Historical Perspective of Accelerometer Technologies

Torben R. LICHT

Patrick R. SCHEEPER

A/S Brüel & Kjaer

DK-2850 Naerum

Denmark

1. Introduction

Basic physical discoveries have in most cases led to useful applications some decades later.

Apart from optical methods making a direct measurement of the kinematic quantity of acceleration possible, the most fundamental physical principle is Newton's second law (1687) relating acceleration to force via a constant of proportionality: The inertial mass.

For measurement of vibration the single degree of freedom system consisting of a mass suspended by a spring plays a crucial role, together with additional physical mechanisms converting (i.e. transducing) the relative motion or force into a directly measurable (mostly electrical) signal.

The most important discoveries for vibration measurements are capacitance (Leyden jar 1745), induction (1819 H.C. Oersted), strain-gauge (1856 Lord Kelvin), piezoelectricity (1880 P. & J. Curie), semiconductor components (transistor 1948 W.B.Shockley), piezoresistivity of semiconductors (1954 C.S.Smith) and electron tunnelling (1983 Binnig).

In parallel to the basic physical principles the technologies have advanced with high speed.

In the period 1920 to 1940 vibration measurements started using inductive, capacitive, strain-gauge and piezoelectric principles. An early accelerometer using a Rochelle salt plate in bending mode was designed by P.V. Brüel in 1943, and manufactured by Brüel & Kjaer afterwards.

The lead titanate zirconate family of piezoelectric materials were found in the forties, and accelerometers using these materials were marketed in the fifties. Construction and material refinements have made these transducers the most widely used general purpose transducers over the last three decades and still today.

The semiconductor strain-gauges or piezoresistive elements came into use as replacements for the classic metal s-

train-gauges in the last half of the fifties and one of the first micromechanical constructions was a pressure sensor using a silicon diaphragm with diffused resistors.

Since then thousands of papers have been published on what today is called micromechanical sensors, mostly based on silicon as material.

One of the first micromechanical accelerometers was also based on piezoresistivity and a bending beam. This was published by Roylance and Angell in 1979⁽¹⁾ and many similar constructions have been investigated in the meantime.

Through the eighties detection and construction principles were proposed and investigated, but the number of commercially available sensors is still very small. Most of these are piezoresistive or capacitive.

The main reason for the small number of commercially available sensors is the very fragmented nature of the market. A large number of small companies are making a huge variety of transducers specialised for each their application. To make micromachined sensors require large investments which can only be paid back if the number of transducers sold is sufficiently large or the prices can be put at a very high level.

Therefore the trend has been transducers for the automotive market, e.g. airbag sensors where the potentially low cost batch fabrication can be exploited, and transducers for very specialised applications, e.g. microgravity measurements where price is unimportant.

To understand the new challenges for the designer of accelerometers it is important to know what happens to various parameters when dimensions are scaled.

2. Scaling

Most classic accelerometers of today have volumes in the order of 1 cm³, seismic masses of 1 to 30 grams and mounted resonance frequencies of 10 to 30 kHz.

To make feasible constructions in silicon wafers these dimensions L have to be reduced by a factor of about 10. The resonance should preferably remain constant.

For the mass-spring system this implies the following scaling properties ($\propto$ denotes proportionality, F the force on the seismic mass due to the acceleration a of the system, k spring stiffness, M seismic mass and A and t cross section and thickness of a simple disc-shaped spring):

$$\begin{array}{lll} \text{Dimensions} & \propto & L \\ \text{Seismic mass} & \propto & L^3 \end{array}$$

The resonance frequency is given by

$$\omega_0 = \sqrt{\frac{k}{M}} \quad \text{----- (1)}$$

and the deformation of the spring is given by the formula (Hooke's and Newton's II law)

$$kx = F = Ma \quad \text{----- (2)}$$

To maintain the resonance frequency ω_0 constant, k must be proportional to M and therefore

$$k \propto \frac{A}{t} \propto M \propto L^3 \quad \text{----- (3)}$$

(For a bender construction with length, width and thickness l, w, t the expression A/t should be substituted by wl^3/t^3)

For a constant vibration input signal this system will then see an input force

$$F \propto M \quad \text{----- (4)}$$

I.e. the sensitivity for force sensitive transduction principles at $\omega \ll \omega_0$ follows

$$S_f \propto M \propto L^3 \quad \text{----- (5)}$$

The displacement is

$$x = \frac{M}{k} a = \frac{a}{\omega_0^2} \quad \text{----- (6)}$$

which means that the displacement sensitive device keep its sensitivity constant when M is changed, provided ω_0 is kept constant, or

$$S_x = \text{Const.} \quad \text{----- (7)}$$

This also implies that the spring dimensions must change considerably to fulfil

$$\frac{A}{t} \propto L^3 \quad \text{----- (8)}$$

Finally a scaling property which is important, especially for active transducers like piezoelectric accelerometers, is the amount of energy available for transduction when a certain vibration is present. For a certain instantaneous velocity v of the accelerometer the kinetic energy stored in the motion of the seismic mass M is

$$E_{\text{Kinetic}} = \frac{1}{2} M v^2 \quad \text{----- (9)}$$

These scaling properties makes it necessary to look at effects which were not of any concern hitherto.

<2.1> Thermal noise

Mechanical motion The thermal motion (displacement) of the seismic mass can be described as follows (see e.g.⁽²⁾):

$$\sigma_{dT}^2 = 4KT\Delta f \frac{1}{M\omega_0^2 Q} \frac{1}{\left[1 - \left(\frac{\omega}{\omega_0}\right)^2\right]^2 + \left(\frac{\omega}{Q\omega_0}\right)^2} \quad \text{----- (10)}$$

$$\omega \ll \omega_0 \Rightarrow$$

$$\sigma_{dT}^2 = 4KT\Delta f \frac{1}{M\omega_0^2 Q} \quad \text{----- (11)}$$

or if using

$$F = kx = \omega_0^2 M \quad \text{----- (12)}$$

it can be expressed as force:

$$\sigma_{FT}^2 = 4KT\Delta f \frac{\omega_0^2 M}{Q} \quad \text{----- (13)}$$

Where K is Boltzmanns constant, Δf the frequency band of observation centered around ω , T the absolute temperature and Q the quality factor of the resonance.

This gives the following proportionalities:

$$\sigma_{dT} \propto \frac{1}{\sqrt{M}} \propto \frac{1}{\sqrt{L^3}} \quad \text{----- (14)}$$

and

$$\sigma_{FT} \propto \sqrt{M} \propto \sqrt{L^3} \quad \text{----- (15)}$$

and in both cases the signal to noise ratio becomes:

$$\frac{S}{N} \propto \sqrt{L^3} \quad \text{----- (16)}$$

which means that by scaling the transducer size as proposed it becomes increasingly difficult to get a proper signal to thermal noise ratio. Till today the electronic noise has been by far the dominant factor, but going to the micromechanical magnitudes, we will reach the thermal limit provided the electronic noise does not increase, which is normally not the case.

Resistance noise

The well known Johnson noise or thermal noise of a resistor

$$u_n = \sqrt{4KTR\Delta f} \quad \text{----- (17)}$$

should naturally not be forgotten.

<2.2> Quantum effects Having dealt with the ther-

mal noise, we also have to evaluate contributions from quantum effects.

An ideal capacitor in the form of a piezoelectric material or a variable capacitor is limited by the smallest available charge, the electron charge:

$$e = 1.6 \cdot 10^{-19} \text{ C} \quad \text{----- (18)}$$

This might be a limiting factor for piezoelectric transducers which are basically capacitors (assuming no losses).

Linked to a current is the so called shot noise which is valid for cases where the electrons can be considered free

$$I_n = \sqrt{2eI\Delta f} \quad \text{----- (19)}$$

For the tunnelling effect mentioned later this is the most important electrical noise source.

For resistors which are not pure metallic, different more complicated mechanisms produce current noise. An expression including most phenomena will have the following form (adapted from ⁽³⁾)

$$v^2 = \Delta f V_0^2 \left(A \frac{1}{f} + B \frac{1}{1 + (2\pi\tau)^2 f^2} \right) \quad \text{----- (20)}$$

where v is the noise voltage over the resistance, V_0 the DC voltage across the resistor, τ a relaxation time constant and A and B constants depending on dimensions and materials. An interesting observation for scaling of transducers with the resistance kept constant is that

$$A \propto A_c^{-2} \propto L^{-4} \quad \text{and} \quad B \propto l^{-2} \propto L \quad \text{----- (21)}$$

A_c being the cross-section of the resistor and l the length.

This shows that especially the $1/f$ noise can become a serious problem when downscaling is made.

In some cases a term following $1/\sqrt{f}$ will also be present, but this has no clear dimensional dependence.

The electronics used for feedback/readout may or may not be able to reach the limits given by the intrinsic noise stated here.

3. Practical implications

To put these "new" parameters into perspective in conjunction with possible applications of today, we will evaluate the effects if we want to make a dedicated modal analysis accelerometer for light structures.

A set of desirable characteristics could be:

Dynamic range (1Hz bandwidth)	100dB (1 mm/s ² to 100 m/s ²)
Frequency range ($\pm 10\%$)	0.1 Hz to 2 kHz
Nonlinearity	Max. 1 % at 100 ms ²
Max. cross-axis sensitivity	< 5%
Weight	< 1 gram
Maximum shock	> 20000 m/s ²

<3.1> Resonance frequency and Q factor

The res-

onance frequency should be around 5 kHz, which is reasonably above the frequency range of interest. To keep the phase of the measured signal in the whole frequency range of interest lower than approx. 1°, the Q-factor should be >20, which keeps the phase shift "concentrated" around the resonance frequency but does however prevent long duration ringing when the resonance is excited.

<3.2> Noise level specifications

A few words may

be said about the specified noise level.

The requirement for the modal analysis accelerometer considered is that the noise level for all frequencies between 1 Hz and 1 kHz is 10⁻³ m/s²/√Hz or lower. An accelerometer with this spectral noise density and a white-noise spectrum is considered useful. However, one problem is that preamplifiers show a frequency dependent noise which has to be dealt with.

It should be noted that this criterion is only relevant for electrical noise, and not for the mechanical-thermal noise level nor for the shot noise, which both are white in the working range as shown earlier (Eq. 10.).

4. Examples

Six different transduction principles, that are considered as the most relevant for silicon-based accelerometers, will be analysed.

First, the accelerometers based on strain sensing are discussed. These are the piezoresistive, the piezoelectric, the resonant and the optical strain gauge accelerometer.

Next, the displacement-sensing accelerometers are discussed: capacitive accelerometers (with and without forced balancing) and accelerometers based on electron tunnelling.

Finally a comparison of the different transduction principles is made.

<4.1> **Strain-sensing accelerometers** Strain-sensing accelerometers have proven to be successful over many years. Brüel & Kjaer's piezoelectric accelerometers are based on strain sensing (or stress sensing). An advantage of strain sensing is that the accelerometer structures can actually be very stiff, thereby obtaining a robust structure with a high resonance frequency. These properties are typical for Brüel & Kjaer's accelerometers, where it is demonstrated that the stiff structure still allows measurement of low-level vibrations in a high dynamic range. The first silicon micromachined accelerometer, the famous device of Roylance and Angell ⁽¹⁾, which was developed at Stanford University in 1979, was based on another strain-detection principle, piezoresistive detection.

Two relatively new strain detection principles are the resonant principle, where the resonance frequency of a small bridge structure is used as a measure for the strain, and the optical strain gauge based on Bragg grating, giving an optical frequency shift when strained ⁽⁴⁾.

The detection principles will be discussed in the next sections.

Piezoresistive detection

The piezoresistive effect is widely used for silicon sensors. The effect is known to be linear up to stress levels which are close to the yield strength of silicon, whereas in the lower end the resolution of the strain gauge is determined by the Johnson noise in the piezoresistors⁽¹⁾. The linearity of the piezoresistive effect can give a high dynamic range of the accelerometer.

Dynamic range of the Wheatstone bridge

The maximum measurable acceleration is usually determined by other nonlinearities than caused by the piezoresistive effect itself. The piezoresistors are used in a Wheatstone bridge configuration. The output voltage ΔU of a Wheatstone bridge with one active resistor is

$$\Delta U = \frac{\Delta R}{(4R + 2\Delta R)} U_0 \quad (22)$$

where U_0 is the DC supply voltage, R is the resistor value and ΔR is the resistance change. It follows that the Wheatstone bridge is only linear as long as $\Delta R \ll 2R$. This criterion can be written as a criterion for the maximum strain ϵ that can be measured using the definition of the gauge factor G :

$$\frac{\Delta R}{R} = G\epsilon \quad (23)$$

If we take a nonlinearity of 1% on ΔU as the upper limit for detection, then (from 22.) $\Delta R/R \ll 0.02$. The gauge factor is dependent on the doping level of the silicon, and can be up to 200 for monocrystalline silicon. For polycrystalline silicon a typical gauge factor is around 20⁽⁵⁾. This gives a maximum measurable strain of 10^{-3} for polycrystalline silicon strain gauges, and 10^{-4} for monocrystalline silicon.

The resolution of the strain gauge is limited by the thermal noise of the piezoresistors, which typically have values between 1 and 10 k Ω . A typical value for the thermal noise of a piezoresistor is then (eq. 17.) approximately 10 nV/ $\sqrt{\text{Hz}}$. If the DC supply voltage of the Wheatstone bridge is 5V, and assuming one active piezoresistor, then the minimum relative resistance change that can be detected is $8 \cdot 10^{-9}$. Using the same gauge factors for polycrystalline and monocrystalline silicon as mentioned above, the minimum strains that can be resolved are $4 \cdot 10^{-10}$ and $4 \cdot 10^{-11}$, respectively.

From this it follows that the dynamic range of a Wheatstone bridge with one active piezoresistor is limited to 128 dB, because of the thermal resistor noise and the requirement of a maximum nonlinearity of 1%. Note that the Wheatstone bridge gets twice as sensitive by using two active resistors, and four times as sensitive by using four active resistors (it is assumed that four resistors of the same magnitude is used in all cases!). This results in noise levels which is 6 and

12 dB lower, respectively. For the bridge with two active resistors the nonlinearity criterion is the same as for the bridge with one active resistor. The bridge with four active resistors does not show the nonlinearity of the Wheatstone bridge, as mentioned above. In practice, nonlinearities will for example be introduced by the nonideal placement (alignment errors) of the piezoresistors on the bending elements, causing the resistance changes to be slightly different for each strain sensing element.

The suspension beams: Nonlinearity and cross-axis sensitivity

Another type of nonlinearity appears when the piezoresistive strain gauges are used in an accelerometer construction. Usually the strain gauges are used in a beam bender construction, since the highest sensitivity is obtained in this way. (examples: Christel et al.⁽⁶⁾, Roylance and Angell⁽¹⁾, Crazzolaro et al.⁽⁷⁾, Terry et al.⁽⁸⁾). However, such a beam bender construction can put an upper limit to the detection, by introducing a nonlinearity caused by stretching of the beam material at large beam deflections. This is the case when the seismic mass is suspended in a stiff frame by beams at opposite sides. An example of such a construction is shown in Fig. 1 (from Allen et al.⁽¹²⁾).

Fig. 1. Example of piezoresistive accelerometer with four beam suspension. From Allen et al.⁽¹²⁾.

A beam bender construction based on a single cantilever beam is not sensitive to this type nonlinearity. This nonlinearity has been demonstrated by Christel et al. from Lucas Novasensor⁽⁶⁾, where cantilever- and fourbeam accelerometers, both with a 2g range, were compared and showed errors of 1% and 25% respectively at 8g acceleration. A disadvantage of the cantilever construction as used by Christel et al. is that it cannot distinguish between in-plane and out-of-plane accelerations, because the center of gravity of the seismic mass is not positioned in the plane of the cantilever beam. Such cantilever beam constructions can be fabricated, by for example adding an electroplated gold mass on the silicon seismic mass (Heuberger⁽⁹⁾), but this usually complicates the fabrication process. Another way to get around the problem of the nonlinearity of beam bender constructions is to use the piezoresistors in a push-pull configuration, as for example demonstrated by Suminto⁽¹⁰⁾. A disadvantage of using strain gauges in a push-pull configuration is the lower sensitivity which limits the applications to cases where the acceleration

levels are relatively high.

The measurement amplifier

Due to the low output impedance of a piezoresistive accelerometer, the component can be connected directly to a measurement amplifier. The typical noise level of a measurement amplifier, is around $10 \text{ nV}/\sqrt{\text{Hz}}$. However, to obtain this low noise level, the gain of the first amplifier stage should be 40 dB and DC coupled. This implies that the DC component in the accelerometer output signal should be less than 50 mV, in order not to drive the DC output of the first amplifier stage to the rail. Consequently, the sensitivity of the accelerometer should be less than 50 mV/g , to avoid that the earth's gravitational field causes this output signal.

Other causes for a DC component in the accelerometer's output signal are unbalance of the Wheatstone bridge, and nonlinearities in the sensitivity of the device (Christel et al.)⁽⁶⁾. However, the response to these effects can be kept under 50 mV. Of course the DC output can be decoupled with a series capacitor between the accelerometer and the amplifier, but then the noise level is suddenly determined by the 1 MOhm typical input resistor of the measurement amplifier, causing the noise level to increase with 20-30 dB.

Performance of accelerometers from the literature

Since the expected noise level of the piezoresistors and of the preamplifier have to be added, giving $14 \text{ nV}/\sqrt{\text{Hz}}$ ideally, a total noise level of $20 \text{ nV}/\sqrt{\text{Hz}}$ is taken as a fixed value for analysing the performance of accelerometers from the literature. The resolution of the accelerometer is then this noise level, divided by its sensitivity.

To evaluate the performance of an accelerometer in an objective way, not only noise level and dynamic range should be compared, but the resonance frequency should be taken into account as well. In table 1 data from several published accelerometer designs are listed.

Assuming that one can resolve $20 \text{ nV}/\sqrt{\text{Hz}}$, the dynamic ranges of the listed accelerometers are high. For modal analysis purposes, the accelerometers of Crazzolaro and Tschan do have a suitable resonance frequency and sufficient

resolution. The accelerometer of Roylance and Angell does have sufficient resolution, but the construction cannot be made stiffer to increase the resonance frequency. With the accelerometer from Barth et al. the sensitivity can be lowered a factor of 100, thus obtaining a resonance frequency of 5 kHz and a resolution of 0.2 mm/s^2 . The response to the earth's gravitational field is less than 50 mV for accelerometers with a proper resonance frequency.

Temperature sensitivity

A typical problem with piezoresistive sensors is the temperature sensitivity. Both the offset and the sensitivity do change with temperature. Typically the temperature coefficients of offset (TCO) and sensitivity (TCS) are 0.2-0.3 %/°C, when no measures are taken to compensate for this. With temperature compensation used, these values can be brought down to 2% over a temperature range from -40 to $+85^\circ\text{C}$ ⁽¹⁴⁾. However, this requires accurate trimming of each individual component adding to the cost.

Summary

Summarising, piezoresistive accelerometers can have a high dynamic range, in combination with a sufficiently high resonance frequency for modal analysis applications. The resolution that can be obtained in practice is expected to be sufficient, when using a low noise measurement amplifier, that can handle the DC response of the accelerometer. A drawback of piezoresistive detection is the temperature sensitivity, although effective temperature compensation schemes have been developed.

Piezoelectric detection

The only micromechanical piezoelectric accelerometer reported in the literature is the device from Chen et al.⁽¹⁵⁾ developed at Berkeley/Rockwell in 1982, and shown in Fig. 2. It was an open-loop, cantilever-type accelerometer, based on sputtered ZnO, and with an on-chip MOS preamplifier. In 1994, a theoretical study was done at B&K to investigate the possibility of a triaxial ZnO-based accelerometer⁽¹⁶⁾.

The next section is divided into three parts. First, the noise

Authors	Working range (m/s ²)	Sensitivity (5V supply) ($\mu\text{V/ms}^{-2}$)	Resolution (m/s ² /√Hz) *	Dynamic range (dB)	Resonance (kHz)	Type
Roylance and Angell ⁽¹⁾	1960	25	$8 \cdot 10^{-4}$	128	2.3	C
Barth et al. ⁽¹¹⁾	19.6	10000	$2 \cdot 10^{-6}$	140	0.5	C
Crazzolaro ⁽⁷⁾	3330	20	$1 \cdot 10^{-3}$	130	7	D
Allen et al. ⁽¹²⁾	196	255	$8 \cdot 10^{-5}$	128	2	Q
Tschan ⁽¹³⁾	490	102	$2 \cdot 10^{-4}$	128	5	Q

Table 1. Performance of accelerometers reported in the literature

* = a noise level of $20 \text{ nV}/\sqrt{\text{Hz}}$ has been assumed

Type: C = cantilever, D = two beam bridge type, Q = four beam bridge type

equivalent stress of a piezoresistive and a piezoelectric strain gauge are compared. Next, previous calculations on ZnO accelerometers are critically reviewed, and finally PZT benders are investigated further.

Fig 2. Piezoelectric ZnO based accelerometer, after ⁽¹⁵⁾.

Piezoresistive vs. piezoelectric strain gauges

A comparison, valid for some applications, between the piezoresistive and the piezoelectric transduction principle can be made by comparing the signal-to-noise ratio of two strain gauges. A rectangular strain gauge is assumed, which is loaded by a uniform, uniaxial stress σ_x .

Piezoelectric

For the piezoelectric strain gauge it can easily be shown that the generated charge Q in the z -direction, per stress-unit (i.e. Pa), is

$$\frac{Q}{\sigma_x} = \frac{d_{31}tC_z}{\epsilon_{33}} \quad (24)$$

where t is the thickness of the piezoelectric material, C_z is the capacitance of the piezoelectric layer, and ϵ_{33} is the permittivity of the piezoelectric film in the z -direction ($\epsilon_{33} = \epsilon_0 \cdot \epsilon_n$).

In principle, if the leakage current through a piezoelectric transducer is low, all the noise is generated by the charge amplifier. The amplifier noise is usually modelled as a white-noise voltage source e_n at the amplifier input node. The equivalent charge noise q_n at the amplifier input can then be calculated. The result is

$$q_n = e_n(C_f + C_z) \quad (25)$$

where C_f is the feedback capacitance in the charge amplifier, which determines the gain. At a high-gain setting, the feedback capacitance is usually much smaller than C_z . Note that parasitic capacitances have been neglected.

Assuming no parasitic capacitances, and a high-gain setting of the charge amplifier, the noise equivalent stress in Pa becomes

$$N = \frac{\sigma_x q_n}{Q} = \frac{\epsilon_{33} e_n}{d_{31} t} \quad (26)$$

Note that the important parameter for the piezoelectric material is the ratio d_{31}/ϵ_{33} . This ratio is approximately $-0.060\text{-m}^2/\text{C}$ for ZnO, and $-0.015\text{-m}^2/\text{C}$ for PZT2 (data from Berlincourt ⁽⁹⁾).

If a B&K 2626 charge amplifier is used, the noise source e_n is approximately $1.2 \times 10^{-8} \text{V}/\sqrt{\text{Hz}}$. Assuming a $1\mu\text{m}$ thick piezoelectric layer, the noise becomes $0.2 \text{Pa}/\sqrt{\text{Hz}}$ for ZnO, and $0.8 \text{Pa}/\sqrt{\text{Hz}}$ for PZT2.

Piezoresistive

The output signal of the piezoresistive strain gauge can easily be found with the equations given in the previous section. Assuming a Wheatstone bridge with 4 active element (highest sensitivity), the output signal u per unit stress becomes

$$\frac{u}{\sigma_x} = \frac{GU_0}{E} \quad (27)$$

where G is the gauge factor, U_0 is the supply voltage and E is the Young's modulus of the strain gauge material ($1.7 \times 10^{11} \text{N/m}^2$ for silicon in the $\langle 110 \rangle$ -direction).

The noise level is given by the well-known Johnson noise equation stated in the previous section. The noise equivalent stress in $\text{Pa}/\sqrt{\text{Hz}}$, at the Wheatstone bridge output becomes

$$N = \frac{u_n \sigma_x}{u} = \frac{E\sqrt{4KTR}}{GU_0} \quad (28)$$

where K is Boltzmann's constant ($1.38 \times 10^{-23} \text{J/K}$), T is the absolute temperature, and R is the bridge resistance.

At a supply voltage of 5V, a temperature of 293K, and assuming a gauge factor of 200 and a resistance of $1\text{k}\Omega$, the noise equivalent stress becomes $0.67 \text{Pa}/\sqrt{\text{Hz}}$.

Conclusion

Comparing the noise equivalent stress of the piezoelectric and the piezoresistive strain gauges, it can be concluded that a PZT-based piezoelectric strain gauge is as good as a piezoresistive strain gauge (other types of PZT with a higher d_{31}/ϵ_{33} ratio do exist, e.g. $-0.021\text{m}^2/\text{C}$ for PZT-5H ⁽¹⁷⁾).

It appeared that zinc oxide-based strain gauges should be able to give lower noise levels.

In a theoretical study, calculations of sensitivity and non-linearity were done on a ZnO-based triaxial accelerometer ⁽¹⁶⁾. Noise level simulations were done on an ASIC charge amplifier, that should be built together with the accelerometer in a package. In the study, it was concluded that the noise level requirements could be fulfilled.

However, this conclusion was based on a noise value at 100Hz. Calculated noise spectra show that the noise below 1000Hz shows a $1/f$ dependence, implicating that at 1Hz the dynamic range is 40dB lower. Thus, the accelerometer can possibly fulfil broadband-noise requirements, but for modal analysis it is more desirable that the noise-level is low for all frequencies.

Reducing the noise by 40dB is difficult. Increasing the seis-

mic mass and lowering the capacitance did not prove it possible to gain more than maybe 20 dB at the most.

It can be concluded that measuring vibrations below 100 Hz is not possible down to the required detection level with a ZnO-based accelerometer as described in ⁽¹⁶⁾. The reported designs can be optimised, but not enough to achieve a 40 dB improvement of the signal-to-noise-ratio. A comparison between PZT- and ZnO-strain gauges shows that a PZT-based accelerometer will show a considerably higher sensitivity, but not a better resolution.

Further comments

These considerations have all been based on the assumption that the piezoelectric material is used as a strain gauge in the same way as a piezoresistive element. This is however somewhat misleading, because the piezoresistive element is supplied with a current to measure the deformation, and is then best used when placed at maximum strain points on a given structure. The structure can then be designed for resonance frequency etc.

When using piezoelectric transduction it must however be remembered that it is an active transduction of vibration energy into electrical energy, and therefore as much as possible of the energy must be "guided" to the piezoelectric material. In this case it means that the beam shall be made from the piezoelectric material to as large an extent as possible.

To evaluate the possibilities we can use the formulas given for macroscopic bimorphs and see what the scaling could bring.

C.P. Germano ⁽¹⁸⁾ gives the formulas

$$S_v = \frac{3}{2} \frac{g_{31} l}{wt} \quad \text{----- (29)}$$

$$k = \frac{Y_{11}^D wt^3}{4l^3} \quad \text{----- (30)}$$

and

$$C_e = \frac{\epsilon_{33} h v}{t} \quad \text{----- (31)}$$

for the voltage sensitivity, spring stiffness and capacitance of a series connected rectangular bimorph. l , w and t are the length, width and thickness of the bender in the one end clamped condition. $g_{31} = d_{31}/\epsilon_{33}$. Y_{11}^D is the elastic constant (N/m^2). If we use the dimensions $l, w, t = 1\text{ mm}, 0.2\text{ mm}, 0.2\text{ mm}$ and $10 \cdot 10^{-3} \text{ Vm/N}$ for g_{31} , $6.9 \cdot 10^{10} \text{ N/m}^2$ for Y_{11}^D and $2000 \cdot 8.85 \text{ pF/m}$ for ϵ_{33} we get $S_v = 375 \text{ V/N}$ which with a seismic mass of 3 mg gives a vibration sensitivity of 11 mV/g. The capacitance becomes $C_e = 18 \text{ pF}$ and the resonance frequency 15.3 kHz.

Using a maximum strain of $20 \cdot 10^6 \text{ N/m}^2$ the maximum vibration level before fracture can be calculated to be approximately 900g using the formula for strain from earlier.

This leads to a noise level of approximately 1 μg , and therefore nearly 180 dB useful dynamic range for the transducer

alone (nonlinearity will reduce this number 10-20 dB if a 2% limit is used). If the swing is limited to 5 V and 18 pF is used as feedback, then the useful dynamic range at the output of a charge amplifier is 168 dB, and the noise level 0.02 mm/s². This is sufficiently low to fulfil the modal accelerometer requirements stated in the beginning.

To fulfil the 1 Hz, 10% lower limiting frequency requirement a 30 G Ω resistor is needed in parallel to the 18 pF feedback capacitor. This could be difficult to realise. A suitable trade-off between 1/f-noise, lower limiting frequency and feedback resistor must be made.

The noise level of 1 μg approaches also the quantum noise level stated earlier, because the stated sensitivity corresponds to a charge sensitivity of 0.2 pC/g and this gives a charge of $2 \cdot 10^{-19} \text{ C}$ at 1 μg which is close to the electron charge of $1.6 \cdot 10^{-19} \text{ C}$.

Conclusion

The conclusion to be drawn from this section is that PZT benders in miniature format should be a viable choice for a modal accelerometer.

This part of micromachining is however not studied in sufficient detail to guarantee that the technology can be scaled without problems. Most of the work done on PZT materials is aimed at thin films which are much thinner than the dimensions indicated here (typically < 1 μm).

Resonant strain detection

The third way to detect strain is the resonant strain gauge.

The strain is indicated by a change in the resonance frequency of a vibrating beam or similar structure. The resonant frequency is typically in the range 50 to 500 kHz.

The modulation can be up to 25% without introducing serious distortion although the effect is quadratic.

Quality factors (Qs) in the range from 200 in air to 100,000 in vacuum has been used. A high Q is naturally desirable if high resolution is needed. This imposes special requirements to the packaging which might be difficult to obtain for extended timespans.

The resonant gauge has the potential of increasing the resolution because of the basic FM modulation property giving a signal to noise improvement proportional to the ratio of modulation to detection bandwidth i.e.

$$\frac{S}{N} \dots \alpha \dots \frac{\Delta f_{\text{Modulation}}}{\Delta f_{\text{Detection}}} \quad \text{----- (32)}$$

If a 2kHz bandwidth is needed and 20kHz (i.e. $\pm 10\text{kHz}$) modulation can be obtained a 20 dB higher S/N ratio can be obtained.

Most reports are giving DC measurements, maybe corresponding to a bandwidth of 0.1Hz and therefore a very high resolution and large dynamic range.

The only vibration test reported to our knowledge were made at 200 Hz and gave a 60 dB dynamic range and 50 mg resolution ⁽¹⁹⁾. The principal part of the construction is shown

in Fig. 3.

The FM demodulators used today will hardly give more than 80 to 100 dB dynamic range for the desired 2 kHz bandwidth.

To our knowledge, the only commercially available resonant accelerometer is made by the Norwegian company Sen

Fig. 3. Hinged mass with resonant beam construction, after Roszhart et al. ⁽¹⁹⁾.

soNor ^(20,21). Their SA30 accelerometer is a component aiming at the airbag sensor market. With this component, the resonant detection principle is primarily used to provide a continuous self test during operation of the device, thereby considerably improving the reliability of the accelerometer.

Results of development of high-performance resonant accelerometers were presented by Roszhart et al. at Kearfott/IC Sensors ⁽¹⁹⁾ and Honeywell ⁽²²⁾. These accelerometers are developed for navigation purposes, where a high resolution is required, whereas the required frequency range is low.

Optical detection

Several different schemes have been devised to use light to detect displacements, most of which falls short of fulfilling our wish to resolve 10^{-12} m or 10^{-6} of a wavelength related to detection of acceleration.

However, a promising method based on a Bragg grating embedded in an optical waveguide acting as a strain gauge, has recently been described ⁽⁴⁾.

The principle of operation is based upon the Bragg condition for the dominant wavelength component λ_b of the light reflected by the grating. It is given by the formula

$$\lambda_b = 2\Lambda n_{\text{eff}} \quad (33)$$

where Λ is the grating period and n_{eff} is the effective index of refraction.

The strain sensitivity is given by

$$\frac{\Delta\lambda_b}{\lambda_b} = (1 - p_e)\epsilon \quad (34)$$

Where p_e is the photoelastic strain coefficient (0.22 for pure silica). The expected sensitivity for an accelerometer with a 13 kHz resonance frequency is $1.7 \cdot 10^{-14}$ nm/ms⁻², and the resolution which can be obtained is $4.4 \cdot 10^{-3}$ ms⁻². Reducing the resonance frequency to half the value will give the desired

resolution of 10^{-3} ms⁻².

The principle including the detection method is quite new (1992) and many technological problems have to be solved before a product can reach the market.

< 4.2 > Capacitive detection with force-balancing

With the capacitive detection principle, the displacement of the seismic mass is measured, instead of the strain in a suspension beam. Advantages of capacitive detection are the high sensitivity and the low temperature sensitivity. Since the capacitance between the seismic mass and a second fixed plate is inversely proportional to the distance, a capacitive accelerometer is non-linear. Capacitive accelerometers are often used in combination with a force-balancing system for linearization. Another advantage of the feedback system is the extended frequency range. Without the use of a feedback system, the strong damping due to the narrow airgap causes the bandwidth to be much less than the resonance frequency of the seismic mass. An example of a capacitive accelerometer construction is shown in figure 4.

Fig. 4. Capacitive accelerometer, after Rudolf et al. ⁽²⁴⁾.

Working range

One upper limit of the dynamic range is set by the feedback system. The maximum acceleration is given by the force acting on the seismic mass (mass $\times$ acceleration) that still can be compensated by electrostatic forces, which depend on the maximum voltage ΔV that the feedback amplifier can generate, and on geometrical factors (the electrode area A and the plate separation d). The following expression for the maximum acceleration is given by Kühnel and Sherman ⁽²³⁾:

$$a = \frac{2\epsilon_0 A}{Md^2} V_p \Delta V \quad (35)$$

where V_p is the DC polarisation voltage between the capacitor plates, and ϵ_0 is the vacuum permittivity.

In Table 2 data are given for the ADXL50 accelerometer from Analog Devices ⁽²³⁾ and the A-SMAC accelerometer from

CSEM.⁽²⁴⁾ These accelerometers have measured working ranges of 80 g and 0.1 g, respectively. The calculated limits set by the feedback system for these accelerometers are 250 g and 0.2 g. The large difference between the calculated and the measured working range of the ADXL50 is due to the amplifier output voltage, which is limited to the output voltage generated at 80g acceleration.

Transducer type.	A-SMAC	ADXL50
Seismic mass (mg)	14.7	$0.16 \cdot 10^{-3}$
Air gap (μm)	7	1.3
Capacitance (pF)	20	0.1
Polarisation voltage (V)	2.5	1.6
Voltage swing (V)	2.0	1.6
Measured data:		
Sensitivity (V/g)	5	0.020
Working range (g)	0.1	80
Noise level (g/ $\sqrt{\text{Hz}}$)	$0.1\text{--}0.5 \cdot 10^{-6}$	$4.9 \cdot 10^{-3}$
Resonance freq. (kHz)	0.126	20
Q factor	not specified	5
Calculated data:		
Sensitivity (V/g)	4.5	0.0001
Working range (g)	0.2	250
Thermal noise (g/ $\sqrt{\text{Hz}}$)	$0.095 \cdot 10^{-6}$ (Q=1 assumed)	$0.16 \cdot 10^{-3}$

Table 2. Data for two capacitive accelerometers.

Noise level

The resolution of the accelerometer is given by the preamplifier noise divided by the sensitivity. To calculate the resolution of the accelerometer, the preamplifier noise has to be calculated. A full discussion of preamplifier noise is beyond the scope of this paper. However, a general rule for preamplifiers for this type of capacitive sensors is that the preamplifier noise decreases if the input capacitance increases, where the input capacitance is the sum of the accelerometer capacitance and the parasitic capacitances. It should be noted that when one keeps the input capacitance constant, and changes the ratio between the accelerometer capacitance and the parasitic capacitance, the noise will remain constant, but the signal-to-noise ratio gets worse if the parasitic capacitance becomes larger. The CSEM preamplifier shows a noise level of $3 \mu\text{V}/\sqrt{\text{Hz}}$ for an input capacitance of 20 pF⁽²⁵⁾. We believe that this noise level can be considered as representative for the current state-of-the-art, and we will use this value for further analysis.

Kühnel and Sherman⁽²³⁾ derive that the sensitivity S of a capacitive accelerometer, that is used in a differential readout configuration, is:

$$S = \frac{V_p}{\omega_0^2 d} \quad \text{--- (36)}$$

where ω_0 is the vacuum resonance frequency. As can be seen in Table 2, this equation predicts a sensitivity of $100 \mu\text{V/g}$ for the ADXL50 accelerometer, and a sensitivity of 4.5 V/g for the A-SMAC. The later value agrees well with the measured value of 5 V/g , whereas the measured sensitivity of the ADXL50 of 20 mV/g is considerably higher than the predicted value. The difference is caused by signal attenuation by stray capacitances (6 dB), and by an additional amplifier stage adding 52 dB gain.

Analysis of the performance

To evaluate the performance, both dynamic range and frequency range should be taken into account. From the expression that Kühnel and Sherman give for the sensitivity, it can directly be seen that a trade-off has to be made concerning sensitivity and resonance frequency. The product of the sensitivity and the resonance frequency squared is the electrical field strength in the air gap. The amplifier output voltage swing and the air-gap thickness are thus key parameters for optimising the sensitivity and bandwidth of the accelerometer. The design trade-off between the sensitivity and bandwidth can be seen from the data listed in Table 2. The ADXL50 has a high resonance frequency of 20 kHz, and consequently a low sensitivity. In contrast, the A-SMAC shows a high sensitivity of 5 V/g , which is needed for the desired μg resolution, at the cost of a low resonance frequency of 126 Hz.

We will investigate with a design example if an accelerometer for modal analysis can be based on capacitive detection in combination with a feedback system. If we assume that the output voltage swing is limited to $\pm 2 \text{ V}$, and the desired working range 100 m/s^2 , then the maximum sensitivity of the accelerometer becomes 20 mV/m/s^2 . If the preamplifier noise level is $3 \mu\text{V}/\sqrt{\text{Hz}}$, this corresponds to a resolution of $1.5 \cdot 10^{-4} \text{ m/s}^2/\sqrt{\text{Hz}}$. If the desired resonance frequency is 5 kHz, and the polarisation voltage V_p is 2 V, equation (36.) gives an air-gap thickness of $0.10 \mu\text{m}$. This is a value which is technologically very hard to realise, and has to our knowledge not been demonstrated in any of the published silicon accelerometers. The noise level can be increased with a factor of 7, to $1 \cdot 10^{-3} \text{ m/s}^2/\sqrt{\text{Hz}}$. This allows the air-gap thickness to be increased to $0.7 \mu\text{m}$, which is still a factor of 3 smaller than the smallest practically used values in accelerometer constructions fabricated by wafer bonding techniques. As can be seen from equation (36.), the limiting factor for improvement of the sensitivity - bandwidth product is the electrical field strength. So either the airgap can be made smaller, or the electronics should be adjusted so that the used polarisation voltage can be increased. Apart from possible complications with increasing the voltages in the electronics with a factor 3, other problems of an increased electrical field strength in the airgap have to be considered, such as the risk for electrical discharge, and electrostatic collapse,

where the seismic mass gets unstable and is attracted to one of the electrodes.

Recent developments with capacitive force-balanced accelerometers have resulted in completely digital devices (e.g. ^{(26), (27)}), where a sigma-delta converter detects if the position of the seismic mass differs from its rest position, and gives a feedback signal as a series of voltage pulses, which at the same time is the bitstream output of the accelerometer. This integrates the AD conversion with the accelerometer electronics. Other digital techniques have been presented, as for instance Pulse Width Modulation feedback (PWM) by Suzuki et al. ⁽²⁸⁾. We will not further analyse the performance of the digital accelerometers, but we believe the physical limitations to the performance are the same as in the analog accelerometers (buffer amplifier noise, accelerometer sensitivity and the electrostatic forces in the feedback system).

Summarising, capacitive force-balanced accelerometers have been demonstrated with low detection levels and a dynamic range of 100 dB. However, the sensitivity-bandwidth product of these accelerometers is limited, which makes them less suitable for modal analysis applications where both high sensitivity and bandwidth are required.

<4.3> Electron tunnelling detection In 1989, Waltman and Kaiser ⁽²⁹⁾ from Jet Propulsion Laboratories presented the first electron tunnelling accelerometer. This accelerometer was machined in aluminium, and the tunnelling current between two gold electrodes was controlled with a piezoelectric cantilever. To sustain a tunnelling current, the two gold electrodes have to be brought at a distance of typically 10 Å. The used tunnelling current is typically around 1 nA, and depends exponentially on the electrode separation:

$$I_t = I_0 e^{-\frac{x}{x_0}} \quad (37)$$

where I_0 is a function of the applied voltage between the conductors and

$$x_0 = \frac{1}{\alpha \sqrt{\phi}} \quad (38)$$

where $\alpha = 1.025 \cdot 10^{10} \text{ m}^{-1} \text{ eV}^{-1/2}$ and ϕ is the effective work function of the surfaces. For gold to gold tunnelling, ϕ is about 0.5 eV and x_0 becomes approximately 1 Å. The tunnelling current is a strongly nonlinear function of the distance x . For each Ångström displacement, the current changes by nearly a factor of three. The sensitivity of a tunnelling accelerometer is the derivative of equation (37):

$$-\frac{\partial I}{\partial x} = \frac{I_t}{x_0} \quad (39)$$

With $I_t = 1.3 \text{ nA}$ and $x_0 = 1 \text{ Å}$, the sensitivity becomes 9.4 A/m.

Noise level, dynamic range and Bandwidth

Due to the high sensitivity of the tunnelling accelerometer,

very low accelerations can be detected even when an amplifier with a normal noise performance ($10^{-14} - 10^{-15} \text{ A}/\sqrt{\text{Hz}}$) is used. In practice, the amplifier noise may turn out to be less relevant than other noise sources, as for instance the shot noise in the tunnelling junction, or the mechanical-thermal noise level of the accelerometer. The shot noise current, is given by equation (19.). For a tunnelling current and gap given above the shot noise becomes $2 \cdot 10^{-14} \text{ A}/\sqrt{\text{Hz}}$. This corresponds to a displacement noise of $2 \cdot 10^{-15} \text{ Å}/\sqrt{\text{Hz}}$.

The upper detection limit is set by the feedback system. If a similar electrode configuration is used as with the capacitive feedback accelerometer, the same expression as in the previous section can be used for calculating the working range.

The influence of the three noise sources is shown in a design example, of which the data are shown in Table 3.

Seismic mass	$1 \cdot 10^{-7} \text{ kg}$
Spring constant	100 N/m
Resonance frequency	5 kHz
Air gap thickness	5 μm
Electrode area	3.5 mm ²
Max. feedback voltage	2 V
Acceleration sensitivity	$9.4 \cdot 10^{-9} \text{ A/m/s}^2$
Working range	100 m/s ²
Noise level:	
Thermal (Q=20 assumed)	$1.6 \cdot 10^{-5} \text{ m/s}^2/\sqrt{\text{Hz}}$
HzShot noise	$2.3 \cdot 10^{-6} \text{ m/s}^2/\sqrt{\text{Hz}}$
Preamplifier	$1.1 \cdot 10^{-6} \text{ m/s}^2/\sqrt{\text{Hz}}$

Table 3. Tunnelling accelerometer design example.

The example above shows that, due to the high sensitivity, the mechanical-thermal noise determines the noise level, rather than shot noise or preamplifier noise. Note that the shot noise was based on a current of 1.5 nA, and the assumed preamplifier noise was $1 \cdot 10^{-14} \text{ A}/\sqrt{\text{Hz}}$, which is a conservative value. Although this is just a simple example, it shows that the preamplifier noise is lower than the contributions of other noise sources. Therefore, researchers at JPL and Stanford University working with tunnelling sensors use simple opamp-based electronics, and do not focus on noise optimisation of the amplifiers ⁽³⁰⁾.

Until now, the work from the researchers at JPL and Stanford University on accelerometers was primarily focused on devices with a limited frequency range (100 Hz), but with a detection level lower than $10^{-8} \text{ g}/\sqrt{\text{Hz}}$ ⁽³¹⁾. A drawing of such an accelerometer is shown in figure 5. Using a 100 mg seismic mass and resonance frequency of 100 Hz, a noise level of $1.5 \cdot 10^{-7} \text{ g}/\sqrt{\text{Hz}}$, at a frequency of 10 Hz, were obtained. The accelerometers sensitivity was 15 kV/g ⁽³¹⁾.

Fig. 5. Tunnelling accelerometer, after Rockstadt et al. ⁽³¹⁾.

Research on electron tunnelling accelerometers started only a few years ago, and electron tunnelling based sensors are not yet on the market. However the results presented up to now look promising. An infrared tunnelling sensor has been reported to operate continuously for more than 18 months (by the time the publication was written), Kenny ⁽³⁰⁾. Many subjects are still under investigation, as for example the influence of tunnel electrode cleanliness (important for long-term stability, ⁽³⁰⁾), noise performance ⁽³²⁾, and the feedback control system ⁽³³⁾. From the design example given above, the expected dynamic range, noise level and bandwidth are sufficient for modal applications.

5. Discussion

Piezoresistive and piezoelectric accelerometers exhibit similar dynamic ranges (both approximately 130 dB), in combination with a sufficient frequency range for modal applications. The piezoresistive accelerometers are produced today, whereas piezoelectric accelerometers only have been presented in a few publications. Since the piezoresistors only have to cover a small area to sense strain, piezoresistive accelerometers can easily be scaled down to dimensions suitable for silicon micromachining, whereas scaling piezoelectric accelerometers is limited, since piezoelectric strain gauges have to cover a certain area, needed for generating sufficient charge for the charge amplifier to measure.

The Bragg grating strain detection will probably find special application areas due to the inherent immunity to EMI and other disturbances, but much work remains to be done, also on the demodulation technique.

Scaling resonant strain gauges is inherent to the detection principle, since resonator frequencies of the order of 100 kHz are desired for operation. We believe the applications of resonant accelerometers are limited to applications at low frequencies, such as inertial navigation. With a limited bandwidth, the resonant accelerometer shows a high dynamic range and resolution.

The capacitive feedback technology has reached the same state as the piezoresistive technology, and a number of these accelerometers is commercially available today. With capacitive feedback accelerometers, the sensitivity-bandwidth prod-

uct is limited by the electrical field strength in the airgap. The commercially available accelerometers therefore show either a high resolution and dynamic range in a limited bandwidth, as for example the CSEM accelerometer for inertial navigation, or a low resolution in an extended bandwidth, as for instance the ADXL50 for airbag applications. We believe that the limited sensitivity-bandwidth product makes these accelerometers less suitable for modal and more general purpose type applications.

Electron tunnelling accelerometers are very well suitable to downscaling, since the active area for sensing is a very small tunnelling tip, thus not having scaling restrictions to the area as with piezoelectric or capacitive accelerometers. Tunnelling accelerometers have been demonstrated with noise levels less than 0.1 μg in a 100 Hz bandwidth. We expect that these accelerometers can show a low noise level in combination with a high bandwidth. However, issues as long-term stability and shock protection of these accelerometers still have to be clarified. The next decade will show if these sensors find their way to the market.

In the 17 years after the first micromachined silicon accelerometer from Roylance and Angell, silicon accelerometers based on many different transduction techniques have been developed, and some found their way to the market. Silicon micromachining allowed downscaling of dimensions, thus making massfabrication of reliable and low cost accelerometers for airbag applications possible. The high precision and new possibilities of the micromachining technology resulted in new types of accelerometers, as for instance the resonant accelerometer and the force-balance capacitive accelerometer. These accelerometers can further increase reliability, for example by the continuous self-test of a resonant accelerometer.

To take advantage of the potentially low price of the accelerometer chip new more cost effective methods of encapsulation and connection have to be developed if the accelerometers shall be suitable for the general purpose market of today.

New technologies are being investigated, such as the electron tunnelling accelerometer, and the optical accelerometers. The new transduction principles will lead to new limits for miniaturisation and noise level. Some of the new transduction principles will make new accelerometers find their way to the market in the next decade, but a revolution as in the computer world does not seem imminent.

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