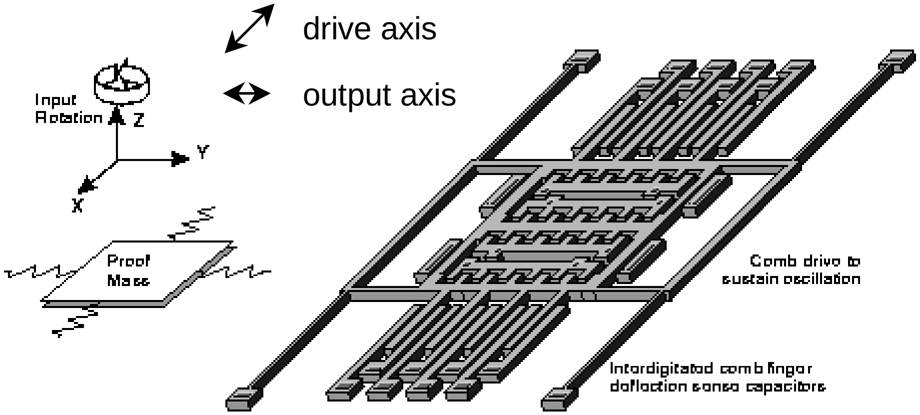
Capacitive Position Sense Circuits

Bernhard E. Boser
Berkeley Sensor & Actuator Center
Dept. of Electrical Engineering and Computer Sciences
University of California, Berkeley

Gyroscope Example



drive axis

output axis

$$x(t) = X_0 \sin\omega_x t$$

$$\ddot{y}_{coriolis} = 2\Omega_z \times \dot{x}(t)$$

$$= 2\Omega_z X_0 \omega_x \cos\omega_x t$$

$$= 4.4\times10^{-3}\left[\frac{m}{s^2}\right]\times\cos\omega_x t$$

$$m\ddot{y} = k_y y$$

$$y = \frac{\ddot{y}}{\omega_y^2}$$

$$= \underline{0.003\text{Å}}\times\cos\omega_x t$$

with Ω_z = 1deg/s, X_0 = 1µm,

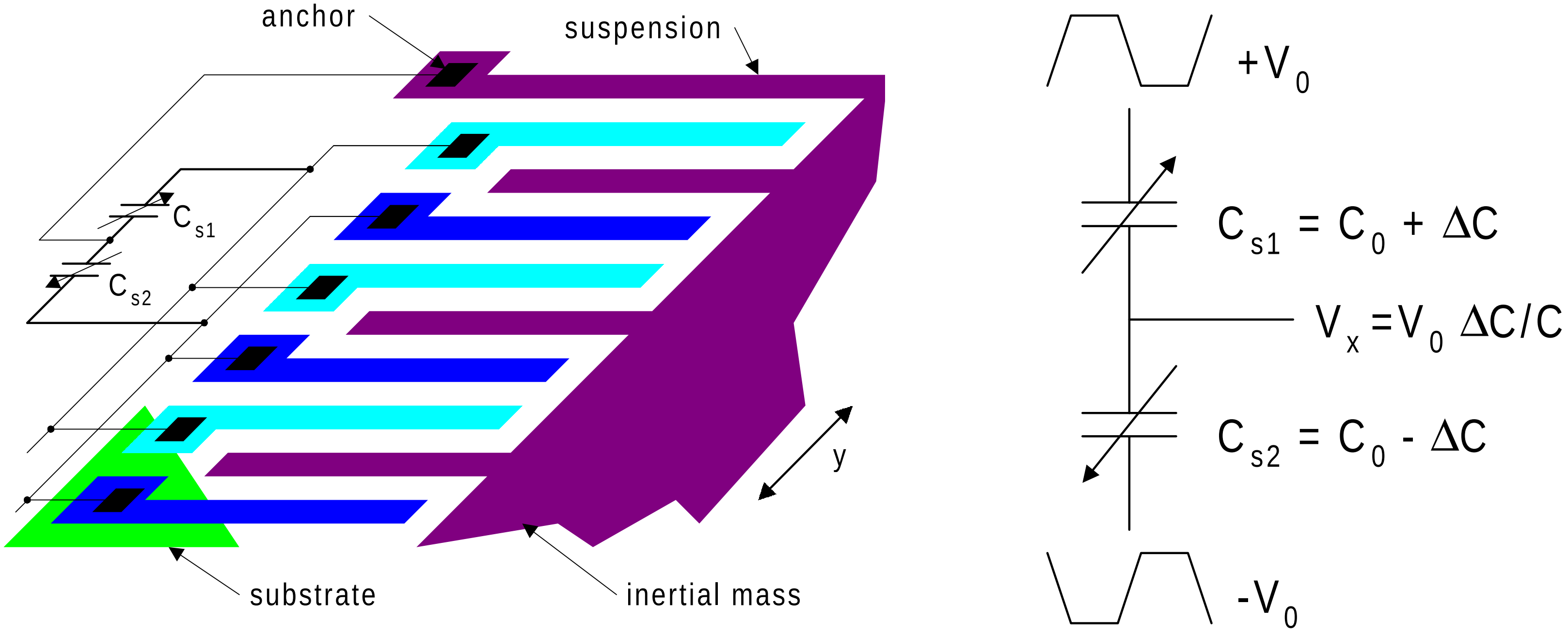
$\omega_x = \omega_y = 2\pi$ x 20krad/s, Q=1

Ref.: W. Clark, “Surface Micromachined Z-Axis Vibratory Rate Gyroscope”, Solid-State Sensor and Actuator Workshop, Hilton Head, pp. 283-287, June 1996.

Outline

☞Capacitive Sensor Interface
- Bootstrapping
- Offset & Drift Reduction
- Charge Sensing
- Correlated Double Sampling
- Noise

Capacitive Interface

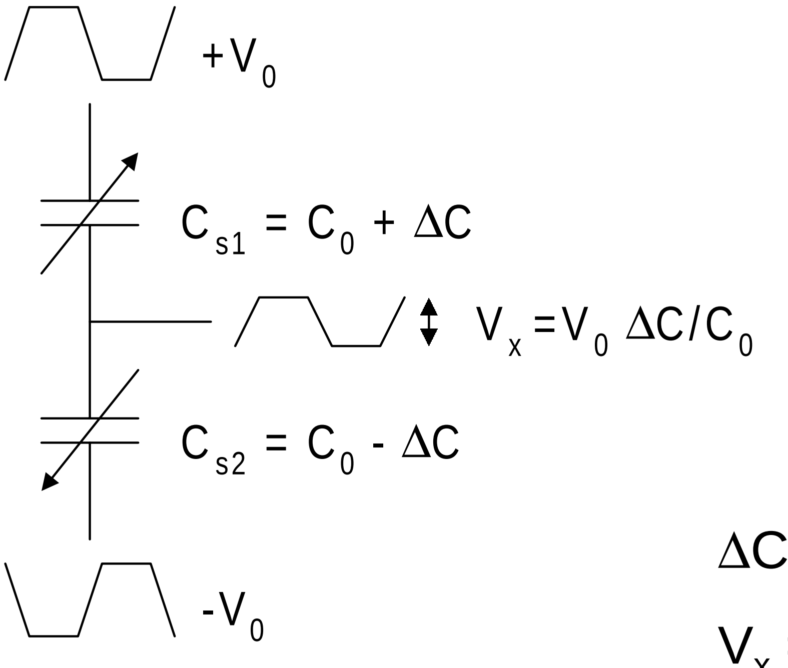


$C_{s1} = C_0 + \Delta C$

$V_x = V_0 \, \Delta C / C$

$C_{s2} = C_0 - \Delta C$

Output Signal



$+V_0$

$C_{s1} = C_0 + \Delta C$

$V_x = V_0 \, \Delta C / C_0$

$C_{s2} = C_0 - \Delta C$

$-V_0$

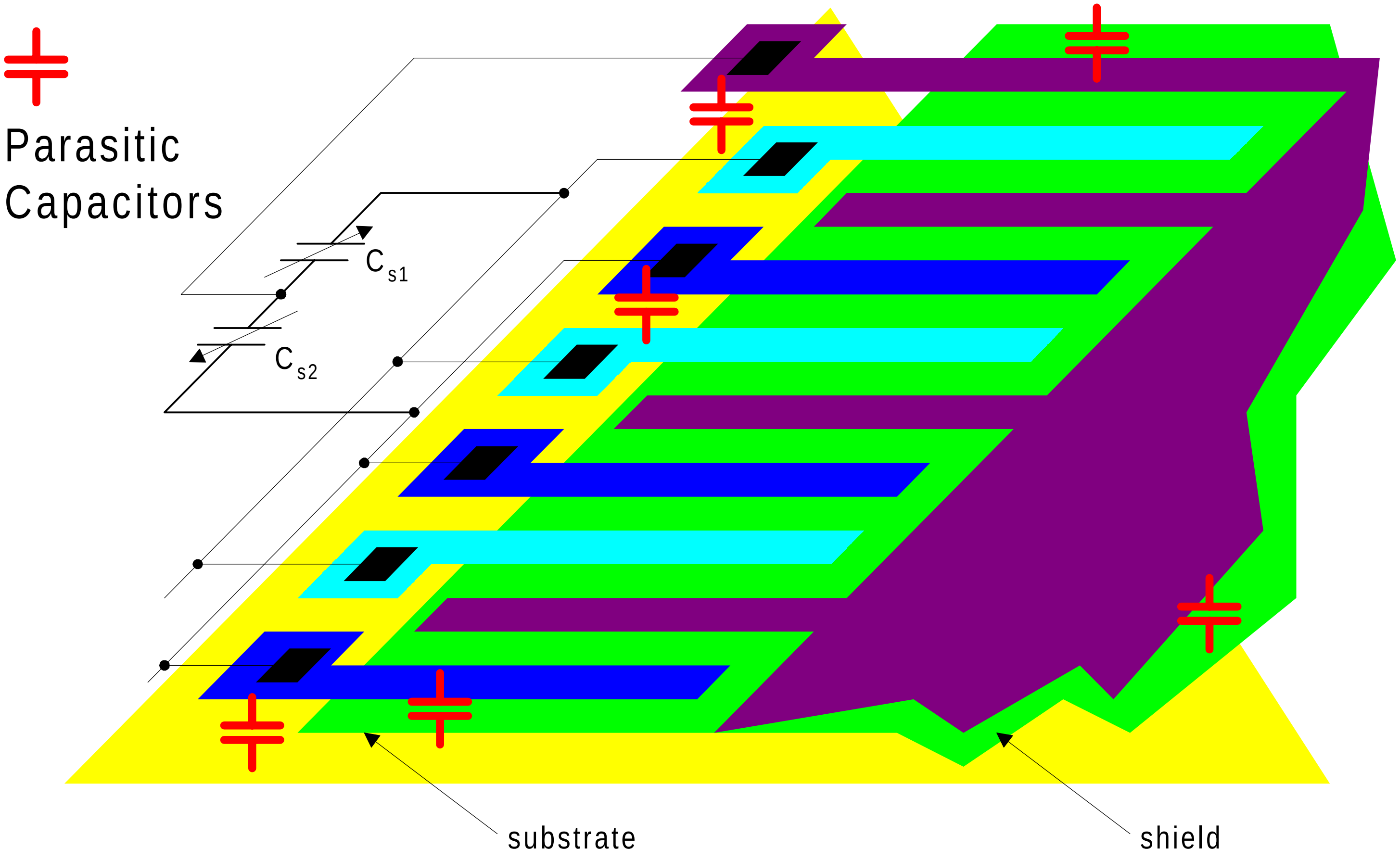
for C_0 = 100 fF
dC/dy = 10 aF/Å
y = 0.003 Å
V_0 = 5 V

⇒

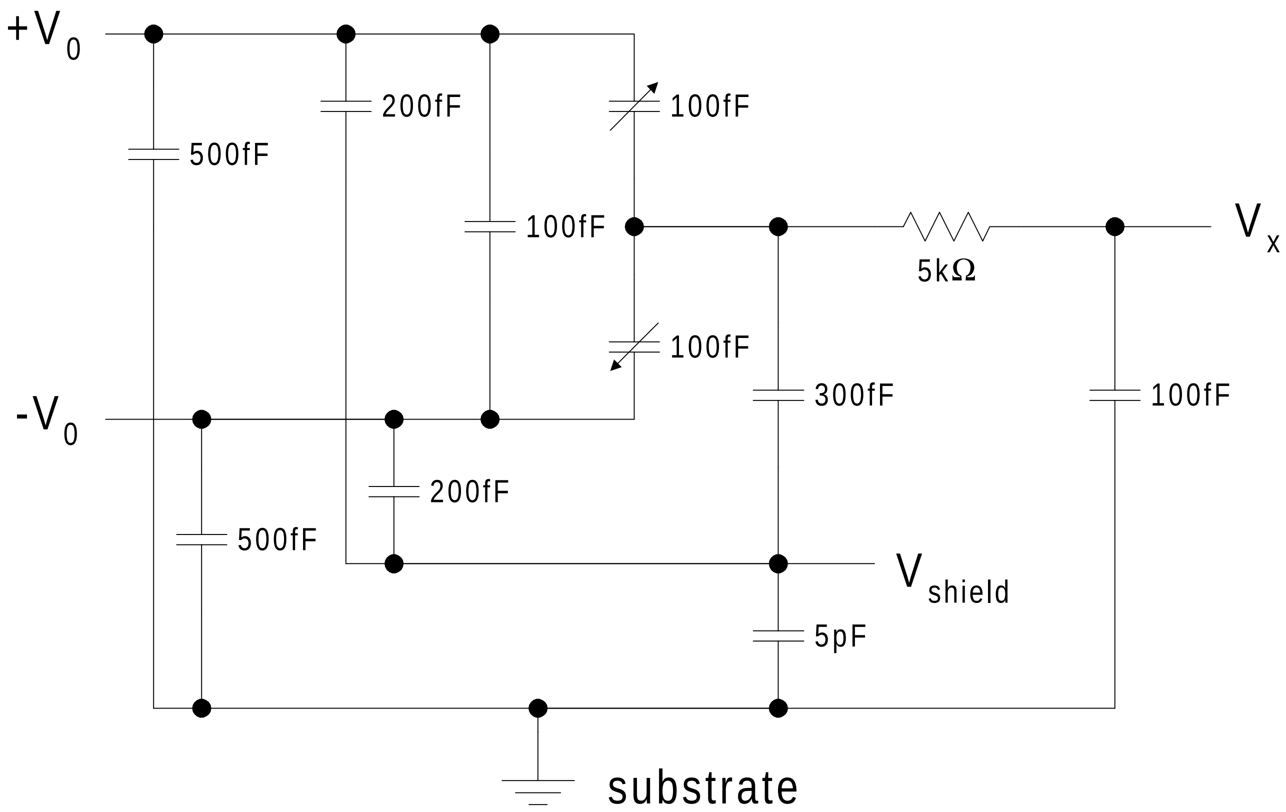
ΔC = 10aF/Å x 0.003Å = 0.03 aF

V_x = 5V x 0.03aF / 100fF = <u>0.3 µV</u>

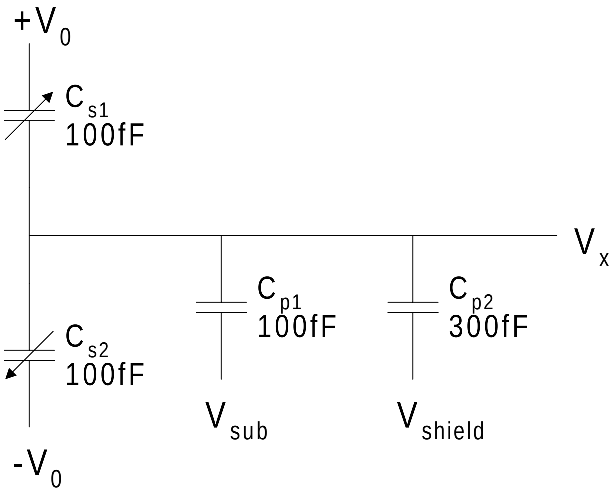
Electrical Interface



Electrical Interface Model



Impact of Parasitics



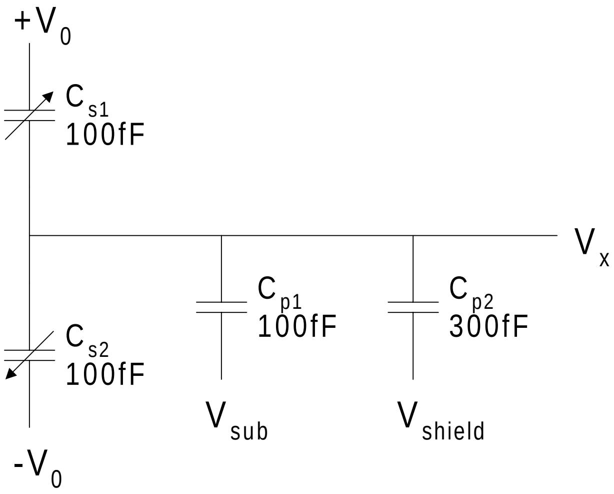
$$V_x = \frac{\overbrace{2\Delta C V_0}^{signal} + \overbrace{C_{p1} V_{sub} + C_{p2} V_{shield}}^{error}}{C_T}$$

$$C_T = 2C_0 + C_{p1} + C_{p2}$$

Impact of Parasitics (cont.)

1) signal attenuation by

$$\frac{1}{1+\frac{C_{p1}+C_{p2}}{2C_0}} = 1/3$$



2) couple V_{shield} & V_{sub} to V_x

e.g. V_{sub} = 1mV

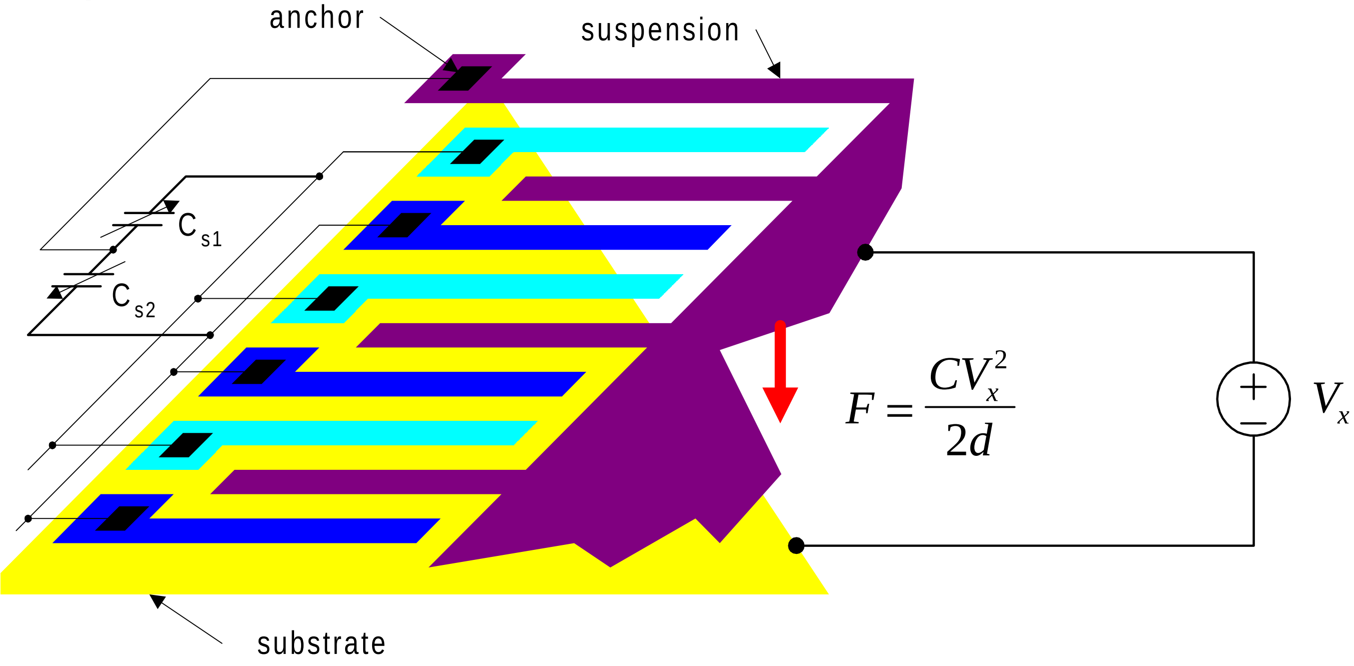
V_x = 250µV >> 0.3µV

<u>keep substrate quiet!</u>

3) parasitic electrostatic forces
check voltage across <u>all</u> capacitors!

Parasitic Electrostatic Force

Example:



$$F = \frac{CV_x^2}{2d}$$

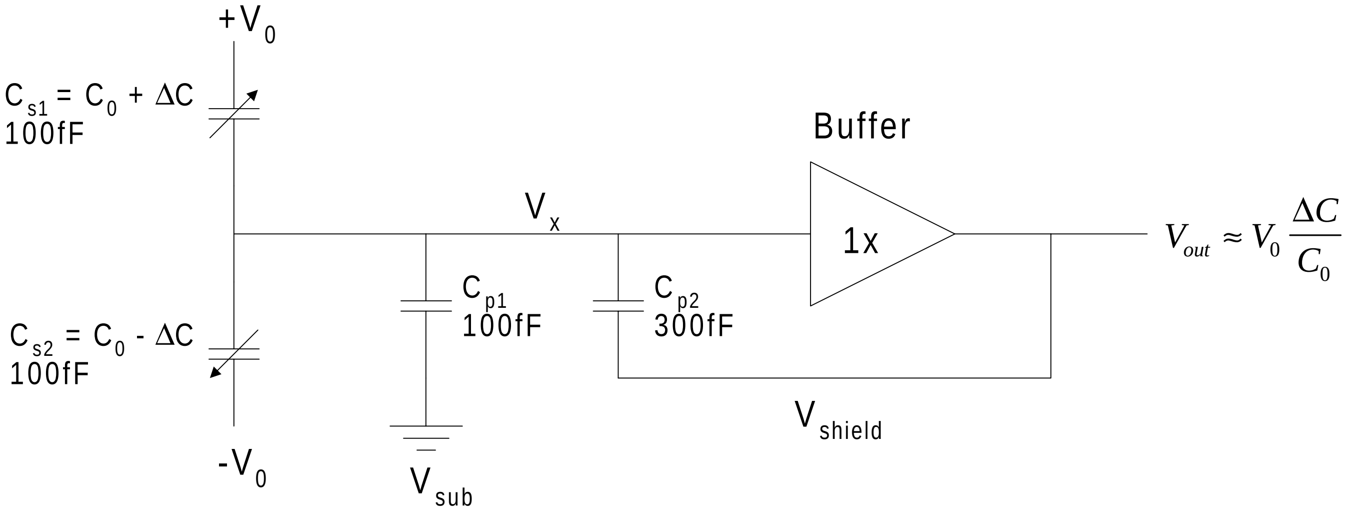
Outline

- Capacitive Sensor Interface

☞Bootstrapping

- Offset & Drift Reduction
- Charge Sensing
- Correlated Double Sampling
- Noise

Bootstrapping



- $V_{shield} = V_{out} = V_x$ ---> voltage across C_{p2} is always zero

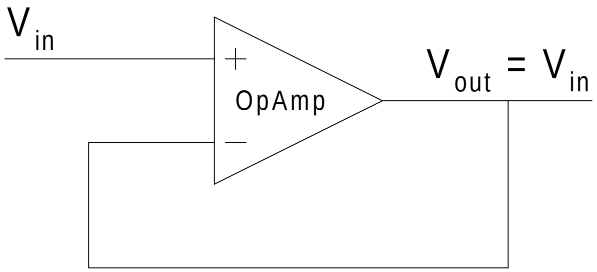
$$V_x = \frac{2\Delta C V_0 + C_{p1} V_{sub}}{2C_0 + C_{p1}}$$

C_{p2} effectively removed

- *Caveats:* instability if buffer gain > 1, increased noise

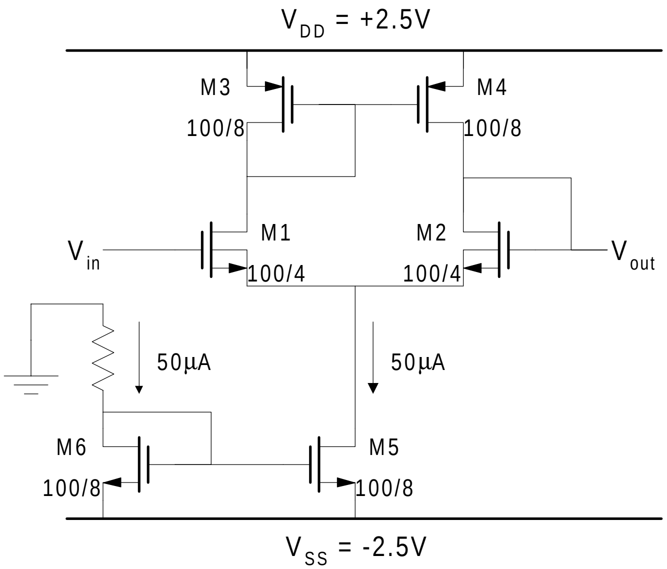
Unity Gain Buffer Circuits

Breadboard Setup:



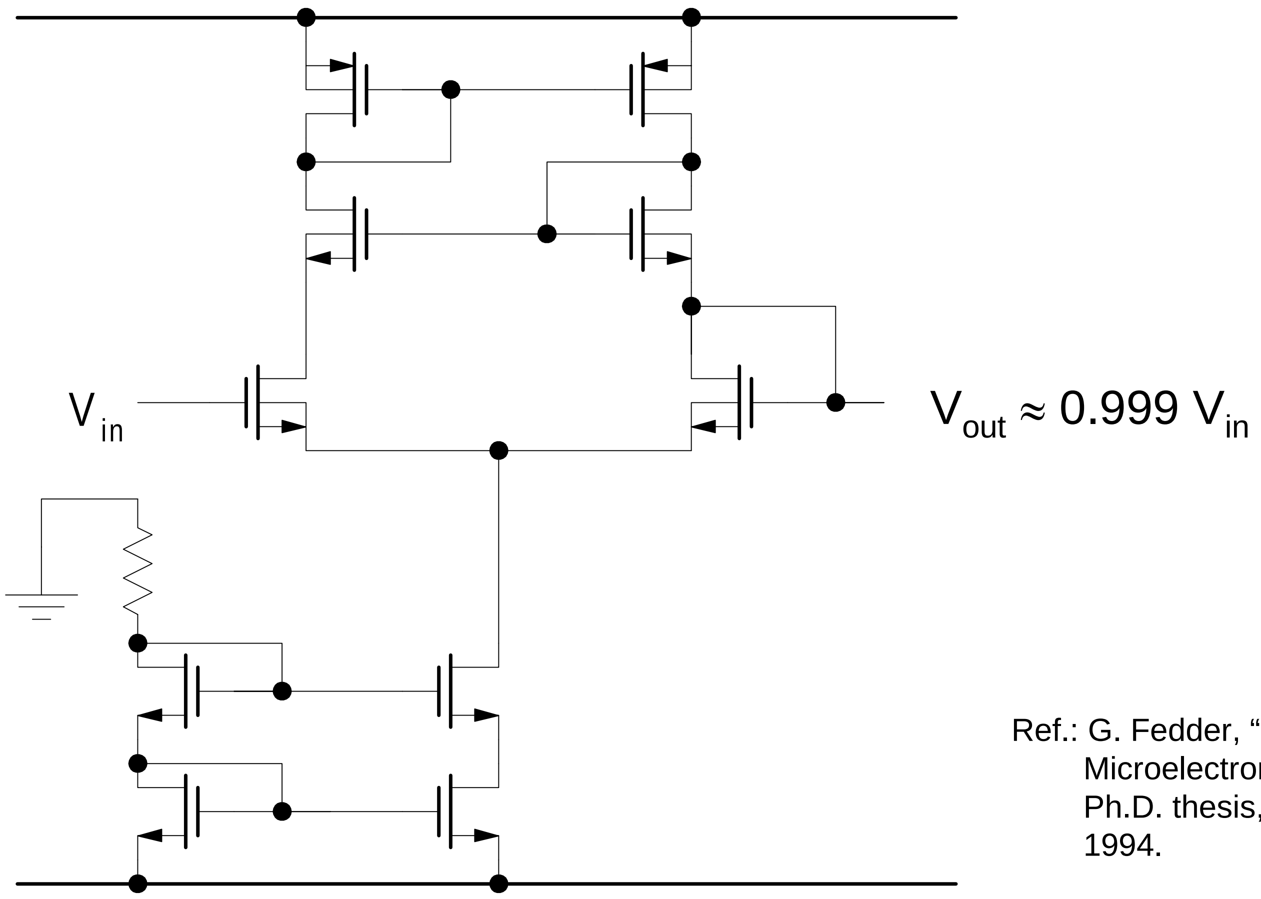
JFET inputs (negligible input current)
e.g. LF356 or LH0022

Integrated Circuit:



$V_{out} \approx 0.98\ V_{in}$

Improved Buffer



Ref.: G. Fedder, “Simulation of Microelectromechanical Systems”, Ph.D. thesis, UC Berkeley, EECS, 1994.

Outline

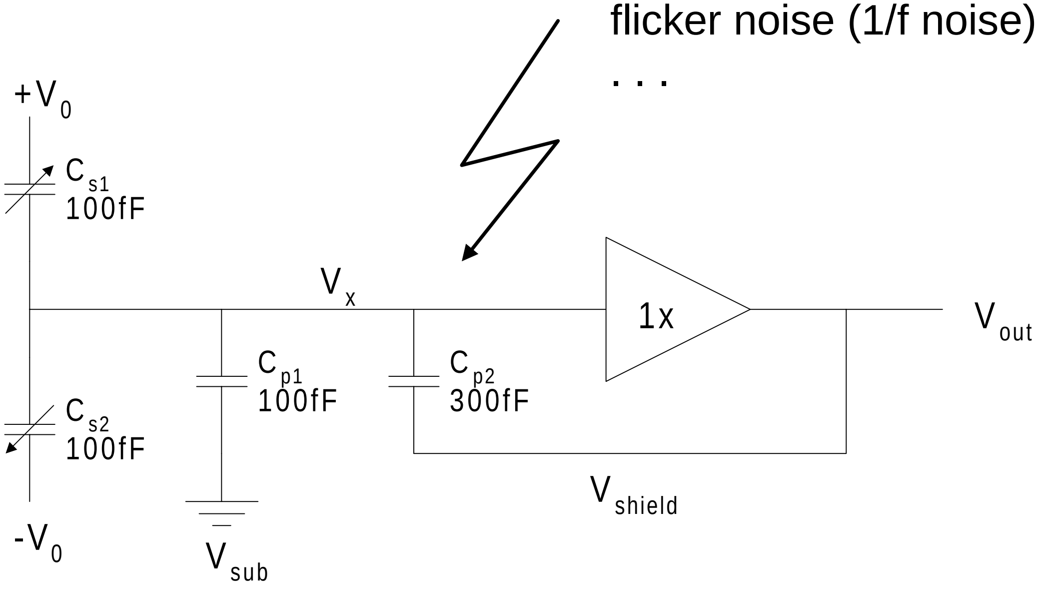
- Capacitive Sensor Interface
- Bootstrapping

☞Offset & Drift Reduction

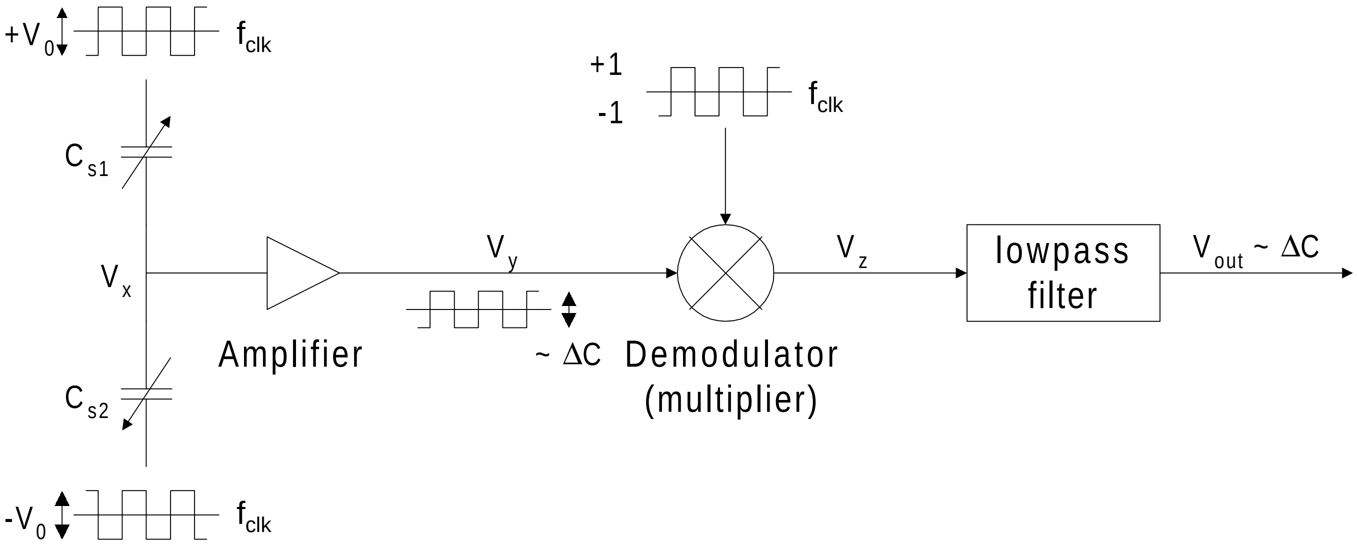
- Charge Sensing
- Correlated Double Sampling
- Noise

Offset & Drift Reduction

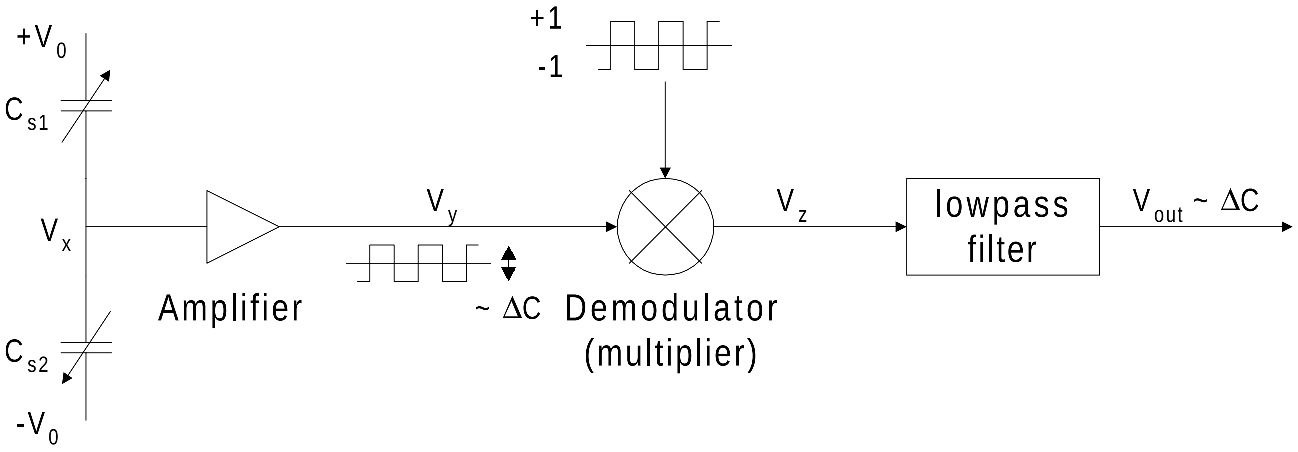
EMI (Electromagnetic Interference)
substrate coupling
carrier generation
amplifier offset & drift
flicker noise (1/f noise)
. . .

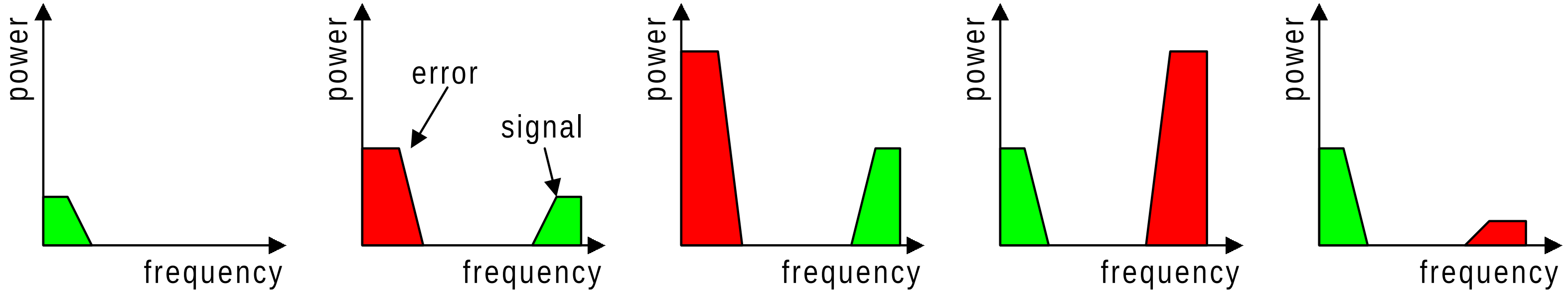


Modulation & Demodulation



Frequency Domain Interpretation





ΔC ~ Δx ~ input V_x V_y V_z V_{out}

Modulation (cont.)

- like AM radio

  different name: “chopper stabilization”
- modulation frequency

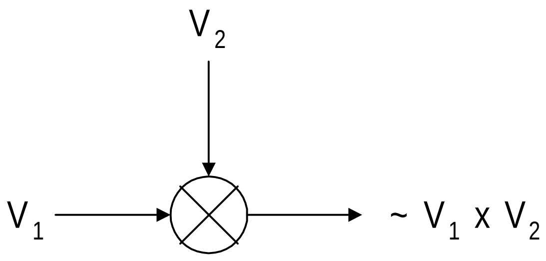
  f_{clk} = 100kHz … 10MHz (typical)
- wave-form:

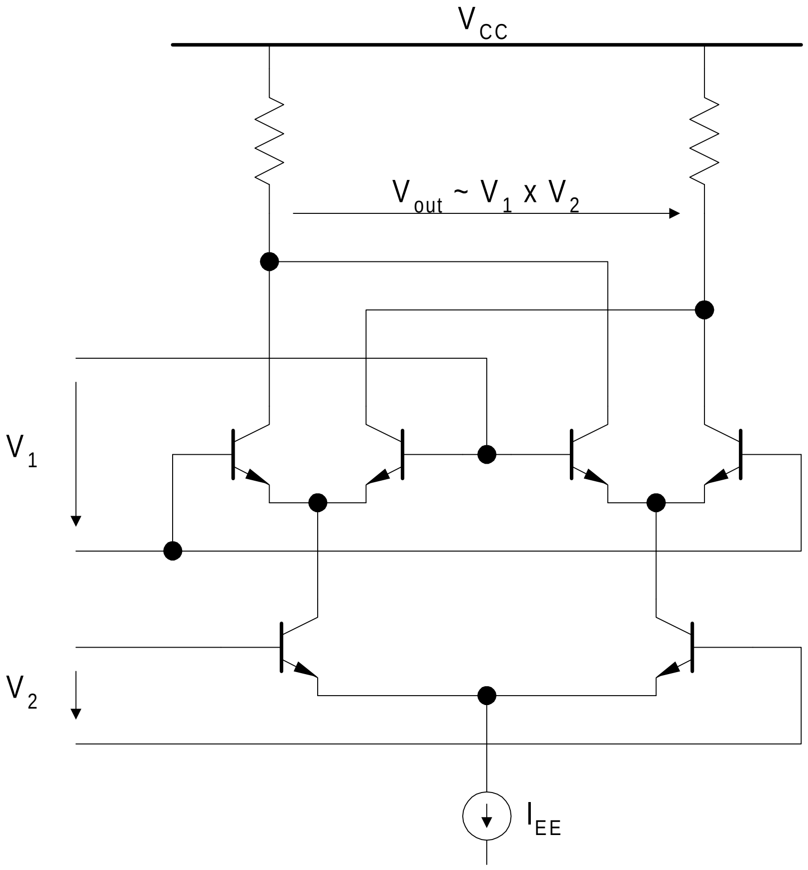
  sinusoidal or

  square-wave (avoid distortion!)

Demodulator Circuit

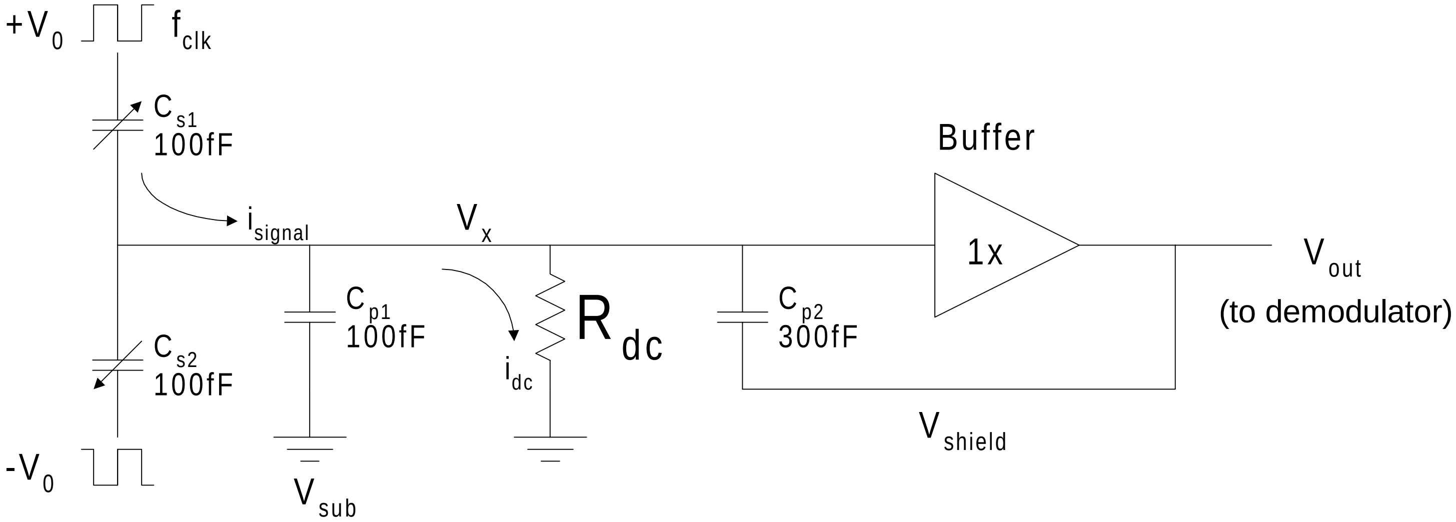
Gilbert Multiplier:





(MOS circuit identical)

DC Potential



Dimensioning R_{dc}:

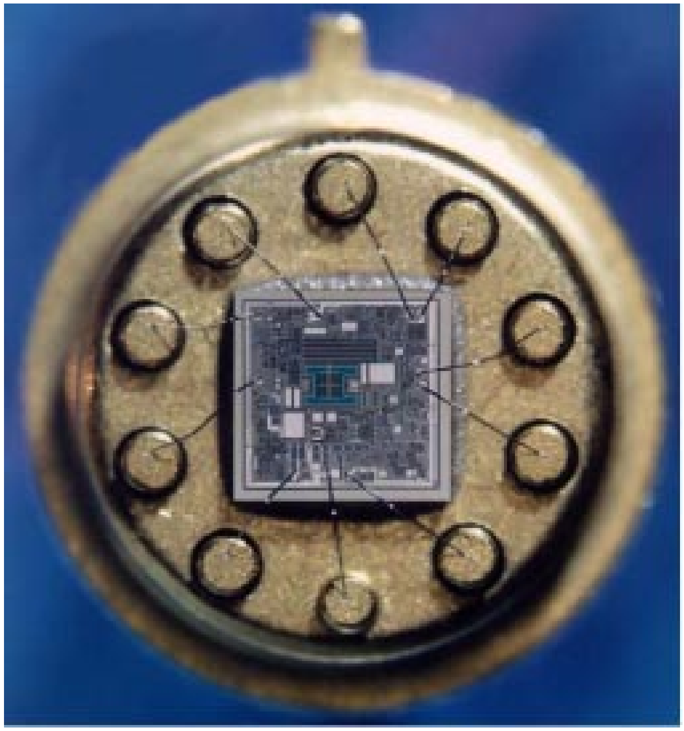
$$\left|i_{signal}\right| >> \left|i_{dc}\right| \longrightarrow 2\pi f_{clk} C_{s1} V_x >> \frac{V_x}{R_{dc}} \longrightarrow R_{dc} >> \frac{1}{2\pi f_{clk} C_{s1}} = \underline{1.6 M\Omega}$$

for f_{clk} = 1MHz and C_{s1} = 100fF

Realizing R_{dc}

- physical resistor
  - low-doped polysilicon
  - thin-film resistor
  - avoid large parasitic capacitance (e.g. diffusion)
- back-to-back diodes
  - nonlinear distortion
- MOSFET in weak inversion
  - poor control
- Switched Capacitors

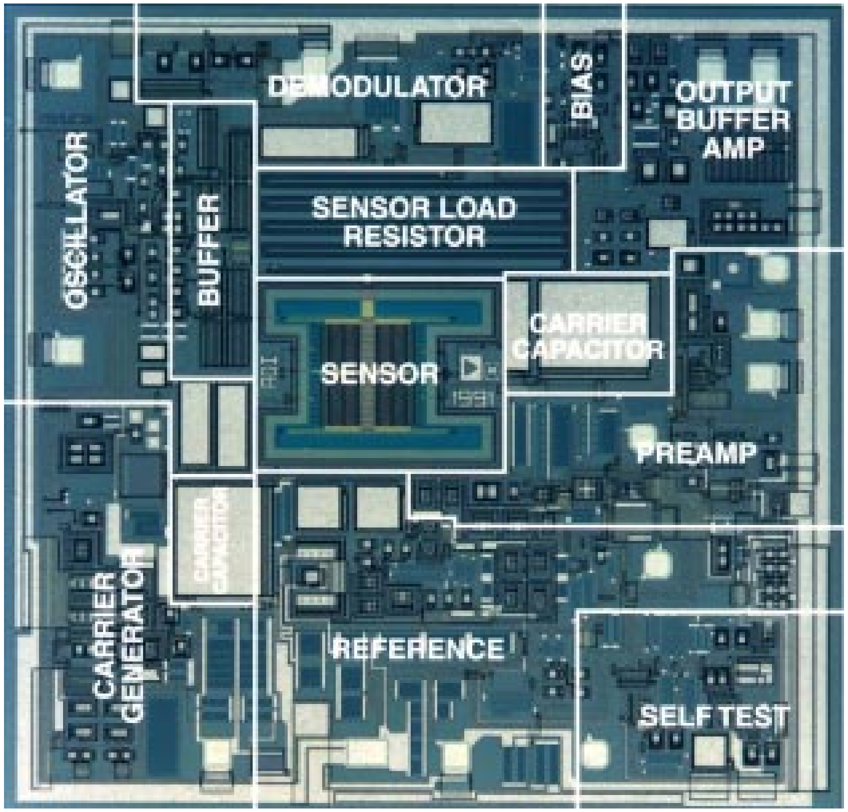
Example: ADXL-50



full-scale input	50 g
noise floor	6.6 mG/rt-Hz
linearity	0.2 % of F.S.
bandwidth	1 kHz
power dissipation	50 mW

Ref.: Analog Devices ADXL-50

ADXL 50 Layout



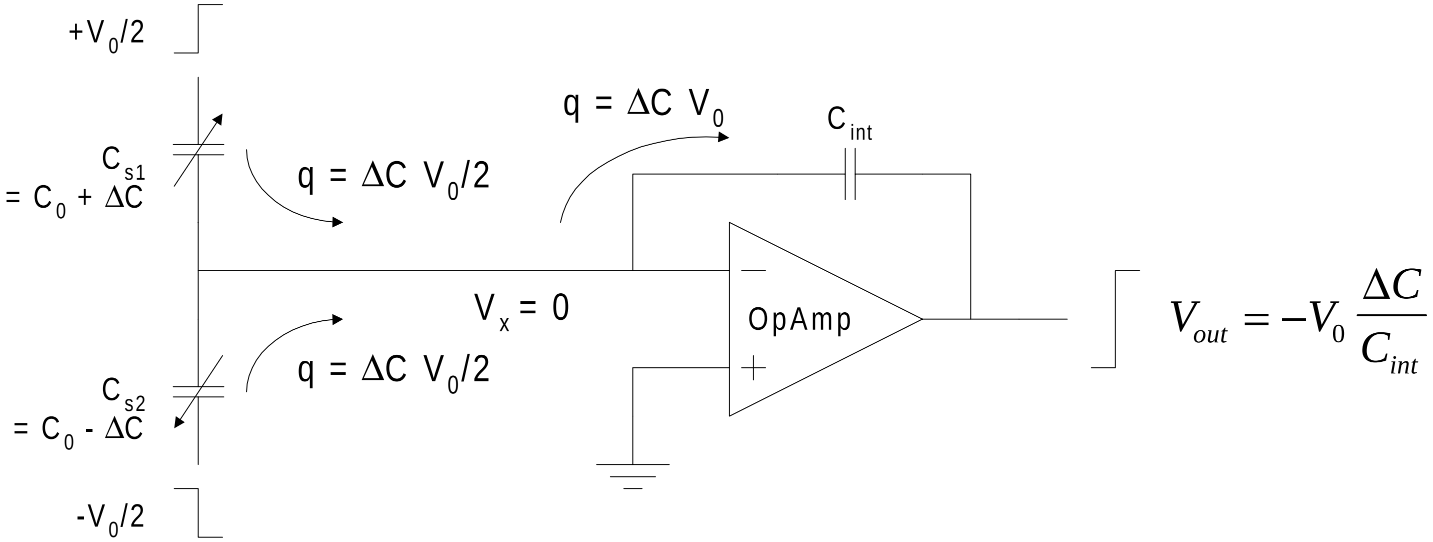
Outline

- Capacitive Sensor Interface
- Bootstrapping
- Offset & Drift Reduction

☞Charge Sensing

- Correlated Double Sampling
- Noise

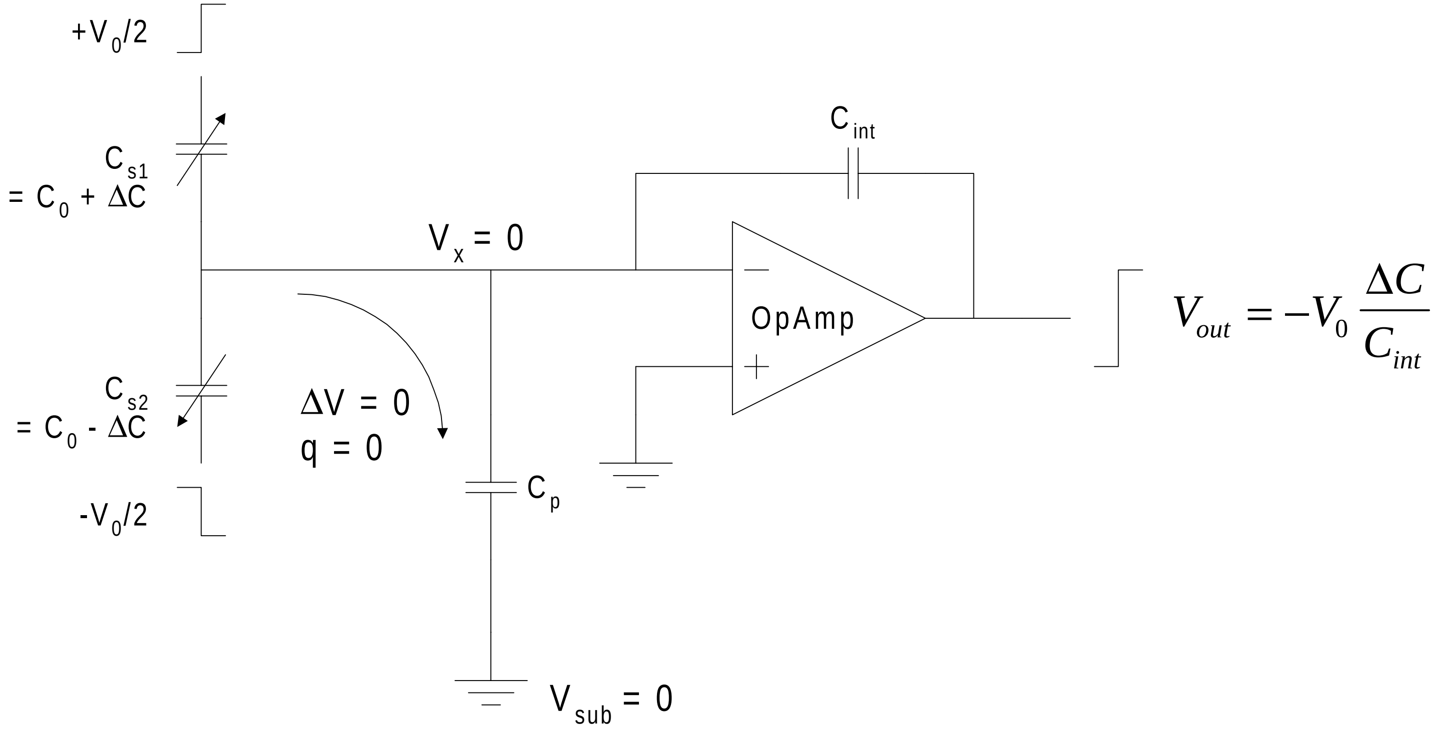
Charge Sensing Concept



$$V_{out} = -V_0 \frac{\Delta C}{C_{int}}$$

V_x is virtual ground

Charge Sensing: Parasitics

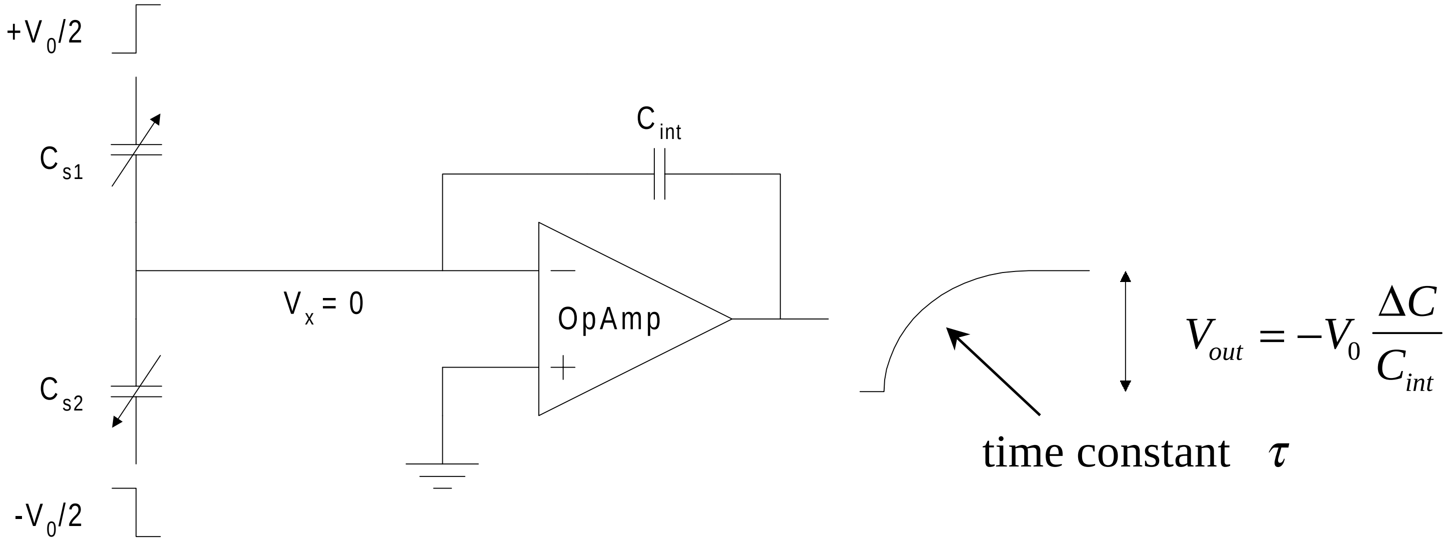


$$V_{out} = -V_0 \frac{\Delta C}{C_{int}}$$

Charge Sensing: Parasitics

- virtual ground shorts C_p
- no need to drive shield
- *CAVEATS:*
  - V_{sub} still couples to V_{out}
  - avoid undesired electrostatic forces
  - C_p increases electronic noise

Choosing C_{int}



$$V_{out} = -V_0 \frac{\Delta C}{C_{int}}$$

time constant τ

Choosing C_{int} (cont.)

- Gain:

$$V_{out} \propto \frac{1}{C_{int}} \qquad \rightarrow$$

$$C_{int} \rightarrow 0 \quad \Leftrightarrow \quad V_{out} \rightarrow \infty$$

- Time-constant (settling time):

$$\tau \propto \frac{1}{C_{int}} \qquad \rightarrow$$

$$C_{int} \rightarrow 0 \quad \Leftrightarrow \quad \tau \rightarrow \infty$$

- Gain-Speed Tradeoff:

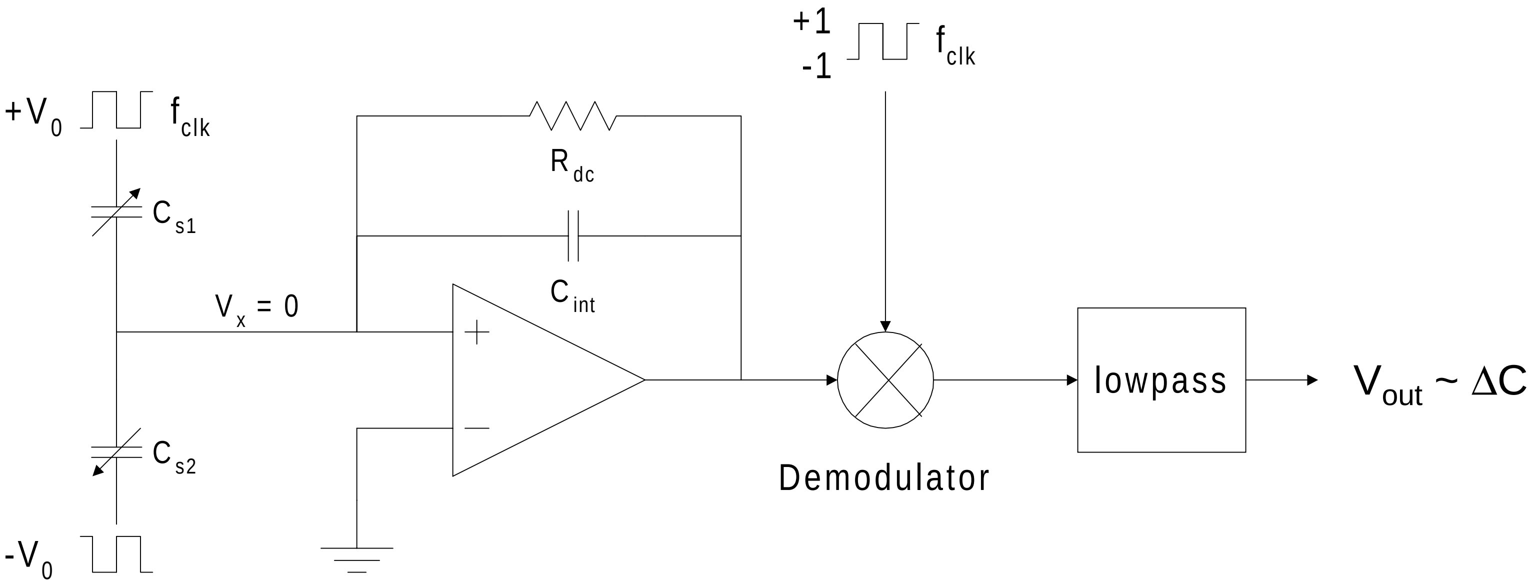
$$typical: \quad C_{int} \approx 2C_0$$

Offset & Drift Reduction

- Chopper Stabilization

- Correlated Double Sampling (CDS)
  - no low-pass filter (increased bandwidth)
  - no need for leak resistor (R_{dc})
  - disadvantage: needs clocks

Chopper Stabilization



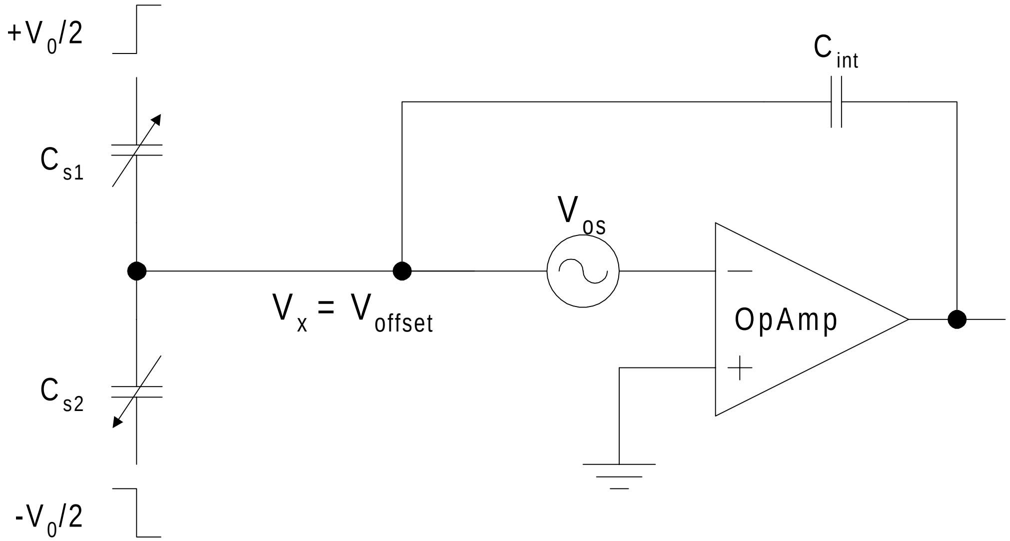
Outline

- Capacitive Sensor Interface
- Bootstrapping
- Offset & Drift Reduction
- Charge Sensing

☞Correlated Double Sampling

- Noise

CDS Problem



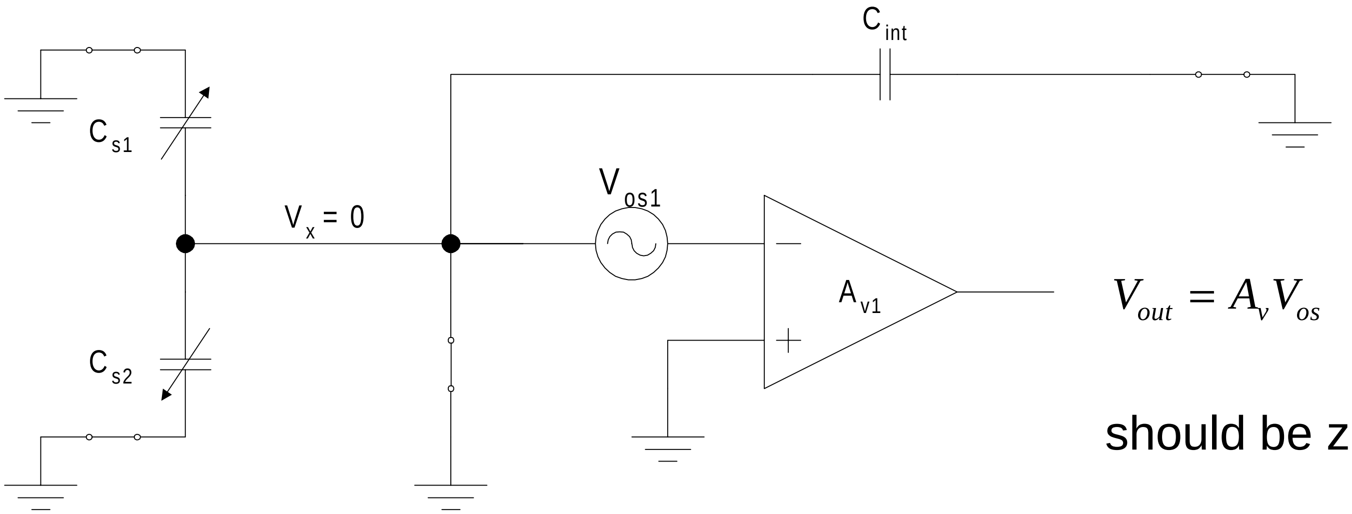
$$V_{out} = V_0 \frac{\Delta C}{C_{int}} + \underbrace{V_{os}\left(1+\frac{C_{s1}+C_{s2}}{C_{int}}\right)}_{error}$$

CDS Concept

Two Phase Operation:

1) measure and cancel offset
2) measure signal

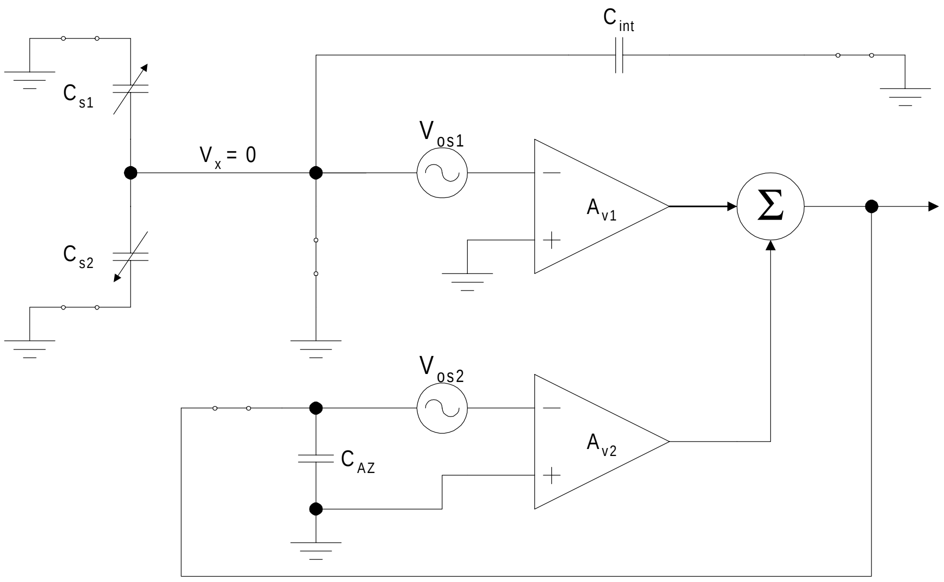
CDS Phase 1a: Measure Offset



$$V_{out} = A_v V_{os}$$

should be zero

CDS Phase 1b: Cancel Offset



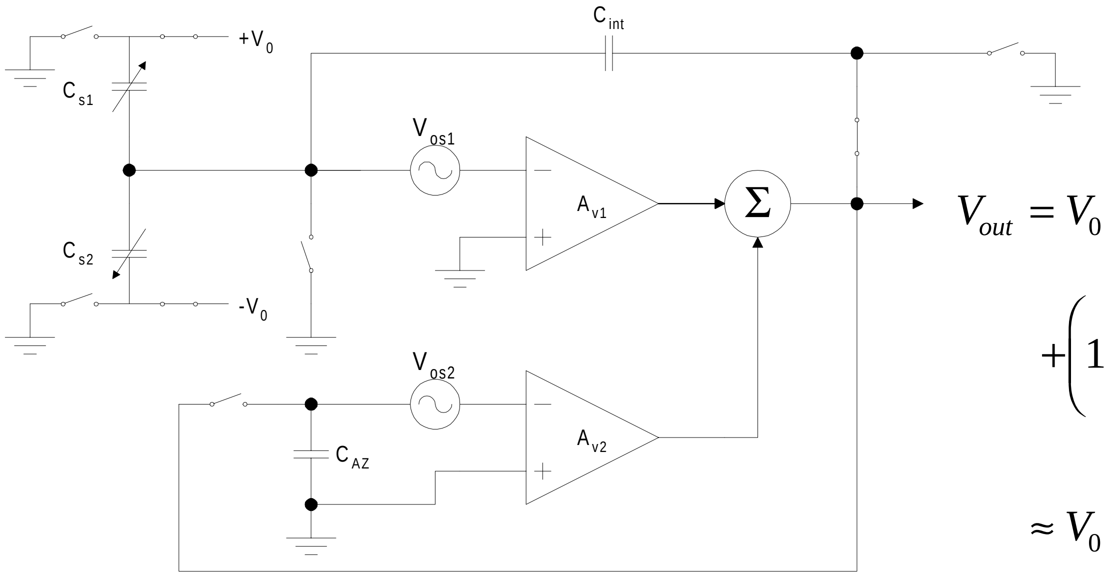
$$V_{out,off} = V_{os1}\frac{A_{v1}}{\underbrace{A_{v2}+1}_{<<A_{v1}}} + V_{os2}\frac{A_{v2}}{\underbrace{A_{v2}+1}_{\approx 1}}$$

input referred amplifier offset:

$$V_{in,off} = \frac{V_{out,off}}{A_{v1}}$$

$$\approx \frac{V_{os1}}{A_{v2}} + \frac{V_{os2}}{A_{v1}} << V_{os1}$$

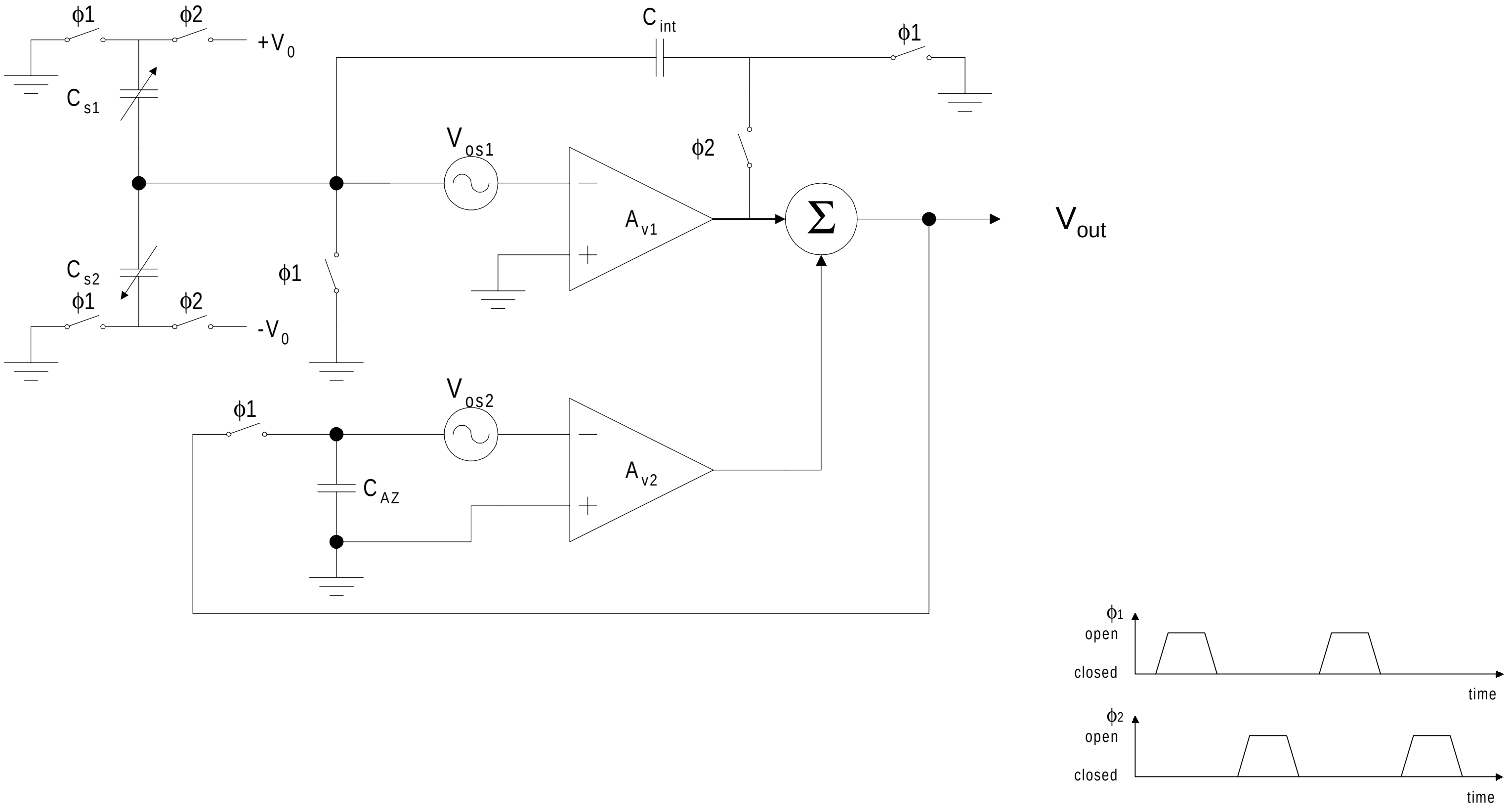
CDS Phase 2: Measure Signal



$$V_{out} = V_0 \frac{\Delta C}{C_{int}} + \left(1 + \frac{C_{s1} + C_{s2}}{C_{int}}\right) \underbrace{\frac{V_{in,off}}{A_{v1}}}_{\approx 0}$$

$$\approx \underbrace{V_0 \frac{\Delta C}{C_{int}}}_{\text{signal}} + \underbrace{\left(1 + \frac{C_{s1} + C_{s2}}{C_{int}}\right)\left(\frac{V_{os1}}{A_{v2}} + \frac{V_{os2}}{A_{v1}}\right)}_{\text{offset}}$$

Complete CDS Circuit



CDS References

- General Principle

  M. Degrauwe et al, “A Micropower CMOS-Instrumentation Amplifier”, IEEE J. Solid-State Circuits, pp. 805-807, June 1985.

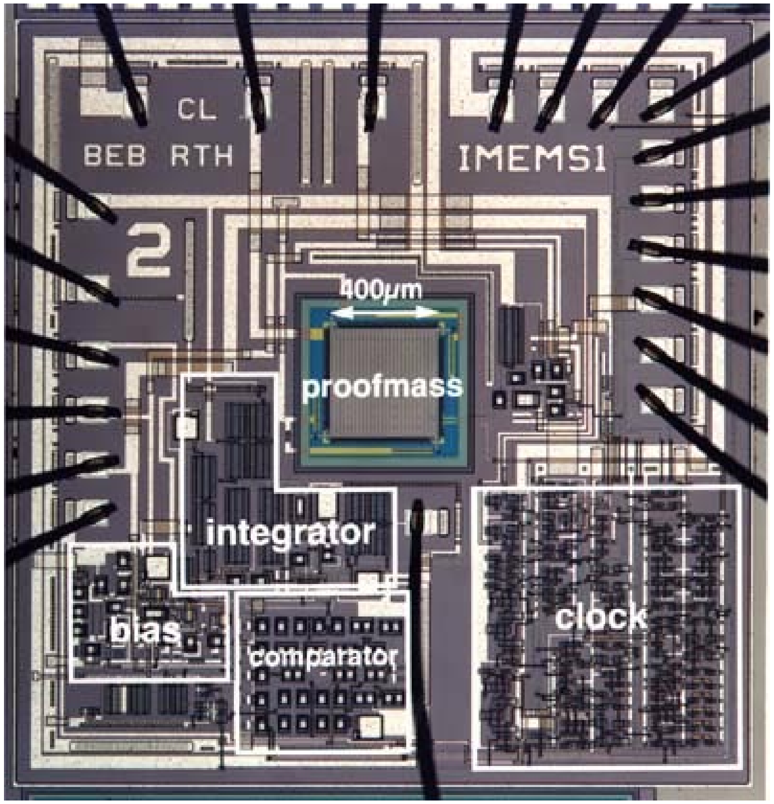
- Capacitive Sensor Interfaces with CDS

  T. Smith et al, “A 15b Electromechanical Sigma-Delta Converter for Acceleration Measurements”, in Digest ISSCC 94, pp. 160-161, February 1994.

  C. Lu et al, “A Monolithic Surface Micromachined Accelerometer with Digital Output”, IEEE J. Solid-State Circuits, pp. 1367-1373, December 1995.

  M. Lemkin et al, “A Micromachined Fully Differential Lateral Accelerometer”, in Digest CICC 96, pp. 315-318, May 1996.

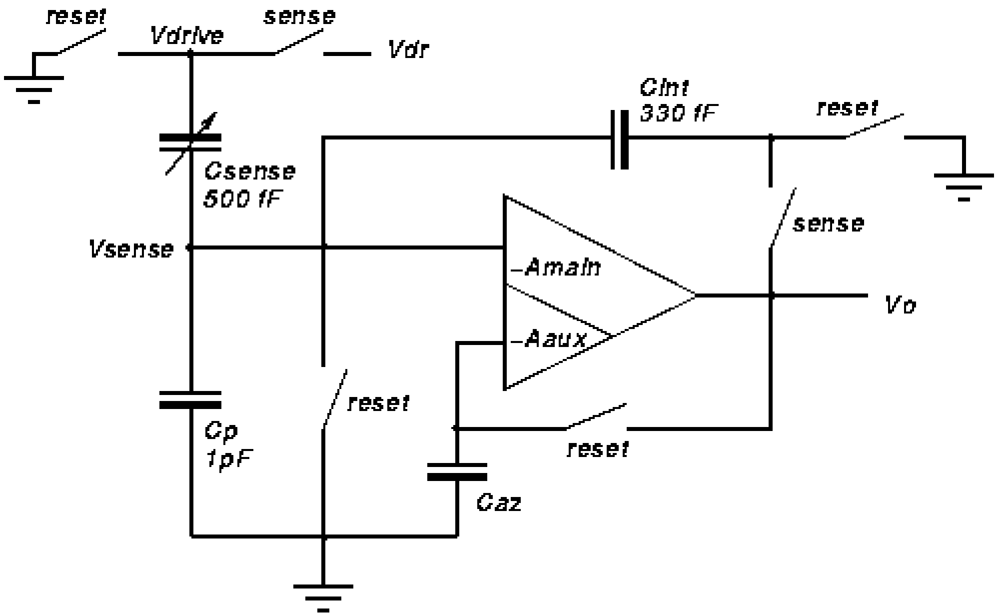
Example 1: CDS Z-Axis Accelerometer



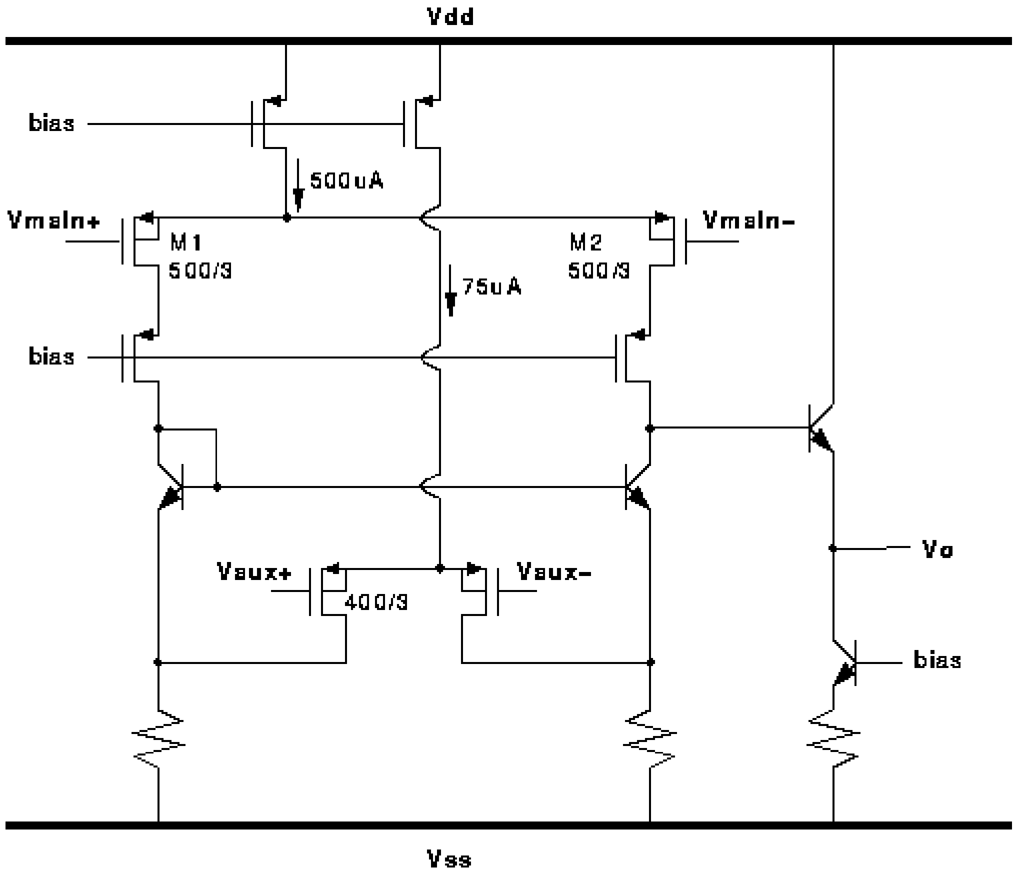
Die size	3 x 3 mm^2
Mass	0.5 μgram
Resonant frequency	4.7 kHz
Air gap	1.6 μm
Sense capacitance	500 fF
Feedback capacitance	300 fF
Noise floor	1.6 mG/rt-Hz

Ref.: C. Lu et al, “A surface micromachined accelerometer with digital output”, J. Solid-State Circ., pp. 1367-1373, Dec. 1995.

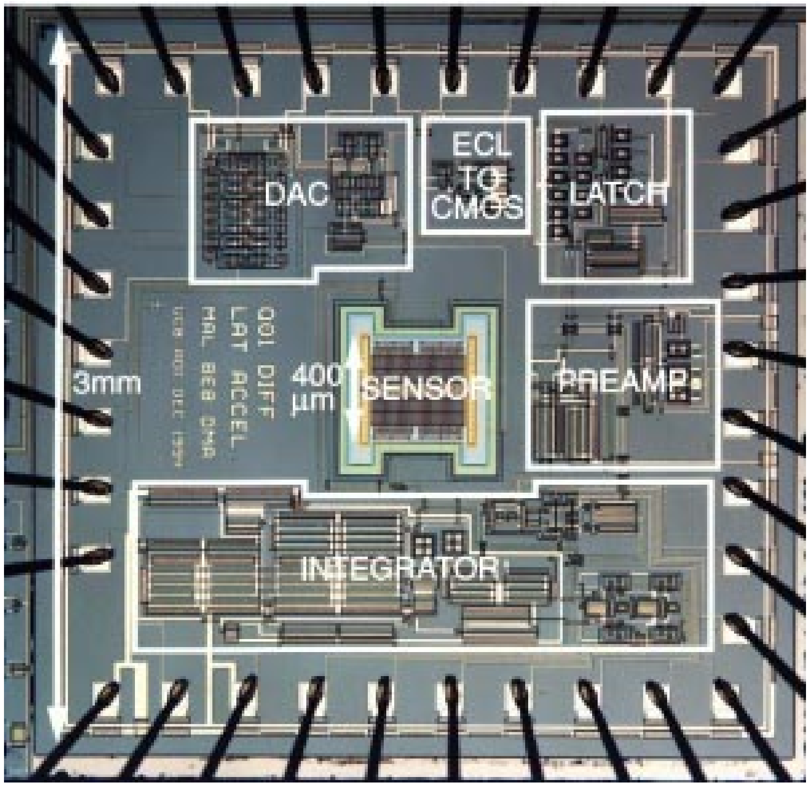
Example 1: CDS Position Sensor



Example 1: CDS Amplifier



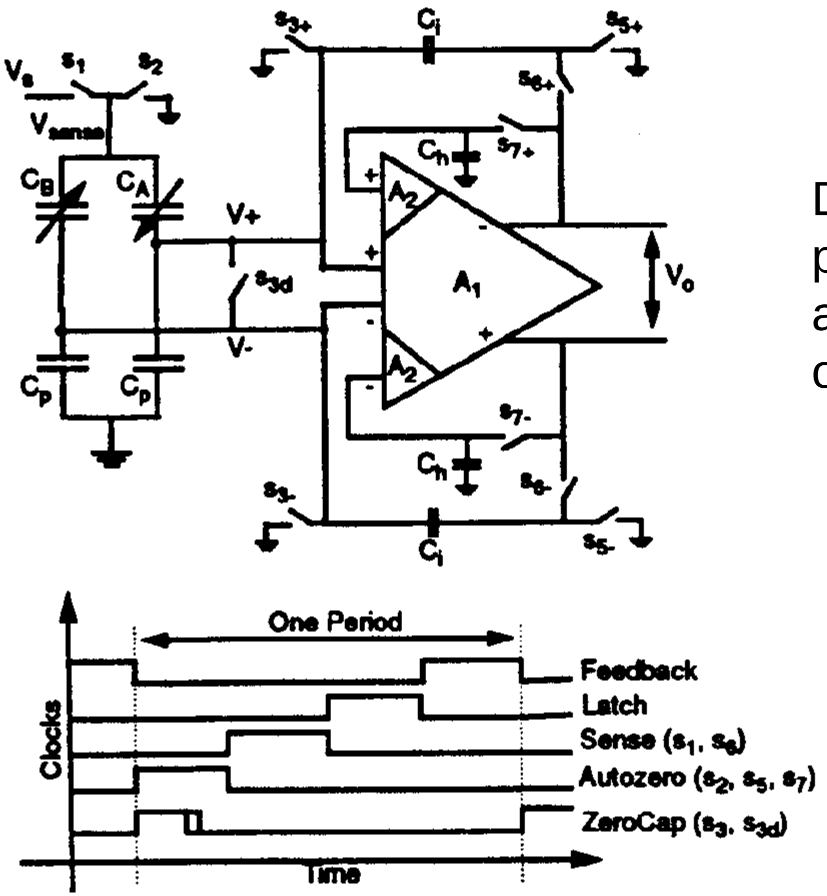
Example 2: Differential CDS Accelerometer



Process	3 µm BiCMOS
Power	3.7 mA @ 5V
Sampling rate	500 kHz
Full scale	±3.5 G
Noise floor	500 µG/rt-Hz
Resonant freq.	8.1 kHz
Proof mass	0.2 µgram
Sense cap.	84 fF per side
Sensitivity	64 fF/µm

Ref.: M. Lemkin et al, "A Micromachined Fully Differential Lateral Accelerometer", in Digest CICC 96, pp. 315-318, May 1996.

Example 2: Differential Position Sense Amp



Differential interface improves power supply and EMI rejection, and achieves first-order cancellation of switch charge injection errors.

Outline

- Capacitive Sensor Interface
- Bootstrapping
- Offset & Drift Reduction
- Charge Sensing
- Correlated Double Sampling

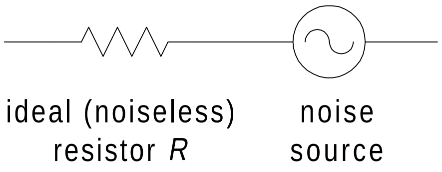
☞ Noise

Noise Sources

- “man-made”, e.g.
  - power supply coupling
  - clock feed-through
  - electromagnetic interference (EMI)
  - → *make arbitrarily small*
- “correlated”, e.g.
  - offset
  - 1/f noise
  - → *low frequency: measure and cancel*
- “thermal” (synonyms: Johnson, Brownian, …)
  - from thermodynamics
  - ***sets ultimate performance limit***

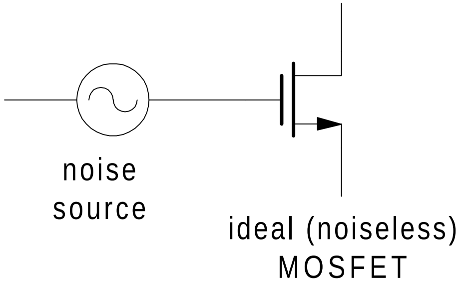
Device Thermal Noise Models

Resistor



$$\overline{v_n^2} = 4k_B T R \Delta f$$

MOSFET



$$\overline{v_n^2} = 4k_B T \frac{2}{3g_m} \Delta f$$

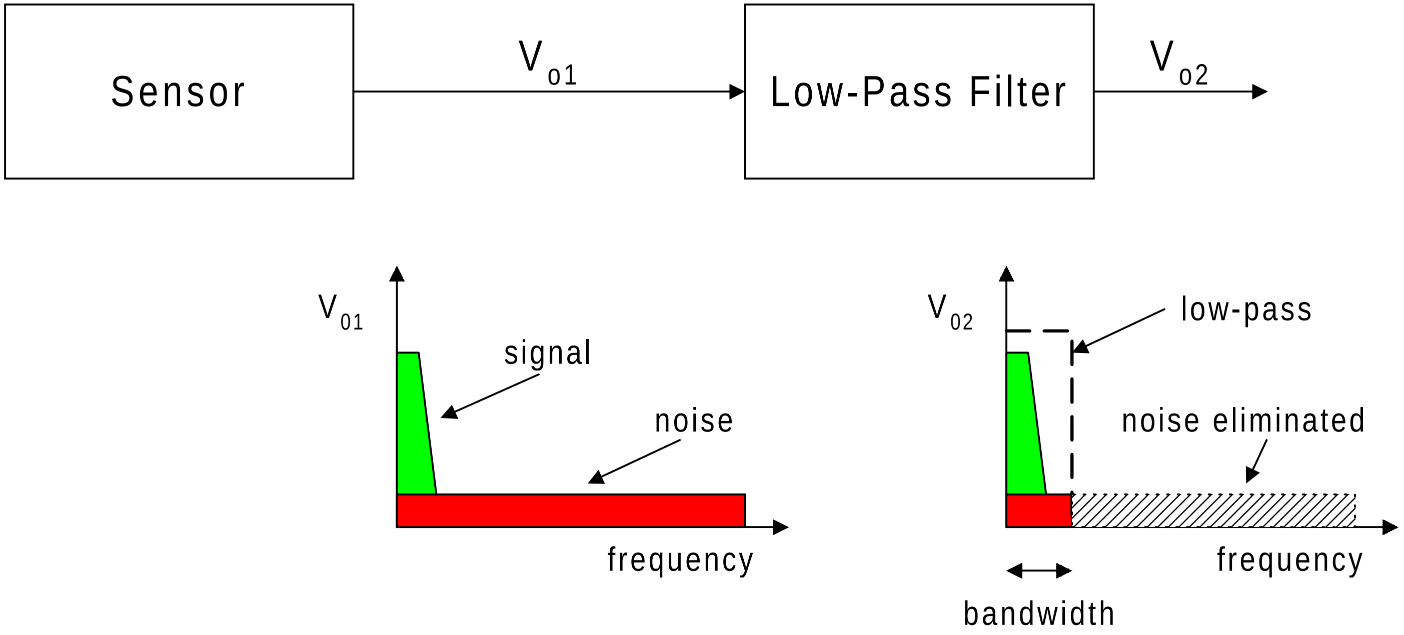
$k_B = 1.38 \times 10^{-23}\ {}^{J}\!/_{{}^0K}$ (Boltzmann's constant)

$k_B T = 4.1 \times 10^{-21}\ J$ at room temperature ($T = 300^0$ K)

$$e.g.\quad R = 1M\Omega,\quad \Delta f = 1kHz \quad \rightarrow \quad \underline{\sqrt{\overline{v_n^2}} = 2\mu V}$$

Noise Bandwidth

Typical sensor application:



Sensor

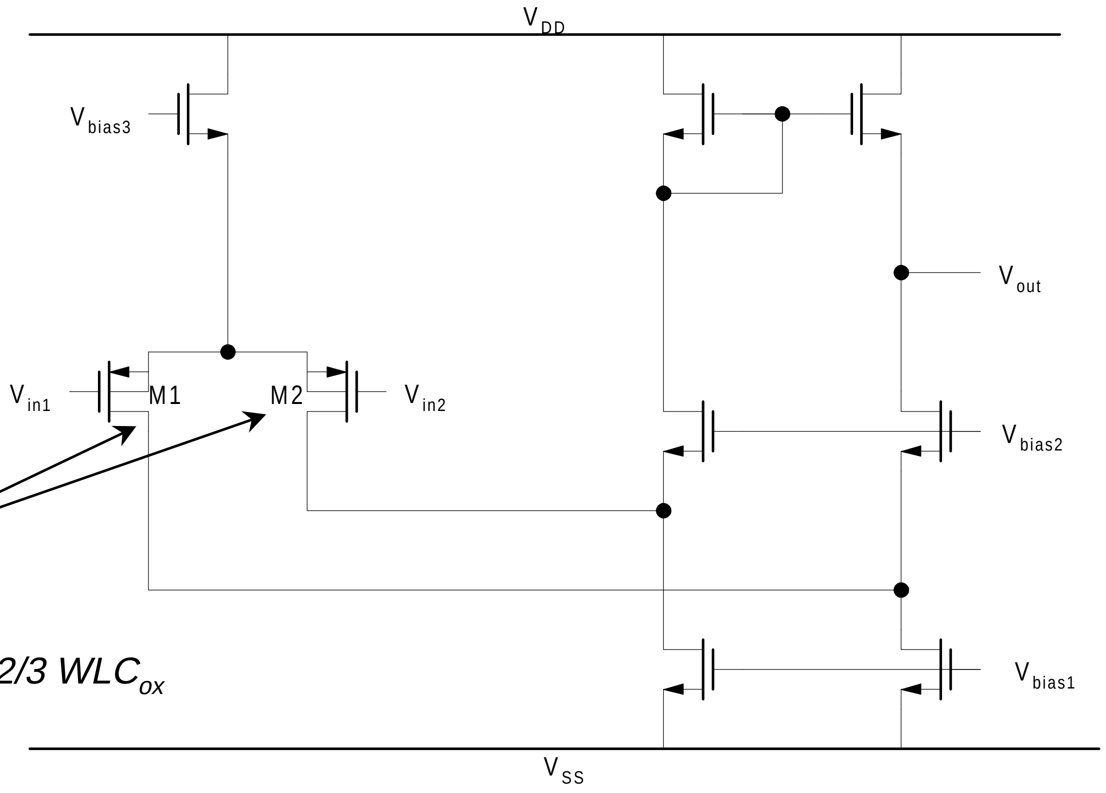
V_{o1}

Low-Pass Filter

V_{o2}

Noise power is proportional to sensor bandwidth (assuming flat spectrum).

MOS Amplifier Noise

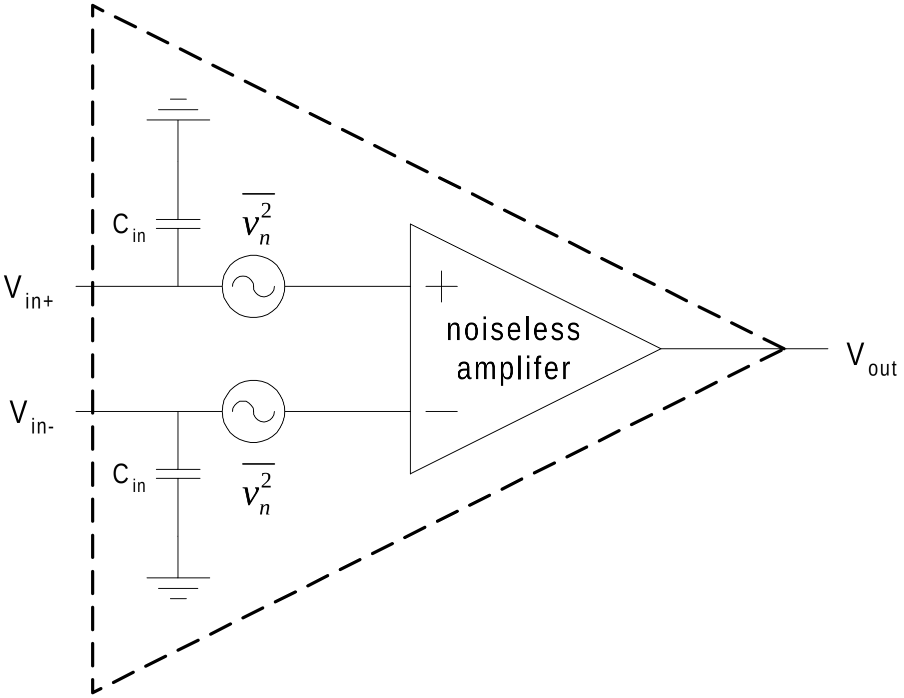


input transistors

transconductance g_m

gate capacitance $C_{gs} = 2/3\ WLC_{ox}$

MOS Amplifier Noise Model



$$\overline{v_n^2} \approx 4k_BT\frac{2}{3g_m}\Delta f(1+F)$$

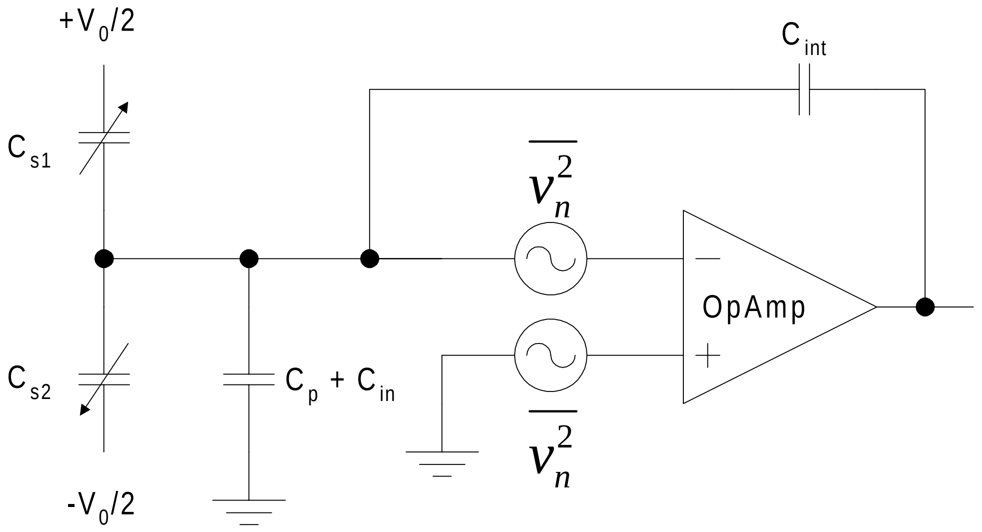
with $F > 0$

(typical: $F = 2 \ldots 5$*)*

$$C_{in} = C_{gs1} = C_{gs2}$$

(assumption: transconductance amplifier - high impedance output)

Noise Calculation



$$\overline{v_{n}^{2}}$$

$$\overline{v_{n}^{2}}$$

$$\overline{v_{o,n}^{2}} = 2\overline{v_{n}^{2}}\left[1 + \frac{C_T + C_{in}}{C_{int}}\right]^2$$

$$with \quad C_T = C_{s1} + C_{s2} + C_p$$

Input Transistor Sizing

- increasing g_m reduces amplifier noise

$$\overline{v_n^2} \propto \frac{1}{g_m}$$

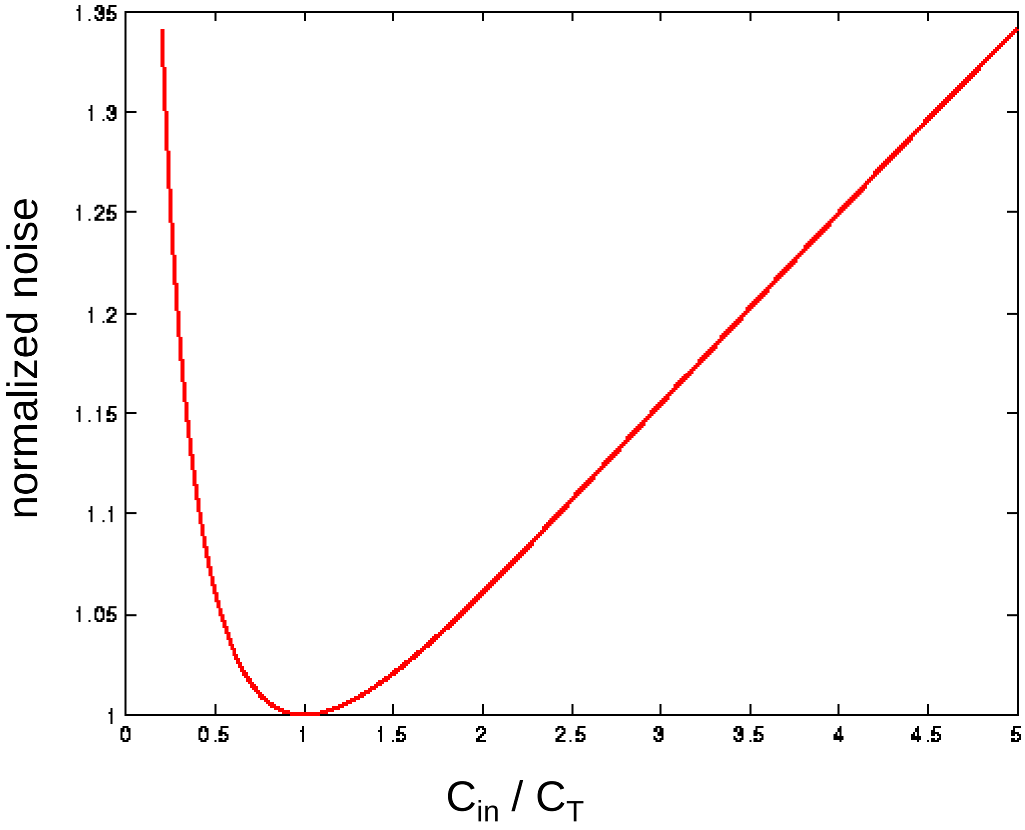
- *CAVEAT:*

$$C_{in} \approx C_{gs} = \frac{g_m}{\omega_T} \qquad \text{typical } \omega_T = 2\pi[1\ \ldots\ 10\ Grad/sec]$$

implies C_{in} and hence noise increase with g_m

➡ *OPTIMUM*

Noise Optimum



- shallow optimum for $C_{in} = C_T$
- C_{in} (transistor size) also affects speed: in practice often choose $C_{in} > C_T$

Minimum Detectable Signal (MDS)

- Definition: signal power = noise power

$$\sqrt{\frac{\overline{\left(\Delta C / C_0\right)^2}}{\Delta f}} \approx \frac{8}{V_0}\sqrt{\frac{2k_B T}{3\omega_T C_0}\left(1+\frac{C_p}{C_0}\right)}$$

Assumption: $C_{in} = C_T$ (optimal transistor sizing)

Ref: B. Boser et al, “Surface micromachined accelerometers”,
IEEE J. Solid-State Circuits, pp. 366-375, March 1996.

Example: Accelerometer

- proof-mass displacement

$$x = \frac{a_{in}}{\omega_r^2} \qquad \omega_r = 2\pi f_r = \text{ resonant frequency}$$

- parallel plate capacitor

$$\frac{dC}{dx} = \frac{C_0}{x_0}$$

- acceleration noise floor

$$\sqrt{\frac{\overline{a_{in}^{\,2}}}{\Delta f}} \approx \frac{32\pi f_r^{\,2} x_0}{V_0} \sqrt{\frac{2k_B T}{3\omega_T C_0}\left(1 + \frac{C_p}{C_0}\right)}$$

E.g. $C_0 = 500fF$, $\omega_T = 2\pi \times 500MHz$, $V_0 = 500mV$, $f_r = 5kHz$, $x_0 = 1\mu m$

$$\sqrt{\frac{\overline{a_{in}^{\,2}}}{\Delta f}} \approx 1\mu G / \sqrt{Hz}$$

Summary

- Capacitive Sensor Interface
- Bootstrapping
- Offset & Drift Reduction
- Charge Sensing
- Correlated Double Sampling
- Noise