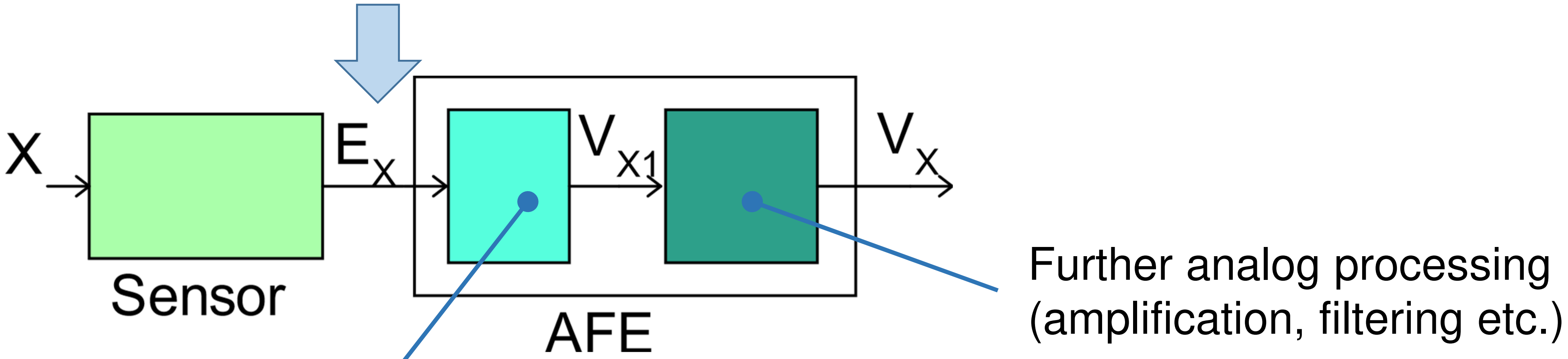
Sensor Interfaces



Sensor Interface: directly connected to the sensor. Detects the typically small variations of the electrical quantity E_X and convert it into a voltage, if required.

Sensors that produce a voltage

Output electrical quantity	Sensor type	Input physical or chemical quantity to be sensed.
Voltage	Thermoelectric sensors	**Temperature difference** Temperature Fluid flow rate Infrared radiation (bolometers) Gas concentration (catalytic sensors)
	Electrochemical sensors	**Ion concentration in electrolytes** Gas concentration (e.g.. “lambda probes”)
	Hall sensors	**Magnetic Field** Position (Proximity) Current measurement
	Piezoelectric sensors	**Force (ac detection)** Acoustical pressure, acceleration.

Sensors that produce a current or a charge

Output electrical quantity	Sensor type	Input physical or chemical quantity to be sensed.
Current	Optical sensors (photodiodes)	**Infrared, visible and Ultraviolet radiation** Imagers Proximity Opacity (e.g. smoke detectors)
Charge	CCD imagers	**Visible radiation**
	High energy particle detectors	**Ionizing radiation and particle detection**

Resistive sensors

Output electrical quantity	Sensor type	Input physical or chemical quantity to be sensed
Resistance	Thermistor and RTDs (Resistive Temperature Detectors)	**Temperature** Fluid flow rates Fluid velocity (e.g. hot wire anemometers) Gas concentration (catalytic sensors) Proximity
	Piezo-resistors	**Strain (strain gauges)** Force (e.g. electronic scales) Pressure (barometers) Altitude Acceleration
	Chemi-resistors	**Gas or vapor concentration**
	Magneto-resistors	**Magnetic field** Proximity Orientation (e.g. electronic compass)
	Photo resistors	**Visible radiation**

Capacitive sensors

Output electrical quantity	Sensor type	Input physical or chemical quantity to be sensed.
Capacitance	Capacitive sensors (mechanical)	Acceleration Angular velocity (gyroscopes) Pressure
	Capacitive sensors (chemical)	**Gas concentration (e.g. humidity sensors)**

Frequently used sensor interfaces

Output quantity	Type of Interface	Notes
Voltage	Instrumentation Amplifier (In-Amp)	
Resistance	Instrumentation Amplifier	Resistors should be mounted in a Wheatstone bridge configuration or biased by a current.
	Trans-Impedance Amplifier (TIA)	Resistor must be biased with a voltage in order to produce a current
Current	Trans-Impedance Amplifier (TIA)	
Capacitance	Trans-Impedance Amplifier (TIA)	Converting capacitance into a current by means of a periodic voltage waveform
	Charge amplifier (switched capacitor)	
Charge	Charge amplifier (switched capacitors)	

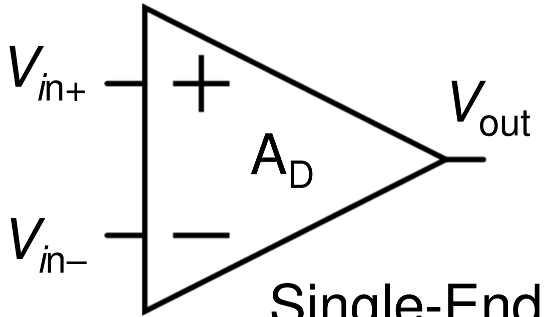
Instrumentation Amplifiers (In-Amps)

Required features:

- **Precise gain (A_D)**
- **High input resistance**

Other important features

- **Differential input**
- Low input referred offset voltage
- Low bias currents
- Low input referred voltage and current noise
- High CMRR (for differential amplifiers)
- Large bandwidth

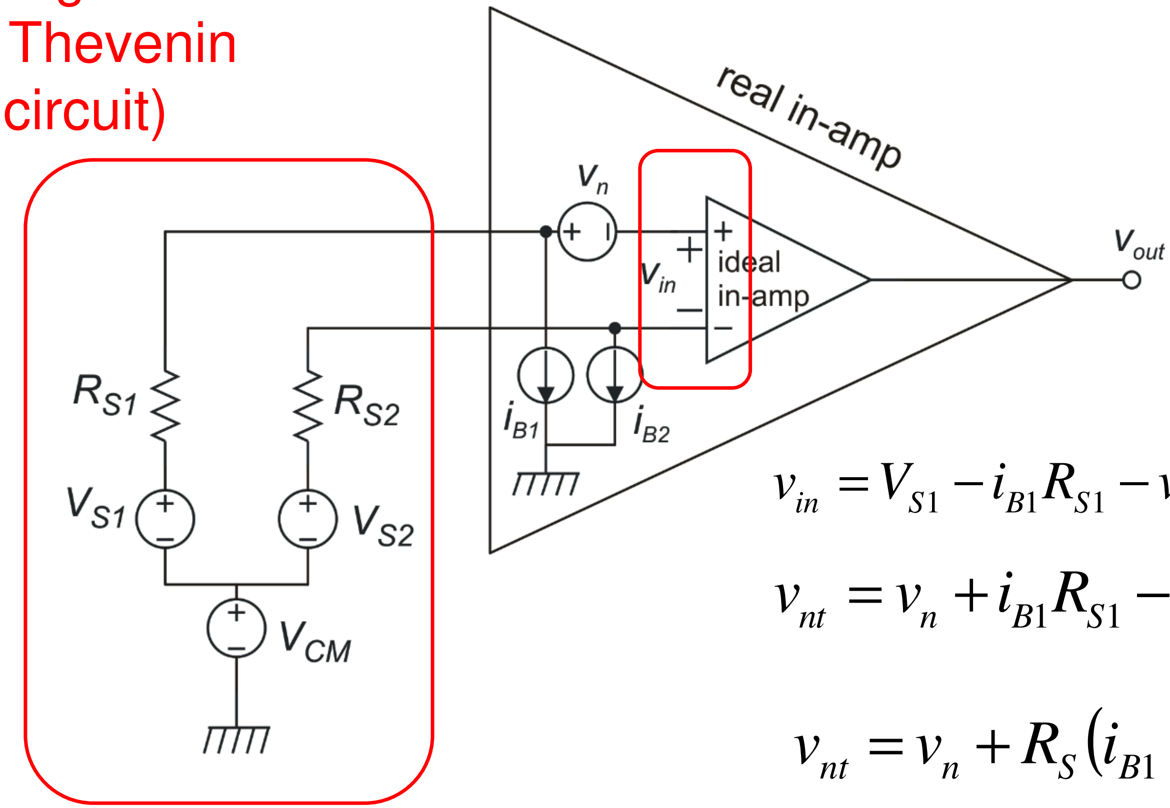


Single-Ended Instrumentation Amplifier

In-amps: errors due to the input noise/offset voltage and currents

Useful Signal: $V_{S1} - V_{S2}$

Differential signal source (Simplified Thevenin equivalent circuit)



$$v_{in} = V_{S1} - i_{B1}R_{S1} - v_n - \left(V_{S2} - i_{B2}R_{S2}\right) = V_{S1} - V_{S2} - v_{nt}$$

$$v_{nt} = v_n + i_{B1}R_{S1} - i_{B2}R_{S2}$$

$$v_{nt} = v_n + R_S\left(i_{B1} - i_{B2}\right)$$

Balanced source case ($R_{S1}=R_{S2}=R_S$)

Total voltage noise expressions

Balanced source case ($R_{S1}=R_{S2}=R_S$)

$$v_{nt} = v_n + R_S\left(i_{B1} - i_{B2}\right)$$

DC components:

$$v_n = v_{io} \quad i_{b1} = I_{B1};\; i_{b2} = I_{B2};\quad \left(i_{B1} - i_{B2}\right) = I_{B1} - I_{B2} = I_{io}$$

$$v_{iot} = v_{io} + R_S I_{io}$$

Noise components:

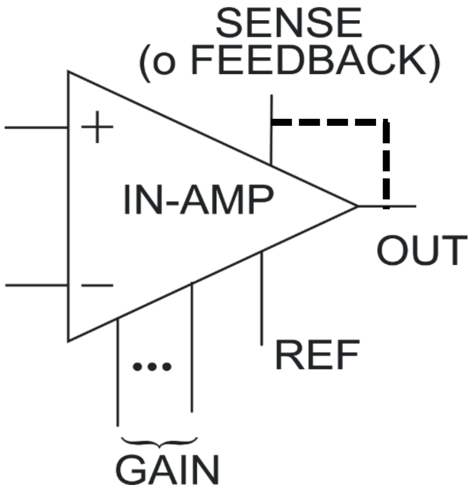
$$v_n \Rightarrow S_{vn};\; \begin{cases} i_{b1} = i_{n1} \Rightarrow & S_{I1}(f) \\ i_{b2} = i_{n2} \Rightarrow & S_{I2}(f) \end{cases} \quad S_{I12}(f)$$

$$S_{vnt} = S_{vn} + R_S^2\left(S_{I1} + S_{I2} - 2S_{I1I2}\right)$$

If i_{n1} and i_{n2} are uncorrelated and $S_{I1} = S_{I2} = S_I$:

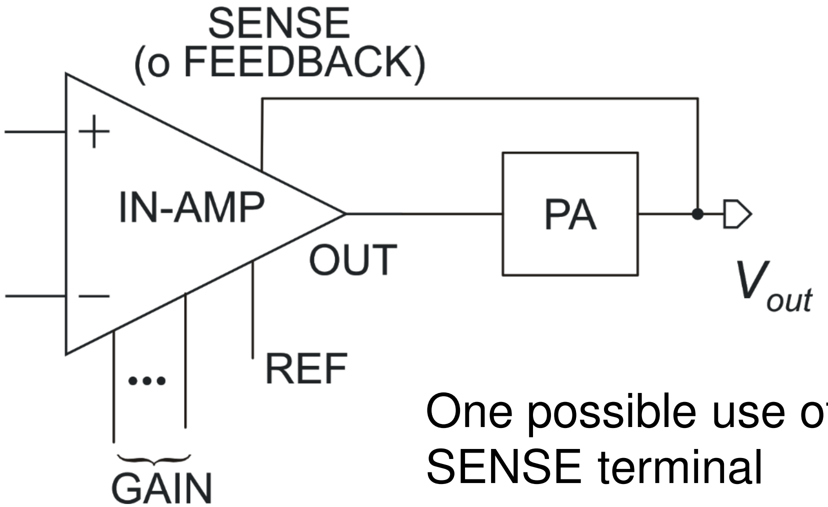
$$S_{vnt} = S_{vn} + 2R_S^2 S_I$$

Discrete monolithic In-Amps



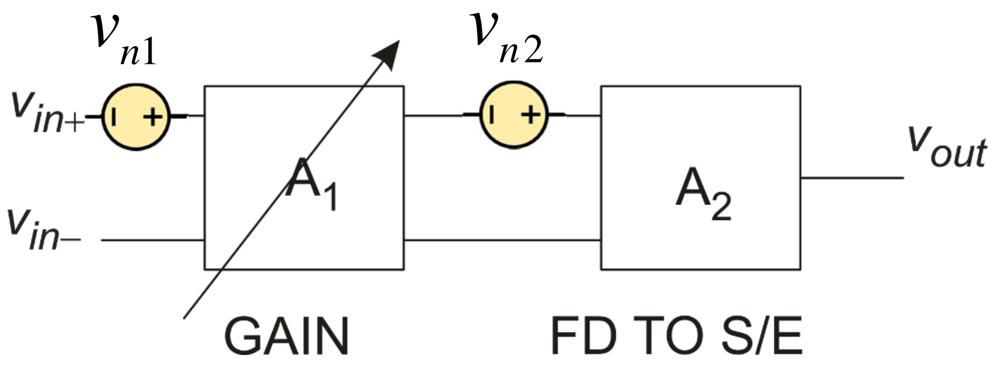
Typical pin configuration

$$V_{out} = G(V_{IN+} - V_{IN-}) + V_{REF}$$



One possible use of the SENSE terminal

In-Amp with variable gain: Input and Output offset and noise



Typical two-stage architecture of In-amps $\quad G = A_1 A_2$

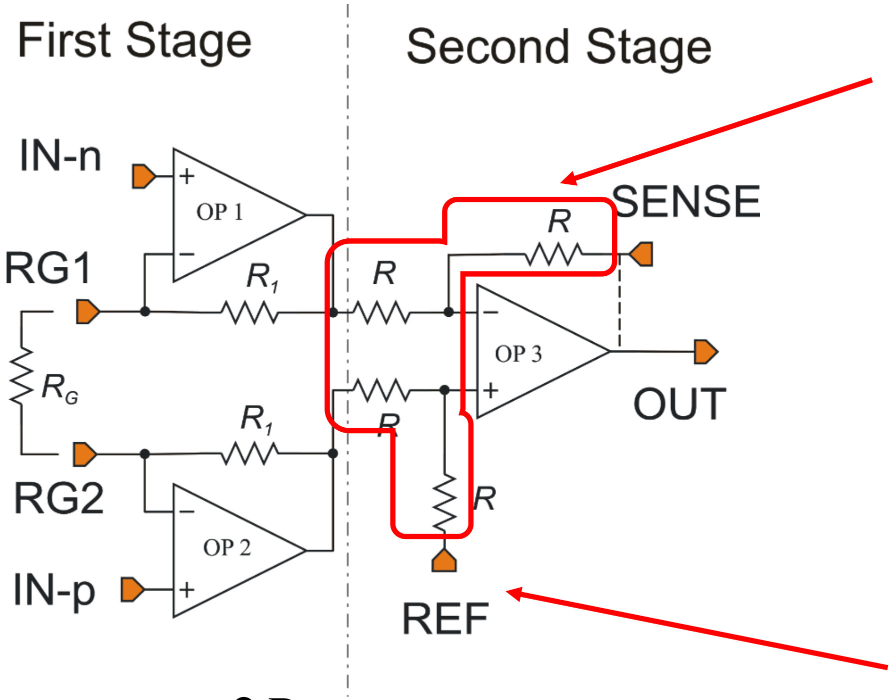
$$v_{n-out} = v_{n1} A_1 A_2 + v_{n2} A_2 \qquad v_{nRTI} = \frac{v_{n-out}}{G} = v_{n1} + \frac{v_{n2}}{A_1}$$

Generally, A_2= 1, thus: $A_1 = G \qquad v_{nRTI} = v_{n1} + \dfrac{v_{n2}}{G}$

v_{n1} : input noise

v_{n2} : output noise

Three-Op-Amp Instrumentation Amplifier

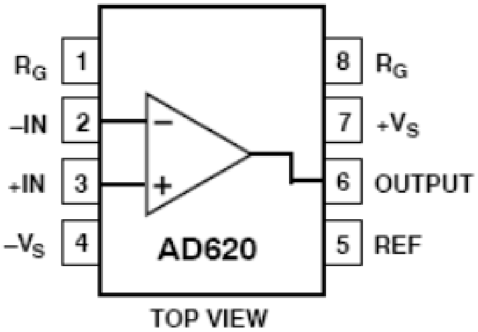


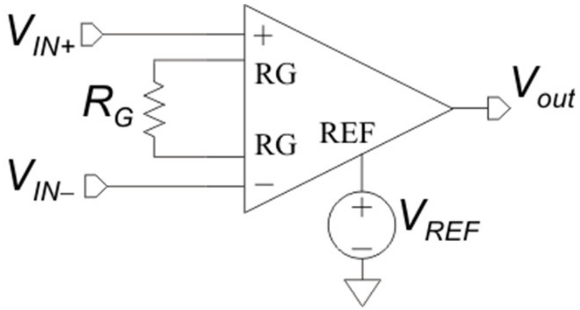
$$G = 1 + \frac{2R_1}{R_G}$$

Mismatch of these resistors degrades the CMRR of the second stage. Resistor trimming is necessary for CMRRs > 60 dB

The circuit that provides V_{REF} must have a very low output resistance (<<R)

AD 620





$$V_{out} = G(V_{IN+} - V_{IN-}) + V_{REF}$$

$$G = 1 + 49.4k\Omega / R_G$$

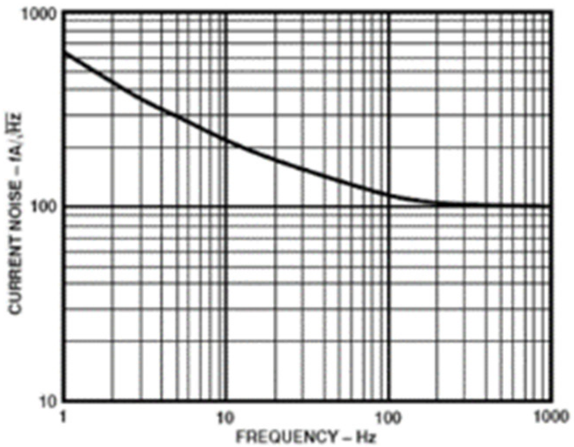


Figure 9. Current Noise Spectral Density vs. Frequenc

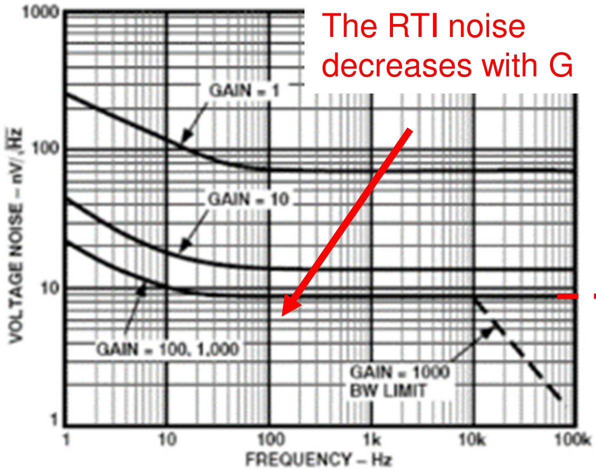


Figure 8. Voltage Noise Spectral Density vs. Frequency, (G = 1–1000)

The RTI noise decreases with G

At G≥100 the amplifier BW is larger than 100kHz. The noise density starts to fall for f> 100kHz

AD 620

GBW does not increase much beyond the G=10 case: BW affected by second stage

Nearly constant GBW product: BW determined by first stage

AD620

Model	Conditions	AD620A Min	AD620A Typ	AD620A Max	AD620B Min	AD620B Typ	AD620B Max	AD620S[1] Min	AD620S[1] Typ	AD620S[1] Max	Units
DYNAMIC RESPONSE											
Small Signal –3 dB Bandwidth											
G = 1			1000			1000			1000		kHz
G = 10			800			800			800		kHz
G = 100			120			120			120		kHz
G = 1000			12			12			12		kHz
Slew Rate		0.75	1.2		0.75	1.2		0.75	1.2		V/µs
Settling Time to 0.01%	10 V Step										
G = 1–100			15			15			15		µs
G = 1000			150			150			150		µs
NOISE											
Voltage Noise, 1 kHz	$Total\ RTI\ Noise = \sqrt{(e^2_{ni}) + (e_{no}/G)^2}$										
Input, Voltage Noise, e_{ni}			9	13		9	13		9	13	nV/√Hz
Output, Voltage Noise, e_{no}			72	100		72	100		72	100	nV/√Hz
RTI, 0.1 Hz to 10 Hz											
G = 1			3.0			3.0	6.0		3.0	6.0	µV p-p
G = 10						0.55	0.8		0.55	0.8	µV p-p
G = 100–1000						0.28	0.4		0.28	0.4	µV p-p
Current Noise	f = 1 kHz		100			100			100		fA/√Hz
0.1 Hz to 10 Hz			10			10			10		pA p-p

Slew -Rate

Settling times

Output noise >> input noise

Broad-Band Noise: $\sqrt{S_{BB}}$

Broad-Band Noise: $\sqrt{S_{BB}}$

Current

Low Frequency Noise
Integrated over 0.1-10 Hz

Low Frequency Noise
Integrated over 0.1-10 Hz

AD 620

The effective input referred offset (RTI) is a combination of the input and output offset

Input offset

Output offset

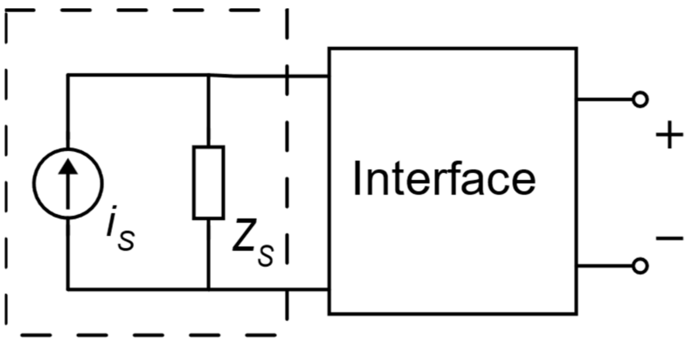
VOLTAGE OFFSET	(Total RTI Error = V_{OSI} + V_{OSO}/G)										
Input Offset, V_{OSI}	V_S = ±5 V to ±15 V		30	125		15	50		30	125	μV
Over Temperature	V_S = ±5 V to ±15 V			185			85			225	μV
Average TC	V_S = ±5 V to ±15 V		0.3	1.0		0.1	0.6		0.3	1.0	μV/°C
Output Offset, V_{OSO}	V_S = ±15 V		400	1000		200	500		400	1000	μV
	V_S = ±5 V			1500			750			1500	μV
Over Temperature	V_S = ±5 V to ±15 V			2000			1000			2000	μV
Average TC	V_S = ±5 V to ±15 V		5.0	15		2.5	7.0		5.0	15	μV/°C
Offset Referred to the Input vs. Supply (PSR)	V_S = ±2.3 V to ±18 V										
G = 1		80	100		80	100		80	100		dB
G = 10		95	120		100	120		95	120		dB
G = 100		110	140		120	140		110	140		dB
G = 1000		110	140		120	140		110	140		dB
INPUT CURRENT											
Input Bias Current			0.5	2.0		0.5	1.0		0.5	2	nA
Over Temperature				2.5			1.5			4	nA
Average TC			3.0			3.0			8.0		pA/°C
Input Offset Current			0.3	1.0		0.3	0.5		0.3	1.0	nA
Over Temperature				1.5			0.75			2.0	nA
Average TC			1.5			1.5			8.0		pA/°C

The input bias current and input offset current are similar, since such a small bias current is the result of internal bias current cancellation

POWER SUPPLY											
Operating Range[4]		±2.3		±18	±2.3		±18	±2.3		±18	V
Quiescent Current	V_S = ±2.3 V to ±18 V		0.9	1.3		0.9	1.3		0.9	1.3	mA
Over Temperature			1.1	1.6		1.1	1.6		1.1	1.6	mA

The AD 620 in-amp represents a good trade-off between input noise voltage, input bias currents and supply current (quiescent current)

Interfacing a sensor whose output is a current



Sensor
(Norton equivalent circuit)

$$v_{out} = Z \cdot i_s$$

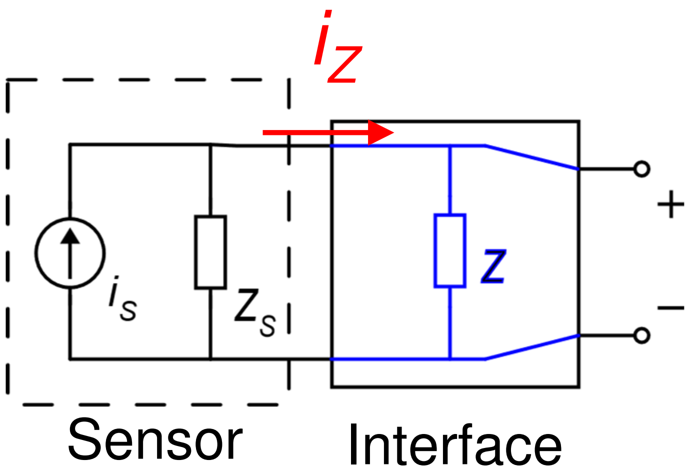
Goal:
to obtain a voltage proportional to the sensor current

Z is a property of the interface and must be accurate.
In many cases *Z* is a resistance

I_S is the quantity to be read, since it contain the useful information

Z_S is the unavoidable output impedance of the sensor.
The ideal case is an infinite Z_S (ideal current source)

The simplest solution



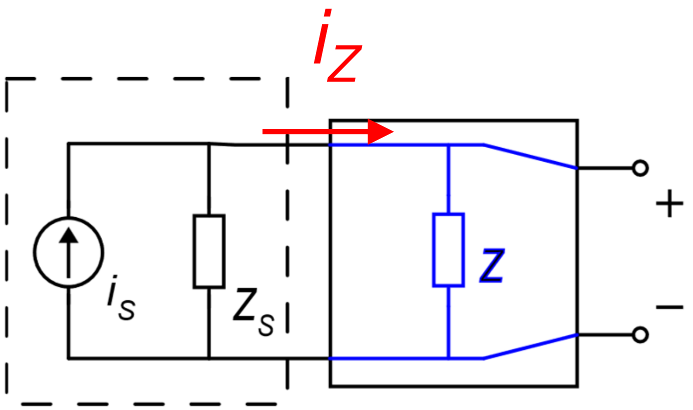
Simply connect a known impedance cross the sensor terminals

$$V_{out} = Z \cdot i_Z$$

$$i_Z = \frac{Z_S}{Z_S + Z} i_S = \frac{1}{1 + \frac{Z}{Z_S}} i_S$$

$$i_Z \neq i_S$$

Sensitivity vs Accuracy



$$v_{out} = Z \cdot i_Z$$

$$i_Z = \frac{1}{1+\dfrac{Z}{Z_S}} i_S \cong \left(1 - \frac{Z}{Z_S}\right) i_S$$

$$v_{out} \cong Z\left(1 - \frac{Z}{Z_S}\right) i_S$$

$$v_{out} = Z \cdot i_S$$

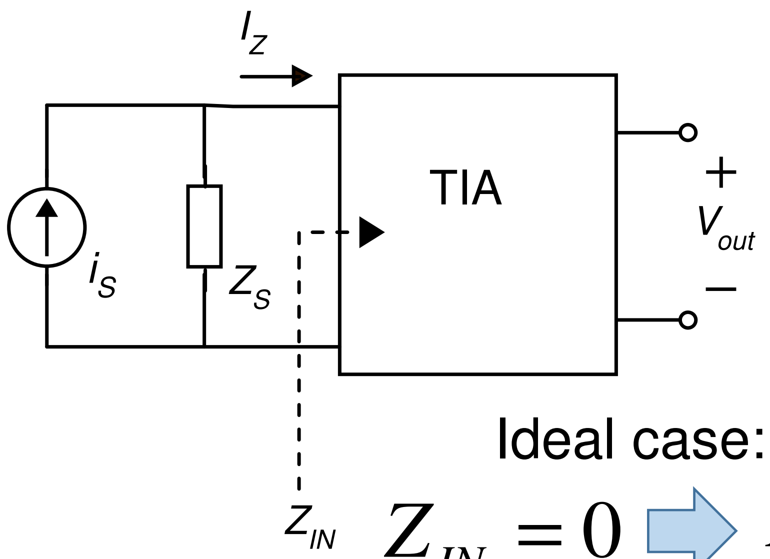
Nominal law

Z should be large (magnitude) to obtain a high sensitivity

Z should be small (magnitude) to reduce the relative error (high accuracy)

For sensor marked by small values of the Z_S impedance, finding a value of *Z* that satisfies both requirements is often impossible

A better interface: the Trans-Impedance Amplifier (TIA)

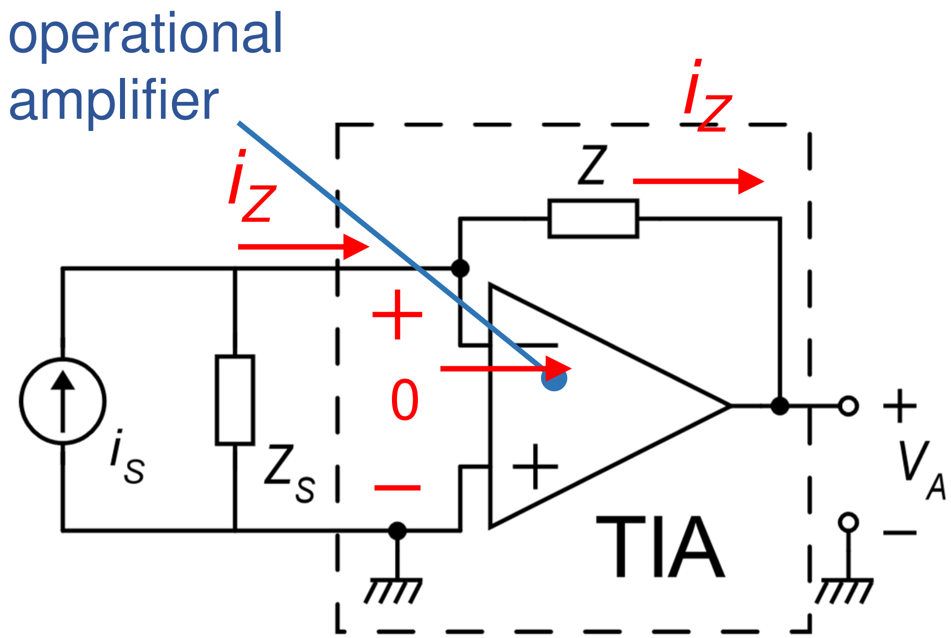


$$v_{out} = Z \cdot I_Z$$

Ideal case:

$$Z_{IN} = 0 \Rightarrow I_Z = i_S \Rightarrow v_{out} = Z \cdot i_S$$

The op-amp based TIA



Ideal case (perfect virtual gnd)

$$V_{IN} = 0 \Rightarrow Z_{IN} = 0$$

$$I_Z = i_S$$

$$v_A = -Z \cdot I_Z = -Z \cdot i_S$$

Sensitivity of the TIA stage (transimpedance)

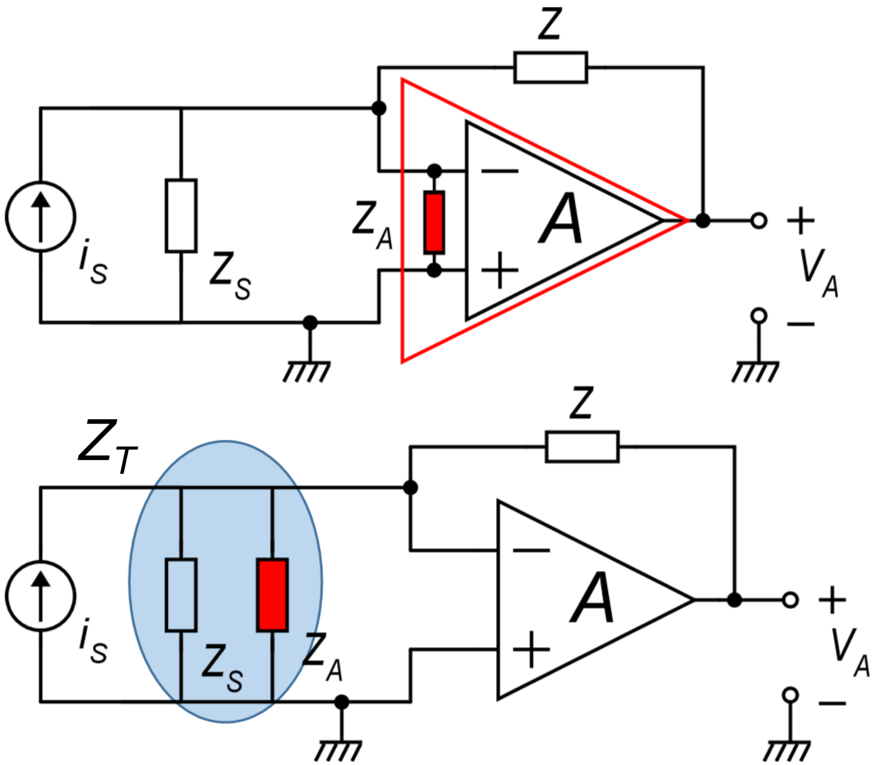
$$k_{TIA} = \frac{v_{An}}{i_s} = -Z$$

TIA non-idealities

- Finite gain
- Finite input impedance

$$A = \frac{A_0}{1 + j\dfrac{f}{f_p}}$$

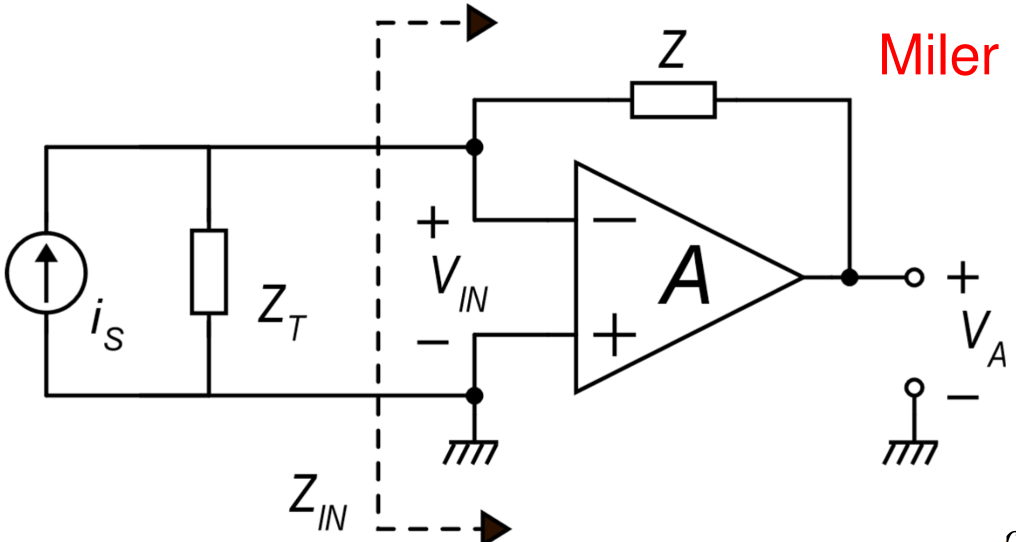
Typical dominant-pole frequency response



$$Z_T = Z_S // Z_A$$

TIA: input impedance Z_{IN}

$$Z_T = Z_S // Z_A$$



Miler Effect:

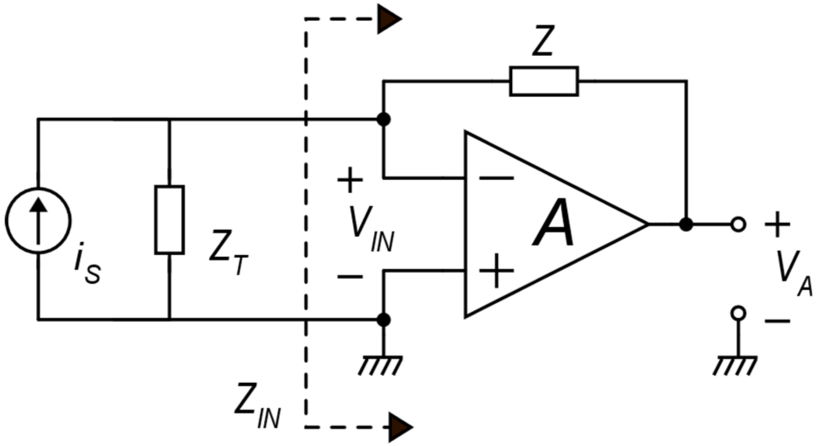
$$Z_{IN} = \frac{Z}{1-K_M} \qquad K_M = \frac{v_{out}}{v_{in}} = -A$$

$$Z_{IN} = \frac{Z}{1+A}$$

$$Z_{IN} = Z\frac{1}{1+\dfrac{A_0}{1+j\dfrac{f}{f_p}}} = Z\frac{1+j\dfrac{f}{f_p}}{1+A_0+j\dfrac{f}{f_p}} = \frac{Z}{1+A_0}\left[\frac{1+j\dfrac{f}{f_p}}{1+j\dfrac{f}{\underline{(1+A_0)f_p}}}\right]$$

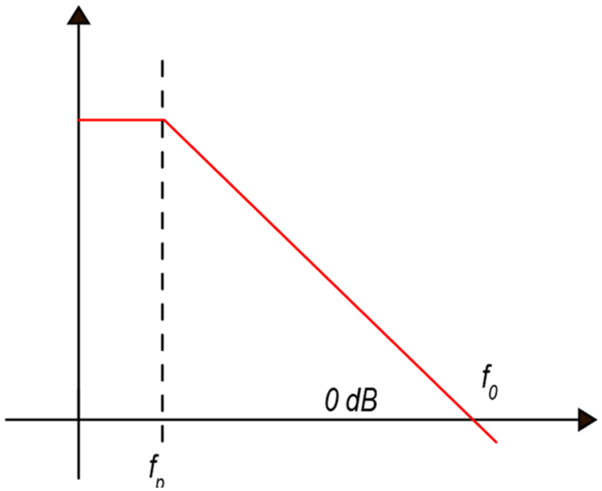
$\cong f_0$

TIA: input impedance Z_{IN}



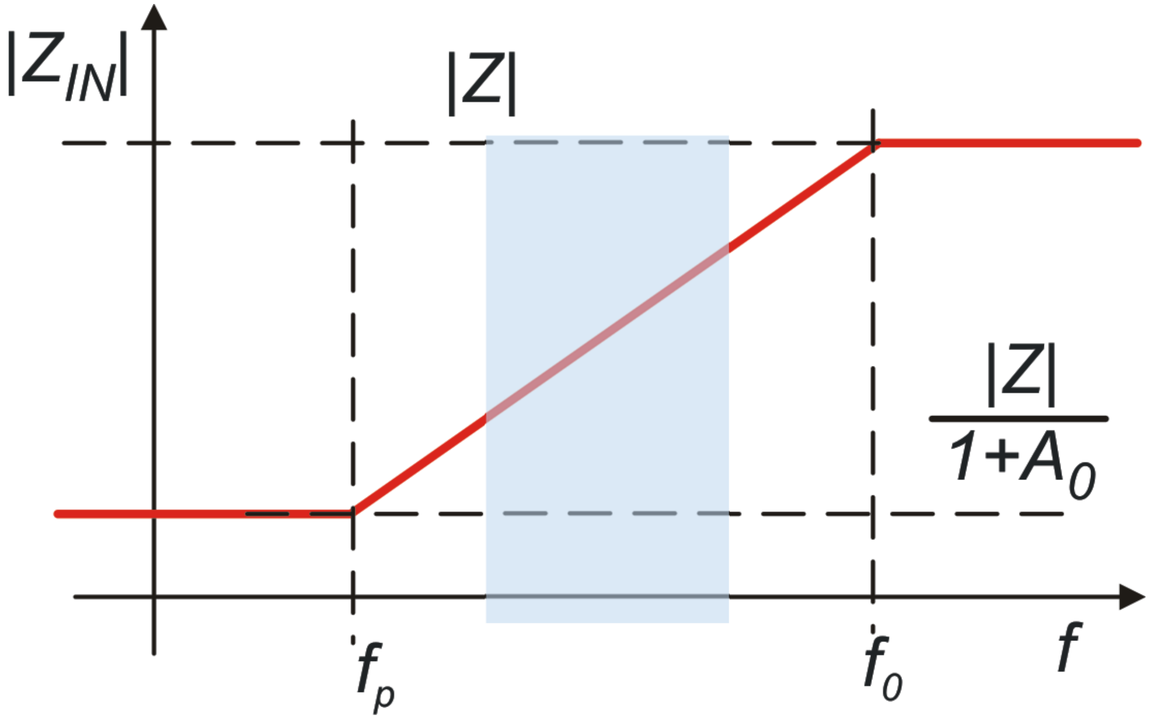
$$Z_{IN} = \frac{Z}{1+A_0}\frac{1+j\dfrac{f}{f_p}}{1+j\dfrac{f}{f_0}}$$

$$f_0 \cong A_0 f_p \cong \left(1+A_0\right) f_p$$



f_0 is defined as the frequency at which the amplifier gain magnitude is unity (0 dB)

TIA input impedance



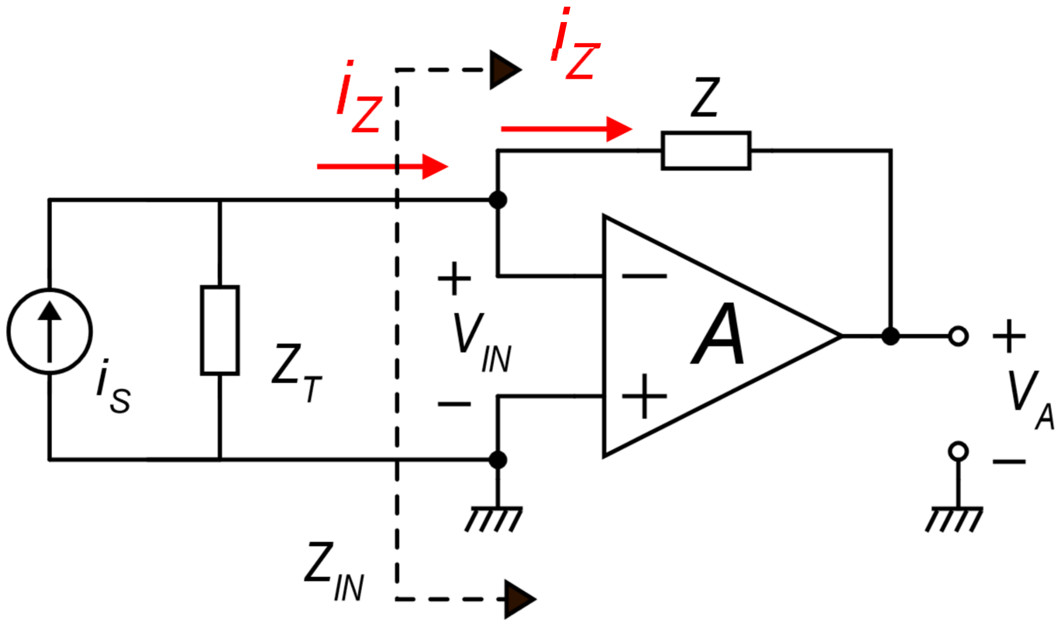
$$Z_{IN} = \frac{Z}{1+A_0}\frac{1+j\dfrac{f}{f_p}}{1+j\dfrac{f}{f_0}}$$

$$f_P \ll f \ll f_0$$

$$Z_{IN} = jZ\frac{f}{f_0}$$

Error due to the finite input impedance: (1) error on I_Z

Ideal case: $V_A = -Z \cdot I_S$



$$I_Z = I_S \frac{Z_T}{Z_{IN} + Z_T} = \frac{1}{1 + \dfrac{Z_{IN}}{Z_T}}$$

if $|Z_{IN}| << |Z_T|$ $\quad I_Z \cong I_S \left(1 - \dfrac{Z_{IN}}{Z_T}\right)$

$$\varepsilon_A \cong \left|\frac{Z_{IN}}{Z_T}\right|$$

Error on current I_Z

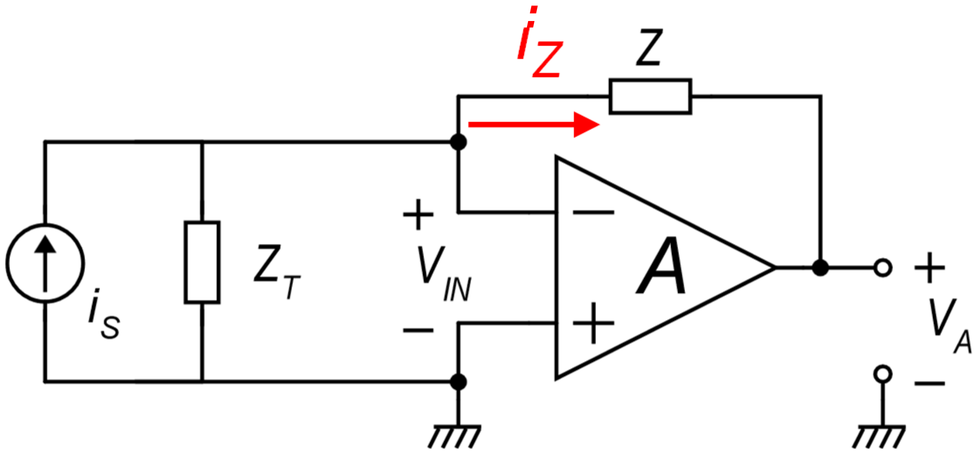
Error due to the finite input impedance: (2) error due to $V_{IN} \neq 0$

Ideal case: $V_A = -Z \cdot I_S$

$$V_A = v_{in} - I_Z Z \qquad v_{in} = -\frac{V_A}{A}$$

$$V_A = \frac{-V_A}{A} - I_Z Z \Rightarrow V_A\left(1 + \frac{1}{A}\right) = -I_Z Z$$

$$V_A = -\frac{I_Z Z}{1 + \frac{1}{A}} \cong -I_Z Z\left(1 - \frac{1}{A}\right)$$



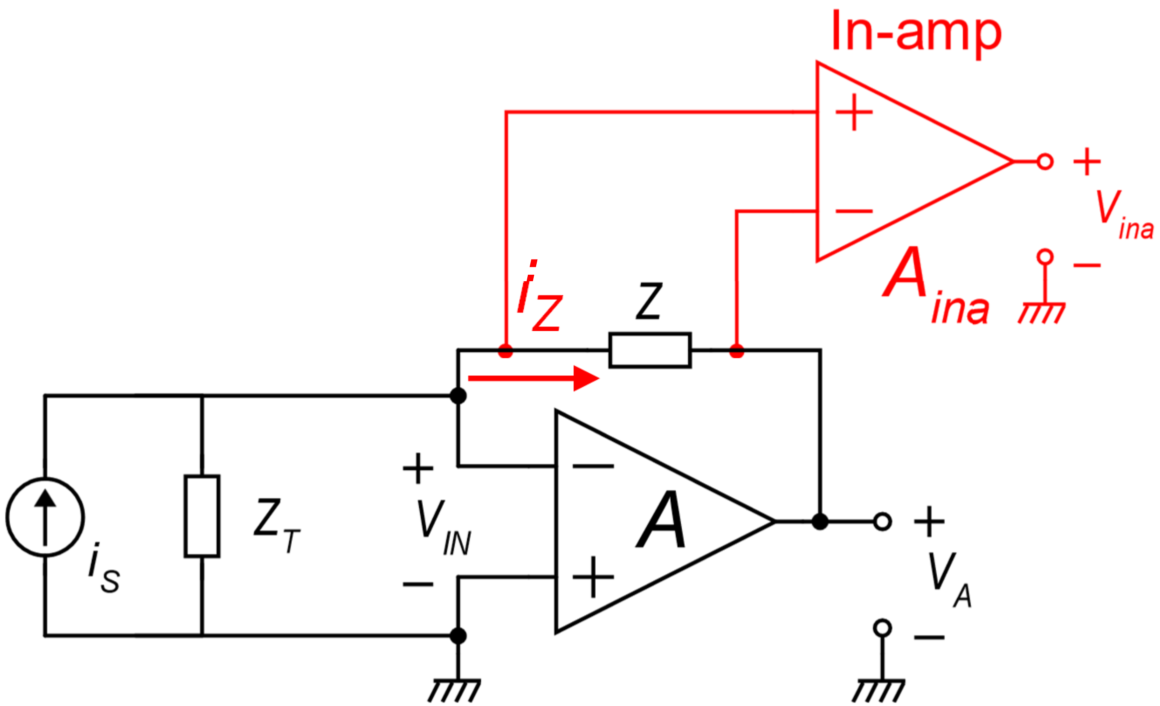
$$\varepsilon_A \cong \left|\frac{Z_{IN}}{Z_T}\right| \qquad \frac{1}{|A(f)|}$$

Relative
error contributions

Reducing the 1/A error

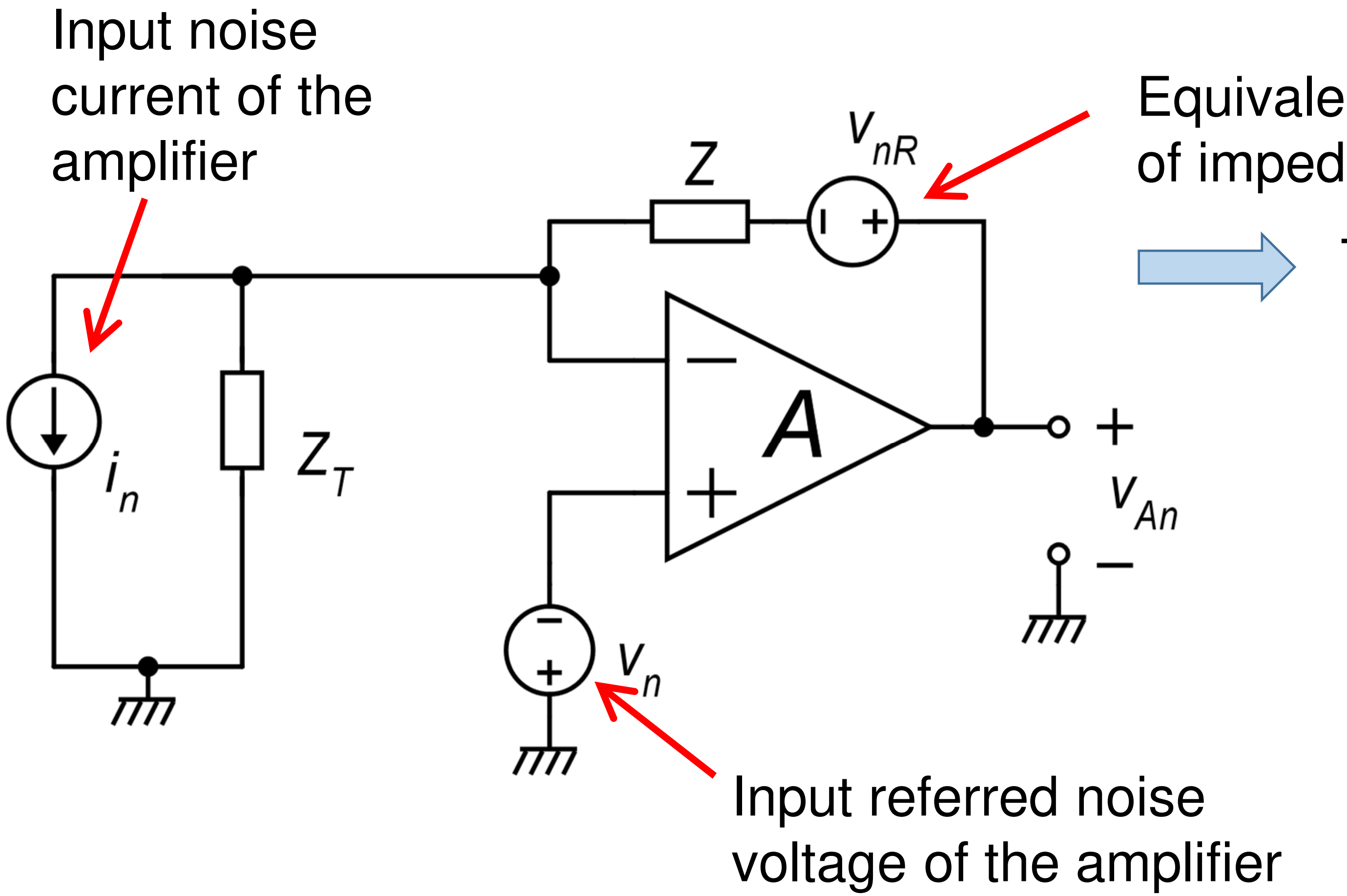
Ideal case: $V_A = -Z \cdot I_S$

This voltage is affected only by the error due to: $I_S \neq I_Z$



$V_{ina} = A_{ina} I_Z Z$

TIA non-ideality: Noise and offset

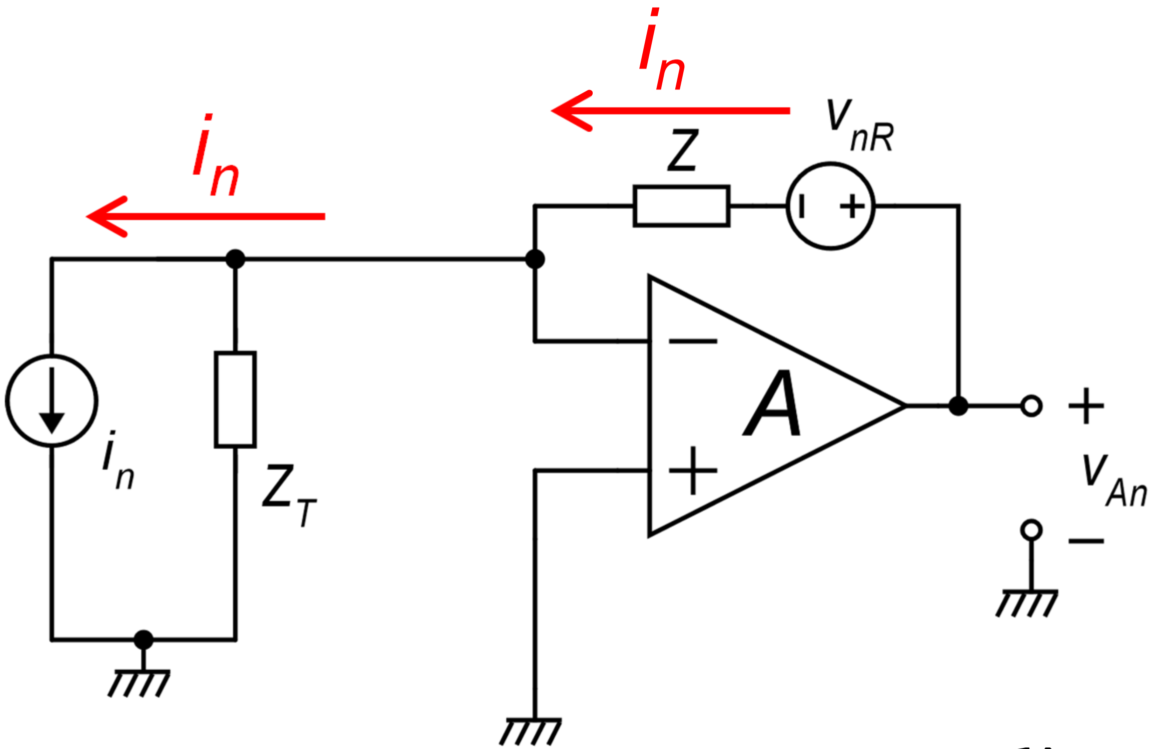


Thermal Noise:

$$S_{VnR} = 4kT\,\mathrm{Re}(Z)$$

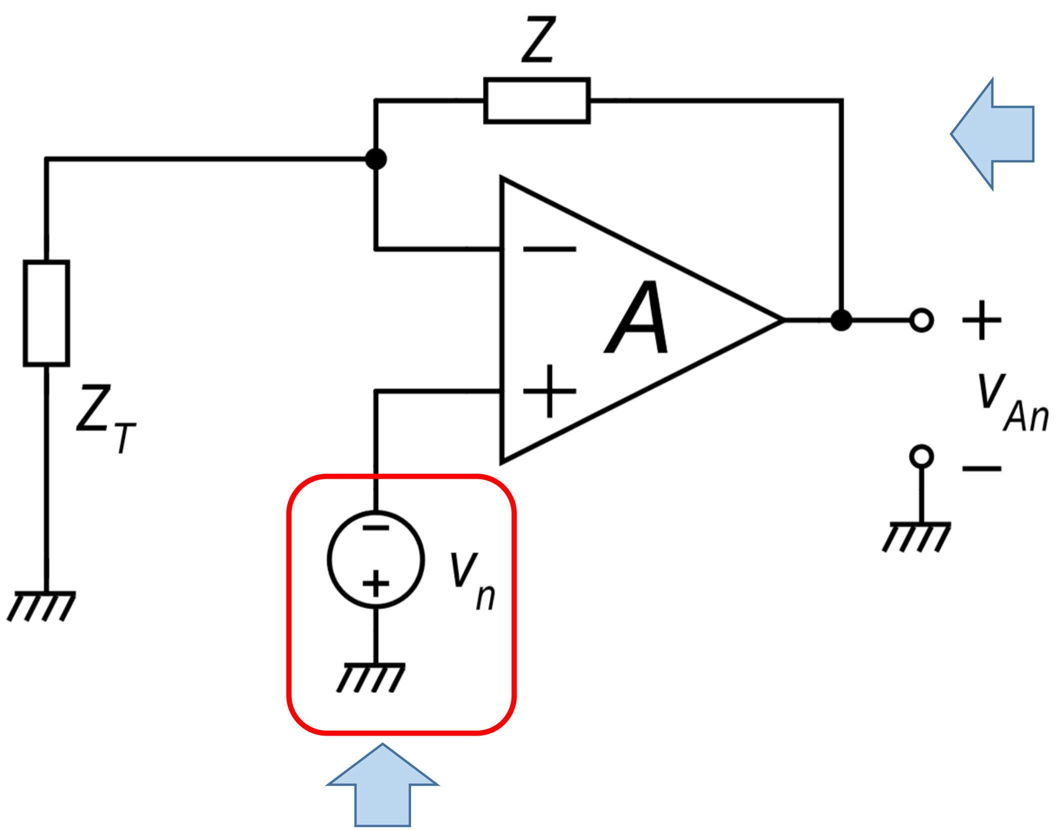
For noise calculations, we will neglect finite gain effects (hypothesis of perfect virtual *ground*)

Effect of i_n and v_{nR}



$$v_{An} = i_n \cdot Z + v_{nR}$$

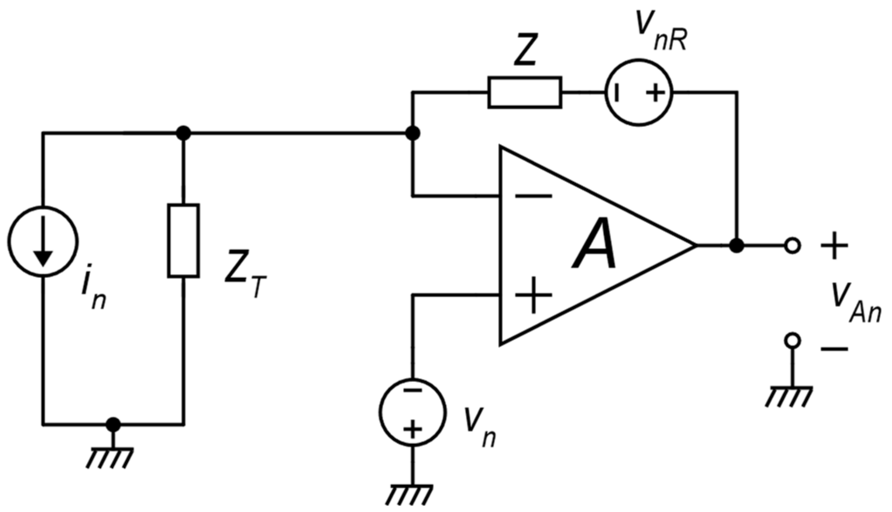
Effect of the voltage source



This is the classical non-inverting amplifier with $-v_n$ as input signal.

$$v_{An} = -v_n\left(1+\frac{Z}{Z_T}\right)$$

Total output noise voltage and input referred noise current



$$v_{An} \cong Zi_n - v_n\left(1+\frac{Z}{Z_T}\right)+v_{nR}$$

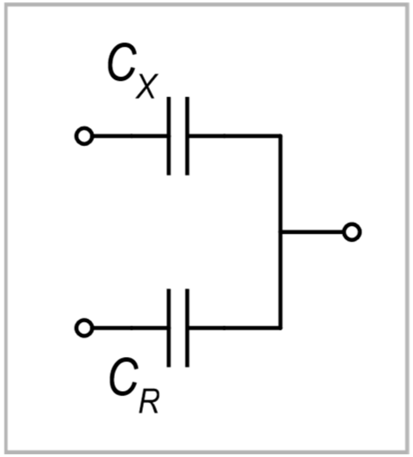
The quantity of interest is the total input-referred current noise: $i_{n\text{-}RTI}$

$$i_{n-RTI}=\frac{v_{An}}{k_{TIA}} \qquad k_{TIA}=\frac{v_A}{i_s}=-Z$$

$$i_{n-RTI} \cong -i_n + v_n\left(\frac{1}{Z}+\frac{1}{Z_T}\right)-\frac{v_{nR}}{Z}$$

$$for\ Z \to \infty : i_{n-rti} = -i_n + \frac{v_n}{Z_T}$$

TIA used to read capacitive sensors



Differential capacitive sensor

X: physical quantity to be sensed

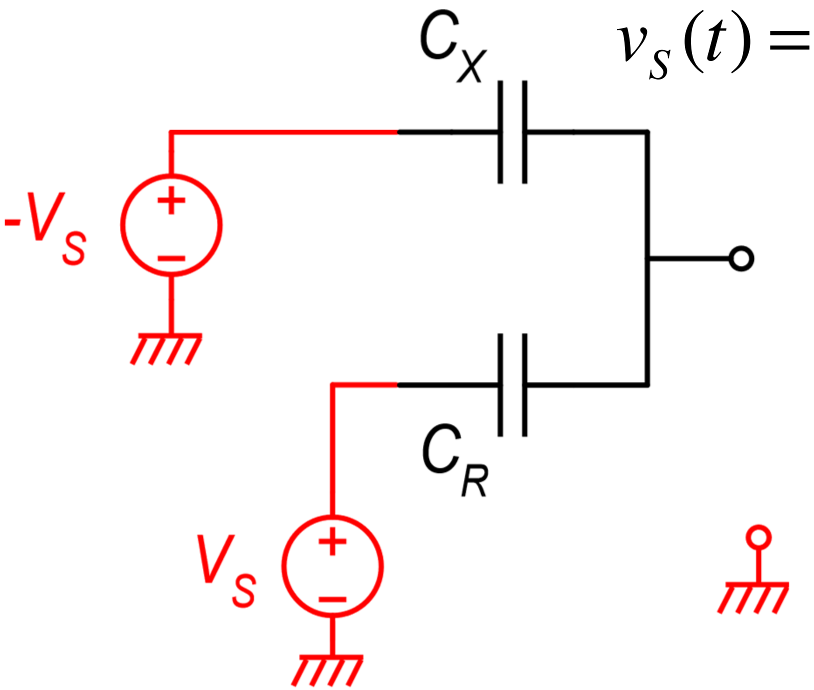
case 1
$$\begin{cases} C_X = C_0 + f(X) \\ C_R = C_0 - f(X) \end{cases}$$
Balanced differential sensor (e.g. MEMS accelerometer)

case 2
$$\begin{cases} C_X = C_0 + f(X) \\ C_R = C_0 \end{cases}$$
Pseudo differential sensor (e.g. pressure sensor)

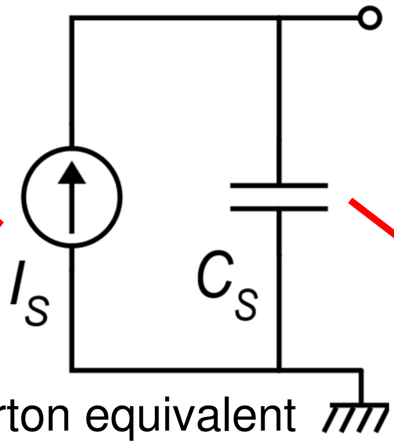
C_0 is typically a large capacitance which does not vary with X, but is widely affected by process spread and often depends on temperature

The interface must read: $\Delta C = C_X - C_R$

Transformation of ΔC into a current



$$v_S(t) = V_{SM}\cos(\omega_S t)$$



Norton equivalent circuit of the sensor

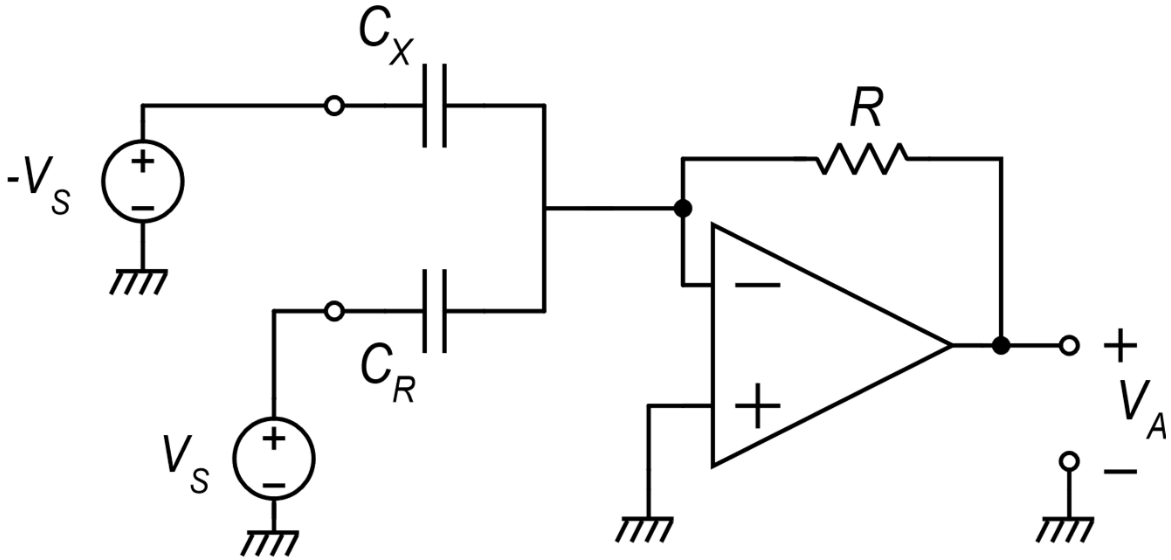
$$C_S = C_X + C_R$$

$$Z_S = \frac{1}{j\omega_S\left(C_X + C_R\right)}$$

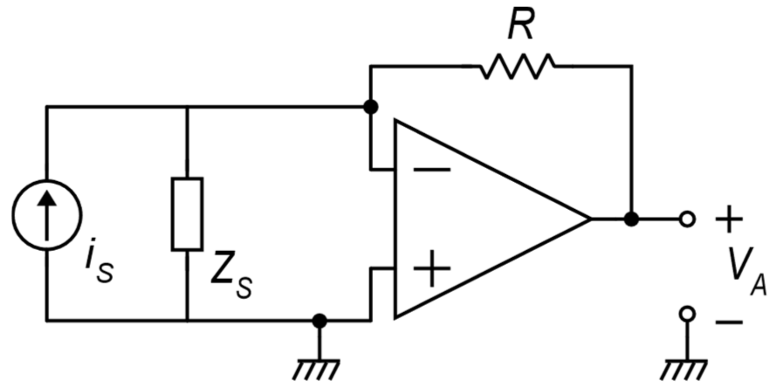
short-circuit current

$$i_S(t) = C_R\frac{dv_S}{dt} - C_X\frac{dv_S}{dt} = \left(C_X - C_R\right)V_{SM}\,\omega_S\sin(\omega_S t)$$

Use of the TIA to read current I_S



Substituting the Norton equivalent circuit to the sensor, we have the usual TIA configuration, with *Z*=*R*



$$v_A(t) = -R \cdot i_S(t) = -\Delta C(t) \cdot R V_{SM} \omega_S \sin(\omega_S t)$$

The useful signal ΔC(t) is modulated by $\sin(\omega_S t)$:

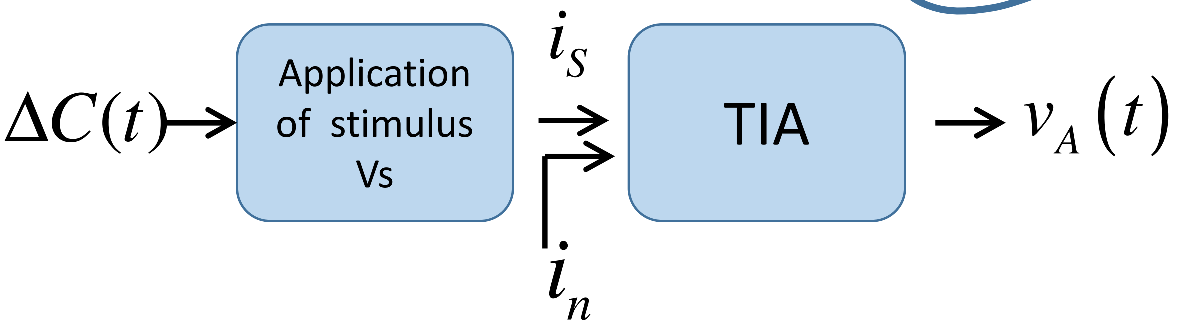
Demodulation is required to extract ΔC(t)

Synchronous demodulation allows detecting also the sign of ΔC(t)

A few general considerations

$$i_S(t) = \Delta C(t) \cdot V_{SM}\,\omega_S \sin(\omega_S t)$$

It is desirable to increase this factor ($V_{SM}\,\omega_S$) to have a larger signal -to-noise ratio



$$\varepsilon_{tot} \cong \left|\frac{Z_{IN}}{Z_T}\right| + \frac{1}{|A(f)|}$$

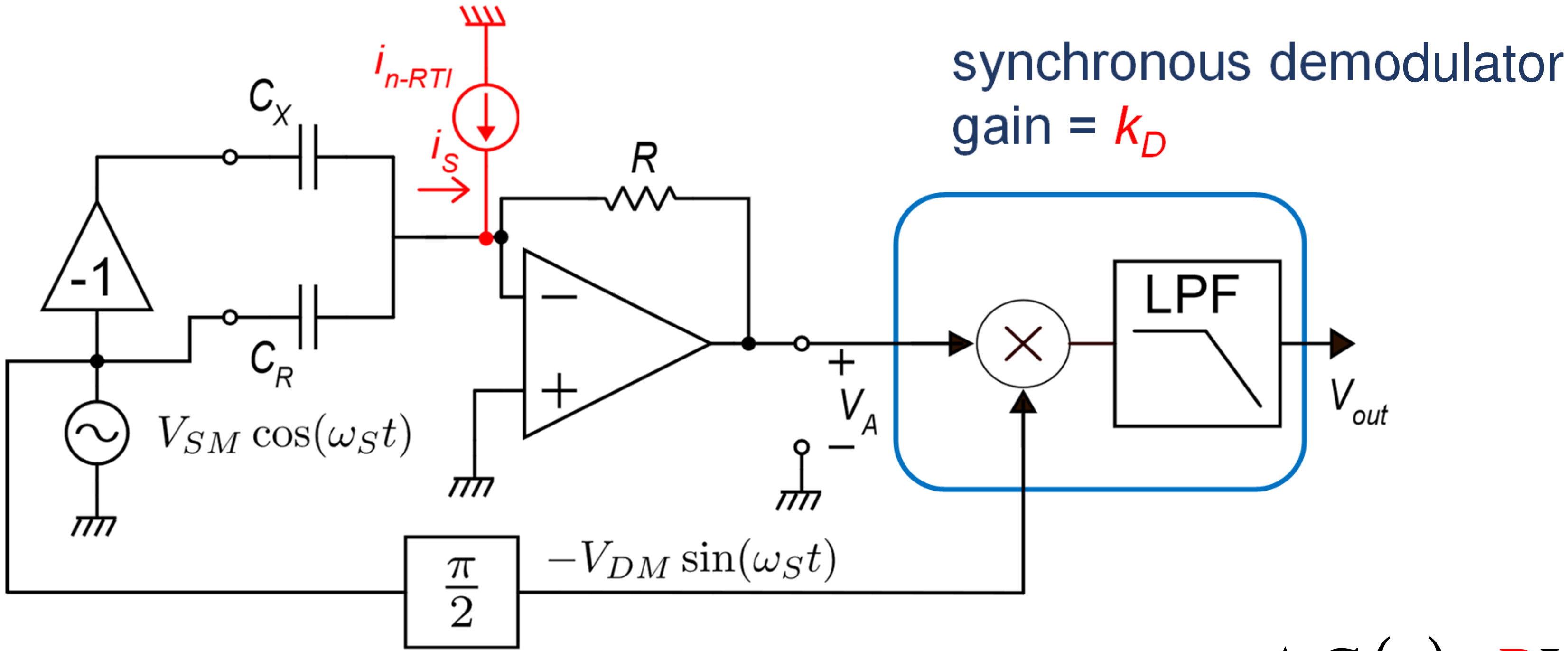
V_{SM} must be as large as possible (limited by V_{dd})

ω_S cannot be too large because the error due to the finite impedance becomes not acceptable.

if ω_S increases:

- $|A(f_S)|$ decreases
- $|Z_{IN}| = R\dfrac{f_S}{f_0}$ increases
- $|Z_T| = \dfrac{1}{\omega_S (C_S + C_A)}$ decreases

Complete interface

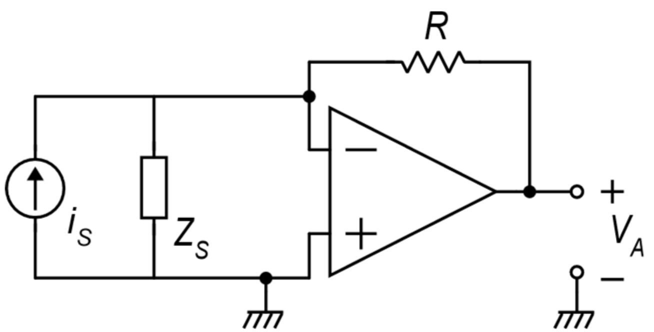


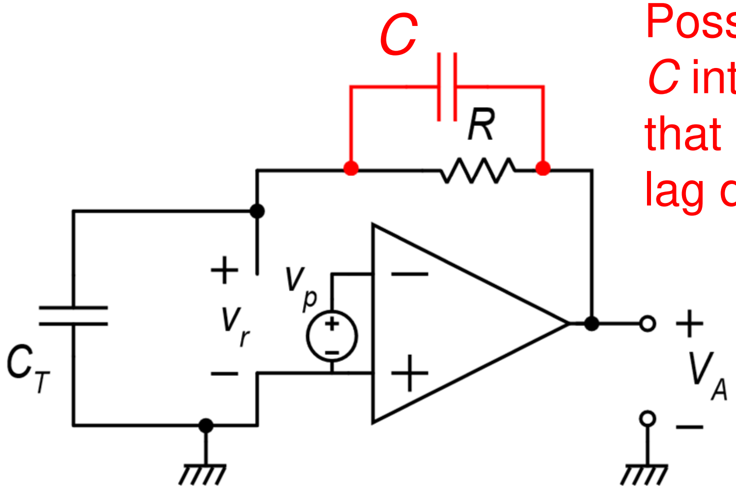
$$v_{out} = \Delta C(t) \cdot R V_{SM} \omega_S k_D$$

$$i_S = \Delta C(t) \cdot V_{SM} \omega_S \sin(\omega_S t)$$

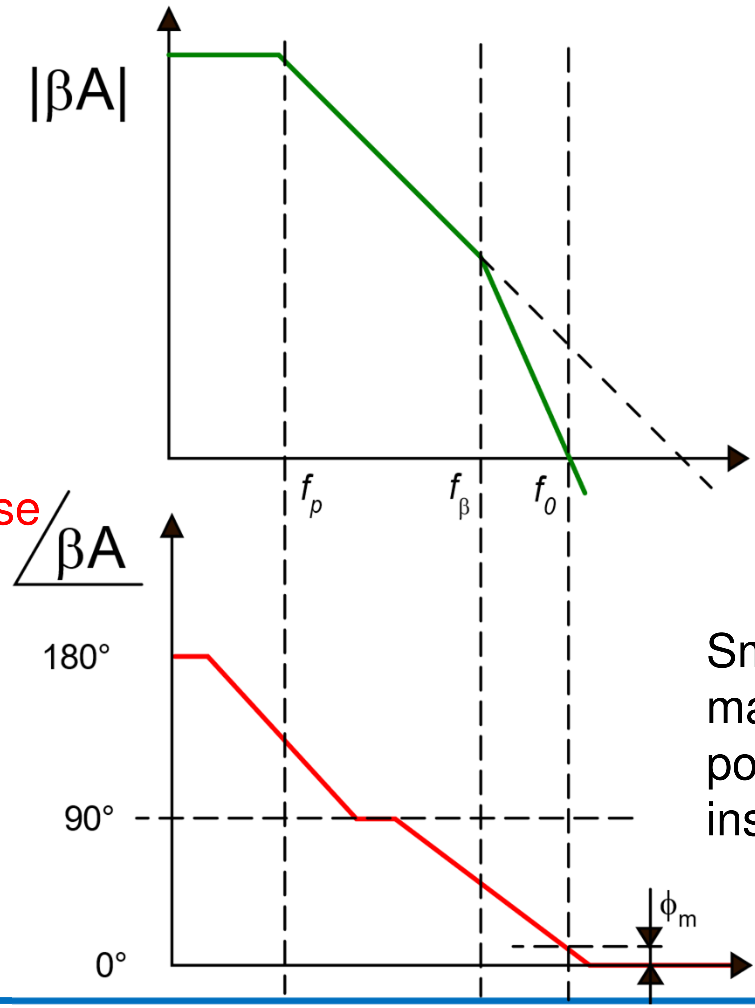
$$v_A = -\Delta C(t) \cdot R V_{SM} \omega_S \sin(\omega_S t)$$ (ideal TIA)

Brief mention to TIA stability issues





Possible solution:
C introduces a zero that reduces the phase lag due to f_β



Small phase margin: potential instability