Smart Sensors for the IoT

5. Signal Conditioners

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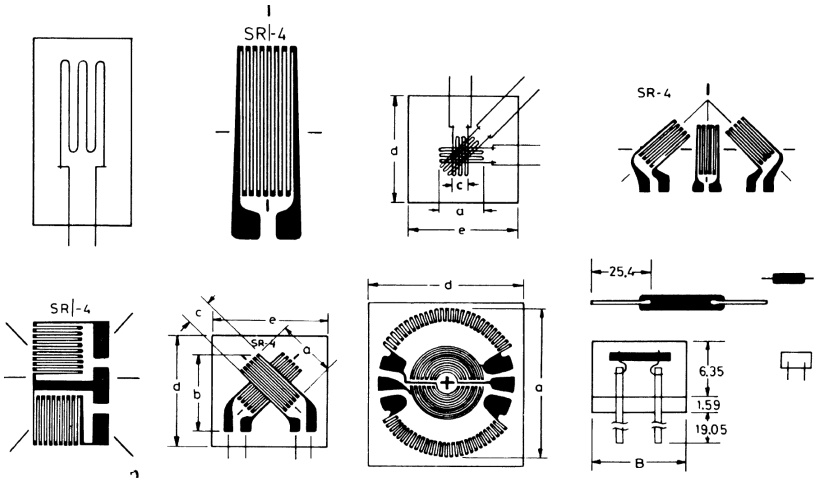
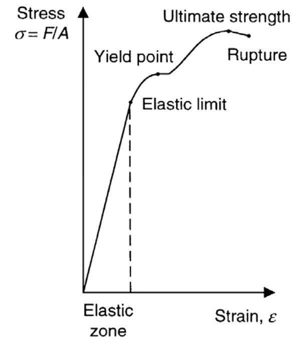
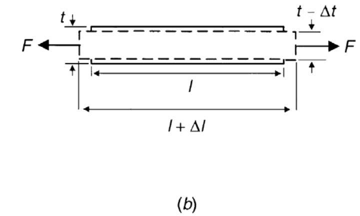
Outline

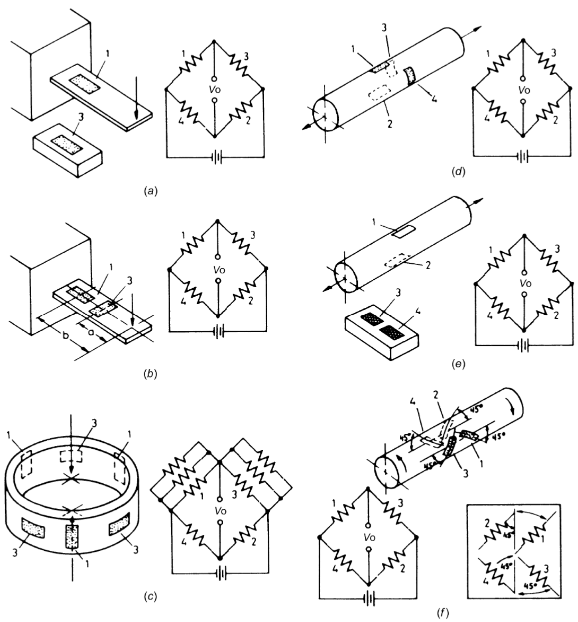
- **Signal conditioners for resistive sensors**
- Signal conditioners for capacitive sensors (capacitive sensor interfaces)

Examples of Resistive Sensors – Strain Gages





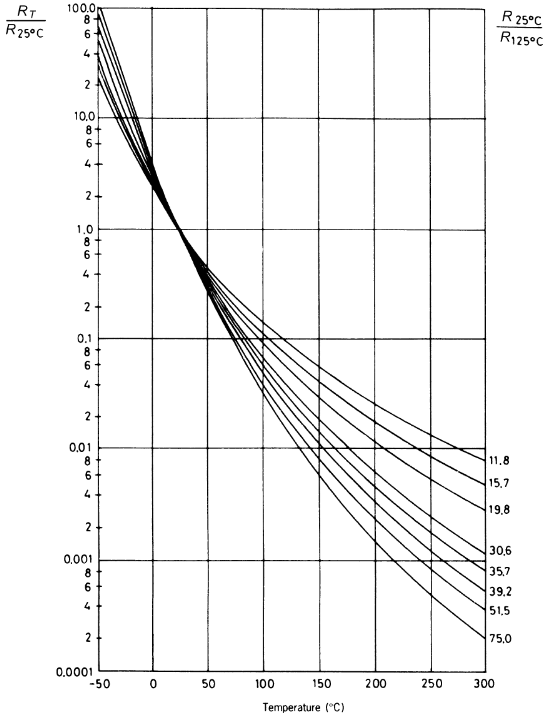




- For small variations, the resistance of the metallic wire is
$$R \cong R_0 \cdot (1 + G \cdot \varepsilon)$$
- where $\varepsilon = d\ell/\ell$ is the **strain** (dimensionless) and G the gage factor given by
$$G = 1 + 2v + C \cdot (1 - 2v)$$
- G ranges between 2 and 6 depending on the material and v is the Poisson ratio according to the Poisson law
$$v = -\frac{dt/t}{d\ell/\ell}$$

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Examples of Resistive Sensors – Thermistors



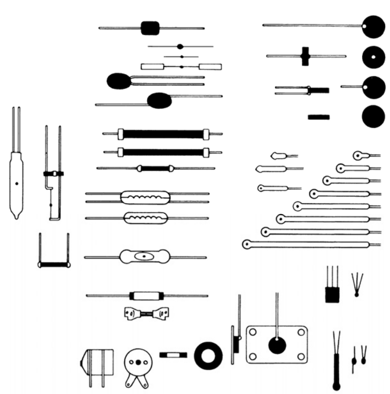


Figure 2.19 Different shapes for NTC thermistors (from Fenwal Electronics).

- Nonlinear characteristic given by
  $$R_T = R_0 \cdot e^{B \cdot (1/T - 1/T_0)}$$
- where R_0 is the resistance at 25°C (or other reference temperature) and T_0 is this temperature in kelvins
- B is called the **characteristic temperature** of the material value, which is temperature-dependent (it increases with temperature) and usually ranges from 2000 K to 4000 K

TABLE 2.4 General Characteristics of Frequently Used NTC Thermistors

Parameter	
Temperature range	−100 °C to 450 °C (not in a single unit)
Resistance at 25 °C	0.5 Ω to 100 MΩ (1 kΩ to 10 MΩ is common)
Characteristic temperature, B	2000 K to 5500 K
Maximal temperature	>125 °C (300 °C in steady state; 600 °C intermittently)
Dissipation constant (δ)	1 mW/K in still air 8 mW/K in oil
Thermal time constant	1 ms to 22 s
Maximal power dissipation	1 mW to 1 W

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Resistance Measurement Methods

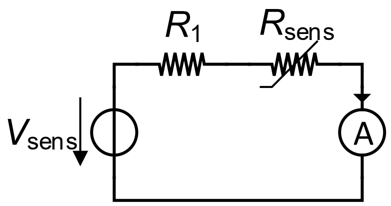
- Resistance measurement can be classified in two different methods
  - **Deflection** methods and
  - **Null** methods
- Deflection methods sense the drop in **voltage** across the resistance to be measured or the **current** through it or both
- Null methods are based on measurement **bridges**

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

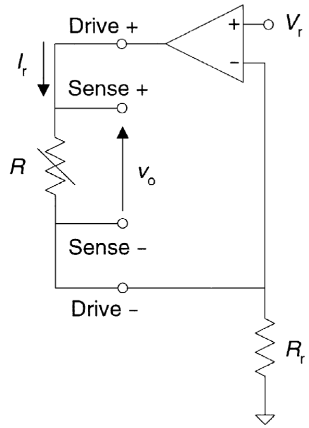
Resistance Measurement

Current sensing

(imposing constant voltage)

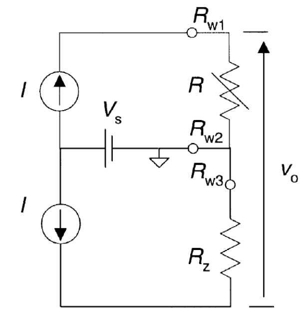


$$R = R_0 \cdot (1 + x)$$



Voltage sensing

(imposing constant current)



- The output voltage for a linear sensor is

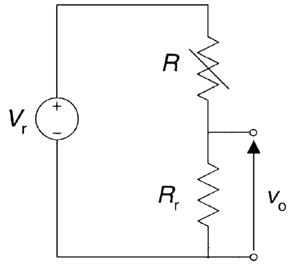
$$v_0 = I_r \cdot R = \frac{V_r}{R_r} \cdot R_0 \cdot (1 + x)$$

- which consists of small fluctuations (due to x) superimposed on a very large offset voltage corresponding to $x = 0$. The latter can be canceled with the circuit shown on the right

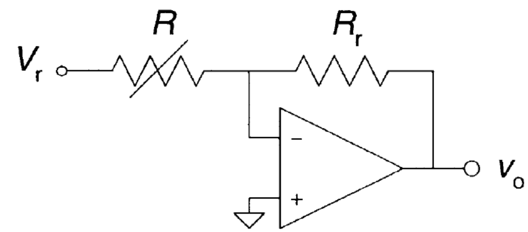
$$v_0 = I \cdot (R - R_z) = I \cdot R_0 \cdot x$$

- which is linear assuming $R_z = R_0$. It additionally delivers a differential voltage
- Using a four-wire sensor, wire resistances does not contribute to the output voltage

Voltage Dividers



$$v_0 = \frac{R_r}{R_r + R} \cdot V_r$$

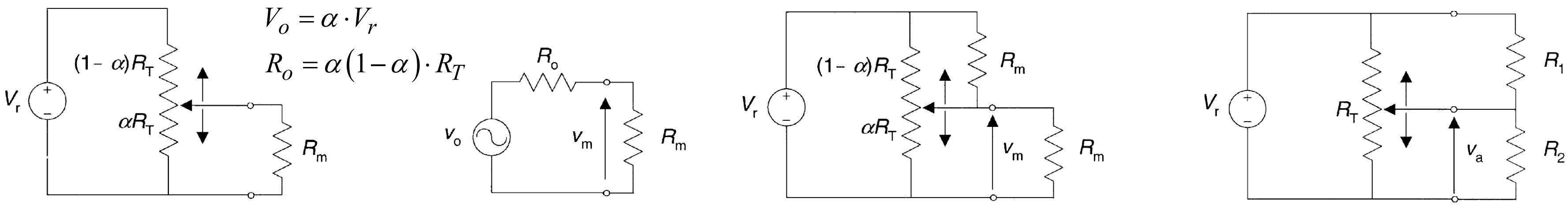


- Input resistance of voltmeter much higher than R we have
$$v_0 = \frac{R_r}{R_r + R} \cdot V_r$$
- from which we can calculate
$$R = R_r \cdot (V_r - v_0)/v_0$$
- or
$$R = R_r \cdot v_0/(V_r - v_0)$$
- if R and R_r are swapped
- Well suited to sensors with large resistance variation and also nonlinear sensors such as NTC thermistors because the nonlinearity of the relationship between v_0 and R permits thermistor linearization

- The voltage-resistance relation in voltage dividers is **not linear** because the current in the circuit depends on the unknown resistance
- If that current were constant, the drop in voltage across a linear resistive sensor would be a linear function of the measurand
- For sensors whose resistance decreases with the applied input, the circuit on the right injects a constant current into the sensor and yields an output voltage
$$v_0 = -\frac{R_r}{R} \cdot V_r$$
- which increases with the measurand (i.e. R decreasing)

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Potentiometers



a) b) c) d)

- Assuming the input resistance of the voltmeter is given by R_m, the measured voltage is

$$v_m = \frac{R_m}{R_0 + R_m} \cdot v_0 = \frac{\alpha}{1 + \frac{\alpha(1-\alpha)}{k}} \cdot V_r \cong \alpha \cdot V_r \text{ for } k \triangleq \frac{R_m}{R_T} \gg 1$$

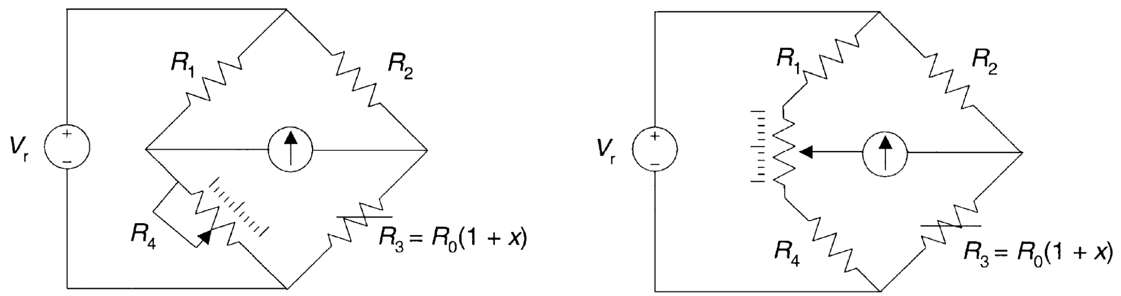
- which is linear only if $k \gg 1$ or $R_m \gg R_T$
- A simple way to reduce the loading error without increasing R_m is to place a resistor equal to R_m on the top side of the potentiometer (c). The measured voltage is

$$v_m = \frac{\alpha(k + 1 - \alpha)}{2\alpha(1-\alpha) + k} \cdot V_r$$

- The added resistor forces $v_m / V_r = 2$ at the central position $\alpha = 0.5$ thus achieving zero error at that point
- By using two different resistors (d) we can obtain zero error at any desired point

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

The Wheatstone Bridge



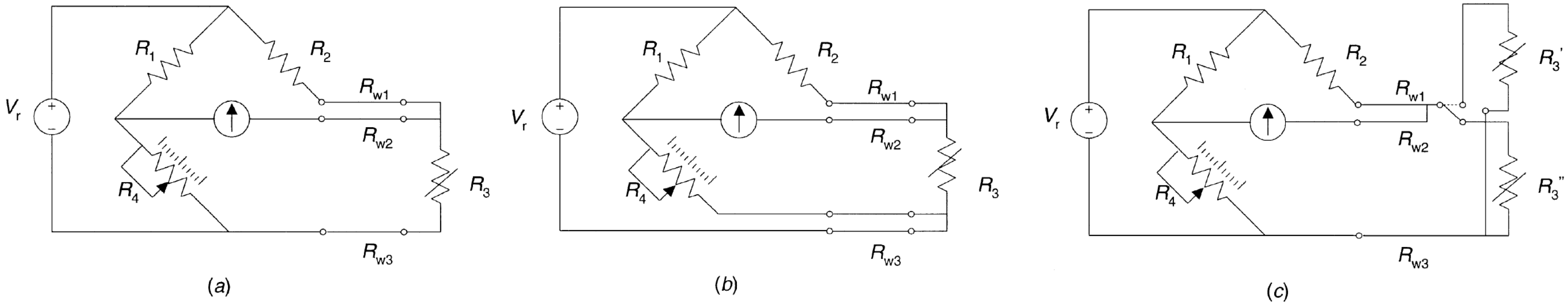
- Principle is to adjust the value of a standard resistor R_4 in order to achieve a balanced condition by means of a feedback system (electric or manual) until the current through the galvanometer or other null indicator is zero resulting in

$$R_3 = R_4 \cdot \frac{R_2}{R_1}$$

- The changes in R_3 is then proportional to the corresponding changes imposed on R_4 in order to balance the bridge
- The right figure shows an arrangement for eliminating the influence that the contact resistance in the adjustable arm has on the measurement by including that resistance in series with the central arm (bridge), through which there is no current when reaching the balance

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Wheatstone Bridge with Long Leads



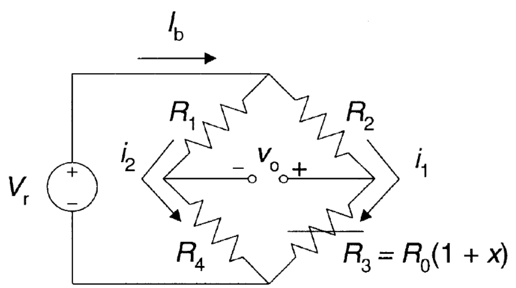
- The resistance and temperature dependence of long wires can result in important errors. The three-wire method (a) solves this problem. Wires 1 and 3 must be equal and undergo the same temperature changes
- The characteristics of wire 2 are irrelevant because in the balance condition there is no current through the bridge central arm
- The relative error in the measurement of R_3 is

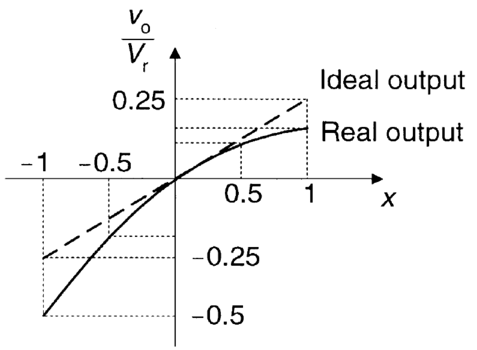
$$\varepsilon = \frac{R_4 R_2/R_1 - R_3}{R_3} = \frac{R_w}{R_3} \cdot \left(1 - \frac{R_4}{R_1}\right) \text{ where } R_w = R_{w1} = R_{w3}$$

- Figure b shows an alternative circuit with the same objective having a similar error which decreases for $R_3 \gg R_w$
- Figure c shows how to apply this method to several sensors using a single set of three long wires

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Deflection Measurement of Wheatstone Bridge





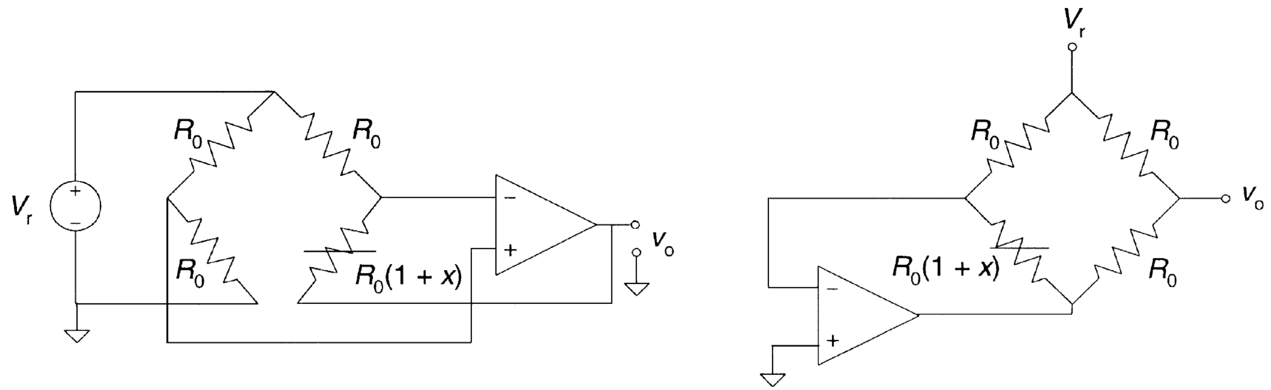
- Instead of measuring the action needed to restore balance on the bridge, the deflection method measures the voltage difference between both voltage dividers or the current through a detector bridging them
- The voltage difference between both branches is

$$v_0 = \left(\frac{R_3}{R_2+R_3} - \frac{R_4}{R_1+R_4}\right) \cdot V_r = \frac{k \cdot x}{(k+1)(k+1+x)} \cong \frac{k \cdot x}{(k+1)^2} \text{ for } x \ll k+1$$

- where $k = R_1/R_4 = R_2/R_0$ corresponding to the situation when the bridge is balanced ($x = 0$)

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Analog Linearization of Resistive Sensor Bridges



(a) (b)

- The Wheatstone bridge can be modified to maintain the current constant through it
- For an ideal op amp, the output is given by

$$v_0 = -x \cdot \frac{V_r}{2}$$

- This method, however, requires the bridge to have five terminals accessible. The bridge must be opened at one of the junctions where the sensor is connected
- The pseudo-bridge in Figure b overcomes that limitation. The opamp must have low offset voltage, input currents, and drifts
- If x can be negative, the op amp must operate on a dual (split) power supply

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Outline

- Signal conditioners for resistive sensors
- **Signal conditioners for capacitive sensors**

Introduction

- Capacitive sensors cover a wide range of applications
- They are well suited for high integration
- They consume little power
- However they are prone to parasitics

Basic Capacitive Sensors

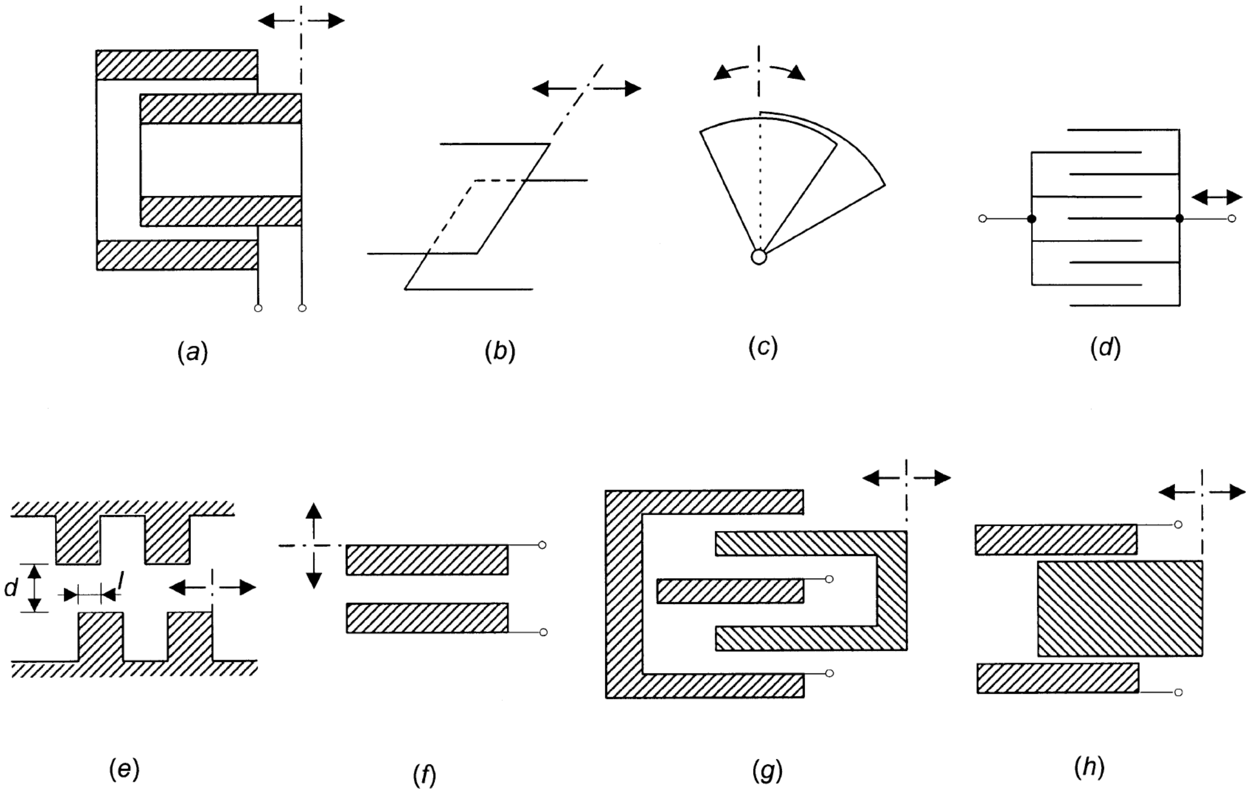
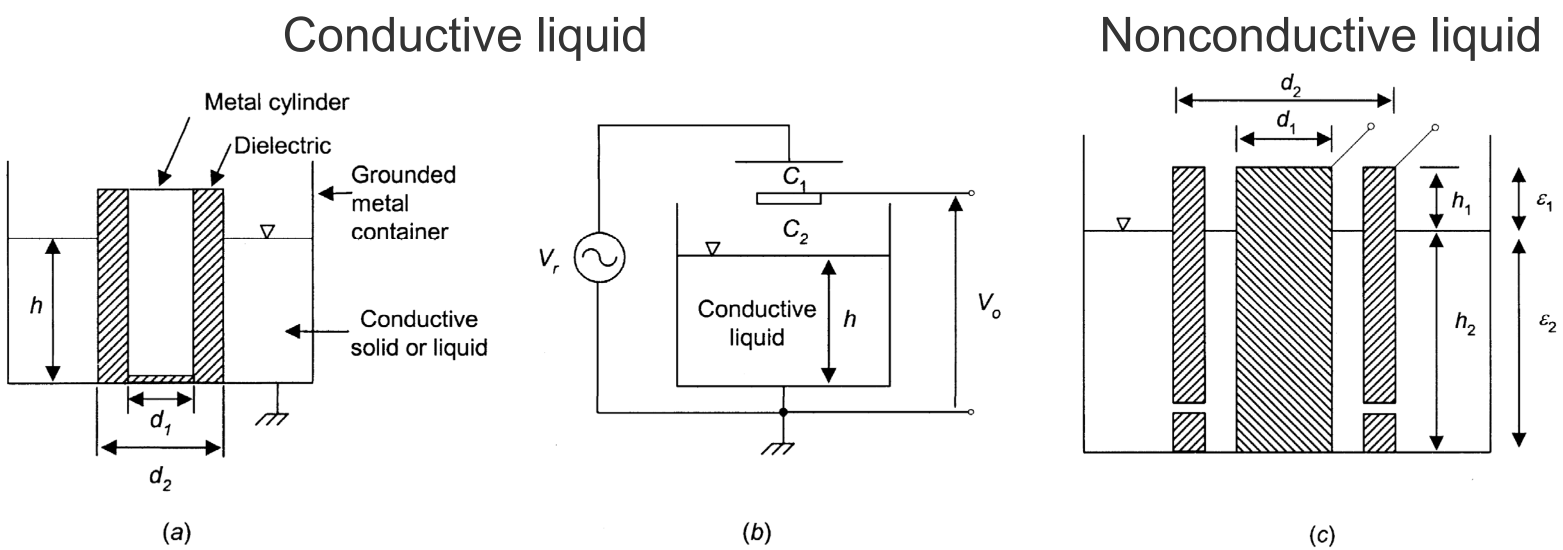


Figure 4.3 Different arrangements for capacitive sensors based on (*a–e*) a variation of area, (*f*) plate separation, and (*g*, *h*) dielectric.

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Capacitive Level Gages



- Capacitance change through area variation
- Linear with height h

$$C = \frac{2\pi \cdot \varepsilon \cdot h}{\ln\left(\frac{d_2}{d_1}\right)}$$

- Capacitance change due to distance variation
- Highly conductive liquid (mercury, water, etc.)

$$v_o = \frac{C_1}{C_1 + C_2} \cdot v_r$$

$$C_1 = const.$$

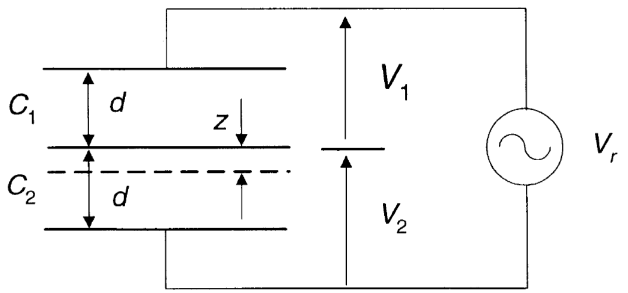
$$C_2 \propto 1/h$$

- Capacitance change due to dielectric variation
- Conductive cylinders are coaxial

$$C \cong \frac{2\pi\left(\varepsilon_1 h_1 + \varepsilon_2 h_2\right)}{\ln\left(\frac{d_2}{d_1}\right)}$$

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

Differential Capacitors

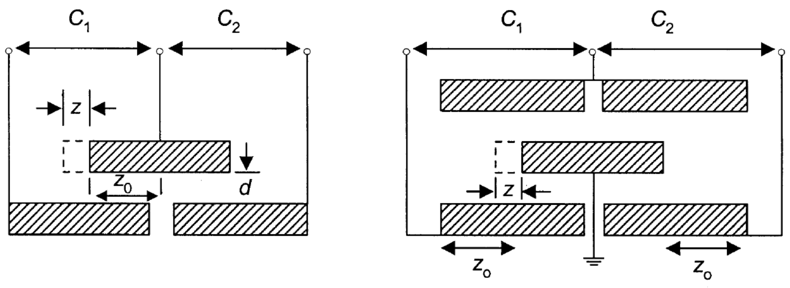


$$C_1 = \frac{\varepsilon A}{d+z} \qquad V_1 = \frac{C_2}{C_1+C_2}\cdot V_r = \frac{d+z}{2d}\cdot V_r$$

$$C_2 = \frac{\varepsilon A}{d-z} \qquad V_2 = \frac{C_1}{C_1+C_2}\cdot V_r = \frac{d-z}{2d}\cdot V_r$$

- The differential voltage $V_1 - V_2$ is **linear** with displacement z

$$V_1 - V_2 = \frac{z}{2d}\cdot V_r$$



$$C_1 = \varepsilon\frac{w(z_0 - z)}{d} = \varepsilon\frac{w}{d}z_0\frac{z_0 - z}{z_0} = C_0\frac{z_0 - z}{z_0}$$

$$C_2 = \varepsilon\frac{w(z_0 + z)}{d} = \varepsilon\frac{w}{d}z_0\frac{z_0 + z}{z_0} = C_0\frac{z_0 + z}{z_0}$$

$$V_1 = \frac{C_2}{C_1+C_2}\cdot V_r = \frac{z_0+z}{2z_0}\cdot V_r$$

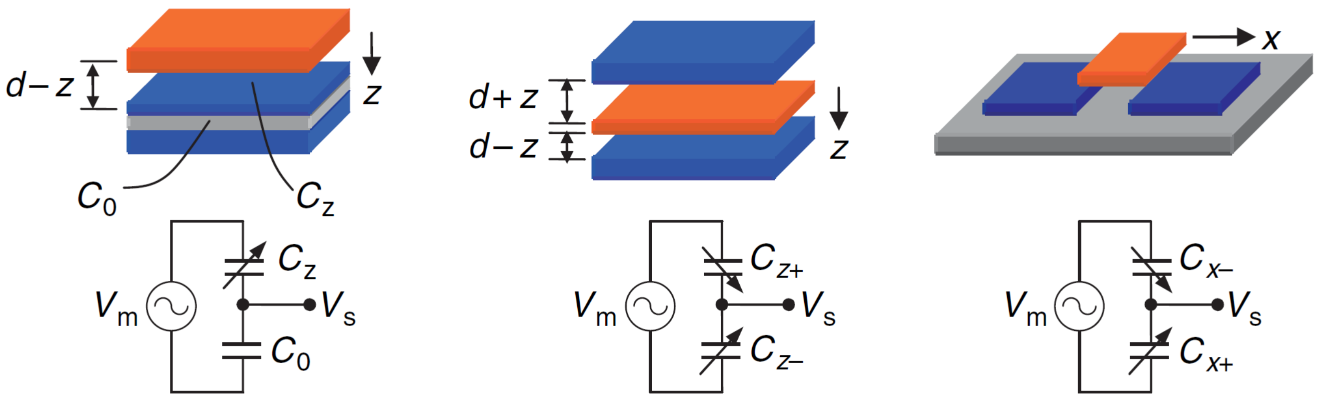
$$V_2 = \frac{C_1}{C_1+C_2}\cdot V_r = \frac{z_0-z}{2z_0}\cdot V_r$$

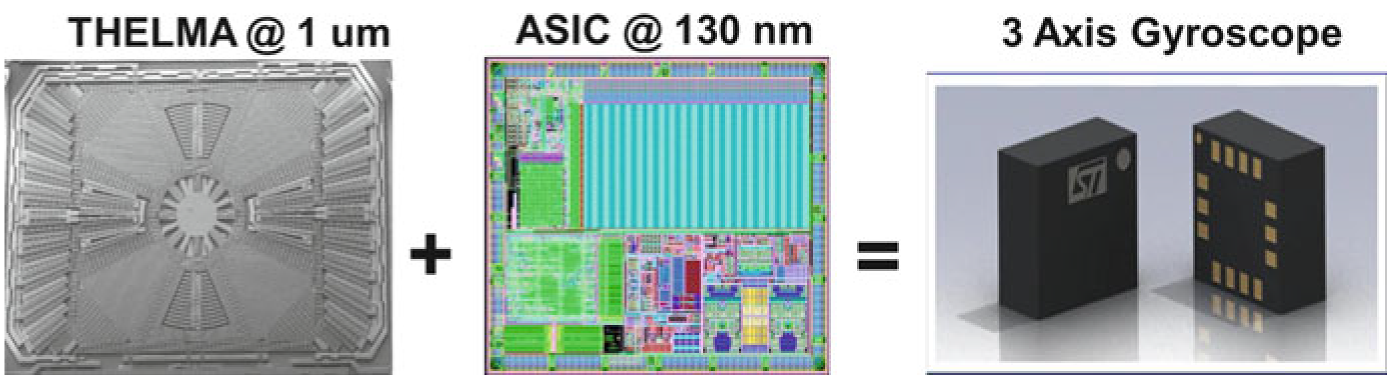
- The differential voltage $V_1 - V_2$ is again **linear** with displacement z

$$V_1 - V_2 = \frac{z}{z_0}\cdot V_r$$

R. Pallas-Areny, J. G. Webster, *Sensors and Signal Conditioning*, 2nd ed., Wiley, 2001.

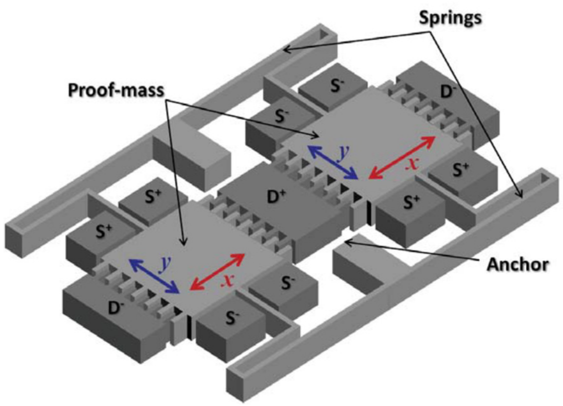
Accelerometers

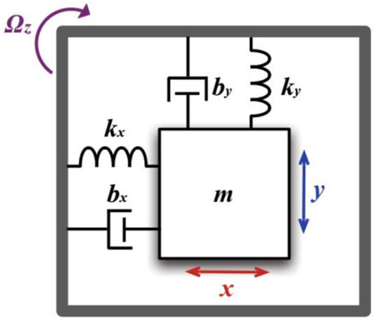


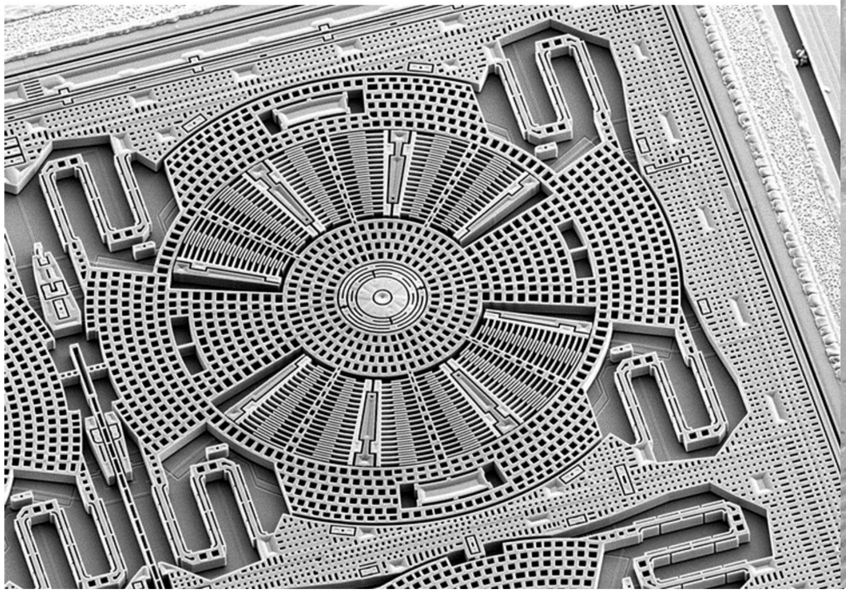


H. Xie, G. K. Fedder, R. E. Sulouff, *Accelerometers*, Ed.: Y. Gianchandani, O. Tabata, H. Zappe, Comprehensive Microsystems, Elsevier, 2008.

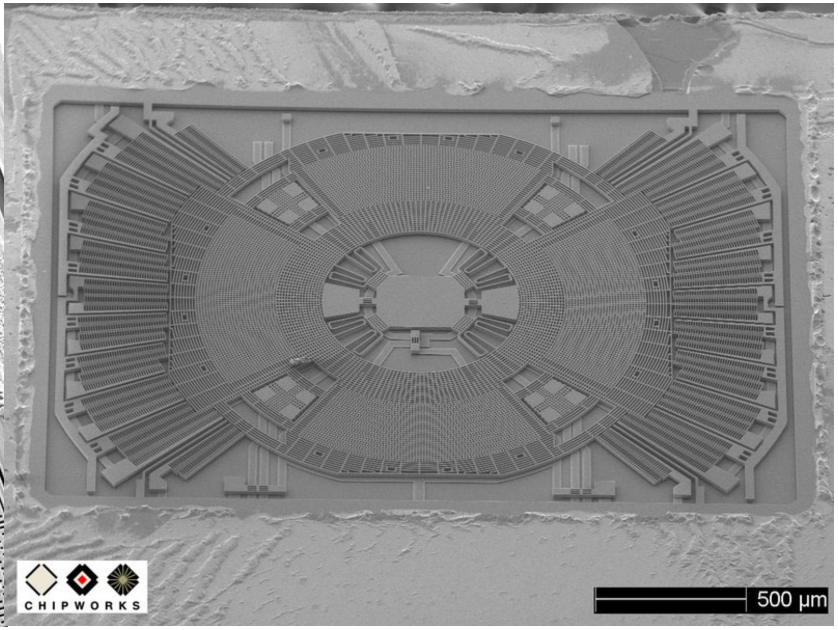
Gyroscopes







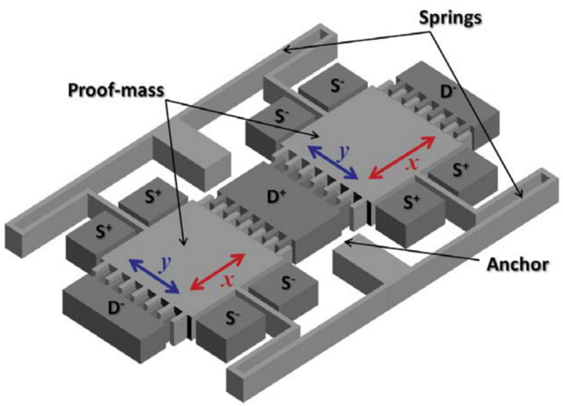
Analog Devices ADXRS295 Gyroscope

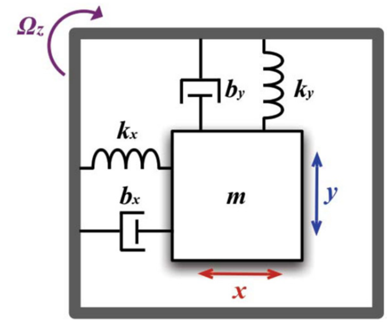


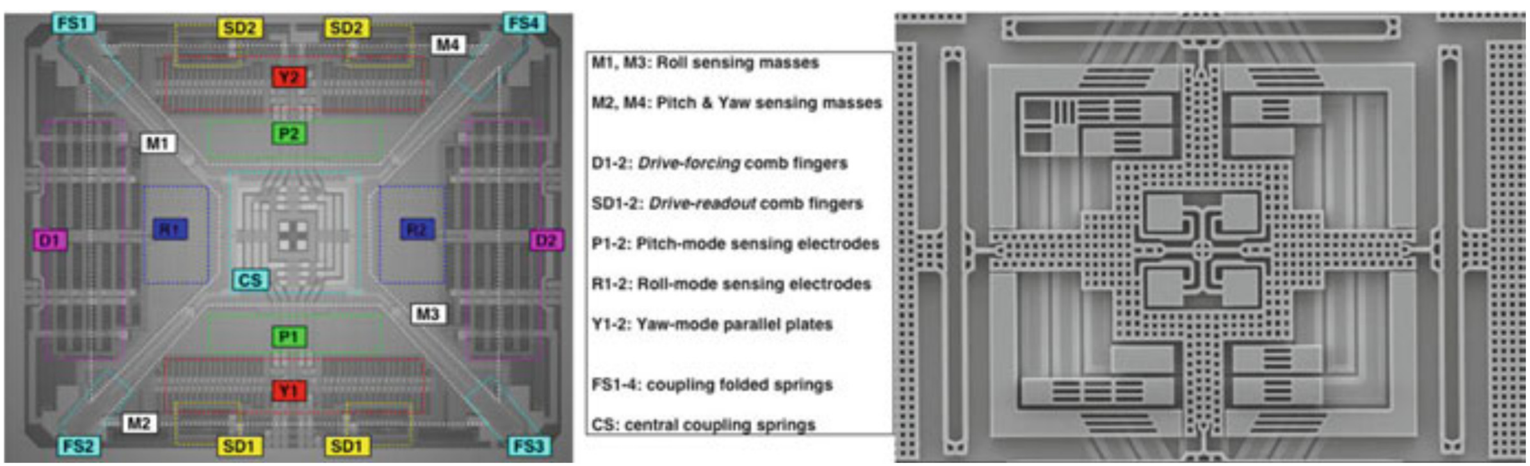
ST LYPR540AH Tri-axis MEMS gyroscope

D. E. Serrano, "Design and Analysis of MEMS Gyroscopes, IEEE Sensors, 2013.

Tri-axial Coriolis Gyroscope

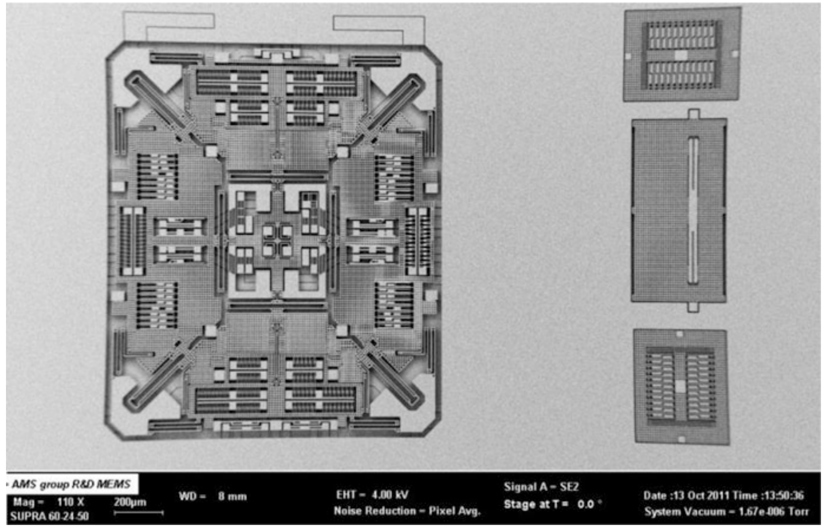


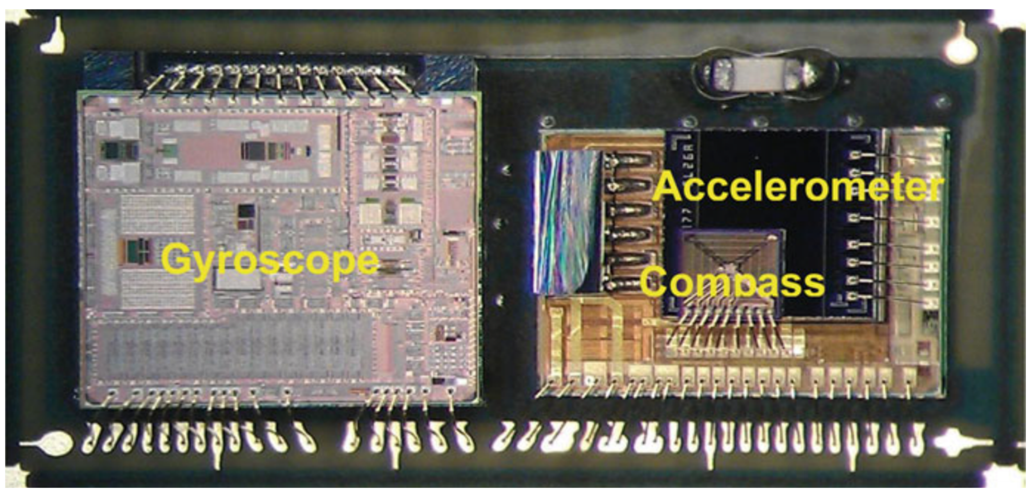




📖 D. E. Serrano, "Design and Analysis of MEMS Gyroscopes, IEEE Sensors, 2013.
📖 B. Vigna, E. Lasalandra, and T. Ungaretti, Motion MEMS and Sensors, Today and Tomorrow, in Advances in Analog Circuit Design, Springer, 2012.

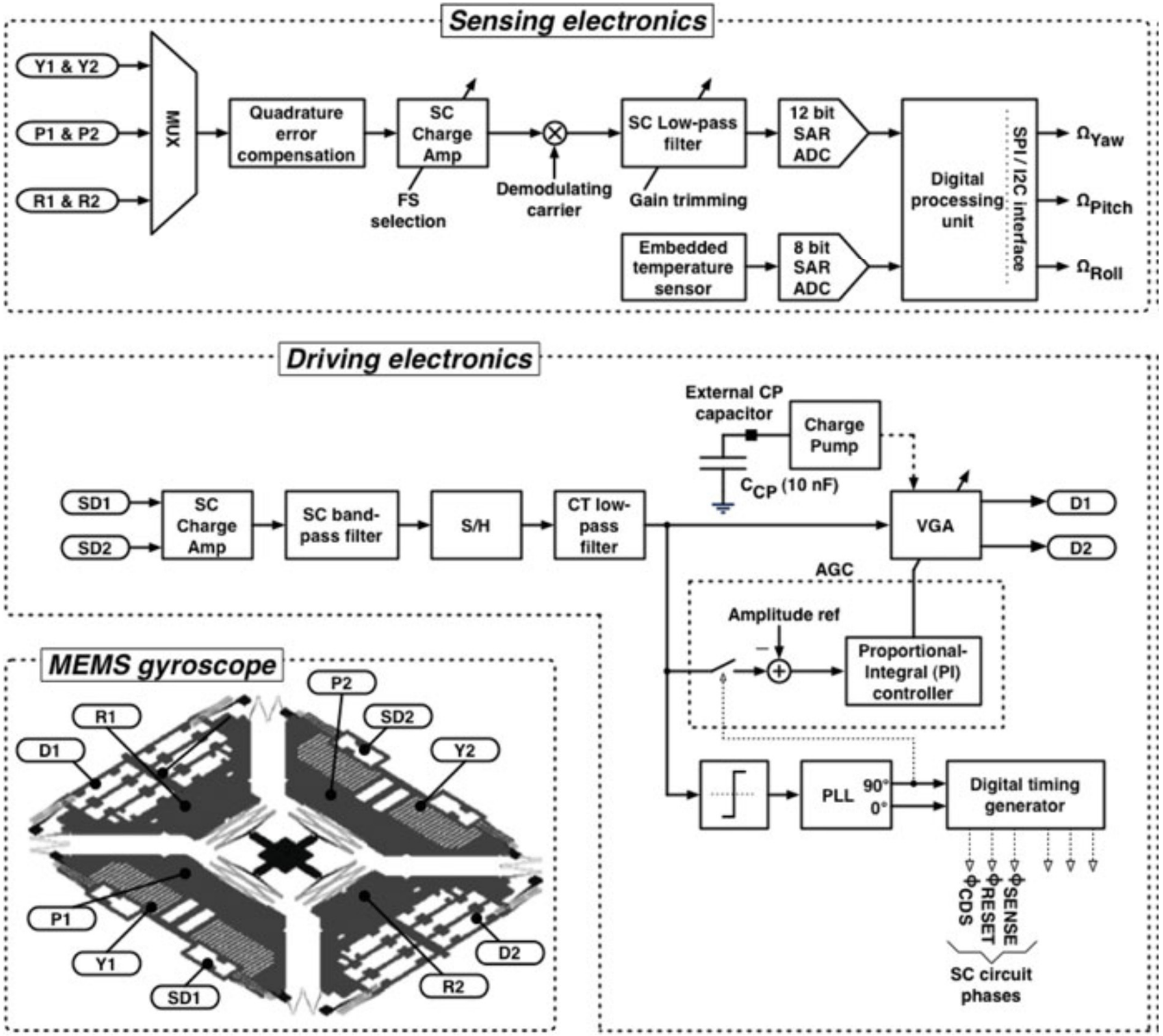
Six and Nine Degree of Freedom Motion Sensors





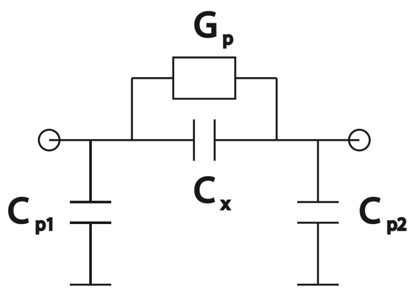
B. Vigna, E. Lasalandra, and T. Ungaretti, Motion MEMS and Sensors, Today and Tomorrow, in Advances in Analog Circuit Design, Springer, 2012.

System Architecture Block Diagram

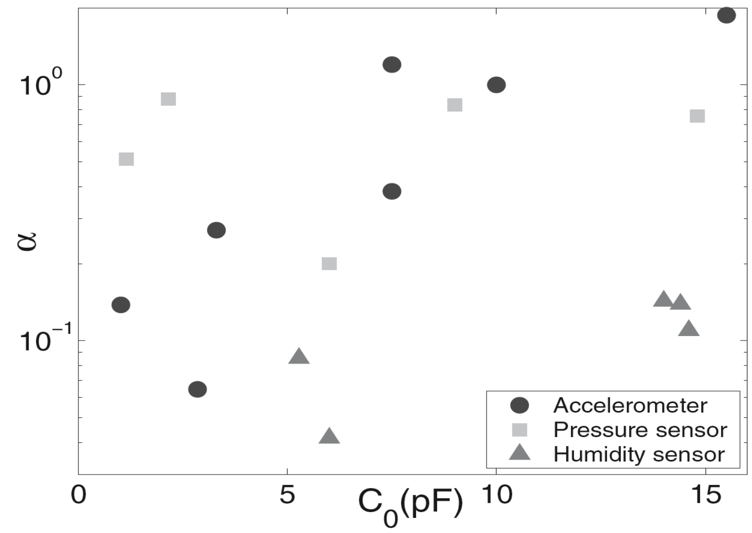


B. Vigna, E. Lasalandra, and T. Ungaretti, Motion MEMS and Sensors, Today and Tomorrow, in Advances in Analog Circuit Design, Springer, 2012.

Mean Capacitance and Capacitance Variation



Single capacitive sensor model with sense capacitance C_x, parasitic shunt conductance G_p and parasitic capacitances C_{p1} and C_{p2}



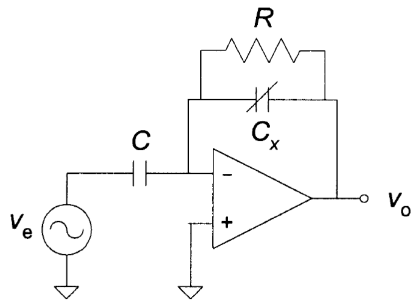
- The mean capacitance C_0 and the relative full-scale deviation α are defined as

$$C_0 = \frac{C_{x,min} + C_{x,max}}{2} \text{ and } \alpha \triangleq \frac{\Delta C}{C_0}$$

- Various capacitive transducer applications show a wide range of mean capacitance C_0 and relative full scale deviations α which are represented above
- Values of α and C_0 depend on the excitation (acceleration, pressure, humidity), the physical input range of the application, the sensor structure and the technology

W. Bracke, R. Puers and C. van Hoof, *Ultra Low Power Capacitive Sensor Interfaces*, Springer, 2007.

Capacitance-to-Voltage Converters

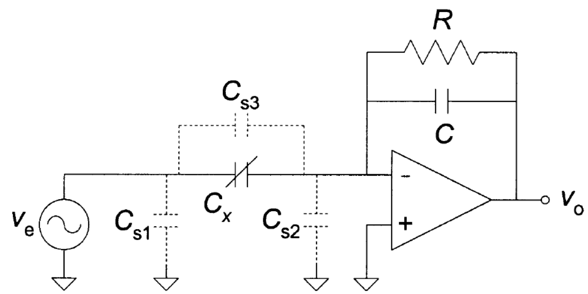


- The capacitance change with the distance between the two plates according to

$$C_x = \epsilon \frac{A}{d(1+x)} = \frac{C_0}{1+x}$$

- where $C_0 = \epsilon A/d$
- If the op amp is assumed ideal and R is disregarded, the output voltage is

$$v_o = -\frac{Z_x}{Z} \cdot v_e = -\frac{C}{C_0}(1+x)v_e$$

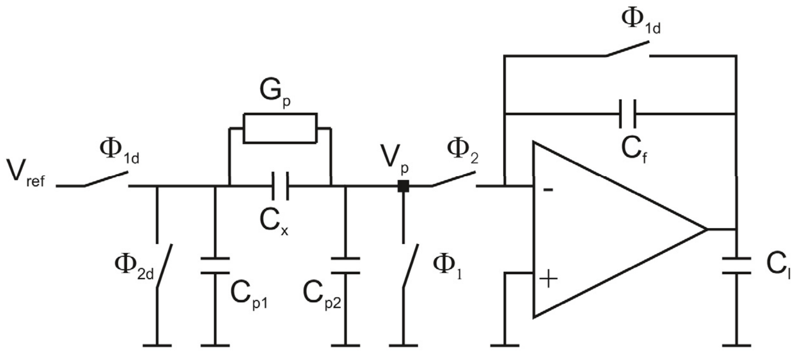


- Applies a voltage to the sensor and measures the resulting current by converting it to a voltage through C
- Disregarding R (op amp bias) and stray capacitance C_{s3}, the output voltage is

$$v_o = -\frac{C_x}{C} \cdot v_e$$

- which is proportional to the sensor capacitance C_x
- Note that stray capacitances C_{s1} and C_{s2} do not contribute to the output

Capacitance-to-Voltage Converters



$$V_{out} = \frac{C_x}{C_f} \cdot V_{ref}$$

- The conventional non-inverting SC amplifier can work as a capacitance-to-voltage converter and is known to be **insensitive** to parasitic capacitors C_{p1} and C_{p2}
- During phase Φ_1, the sense capacitor C_x is charged to V_{ref}, while during phase Φ_2, this charge is transferred to the feedback capacitor C_f leading to an output voltage

$$V_{out} = \frac{C_x}{C_f} \cdot V_{ref}$$

- In reality the charge transfer is affected by the leakage through the conductance G_p, the OTA finite transient response, offset and finite gain

Main Capacitive Readout Circuits

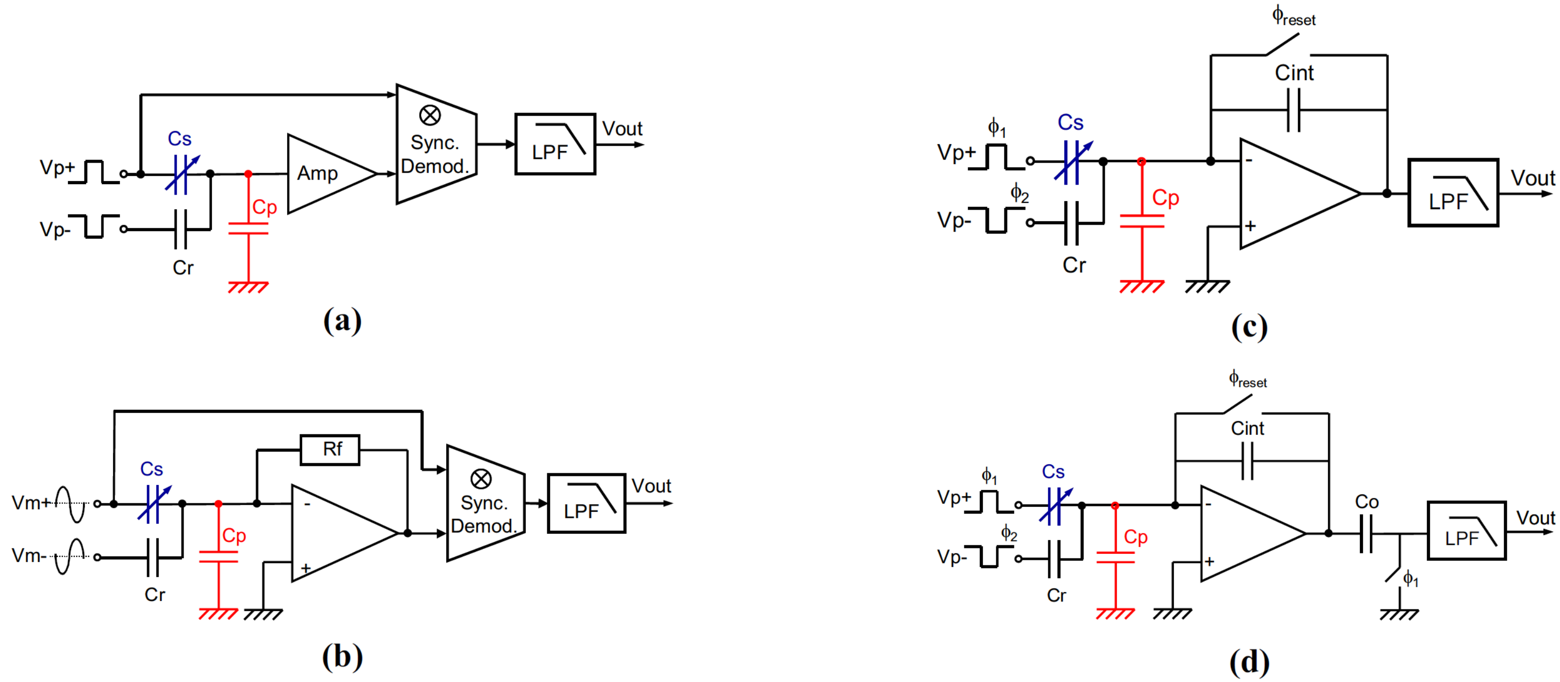
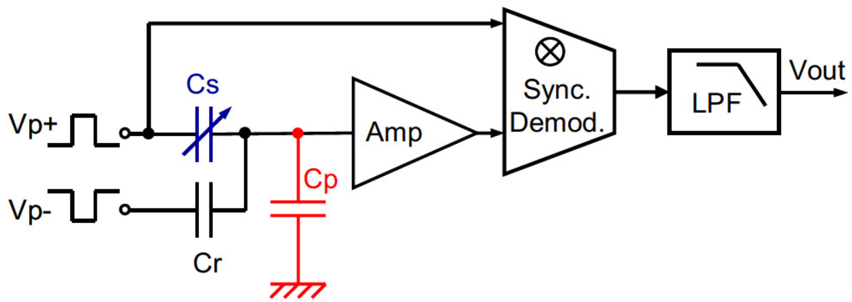


Figure 1: Simplified block diagram of various capacitive readout circuits: **(a)** ac-bridge with voltage amplifier; **(b)** Transimpedance amplifier; **(c)** Switch-capacitor circuit; **(d)** Switched-capacitor circuit with kT/C noise reduction.

- Capacitive readout circuits can be divided into three groups:
  - ac-bridge with voltage amplifier,
  - transimpedance amplifier,
  - switched-capacitor circuits

N. Yazdi, H. Kulah, and K. Najafi, "Precision readout circuits for capacitive microaccelerometers", IEEE Sensors Conference, 2004.

Square Wave Driven AC-bridge with Voltage Amplifier



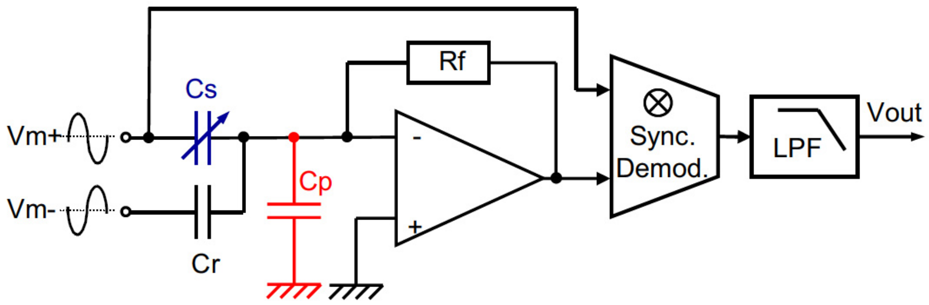
- The sense and reference capacitors C_s and C_r are driven by square waves of opposite signs
- The amplitude of the half bridge output which is proportional to $\Delta C = C_s - C_r$ is then amplified by a voltage amplifier and demodulated resulting in

$$V_{out} = \frac{\Delta C}{2C_0 + C_p} \cdot A_v \cdot V_{ref} \text{ where } C_0 = \frac{C_s + C_r}{2}$$

- This architecture does not eliminate the effect of the parasitic capacitance C_p which can significantly degrade the performance

N. Yazdi, H. Kulah, and K. Najafi, "Precision readout circuits for capacitive microaccelerometers", IEEE Sensors Conference, 2004.

Harmonic AC-bridge with Transimpedance Amplifier



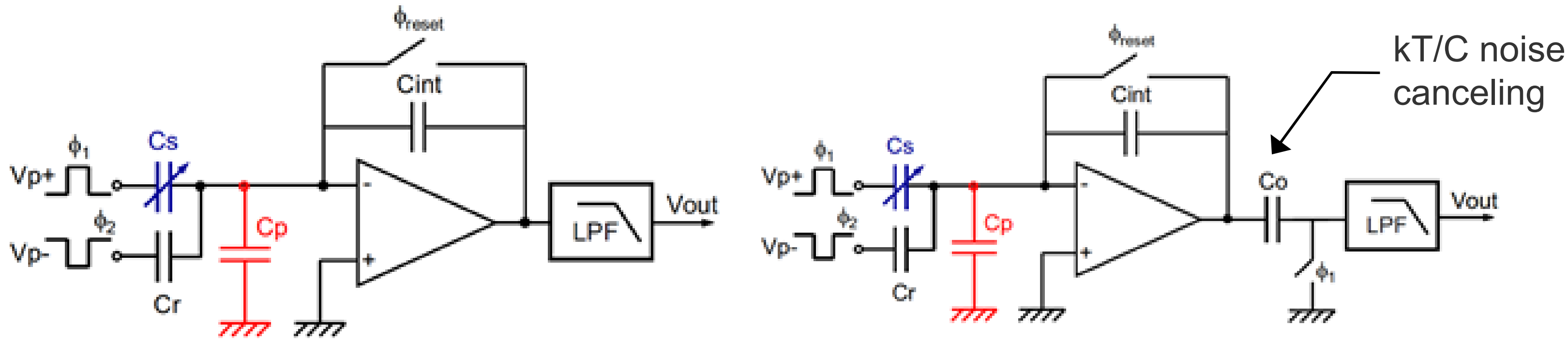
- The bridge output is held at virtual ground by an op-amp with resistive feedback (transimpedance amplifier), which reduces the effect of C_p
- The drive signals are sinusoidal, with frequency f_{drive}, to avoid errors induced by distortion
- If f_{drive} is smaller than the amplifier bandwidth, the output voltage after demodulation equals

$$V_{out} = 2\pi f_{drive} \cdot V_m \cdot R_f \cdot \Delta C$$

- Requires harmonic oscillator

N. Yazdi, H. Kulah, and K. Najafi, “Precision readout circuits for capacitive microaccelerometers”, IEEE Sensors Conference, 2004.

Switched Capacitor Interface



- Capacitor C_s is connect to V_{p+} during phase Φ_1 while the integration capacitor C_{int} and the parasitic capacitor C_p are discharged
- Capacitor C_s is then connect to ground while C_r is connected to V_{p-}
- The charge difference $\Delta C \cdot V_{ref}$ is integrated in the feedback capacitor leading to

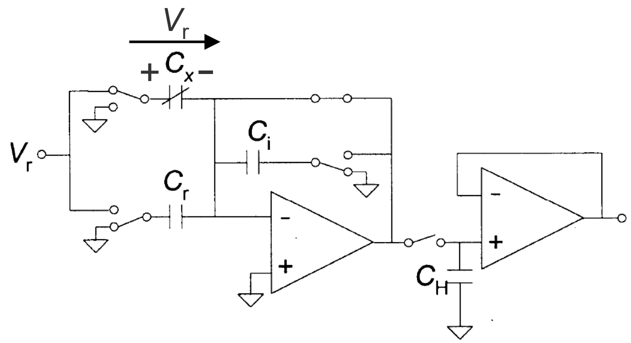
$$V_{out} = \frac{\Delta C}{C_{int}} \cdot V_p \text{ with } \Delta C = C_s - C_r$$

- This circuit also eliminates the effect of the parasitic capacitances and does not need complex driving voltages
- The offset and 1/f noise is cancelled by autozero
- The kT/C_{int} thermal noise can be eliminated by sampling and deducting from the output

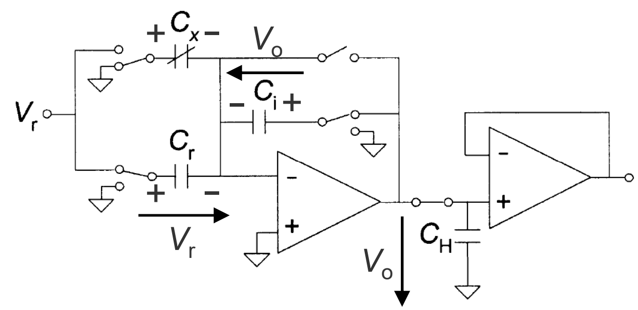
W. Bracke, R. Puers and C. van Hoof, *Ultra Low Power Capacitive Sensor Interfaces*, Springer, 2007.

Switched Capacitor Charge Redistribution Method

Autozero phase



Measurement phase



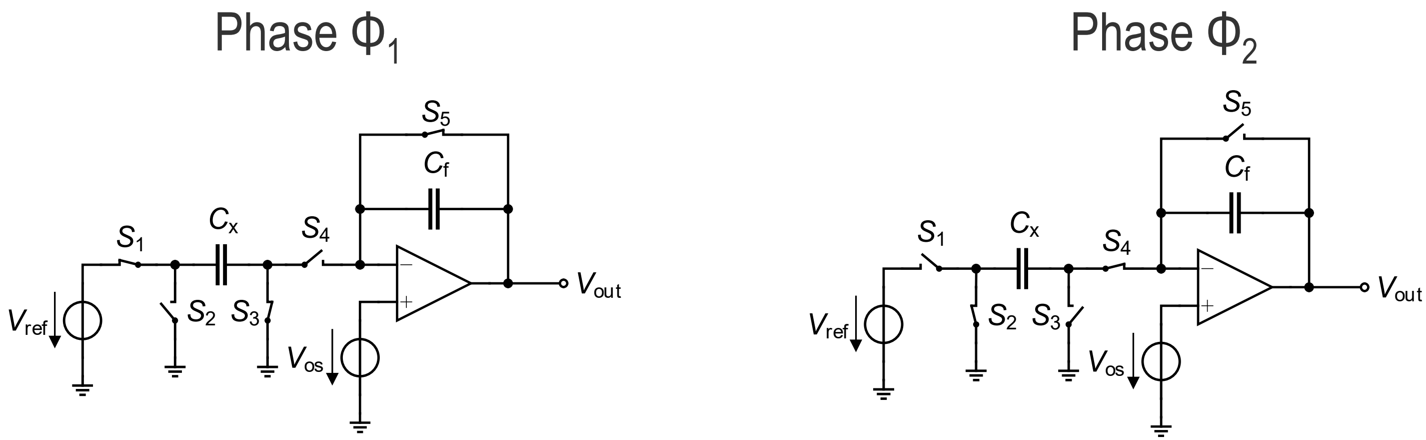
- In the autozero phase the sensor C_x is charged at V_r while the other capacitors C_r and C_i are discharged
- In the measurement phase, C_x is discharged, C_r is connected to V_r
- Applying the charge conservation at the virtual ground between the autozero and measurement phases leads to

$$-C_x \cdot V_r + 0 + 0 = 0 - C_r \cdot V_r - C_i \cdot V_o$$

- Resulting in the output voltage proportional to the difference between C_x and C_r

$$V_o = \frac{C_x - C_r}{C_i} \cdot V_r$$

Effect of the Amplifier Offset Voltage



- The amplifier offset voltage introduces an offset voltage at the output

$$V_{out} = \frac{C_x}{C_f} \cdot V_{ref} + \left(1 + \frac{C_x}{C_f}\right) \cdot V_{os}$$

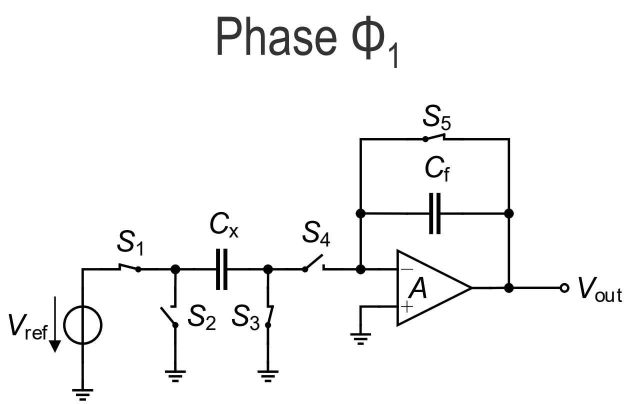
- And results in a smaller value estimation of the capacitance C_x

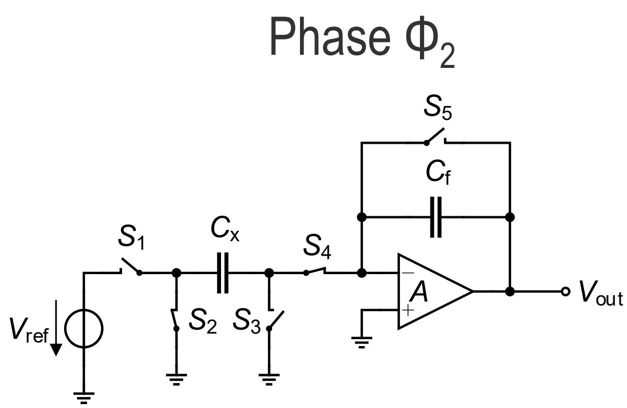
$$\widetilde{C_x} = C_x + (C_x + C_f) \cdot \frac{V_{os}}{V_{ref}}$$

- Or a relative error

$$\varepsilon \triangleq \frac{\widetilde{C_x} - C_x}{C_x} = \left(1 + \frac{C_x}{C_f}\right) \cdot \frac{V_{os}}{V_{ref}}$$

Effect of the Amplifier Finite Gain





- The amplifier finite gain A introduces a reduction of the gain

$$V_{out} = \frac{C_x/C_f}{1 + \frac{1 + C_x/C_f}{A}} \cdot V_{ref}$$

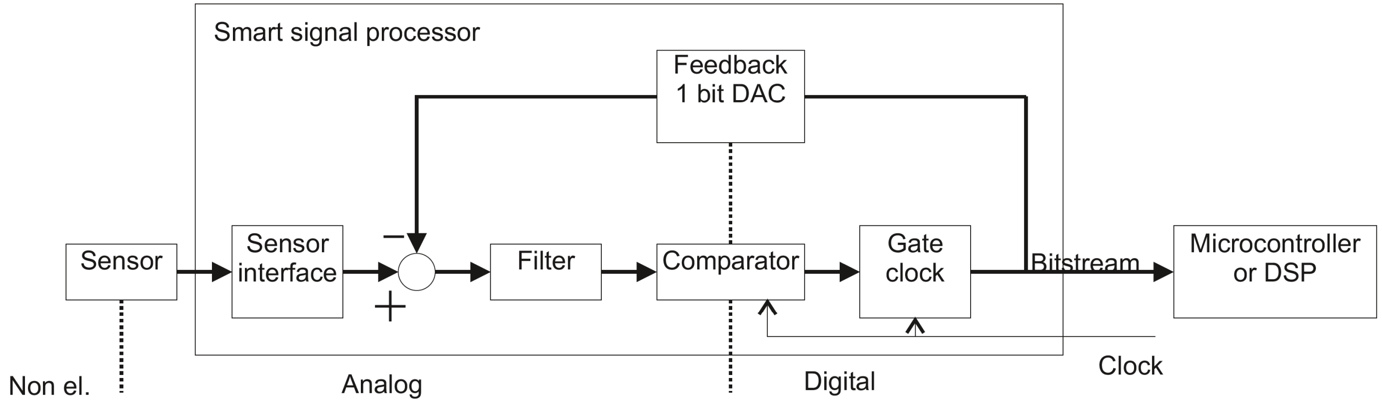
- And results in a smaller value estimation of the capacitance C_x

$$\widetilde{C_x} = \frac{C_x}{1 + \frac{1 + C_x/C_f}{A}}$$

- Or a relative error

$$\varepsilon \triangleq \frac{\widetilde{C_x} - C_x}{C_x} \cong \frac{1 + C_x/C_f}{A}$$

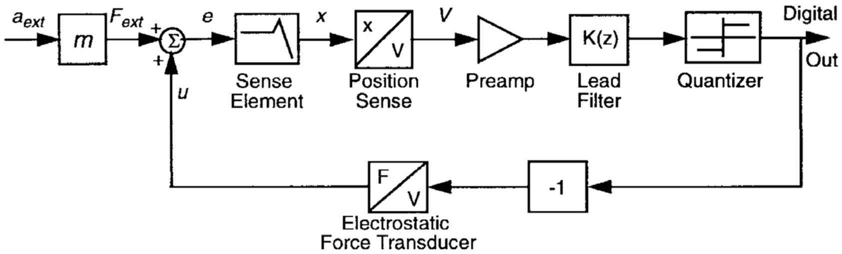
Smart Sensor System with First Order ΣΔ Modulator

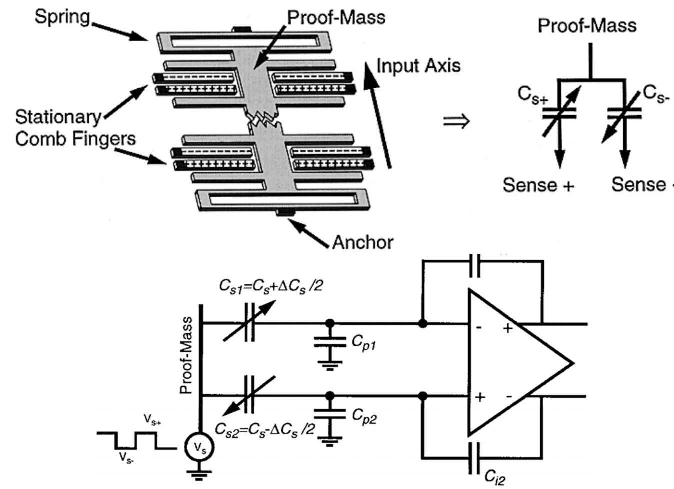


- Many capacitive sensor interfaces use the sensor directly in an oversampled ΣΔ modulator
- High power consumption, because the capacitors must be charged and discharged at the rate of the high oversampling clock of the modulator
- Most of these sensor interfaces are designed for closed loop accelerometers where the electrostatic feedback force is used to keep the sensor mass in its balanced position for high linearity
- The mechanical transfer characteristic acts as a 2nd-order filter in the ΣΔ loop, which makes the stability strongly dependent on the specific sensor
- Moreover, these systems need an important start-up time to bring the sensor mass close to equilibrium
- Not very suited for low-power (except if they are used as incremental ADC)

W. Bracke, R. Puers and C. van Hoof, *Ultra Low Power Capacitive Sensor Interfaces*, Springer, 2007.

Accelerometer Force-feedback Interface

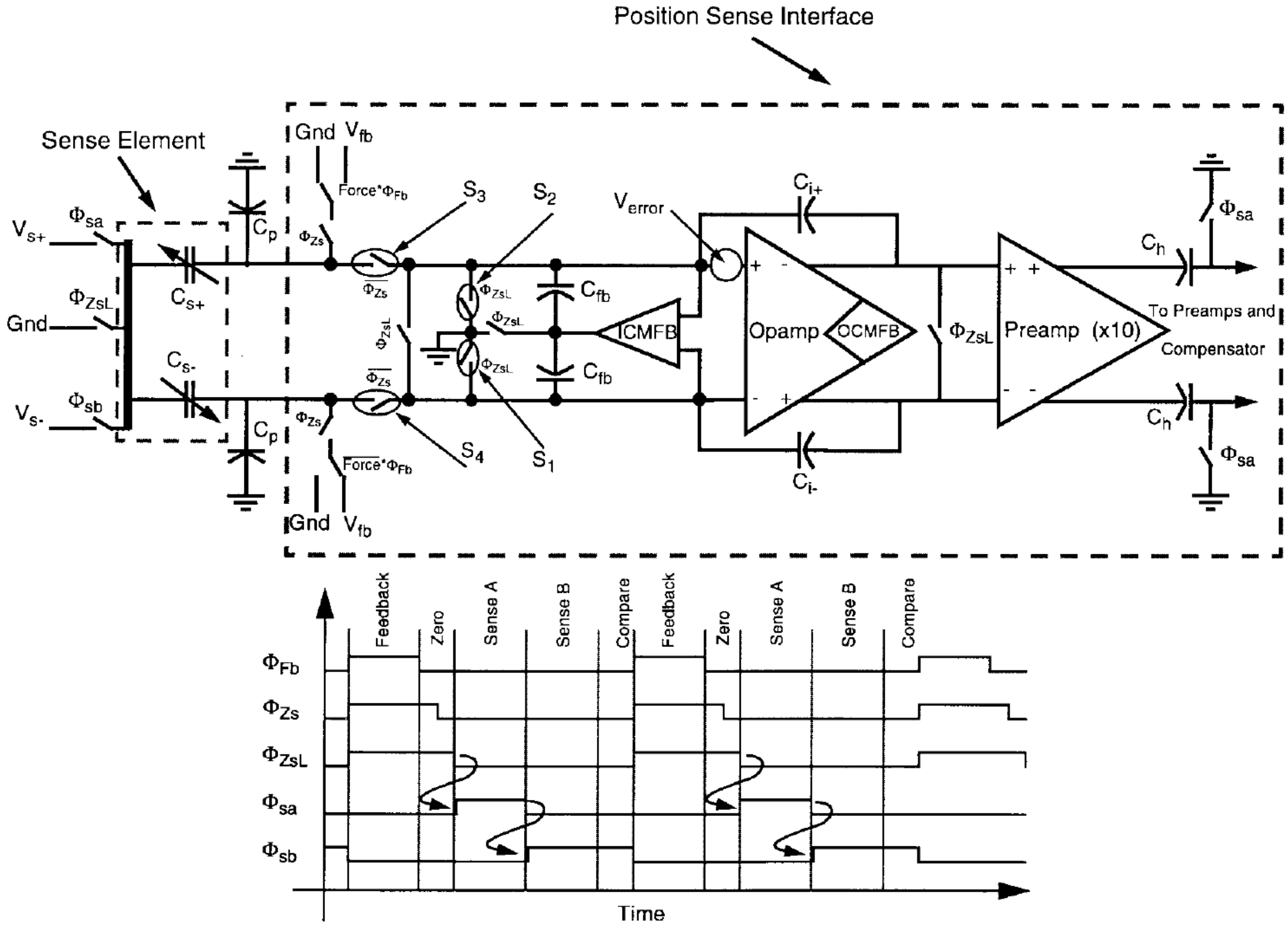




- Many MEMS accelerometers use the sensor directly in a ΣΔ loop
- The electrostatic feedback force is used to keep the sensor mass in its balanced position to improve its linearity
- The accelerometer usually has a 2nd-order transfer function, which combined with a 1st-order ΣΔ modulator gives a 3rd-order system which stability needs to be carefully studied
- Since the ΣΔ modulator operates at a high oversampling clock frequency, these systems usually show a fairly high power consumption

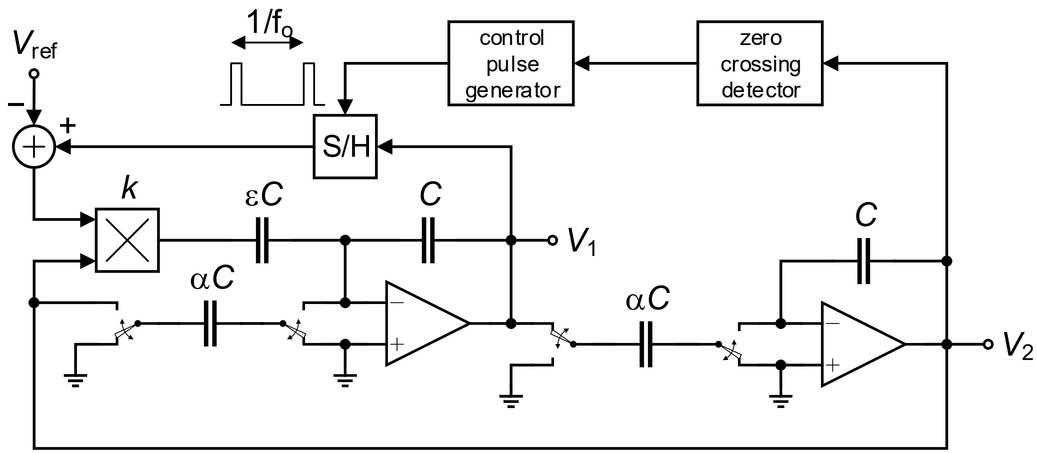
M. Lemkin and B. E. Boser, JSSC, April 1999.

Accelerometer Force-feedback Interface



M. Lemkin and B. E. Boser, JSSC, April 1999.

Capacitance-to-Frequency Interface



- Quadrature SC oscillator combined with a automatic gain control (AGC) circuit for controlling the amplitude
- Oscillation output voltage V_1 is sampled around the peak and compared to a reference voltage V_{ref}
- Error signal controls the gain of an additional feedback path made of a multiplier and a capacitor εC which simulates positive or negative losses in the oscillator
- If $\hat{V}_1 < V_{ref}$ the losses are made negative and the amplitude increases, whereas when $\hat{V}_1 > V_{ref}$ the losses are made positive and the amplitude decreases
- A steady-state amplitude is achieved when $\hat{V}_1 = V_{ref}$

F. Krummenacher, "A High-Resolution Capacitance-to-Frequency Converter," JSSC, June 1985.

Capacitance-to-Frequency Interface

- It can be shown that the oscillation frequency is given by

$$f_o = \frac{f_c}{\pi} \cdot \sin^{-1}\left(\frac{\alpha}{2}\right) \cong \frac{f_c}{2\pi} \cdot \alpha$$

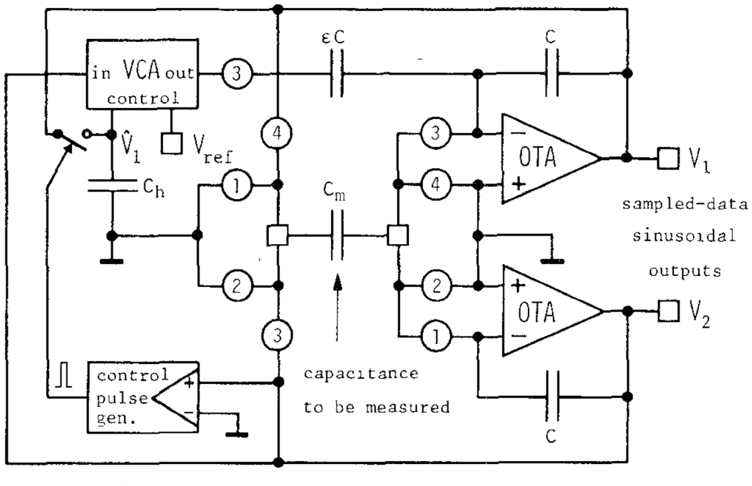
- where f_c is the clock frequency
- For small values of the capacitance ratio α the above relation becomes linear

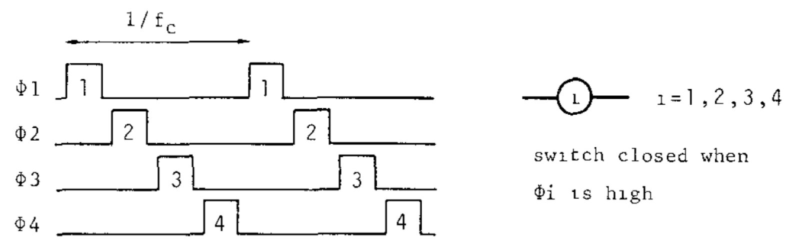
$$f_o \cong \frac{f_c}{2\pi} \cdot \alpha$$

- The frequency is therefore proportional to the capacitor ratio α to be measured
- The capacitor to be measured C_m must assume the function of each of the switched capacitor αC of the oscillator
- This can be accomplished by using an appropriate multiplexing as shown in the following schematic

F. Krummenacher, "A High-Resolution Capacitance-to-Frequency Converter," JSSC, June 1985.

Capacitance-to-Frequency Interface





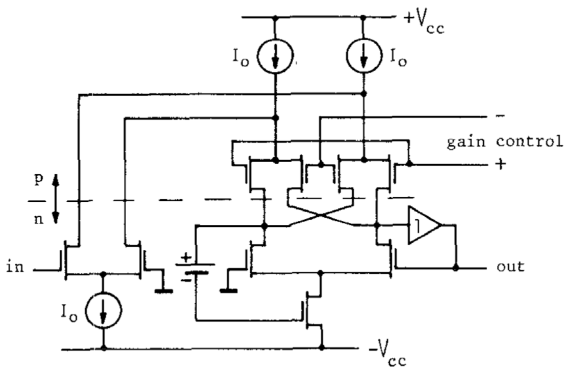


Fig. 5. CMOS voltage-controlled amplifier.

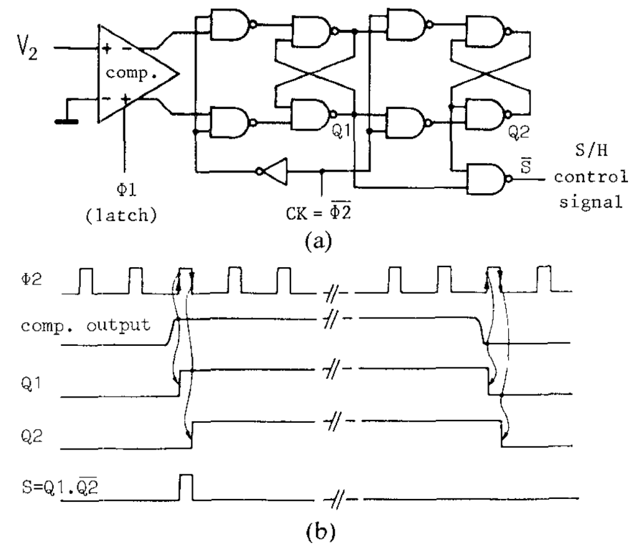
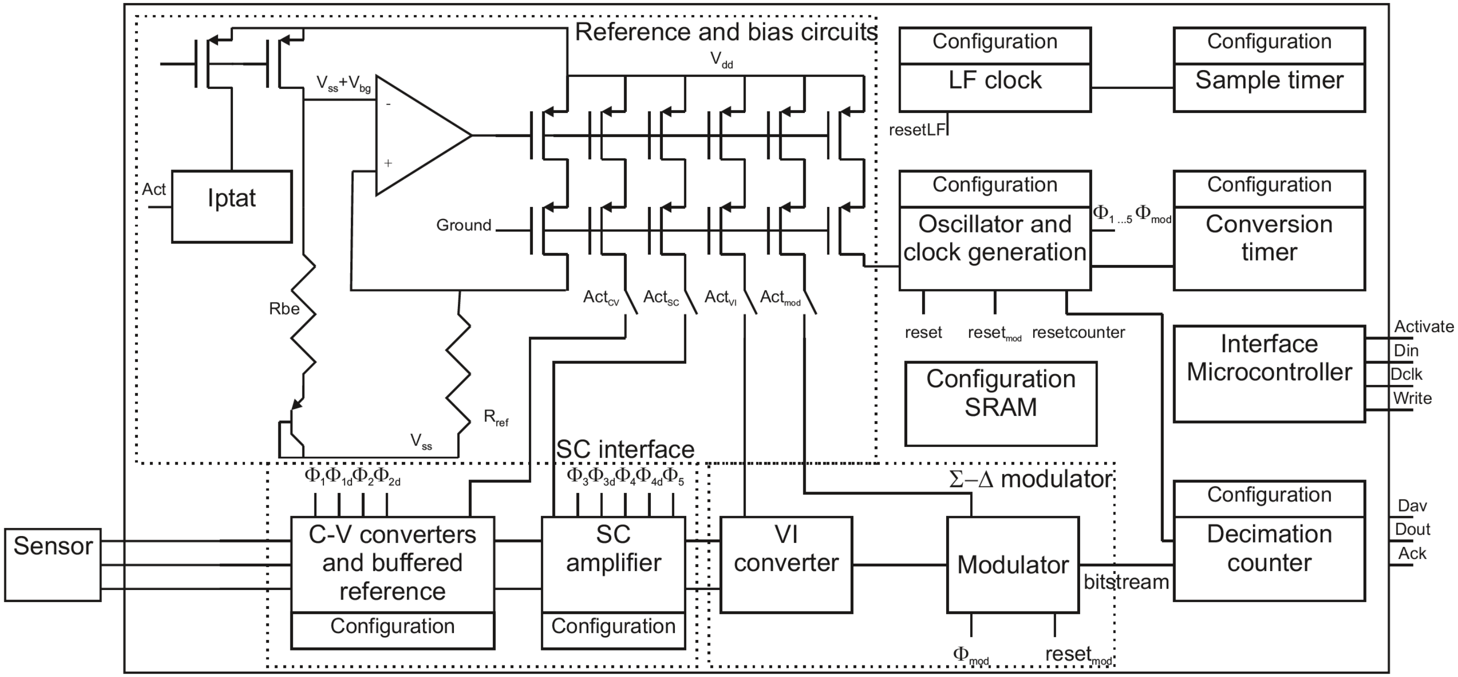


Fig. 6. Control pulse generator for the acquisition and the memorization of $\hat{V}_1$. (a) Circuit schematic. (b) Timing diagram.

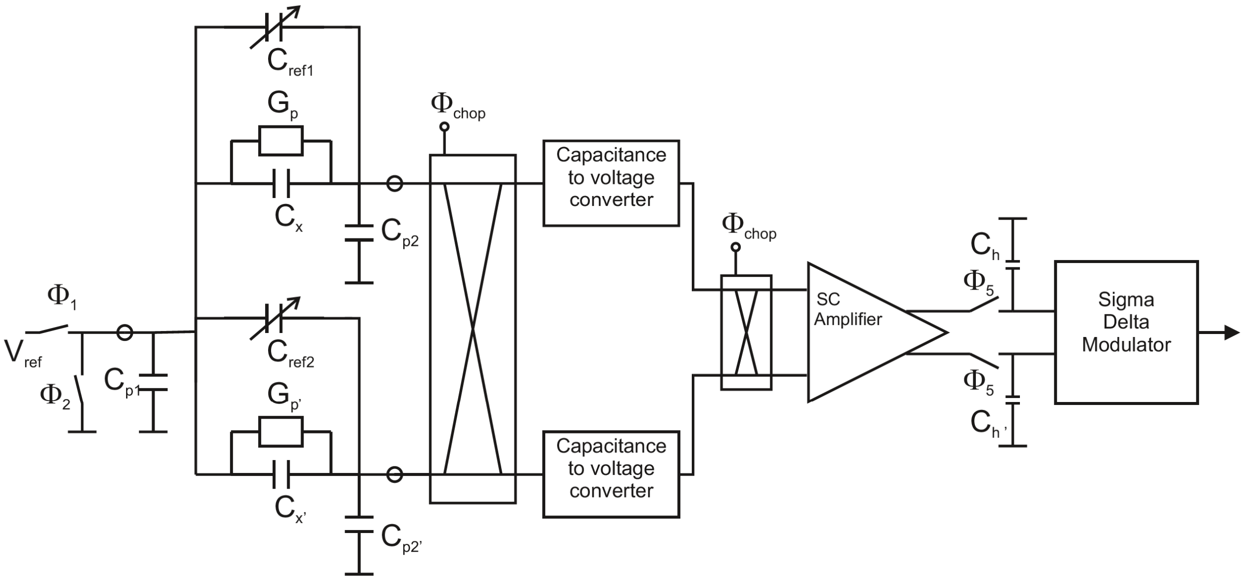
F. Krummenacher, "A High-Resolution Capacitance-to-Frequency Converter," JSSC, June 1985.

Generic Sensor Interface Chip for Capacitive Sensors



W. Bracke, R. Puers and C. van Hoof, *Ultra Low Power Capacitive Sensor Interfaces*, Springer, 2007.

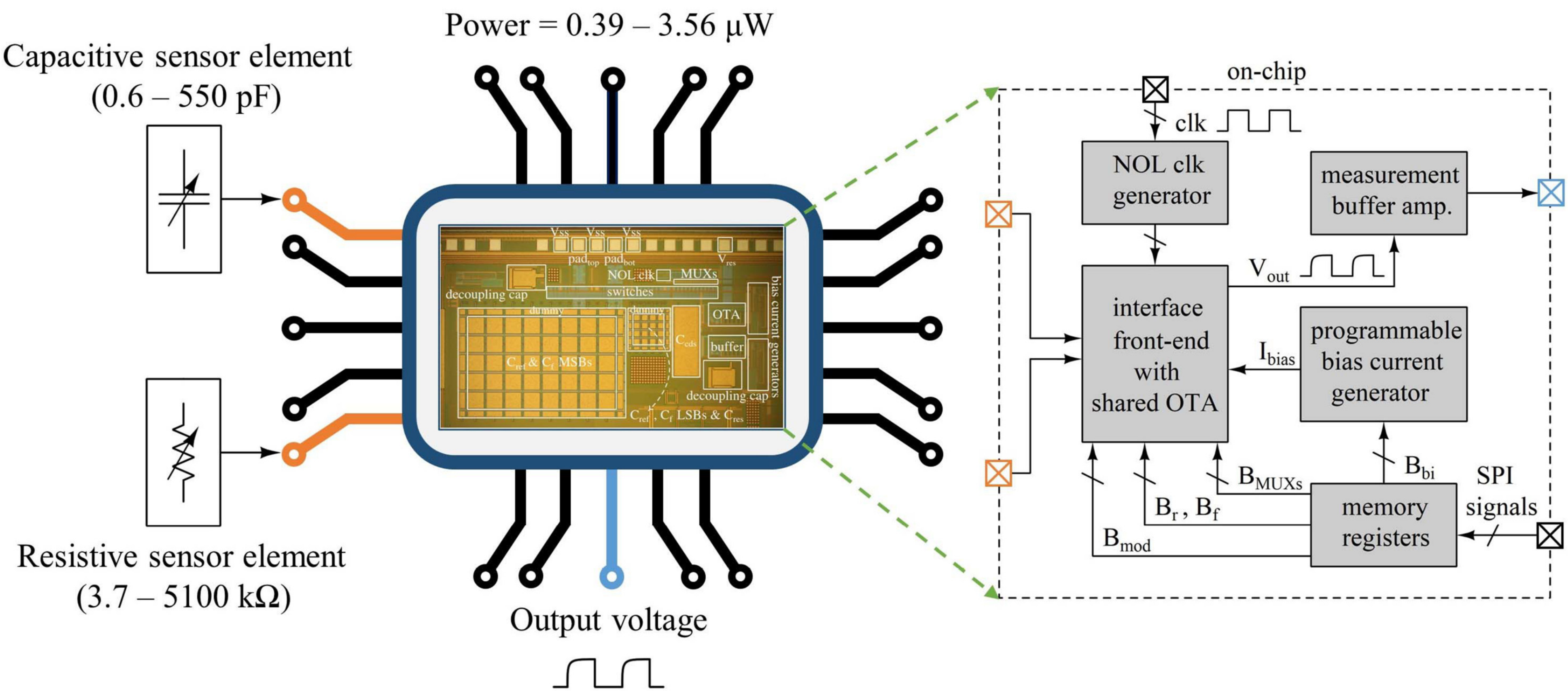
Capacitive Sensor Readout Architecture



- Low frequency for the SC interface and higher frequency for the ΣΔ modulator
- Reducing the clock frequency of the SC interface increases the influence of the parasitic shunt conductance which can cause a serious reliability problem

W. Bracke, R. Puers and C. van Hoof, *Ultra Low Power Capacitive Sensor Interfaces*, Springer, 2007.

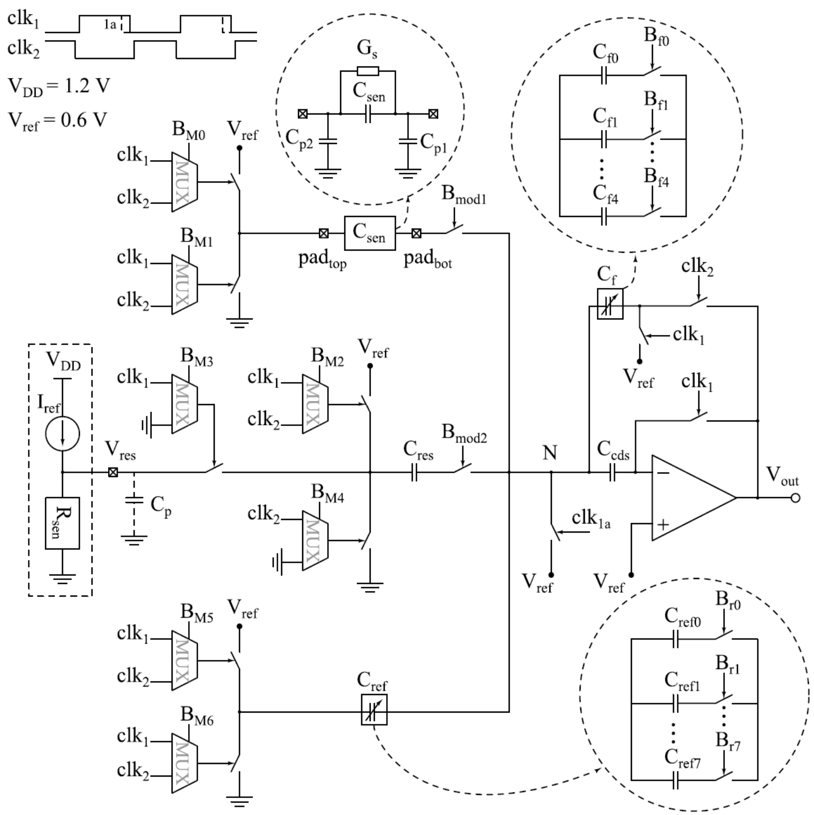
Universal Multi-sensor Interface (UMSI)



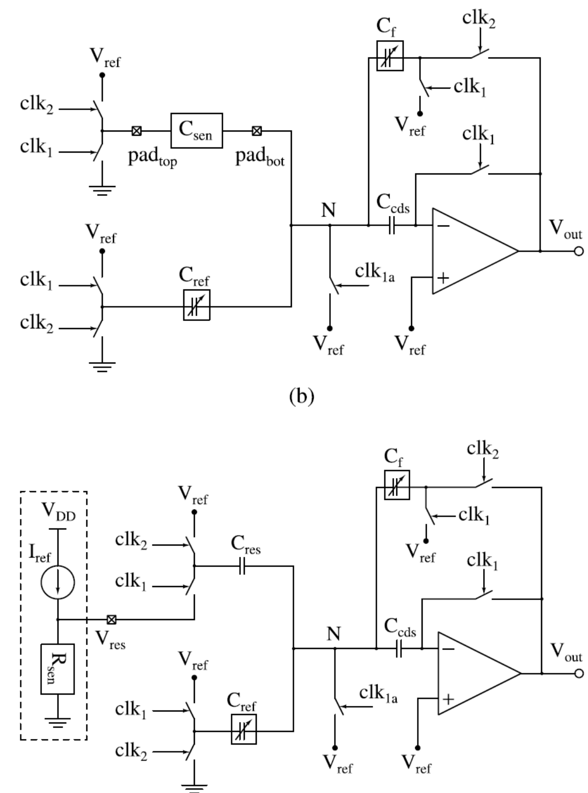
M. M. Moayer, J. Salomaa and K. A. I. Halonen, IEEE Sensors Journal, vol. 20, no. 20, pp. 12262-12273, Oct. 2020.

Universal Multi-sensor Interface (UMSI)

System configured for capacitance measurement



System configured for resistance measurement



📖 M. M. Moayer, J. Salomaa and K. A. I. Halonen, IEEE Sensors Journal, vol. 20, no. 20, pp. 12262-12273, Oct. 2020.