerometers. Figure 11.19 [21] shows a piezoresistive silicon accelerometer with micro-machined beams and suspended mass. Bulk micromachining allows a very precise

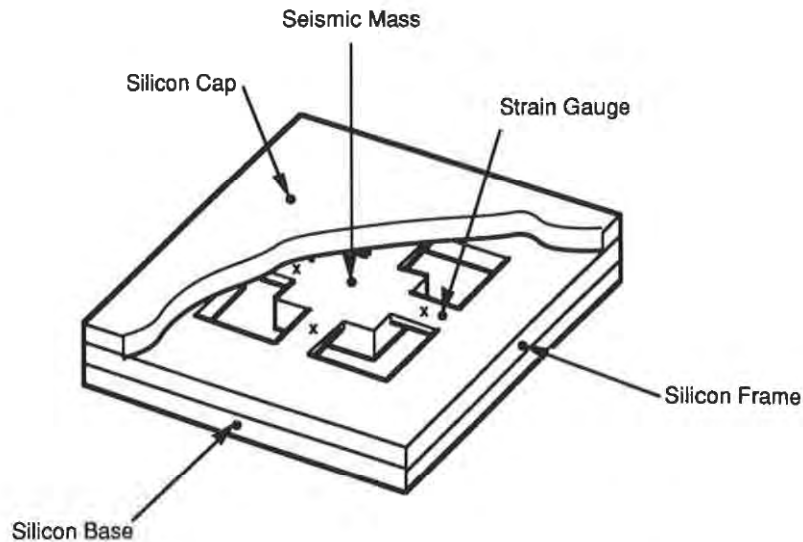


Figure 11.19 Piezoresistive silicon accelerometer.

mass and support structure to be consistently produced in a batch process with several thousand devices in a single-wafer lot. The piezoresistive element(s) of the strain gauge are ion implanted into the suspension arm for maximum sensitivity. The output is in mV/g and the resistive elements are temperature sensitive so additional circuitry is required to provide calibration, temperature compensation, and allow interface with external circuits.

Capacitive approaches for semiconductor accelerometers are also possible, such as the structure shown in Figure 11.20 [22]. In this case, the working structure is obtained by the deposition of sacrificial and structural layers on the top of a silicon substrate. Selective chemical etching is used to etch the sacrificial layer separating the conductive layers to produce the air gap capacitor. Acceleration perpendicular to the substrate surface causes the movable plates of the capacitor to move, producing an output that is directly proportional to acceleration. The output (in fF/g) is converted into a frequency for direct interface to a microcontrol unit.

One of the major advantages of micromachined silicon accelerometers is the ability to integrate self-test features. An electrostatic field applied to an integral capacitor built into a silicon accelerometer can be used to deflect the beam and provide an indication that the accelerometer is working properly and even check the calibration. The ability to check the calibration can be used to test the accelerometer in wafer form, packaged form, and in the final assembly. In the application, the device can be tested for drift and lifetime measurements without having to apply external

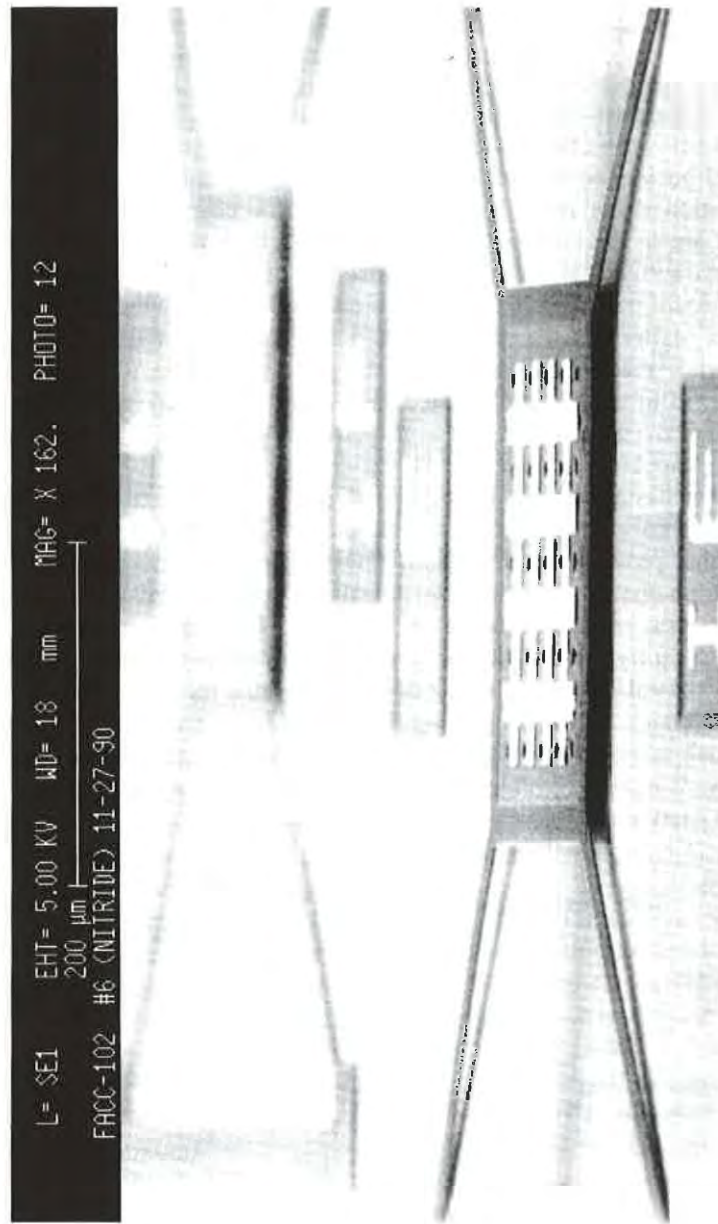


Figure 11.20 Capacitive accelerometer.

forces or the occurrence of a crash. More details on accelerometers can be found in Chapter 12.

11.2.5.4 Knock Sensors

The use of lower octane, lead-free fuels and the improved fuel economy gained by operating just below an engine's knock limit make sensing engine knock important. In addition, turbocharged engines can experience significant damage at high speeds due to detonation if the spark is not controlled. The characteristic frequencies of engine knock are in the range of 3 to 9 kHz [23]. Both resonant and nonresonant (flat bandwidth within 0 to 10 kHz) knock sensors are used in automotive applications. The resonant units detect secondary mode knock that can occur above 10 kHz and provide a higher signal-to-noise ratio. The knock sensor uses a piezoelectric ceramic material, such as barium titanate, to generate a voltage when the sensing mass is subjected to a force due to the acceleration that occurs with engine knock. The amplitude of the generated signal is proportional to the acceleration and therefore to the level of knock.

The sensor has a simple construction (Figure 11.21) and can be manufactured at relatively low cost. The ceramic sensing element is captured between two conductive caps that provide a current path. Engine-generated knock is transmitted to the element from the conductive metal housing through a Belleville washer. The output level is in the range of hundreds of millivolts. The sensitivity to noise requires amplification circuitry external to the sensor. In addition, the need for filtering in the electronic control unit for different engine types and the possibility of requiring more than one knock sensor is spurring development for more direct measurement techniques, such as cylinder pressure, combustion temperature and in-cylinder flame measurements. It is also possible that an integrated sensor could perform the total function even more economically.

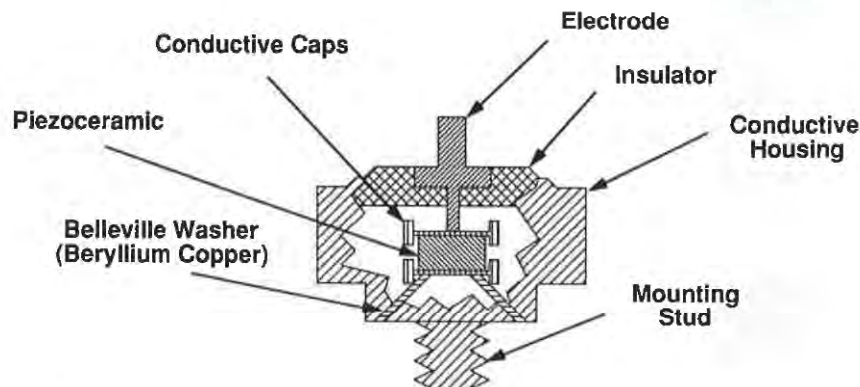


Figure 11.21 Knock sensor.

11.3 FUTURE AUTOMOTIVE SENSING

The increasing complexity of vehicle systems combined with the need for additional sensing inputs with improved accuracy and the automotive environment pose several challenges for sensor designers. Some of the areas that will affect the types of sensors and how sensors will be used in future systems are covered in the remaining sections.

11.3.1 High-Temperature Operation

Higher temperature operation of automobiles, especially for underhood applications ($\geq 125^{\circ}\text{C}$), is pushing the maximum operating temperature of silicon. Silicon sensors are being used more frequently in higher operating temperatures. Alternate semiconductor materials, such as GaAs (gallium arsenide), GaP (gallium phosphorus), GaAs with silicon, and SiC (silicon carbide) are among the possible solutions. These wide-bandgap materials can operate at temperatures that exceed the maximum operating temperature of silicon. Silicon bonding technology is also being pursued as a potential means of obtaining silicon sensors that can survive higher temperature operation. In addition to performance at higher operating temperatures, the stresses placed on the package require sensor manufacturers to address thermal cycling capability with high-temperature extremes.

11.3.2 Liquid Level

Various liquid levels can be measured in a vehicle: engine oil, transmission oil, coolant, windshield-washer fluid, battery, power-steering fluid, brake fluid, and fuel. All of these sensors, except fuel level, could be satisfied by a switch that simply detects that when a predefined minimum level of liquid has been reached so that a driver indication can be provided, either directly illuminating a dash lamp or through a body computer activating an output driver. One way to sense a minimum level of fluid uses a Hall-effect sensor (Section 11.2.3.2) [11]. The magnet is located in a float assembly that keeps the Hall device open as long as the fluid level is sufficiently high. When the minimum level is reached, the Hall switch closes and the integrated Darlington output stage allows current up to a 900 mA peak to flow. Internal current limiting allows the integrated sensor to turn on a dash lamp, such as a Sylvania #168. The circuit also incorporates thermal shutdown, shortcircuit protection, fly-back voltage protection, and reverse battery protection. The float and housing material can be chosen to withstand the effects of the fluid being monitored, and the plastic packaged Hall sensor is isolated from the fluid.

Sensing the fuel level in the gas tank is one of the more difficult fluid-level applications. In this case an analog output is required with good accuracy over the entire fuel level. Present systems utilize a potentiometer with a float mechanism. A semiconductor pressure sensor provides an excellent solution for meeting improved

performance requirements and avoiding wearout mechanisms; further, they can provide additional features. However, in addition to exposure to gasoline, the sensor package must withstand alcohol and sour gas. Coping with these harsh media has been a problem for sensors that do not incorporate a stainless steel diaphragm; however, stainless steel would not meet the system cost requirements. The use of new materials and techniques continues to be pursued to solve the media compatibility problem.

Other alternatives for fuel tank sensors such as capacitive sensing and ultrasonics have been effective in other harsh liquid level applications. These techniques are also being developed for automotive use.

11.3.3 Chemical Sensing

Chemical sensors can provide an indication of CO or noxious odors in the passenger compartment. The presence of CO could trigger a vehicle alarm, and the presence of noxious odors could be used to automatically change the mode of the heating and air conditioning system from a recirculation of passenger compartment air to allow fresh air into the passenger compartment or the reverse situation depending on the location of the offensive odor. Chemical sensors are being developed for vehicle applications based on thin-film of stannic oxide (SnO_2) whose conductivity changes with the concentration of a specific gas such as CO, H_2S , methane, and other hydrocarbons. The selectivity, operating temperature, and sensitivity of these devices as well as cost are issues that will determine their acceptance in automotive applications.

Chemical sensing already plays an important role in the emissions control system. An oxygen sensor mounted in the exhaust gas provides the closed loop information necessary for the MCU to control vehicle emissions. Sensing NO_x is an area that continues to be investigated as a means of achieving tighter emission controls.

11.3.3.1 Oxygen Sensor

An oxygen sensor is used to monitor the exhaust gas mixture and provide information on the status of the exhaust gas content to the MCU [24]. The control strategy allows the MCU to vary the air-fuel mixture around stoichiometry (14.6 A/F) to increase catalytic converter efficiency, prolong the life of the converter, and provide increased fuel economy. To ensure the fastest response time the oxygen sensor is placed in the exhaust manifold at a point where it can respond quickly to exhaust changes with minimal lag time. The operating temperature inside the manifold can easily exceed 900°C , and in fact, a minimum operating temperature (about 300°C) is necessary to ensure proper performance of existing sensors. External electrical power is frequently applied to an integral heating element in the sensor so that closed-loop operation can be initiated within 30 seconds of starting and during idle.

The sensor is a hollow cone shaped zirconia ceramic (ZrO_2) covered with a porous platinum electrode on both sides (Figure 11.22(a) [25]). The overall effect

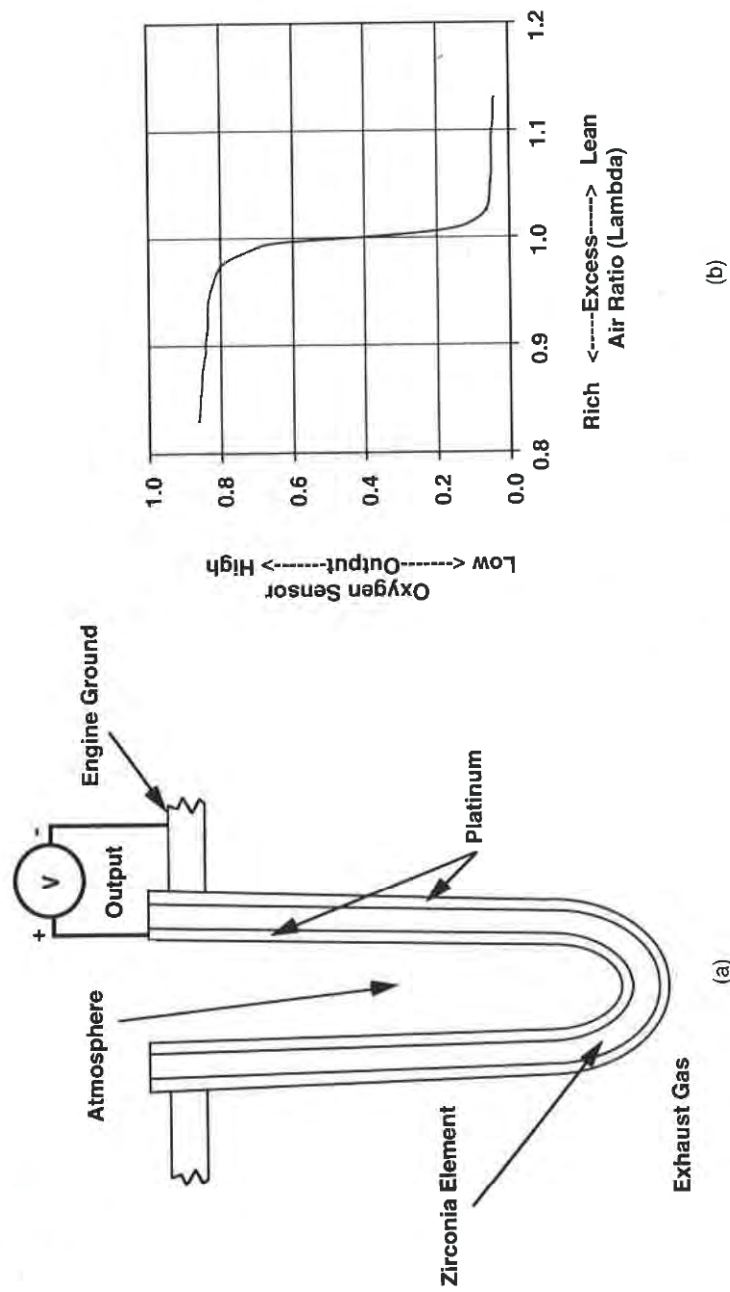


Figure 11.22 Oxygen sensor: (a) construction; (b) output versus air ratio.

of the plating is to create an electrochemical cell that develops a potential difference between the electrodes. The sensor's output voltage (Figure 11.22(b) [25]) is a logarithmic function of the partial pressures of oxygen on either side of the cell wall. One side of the cell is exposed to exhaust gas and the other to atmospheric air. When the engine is operating in a lean mode, the exhaust gas oxygen increases and the partial pressure of oxygen on both sides of the ceramic are almost equal, causing the sensor's output voltage to be near zero volts. As the exhaust gas is enriched, it passes stoichiometry, which is the ideal burn point. Just beyond stoichiometry, oxides (like CO and hydrogen) increase dramatically, which increases the pressure differential across the electrolyte, causing a voltage to be generated (approximately 1V). Recent developments include efforts to increase the response time (≤ 10 ms) so that individual cylinders and not just the average composition can be detected. One approach uses a thin ($1\text{ }\mu\text{m}$) film semiconducting metal-oxide which is deposited by reactive cathode sputtering on an aluminum oxide substrate [26].

11.3.4 Oil Quality

Oil life sensors, such as the one used on 1991 Buicks [27], are intended to alert drivers to the need to change engine oil and avoid costly engine failures. The Buick sensor was developed by GM Research Labs and uses accumulated engine rpm, time, and engine temperature as inputs to a microcontroller that calculates an effective oil quality and provides an indication. The calculation is really an inferred oil quality and not a direct indication of the actual oil quality (see Section 11.3.5). Ford has also developed a capacitive sensor that measures the change in the dielectric constant of the oil as an indicator to viscosity and therefore oil condition [28]. However, the dielectric constant of different oils varies considerably and as a result the production usage of this approach has not occurred to date. Another approach measures oil viscosity by determining the time it takes for a sample to flow through a small chamber, which is a more direct indication of the actual state of the engine's oil. However, it has proven to be too costly for automobiles but may be applicable for trucks.

11.3.5 Capability to Compute Rather than Sense

The MCU and digital control have been significant driving forces for developing microelectronic sensors. At the same time, an MCU can also limit the number of sensors used in a particular system. The ability of the MCU to calculate from information provided by existing sensors and obtain an inferred measurement can avoid the cost of adding a sensor even though the sensor could provide more accurate data. A paradox occurs when a need exists for sensing redundancy. The cost to benefit must be evaluated for every new sensor that could be added to a system. To demonstrate the importance of this situation in an extremely cost sensitive environment, three specific cases will be examined.

Monitoring engine misfire is one of the requirements of the California Air Resources Board (CARB) regulation for on-board vehicle diagnostics (OBDII), which will be required on 1996 vehicles. California vehicle emissions standards are usually adopted by the federal government's Environmental Protection Agency. As a result, considerable research is being conducted to determine the most cost-effective way to monitor engine misfire [29]. One technique to measure misfire of an individual cylinder utilizes a magnetic sensor in the crankshaft to measure slight deviation in rpm and measurement of torque based on a nonuniformity index. The resulting values are compared to predicted values of power and torque [30]. Another system uses the difference in angular displacement at each end of the crankshaft. The twist of the shaft is determined by measuring the phase differences between the two encoders; this measurement is proportional to torque [31]. In-cylinder pressure measurement is the most straightforward and accurate method of measuring misfire. However, the technique requires relatively expensive sensors, which not only must measure high levels of pressure but withstand combustion temperatures. Also, a pressure sensor must be used in each cylinder [32].

Early requirements to measure manifold absolute pressure and barometric pressure utilized separate sensors. By reading the barometric pressure when the input to the MAP sensor is not affected by manifold pressure (during cranking and at wide-open throttle conditions that other sensors or switches indicate to the MCU) the additional barometric sensor has been eliminated. It should be noted that the manifold absolute pressure measurement is also an indirect measurement compared to measuring mass air flow.

Electronic antilock brake systems demonstrate an alternative to sensing hydraulic pressure. ABS is currently offered on selected models by every auto manufacturer. Use of ABS by some manufacturers is projected for availability on 100% of the vehicles by the mid 1990s. ABS is regarded as the major safety and driving option on present production automobiles. A typical ABS system uses a microprocessor, wheel-speed sensors, semiconductor power control devices, and hydraulic components to automatically pump the brakes rapidly and repeatedly (up to 15 times per second) [33]. This allows the driver to maintain steering ability and retain vehicle stability under difficult road surfaces, such as ice or wet asphalt, and under hard braking conditions.

Most ABS systems use fast-acting solenoid valves activated by power transistors to control the hydraulic pressure applied to the wheels and avoid locking up on slippery surfaces or during panic stops. Sensors measure the speed of the wheels and provide a signal to the microprocessor. The frequency of the signal is proportional to the speed of the wheel. When a wheel lock-up is detected, the MCU drives power FETs to activate solenoid valves that keep each wheel in an optimum slip range.

GM's ABS-VI systems uses three dc electric motors to move pistons inside the hydraulic chambers to bleed off brake fluid when wheel lockup is detected [34]. One of the major advantages of this approach is the capability to obtain information on brake pressure without requiring transducers for measuring hydraulic pressure. In

ABS-VI, the electric current going to the motors is proportional to brakeline pressure. This allows the system to apply optimum brake pressure to stop each wheel without locking. Partial excursions of the piston can be initiated instead of the on-or-off action of solenoids. As a result of this approach, the ABS-VI system has a very low level of brake pedal pulsation. Systems with solenoids can also benefit from information on line pressure because the solenoids could maintain the system pressure at the same level before and after release. However, pressure transducers have to be added to make the measurement because the system does not inherently provide this information.

11.3.6 Vehicle Diagnostics

Vehicle diagnostics are important not only for on-board detection and warning of system faults but for subsequent service for fault conditions. The complexity of electronic systems and the problems of diagnosing system faults have caused the auto industry to develop service diagnostic equipment. Ford's Service Bay Diagnostics System, Chrysler's Mopar Diagnostic System, and GM's Computerized Maintenance System are designed to provide service technicians with the ability to diagnose and correct malfunctions in electronically controlled vehicles on the first try [35]. These systems also allow reduced repair time and increased shop productivity. Continued growth of sophisticated vehicle systems will rely even more on these types of diagnostic units. On-board vehicle sensors provide information through the vehicle's electronic control units. Sensors on the diagnostic equipment provide additional information for verifying calibration and proper function. For example, fault sensing in an inflatable restraint system can provide information on a failed accelerometer that is incapable of providing any signal in the event of a crash and an extremely out of calibration accelerometer that is incapable of providing the correct firing point. An indication to the driver of a system problem and the driver's reaction to the warning by obtaining service to correct the fault are essential.

As indicated in Section 11.3.5, in certain instances sensors provide sufficient information, based on secondary parameters, for MCU algorithms to determine that a fault condition exists and still react to the inputs from other sensors to provide an estimate of how the system should respond rather than strictly going into a default mode. This can provide substantially better performance and safety than simply disabling the system until it is repaired. For example, in an ABS system with four wheel sensors, an open or shorted sensor providing no input can possibly be ignored for certain driving situations especially if other vehicle sensors, such as an indication of braking actuation or deceleration sensor, are providing system inputs that allow alternative calculations to be performed.

The SAE has developed a standard for vehicle diagnostic interfaces as indicated earlier. The International Standards Organization also has a standard for diagnostic interfacing.

11.3.7 RF Sensor Applications

The use of radio frequency (RF) sensing techniques is necessary in some of the most sophisticated systems being developed for future vehicles. Three of these systems are collision avoidance, near obstacle detection (NOD), and intelligent vehicle highway systems (IVHS).

Artificial intelligence simulating human visual, sensory, and judgment functions is the basis of a collision avoidance system [36]. Eight ultrasonic sensors (four in the front and four at the rear of the vehicle) provide information for the range finder in the fuzzy logic control. In addition, two infrared and two touch sensors are mounted on the front and rear bumpers and a color video camera provides long range image detection. When an object is detected that could cause a collision the vehicle's speed is reduced. Blind spot detection and avoiding problems at night and in fog or due to driver fatigue are just a few of the frequently encountered causes for accidents that could be avoided with the use of such a system.

Near obstacle detection is a simplification of a full collision avoidance system. In low-speed traffic maneuvers, such as backing out of a driveway or parking, the presence of an object in the vehicle's path could trigger a warning alarm or automatically apply the brakes. Only short-range ultrasonic, infrared, or radar sensors would be required for this system [37]. A head-up display (HUD) may be part of blind spot detection provided to the driver.

Intelligent vehicle highway systems or smart highways are being evaluated worldwide in an effort to reduce traffic congestion and improve traffic flow in the most populated cities. Improved traffic flow results from the use of traffic sensors and communication links that provide information to the driver regarding the best route to select to minimize the distance traveled. The sensors are 6-ft loops of heavy-gauge wire embedded beneath the pavement that are used to count the traffic flow or receivers that detect infrared signals transmitted from beacons at key intersections. Another aspect of IVHS is driver information that involves bidirectional RF communication with the vehicle. Several potential areas are shown below [38].

- Dead reckoning map matching navigation system with GPS (global positioning satellites);
- Digital traffic information and network link times receiver;
- Dynamic route selection for minimal time/distance travel;
- Color video display for maps, traffic info, guidance and road signs;
- Synthesized voice for traffic info and guidance;
- Map database including turn restrictions, traffic signs and nominal link times;
- Business directory and travel info database integrated with map database;
- Electronic vehicle identification for toll debiting;
- Digital cellular telephone; and
- Semiautomatic MAYDAY using cellular telephone.

Several test systems have been implemented to demonstrate various aspects of IVHS, such as the ADVANCE (advanced driver and vehicle advisory navigation concept) in the Chicago, Illinois, area. Figure 11.23 shows the interaction of the vehicle with global positioning satellites, embedded road sensors and two-way RF communication.

Dead reckoning will require precision magnetic field sensors to provide the actual vehicle location for many of these feature to be implemented. Collision avoidance or near obstacle detection could ultimately be part of IVHS.

In brief, the RF sensing techniques for future vehicles include: SAW (surface acoustic wave) sensors, Doppler radar, sonar, ultrasonics and microwave sensors. Some of these RF sensors are already used in anti-theft and remote vehicle entry systems.

11.3.8 Future Sensor Requirements

The number of systems that are being developed for vehicle usage and the number of sensors (and actuators) required for the proper operation of these systems (Table 11.2) will require changes in the way that future sensors are applied to vehicles. Furthermore, the effect of new computing approaches and the resulting complexity of the vehicle under fault conditions will also have an impact on sensing technology. Figure 11.24 [39] shows the interaction of the various systems with some of the required measurements (sensors).

The combination of signal conditioning with sensing elements that allows a direct interface to and bidirectional communication with the microprocessors is called *smart sensors* [5]. Smart sensors will have built-in intelligence with the capability of performing logic functions. The smart sensors and even smart systems are typically regarded as monolithic microsensors that offer fewer components, improved reliability, and system-level cost reduction but may be sensors and additional interface ICs or a custom MCU combined in a single package. Smart sensors can relieve a heavily used MCU of some of its activity and provide more information than a standard sensor. In addition to their ease of interface, smart input devices can also provide protection, fault detection (diagnostics), and status reporting. Based on the increasing complexity of vehicle systems, diagnostics for end-of-line assembly, driver, and service are being viewed as mandatory for manufacturability, safety, and first-time success on service repairs (see Section 11.3.6). Smart sensors will be an integral part both of more complex control strategies and new system architectures, such as fuzzy logic and neural networks.

Combinations of sensors and sensors with actuators will be necessary on future vehicles. Pressure and temperature and flow and temperature are already possible in monolithic microsensors. The combination of a sensor and an actuator has been classified, in general, as the area of mechatronics, and devices have also been called *sensactors* [5]. Use of such devices will outboard some of the control to decentralized control points in the vehicle.

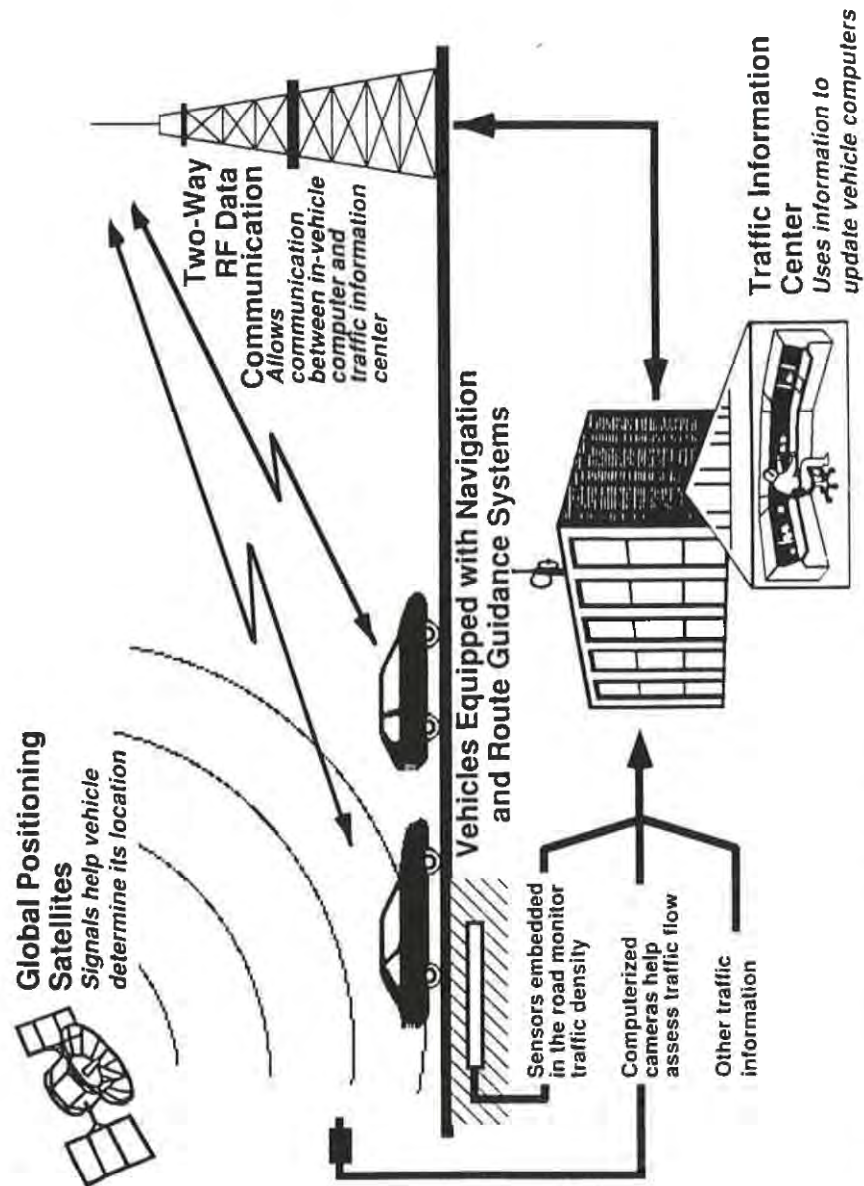


Figure 11.23 Advanced driver and vehicle advisory navigation concept.

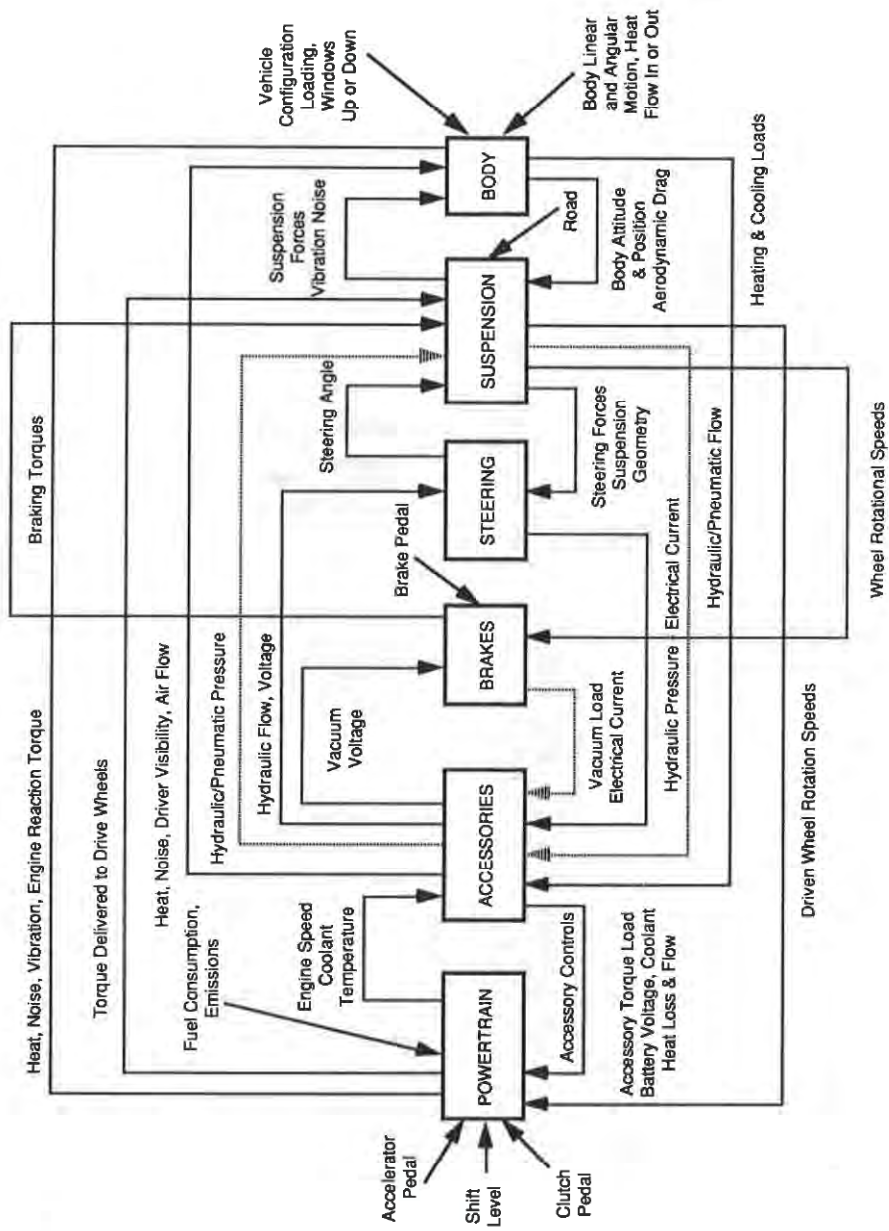


Figure 11.24 Sensor interaction in future vehicle systems.

Multiplexing and multiplex wiring have been implemented in low-volume production vehicles and are considered an important aspect of future vehicles. The SAE has established SAE J1850 as a recommended network interface for MUX systems. A MUX system represents the ultimate integration of vehicle control systems. By utilizing a microprocessor to transfer pulsewidth-modulated input commands over a single data wire at rates of over 5 kB/s to various distributed output devices, the number of wires, weight, and complexity of wiring harnesses can be reduced significantly. This results in easier assembly, higher reliability due to the reduced number of interconnections, and improved diagnostic capability. A fully multiplexed vehicle can easily save over 25 lbs of copper wiring and weight—a saving that translates into improved fuel economy. In addition, system-level simplification will result in a reduced number of components by the sharing of the output from certain sensors, as well as by reduced procurement and vehicle assembly costs. Improved long-term reliability and owner cost will result from the reduction of potential connector failures to half of previous levels. The use of smart sensors will be essential when MUX concepts are applied to a greater portion of the fleet.

MCUs with faster computation capability, higher resolution, fuzzy logic, and neural networks will place additional requirements on the accuracy expectations of vehicle sensors. Systems with 32-bit MCUs are already being pursued by some manufacturers. Fuzzy logic is being applied to collision avoidance systems, cruise control, antilock brake systems, and automatic transmissions [40]. The use of fuzzy sets requires that sensor inputs be assigned values such as high, very high, medium, low, or very low. Each value is given a weight, which allows a decision to be made under circumstances that are imprecise. Neural networks using failure detection filters are being investigated as a real time means of identifying vehicle faults. The computer is trained to recognize the difference in the system between normal operation and under fault conditions, with the capability of observing the whole vehicle system and the dynamic interactions that occur. The full impact of these advanced computational capabilities on sensors is not known at present but the way that sensors are specified in the future is certain to change.

11.4 SUMMARY

The automobile's hostile environment, the high-reliability requirement of automotive systems, and the need for large-volume, low-cost components present both opportunities and significant challenges to sensor manufacturers and system designers. The capabilities of existing sensors are improving at a rapid rate and new technologies are being added or adapted to provide additional approaches to vehicle sensing requirements. The transition of high-precision, low-volume-high-cost sensors to moderate-precision, high-volume-low-cost sensors will provide new sensing approaches for nonautomotive applications that rely on the volume of automotive sales to develop cost-effective technologies that are also rugged and reliable.

This chapter has described a number of vehicle applications for several types of microelectronic sensors in existing and future vehicles. The problems that have been solved and those that must be solved for these devices to be used have been addressed.

REFERENCES

- [1] "New and Emerging Markets for Automotive Sensors in North America," Market Intelligence Research Corporation (MIRC) Rep., 1990.
- [2] *SAE Handbook*, Vol. 2, *Parts and Components*, Warrendale, PA, SAE, latest edition.
- [3] S. Middelhoeck, P. J. French, J. H. Huijsing, and W. J. Lian, "Sensors with Digital or Frequency Output," *Sensors and Actuators*, Vol. A15, 1988, pp. 119-133.
- [4] "Motorola Semiconductors for Automotive Electronic Systems," Motorola brochure SGBR932S/D, 1990.
- [5] J. J. Paulsen, "Powertrain Sensors and Actuators: Driving Toward Optimized Vehicle Performance," Int. Cong. on Transportation Electronics Proc., 1988, Dearborn, MI, pp. 43-63.
- [6] W. D. Siuru, Jr., "Sensing Tire Pressure on the Move," *Sensors*, 1990, pp. 16-19.
- [7] R. G. Wells, "Non-Contacting Sensors: An Update," *Automotive Engineering*, 1988, pp. 39-45.
- [8] Discrete and Special Technologies Group Reliability Audit Report Second Quarter 1990, Motorola Brochure BR923/D Rev 1, p. 35.
- [9] R. K. Jurgen, "Detroit '88: Driver-Friendly Innovations," *IEEE Spectrum*, 1987, pp. 53-57.
- [10] *Integrated and Discrete Semiconductors*, Allegro MicroSystems Inc., Data Book AMS-500, 1991, pp. 4-64-4-65.
- [11] R. Vig, "Protected Hall Effect Contactless Switching of Incandescent and Inductive Loads," *Sensors Expo Proc.*, Chicago, 1990, pp. 205A-1-205A-12.
- [12] F. Goodenough, "Multiple Hall-Sensor ICs Talk to Host via Two-Wire, Two-Way Power Signal Bus," *Electronic Design*, 1989, p. 121.
- [13] D. J. Strycharz and J. H. Houldsworth, "Contactless Position Sensor for Automotive Applications," 1990 IEEE Workshop on Electronic Applications in Transportation, pp. 32-34.
- [14] S. Mizutani and T. Ohtake, "Recent Sensor Technology in Japan," SAE 860410, pp. 35-44.
- [15] R. K. Jurgen, "Global '90 Cars: Electronics-Aided," *IEEE Spectrum*, 1989, p. 45.
- [16] K. Takahashi, S. Tsuruoka, Y. Nishimura, N. Arai, and H. Tokuda, "Hot Wire Air Flow Meter for Engine Control Systems," SAE SP-805, *Sensors and Actuators*, pp. 1-5.
- [17] K. W. Lee and B. E. Walker, "Silicon Micromachining Technology for Automotive Applications," SAE Int. Cong. and Exposition, *Sensors and Actuators*, Vol. SP-655, 1986, pp. 45-53.
- [18] M. Baker, L. C. Puhl, E. A. Dabbish, M. Danielson, and R. Frank, "Sensing and Systems Aspects of Fault Tolerant Electronics Applied to Vehicle Systems," SAE P-233 *Proc. Int. Cong. on Transportation Electronics*, Dearborn, MI, 1990.
- [19] P. Alling, "Solid-State Sensors to Emerge in Mid-90's," *Automotive Electronics J.*, 1990, p. 14.
- [20] D. E. Bergfried, B. Mattes, and M. Rutz, "Electronic Crash Sensors for Restraint Systems," SAE P-233 *Proc. Int. Cong. on Transportation Electronics*, Dearborn, MI, 1990, pp. 169-177.
- [21] W. C. Dunn, "Accelerometer Design Considerations," MICRO SYSTEM Technologies 90, Berlin, 1990.
- [22] W. Dunn, "Automotive Sensor Applications," 1990 IEEE Workshop on Electronic Applications in Transportation, IEEE CAT 90TH0310-3, pp. 25-31.
- [23] W. G. Wolber and P. J. Ebaugh, "Engine Control Sensors: A Status Report," *Automotive Engineering*, Vol. 93 No. 7, 1985, pp. 29-37.

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- [24] D. L. Hittler, M. A. Boguslawski, L. P. Carrion, and R. K. Frank, "A Microcomputer Emissions Control System," Proc. Convergence 80, SAE SP90, B-5 2.
 - [25] H. Shiga and S. Mizutani, *Car Electronics*, Japan, ACLA Inc., 1988, pp. 70-71.
 - [26] John Gosch, "Automotive Gas Sensor Works in 10 mS at a Temperature up to 1000°C," *Electronic Design*, 1991, p. 34.
 - [27] *The Hansen Report on Automotive Electronics*, Vol. 4, No. 1, 1991, p. 7.
 - [28] A. Greenberg, "Cost Puts Direct Oil Condition Sensing in Back Seat," *Automotive Electronics J.*, 1990, pp. 9 and 12.
 - [29] T. Inoue, K. Aoki, and T. Suzuki, "Future Engine Control," SAE P-233, *Proc. Int. Cong. on Transportation Electronics*, Dearborn, MI, 1990, pp. 285-298.
 - [30] K. G. Tatterson, "U of M Aims to Silence Skeptics with Misfire Measurement Unit," 1990, p. 8.
 - [31] M. G. Sheldrick, "Interest in Misfire Detection Technology Grows," *Automotive Electronics J.*, 1989, p. 12.
 - [32] A. Greenberg and M. Boretz, "In-Cylinder Combustion Sensors: Not for Near Term," *Automotive Electronic News* (April 24, 1989), p. 9.
 - [33] B. Nadel, "Antilock Brakes for \$400," *Popular Science*, 1990, pp. 78-81.
 - [34] P. Frame, "GM's Electric-Motor Powered ABS-VI Bucks Conventional Technology," *Automotive News*, 1990, 14i-15i.
 - [35] F. J. Gawrowski, "Ford, Chrysler Offer Dealer Diagnostic Systems," *Automotive News*, 1991, p. 16.
 - [36] B. Harnell, "Fuzzy Logic Guidance Developed by Mazda," *Automotive Electronic News*, 1989, p. 25.
 - [37] D. H. Walsh, "Electronics for Vehicle Safety in the Near and Intermediate Future," SAE P-233, *Proc. Int. Cong. on Transportation Electronics*, Dearborn, MI, 1990, pp. 165-168.
 - [38] J. H. Rillings and R. J. Betsold, "An Advanced Driver Information System for North America," SAE P-233, *Proc. Int. Cong. on Transportation Electronics*, Dearborn, MI, 1990, pp. 85-102.
 - [39] N. A. Shilke, S. M. Rohde, R. D. Fruechte, and J. H. Rillings, "An Automotive Systems Approach," *Automotive Engineering*, 1989, pp. 165-170.
 - [40] A. G. King, "Nissan Patents 'Fuzzy Logic' ABS, Gearbox," *Automotive Electronic News*, 1989, p. 17.

Chapter 12

Signal Processing for Micromachined Sensors

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Motorola

12.1 INTRODUCTION

Silicon, because of its excellent electrical characteristics, is the material of choice in virtually all of the computational and control electronics in use. In addition to that, because of its outstanding mechanical qualities such as high modulus of elasticity, good elastic properties (no hysteresis), machinability in a batch mode, and low cost, silicon is also the prime element of choice for micromachined sensors.

Combining the best of the IC world and the micromachined world has led us to integrated systems—integration of sensors and electronics. The integration process of sensors and IC circuits deals with a broad spectrum of problems, and because of the limits of this chapter, we will focus on some of the problems related to signal processing micromachined sensors. The largest application for micromachined sensors is in the area of sensing mechanical forces such as pressure or acceleration. To sense these measurands, micromechanical structures must be devised for efficient conversion of the measurand into an electrical signal for amplification and conditioning. A number of silicon micromachined [1,2] structures have been devised to that end, using piezoresistive [3], piezoelectric [4], or capacitance [5,6] techniques. Also the corresponding signal conditioning circuits have been designed around these micromachined structures. This chapter will be dealing with those circuits.

12.2 SENSING METHODS

As an introduction to the problem, it is useful to compare devices based on the three basic principles used for the sensing of mechanical forces. These are shown in Table 12.1.

Table 12.1
Methods of Sensing Mechanical Forces

	<i>Piezoelectric</i>	<i>Piezoresistive</i>	<i>Capacitive</i>
Size	Medium	Medium	Small
Sensitivity	Medium	Medium	High
dc response	No	Yes	Yes
Loading effects	High	Low	High
Temperature range	Wide	Medium	Wide
Shift due to shock	Yes	No	No
Self-test features	Possible	Capacitive	Yes
Damping available	No	Yes	Yes

12.2.1 Piezoelectric Sensing

Piezoelectric devices can be cost effective in some accelerometer and pressure sensor applications, but can suffer from offset and temperature-dependent operation. Self-test features are difficult to implement for these devices. The devices usually have a high Q and low damping, which can give them a long settling time. Offsets and temperature effects prevent their use in the frequency range <5 Hz. The piezoelectric device also has a very high output impedance, which makes the device sensitive to load and electromagnetic radiation.

The basic principle of piezoelectric devices is that a piezoelectric material induces a charge or develops a voltage across itself when it is deformed by stress [4]. An accelerometer as shown in Figure 12.1 can be made by placing a layer of piezoelectric material between a mass and a mounting base, with top and bottom contacts. In a similar fashion, a pressure sensor is feasible. The main advantages of piezoelectric sensing are the wide operating temperature range (up to 300°C) and high operating frequency range (up to 100 kHz).

Figure 12.2 shows a typical circuit [7] and the trimming network as used with piezoelectric sensors. The output from the sensor is amplified in a charge amplifier,

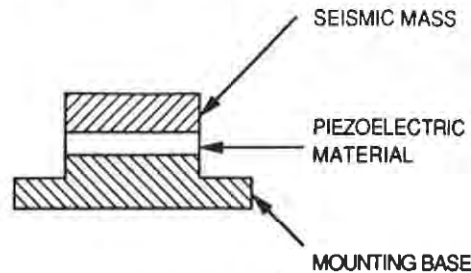


Figure 12.1 Piezoelectric accelerometer.

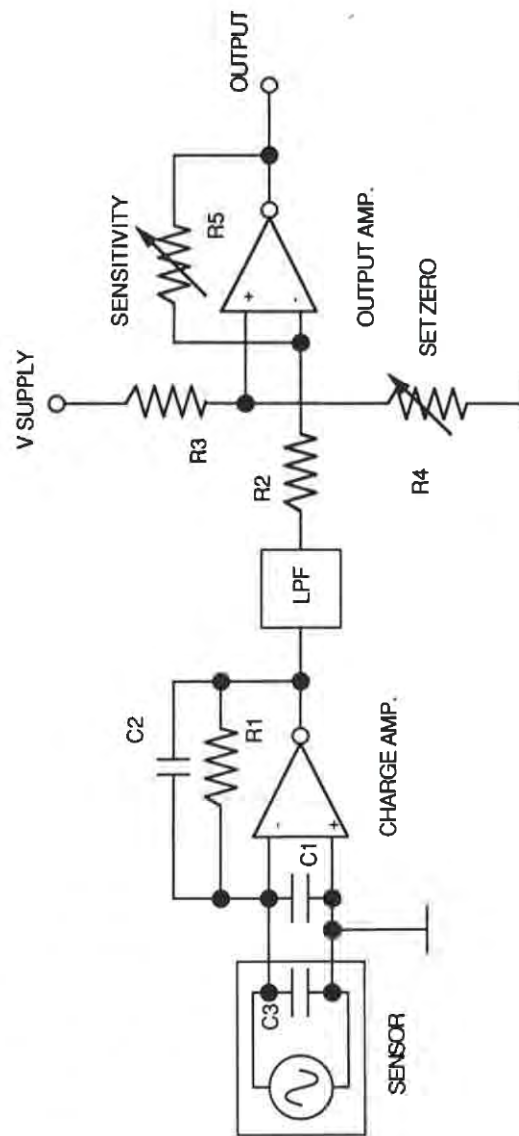


Figure 12.2 Piezoelectric charge amplifier.

which converts the charge generated by the transducer sensor into a voltage that is proportional to the charge. The circuit is a modified voltage amplifier, with feedback via capacitor $C2$ to maintain the input at 0 or ground potential. This eliminates the effect of stray or ground capacitance, $C1$. The output voltage from the charge amplifier is fed via a low-pass filter (LPF) to an output amplifier, where it is trimmed for offset by $R4$ and trimmed for sensitivity by $R5$. If the input of the amplifier is overloaded due to excessive charges generated by high transient forces on the sensor, the amplifier goes into saturation. The recovery time is long, being set by $C2 \times R1$, which can have a time constant of 1 or 2 seconds. The sensitivity of the sensor is set by the piezoelectric material and the size of the mass (1–200 pC/g). The deviation in sensitivity with temperature is higher with higher sensitivity materials. A typical curve of the sensitivity as a function of temperature for a medium sensitivity material is shown in Figure 12.3. The sensitivity varies $\pm 10\%$ from -40°C to $+150^\circ\text{C}$ [8]. Another phenomenon encountered in piezoelectric materials is the pyroelectric effect; that is, changes in the output of the sensor caused by temperature variations. These changes can also be large enough to overload the amplifier at low frequencies (<5 Hz).

12.2.2 Piezoresistive Sensing

The property that silicon changes its resistivity when exposed to stress is called the *piezoresistive effect*. This effect has been in use in pressure sensors for a number of years. The sensing resistors can be either p - or n -type doped regions. The resistance of piezoresistors is very sensitive to strain, and thus to pressure, when correctly placed on the diaphragm of a pressure sensor. Using (100)-orientated silicon material the piezoresistive effect can be described as [3]

$$E = r_0 I \pi_x S \quad (12.1)$$

where E = electric field; r_0 = unstressed bulk silicon resistivity; I = excitation current density; π_x = piezoresistive coefficient; and S = stress tensor due to force.

It is important to note that π_x depends on crystal orientation. For example, the piezoresistive coefficient in the (111) plane ($\pi_x = \pi_{11}$) can be opposite in sign to π_{12} , where π_{12} is the piezoresistive coefficient in the (112) plane. This anisotropy is used to build a strain gauge in the form of a resistor bridge with four correctly oriented resistors. Because the resistor elements are not positioned at the same location, π_x and S will be slightly different for each resistor, giving rise to nonlinearities in sensitivity to both strain and temperature.

An alternative to the resistor bridge is the transverse voltage strain gauge (Xducer, which is a trademark of Motorola), shown in Figure 12.4. It is a single resistive element on a diaphragm, with voltage taps $S+$ and $S-$ centrally located on

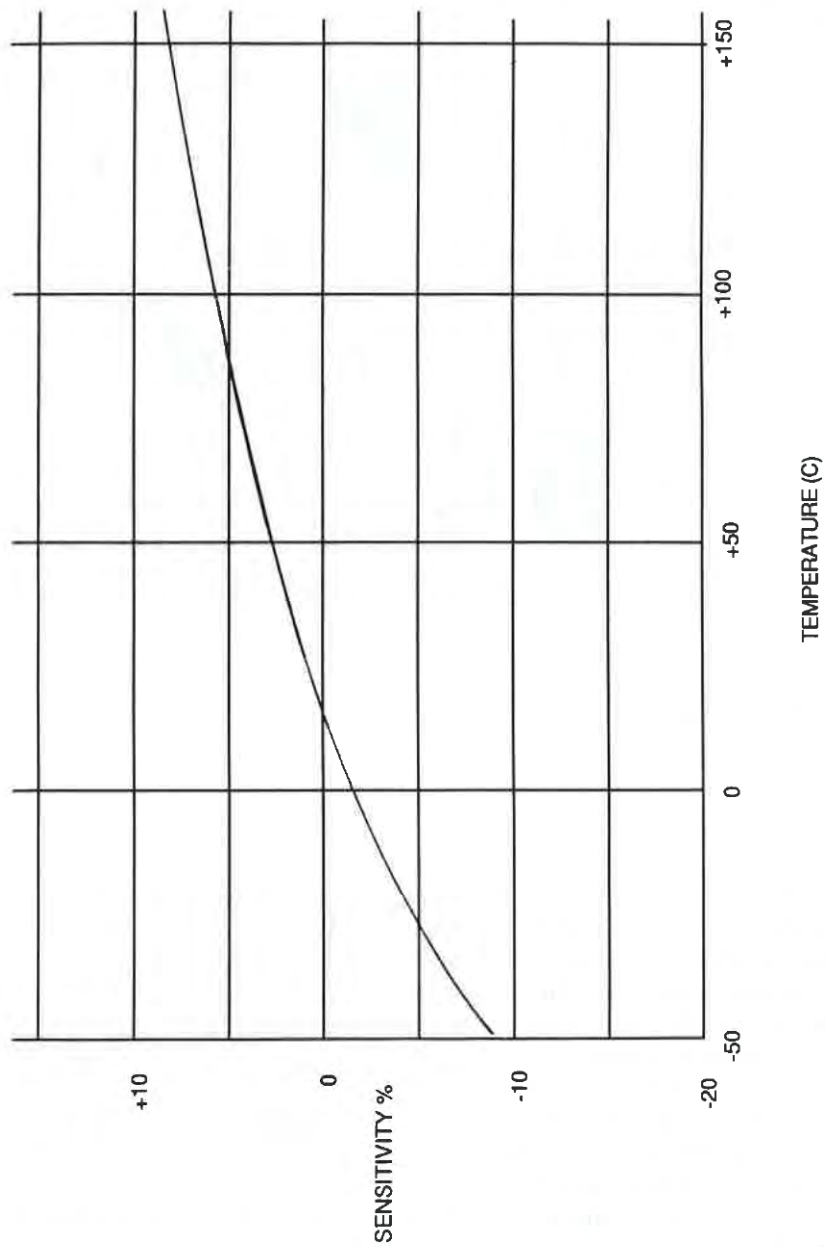


Figure 12.3 Sensitivity versus temperature in a piezoelectric device.

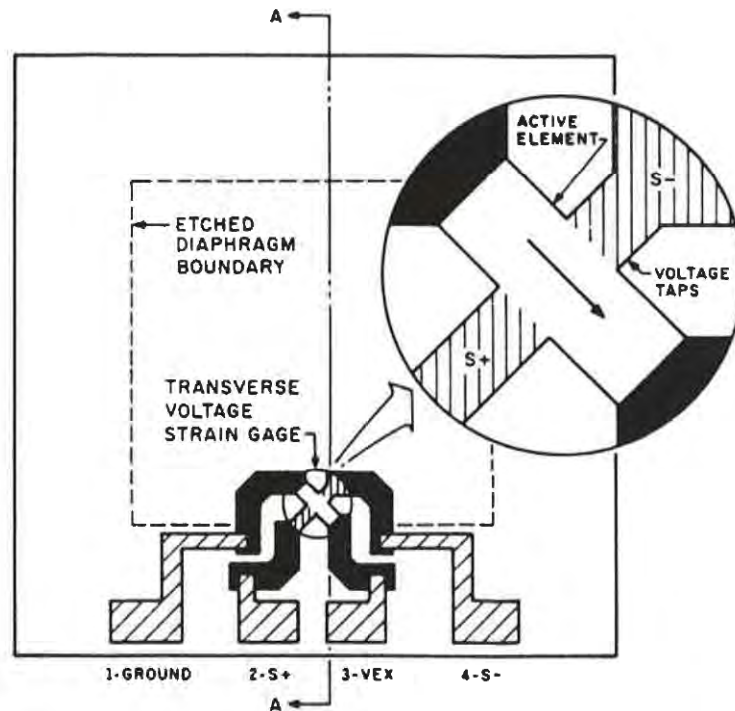


Figure 12.4 Transverse piezoresistive strain gauge.

either side of the resistor. When a current is passed through the resistor, the voltages at the taps are equal when the element is not under strain, but when the element is under strain, a differential voltage output appears. The element is oriented to rely on only one piezoresistive coefficient (π_{44} , which is the shear stress at 45° to the 100-crystal plane). This gives a linear output with strain, as well as improved temperature characteristics compared to the bridge structure. Trimming of the signal conditioning circuits is necessary to set the zero operating point, to set the gain of the system, and to offset temperature effects. The maximum operating temperature is limited to about 100°C to maintain good linearity and accuracy with temperature. This is because of junction leakage currents in the sensing elements and nonlinearity with temperature of the piezoresistive coefficient. Higher operating temperatures have been obtained using oxide-isolated strain gauges [4].

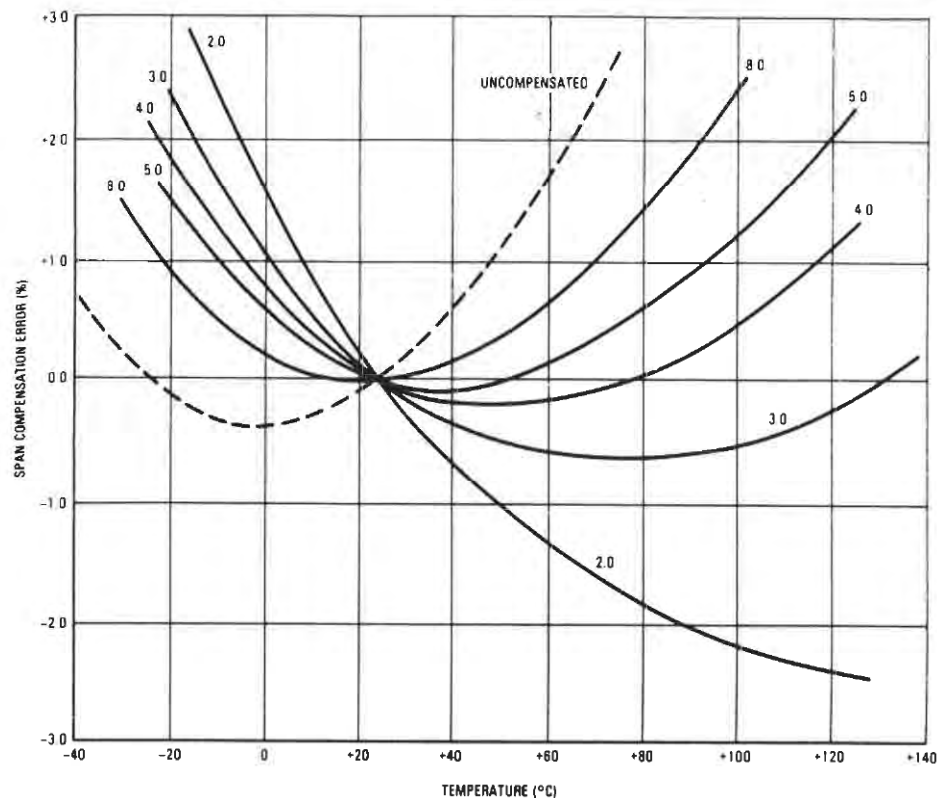


Figure 12.5 Sensitivity vs. temperature in a piezoresistive strain gauge.

Figure 12.5 shows the variation in sensitivity of an X-ducer piezoresistive strain gauge vs. temperature for an uncompensated device and for devices with different values of series resistors added for compensation [9]. The best compensation ratio of resistor value to X-ducer value is between 3 and 4, which reduces the error in the uncompensated strain gauge from about 3%, over the temperature range -20°C to $+80^{\circ}\text{C}$, to about 1.5%, over the temperature range -20°C to $+110^{\circ}\text{C}$. This error can be further reduced to about $\pm 0.5\%$, using a resistor and thermistor combination. A balanced series span compensation is shown in Figure 12.6. The balance resistor network is split so that the output from the X-ducer is at half the supply voltage, enabling direct connection to an amplifier, and the resistor thermistor values are chosen so that their temperature coefficient compensates that of the X-ducer.

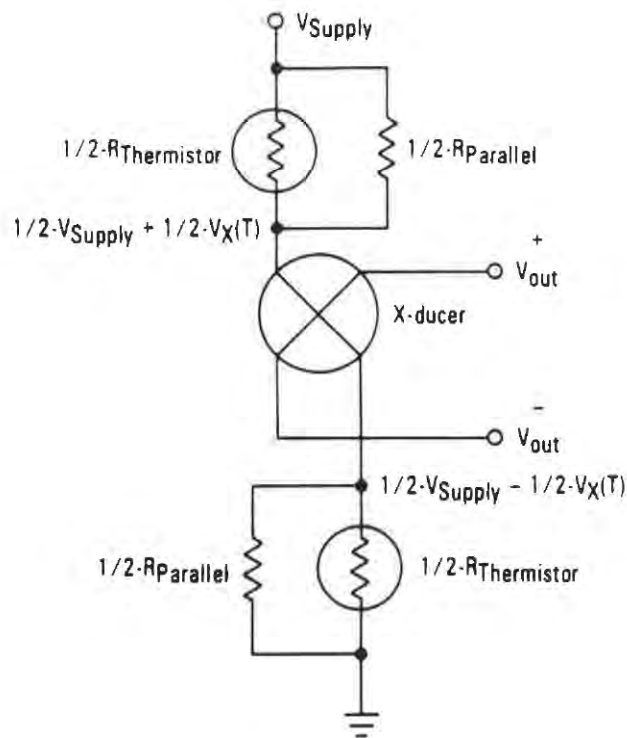


Figure 12.6 Piezoresistive sensor compensation.

12.2.3 Capacitive Sensing

Capacitive sensing has the following advantages over piezoresistive sensing: small size of elements, wider operating-temperature range, ease of trimming, good linearity, and compatibility to CMOS signal conditioning. Capacitive sensing [10] can be done in two different modes of operation, open-loop and closed-loop. Sensing in the open-loop configuration can be accomplished by single capacitance sensing, whereas the differential capacitor measurements can be used in both open- and closed-loop systems. When the differential capacitance sensing method is used to meet the sensitivity requirements, a charge redistribution technique [11] is normally used. Table 12.2 gives a comparison of piezoresistive and capacitive sensing.

Capacitance and capacitive changes can be measured either in a bridge circuit or using switched-capacitor techniques. It is possible to measure capacitance changes of 0.05 fF and lower, under ideal conditions, using switched capacitor techniques

Table 12.2
Comparison of Capacitive and Piezoresistive Sensing
Characteristics and Requirements

	<i>Capacitive</i>	<i>Piezoresistive</i>
Trimming	Zero and span	Complex
Temperature compensation	Not required	Required
Temperature range	Wide	Limited
Signal condition	Easy	More complex
Linearity	Good	Poor with temperature
Sensitivity	High	Low
Self-test	Yes	With capacitor

[11]. However, in an industrial environment the lower limit becomes 0.5 fF. A system such as an accelerometer or pressure sensor with a resolution of 1% will, therefore, have a full-scale capacitive change of 50 fF. If the measurement range of the structure is limited to 10% of its allowable movement (for linearity reasons), the actual capacitance could be 500 fF, or 250 fF + 250 fF using differential sensing. Small air gap spacings between two plates on the order of 2 μm or less can be obtained by using surface micromachining, from which a capacitor or plate size of $336 \mu\text{m} \times 336 \mu\text{m}$ (or $237 \mu\text{m} \times 237 \mu\text{m}$ with differential sensing) can be obtained with air as the dielectric.

Any of the capacitive sensing techniques used in a micromachined structure require an ac voltage across the capacitor being measured. This in turn produces an electrostatic field, which generates an attractive force between the capacitor plates, as shown in the single-sided capacitive cantilever structure in Figure 12.7(a). Other forces can be generated by the sensing-current flow, which will produce an electromagnetic field, or by system noise. In the case of system noise, the average noise voltage generated is 0; however, the electrostatic force is proportional to the square of the noise voltage, which is not 0. Using single-capacitor sensing, the sensor capacitance can be compared to a fixed capacitor. The electrostatic field produced by the sensing waveform can be substantial when small capacitor spacings are used. In the case of an accelerometer, this may cause both a deflection of the seismic mass and a zero offset. If the voltage amplitude of the driving waveform is too large, the electrostatic forces can pull and lock the mass down to the substrate, rendering the device inoperable.

Differential sensing, using a movable plate between two fixed plates, as shown in Figure 12.7(b), can be operated in either an open-loop or a closed-loop signal conditioning configuration. In both cases, the electrostatic forces from the two non-movable top and bottom plates on the movable middle plate will cancel, when the middle plate is in a central position. This greatly reduces any tendency for latch up,

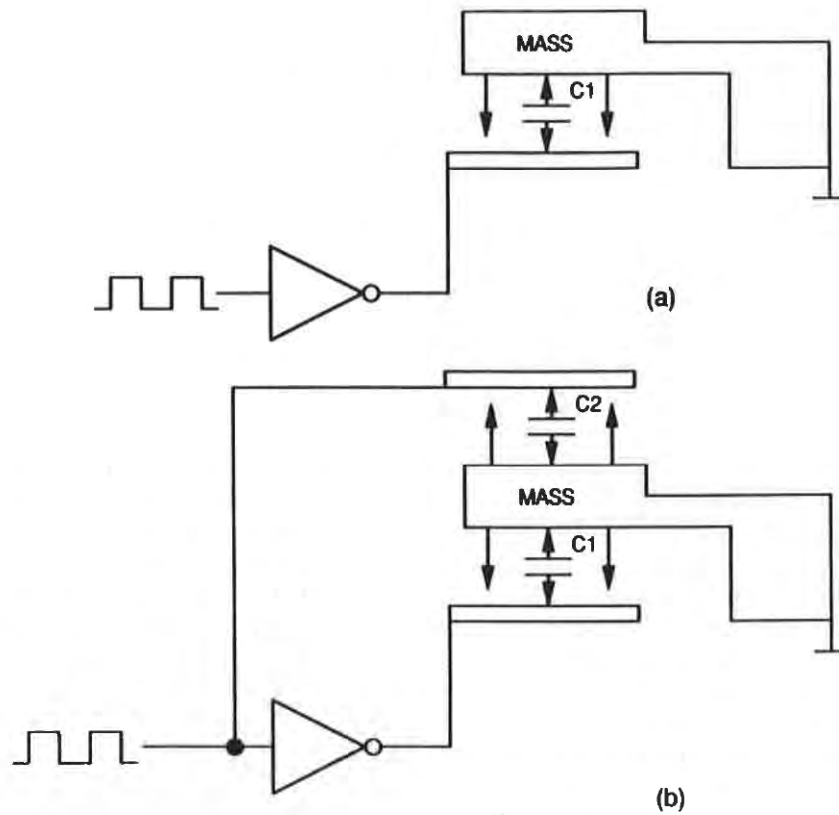


Figure 12.7 Capacitive sensing with electrostatic forces: (a) single-sided capacitive cantilever structure; (b) a movable plate between two fixed plates.

which exists in the open-loop configuration. Also, any electrostatic forces created by noise will tend to cancel. Differential sensing also greatly improves the system linearity and minimizes temperature effects.

Figure 12.8 shows the signal conditioning circuit [12] and trimming scheme used with single-ended capacitive sensing. The first stage of the CMOS signal conditioning circuit uses switched-capacitor filter techniques for converting the capacitance change of the transducer C_1 into a voltage. Initially, switch S_1 is at ground potential and switches S_2 , S_3 , S_4 are closed. Each integrator has unity gain so that each of the negative inputs will be at ground potential. Switches S_2 , S_3 , S_4 are then opened, and S_1 is connected to V_{REF} . The output of the integrators is proportional

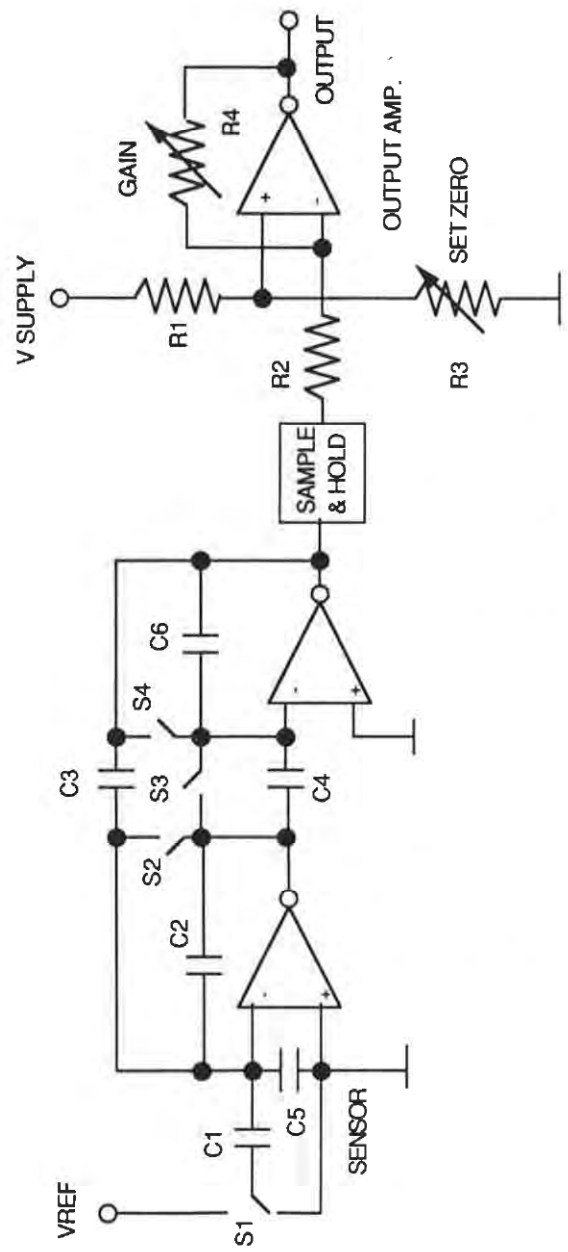


Figure 12.8 Single capacitor sensing block diagram.

to the reciprocal of the difference of the feedback capacitors $C2$ and $C3$, and the output voltage is given by

$$V_{\text{out}} = V_{\text{ref}} \frac{C1}{(C2 - C3)} \quad (12.2)$$

In this converter the input voltage is also maintained at zero potential by feedback via capacitor $C2$, so that the effect of stray or ground capacitance $C5$ is eliminated. In this case, because the input or sensing capacitor is switched, saturation of the amplifier or recovery time is no problem. The output voltage of the converter V_{out} , is fed to the output stage via a sample-and-hold circuit, where the signal is amplified and trimmed for zero offset by R_3 , and the gain is set by R_4 .

In the closed-loop system, the seismic plate is maintained in its central position by electrostatic forces, so that only minimal movement occurs during acceleration. This is achieved by monitoring the capacitance and applying an appropriate voltage to the top or bottom capacitor plate to produce sufficient electrostatic force on the central plate (seismic mass) to balance any acceleration forces on the seismic mass. The servo system feedback is shown in the block diagram in Figure 12.9, which contains the mechanical, electrical, and electrostatic blocks. When the gain of the electrical block is high, the transfer function can be modified, so that the output (V_{out}) is both proportional to the acceleration and independent of the mechanical parameters, except for the mass

$$V_{\text{out}} = \frac{m \times \text{Acc} \times g \times d}{C \times V^2} \quad (12.3)$$

where m = the mass; g = gravitational constant; C = plate capacitance; Acc = acceleration; d = plate separation; and V = plate voltage.

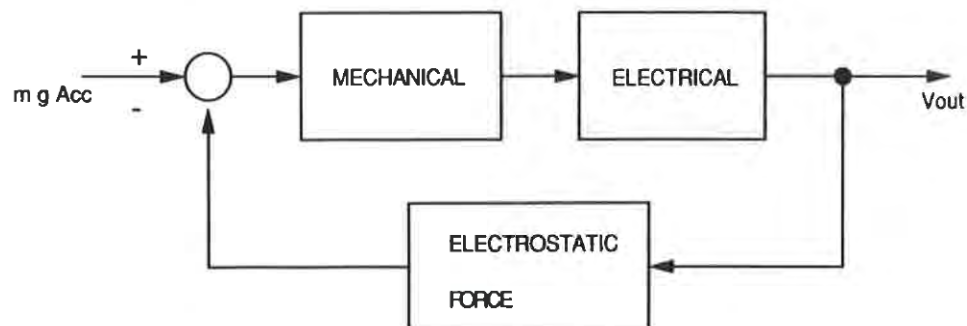


Figure 12.9 Servo loop block diagram.

The closed-loop technique has the following advantages over open-loop sensing: minimized fatigue in the suspension arms, a wide dynamic operating range, a wide temperature range, improved linearity, a sensitivity that depends only on the mass, a reduced bandwidth dependence on film damping. With proper system design, all of the mechanical parameters and variables, except the mass, are reduced to second-order terms, thus both greatly improving the performance of the system and making the sensitivity, bandwidth, and any temperature effects dependent upon the signal-conditioning electronics.

12.3 OPEN- AND CLOSED-LOOP SYSTEMS

In the suspended-mass type of accelerometer, the sensitivity and the bandwidth in the open-loop system depend upon a number of process variables: the width of the suspension arms, the thickness of the suspension arms, Young's modulus, the weight of the seismic mass (density, thickness, dimensions), the spacing between plates, and the damping (slots, gas, pressure). As can be seen, these parameters can give wide spreads in zero offset and sensitivity and in temperature drift. However, in the closed-loop system the process variables become second-order terms, so that in this system, sensitivity, bandwidth (damping), and drift with temperature are minimized.

12.3.1 Different Micromachined Structures—Mechanical Issues

To describe open-loop and closed-loop systems we will use an accelerometer as an example. A number of shapes and configurations for micromachined accelerometers is possible. However, in considering the alternatives, the following requirements must be taken into consideration: sensitivity, stability, material fatigue, resistance to shock, damping, cross-axis sensitivity, linearity, operating temperature range, operating frequency range, and reproducibility.

Three basic methods of suspension are shown in Figure 12.10. These structures have the ability to be designed to meet the preceding requirements. Shown are the cantilever, Figure 12.10(a); the torsion-bar suspension, Figure 12.10(b); and a seismic mass suspended by four cantilevers from a central pillar, Figure 12.10(c). The simple cantilever is the structure with the highest sensitivity for a given size of suspension arm, and thus it can be made smaller for a given sensitivity than any of the other configurations. This is shown in (12.4) and (12.5), which are the acceleration-sensitivity equations for the two-arm cantilever and the four-arm suspension configurations shown in Figure 12.10(a) and (c) [13], respectively. The deflections Δ_{ta} and Δ_{fa} of the mass in each case is given by

$$\Delta_{ta} = \frac{4 \times m \times \text{Acc} \times g \times L^3}{Y \times 2 \times a \times b^3} \quad (12.4)$$

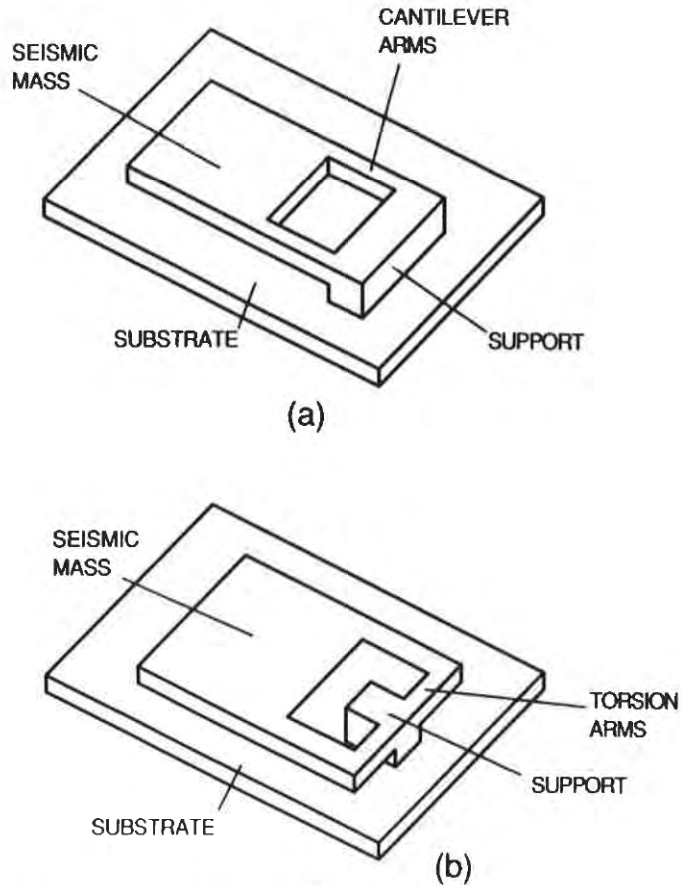
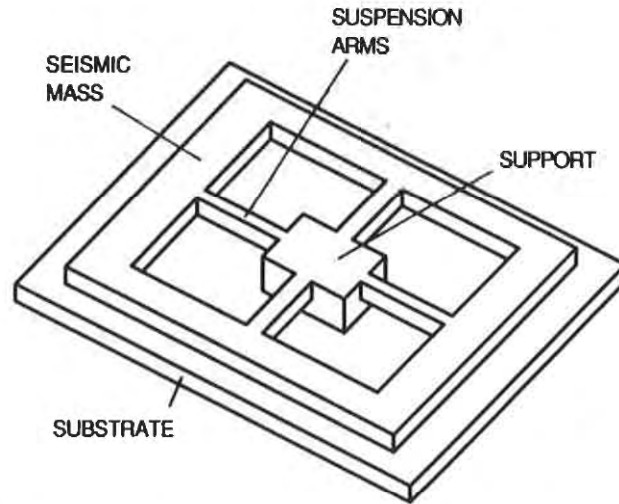


Figure 12.10 Various accelerometer configurations: (a) the cantilever; (b) the torsion-bar suspension; and (c) a seismic mass suspended by four cantilevers from a central pillar.

$$\Delta_{ia} = \frac{m \times \text{Acc} \times g \times L^3}{Y \times 4 \times a \times b^3} \quad (12.5)$$

where Δ_{ia} = two-arm deflection of the mass; Δ_{fa} = four-arm deflection of the mass; L = length of arms; Y = Young's modulus; a = width of arms; and b = thickness of arms. For equal mass and arm size, the cantilever structure is eight times more sensitive than the four-arm structure. In torsion-bar suspension, Figure 12.10(b) [13], the displacement Δ_{ib} of the mass is given by



(c)

Figure 12.10 continued.

$$\Delta_{tb} = \frac{m \times \text{Acc} \times g \times 2 \times L^2}{G \times ab(a^2 + b^2)} \quad (12.6)$$

where Δ_{tb} = torsion-bar displacement of the mass and G = shear modulus. It is difficult to get a comparison of the sensitivity of the torsion-bar suspension versus the cantilever suspension, but in practical devices the size of the torsion-bar devices is usually found to be larger than the cantilever devices.

Two suspension arms are sometimes used, as shown in Figure 12.10(a), to improve the cross-axis sensitivity of the device. The cross-axis sensitivity is defined as the ratio of the output from transverse acceleration to the normal acceleration, and it is typically required to be less than 3%. It can be seen from the preceding equation that, to increase the sensitivity in a structure, the width and thickness of the suspension arms must be minimized, the length increased, the capacitor spacing minimized, and the mass increased. In a flat structure, the long suspension arms required for good sensitivity can be bent, as shown in Figure 12.11, to reduce the overall size of the structure, but the suspension-arm equation must be modified by a torsional term. Under adverse conditions the deflection can become excessive, and the multiple-arm structure can be driven close to one of the fixed plates, in which case the electrostatic force due to the drive voltage becomes very large, as is shown in Figure 12.12. If open-loop sensing is used with this type of structure, latch-up

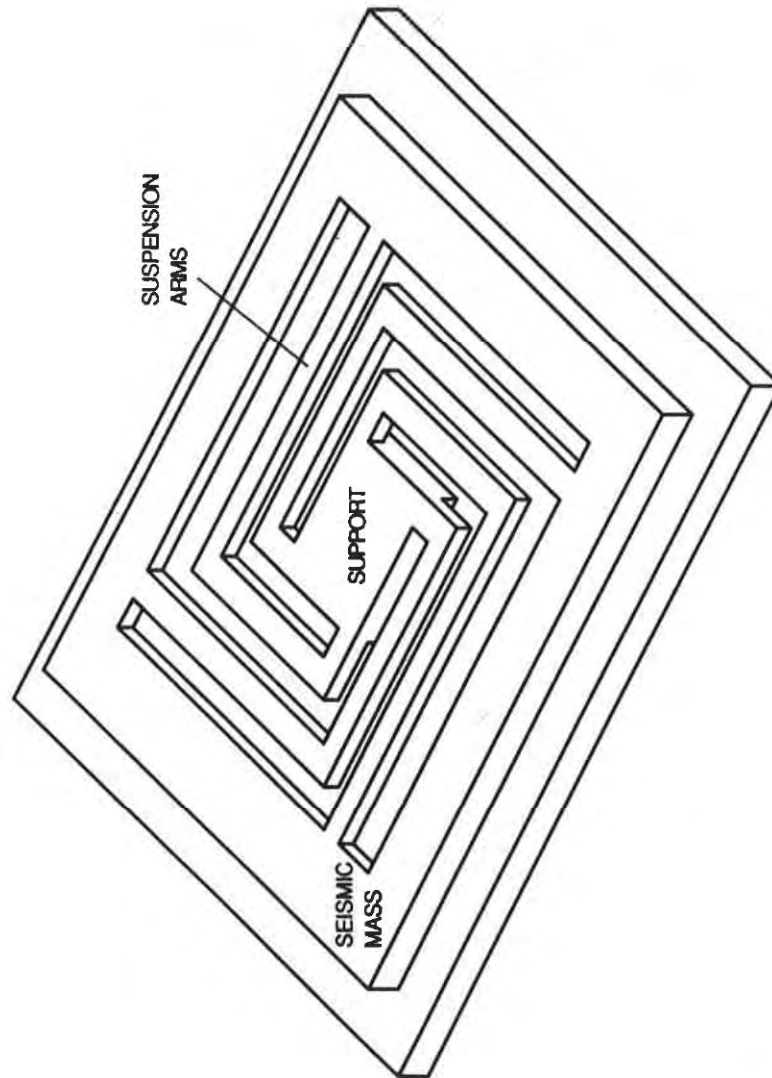


Figure 12.11 Flat accelerometer with curved arms.

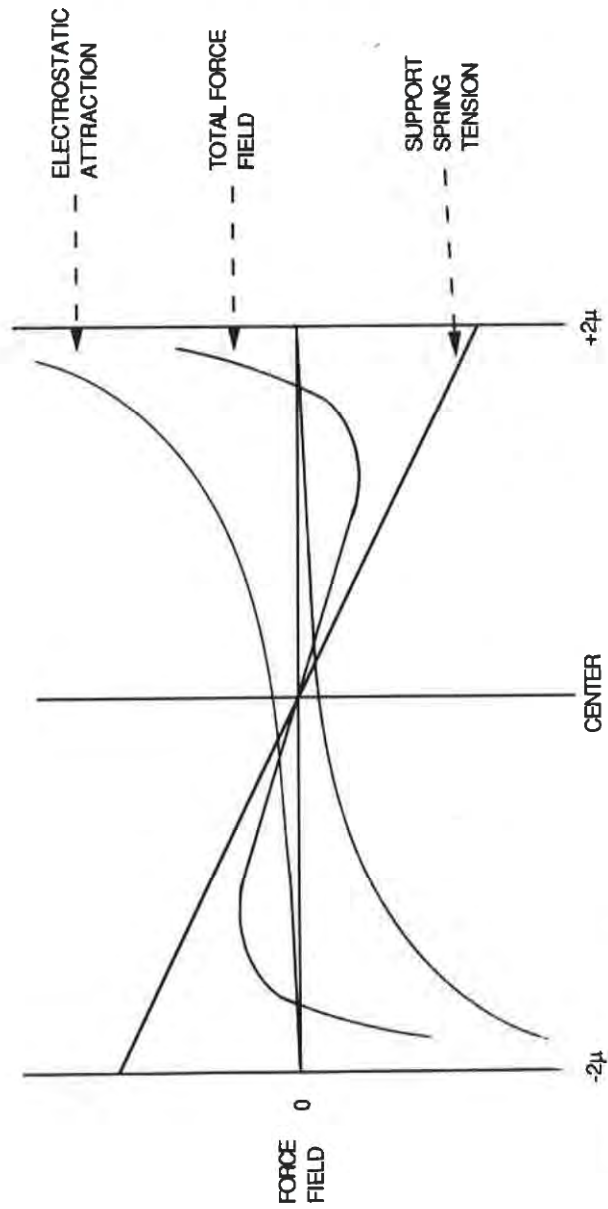


Figure 12.12 Electrostatic and mechanical forces.

can occur unless the deflection is limited by the use of end stops. Figure 12.12 shows the plate displacement vs. spring tension and the electrostatic fields encountered with capacitor spacings of $2\text{ }\mu\text{m}$. As can be seen, for good linearity the center plate must move less than 10% of the possible deflection for full-scale output. In both the simple-cantilever and the torsion-bar suspension devices, the deflection should also be limited to 10% for good linearity. The moving plate is angled down, and end stops can be used to prevent further movement of the seismic mass, thus limiting the minimum plate separation and hence the electrostatic forces. In this way, latch-up can be prevented with proper design and drive voltage.

12.3.2 Open-Loop Capacitive Sensing

The CMOS signal-conditioning circuit for open-loop differential sensing and the cross section of a capacitive torsion-bar sensor are shown in Figure 12.13 and 12.14, respectively. CMOS has a number of advantages over bipolar, when sensing small capacitance changes using switched-capacitor techniques: very high input impedance, good switch characteristics, low power requirements, and small size.

The output voltage from the converter section in Figure 12.13 is given by

$$V_{\text{out}} = \frac{V_{\text{ref}}(C1 - C2)}{C3} \quad (12.7)$$

which gives an output voltage proportional to the differential capacitance of $C1$ and $C2$. Here, again, the input is referenced to ground by feedback capacitor $C3$, so that

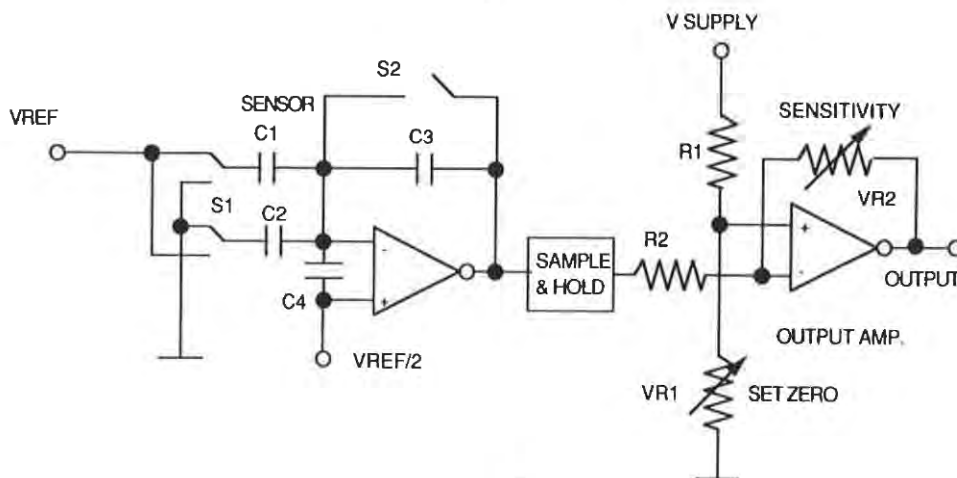


Figure 12.13 Open-loop differential sensor block diagram.

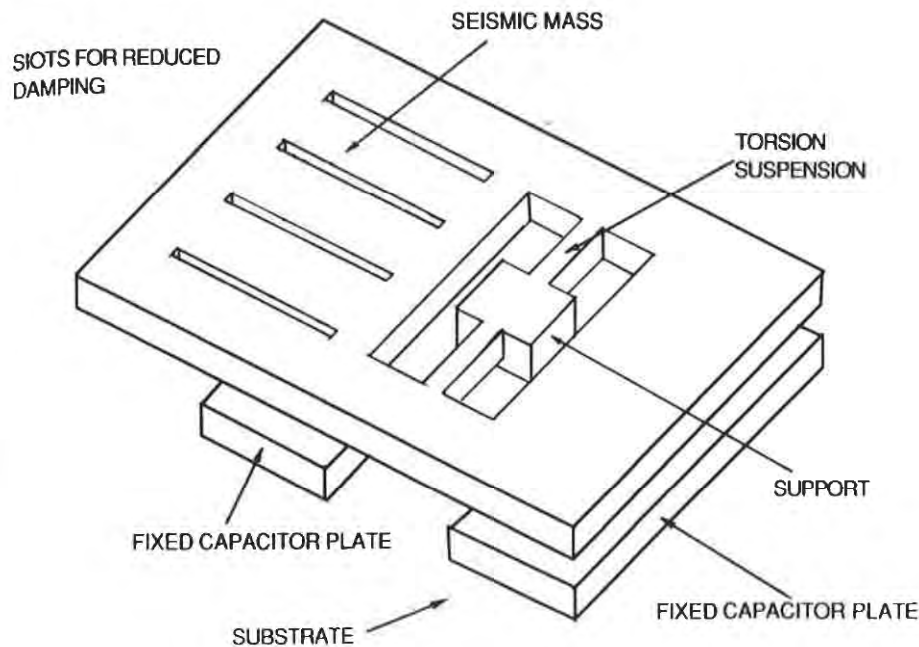


Figure 12.14 Differential capacitive accelerometer.

the effects of stray capacitance are eliminated. The first-stage output is fed via a sample-and-hold circuit to the output stage, where the signal can be amplified to give a voltage or a current output, depending upon the system requirements. The output can also be ratiometric (proportional to the supply voltage), which is sometimes required when the analog output is digitized by an external A/D converter. If single-capacitor sensing is required, the circuit shown in Figure 12.8 can be used. To offset processing tolerances of the sensing capacitors, as well as variations in system sensitivity, both zero and gain trimming are required; and this is normally performed in the output stage. However, an alternative is to individually adjust the amplitude of the V_{ref} , which is the driving voltage to $C1$ and $C2$. If the circuit is required to operate over a very wide temperature range and with an accuracy of no better than 2%, no temperature adjustment is required to correct for changes in the physical properties of the material. Pressure changes can cause changes in the bandwidth of the device, unless it is hermetically sealed.

The driving waveforms for the sensor and the signal conditioning circuits are shown in Figure 12.15. The clock, which typically has a frequency of 250 kHz, is divided into four phases, which generate three nonoverlapping phases. These perform the following functions.

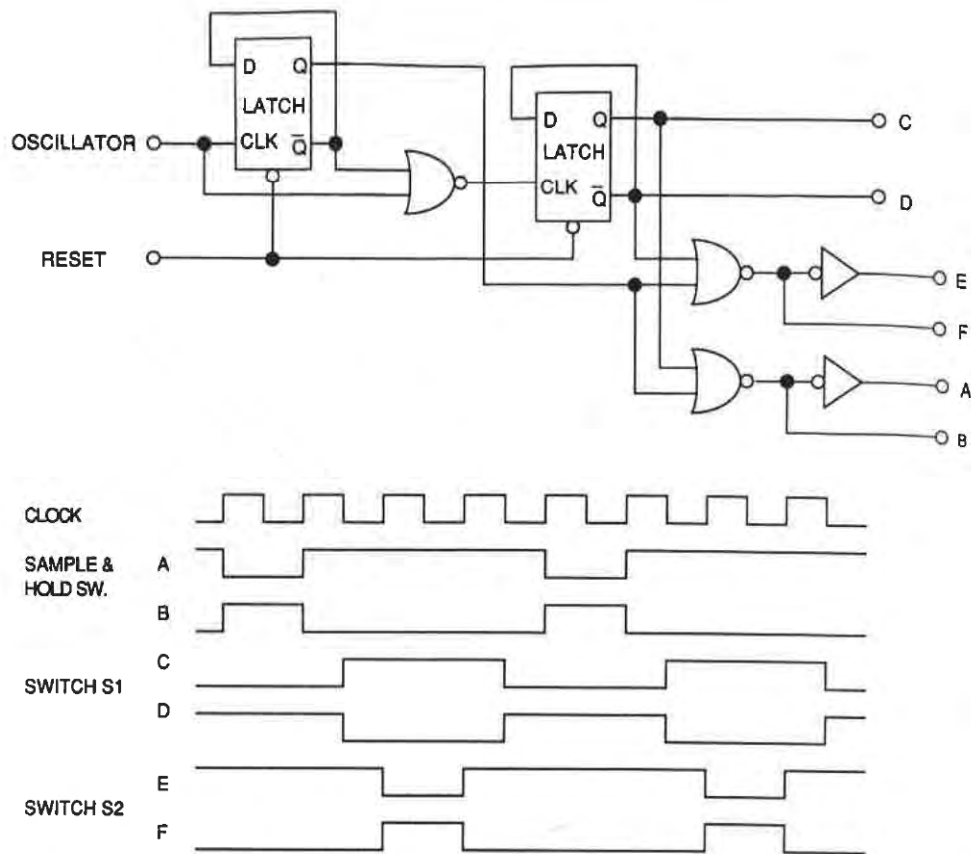


Figure 12.15 Open-loop differential sensing waveforms.

- **Reset integrator:** sets the input voltage of the integrator to $V_{\text{ref}}/2$ by shorting the input to the output.
- **Sensor drive:** drives the fixed top and bottom capacitor plates in the sensor. The integrator circuit converts the difference in capacitance into a voltage.
- **Sample and hold:** samples the output of the integrator and stores the voltage on a capacitor to supply dc voltage to the output stage.

12.3.3 Closed-Loop Capacitive Sensing

In a closed-loop system a force due to acceleration can be sensed either with piezoresistors or with capacitive techniques. However, the capacitive sensing has advantages and we will focus on this technique. Figure 12.16 shows the block diagram of a

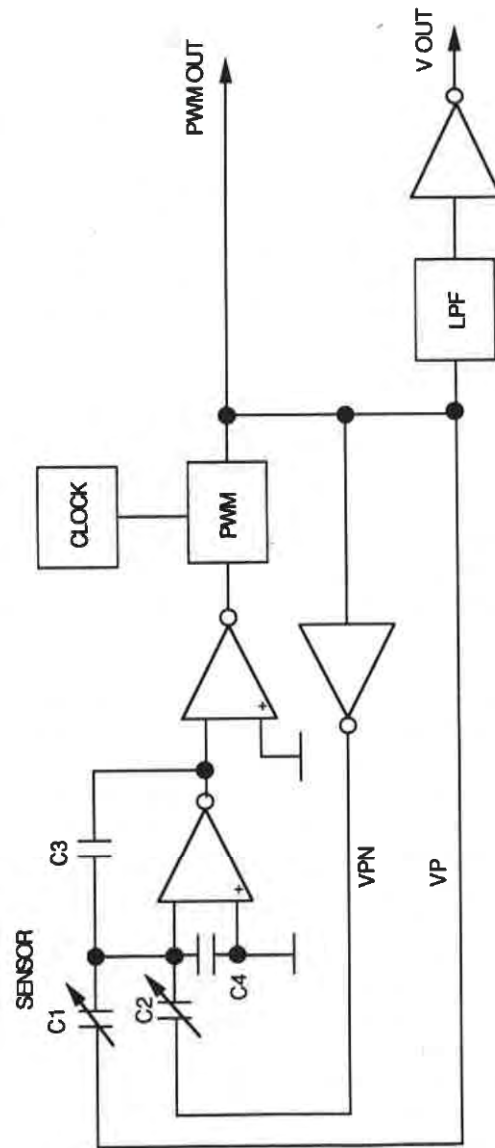


Figure 12.16 PWM block diagram.

pulsewidth modulator [14] (PWM) for use in a closed-loop system. The circuit waveforms are shown in Figure 12.17. The difference in value between the two sensing capacitors is converted into a voltage using the same technique as in the open-loop case. The output signal from the converter is amplified and converted into a PWM signal (VP). This signal is inverted (VPN) and the true and inverted signals used to drive the top and bottom capacitor plates. With no acceleration, the output from the PWM circuit has a 50% duty cycle, so that with the seismic mass referenced to ground potential, the electrostatic forces on the center plate are equal and opposite. When an external force is applied, a change in capacitance is sensed, and the duty cycle of the driving waveform changes, increasing the duty cycle on one plate and reducing it on the other, as shown by the waveforms. The net effect is to produce an electrostatic force to counterbalance the externally applied force on the center plate. The closed-loop transfer function of the PWM system with a high-gain amplifier is given by

$$\text{Duty cycle} = \frac{\text{Acc} \times g \times d}{C \times V^2} \quad (12.8)$$

That is, the duty cycle is proportional to the acceleration. The output from the PWM section is also fed to the output via a low-pass filter, which will convert the PWM signal into an analog signal. The output stage as shown is normally trimmed for offset and gain.

Another example of closed-loop operation is the delta modulator [15], the circuit for which is shown in Figure 12.18, with the waveforms shown in Figure 12.19. This method uses digital techniques. A reference voltage is switched across the plates of capacitors C1 and C2, and when an external force is applied, the resulting charge on the center plate caused by a difference in the capacitance of C1 and C2 is converted into a voltage, as in the circuit shown in Figure 12.13. This voltage is then clocked into a latch, where it sets up a 1 or 0 depending upon the sign of the sensed capacitor charge. The output from the latch is then used to apply a voltage to the appropriate capacitor plate to force the center plate back to its zero position. With no external forces applied, the center plate vibrates about its center position. Clocking for this type of system is nominally about 2 MHz. The datastream from the latch also goes to a decimation circuit, which consists of a low-pass filter (second-order Butterworth) that is used to convert the 2MHz, 1-bit serial signal into an 8-bit word that can be clocked onto an 8-bit MPU bus at an 8 kHz clock rate. Alternatively, a low-pass filter can be used to convert the data into an analog output. The waveforms shown in Figure 12.19 compare the displacement of the seismic mass in an open-loop system and a closed-loop system. The movement with the closed-loop system is minimal. Also shown is the modulated data output from the delta modulator. Trimming the zero offset and sensitivity is required.

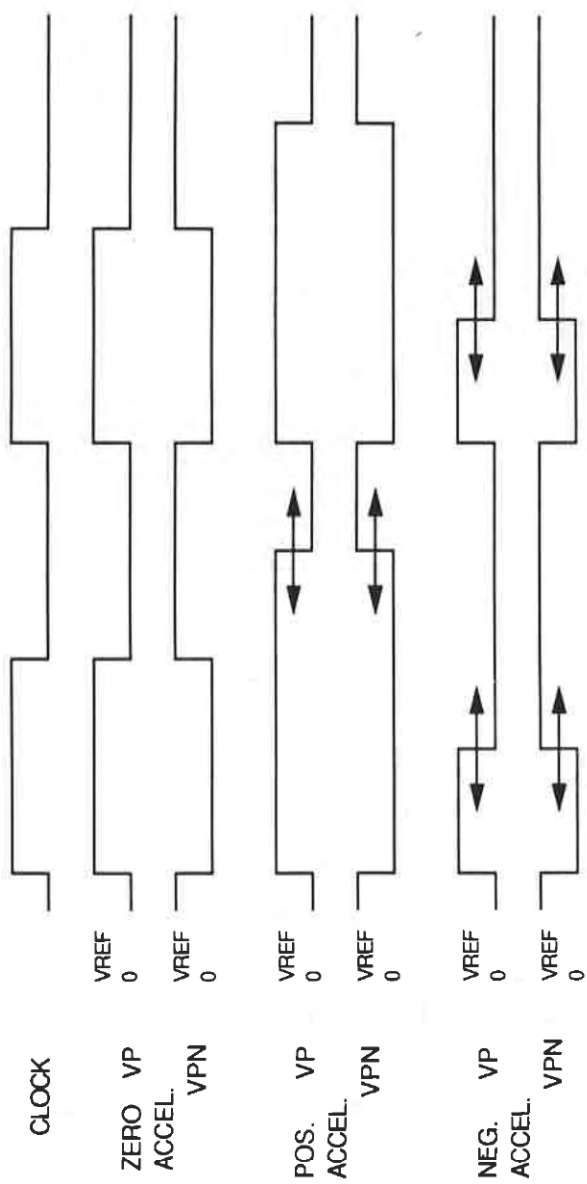


Figure 12.17 PWM waveforms.

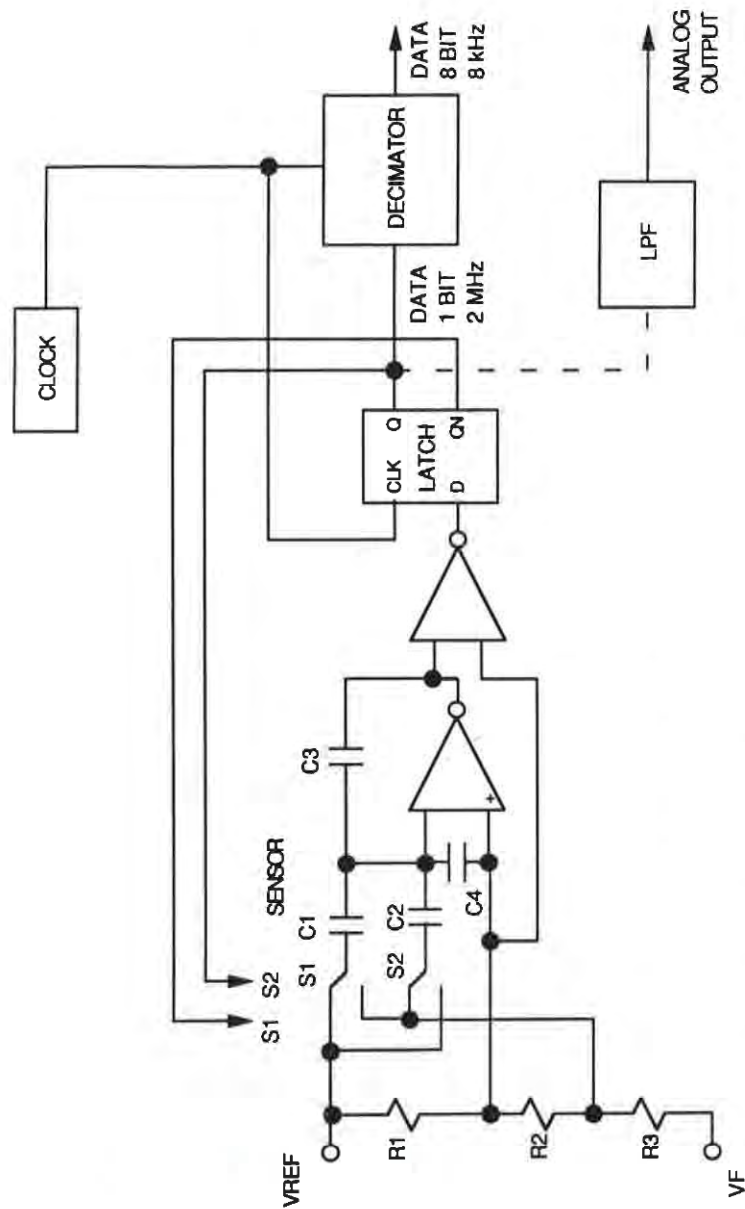


Figure 12.18 Delta modulator block diagram.

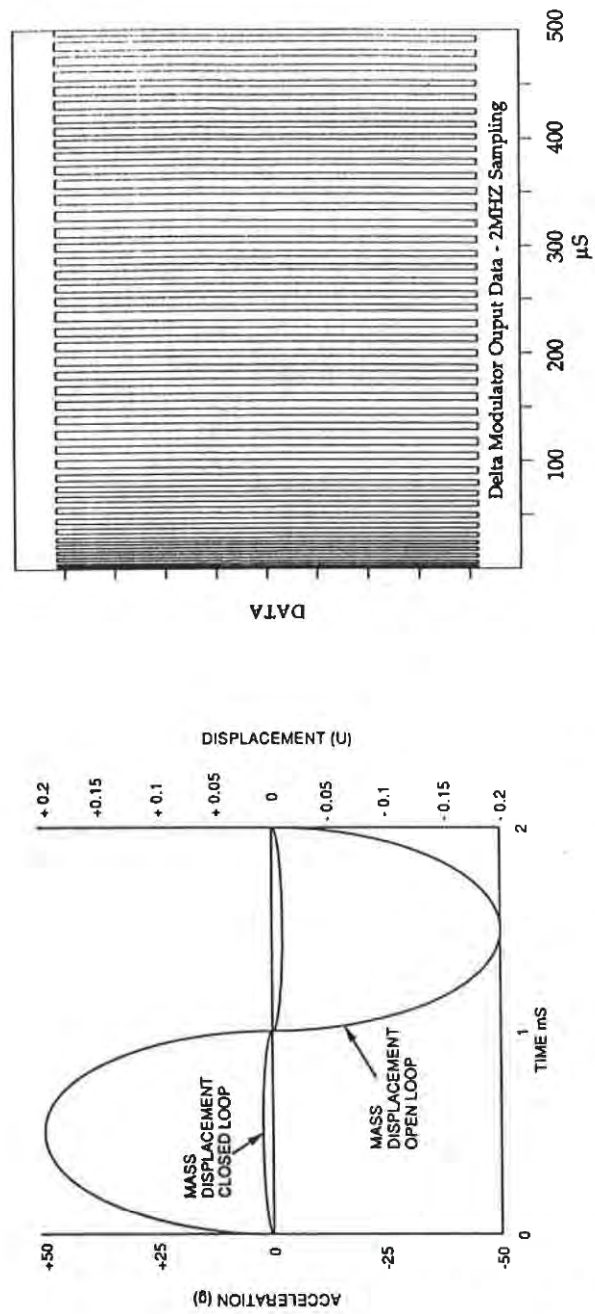


Figure 12.19 Delta modulator waveforms.

Closed-loop capacitive sensing can be used at high temperatures ($>150^{\circ}\text{C}$), and, because the capacitors are dielectrically isolated, temperature correction is not required.

12.4 INTEGRATION

For monolithic integration of sensors and microcircuits, the processing must have compatible steps, and the integrated device has to be cost effective. Any processing steps specific for one device say, a sensor circuit, must not adversely affect the other. The starting material must be compatible to both. Integration usually provides better temperature sensing and control, reduces the silicon area, and grounds capacitances by eliminating the chip-to-chip bonding pads. Other advantages are better noise immunity, reduced lead inductance, and better reliability.

12.4.1 Integrated Pressure Sensor

Figure 12.20 shows a three-stage bipolar amplifier and its trimming network used with the MPX5100D integrated X-ducer strain gauge. The trimming network compensates for the characteristics of the strain gauge. The bipolar amplifier is trimmed using thin-film resistors. The span is compensated by resistor R_s , Figure 12.6. The temperature and offset calibration circuit uses two operational amplifiers, OA1 and OA2. Resistors $R1$ and $R2$ are used to set the operating point of the amplifiers to match that of the X-ducer. R_o is then set to match the shift in output voltage of the X-ducer over temperature. Amplifier OA3 is the main gain stage. The zero offset is adjusted by $R8$ and $R9$ and the gain or sensitivity is set by R_g . The sequence of trimming is critical to prevent interaction between the adjustments. Using this approach accuracies of $\pm 1\%$ over the temperature range -40°C to $+100^{\circ}\text{C}$ can readily be achieved.

12.4.2 Integrated Accelerometer

An integrated accelerometer was described in [5]. The device uses CMOS signal conditioning circuits and capacitive micromachined sensing structures of the type shown in Figure 12.10(b). The conditioning circuits use switched-capacitor sensing techniques in an open-loop configuration (Figure 12.13) and can operate over the temperature range -55 to $+125^{\circ}\text{C}$. The output stage is a voltage-to-frequency converter, so that a frequency output is obtained proportional to the sensed acceleration. Trimming is done off-chip.

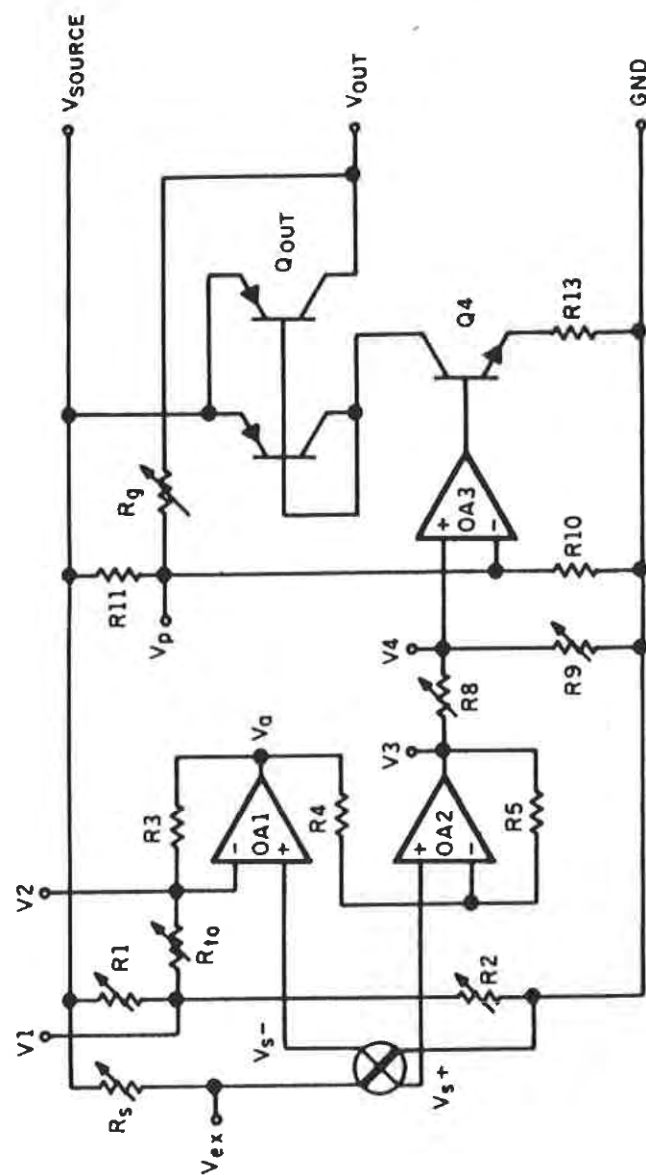


Figure 12.20 Integrated X-ducer sense amplifier.

12.4.3 Dynamic Considerations

No section on sensing and control circuits for micromachined structures would be complete without looking at dynamic characteristics. For instance, in the accelerometer air-film damping is desirable to reduce the effect of shock under high-g forces [16,17], but reduces the frequency response of the system. In the case of the cantilever device, the natural resonant frequency (f_0) of the undamped accelerometer is given by [18]

$$f_0 = \frac{(Yab^3)^{1/2}}{2\pi(4L^3m)^{1/2}} \quad (12.9)$$

A typical f_0 for the undamped cantilever shown in Figure 12.10(a) would be about 5 kHz, thus giving a high peak output at resonance. Such a structure, when critically damped [19–21] (the damping constant ϑ is 0.7), will give a response that is flat to about 2 kHz. Typical response curves for both an undamped and a critically damped cantilevers are shown in Figure 12.21. The damping depends upon the area of the seismic mass, the thickness, and the pressure and type of gas in the air film. In

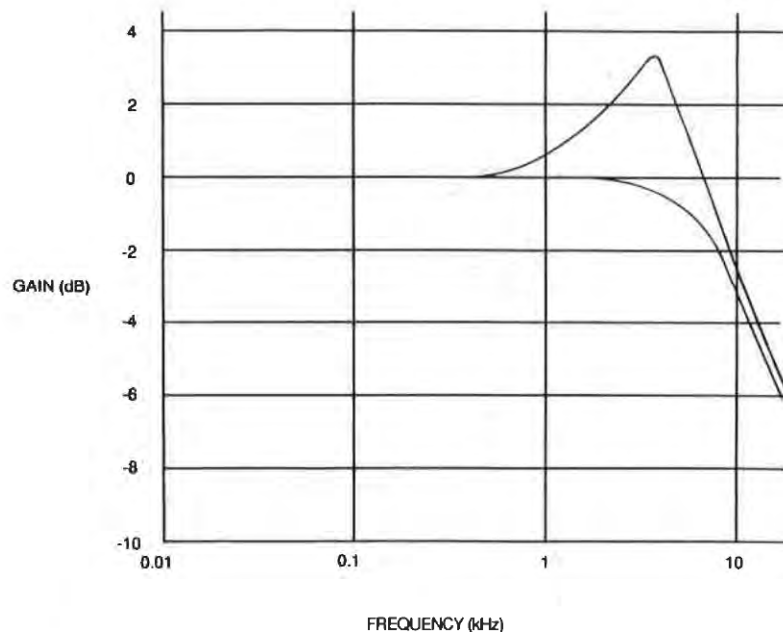


Figure 12.21 Cantilever response, undamped and critically damped.

applications where a high-frequency response is required, damping can be reduced by increasing the thickness of the air gap. On the other hand, this is undesirable, because to maintain the value of the sensing capacitance, the size of the device must be increased. The pressure of the gas in the accelerometer can be reduced, but this necessitates sealing the device. Also, holes can be put into the moving plate to reduce the effect of the air damping. Using this approach the frequency (f) versus gain is given by [19]

$$\text{Gain} = \text{Force} / \{(f_0^2 - f^2)^2 + 4\delta^2 f^2\}^{1/2} \quad (12.10)$$

where δ is the damping constant. With the cantilever mass as a solid plate and a $2 \mu\text{m}$ air gap, the damping at atmospheric pressure gives the ratio [13,16] $\delta/f_0 = 16$. If slits are used in the plate to reduce the effect of film damping, as shown in Figure 12.14, the ratio, $\delta/f_0 = 7.5, 3$, and 1 for one, three, and five slits, respectively. If these numbers are put into (12.10) and plotted, a set of response curves similar to those shown in Figure 12.22 will be obtained for frequency vs. gain. As can be seen with a solid seismic mass the frequency response is flat to about 60 Hz, increasing to 700 Hz with five slits.

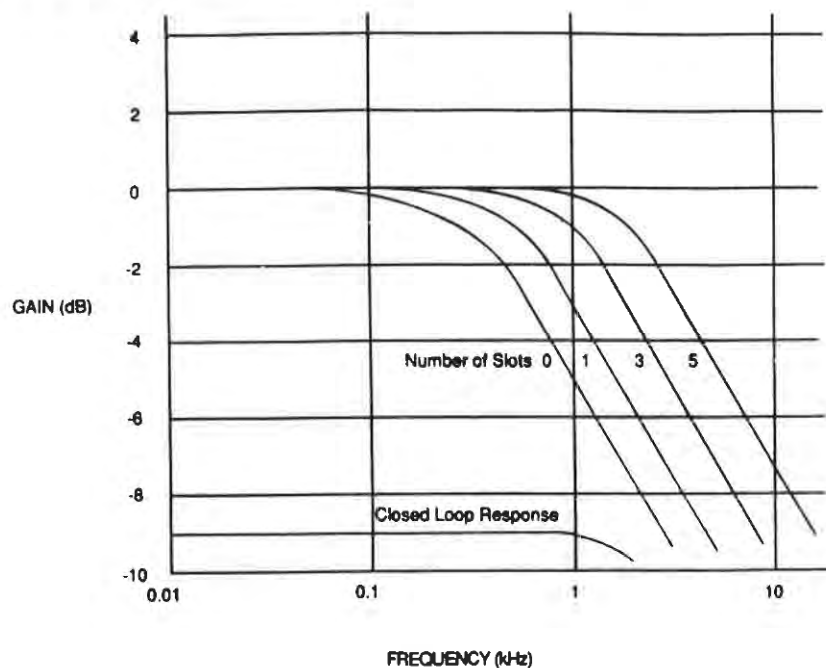


Figure 12.22 Cantilever frequency response, with and without slots.

When considering the dynamic response, the circuit configuration needs also to be taken into account. The curves plotted in Figures 12.21 and 12.22 apply to open-loop systems. In a closed-loop system, where the seismic mass is maintained in its central position, the film-damping effects are greatly reduced and will become of only second order, as shown in [14]. This enables the frequency response to be controlled by the signal-conditioning circuits. The effect is also shown in Figure 12.22, where the closed-loop curve shows the reduced gain required from the mechanical transducer, along with the corresponding increase in frequency response. For a solid plate under atmospheric pressure, the frequency response is increased from 60 Hz to 1 kHz, for the closed-loop configuration.

12.4.4 Shock Considerations

To ensure survivability a system must be designed to tolerate shock. A common criterion for micromachined devices is the ability to survive when dropped onto a concrete floor from a height of 3 ft, which is equivalent to a shock of 5,000g. The problem of survivability becomes most difficult in devices with high sensitivity, because of the increased mass and minimal strength in the suspension arms. The acceleration waveform due to shock [17] is shown in Figure 12.23, and although very

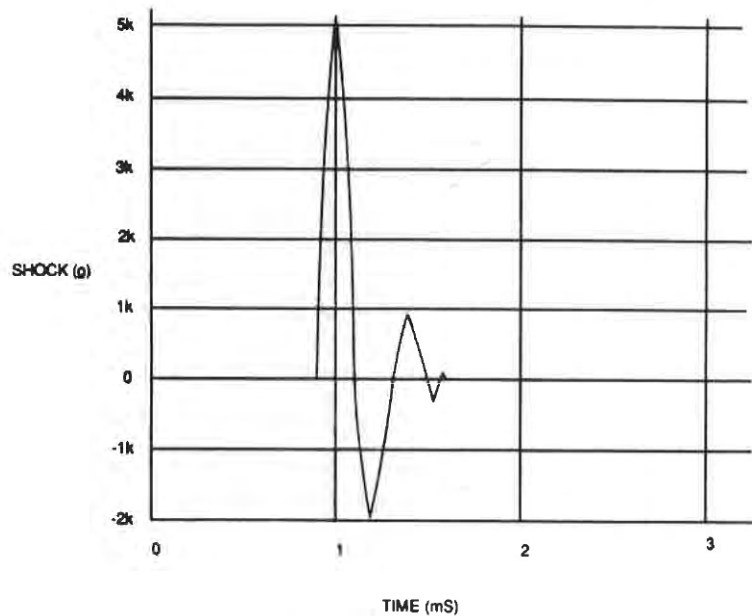


Figure 12.23 Shock waveform.

high amounts of g are involved, their duration is short (0.1 ms), so that film damping can hold the seismic mass in its central position.

An accelerometer can be considered as having three major axis of movement. These are shown in Figure 12.24. The device is most sensitive in the Z -direction by design, and in this direction film damping is used for protection. It should be noted that because of the higher damping factor that can be used with closed-loop operation, a high tolerance to shock is obtained. In the Y -direction there is no film damping, but the device is very rugged, because of the high tensile strength of silicon. The movement depends upon the size of the mass and on the width and thickness of the suspension arms. In this direction an accelerometer is normally capable of withstanding a shock of several thousand g .

In the X -direction the two parallel cantilever beams as shown in Figure 12.24, are used to give good shock tolerance, which will be much higher than the tolerance of a single-support beam. Some twisting of the plate can occur under shock, so that some film damping can occur. But the twisting during normal operation is prevented by the two suspension arms, and good cross-axis sensitivity is obtained. In this direction the device can also be designed to withstand several thousand g . It is worth mentioning that end stops can also be used to prevent excessive travel of the seismic plate in the vertical and transverse directions, a problem that could arise during extended periods of high acceleration. In addition, the end stops can be extended to support the upper capacitor plate over the moving plate, which is a configuration required for differential sensing of the movement of the mass.

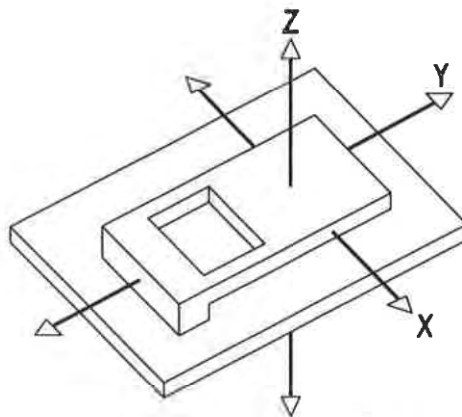


Figure 12.24 Major axis.

12.4.5 Self-Test Features

In many applications self-testing [5] is a requirement. The purpose of the self-test feature is to check that the device and the system are operating correctly. In the case of the accelerometer, this can be achieved with an additional, fixed capacitor plate, as shown in Figure 12.25. Self-testing can be used with either piezoresistive or capacitive sensing. In either case, the self-test feature uses electrostatic forces to deflect the mass and verify operation of the device. By applying a dc voltage between the plates of the test capacitor and the center plate, the mass can be deflected in a manner simulating an externally applied force. Figure 12.26 shows the relation between the plate voltage, and the equivalent g force for an accelerometer with a $2\text{ }\mu\text{m}$ spacing. The electrostatic forces are high for these spacings, and accelerations of 20g and higher can be simulated with only 5V using a test capacitor plate as small as $5,000\text{ }\mu\text{m}^2$. The deflection can then be measured using the normal sensing circuits. Other advantages of the self-test feature are as follows:

1. The functionality of the device can be tested at the probe level.
2. Assembled devices can be checked.

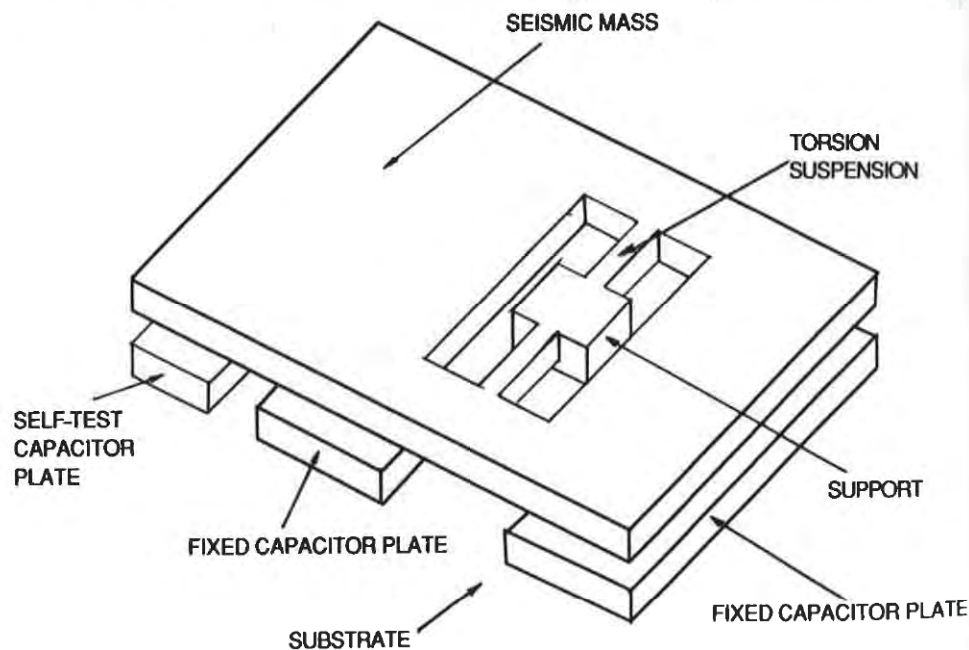


Figure 12.25 Self-test capacitor.

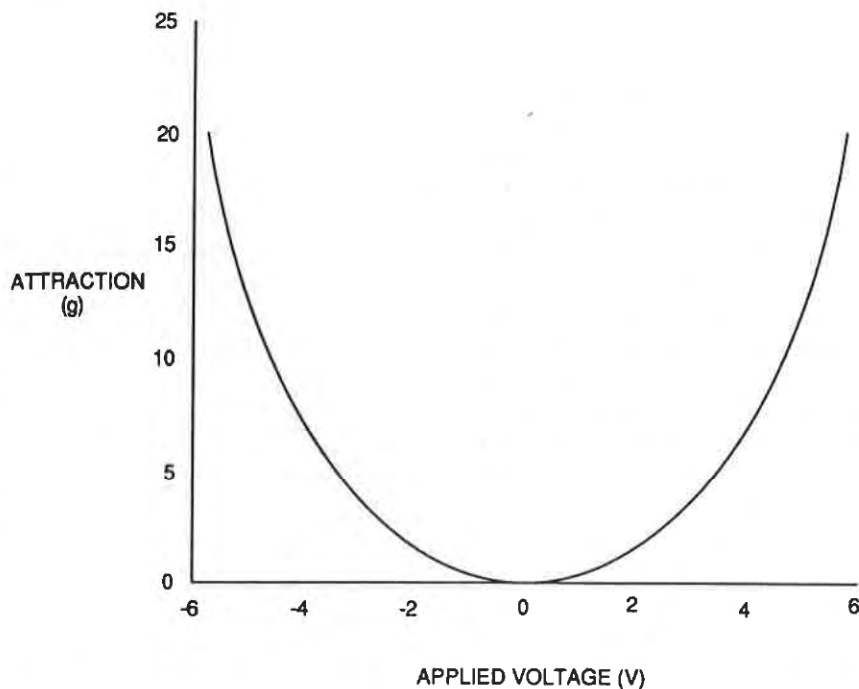


Figure 12.26 Test capacitor voltage versus g force.

3. The device can be exercised for drift and lifetime measurements without having to apply mechanical forces.
4. Operating conditions can be simulated to provide a system test.
5. Devices can be evaluated as functions of frequency and temperature.

The self-test feature can also be used for the development of a smart sensor. The self-test voltage can be trimmed, in order to give a very accurate check of the operation of an accelerometer.

12.4.6 Testing and Trimming

Mass-production testing of dynamic mechanical devices, such as accelerometers, imposes many specific problems in testing and requires special test equipment. For example, it is difficult to check for functionality by probing a wafer of accelerometers on a shaker table. Hence, the self-test capacitor is required to simulate externally applied forces at the test level. The capacitance between the plates of an

accelerometer, with air as an insulator, can be designed to be as low as 0.5 pF. However, the change in capacitance can be several pF when the moving plate is pulled to within close proximity to the fixed plate. This capacitance change can be measured at a probe station, providing care is taken to minimize stray capacitance. When the capacitance is integrated with the signal-conditioning circuit, testing is simplified, although the circuit has to be trimmed for zero offset and sensitivity (and over temperature in the case of piezoresistive sensing). Although the self-test capacitor can be used to check functionality, system bandwidth, and to exercise the device for accelerated life testing, it cannot be used for absolute calibration of sensitivity, unless it is also trimmed and calibrated.

A shaker table is normally used for final testing of accelerometers. The zero offset is trimmed with the device stationary, and then the sensitivity is measured and the device calibrated during acceleration. Usually the calibration is easier at lower acceleration than at higher acceleration, because of the higher velocities and stroke required from the shaker table. Another approach is to use a centrifuge, which can be used for device evaluation but is impractical for final testing, because of the burden of loading and unloading times. During testing, the offset and sensitivity are measured and the biasing resistors trimmed to enable the device to meet the specifications. It is interesting to mention that several methods of trimming are available: NiCr or CrSi resistor biasing chains can be deposited on the die and trimmed using a laser beam to cut shorting links, which will increase the values of the biasing resistors. This method can be used to accurately trim over wide resistor ranges and high resistance values. The main disadvantage is the extra process steps required to deposit the resistors.

Alternative trimming methods using available on-chip components are zener zapping or fusible link technology, either trimming scheme can be performed during the probe operation. In either case a number of pads with biasing resistors connected between them are required. Zener zapping uses zener diodes, which can be broken down and shorted by passing a current through them, these diodes are connected between the pads to short out the appropriate biasing resistors. Fusible link technology uses of metal or polysilicon links between the pads, these links can be blown by passing a high current through them, and the appropriate resistors inserted into the bias chain. The resistors are normally binary weighted with respect to each other and are normally used for course trimming.

Finally, for fine trimming over a limited working range either zipper or polysilicon trimming can be used. With zipper trim a n^+ diffusion or resistor is partially covered with metal that can be made to migrate either way along the resistor to increase or decrease its value. This is accomplished by passing an excessive current through the n^+ and metal combination. In the case of the poly resistor, its value can be reduced by passing a high current through it, thereby changing its grain structure and hence its resistance.

12.5 SUMMARY

This chapter has been devoted to the issues dealing with both the conversion of mechanical forces into electrical signals and the signal processing of those signals, as they apply to mechanical sensors. The basic methods of sensing mechanical forces have been discussed, together with their pros and cons, and the various signal-conditioning schemes used to give an electrical output have been presented. The accelerometer has been used as an example to show how it can be done in an open- or closed-loop system configuration. Dynamic and shock considerations have been discussed, as well as self-test features. Finally, sensor integration has been discussed together with methods of trimming the signal conditioning circuits to obtain zero offset and to set the sensitivity.

REFERENCES

- [1] W. C. Tang, T. C. H. Nguyen, and R. T. Howe, "Laterally Driven Polysilicon Resonant Microstructures," IEEE Micro Mechanical Systems Workshop, 1989.
- [2] K. Peterson and P. Barth, "Silicon Fusion Bonding: Revolutionary New Tool for Silicon Bonding and Microstructures," *Wescon*, 1989, pp. 220-224.
- [3] I. Basket, R. Frank, and D. Slocum, "Integrated Sensors Today and Tomorrow," *Sensors*, Vol. 8, No. 3, 1991, pp. 32-40.
- [4] T. Tamagawa, P. Schiller, H. Yoon and D. L. Polla, "Micromachined Zinc Oxide Thin Film Sensors," *J. of the Electrochemical Society*, 1990, pp. 1084-1085.
- [5] J. C. Cole, "A New Capacitive Technology for Low-Cost Accelerometer Applications," *Sensors Expo. Int.*, 1989.
- [6] R. P. Payne and K. A. Dinsmore, "Surface Micromachined Accelerometer: A Technology Update," SAE Int. Cong. and Exposition, Detroit, 1991, pp. 127-135.
- [7] D. E. Bergfried, B. Mattes, and R. Rutz, "Electronic Crash Sensors for Restraint Systems," *Proc. Int. Cong. on Transportation Electronics*, Detroit, 1990, pp. 169-177.
- [8] P. L. Chen et al., "Integrated Silicon Microbeam PI-FET Accelerometer," *IEEE Trans. Electron Devices*, Vol. ED-29, 1982, p. 27.
- [9] C. Swartz, C. Derrington, and J. Gragg, "Temperature Compensation Methods for the Motorola X-ducer Pressure Sensor," Motorola Application Note AN-840, 1990.
- [10] K. Watanabe and W.-S. Chung, "A Switched Capacitor Interface for Intelligent Capacitive Transducers," *IEEE Trans. on Inst. and Meas.*, Vol. IM35, No. 4, 1986.
- [11] J. T. Kung, H.-S. Lee, and R. T. Howe, "A Digital Readout Technique for Capacitive Sensor Applications," *IEEE J. Solid-State Circuits*, Vol. SC-23, No. 4, 1988, pp. 972-977.
- [12] H. Kandler, J. Eichholz, Y. Manoli, and W. Mokwa, "CMOS Compatible Capacitive Pressure Sensor with Read-out Electronics," *Micro System Technologies*, Vol. 90, 1990, pp. 574-580.
- [13] E. P. Popov, *Mechanics of Materials*, Englewood Cliffs, NJ, Prentice-Hall, 1952, pp. 580-581.
- [14] S. Suzuki et al., "Semiconductor Capacitance-Type Accelerometer with PWM Electrostatic Servo Technique," *Sensors and Actuators*, Vols. A21-A23, 1990, pp. 316-319.
- [15] H. Leuthold and F. Rudolf, "An ASIC for High-Resolution Capacitive Microaccelerometers," *Sensors and Actuators*, Vols. A21-A23, 1990, pp. 278-281.
- [16] H. V. Allen, S. C. Terry, and D. W. De Bruin, "Accelerometer Systems with Self-Testable Features," *Sensors and Actuators*, Vol. A20, 1989, pp. 153-161.

- [17] M. Mutoh et al., "Development of Integrated Semiconductor-Type Acceleration," 1990 IEEE Workshop on Electronic Applications in Transportation, 1990, pp. 35-38.
- [18] H. F. Schlaak et al., "Micromechanical Capacitive Acceleration Sensor with Force Compensation," *Micro System Technologies*, Vol. 90, 1990, pp. 617-622.
- [19] U. E. Gerlach-Meyer, "Capacitive Accelerometer Made by Silicon Micromechanics," *Micro System Technologies*, Vol. 90, 1990, pp. 623-628.
- [20] *American Institute of Physics Handbook* 3d ed., McGraw-Hill Book Co., 1972, p. 9.
- [21] W. T. Thomson, "Vibration," *Mark's Standard Handbook for Mechanical Engineers*, 8th ed., McGraw-Hill Book Co., 1978, pp. 5-67.

Chapter 13

Controlled Oscillators and Their Applicability to Sensors

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13.1 INTRODUCTION

Sinusoidal oscillators and multivibrators represent a widely used group of electronic circuits. A lot of work has been done to investigate and design the circuits where the oscillation parameters (amplitude, frequency, phase, and in multivibrators, the duty cycle) provide information on the values of the passive (resistors, capacitors, and inductors) and active (current and voltage sources) elements used as parts of the oscillating circuits themselves. These values can be functions of external mesurands (mechanical force, electric or magnetic field, temperature, and so on). Then, the modification of oscillation parameters gives information about the external factor.

Many integrated sensors can be used exactly as the aforementioned elements and the oscillating circuits driven from sensors become important interface circuits [1]. Here the circuits of this kind, especially voltage and current controlled oscillators and multivibrators, are reviewed. Preference is given to oscillating circuits with linearly controlled frequency. The frequency output, being noise immune, is required in most cases.

The main attention is given to multivibrator circuits (only one circuit of a voltage-controlled sinusoidal oscillator is considered). A current trend in the realization of integrated sensors is the demand for communication with a microprocessor. The rectangular multivibrator output is easier convertible in the final digital form. Yet, the duty cycle modulation available in a sensor-driven multivibrator can be a very robust form of signal as well and is used in some interesting applications.

The chosen circuits are treated mostly on the level of operational amplifiers, comparators, multivibrators and switches considered as building blocks. Not only the operation principle of the circuits is described. The main goal of the approach

is to present the control characteristics in a form allowing estimation of the imperfections (nonlinearity, instability, drift, and offset) of control characteristics. Special attention is given to the slew-rate limitation of operational amplifiers and comparators. This nonideality, which is not "visible" in the ordinary self-oscillating circuits, can be a main source of errors in sensor applications.

Voltage- and current-controlled multivibrators that can be designed for high-frequency operation are considered on the device level. Here, again, attention is paid mainly to the derivation of the control characteristics to obtain the result providing correct calculation of the oscillation frequency and also for estimation of the control characteristics imperfections.

The scope is limited by continuous-time oscillators and multivibrators. The application of switched-capacitor self-oscillating circuits to the conversion of RLC parameters into a frequency or digital signal can be found in [2].

13.2 CONTROLLED SINUSOIDAL OSCILLATORS

13.2.1 The Oscillator Circuit and Its Control Characteristics

Circuits of RC-sinusoidal oscillators with electronic control of frequency have been known for some years [3–5] and were even systematized in [6]. Their later versions [7–9] are based on multipliers. Practical applications of these circuits seem limited to laboratory experiments even though the authors [8,9] underline the useful features of these oscillators; for example, linear control of frequency.

One of the best oscillators with electronic amplitude and frequency control [8] is shown in Figure 13.1. The circuit includes the main positive feedback loop (operational amplifiers A_1 , A_2 , and A_4 and multipliers M_1 and M_2). The oscillation frequency is controlled by the voltage V_f . The sinusoidal voltages v_o and v_d are shifted by 90° . The amplitude control circuit includes the multipliers M_3 and M_4 operating as squarers, the error amplifier A_3 , and the multiplier M_5 , which controls the main loop damping. The oscillation amplitude is controlled by the voltage V_a .

The circuit can be described by a simplified two-loop block diagram (Figure 13.2), including two integrators in the main loop and the negative feedback loop around one of the integrators. From this block diagram one obtains that

$$v_o = -\frac{K_2 V_f}{T_2} \int v_d dt \quad (13.1)$$

and

$$v_d = -\frac{1}{T_1} \int \left(\frac{R_1 K_5 V_e}{R_8} v_d - H K_1 V_f v_o \right) dt \quad (13.2)$$

where $H = R_9/R_{10}$, $T_1 = R_1C_1$, and $T_2 = R_2C_2$. Eliminating the voltage v_o from (13.1) and (13.2), one can establish the oscillator differential equation

$$\frac{d^2v_d}{dt^2} + \frac{R_1K_5V_e}{T_1R_8} \frac{dv_d}{dt} + \frac{HK_1K_2V_f^2}{T_1T_2} v_d = 0 \quad (13.3)$$

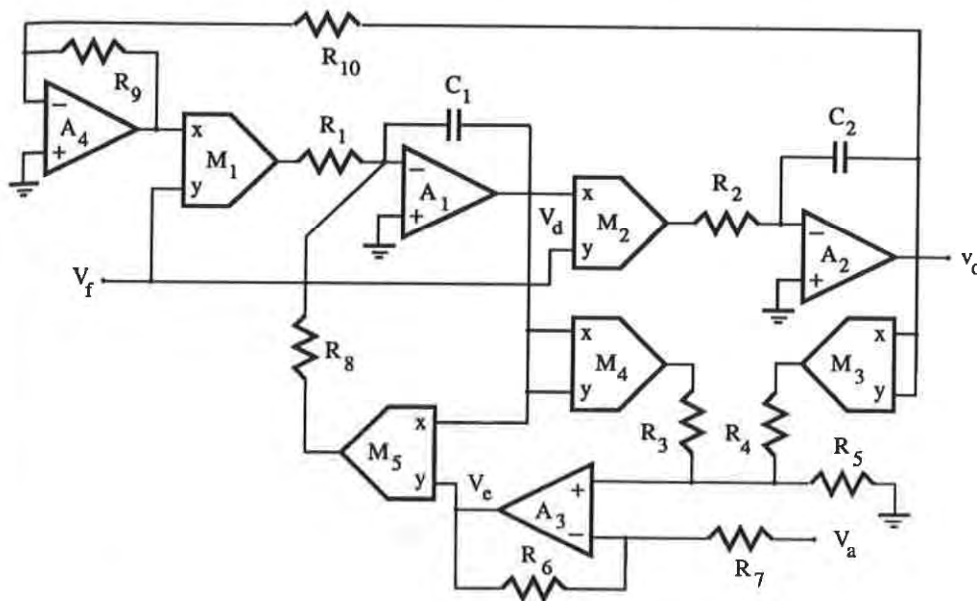


Figure 13.1 Sinusoidal oscillator with electronic amplitude and frequency controls.

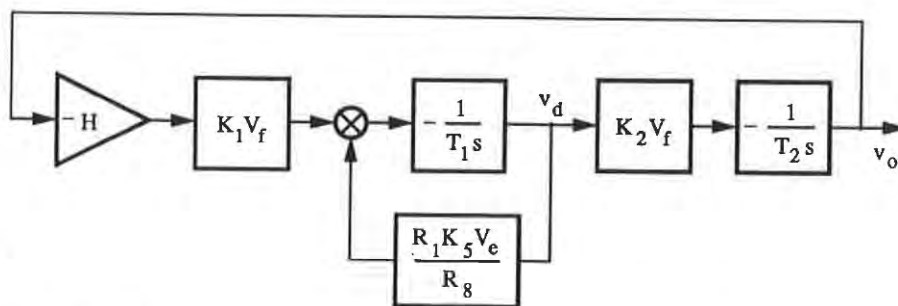


Figure 13.2 Feedback loops in the oscillator.

In the steady state oscillations $V_e = 0$ and the oscillation frequency as follows from (13.3) is

$$\omega_0 = V_f \sqrt{\frac{HK_1K_2}{T_1T_2}} \quad (13.4)$$

The resistor and the timing capacitor values in the main loop are usually chosen so that $H = 1$, $R_1 = R_2 = R_f$, $C_1 = C_2 = C_f$. Then $T_1 = T_2 = T_f$. The multiplying constants are also usually set equal. If $K_1 = K_2 = K$, then (13.4) is simplified to

$$\omega_0 = \frac{KV_f}{T_f} = KV_f\omega_f \quad (13.5)$$

Therefore, the frequency control characteristic is linear. This linear frequency control was experimentally verified. The theoretical and experimental characteristics coincide very well [10] when the circuit is realized with standard operational amplifiers (LF356) and multipliers (ICL8013). They even preserve linearity for the control voltages V_f , which are above the multiplier full voltage range of 10V.

In this oscillator it is difficult to obtain the frequency control of more than two decades for one capacitor setting. The decreasing control voltage V_f opens the main loop, and the level of nonlinear distortions, caused by the operational amplifiers offset and drift, increases. If the standard [11,12] offset compensation circuits are used both for operational amplifiers and multipliers the total harmonic distortion (THD) can be reduced to -80 dB at $V_f = 10$ V. In case of offset compensation circuits in multipliers, the THD becomes about -60 dB only at $V_f = 10$ V [10], it increases approximately linearly when V_f decreases, and at $V_f = 0.1$ V it becomes -30 dB.

In the steady state oscillation $v_d = V_{dm} \sin \omega_0 t$, and then one obtains from (13.1) and (13.4), with the aforementioned choice of parameters in the main loop,

$$v_o = V_{om} \cos \omega_0 t = V_{dm} \cos \omega_0 t \quad (13.6)$$

so that the amplitudes V_{om} and V_{dm} become equal. Then one finds that

$$V_e = \frac{R_5(K_4R_4V_d^2 + K_3R_3V_o^2)}{(R_3R_4 + R_4R_5 + R_3R_5)} \left(1 + \frac{R_6}{R_7}\right) - \frac{R_6}{R_7} V_a \quad (13.7)$$

When $V_{om} = V_{dm}$ and, in addition, in the amplitude control circuit $R_3 = R_4 = R_a$ and $K_3 = K_4 = K$ as well, then the condition $V_e = 0$ gives the oscillation amplitude

$$V_{om} = \left\{ \frac{V_a[2 + (R_a/R_5)]}{K[1 + (R_7/R_6)]} \right\}^{1/2} \quad (13.8)$$

The experimental and theoretical amplitude control characteristics for this oscillator are in a close agreement as well [10]. The square root law of (13.8) can be transformed into a linear law using an additional multiplier.

13.2.2 Oscillator Amplitude and Frequency Transients

The oscillator amplitude transients are also predicted using the model shown in Figure 13.2 and the assumption that the amplitude dependent feedback around the first integrator changes slowly within an oscillation period. Then, if the time constants in the integrators are equal and the multiplying coefficients are equal, the oscillator is described by the characteristic equation

$$s^2 + \frac{R_f K V_e}{T_f R_8} s + \frac{H K^2 V_f^2}{T_f^2} = 0 \quad (13.9)$$

Deviations of V_e pull the complex conjugate poles of this equation to the right half or the left half of the s -plane from their position on the $j\omega$ -axis, thereby producing growth or decay of the oscillation amplitude. The oscillation amplitude can then be given [13] as

$$V_{dm} = V_{dm0} \exp\left(\int_0^t \delta p_r(t) dt\right) \quad (13.10)$$

where $\delta p_r(t)$ is the real component of the change in the $j\omega$ -poles, and V_{dm0} is the initial steady state value of amplitude, when $\delta p_r(t) = 0$. It follows from the (13.9) that

$$\delta p_r = -\frac{R_f K}{2T_f R_8} \delta V_e \quad (13.11)$$

and, hence,

$$V_{dm} = V_{dm0} \exp\left[-\frac{K}{2C_f R_8} \int_0^t \delta V_e(t) dt\right] \quad (13.12)$$

The following analysis considers only one type of amplitude transient response; namely, the transient response for a step change in the reference voltage. In this case the oscillation amplitude changes from the initial value V_{dm0} to the new steady state value V_{dm1} (both of these values can be calculated from (13.7) and the condition $V_{dm} = V_{om}$) when the amplitude control voltage changes from V_{a0} to V_{a1} . Such a restriction allows integration in closed form of the amplitude control differential

equation, which follows. Considering the steps of different amplitudes at different initial conditions, one can obtain adequate information on the oscillator amplitude dynamics [10,14]. As follows from (13.7) for a step change in the amplitude control voltage,

$$\delta V_e = \frac{[1 + (R_6/R_7)]}{[2 + (R_a/R_5)]} K V_{dm}^2 - \frac{R_6}{R_7} V_{a1} \quad (13.13)$$

The inversion of (13.8) gives

$$V_{a1} = \frac{[1 + (R_7/R_6)]}{[2 + (R_a/R_5)]} K V_{dm1}^2 = \frac{[1 + (R_7/R_6)]}{[2 + (R_a/R_5)]} K V_{dm1}^2 \quad (13.14)$$

and the substitution of (13.13) and (13.14) into (13.12) results in

$$V_{dm} = V_{dm0} \exp \left[\int_0^t a_0 (V_{dm1}^2 - V_{dm}^2) dt \right] \quad (13.15)$$

Here $a_0 = \{[1 + (R_6/R_7)]K^2\}/\{2[2 + (R_a/R_5)]R_8C_f\}$. Differentiating the (13.15) one obtains the amplitude control differential equation

$$\frac{dV_{dm}}{dt} = a_0 V_{dm} (V_{dm1}^2 - V_{dm}^2) \quad (13.16)$$

This equation can be integrated [12]. The final result of integration is

$$V_{dm} = \frac{V_{dm1}}{\{1 + [(V_{dm1}^2/V_{dm0}^2) - 1] \exp(-2a_0 V_{dm1}^2 t)\}^{1/2}} \quad (13.17)$$

The calculation shows that it is easy to choose the parameters (say, for example, $R_a = R_f = R_6 = R_8 = 100 \text{ k}\Omega$, $R_5 = R_7 = 62 \text{ k}\Omega$) so that if, for example, $V_{dm0} = 3\text{V}$ and $V_{dm1} = 10\text{V}$ the amplitude transient is about one oscillation period. If the amplitude decreases by a step, and $V_{dm0} = 10\text{V}$ and $V_{dm1} = 3\text{V}$, the transient is about three oscillation periods. The experimental transients are in a good agreement with calculations [10] and show that this oscillator has a very fast amplitude response. It is possible to create noncontrolled oscillators with fast amplitude transients [13–20] but the solutions for voltage controlled oscillators are very rare, and this criterion was decisive in the choice of the considered circuit. In any real oscillator the controls of frequency and amplitude interact (for the considered circuit this is considered later) and the frequency variation produces perturbations in the am-

plitude control systems. These small perturbations can cause long amplitude transients [21].

The oscillator also has a fast frequency response. Equation (13.5) shows that the new value of the oscillation frequency should be established instantly when the frequency control voltage V_f changes by a step. Experimentation [10] confirms this assumption, even, for example, for the steps of V_f from 10V to 1.25V and vice versa, resulting in an eightfold change of the oscillation frequency with no noticeable transients.

13.2.3 Interaction of the Frequency and Amplitude Controls

The frequency and amplitude controls in the considered oscillator should be independent, as follows from (13.5) and (13.8). In practice, a weak interaction exists between them. This interaction is more noticeable when the oscillator operates at high frequencies. The mechanism of this interaction deserves some attention because it is also the cause of frequency errors in some sensor-controlled multivibrators.

Assume that an operational amplifier has a slew rate S_r limitation but is ideal otherwise. Let this amplifier operate as an integrator (Figure 13.3), of which the input voltage v_i is a rectangular wave with an amplitude of V_i and frequency of $f_o = 1/T$. The output voltage v_o is a triangular wave of the amplitude $V_o = 4V_i/T$. The relationship between v_o and v_i is

$$v_o = -\frac{1}{R_i C_i} \int v_i dt \quad (13.18)$$

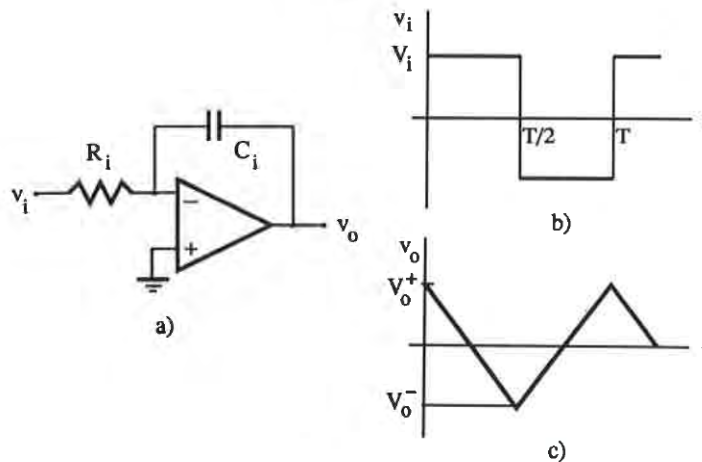


Figure 13.3 Integration of a rectangular wave: (a) integrator; (b) input voltage; (c) output voltage.

which, in this case, can be rewritten as

$$v_i = -R_i C_i \frac{dv_o}{dt} \quad (13.19)$$

If the input amplitude changes and achieves the value V_{IM} , such that the output wave is limited by S_r , then

$$V_{IM} = -R_i C_i S_r \quad (13.20)$$

The intrinsic time constant $R_i C_i$ can be eliminated now, and one can write that

$$v_i = -\frac{V_{IM}}{S_r} \frac{dv_o}{dt} \quad (13.21)$$

Let the circuit shown in Figure 13.3, become an intrinsic stage, causing a slew rate limitation of another operational amplifier (Figure 13.4(a)). Then, to obtain a triangular output wave at the output of this more complicated operational amplifier, one has to apply a voltage v_i given by (13.21) directly to the amplifier input. However, if this amplifier operates with an external feedback circuit (Figure 13.4(b)) and

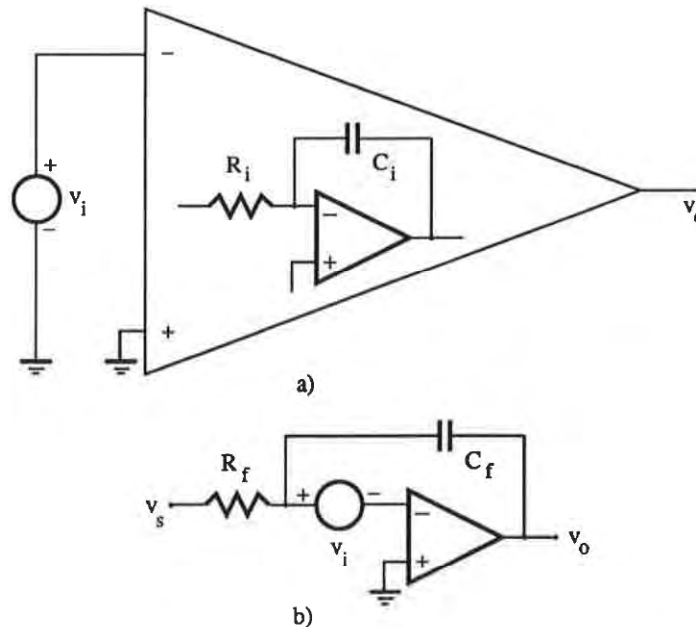


Figure 13.4 Slew-rate error voltage at: (a) the limiting stage input; (b) the slew-rate limited amplifier input.

one considers that the amplifier output voltage is caused by this external generator v_s , the compensating voltage v_i (note the polarity) should be introduced in the input terminal (like an offset voltage in dc analysis). Then one can write, from Figure 13.4(b), that

$$v_o = -\frac{1}{R_f C_f} \int v_s dt + v_i + \frac{1}{R_f C_f} \int v_i dt \quad (13.22)$$

Substituting (13.21) in (13.22) and differentiating one obtains that

$$\frac{dv_o}{dt} = -\frac{1}{R_f C_f} v_s - \frac{V_{IM}}{S_r} \frac{d^2 v_o}{dt^2} - \frac{1}{R_f C_f S_r} \frac{dv_o}{dt} \quad (13.23)$$

Thus, the transfer function of the integrator using an operational amplifier with a slew rate limitation is

$$\frac{v_o(s)}{v_s(s)} = -\frac{\omega_f}{s(1 + \eta\omega_f + \eta s)} \approx -\frac{\omega_f}{s(1 + \eta s)} \quad (13.24)$$

where $\omega_f = 1/(R_f C_f)$ and $\eta = V_{IM}/S_r$. The slew rate limitation introduces an additional pole located at $s = -1/(\eta R_f C_f) - 1/\eta \approx -1/\eta$.

The circuit (Figure 13.5) that includes two such integrators and an inverter with the same limitation will oscillate. Assuming that all operational amplifiers in this circuit are identical, one can find that the characteristic equation of this circuit is

$$s^2(1 + \eta\omega_f + \eta s)^2(1 + \eta s) + \omega_f^2 = 0 \quad (13.25)$$

If η is small and the circuit oscillates at the frequency $\omega_f \ll 1/(2\eta)$ the dominant roots satisfy the equation

$$3\eta s^3 + s^2 + \omega_f^2 = 0 \quad (13.26)$$

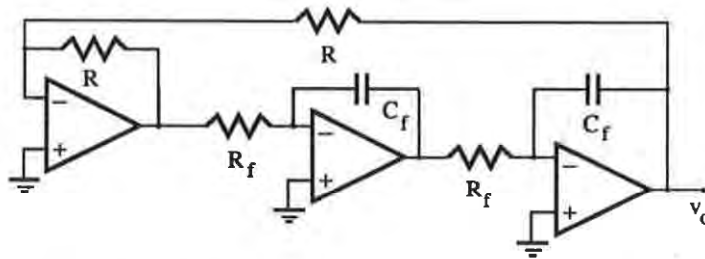


Figure 13.5 Oscillating subcircuit.

which can be rewritten as

$$(3\eta s + 1)(s^2 - 2\eta\omega_f^2 s + \omega_f^2) = 0 \quad (13.27)$$

Thus, the circuit has two roots with positive real parts located near the points $\pm j\omega_f$. The circuit will oscillate, of course, at the frequency very close to ω_f .

If the circuit of Figure 13.5 is used with the same amplitude and frequency control circuits as in Figure 13.1 the circuit characteristic equation will be

$$s^2(1 + 4\eta s) + KV_e\omega_f s(1 + 3\eta s) + (KV_f\omega_f)^2 \approx 0 \quad (13.28)$$

which can be represented as

$$(1 + 4\eta s)\{s^2 - [KV_e\omega_f - 4\eta(KV_f\omega_f)^2]s + (KV_f\omega_f)^2\} \approx 0 \quad (13.29)$$

The circuit will oscillate at the frequency $KV_f\omega_f$ if

$$V_e = 4\eta\omega_f KV_f^2 \quad (13.30)$$

Then, using (13.7) one can find that the oscillation amplitude will be

$$V_{om} = \left\{ \frac{(2 + R_6/R_5)}{K(1 + R_6/R_7)} \left(\frac{R_6}{R_7} V_a + 4\eta\omega_f k V_f^2 \right) \right\}^{1/2} \quad (13.31)$$

This result shows that if the operational amplifiers have slew-rate limitation, the oscillator will have interacting controls and the oscillation amplitude will depend on both V_a and V_f . This conclusion was verified experimentally [10]. The oscillation frequency still changes linearly with the control voltage V_f .

13.2.4 Sinusoidal Oscillators in Applications

Thus, it is feasible to design an oscillator with linear dependence of the oscillation frequency on the control voltage. The transients will be practically absent even when one has fast control voltage changes. The frequency and amplitude control interaction is weak when the oscillation frequencies are low, but at high frequencies this interaction is noticeable. The oscillation amplitude increases, and this can introduce the measurements errors if the oscillator output is used in a phase-locked system [22]. The lower oscillation frequency is also limited by increasing nonlinear distortions. Yet, a two decade variation of frequency can be easily achieved with one capacitor value. The circuit has to include rather bulky offset compensation circuitry, which makes difficult circuit integration. These circuits are more complicated than the

multivibrator-based circuits designed for frequency control. In addition, their useful properties, such as independent control of frequency and amplitude, are essentially deteriorating when they are used in sensors, that is, with low voltage power supplies and with small capacitors.

13.3 CONTROLLED MULTIVIBRATORS

13.3.1 Multivibrators versus Sinusoidal Oscillators

The sinusoidal oscillators with control of oscillation parameters represent rather complicated circuits, although they can be simplified when only one oscillation parameter (usually frequency) must be controlled. Even in that case, however, amplitude stabilization is required, otherwise the circuit will have an oscillation frequency different from that designed [23,24]. The simplest amplitude stabilization system usually includes an amplitude detector and filter, and the need for small amplitude modulation increases the filter capacitor size and reduces the amplitude stabilization speed. A perturbation of this stabilization system results in a transient with a duration of tens and even hundreds of oscillation periods. The frequency control in sinusoidal RC-oscillators usually requires two changing elements (resistors or capacitors). Even though the voltage controlled oscillators do not require tightly controlled components (as is the case of tuned RC oscillators [25]), these difficulties prevent wide use of sinusoidal oscillators in sensor applications. There are sensors where the oscillating system or the oscillator feedback circuit is a highly selective structure (mechanical or quartz resonator) [26,27], allowing one to neglect the influence of nonlinear distortions on the oscillation frequency. Unfortunately, such sensors are not compatible with microelectronic technology. The oscillators with distributed structures in active feedback [28,29] are not frequently realized in sensor applications [30]. Here the condition of oscillation is sensitive to the feedback structure parameters (which is used for sensing) and the feedback structure has high selectivity, allowing one to omit amplitude stabilization and use the power supply voltage limitation.

The amplitude stabilization by power supply voltage imposes no requirements on the multivibrator structure. The circuits of amplitude stabilization are very simple (a resistor and a temperature compensated Zener diode [8]) or even omitted (if switches and the power supply voltage are used for this purpose). The amplitude and frequency transients are very short in multivibrators as well, usually their duration is about one oscillation period. In addition, frequency control is usually achieved by modification of one element (resistor or capacitor). All these advantages of multivibrators resulted in their use in much wider sensor applications.

Frequency modulation by the sensor signal of the multivibrator output voltage or current is used more frequently than with resistor or capacitor modulation. This is because in many cases the multivibrator is preceded by a voltage or current amplifier.

The method of conversion of the frequency-modulated rectangular signal into a useful analog or digital form is highly immune to noise and interference signals. Here the phase-lock loop techniques are the most important [31,32]. They allow not only easy transformation of the signal in analog form but the frequency conversion and multiplexing as well. The signal can be put in the frequency band suitable for telecommunication transmission.

13.3.2 Multivibrators with Operational Amplifiers

A frequently used multivibrator circuit is shown in Figure 13.6(a). It includes an integrator and a Schmitt trigger. The output voltage v_o is a rectangular wave limited by the values V_o^+ and V_o^- (Figure 13.6(b)) and the integrator output voltage v_1 is a triangular wave achieving the values V_1^+ and V_1^- (Figure 13.6(c)). The transition of the multivibrator from one quasi-stable state to another is determined by the balance of voltage at the positive terminal of the Schmitt trigger operational amplifier. At the end of the interval T_1 , one has

$$V_1^- \frac{R_b}{R_a + R_b} + V_o^+ \frac{R_a}{R_a + R_b} = 0 \quad (13.32)$$

which gives $V_1^- = -V_o^+ (R_a/R_b)$. By the same token, at the end of interval T_2 , one has $V_1^+ = -V_o^- (R_a/R_b)$. Thus, the capacitor voltage v_c has the swing of $\Delta V_c =$

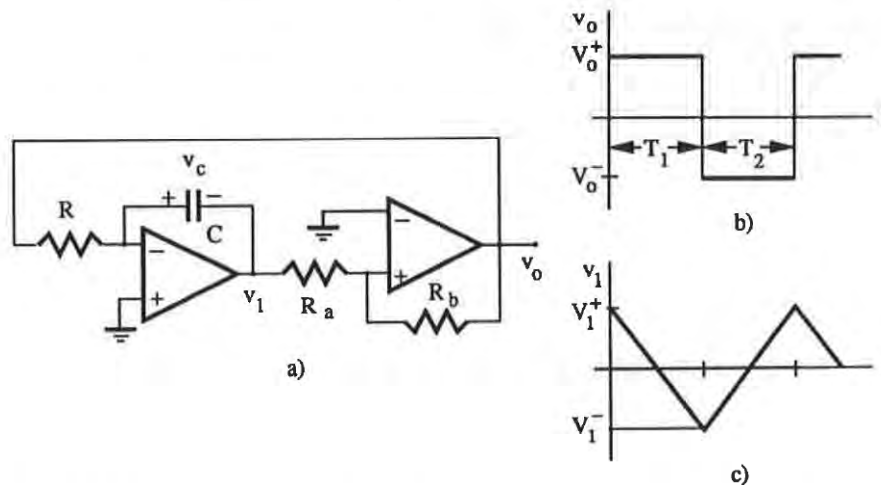


Figure 13.6 Common two-amplifier multivibrator: (a) circuit; (b) output voltage; (c) integrator output voltage.

$V_1^+ - V_1^- = (V_o^+ - V_o^-) (R_a/R_b)$. During the first interval capacitor C is recharged by current $I_1 = V_o^+/R$, thus, the duration

$$T_1 = RC \left(\frac{V_o^+ - V_o^-}{V_o^+} \right) \frac{R_a}{R_b} \quad (13.33)$$

To obtain the duration T_2 it is enough to interchange the plus and minus in the superscripts of (13.33). Hence,

$$T_2 = RC \left(\frac{V_o^- - V_o^+}{V_o^-} \right) \frac{R_a}{R_b} \quad (13.34)$$

and the oscillation period is

$$T = T_1 + T_2 = RC \frac{R_a(V_o^+ - V_o^-)(V_o^- - V_o^+)}{R_b V_o^+ V_o^-} \quad (13.35)$$

If $V_o^+ = -V_o^- = V_o$ then (13.35) is simplified, and the oscillation frequency becomes

$$f_0 = \frac{1}{4RC} \left(\frac{R_b}{R_a} \right) \quad (13.36)$$

The oscillation frequency can be modulated modifying R or C . The triangular wave amplitude can also be modified by R_a , R_b , or their ratio (the frequency, in this case, is modulated as well).

The main source of dynamic frequency errors in this multivibrator is the finite integrator slew rate. The correct frequency value can be calculated the following way. If the error voltage source

$$v_i = -\frac{V_{IM}}{S_{\eta i}} \frac{dv_1}{dt} \quad (13.37)$$

is introduced in the input terminal of the integrator (Figure 13.7(a)) the integrator output voltage, say, when the multivibrator output voltage is V_o^+ , will be

$$v_1 = -\frac{1}{RC} \int V_o^+ dt + v_i + \frac{1}{RC} \int v_i dt \approx -\frac{1}{RC} \int V_o^+ dt + \frac{1}{RC} \int v_i dt \quad (13.38)$$

Differentiating (13.38) and using (13.37) one finds that during this period of time

$$\frac{dv_1}{dt} = -\frac{V_o^+}{RC + \eta_i} \quad (13.39)$$

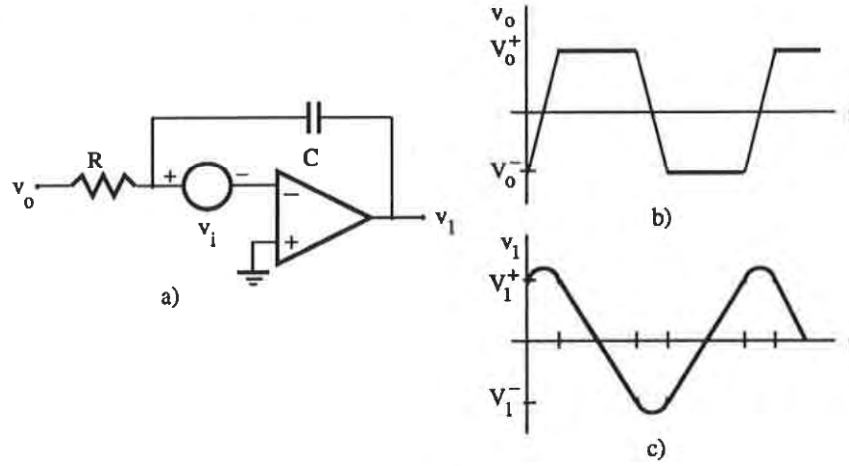


Figure 13.7 Slew-rate error voltage and multivibrator wave distortions: (a) integrator slew-rate error voltage; (b) output wave; (c) integrator output.

where $\eta_i = V_{IM}/S_{rs}$. From the other side, the capacitor voltage swing

$$\Delta V_c = \pm(V_o^+ - V_o^-) \frac{R_a}{R_b} \quad (13.40)$$

is not influenced by the presence of v_i . Thus, the correct oscillation period will be

$$T = T_1 + T_2 = (RC + \eta_i) \frac{R_a(V_o^+ - V_o^-)(V_o^- - V_o^+)}{R_b V_o^+ V_o^-} \quad (13.41)$$

that is, the source v_i introduces the error

$$\frac{\delta f}{f_0} \approx -\frac{\eta_i}{RC} = -\frac{4f_0 \eta_i R_a}{R_b} \quad (13.42)$$

which increases with the frequency.

When the oscillation frequency becomes very high it is necessary to consider the finite slew rate of the Schmitt trigger operational amplifier. The Schmitt trigger has positive feedback, and one can consider that its output voltage always changes with a constant speed of S_{rs} between multivibrator quasi-stable states (Figure 13.7(b)). The oscillation period becomes

$$T = T_1 + T_2 = (RC + \eta_i) \frac{R_a(V_o^+ - V_o^-)(V_o^- - V_o^+)}{R_b V_o^+ V_o^-} + \frac{2(V_o^+ - V_o^-)}{S_{rs}} \quad (13.43)$$

so that the frequency error is

$$\frac{\delta f}{f_0} \approx -\frac{\eta_i}{RC} - \frac{4V_0 f_0}{S_{rs}} \quad (13.44)$$

The integrator output voltage has now visibly smoothed corners (Figure 13.7(b)), and the capacitor voltage swing becomes

$$\Delta V_c = 2V_1 + \frac{V_o^2}{(RC + \eta_i)S_{rs}} \quad (13.45)$$

where the symmetric swing is assumed.

13.3.3 Switches in the Basic Multivibrator Circuits

Some useful modification occurs when additional switches are incorporated in the basic multivibrator circuit. The switches being realized by active devices of minimal geometry are not a heavy burden, either in terms of the multivibrator complexity or in chip area.

13.3.3.1 Duty-Cycle Modulation

The first useful modification is shown in Figure 13.8. The output voltage in this multivibrator controls the switches P_1 and P_2 , connecting the resistors R_1 and R_2 to the integrator input. The durations T_1 and T_2 will be

$$T_1 = CR_1 \frac{(V_o^+ - V_o^-)R_a}{V_{CC}^+ R_b} \quad (13.46)$$

$$T_2 = CR_2 \frac{(V_o^- - V_o^+)R_a}{V_{CC}^+ R_b} \quad (13.47)$$

If the output wave is symmetric and, in addition, $V_{CC}^+ = -V_{CC}^- = V_{CC}$ the oscillation period becomes

$$T = 2(R_1 + R_2)C \frac{V_o R_a}{V_{CC} R_b} \quad (13.48)$$

This circuit is used when the resistor change is transformed into a duty cycle change. For this purpose it is more convenient to consider the quasi-stable state

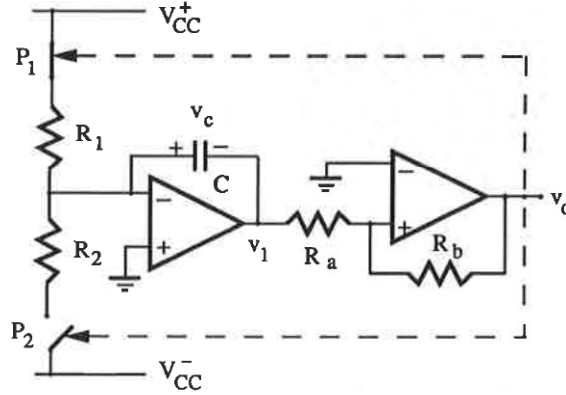


Figure 13.8 Switch-commutated resistors.

duration ratio $D = T_1/T_2$. If $R_1 = R + \delta R$ and $R_2 = R - \delta R$, then the oscillation frequency for the circuit of Figure 13.8 does not change and

$$D = \frac{T_1}{T_2} = \frac{R_1}{R_2} \approx 1 + \frac{\delta R}{R} \quad (13.49)$$

The errors caused by the slew-rate of the circuit operational amplifiers can be calculated individually for T_1 and T_2 , as was outlined in the previous section. In addition, these durations should be augmented by the switching time of the switches.

13.3.3.2 Bridge-to-Frequency Converter

Using switches in the two-operational-amplifier multivibrator one can obtain the circuit where the oscillation frequency is controlled by detuning a resistive bridge. Such a circuit is shown in Figure 13.9. For normal circuit operation the bridge should be detuned (at least, the resistors R_1 and R_2 should be different).

The conditions for the circuit transition from one quasi-stable state to another are $V_1^- = -V_o^+(R_a/R_b)$ and $V_1^+ = -V_o^-(R_a/R_b)$ as in the circuit of Figure 13.6. Let the position of switches shown in Figure 13.8 correspond to the first quasi-stable state, and the output voltage is $v_o = V_o^+$ at this state. Then, at the end of this state the capacitor voltage v_c will be

$$v_c(T_1) = \frac{V_{CC}^+ R_4 + V_{CC}^- R_3}{(R_4 + R_3)} + V_o^+ \frac{R_a}{R_b} \quad (13.50)$$

The voltage v_c at the end of second quasi-stable state (equal to v_c at the beginning

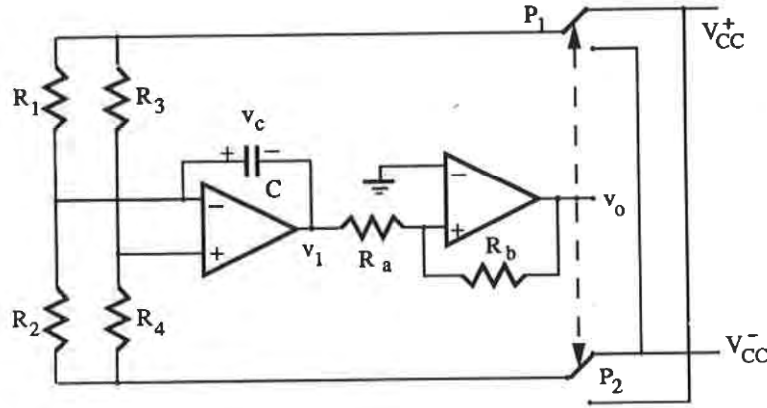


Figure 13.9 Switch-commutated bridge in the two-amplifier multivibrator.

of the first quasi-stable state) is obtained by interchanging the plus and minus in the superscripts (13.50):

$$v_c(0) = \frac{V_{cc}^- R_4 + V_{cc}^+ R_3}{(R_4 + R_3)} + V_o^- \frac{R_a}{R_b} \quad (13.51)$$

Thus, the capacitor voltage swing during the first quasi-stable state is

$$\Delta V_c = \frac{(V_{cc}^+ - V_{cc}^-)(R_4 - R_3)}{(R_4 + R_3)} + (V_o^+ - V_o^-) \frac{R_a}{R_b} \quad (13.52)$$

During this quasi-stable state the capacitor is recharged by the current

$$I_1 = \frac{(V_{cc}^+ - v^+)}{R_1} + \frac{(V_{cc}^- - v^+)}{R_2} = \frac{(V_{cc}^+ - V_{cc}^-)}{(R_3 + R_4)} \left(\frac{R_3}{R_4} - \frac{R_4}{R_2} \right) \quad (13.53)$$

One finds that the duration of the quasi-stable state

$$T_1 = T_2 = \frac{(R_4 - R_3)C}{[(R_3/R_1) - (R_4/R_2)]} + \frac{R_a(R_3 + R_4)C}{R_b[(R_3/R_1) - (R_4/R_2)]} \left(\frac{V_o^+ - V_o^-}{V_{cc}^+ - V_{cc}^-} \right) \quad (13.54)$$

and the oscillation frequency is

$$f_0 = \frac{1}{2C} \left(\frac{R_3}{R_1} - \frac{R_4}{R_2} \right) \frac{(V_{cc}^+ - V_{cc}^-)R_b}{[R_b(R_4 - R_3)(V_{cc}^+ - V_{cc}^-) + R_a(R_4 + R_3)(V_o^+ - V_o^-)]} \quad (13.55)$$

The switches allow the elimination of the feedback resistances of the Schmitt trigger (Figure 13.10) [33]. For this circuit the transition from the first quasi-stable state to the second is determined by the condition $v_1 = V_1^- = V_{CC}^-$ (the condition of the opposite transition is $v_1 = V_1^+ = V_{CC}^+$). The capacitor recharges from

$$v_c(0) = \frac{V_{CC}^- R_4 + V_{CC}^+ R_3}{(R_4 + R_3)} - V_{CC}^+ \quad (13.56)$$

to

$$v_c(T_1) = \frac{V_{CC}^+ R_4 + V_{CC}^- R_3}{(R_4 + R_3)} - V_{CC}^- \quad (13.57)$$

so that

$$\Delta V_c = 2(V_{CC}^+ - V_{CC}^-) \frac{R_4}{(R_3 + R_4)} \quad (13.58)$$

The recharge current is given by (13.53) again and one obtains that

$$T_1 = T_2 = \frac{2CR_4}{[(R_3/R_1) - (R_4/R_2)]} \quad (13.59)$$

and the oscillation frequency is

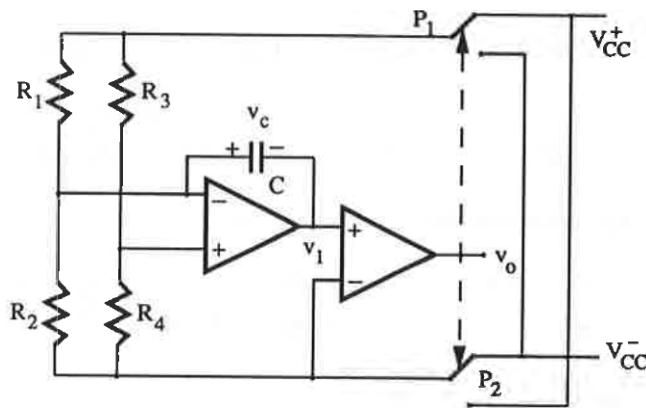


Figure 13.10 Switch-commutated bridge in the amplifier-comparator multivibrator.

$$f_0 = \frac{1}{4CR_4} \left(\frac{R_3}{R_1} - \frac{R_4}{R_2} \right) \quad (13.60)$$

The finite slew rate of the integrator can be a main source of frequency error. For error analysis it is convenient, for this case, to use two separate error voltages V_I^+ and V_I^- for each quasi-stable state and to represent them as

$$V_I^+ = -V_I^- = \frac{V_{IM}}{S_{ri}} \left(\frac{V_1^+ - V_1^-}{T/2} \right) = \frac{2f_0 V_{IM} (V_o^+ - V_o^-) R_a}{R_b S_{ri}} \quad (13.61)$$

Figure 13.11 shows the integrator of the multivibrator with this source of error during the first quasi-stable state. Then one can write that

$$v_c(T_1) = \frac{V_{cc}^+ R_4 + V_{cc}^- R_3}{(R_3 + R_4)} + V_I^+ + V_o^+ \frac{R_a}{R_b} \quad (13.62)$$

Again, the initial value of the capacitor voltage can be obtained by interchanging the signs in the superscripts. This gives

$$v_c(0) = \frac{V_{cc}^- R_4 + V_{cc}^+ R_3}{(R_3 + R_4)} + V_I^- + V_o^- \frac{R_a}{R_b} \quad (13.63)$$

so that

$$\Delta V_c = (V_{cc}^+ - V_{cc}^-) \left(\frac{R_4 - R_3}{R_4 + R_3} \right) + (V_I^+ - V_I^-) + (V_o^+ - V_o^-) \frac{R_a}{R_b} \quad (13.64)$$

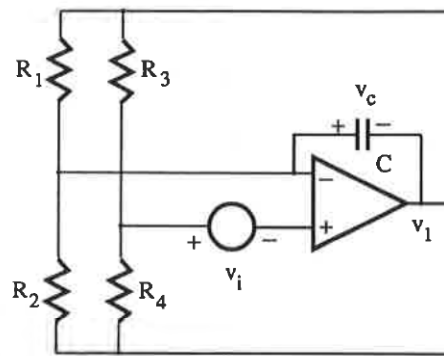


Figure 13.11 Slew-rate error voltage in the integrator with a bridge.

The recharge current in this case will be

$$I_1 = \left(\frac{V_{CC}^+ - V_{CC}^-}{R_4 + R_3} \right) \left(\frac{R_3}{R_1} - \frac{R_4}{R_2} \right) - V_I^+ \left(\frac{1}{R_1} + \frac{1}{R_2} \right) \quad (13.65)$$

Using (13.64) and (13.65) one can find the duration of T_1 . Then, changing the signs of the subscripts in T_1 , the duration of T_2 will be found. After this, one can find the oscillation frequency. These cumbersome expressions are omitted here. In the case of $V_{CC}^+ = -V_{CC}^- = V_{CC}$, $V_o^+ = -V_o^- = V_o$, and $V_I^+ = -V_I^- = V_I$ the result will be

$$f = f_0 + \delta f = \frac{\frac{2V_{CC}}{(R_3 + R_4)} \left(\frac{R_3}{R_1} - \frac{R_4}{R_2} \right) - V_I \left(\frac{1}{R_1} + \frac{1}{R_2} \right)}{4C \left[V_{CC} \left(\frac{R_4 - R_3}{R_3 + R_4} \right) + V_I + V_o \frac{R_a}{R_b} \right]} \quad (13.66)$$

Neglecting the term V_I in the denominator of (13.66) one finds that

$$\frac{\delta f}{f_0} \approx -\frac{V_I}{2V_{CC}} F(R_B) = -\frac{f_0}{S_{ri}} \frac{V_{IM} V_o R_a}{V_{CC} R_b} F(R_B) \quad (13.67)$$

where $F(R_B) = [(R_1 + R_2)(R_3 + R_4)] / (R_2 R_3 - R_1 R_4)$.

It is easy to verify that, for the circuit of Figure 13.10, the frequency error is [33]

$$\frac{\delta f}{f_0} \approx \frac{f_0 V_{IM}}{S_{ri}} F(R_B) \quad (13.68)$$

When the oscillation frequency increases the slew rate of the output comparator in Figure 13.9 or Figure 13.10, t increases the durations of T_1 and T_2 , as in the circuit of Figure 13.6. In addition, the switching time of P_1 and P_2 should be added to T_1 and T_2 as well.

13.3.4 One-Amplifier Multivibrator

The circuit of a one-amplifier multivibrator (Figure 13.12(a)) is well known and can be used in many sensor applications. At low oscillation frequency the output voltage v_o is a rectangular wave swinging from V_o^+ to V_o^- (Figure 13.12(b)) and the capacitor voltage v_c , which consists of the exponent pieces, changes from $V_o^+ \rho$ to $V_o^- \rho$ (Figure

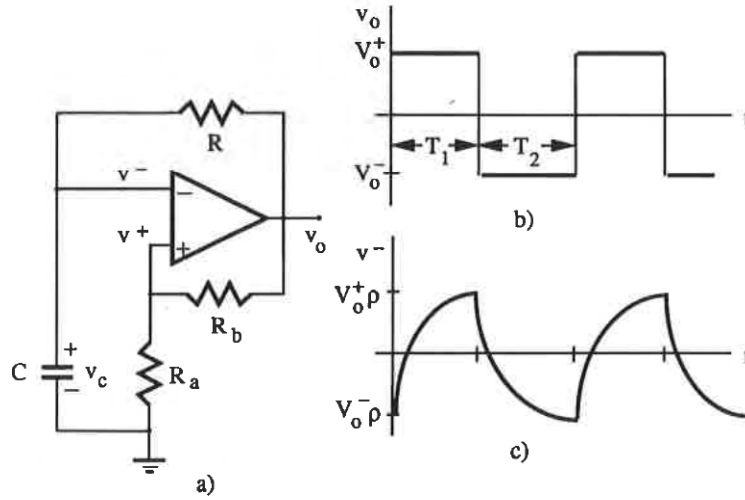


Figure 13.12 One-amplifier (comparator) multivibrator: (a) circuit; (b) output voltage; (c) OpAmp inverting input voltage.

13.12(c)). Here $\rho = R_a/(R_a + R_b)$. At high oscillation frequencies one has to take into consideration the operational amplifier finite slew rate. The output voltage can be approximated by a trapezoidal waveform (Figure 13.13(a)). The capacitor voltage amplitude increases (Figure 13.13(b)).

Assume that the voltage v_c has achieved the value of $V_o^+\rho$ and the voltage v_o starts to decrease. Then, for $0 < t < T_{1s} = (V_o^+ - V_o^-)/S_{rs}$, this voltage is

$$v_o(t) = V_o^+ - S_{rs}t \quad (13.69)$$

Solving the differential equation

$$v_c + \tau \frac{dv_c}{dt} = v_o \quad (13.70)$$

where $\tau = RC$. Using the initial condition $v_c(0) = V_o^+\rho$ one can find that

$$v_c = V_o^+[1 - \exp(-t/\tau)] + V_o^+\rho \exp(-t/\tau) - S_{rs}t + \tau S_{rs}[1 - \exp(-t/\tau)] \quad (13.71)$$

This result, for $0 < t < T_{1s}$, can be approximated as

$$v_c \approx V_o^+\rho + V_o^+(1 - \rho) \frac{t}{\tau} \quad (13.72)$$

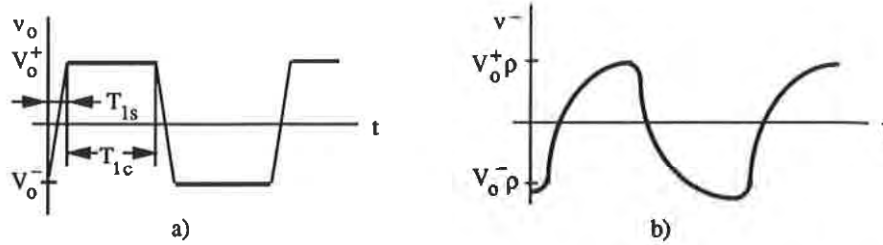


Figure 13.13 Slew-rate wave distortions in one comparator multivibrator: (a) slew-rate limited output voltage; (b) OpAmp inverting input voltage.

so that $v_c(T_{1s}) = V_o^+ \rho + [V_o^+ (1 - \rho)T_{1s}/\tau]$. And for $t > T_{1s}$ the output voltage is $v_o = V_o^-$.

Solving (13.70) again, with a new right side and new initial condition, one finds that

$$v_c = v_c(T_{1s}) + [V_o^- - v_c(T_{1s})]\{1 - \exp[-(t - T_{1s})/\tau]\} \quad (13.73)$$

From (13.73), using the final condition $v_c(T_{1s} + T_{1c}) = V_o^-$, one finds

$$T_{1c} = \tau \ln \left[\frac{V_o^- - v_c(T_{1s})}{V_o^- (1 - \rho)} \right] \quad (13.74)$$

Finally, for the symmetric conditions $V_o^+ = -V_o^- = V_o$, one finds that

$$T = 2(T_{1s} + T_{1c}) = 2T_{1s} + 2\tau \ln \left(\frac{1 + \rho}{1 - \rho} + \frac{T_{1s}}{\tau} \right) \approx \frac{4T_{1s}}{1 + \rho} + 2\tau \ln \left(\frac{1 + \rho}{1 - \rho} \right) \quad (13.75)$$

If the first term in (13.75) is omitted (as is usual for low oscillation frequencies), then

$$f_0 = \frac{1}{2\tau \ln \left(\frac{1 + \rho}{1 - \rho} \right)} \quad (13.76)$$

and, hence, the relative error introduced by the operational amplifier slew-rate limitation is

$$\frac{\delta f}{f_0} \approx -\frac{4f_0 T_{1s}}{(1 + \rho)} = -\frac{8V_o f_0}{S_{rs}(1 + \rho)} \quad (13.77)$$

13.3.5 Voltage-to-Frequency Converters

In the multivibrators considered so far it was assumed that the ratio $D = T_1/T_2$ is about 0.5. The factors that could cause D to differ from this value (like asymmetric output of the operational amplifiers) are considered parasitic. If D is modified on purpose to be strongly different from 0.5, the circuit can be adapted to be a voltage-to-frequency converter. The output voltage of these multivibrators is a train of narrow pulses having the frequency controllable by the voltage. The circuit is usually designed so that the width of these pulses is as narrow as possible; in this case a better linearity of the control characteristic is provided.

13.3.5.1 VFC with an Active Integrator

Such modifications can be easily done, for example, in the circuit of Figure 13.8. The modified circuit [34] is redrawn here in Figure 13.14. During the long part T_1 of the oscillation period, the switch P is open and the timing capacitor C is charged from the input voltage V_i . During the short part T_2 the switch P is closed and the timing capacitor is recharged from the voltage V_d^- . The integrator output voltage v_1 changes from $V_1^+ = V_o^+ \rho$ to $V_1^- = V_o^- \rho$, where $\rho = R_a/(R_a + R_b)$. The duration T_1 can be found from the charge balance equation:

$$T_1 \frac{V_i}{R} = C(V_1^+ - V_1^-) \quad (13.78)$$

The duration T_2 is calculated from the equation

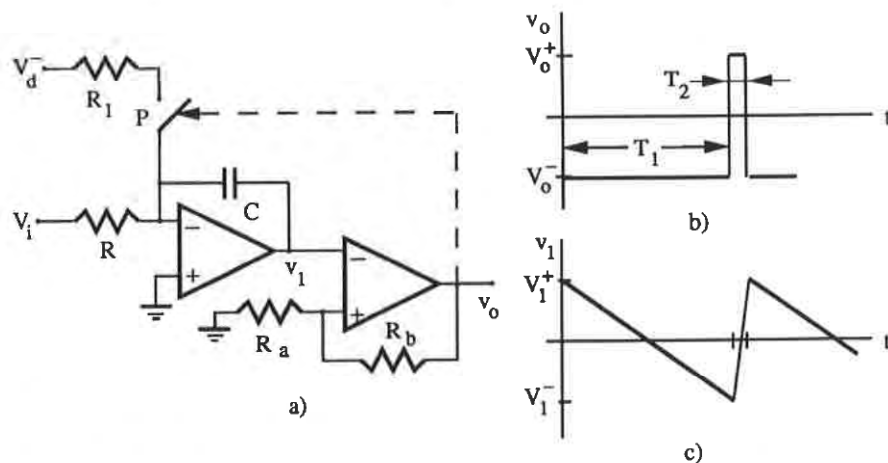


Figure 13.14 VFC with a timing capacitor recharge: (a) circuit; (b) integrator output; (c) VFC output.

$$T_2 \left(\frac{V_i}{R} + \frac{V_d^-}{R_1} \right) = C(V_1^- - V_1^+) \quad (13.79)$$

Then the oscillation frequency f can be found as

$$f = \frac{1}{T_1 + T_2} = \frac{V_i}{RC(V_1^+ - V_1^-)} \left[1 + \frac{V_i R_1}{V_d^- R} \right] \quad (13.80)$$

If the second term in the brackets of (13.80) is small and can be neglected, then the oscillation frequency

$$f_0 = \frac{V_i}{RC(V_1^+ - V_1^-)} \quad (13.81)$$

is proportional to the input voltage. When the input voltage increases, this term increases as well and determines the relative frequency error

$$\frac{\delta f}{f_0} = \frac{V_i R_1}{V_d^- R} \quad (13.82)$$

The additional errors caused by the slew-rate limitation and finite switching time are added (at the upper end of the frequency range) in the same fashion as in the circuit of Figure 13.8.

13.3.5.2 VFC with a Passive RC-Circuit and Full Discharge of the Timing Capacitor

The minimal value of the error term (13.82) is limited by the integrator maximal current. In addition, the auxilliary voltage source V_d^- should provide a high current for the capacitor recharge. These deficiencies are eliminated in the circuit shown in the Figure 13.15(a) [35–37]. Here the active integrator is replaced by a passive RC-circuit. A buffer is introduced between the integrator and the dynamic Schmitt trigger.

The circuit operation can be understood tracking the voltage v^+ (Figure 13.15(b)). During the time interval T_1 the switch P is open and the capacitor C is charging from the input signal V_i . The voltage v_1 follows the voltage at the capacitor C , and the values of R_a , R_b , and C_d are such that the voltage v^+ at the Schmitt trigger positive terminal follows v_1 as well. The capacitor C_d is small, and the initial transient for v^+ is short and even can be completely eliminated (the required parameter relationship is given later). When v^+ attains the value of V_{Ref} the output voltage v_o jumps to V_o^+ and the switch P will be closed. The voltage v_1 drops to 0, yet the voltage

v^+ becomes much higher than V_{Ref} because the voltage at C_d and V_o^+ are operating in series. Then the capacitor C_d starts to recharge and v^+ quickly diminishes. When v^+ drops to the value of V_{Ref} , the output voltage will drop to V_o^- , the switch will be open again, and the cycle will be repeated.

The durations T_1 and T_2 are calculated considering the voltages in the Schmitt trigger feedback circuit. When the switch P is open (Figure 13.15(d)) the buffer output voltage is

$$v_1 = V_i[1 - \exp(-t/RC)] \approx (V_i t)/(RC) \quad (13.83)$$

If C_d is small then

$$i_d = C_d \frac{dv_1}{dt} \approx \frac{V_i C_d}{RC} \quad (13.84)$$

At the end of the first quasi-stable state $v_1(T_1) = V_{\text{Ref}} + i_d R_a$; that is,

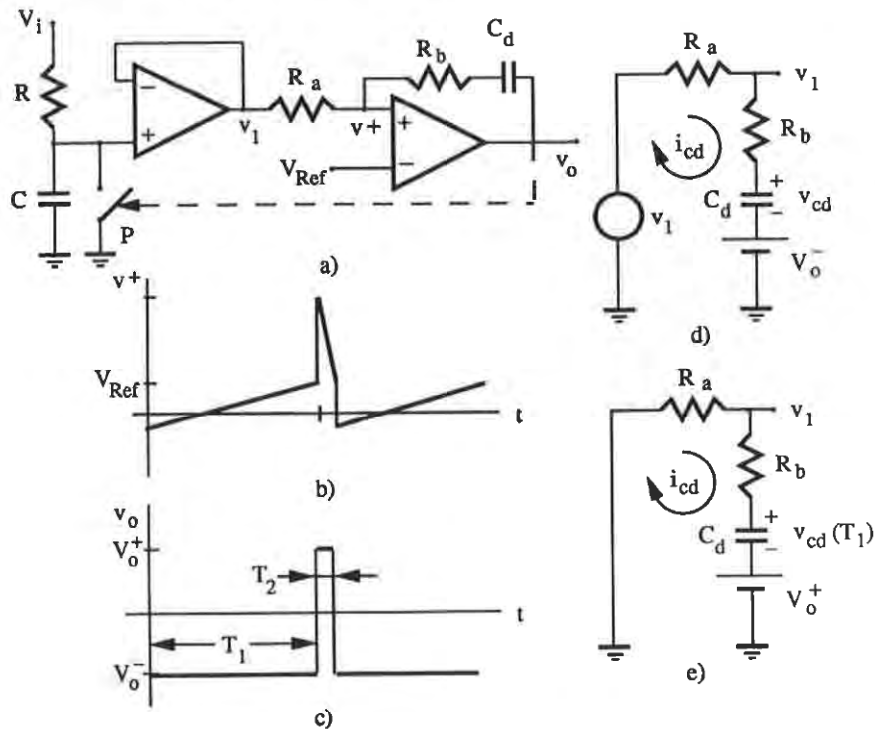


Figure 13.15 VFC with full discharge of the timing capacitor: (a) circuit; (b) comparator input; (c) VFC output; (d) C_d charge; (e) C_d discharge.

$$V_i[1 - \exp(-t/RC)] = V_{\text{Ref}} + V_i(R_a C_d/RC) \quad (13.85)$$

and from (13.85) one finds that

$$T_1 = RC \ln \left\{ \left[1 - \left(\frac{V_{\text{Ref}}}{V_i} + \frac{R_a C_d}{RC} \right) \right]^{-1} \right\} \approx RC \frac{V_{\text{Ref}}}{V_i} + R_a C_d \quad (13.86)$$

At the end of this part of the oscillation period, the voltage v_{cd} at the feedback capacitor becomes $v_{cd}(T_1) = V_{\text{Ref}} - V_o^- - (V_i R_b C_d/RC)$.

When the circuit switches in the second quasi-stable state, the voltage v^+ can be calculated from the circuit of Figure 13.15(e). First of all, it is seen that the condition

$$[V_o^+ + v_{cd}(T_1)]\rho > V_{\text{Ref}} \quad (13.87)$$

should be satisfied to obtain this quasi-stable state. Here $\rho = R_a/(R_a + R_b)$ as usual. Then one can write that

$$v^+ = [V_o^+ + v_{cd}(T_1)]\rho \exp\{-(t - T_1)/[C_d(R_a + R_b)]\} \quad (13.88)$$

for $T_1 \leq t \leq (T_1 + T_2)$, and the condition $v^+(T_1 + T_2) = V_{\text{Ref}}$ gives

$$T_2 \approx C_d(R_a + R_b) \left\{ 1 - \frac{V_{\text{Ref}}/\rho}{[V_o^+ - V_o^- + V_{\text{Ref}} - (V_i R_b C_d)/(RC)]} \right\} \quad (13.89)$$

The oscillation frequency that is

$$f = \frac{1}{T_1 + T_2} = \frac{V_i}{RCV_{\text{Ref}} + V_i(R_a C_d + T_2)} \quad (13.90)$$

includes, for this circuit, the error term increasing with V_i . Indeed, if one denotes

$$f_0 = \frac{V_i}{RCV_{\text{Ref}}} \quad (13.91)$$

then, as follows from (13.90), the relative frequency error is

$$\frac{\delta f}{f_0} \approx -f_0(R_a C_d + T_2) \quad (13.92)$$

Finally, at the end of the second quasi-stable state $v_{cd}(T_1 + T_2) = v_{cd}(0) = V_{\text{Ref}}[1 + (R_b)/(R_a)] - V_o^+$ and, if the condition $V_{\text{Ref}} = V_o^+ \rho$ is satisfied, the voltage v^+ follows the voltage at the timing capacitor without a transient in the first quasi-stable state.

13.3.5.3 VFC with a Passive RC-Circuit and Partial Discharge of the Timing Capacitor

The operation of the circuit [36] shown in Figure 13.16(a) is as follows. Here the voltage V_{Ref} is very small (it is necessary only to prevent the circuit negative latch-up) and the timing capacitor is recharged to a negative voltage when the switch P is closed for a short time by the Schmitt trigger feedback circuit. When this happens the voltage v^- at the negative terminal drops to

$$v^-(0) = V_{\text{Ref}} \frac{C}{(C + C_r)} - (V_o^+ - V_o^-) \frac{C_r}{(C + C_r)} \quad (13.93)$$

At this time the voltage at the feedback capacitor is $v_{cd}(0) = V_o^+ - V_{\text{Ref}}$. After the initial drop the timing capacitor C charges up and v^- slowly increases as

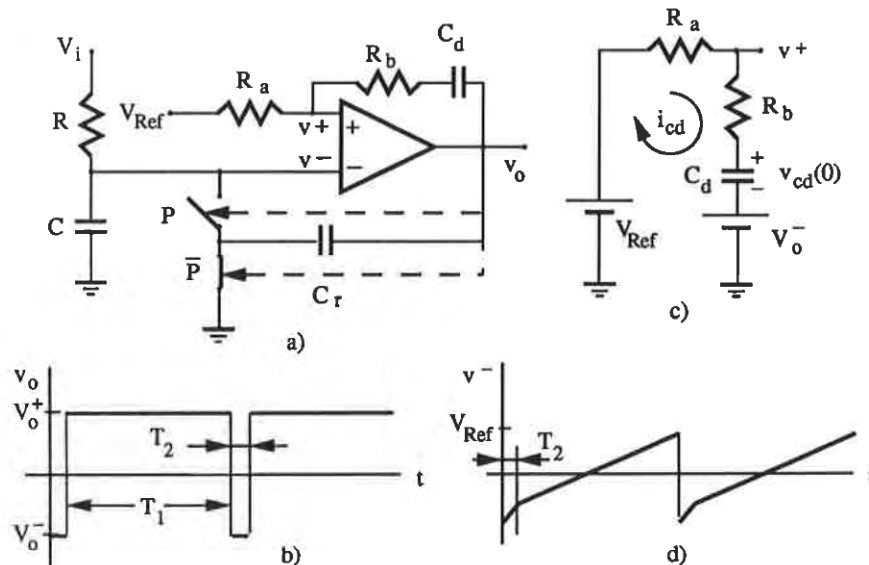


Figure 13.16 VFC with partial discharge of the timing capacitor: (a) circuit; (b) VFC output; (c) C_d charge; (d) comparator input.

$$v^- = V_i + [v^-(0) - V_i] \exp\{-t/[R(C + C_r)]\} \quad (13.94)$$

Simultaneously, the voltage at the positive terminal (which is calculated considering the feedback circuit of Figure 13.16(d)) decreases as

$$v^+ = [V_{\text{Ref}} - (V_o^+ - V_o^-)\rho] \exp\{-t/[C_d(R_a + R_b)]\} \quad (13.95)$$

When $v^+(T_2) = v^-(T_2)$, part T_2 of the oscillation period is finished. This gives

$$T_2 \approx (R_a + R_b)C_d \left[1 - \frac{C_r}{\rho C} \right] \quad (13.96)$$

Assuming that T_2 is small and using (13.94) with C instead of $C + C_r$ (C_r is not connected now) and the conditions $v^-(T_2) \approx v^-(0)$ and $v^-(T_1 + T_2) = V_{\text{Ref}}$, one obtains that

$$T_1 \approx \frac{(V_o^+ - V_o^-)RCC_r}{(V_o^+ - V_o^-)C_r + (V_i - V_{\text{Ref}})C} \quad (13.97)$$

Then the oscillation frequency

$$f = \frac{1}{T_1 + T_2} \approx \frac{(V_o^+ - V_o^-)C_r + (V_i - V_{\text{Ref}})C}{(V_o^+ - V_o^-)RCC_r + T_2(V_i - V_{\text{Ref}})C} \quad (13.98)$$

if the condition $RC \gg T_2$ is satisfied. If

$$f_0 = \frac{(V_i - V_{\text{Ref}})}{RC_r(V_o^+ - V_o^-)} + \frac{1}{RC} \quad (13.99)$$

then the relative frequency error will be

$$\frac{\delta f}{f_0} \approx -f_0 T_2 \quad (13.100)$$

13.3.6 Current-to-Frequency Converters

The previously described VFCs could be transformed into current-to-frequency converters (CFCs) if the voltage source V_i and resistor R were replaced by a current source charging the timing capacitor C . The capacitor discharge was realized via a

circuit of noncritical design, providing fast operation. When the oscillation frequency increases, the discharge time becomes comparable with the charge time and results in the nonlinearity of the control characteristic.

The CFCs considered in this section are based on a different approach. It is assumed here that the timing capacitor is periodically charged and discharged from two tightly matched current sources. The rest of the multivibrator circuit provides redirection of these currents. A transistor current source has wide current and frequency ranges [22,25,39], and redirection is realized by switches that can be very fast. The approach requires more complicated circuits but results in the circuits operating at high frequencies and having very linear control characteristics. The ratio $D = T_1/T_2$ of these multivibrators is normally about 1, making them highly suitable for phase-locked loop applications [22,31,32].

These multivibrators can be divided into two types: CFCs with a grounded timing capacitor and CFCs with a floating timing capacitor. Both types are characterized by a very broad range of control (four decades of the control current) and provide approximately the same results in terms of nonlinearity and stability of control characteristic.

13.3.6.1 Comparator Current-to-Frequency Converters

The floating timing capacitor CFCs that can be realized using comparators and complementary switches are shown in Figure 13.17. The condition of switches shown corresponds to the recharge of the timing capacitor from the polarity shown in parentheses to the polarity shown without parentheses.

At this time the output voltage of the comparator A_1 in Figure 13.17(a) is low (its negative input terminal has a potential higher than that of the positive terminal). The output voltage of A_2 is high. The input terminals of this comparator have constant voltage difference of V_c between them during all this recharging. The left voltage source absorbs both currents I and the capacitor C is recharging via the right current source. When the voltage v_c becomes equal to V_c (with the polarity shown without parentheses) the output voltage of A_1 becomes high and the condition of switches P_1 , P_1 changes. Even if P_2 is not yet open the output voltage of A_1 is preserved as high (especially if the resistor is chosen so that $2IR < V_c$) because the timing capacitor will continue to recharge until the switch P_1 is closed. When P_1 is, finally, closed, the voltage v_c becomes applied to the input terminals of the comparator A_2 , with the polarity providing an abrupt change of the output voltage of this comparator. As a result the switches P_2 , P_2 change their condition. The switching is finished when the switch P_2 is closed (thus, the correct sequence of switch condition changing is P_1, P_1, P_2, P_2). Now one has to consider the second, completely symmetric, part of the oscillation cycle.

In ideal comparators and ideal switches, the capacitor is recharged from $+V_c$ to $-V_c$ by current I . Hence, the duration of a semiperiod will be $T/2 = (2V_c)/I$, and the oscillation frequency is

$$f_0 = \frac{I}{4CV_c} \quad (13.101)$$

and can be linearly controlled by the matched current sources.

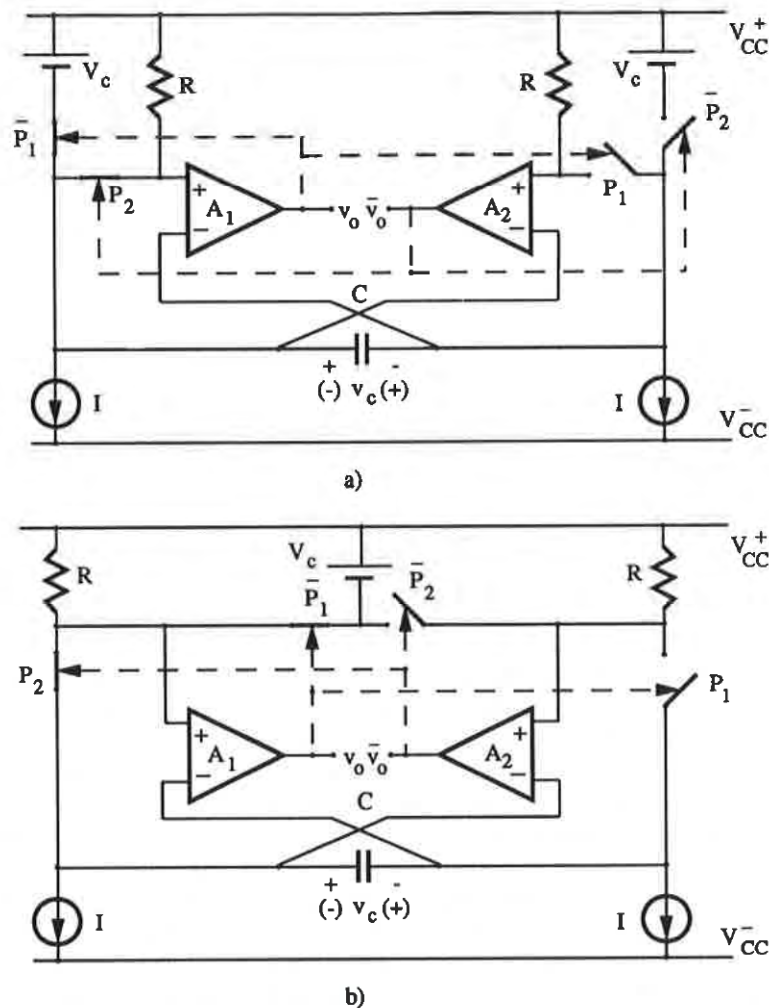


Figure 13.17 Comparator CFCs with: (a) two reference voltages; (b) one reference voltage.

In normal operation of this CFC both voltage sources V_c are never used at the same time. This allows one to modify this circuit into the circuit shown in Figure 13.17(b). The operation of this circuit follows directly from the previous description and the oscillation frequency is given by the (13.101) as well.

The main source of the frequency error in both circuits is the slew-rate limitation of A_1 when P_1 is open and P_1 is not closed yet, and an additional charge is deposited on the timing capacitor and the slew-rate limitation of A_2 when P_2 is still closed and P_2 is still open, and the timing capacitor is discharged via both resistors but not by a current source. If the combined delay introduced by the comparator and two connected complementary switches is denoted T_s , the relative frequency error will be

$$\frac{\delta f}{f_0} \approx -2f_0 T_s \quad (13.102)$$

The additional increase of the v_c voltage swing can be important for some low-voltage circuits.

13.3.6.2 Emitter- and Source-Coupled Multivibrators

The emitter-coupled multivibrator [22,25,40] shown in Figure 13.18(a) is, probably, the first circuit used as a current-to-frequency converter with a wide range (at least, four decades) of linear frequency control. It is similar to the circuit of Figure 13.17. Indeed, the diodes Q_5 and Q_6 play the same role as the voltage sources in that circuit. In addition, they are self-switching, which allows one to have only two additional switches, so that Q_1 and Q_2 are sufficient.

Referring to the circuit of Figure 13.18, the frequency of oscillations can be determined [22] as

$$f_0 \approx \frac{I}{4C V_{be(ON)}} \quad (13.103)$$

Here this result is denoted as an approximate one. Indeed, during the switching from one quasi-stable state to another the circuit has a positive feedback loop and the loop gain depends on the current I . The voltage drop at Q_5 and Q_6 changes as well. The approximate values of $V_c(0) = -V_c(T_1) = V_{be(ON)} = 0.7V$, which, used in the (13.103), results in a large error for f_0 . A more exact result without the overcomplication of the calculations requires the following assumptions [40].

- The exact values $V_c(0) = -V_c(T_1)$ of v_c at the instant of transition from one quasi-stable state to another are most important in the calculation of the oscillation frequency.

- One can consider that recharging of the timing capacitor C from $V_c(0)$ to $V_c(T_1)$ and vice versa is obtained with a constant current I .
- One can assume that the transition from one quasi-stable state to another is an instantaneous process.

To obtain the exact values of $V_c(0)$ and $V_c(T_1)$ the multivibrator of Figure 13.18, during a short period of time near the transition from one quasi-stable state to another, is considered a linear circuit. Then, the loop transfer function for this linear small-signal equivalent circuit is calculated. Equating this transfer function to unity one finds the exact values of the collector current I_{c1} and I_{c2} at the instant of transition. Finally, using these currents one calculates $V_c(0)$ and $V_c(T_1)$.

The simplified small signal equivalent linear circuit describing the multivibrator during the transition is shown in Figure 13.18(b). The duration of a quasi-stable state is much longer than that of a transition, so the voltage on the coupling capacitor practically does not change during this transition, and the capacitor is represented

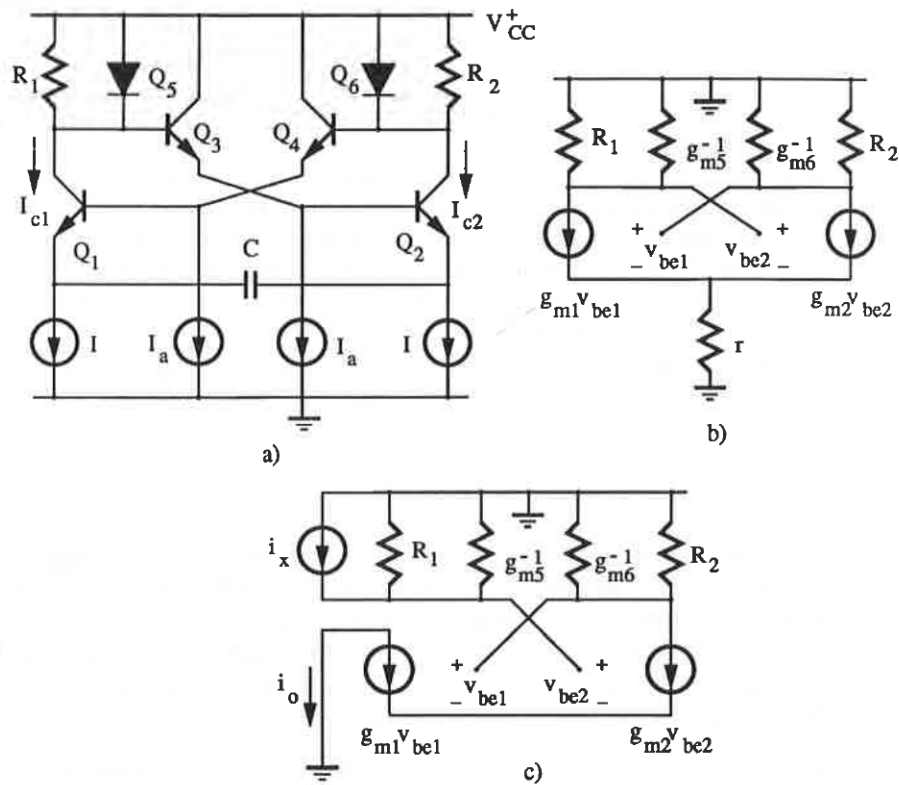


Figure 13.18 Emitter-coupled multivibrator: (a) circuit; (b) small-signal equivalent; (c) open loop.

as a short circuit in the model. Transistors Q_3 and Q_4 of the Figure 13.18(a) are linear buffers and they are not included in this equivalent circuit. The circuit with $R_1 = R_2 = R$ will be considered but the notations R_1 and R_2 are preserved for convenience. The resistor r is introduced to avoid infinite nodal impedance (it can be considered as representing the output resistance of two current sources in the emitter circuits of Q_1 and Q_2 connected in parallel by the shorting coupling capacitor and the assumption $r \rightarrow \infty$ in the final result does not introduce any noticeable error).

The loop transfer function $A_L = i_o/i_x$ can be easily obtained from the circuit shown in Figure 13.18(c) (r is omitted). It is

$$A_L = \frac{g_{m1}g_{m2}R_1}{(1 + g_{m5}R_1)[g_{m1} + g_{m2} - g_{m1}g_{m2}R_2(1 + g_{m6}R_2)^{-1}]} \quad (13.104)$$

At the instant of transition $A_L = 1$, and this gives the condition of transition

$$g_{m1} + g_{m2} = g_{m1}g_{m2}[R_1(1 + R_1g_{m5})^{-1} + R_2(1 + R_2g_{m6})^{-1}] \quad (13.105)$$

In the considered transition, Q_1 is on and changes its state to off. Correspondingly, Q_2 is off and turns on (the currents in the opposite transition are obtained by interchanging the I_{c1} and I_{c2} values). The transition starts when the collector current of Q_1 decreases from the value of $2I$ to that of $2I - \Delta I$ and the collector current of Q_2 increases from 0 to ΔI . At this instant the transconductances will be $g_{m1} = I_{c1}/V_T = (2I - \Delta I)/V_T$ and $g_{m2} = I_{c2}/V_T = \Delta I/V_T$ (here $V_T = 26$ mV at the temperature 300K). These values should be substituted in (13.105) to obtain the equation for calculation of ΔI .

The diodes Q_5 and Q_6 are clamping the base of Q_3 and Q_4 . Then $(2I - \Delta I) > V_{be(ON)}$ and the currents in Q_5 and R_1 are comparable. In this case

$$g_{m5} = \frac{2I - \Delta I - [V_{be(ON)}/R_1]}{V_T} \quad (13.106)$$

When Q_1 is on, the normal circuit operation requires that the condition $V_{R2} < V_{be(ON)}$ be satisfied up to and including the instant of transition (that is, the diode Q_6 is off and one has to set $g_{m6} = 0$ in (13.105)). If all these transconductance values are substituted in (13.107), one obtains

$$\frac{2IV_T}{R} = (2I - \Delta I)\Delta I \frac{[2V_T + R(2I - \Delta I) - V_{be(ON)}]}{[V_T + R(2I - \Delta I) - V_{be(ON)}]} \quad (13.107)$$

Solving (13.107) numerically one can obtain the value of ΔI . If the term $(\Delta I)^3$ is neglected, a quadratic equation gives the approximate value of

$$\Delta I \approx \frac{V_T}{R} \left[1 - \frac{2V_T}{2IR - V_{be(ON)}} \right] \quad (13.108)$$

Then the coupling capacitor voltage V_c can be found from

$$V_{R1} + V_{be3} + V_{be2} - V_c - V_{be1} - V_{be4} - V_{R2} = 0 \quad (13.109)$$

Transistors Q_3 and Q_4 have equal emitter currents, and $V_{be3} = V_{be4}$. Also, in the considered transition point $V_{R2} = I_{c2}R_2 = \Delta IR_2$ and $V_{R1} = I_{c1}R_1 = (2I - \Delta I)R_1$. Then one finds from (13.109) that

$$V_c = (2I - \Delta I)R_1 - \Delta IR_2 - V_T \ln \left(\frac{2I}{\Delta I} - 1 \right) \quad (13.110)$$

In the opposite transition (Q_2 is *on* and turns *off*, Q_1 is *off* and turns *on*) the capacitor C will have the voltage $-V_c$ on it (with $R_1 = R_2 = R$) and so the total voltage swing will be $\Delta V_c = 2V_c$. The recharge time of C , of course, increases when the current is deflecting into Q_2 during the transition. But the calculations show that the maximum value of deflected current ΔI is so small that this increase can be neglected and the charge balance equation can be written as $\Delta V_c C = (IT)/2$, where T is the oscillation period. Then the oscillation frequency is

$$f = \frac{I}{4C \left\{ V_{be(ON)} - V_T \left[\ln \left(\frac{2IR}{V_T} + \frac{4IR}{2IR - V_{be(ON)}} \right) + 1 - \frac{2V_T}{2IR - V_{be(ON)}} \right] \right\}} \quad (13.111)$$

This is the required equation of the frequency control characteristic. It shows that

$$\frac{\delta f}{f_0} \approx \frac{V_T}{V_{be(ON)}} \left[\ln \left(\frac{2IR}{V_T} \right) + 1 \right] \quad (13.112)$$

which varies very slowly when I increases. Equation (13.112) allows one to predict the nonlinearity for a given range of the control current I . It also can be used to evaluate the thermostability of the characteristic and the required compensating means in the bias circuit.

The diode clamping is essential for this circuit to be a current-to-frequency converter. If the transistors Q_5 and Q_6 are absent, the calculation procedure is the same except one has to use $g_{m5} = g_{m6} = 0$ from the onset of calculations [40]. The result will be

$$f = \frac{1}{8RC \left\{ 1 - \frac{V_T}{2IR} \left[\ln \left(\frac{4IR}{V_T} \right) + 1 \right] \right\}} \quad (13.113)$$

That is, the circuit becomes a fixed-frequency multivibrator (the variation of the second term in the denominator is small).

The circuit of Figure 13.17(b) can be also used as a prototype to develop an emitter-coupled multivibrator [41]. The calculation of the oscillation frequency follows the same steps. One of the most precise circuits [42] based on this model is shown in Figure 13.19. In this circuit the voltage source V_c is removed from the path of currents (this path is closed via Q_5 and Q_6). The design key point in this circuit is four matched current sources operating in the emitter and collector circuits of Q_1 and Q_2 and six matched transistors (Q_5 , Q_4 , and the transistors realizing D_1 to D_4). If the last requirement is satisfied, the voltage appearing between the collectors of Q_1 and Q_2 does not depend on the current I , the temperature, the exact value of $V_{be(ON)}$, and the power supply voltage. The voltage at the timing capacitor changes from $-V_c$ to $+V_c$ during one half-cycle and the oscillation frequency is given by (13.101) again. The oscillation frequency stability becomes dependent on the V_c realization and its matching with charging current sources.

The source-coupled multivibrator [44] shown in Figure 13.20(a) represents the CMOS version of the previously considered multivibrator. The frequency calculation

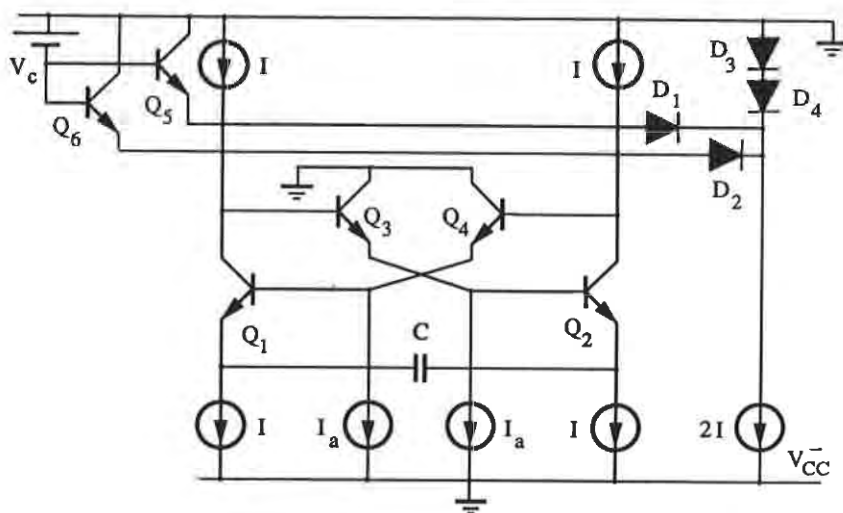


Figure 13.19 Modified emitter-coupled multivibrator.

for this circuit is based on the same assumptions. Yet a profound difference exists between the CMOS multivibrator and its bipolar prototype. This difference is due mainly to the body effect of n -channel transistors. As a result the frequency versus the control current I of the multivibrator becomes nonlinear.

During the short period just before the discontinuous change of the drain currents, the multivibrator can be described by the simplified small-signal equivalent circuit shown in Figure 13.20(b). Investigation of the loop transfer function near a transition point brings up the condition of transition

$$g_{m1} + g_{m2} = g_{m1} g_{m2} R_{oe} \quad (13.114)$$

where $R_{oe} = [R(2 + Rg_{m3})]/(1 + Rg_{m3})$. The condition (13.114) is used to find the drain currents I_{d1} and I_{d2} at this instant. If the transition in which M_1 is on and turns off is considered, then I_{d2} is small and the current through the parallel connection of R and M_3 changes very little up to the instant of discontinuous change ($I_{d1} \approx 2I$). The assumption that R_{oe} is constant can be used to simplify the calculations. To calculate R_{oe} itself, one has to find

$$g_{m3} = 2\beta_p(V_{DD} - V_{g2} - |V_{tp0}|) \quad (13.115)$$

where

$$V_{g2} \approx V_{DD} - |V_{tp0}| - \frac{\sqrt{1 + 4\beta_p R(2IR - |V_{tp0}|)} - 1}{2\beta_p R} \quad (13.116)$$

Here $|V_{tp0}|$ is the threshold voltage for p -channel transistors with the source and substrate connected together. Equation (13.115) assumes that the transistors are described by the square law characteristic, where β_p (or β_n for n -channel transistors) is the proportionality coefficient. Further discussion of β_n and β_p , which should be used for calculations, will be given later. It is also assumed that the transistors are matched in pairs (thus, $\beta_{p3} = \beta_{p4} = \beta_p$), and the voltage swing at the capacitor C is small enough to neglect the difference in the source voltages of M_1 and M_2 (thus, $\beta_{n1} \approx \beta_{n2} \approx \beta_n$). The last assumption allows us to write that $g_{m1} = 2\sqrt{(\beta_n I_{d1})}$ and $g_{m2} = 2\sqrt{(\beta_n I_{d2})}$. If these values are substituted in (13.114) one finds that

$$I_{d1} + I_{d2} = 2R_{oe}\sqrt{\beta_n I_{d1} I_{d2}} \quad (13.117)$$

This equation should be solved together with the condition $I_{d1} + I_{d2} = 2I$, which follows from the circuit topology.

Once the currents I_{d1} and I_{d2} are found, one can calculate the gate voltages

$$V_{g1} = V_{d2} = V_{DD} - I_{d2}R \quad (13.118)$$

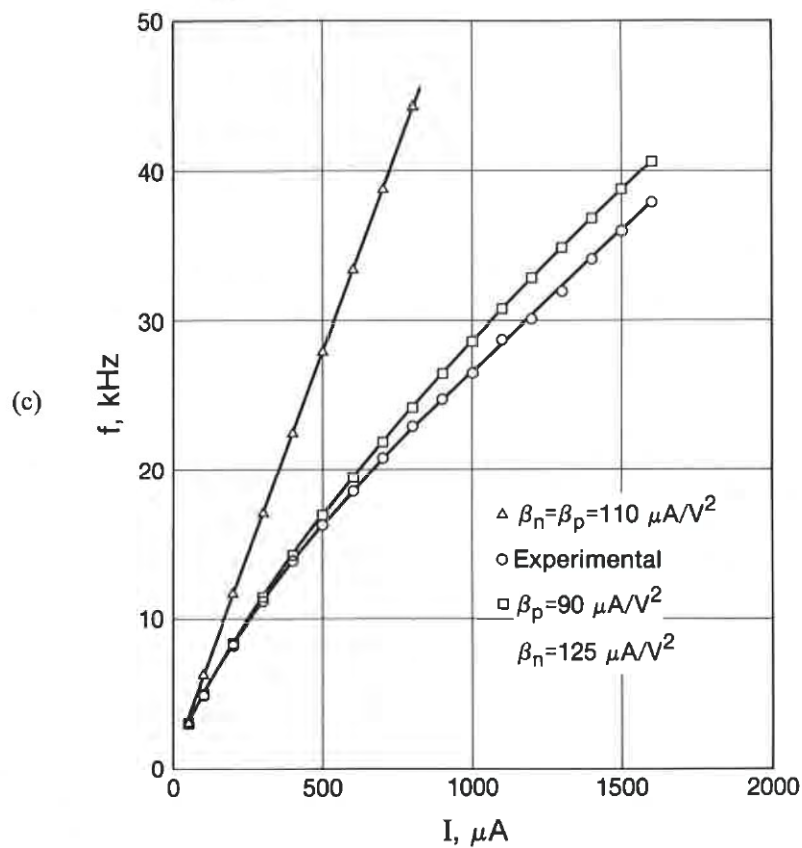
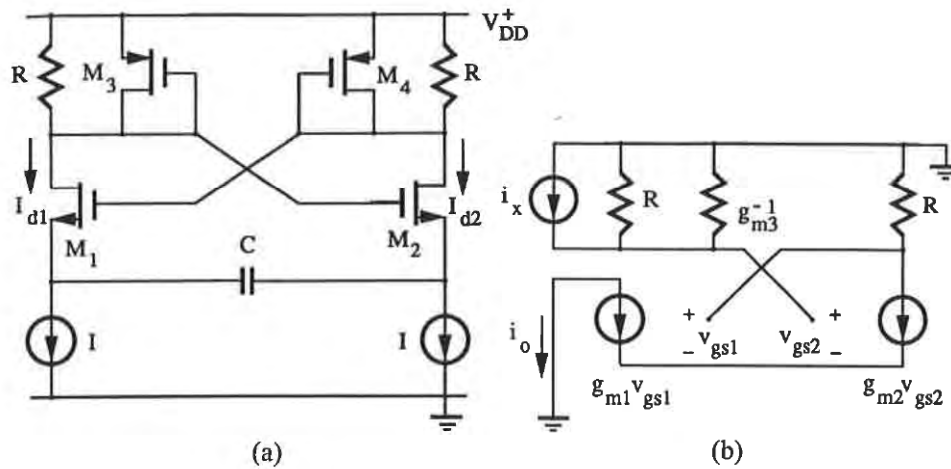


Figure 13.20 Source-coupled multivibrator: (a) circuit; (b) open loop; (c) control characteristic.

$$V_{g2} = V_{d1} = V_{DD} - |V_{tp0}| - \frac{\sqrt{1 + 4\beta_p R(RI_{d1} - |V_{tp0}|)} - 1}{2\beta_p R} \quad (13.119)$$

Then the source voltages are obtained from the equation

$$V_{si} = \frac{1}{2} (2V_{ni} + \gamma^2 - \gamma\sqrt{\gamma^2 + 4V_{ni} + 8|\phi_f|}) \quad (13.120)$$

where $V_{ni} = V_{gi} + 2|\phi_f| - V_{tn0} - \sqrt{(I_{di}/\beta_n)}$ and $i = 1, 2$.

In (13.119) and (13.120), γ is the body-effect coefficient, $|\phi_f|$ is the Fermi voltage, and V_{tn0} is the threshold voltage for n -channel transistors with the source and substrate connected together. The results (13.119) and (13.120) follow directly from the basic relationships of the MOS transistor [45,46] rewritten as

$$I_{di} = \beta_n (V_{gi} - V_{si} - V_{tni})^2$$

(hence, here the coefficients β_n and β_p are twice larger than in the frequently used notation [46]) and

$$V_{tni} = V_{tn0} + \gamma [\sqrt{(2|\phi_f| - V_{si})} - \sqrt{(2|\phi_f|)}]$$

for each transistor.

When V_{s1} and V_{s2} are found and the pinch-off operation of the devices is confirmed ($V_{dsi} \geq V_{gsi} - V_{tni}$, $i = 1, 2$), the voltage amplitude at the capacitor is calculated as

$$V_c = V_{s1} - V_{s2} \quad (13.121)$$

Then the frequency of oscillations is $f = I/(4CV_c)$ as usual, and the linear control characteristic will be obtained if V_c is constant.

There is no close formula, and the frequency control characteristic can be calculated point by point only. Yet it can be approximately analyzed [44]. This analysis shows that

$$V_c \approx |V_{tp0}| + \sqrt{\frac{I_{d2}}{\beta_n}} + \sqrt{2I} \left(\frac{1}{\sqrt{\beta_p}} - \frac{1}{\sqrt{\beta_n}} \right) \quad (13.122)$$

The second term is usually small and can be neglected. It is seen from (13.122) that V_c can be nearly constant if one can match the transistor parameters so that β_p and β_n are equal. The dependence (even though weak) of β_p and β_n on the level of the operating current I and source-substrate voltage makes the matching impossible over a wide range of operating current.

Figure 13.20(c) shows the results of the frequency calculations when the constant values of $\beta_n = \beta_p = 110 \text{ mA/V}^2$ are used (MO415 CMOS transistor array [47], the measurements at the drain current of 50 mA and $V_{sb} = 0$ for each device). The experimental results show strong divergence with the calculated ones, as the operating current increases. As was mentioned already, the main cause of this is the voltage and current dependence of the MOS transistor gain. In a considered circuit this effect is most pronounced for n -channel transistors. If one relies on the average value of β_n , which is in this case about 125 mA/V^2 , the error between the experimental and calculated results is not more than 10% (Figure 13.20(c)). This chosen average value provides the coincidence of all dc levels within the quasi-stable states and the slope of the V_c versus $\sqrt{I}$ characteristics (the calculated slope is $22 \text{ V/mA}^{0.5}$, the experimental one is $20 \text{ V/mA}^{0.5}$ for the considered devices) within the same error band. Yet, the control characteristic is, of course, nonlinear.

13.3.6.3 Schmitt-Trigger-Type Current-to-Frequency Converters

The previously considered CFCs used a floating timing capacitor and two matched current sources. In some integrated circuits it is highly desirable to have one plate of the capacitor grounded. This requirement resulted in the CFCs using the Schmitt trigger and different switching circuits providing redirection of the charging current. The circuits can be divided into three groups.

In the first group of circuits (Figure 13.21(a)) the timing capacitor is charged by the current source I from the top, and when the voltage v_c attains the value of V_{ch} the trigger turns off the current source $\bar{I}$, opens the switch P , closes the switch $\bar{P}$ and turns on the complementary current source I . Now the timing capacitor is discharged, and when v_c attains the value of V_{cl} , the trigger turns off the current source I and the switching operations are repeated in the inverse order.

In the second group (Figure 13.21(b)) the current source I permanently charges the timing capacitor from the top and the trigger redirects the current source $2I$. During the timing capacitor discharge both current sources are connected to the timing capacitor, and it is discharged by the difference of two currents.

Finally, the third group (Figure 13.21(c)) has only one current source, which is redirected by the Schmitt trigger and two complementary switches. When the switch $\bar{P}$ is on, the current is directed into current mirror and charges the timing capacitor from the top; when the switch is on, the current source is directly connected to the capacitor and discharges it.

The oscillation frequency is

$$f_0 = \frac{I}{2C(V_{ch} - V_{cl})} \quad (13.123)$$

for all three groups of circuits.

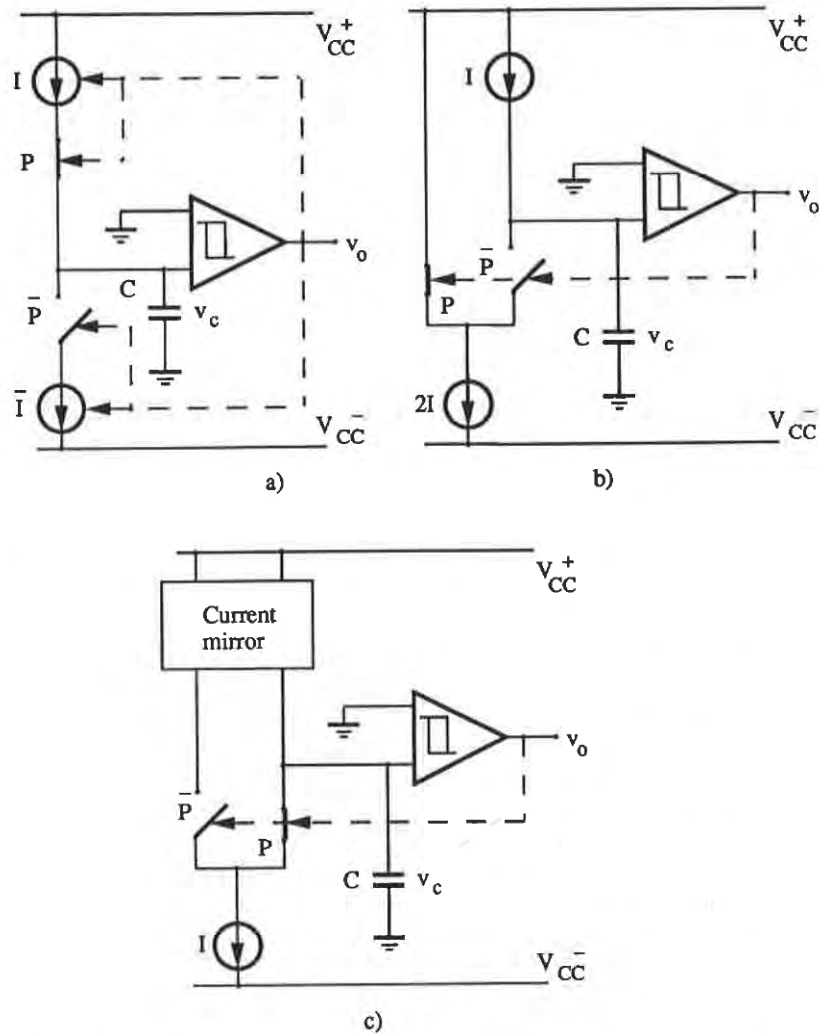


Figure 13.21 CFCs with a Schmitt trigger: (a) complementary sources; (b) I and $2I$ sources; (c) I and a current mirror.

In the circuits of the first group, the switching and trigger subcircuits are operating in series. These circuits are suitable for applications in sensors measuring slow variables (temperature, static fields) [48,49]. The trigger can be realized using a comparator and two-resistor positive feedback. When the currents I and $\bar{I}$ are changing simultaneously one obtains a CFC. The relative frequency error for these converters can be estimated using (13.102), where T_s is the combined delay introduced by the

comparator and the switch. If I and $\bar{I}$ are changing separately but their sum is preserved, one obtains a duty-cycle modulation.

In the second and third groups the switching and trigger subcircuits are driven in parallel. This results in faster circuits, capable of operating at higher frequencies. In addition, these CFCs are structurally divided into two independent subcircuits; namely, the threshold circuit and the charge-discharge circuit. This makes the design more flexible and provides an abundance of circuits adapted for different design conditions. The charge-discharge circuit includes two complementary switches that only redirect the current ($2I$ or I). Nearly all designs use a differential pair, which is able to provide fast switching of its tail current. The Schmitt trigger can be realized, using an operational amplifier in this case as well. The fast converters rely on more simple structures of the Schmitt triggers, also based on switching a differential pair. The whole circuit becomes a nonsaturated multivibrator. The circuits corresponding to the second group allow the widest range of the control current because they can be realized (in bipolar technology) using n - p - n transistors only [50]. The circuits of the third group are slower (the bipolar p - n - p integrated transistors have inferior frequency properties), but their bipolar design can be topologically transferred into CMOS technology (the additional limitations are minor).

The frequency calculation for the circuits of the second and third groups is very similar to that for emitter-coupled multivibrators. In accordance with the first assumption, the upper V_{CH} and lower V_{CL} thresholds of the Schmitt trigger have to be calculated precisely via the currents in the trigger circuit at the instants of transitions. Then one can neglect small variations in the charge and discharge current before the transition from one quasi-stable state to another and consider the transition as an instantaneous process.

As a representative example the circuit with a current mirror feedback Schmitt trigger (Figure 13.22(a)) is considered. The timing capacitor C is charged and discharged by the current source I (the control current), which is redirected by the trigger output voltage v_o via the switching differential pair Q_3, Q_4 , and the current mirror Q_5, Q_6 . The trigger circuit includes the differential amplifier Q_1, Q_2 with positive feedback via the current mirror Q_7, Q_8 and the divider R_1, R_2 .

As in all of the previously considered circuits, before a jump from one quasi-stable state to another this CFC operates as a linear circuit (when both Q_1 and Q_2 are conducting) with positive feedback. The equivalent small-signal circuit for this part of operation is shown in Figure 13.22(b). Investigating the loop transfer function of the circuit at a transition point, one obtains the condition of jump

$$g_{m2} = r_{\pi 2}^{-1} + \frac{1}{R_{oe}} \left(1 + \frac{r_{\pi 2}^{-1} + g_{m2}}{r_{\pi 1}^{-1} + g_{m1}} \right) \quad (13.124)$$

where $R_{oe} = R_1 R_2 / (R_1 + R_2)$. Taking into consideration [25] that $r_{\pi} \approx \beta / g_m$ and, hence, $r_{\pi}^{-1} \ll g_m$ the condition (13.128) can be simplified to

$$g_{m1}^{-1} + g_{m2}^{-1} \approx R_{oe} \quad (13.125)$$

Substituting $g_{m1} = I_{c1}/V_T$ and $g_{m2} = I_{c2}/V_T$ into (13.129), one finds that

$$I_{c1}^{-1} + I_{c2}^{-1} = R_{oe} V_T^{-1} \quad (13.126)$$

The condition $I_{c1} + I_{c2} = I_a$ is also valid. Finding I_{c1} and I_{c2} from (13.126) and this last condition and using [25] $V_{be i} = V_T \ln(I_{c i}/I_s)$ for $i = 1, 2$, one finds further that the voltage between bases of Q_1 and Q_2 at the instant of switching is

$$\Delta V = V_{be1} - V_{be2} = \pm V_T \ln m \quad (13.127)$$

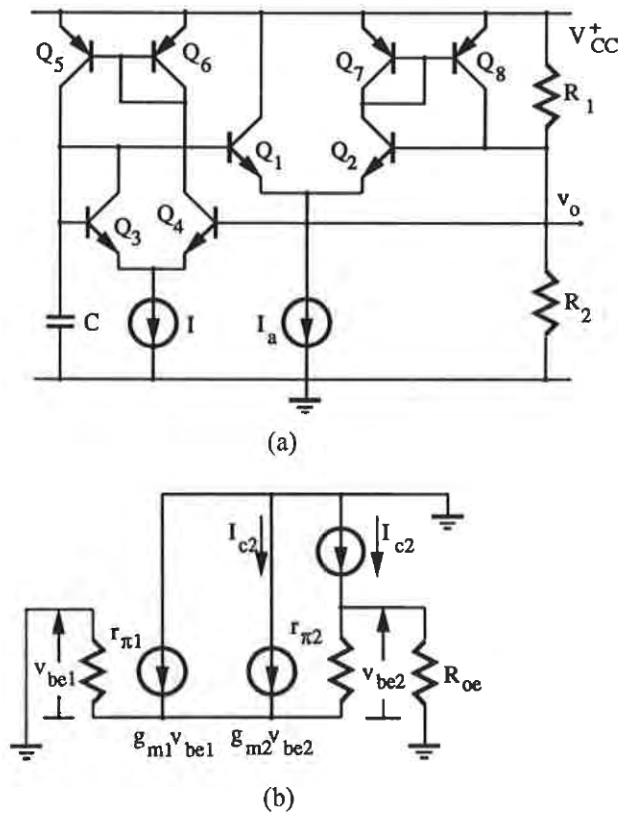


Figure 13.22 Bipolar CFC with a current-feedback Schmitt trigger: (a) circuit; (b) small-signal equivalent; (c) control characteristics.

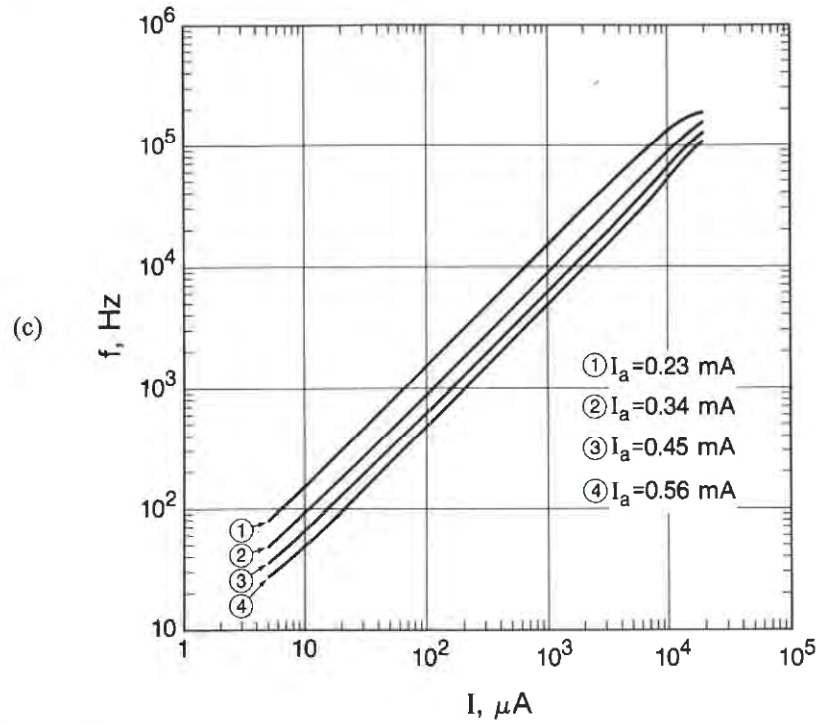


Figure 13.22 continued.

where $m = \{1 + \sqrt{[1 - (4V_T)/(I_a R_{oe})]}\} / \{1 - \sqrt{[1 - (4V_T)/(I_a R_{oe})]}\}$. Now one can calculate

$$V_{cL} = \frac{V_{CC}R_2}{R_1 + R_2} + \Delta V + I_{c2}R_{oe} \quad (13.128)$$

Substituting the value of ΔV with upper sign one finds that

$$V_{cL} \approx \frac{V_{CC}R_2}{R_1 + R_2} + V_T(1 + \ln m) \quad (13.129)$$

where one takes into consideration that m is usually large. Then one finds

$$V_{cH} = \frac{V_{CC}R_2}{R_1 + R_2} + \Delta V + I_{c2}R_{oe} - R_{oe} \frac{I_a}{\beta_{F2}} \quad (13.130)$$

where ΔV has to be taken from (13.127) with the lower sign. Here the term representing the base current of transistor Q_2 is included. Indeed, at high values of I_a , this base current reduces the value of the voltage v_o and, correspondingly, V_{ch} . The calculations gives

$$V_{ch} \approx \frac{V_{CC}R_2}{R_1 + R_2} + I_a R_{oe} - V_T(1 + \ln m) - R_{oe} \frac{I}{\beta_4} \quad (13.131)$$

Now the peak-to-peak voltage at the timing capacitor can be calculated as

$$V_c = V_{ch} - V_{cl} = I_a R_{oe} - 2V_T(1 + \ln m) - R_{oe} \frac{I}{\beta_4} \quad (13.132)$$

If the effect of variation in the charging current during the transition periods from one quasi-stable state to another is neglected, then the timing capacitor is charged, in accordance with this assumption, by the current I so that the charge time is $T_1 = CV_c/I$. The discharging current is $I + I_a/\beta_1$ and the discharge time is $T_2 = CV_c/(I + I_a/\beta_1)$, where the effect of the base current of transistor Q_1 is included. Thus, the oscillation frequency

$$f \approx \frac{I}{C \left[I_a R_{oe} - 2V_T(1 + \ln m) - R_{oe} \frac{I}{\beta_4} \right] \left(2 - \frac{I_a}{I\beta_1} \right)} \quad (13.133)$$

If one denotes

$$f_0 = \frac{I}{2CR_{oe}I_a} \quad (13.134)$$

which shows that this CFC has linear control, then the relative frequency error will be

$$\frac{\delta f}{f_0} \approx \frac{2V_T(1 + \ln m)}{I_a R_{oe}} + \frac{I}{I_a \beta_4} + \frac{I_a}{2I\beta_1} \quad (13.135)$$

Using standard bipolar technology one can obtain the circuit with linear current control characteristic (Figure 13.22(c)) embracing three decades of control current ($R_1 = 5.09 \text{ k}\Omega$, $R_2 = 3.89 \text{ k}\Omega$, $V_{CC} = 3\text{V}$). The full-scale nonlinearity is less than 2% if the trigger current I_a is chosen to optimize the performance. At high values

of the control current, the circuit performance deteriorates due to a low f_T value (about 3 MHz) for integrated $p-n-p$ transistors in standard technologies.

The CMOS multivibrator with a current feedback CMOS Schmitt trigger (Figure 13.23) repeats the structure of its bipolar prototype. The analysis, with minor modifications that are clear from the previous section, follows the same steps. The instant of switching is characterized [52] by the voltage

$$\Delta V = V_{gs1} - V_{gs2} = \pm \sqrt{\frac{I_a}{\beta_{nt}}} (1 - q) \quad (13.136)$$

Here $\beta_{nt} = (\mu C_{ox} W_1)/(2L_1) = (\mu C_{ox} W_2)/(2L_2)$ (that is, it is assumed that M_1 and M_2 are matched), and $q = [\sqrt{(1 + 4\beta_{nt} I_a R_{oe})} - 1]^{-1}$ and is small.

The timing capacitor voltage v_c changes between its highest value of

$$V_{cH} \approx \frac{V_{DD} R_2}{R_1 + R_2} + I_a R_{oe} + (1 - q) \sqrt{\frac{I_a}{\beta_{nt}}} \quad (13.137)$$

and its lowest value of

$$V_{cL} \approx \frac{V_{DD} R_2}{R_1 + R_2} + (1 - q) \sqrt{\frac{I_a}{\beta_{nt}}} \quad (13.138)$$

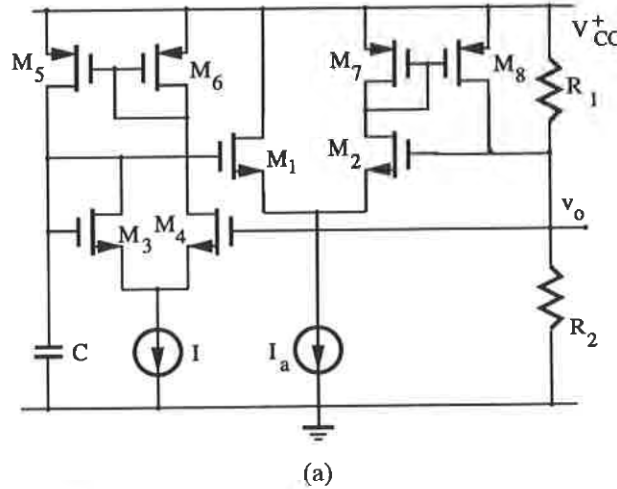


Figure 13.23 CMOS CFC with a current feedback Schmitt trigger: (a) circuit; (b) control characteristics.

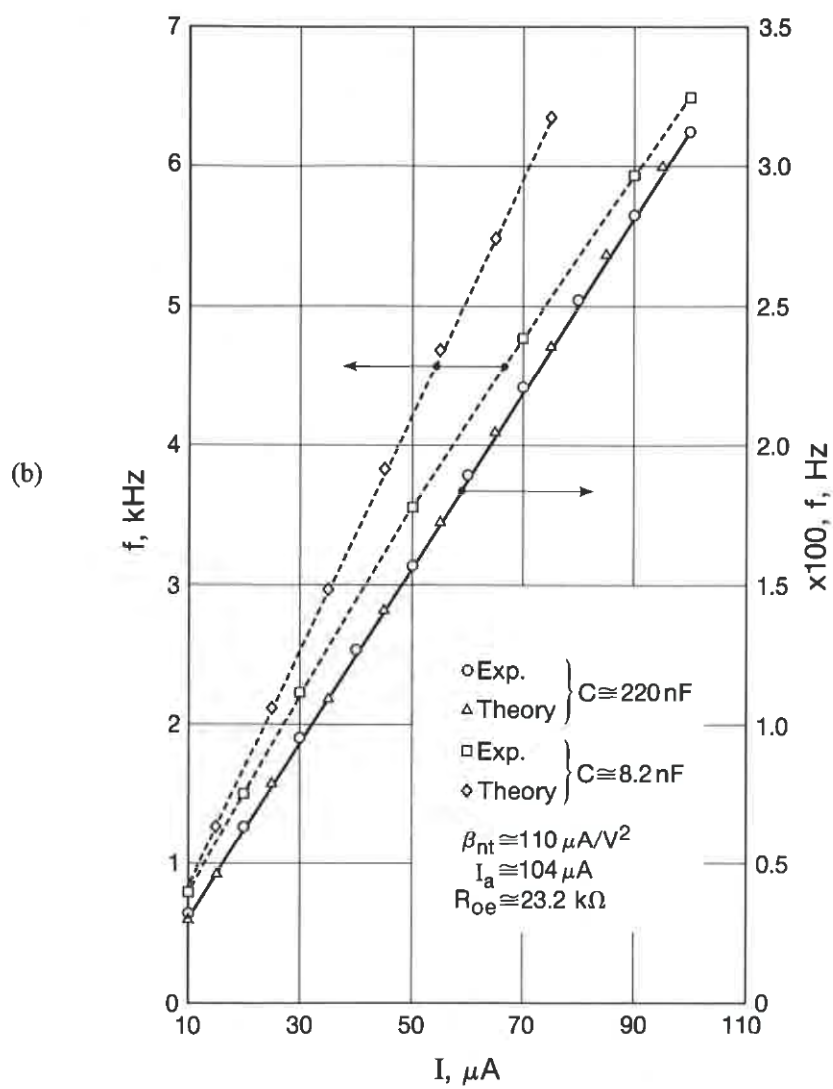


Figure 13.23 continued.

so that the voltage at the timing capacitor has the peak-to-peak value of

$$V_c \approx I_a R_{oe} - 2(1 - q) \sqrt{\frac{I_a}{\beta_{nt}}} \quad (13.139)$$

This voltage, contrary to the source-coupled multivibrator case, does not depend on the control current I or parameters of the devices through which this current flows. The timing capacitor is charged and discharged by the current I during both semiperiods. The oscillation frequency in this case (no terms in (13.139) depend on I) is

$$f_0 = \frac{I}{2CV_c} \approx \frac{I}{2C \left[I_a R_{oe} - 2(1 - q) \sqrt{\frac{I_a}{\beta_{nt}}} \right]} \quad (13.140)$$

Equation (13.140) shows that this multivibrator has linear control characteristic with I as a control current. The nonlinearity of the control characteristic is a result of switching delays. If the switching delays are constant, they can be taken into consideration by a simple correction of the result (13.140); namely, in this case

$$f = \left[\frac{2CV_c}{I} + T_s \right]^{-1} \quad (13.141)$$

where T_s is the total switching delay. Then

$$\frac{\delta f}{f_0} = -f_0 T_s \quad (13.142)$$

This result was confirmed [53] experimentally (breadboarding, MO415 CMOS transistor array devices), where the frequency was measured for the range of I from 10 to 100 μA . Two values of timing capacitor C were used: 220 and 8.2 nF. The test with the large capacitor (right scale) confirms that the theoretical and experimental results correspond and show a linear relationship between frequency and current, as expected. For CFC with the small capacitor (left scale, a higher overall frequency range) the theoretical and experimental results diverge and the experimental characteristic becomes nonlinear. The total switching delay of the circuit was measured to be about 30 μs . Subtracting this delay from the frequency characteristic returns it to a straight line.

The lower limit for the current is set by the level that could be accurately measured using the available test equipment (in fact, the circuit can operate at lower currents as well, which was confirmed for I_c down to $1 \mu\text{A}$), and the upper limit is set by the condition

$$I < \frac{\beta_{nc}}{\beta_{nt}} \quad (13.143)$$

This condition, specific for CMOS integrated circuits, provides that the charge current is entirely switched by the threshold circuit. In (13.143) $\beta_{nc} = (\mu C_{ox} W_3)/(2L_3) = (\mu C_{ox} W_4)/(2L_4)$. If $\beta_{nc} = \beta_{nt}$, the condition (13.143) limits I to I_a .

Using the extended control range configuration [53,54], where a buffer stage is added to the basic configuration of Figure 13.23, allows CFC to be obtained with the control characteristic over a current range of 3 to 1,000 mA with the integral nonlinearity of 1.5%.

13.3.7 Duty-Cycle Modulation

A multivibrator with duty-cycle modulation was mentioned in the previous sections. The ratio T_1/T_2 is determined by the ratio of two elements (as it was in (13.49)), which is an attractive feature allowing wide element tolerance.

As was mentioned previously the frequency in the current-controlled multivibrators can be calculated with sufficient precision if the exact swing of the voltage at the timing capacitor is known. In the emitter-coupled multivibrator this voltage changes with control current, introducing a nonlinearity in frequency control characteristic. A multivibrator with duty-cycle modulation represents an attempt to use a controllable oscillation parameter that is independent of the V_c values.

The circuit of an emitter-coupled multivibrator with nonidentical current sources in the emitters of Q_1 and Q_2 (Figure 13.24(a)) oscillates with a duty cycle different from 50%. Indeed, using the model shown in Figure 13.18 with $g_{m5} = g_{m6} = 0$ and following the same calculation steps, one can find the voltage V_c at the timing capacitor at the instant of switching. The oscillation period consists of two parts, $T_1 = (2V_c)/I_1$ and $T_2 = (2V_c)/I_2$; and their ratio $T_1/T_2 = I_2/I_1$, indeed, does not depend on V_c .

The exact control characteristic, hence, depends on the circuit that controls the ratio I_2/I_1 . If, for example [55], these currents are the output currents of a differential pair driven from a bridge (Figure 13.24(b)), then

$$\frac{T_1}{T_2} = \exp\left(\frac{v_2 - v_1}{V_T}\right) \approx 1 + \frac{2V_{CC}\delta R}{V_T(R_0 + 2R)} \quad (13.144)$$

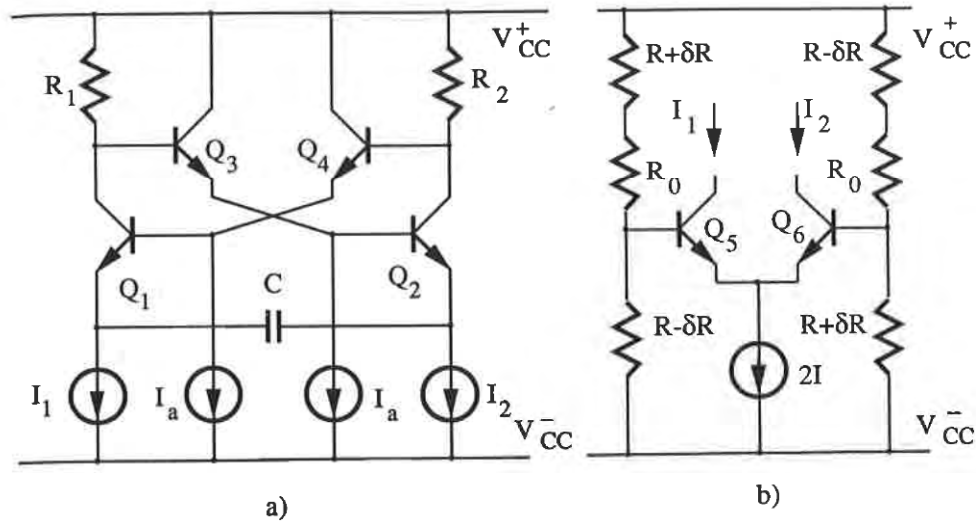


Figure 13.24 Emitter-coupled multivibrator with a controlled duty cycle: (a) multivibrator; (b) control of I_1 and I_2 .

is linearly dependent on the bridge detuning when the differential signal $v_2 - v_1$ is small (which is usually the case). One can estimate the control characteristic non-linearity using the additional term of the exponent expansion.

The oscillation frequency depends on the ratio I_2/I_1 as well, but

$$f = \frac{I}{4V_c} \left[1 - \left(\frac{v_2 - v_1}{2V_T} \right)^2 \right] \quad (13.145)$$

$$\frac{\delta f}{f_0} = - \left(\frac{v_2 - v_1}{2V_T} \right)^2 \quad (13.146)$$

is by the order of magnitude less than the change of T_2/T_1 (the voltage is constant if the sum $I_1 + I_2 = 2I$ and constant). Here $f_0 = I/(4V_c)$.

Cascoding of an emitter-coupled multivibrator with a voltage-controlled differential pair reduces the circuit dynamic range. The reduction can be avoided by using the parallel connection of a Schmitt-trigger-type multivibrator and a voltage-controlled current source (Figure 13.25).

The multivibrator of the circuit is shown in Figure 13.25(a). Ignoring for a while the current source δI , assume that the power supply is turned on. The capacitor C is initially discharged and the transistor Q_1 will be on and Q_3 will turn off. The

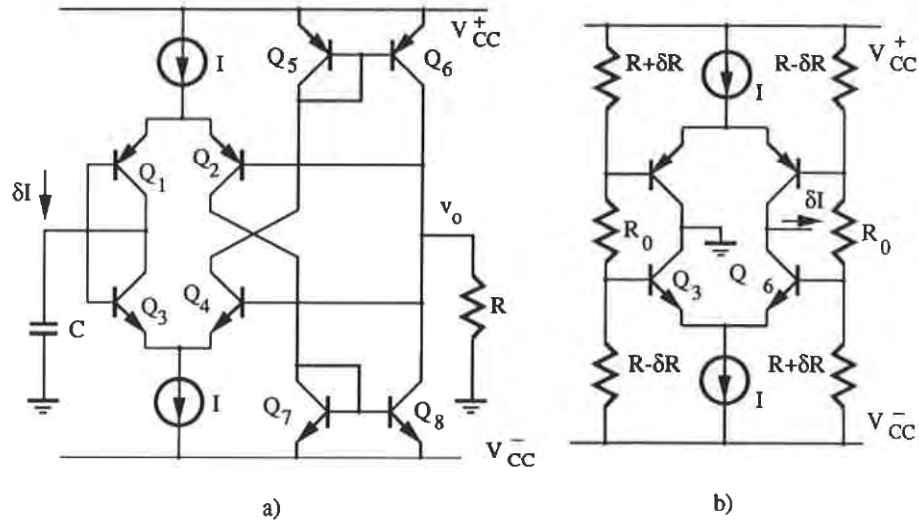


Figure 13.25 Schmitt-trigger multivibrator with a controlled duty cycle: (a) multivibrator; (b) bridge control of δI .

current I from the bottom current source will be taken by Q_4 and repeated by the current mirror Q_5, Q_6 . The voltage v_o at the resistor R will jump to its positive value of $V_{oh} = IR$. The current from the top current source will charge, via Q_1 , the capacitor C . When the voltage v_c attains the value of $V_{ch} \approx IR - 120 \text{ mV}$, the transistor Q_1 starts to turn off, and Q_3 starts to turn on. Positive feedback will accelerate this switching, and the resistor voltage will jump to its negative value of $V_{ol} = -IR$. Now the multivibrator in its steady-state oscillation, the voltage v_c changes from V_{ch} to $V_{cl} \approx -IR + 120 \text{ mV}$ and vice versa.

The detailed calculation that follows the same steps as in the multivibrator with a noncomplementary trigger just considered shows that

$$V_c = V_{ch} - V_{cl} = 2IR - 2V_T(1 + \ln m) \quad (13.147)$$

where $m = \{1 + \sqrt{[1 - (4V_T)/(IR)]}\} / \{1 - \sqrt{[1 - (4V_T)/(IR)]}\}$. The oscillation frequency, hence, is $f_0 = I/(2CV_c)$.

If the parallel current source δI is in operation, the oscillation period becomes $T = T_1 + T_2$, where $T_1 = (CV_c)/(I + \delta I)$ and $T_2 = (CV_c)/(I - \delta I)$. The ratio

$$\frac{T_1}{T_2} = \frac{I + \delta I}{I - \delta I} \approx 1 - \frac{2\delta I}{I} \quad (13.148)$$

and also does not depend on V_c . The oscillation frequency becomes $f = f_0[1 - (\delta I/I)^2]$. Thus, for small values of $\delta I/I$, the ratio T_1/T_2 changes as a linear function of $\delta I/I$, and the frequency variation $\delta f/f_0 = -(\delta I/I)^2$ is one order of magnitude less.

This parallel source δI can be arranged, as in the previous circuit, as the bridge-controlled current source (Figure 13.25(b)). If the base current is neglected, one can obtain $\delta I = (8V_{CC}g_m\delta R)/(R_0 + 2R)$. Here $g_m = I_a/(2V_T)$ is the transconductance of the transistors Q_9 to Q_{12} . Then

$$\frac{T_1}{T_2} = 1 - \left[\frac{8V_{CC}\delta R}{V_T(R_0 + 2R)} \right] \frac{I_a}{I} \quad (13.149)$$

Comparing (13.148) and (13.149) one can see that in this circuit the sensitivity to the bridge detuning can be additionally controlled by the ratio I_a/I . The circuit of Figure 13.25(b) can be, with some modifications, used as a duty-cycle detector [56].

13.3.8 Controlled Multivibrators in Applications

It is feasible to design a multivibrator with a linear dependence of the oscillation frequency on the control voltage. The three decade (four decades in bipolar case) variation of frequency can be achieved with one capacitor value. The offset compensation circuits are simple. The multivibrators do not require amplitude control-stabilization circuits; hence, they are less complicated than the sinusoidal-oscillator-based circuits designed for the frequency control. The circuit integration is less difficult. The factors limiting the linearity of control characteristics are or the same (finite slew rate of the operational amplifiers) or become important at higher frequencies (switching delays).

13.4 SUMMARY

The considered circuits show that there are many ways of conversion from analog signal to frequency, pulse rate, and duty cycle. The conversion of an analog signal in the frequency of a sinusoidal oscillator is used mainly when the amplitude control system can be essentially simplified (a rectifier with a filter of a long time constant) or completely omitted. As a result the sinusoidal oscillators are combined with sensors of slow changing variables (temperature, humidity). The development of CMOS multiplier circuits (which are now in the stage of intensive investigation) with small offsets is vital for this method of conversion. Analog-to-frequency conversion using multivibrators is very flexible. The frequency can be controlled by one element (passive or active), which makes these circuits suitable for applications with many types

of sensors. A resistive bridge can be also a part of multivibrator structure, which makes the multipliers even more flexible in sensor applications. These properties are preserved for multivibrators with duty cycle control. Also, in multivibrators the frequency or pulse output can be easily encoded into digital form.

REFERENCES

- [1] S. Middelhoek, P. J. French, J. H. Huising, and W. J. Lian, "Sensors with Digital or Frequency Output," *Sensors and Actuators*, Vol. A15, 1988, pp. 119–133.
- [2] A. Cichocki and R. Unbehauen, "Application of Switched-Capacitor Self-Oscillating Circuits to the Conversion of RLC Parameters into a Frequency or Digital Signal," *Sensors and Actuators*, Vol. A24, 1990, pp. 129–137.
- [3] A. G. J. Holt and M. R. Lee, "A Class of RC-Oscillators," *Proc. IEEE*, Vol. 55, 1967, p. 1119.
- [4] W. G. Howard and D. O. Pederson, "Integrated Voltage-Controlled Oscillators," *Proc. Nat. Electron. Conf.*, Vol. 23, 1967, pp. 279–284.
- [5] A. B. Grebene, "A High-Frequency Voltage-Controlled Oscillator for Integrated Circuits," *Proc. Nat. Electron. Conf.*, Vol. 24, 1968, pp. 216–220.
- [6] Y. Sun, "Generation of Sinusoidal Voltage (Current) Controlled Oscillators for Integrated Circuits," *IEEE Trans. Circuit Theory*, Vol. CT-19, 1972, pp. 137–141.
- [7] J. G. Graeme, *Applications of Operational Amplifiers*, New York, McGraw-Hill, 1973.
- [8] M. Herpy, *Analog Integrated Circuits*, New York, John Wiley & Sons, 1980.
- [9] V. P. Singh and S. K. Saha, "Linear Sinusoidal VCO," *Int. J. of Electronics*, Vol. 65, 1988, pp. 243–247.
- [10] I. M. Filanovsky, S.-S. Qiu, and G. Kothapalli, "Sinusoidal Oscillator with Voltage Controlled Frequency and Amplitude," *Int. J. of Electronics*, Vol. 68, 1990, pp. 95–112.
- [11] *Intersil Data Book*, 1981.
- [12] *Linear Databook*, National Semiconductor Corporation, 1982.
- [13] D. Meyer-Ebrecht, "Schnelle amplitudenregelung harmonischer Oszillatoren," *Philips Research Rep. Supplement*, Vol. 6, 1974, pp. 1–85.
- [14] I. M. Filanovsky, G. J. Fortier, and L. F. Taylor, "Amplitude Transients in the Twin-T Bridge RC-Oscillator with a Multiplier Amplitude Control System," *Int. J. Electronics*, Vol. 64, 1988, pp. 547–561.
- [15] E. Vannerson and K. C. Smith, "A Low Distortion Oscillator with Fast Amplitude Stabilization," *Int. J. Electronics*, Vol. 39, 1975, pp. 465–472.
- [16] E. Vannerson and K. C. Smith, "Fast Amplitude Stabilization of an RC-Oscillator," *IEEE J. Solid-State Circuits*, Vol. SC-9, 1974, pp. 176–179.
- [17] J. Pahor, J. Fettich, and M. Tavzes, "A Harmonic Oscillator with Low Harmonic Distortion and Stable Amplitude," *Int. J. Electronics*, Vol. 37, 1974, pp. 765–768.
- [18] N. Vannai and L. Pap, "RC Oscillators with Extremely Low Harmonic Distortion," *Periodica Polytechn., Elect. Eng.* (Hungary), Vol. 24, No. 1–2, 1980, pp. 59–65.
- [19] I. M. Filanovsky, "A Wien-Bridge RC-Oscillators with Fast Amplitude Control," *Int. J. Electronics*, Vol. 58, 1985, pp. 817–826.
- [20] I. M. Filanovsky, "Oscillators with Amplitude Control by Restoration of Capacitor Initial Conditions," *IEEE Proc.*, Vol. 134, No. 1, 1987, pp. 31–37.
- [21] D. Meyer-Ebrecht, "Fast Amplitude Control of a Harmonic Oscillator," *Proc. IEEE*, Vol. 60, No. 6, 1972, p. 736.
- [22] P. R. Gray and R. G. Meyer, *Analysis and Design of Analog Integrated Circuits*, 2d ed., New York, J. Wiley & Sons, 1984.

- [23] B. Van der Pol, "The Nonlinear Theory of Electric Oscillations," *Proc. IRE*, Vol. 22, 1934, pp. 1051–1086.
- [24] H. S. Tan and K. W. H. Foulds, "Analytical Determination of the Waveforms of Nonlinear Oscillators," *Electronic Circuits and Systems*, Vol. 1, 1977, pp. 99–102.
- [25] A. Grebene, *Bipolar and MOS Analog Integrated Circuit Design*, New York, J. Wiley & Sons, 1984.
- [26] A. Kindlund, H. Sundgren, and I. Lundström, "Quartz Crystal Gas Monitor with a Gas Concentrating Stage," *Sensors and Actuators*, Vol. A6, 1984, pp. 1–17.
- [27] R. M. Landon, "Resonator Sensors," *J. Phys. E: Sci. Instrum.*, Vol. 18, 1985, pp. 103–115.
- [28] J. D. Maines, E. G. S. Paige, A. F. Saunders and A. S. Young, "Simple Technique for the Accurate Determination of Delay-Time Variations in Acoustic-Surface-Wave Structures," *Electron. Letters*, Vol. 5, No. 26, 1969, pp. 678–680.
- [29] J. Crabb, M. F. Lewis, and J. D. Maines, "Surface-Acoustic-Wave Oscillators: Mode Selection and Frequency Modulation," *Electron. Letters*, Vol. 9, No. 10, 1973, pp. 195–197.
- [30] A. Venema, E. Nieuwkoop, M. J. Vellekoop, M. S. Nieuwenhuizen, and A. W. Barendsz, "Design Aspects of SAW Gas Sensors," *Sensors and Actuators*, Vol. A10, 1986, pp. 47–64.
- [31] R. Best, *Phase-Locked Loops*, New York, McGraw-Hill, 1984.
- [32] D. H. Sheingold, ed., *Transducer Interfacing Handbook*, Norwood, MA, Analog Devices, 1980.
- [33] J. H. Huising, G. A. Van Rossum and M. Van der Lee, "Two-Wire Bridge-to-Frequency Converter," *IEEE J. Solid-State Circuits*, Vol. SC-22, 1987, pp. 343–349.
- [34] G. B. Clayton, *Operational Amplifiers*, 2d ed., London, Newnes-Butterworths, 1979.
- [35] I. M. Filanovsky, V. A. Piskarev and K. A. Stromsmoe, "Nonsymmetric Multivibrators with an Auxiliary RC-Circuit," *IEE Proc.*, Vol. 131, No. 4, 1984, pp. 141–146.
- [36] I. M. Filanovsky, "A Simple Voltage-to-Frequency Converter," *IEEE Circuits and Devices*, Vol. 6, No. 5, 1990, p. 47.
- [37] I. M. Filanovsky and V. A. Piskarev, "A Simple Bridge RC Multivibrator," *Int. J. Electronics*, Vol. 57, No. 2, 1984, pp. 217–226.
- [38] T. M. Frederiksen, *Intuitive Operational Amplifiers*, New York, McGraw-Hill, 1988.
- [39] T. J. Van Kessel and R. J. Van de Plassche, "Integrated Linear Basic Circuits," *Philips Tech. Review*, Vol. 32, No. 1, 1971, pp. 1–12.
- [40] I. M. Filanovsky, "Remarks on Design of Emitter-Coupled Multivibrators," *IEEE Trans. Circ. Syst.*, Vol. CAS-35, 1988, pp. 751–755.
- [41] R. R. Cordell and W. G. Garrett, "A Highly Stable VCO for Application in Monolithic Phase-Locked Loops," *IEEE J. Solid-State Circ.*, Vol. SC-10, No. 10, 1975, pp. 480–485.
- [42] B. Gilbert, "A Versatile Monolithic Voltage-to-Frequency Converter," *IEEE J. Solid-State Circuits*, Vol. SC-11, No. 6, 1976, pp. 852–864.
- [43] F. V. J. Slecckx and W. M. C. Sansen, "A Wide-Band Current-Controlled Oscillator Using Bipolar-FET Technology," *IEEE J. Solid-State Circuits*, Vol. SC-15, No. 5, 1980, pp. 875–881.
- [44] I. G. Finvers and I. M. Filanovsky, "Analysis of a Source-Coupled CMOS Multivibrator," *IEEE Trans. Circ. Syst.*, Vol. CAS-35, 1988, pp. 1182–1185.
- [45] R. Gregorian and G. C. Temes, *Analog MOS Integrated Circuits for Signal Processing*, New York, J. Wiley & Sons, 1986.
- [46] P. E. Allen and D. R. Holberg, *CMOS Analog Circuit Design*, New York, Holt, Rinehart and Winston, 1987.
- [47] *Analog MOS Integrated Circuit Design Manual*, Scotts Valley, CA, Interdesign Corp., 1983.
- [48] A. Nathan, I. A. McKay, I. M. Filanovsky, and H. P. Baltes, "Design of a CMOS Oscillator with Magnetic Field Frequency Modulation," *IEEE J. Solid-State Circuits*, Vol. SC-21, 1987, pp. 230–232.
- [49] C. C. M. Meijer, "A Three-Terminal Wide-Range Temperature Transducer with Microcomputer Interface," *European Solid-State Circuits Conf.*, Delft, The Netherlands, 1986.

-
- [50] J. F. Kukielka and R. G. Meyer, "A High-Frequency Temperature Stable Monolithic VCO," *IEEE J. Solid-State Circuits*, Vol. SC-16, 1981, pp. 639–647.
 - [51] I. M. Filanovsky, "A Current Controlled Multivibrator for Low Voltage Power Supply," *Int. J. Electronics*, Vol. 65, No. 1, 1988, pp. 37–43.
 - [52] I. M. Filanovsky and I. G. Finvers, "A Simple Nonsaturated CMOS Multivibrator," *IEEE J. Solid-State Circuits*, Vol. SC-23, No. 1, 1988, pp. 290–292.
 - [53] I. G. Finvers, "CMOS Current Controlled Oscillators," MSc thesis, University of Alberta, Edmonton, Canada, 1988.
 - [54] I. M. Filanovsky, I. G. Finvers, Lj. Ristic, and H. P. Baltes, "Multivibrators with Frequency Control for Application in Integrated Sensors," *Proc. First Int. Forum on ASIC and Transducer Technology (ASICT'88)*, Honolulu, 1988, p. 65.
 - [55] R. R. Spencer and J. B. Angell, "A Voltage-Controlled Duty-Cycle Oscillator," *IEEE J. Solid-State Circuits*, Vol. SC-25, No. 1, 1990, pp. 274–281.
 - [56] I. M. Filanovsky and V. A. Piskarev, "Duty Cycle Detector," *IEEE Circuits and Devices*, Vol. 7, No. 1, 1991.

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