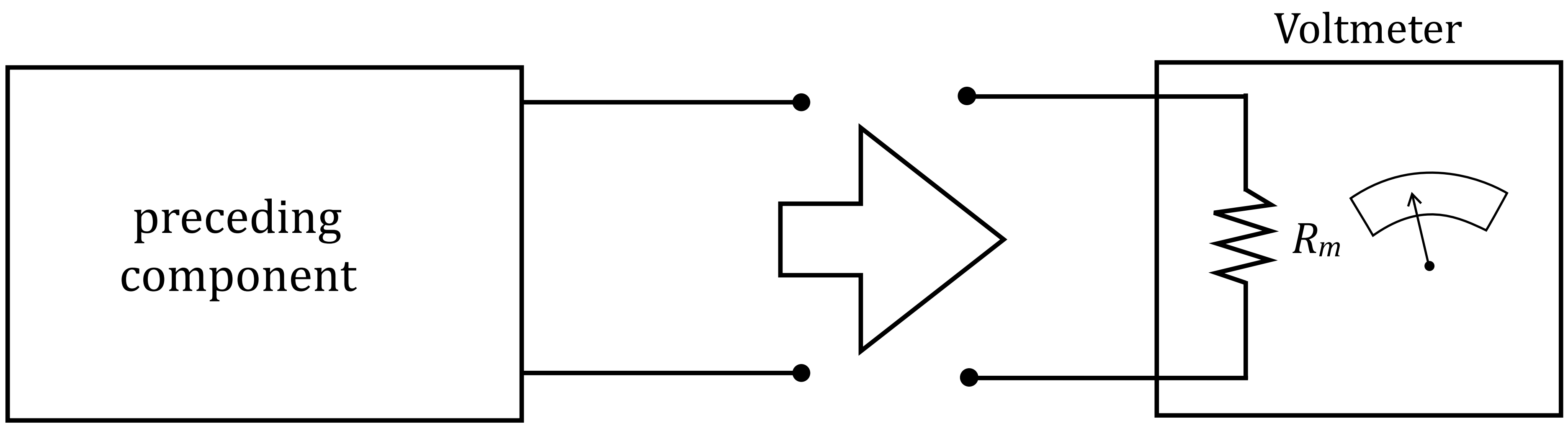
Component Interconnection and Signal Conditioning

ME 473

Professor Sawyer B. Fuller

The loading effect

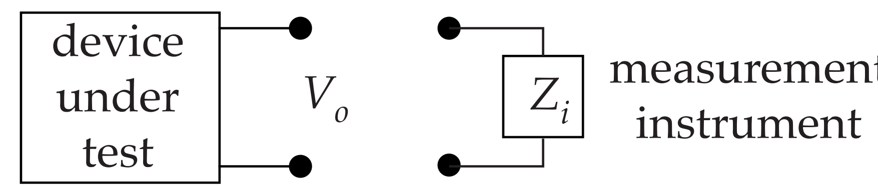


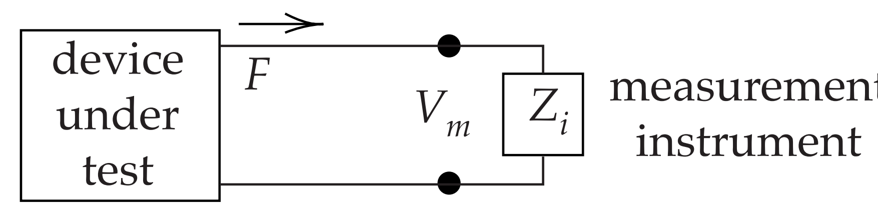
We need to quantify the effect of the effect of interconnecting

- to *match* input to output, e.g. to optimize power transfer efficiency
- to minimize loading for a sensor

Across Variable Measurements

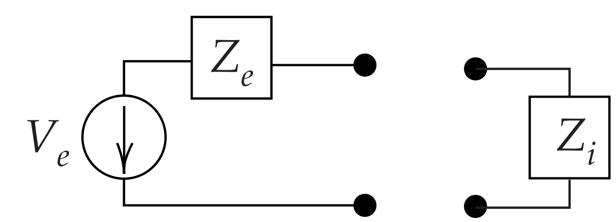
Suppose that we wish to measure an across variable at the output of a "device under test" with a "measurement instrument." The measurement instrument is attached across the terminals of interest. Of course we desired that the measured variable be undisturbed by the connection of the instrument. That is, we want V_m to be as nearly equal to V_o as possible. We say that the measurement instrument should not "load" the device under test.





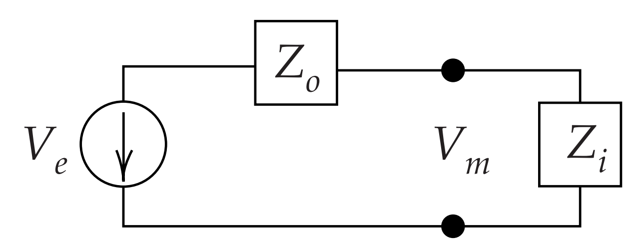
The *output impedance* of the device under test is the equivalent impedance defined by its Thevenin model $Z_o = Z_e$ for the unloaded output terminals.

Similarly, the *input impedance* Z_i of the measurement instrument is the Thevenin equivalent impedance defined for its input terminals.



Connecting the Thevenin model for the device under test to the input impedance of the measurement instrument we have the network at the right.

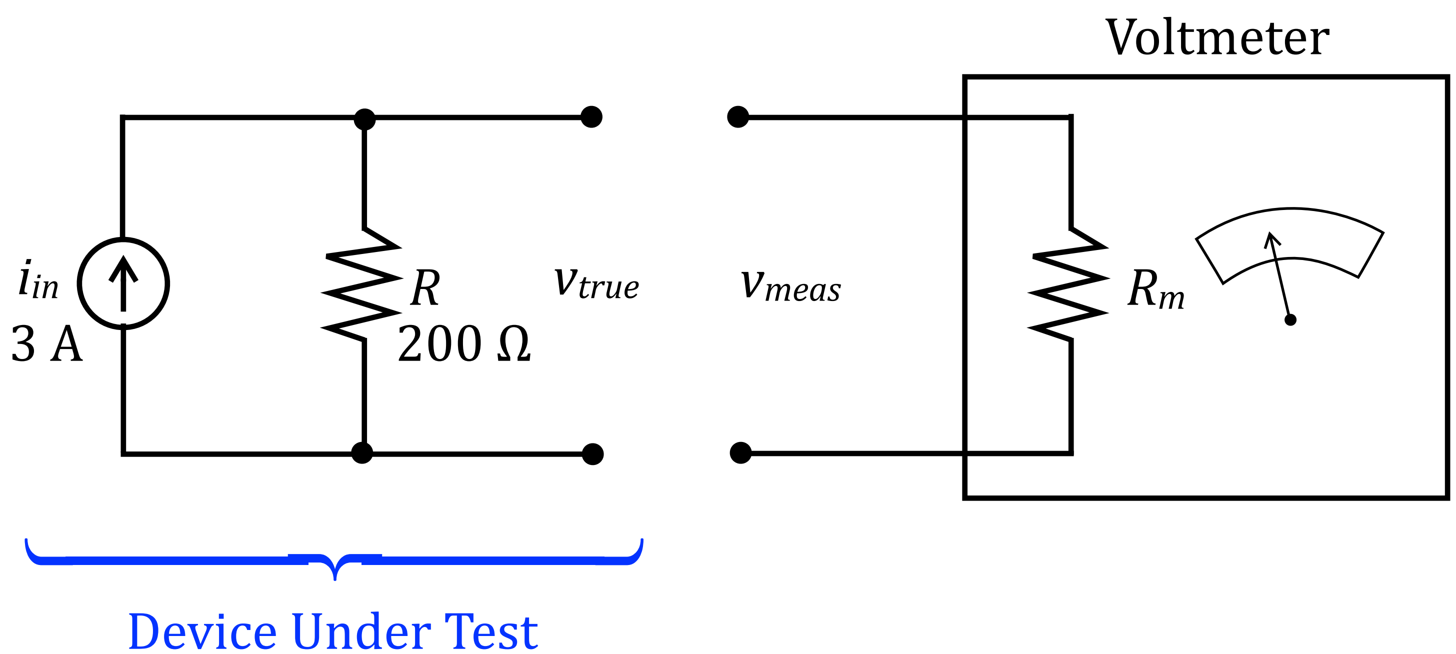
The Thevenin equivalent across variable source is by definition equal to V_o, the value that we wish to measure. Applying the across variable divider rule: $\dfrac{V_m(s)}{V_o(s)} = \dfrac{1}{1 + Z_o/Z_i}$.



Since we desire that the ratio approach unity, the input impedance of the measurement instrument must be large in comparison with the output impedance of the device under test: $Z_i >> Z_o$

Use Thevenin equivalent circuit for across variable measurements so you can use the across variable divider rule.

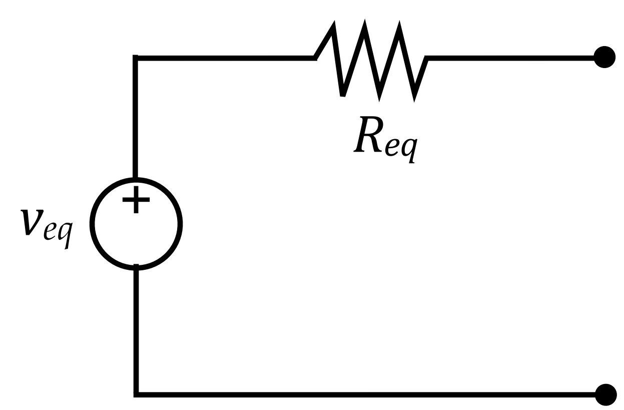
Loading: Example #1



Voltmeter model: purely resistive

What is the smallest R_m such that the voltage v_{meas} displayed on the voltmeter will be within 1 Volt of v_{true}?

Thevenin-equivalent model of Device Under Test:



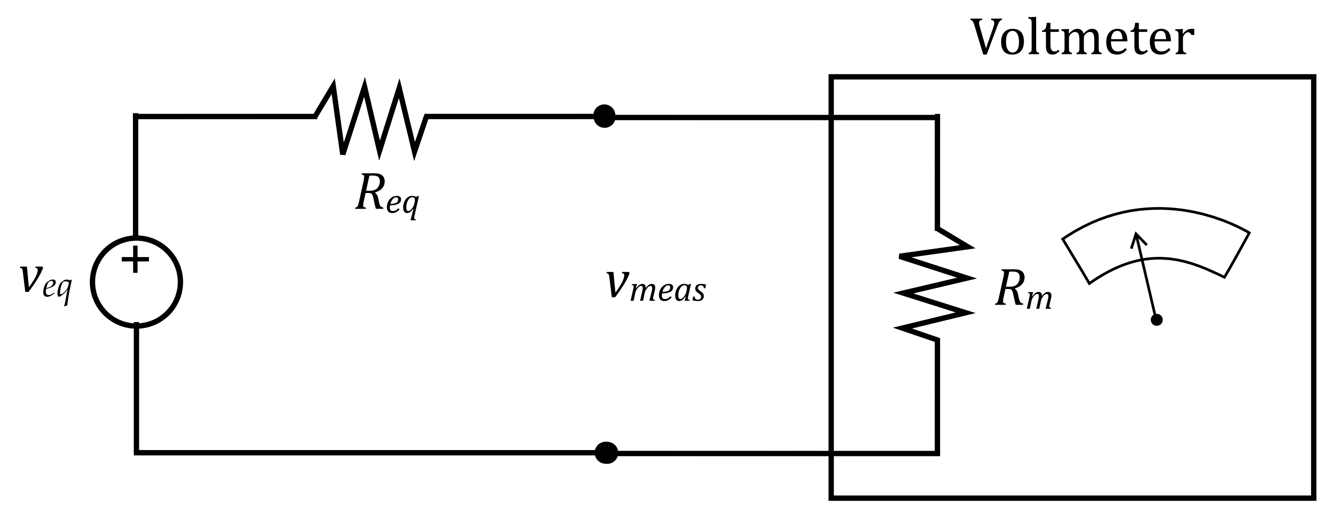
$$v_{eq} = i_{in} R = v_{true} = 600 \text{ V}$$

$$R_{eq} = R = 200 \text{ Ohms}$$

Loading: Example #1

$v_{eq} = i_{in} R$
$= v_{true}$
$= 600$ V

$R_{eq} = R$
$= 200$ Ohms



What is the smallest R_m such that the voltage v_{meas} displayed on the voltmeter will be within 1 Volt of v_{true}?

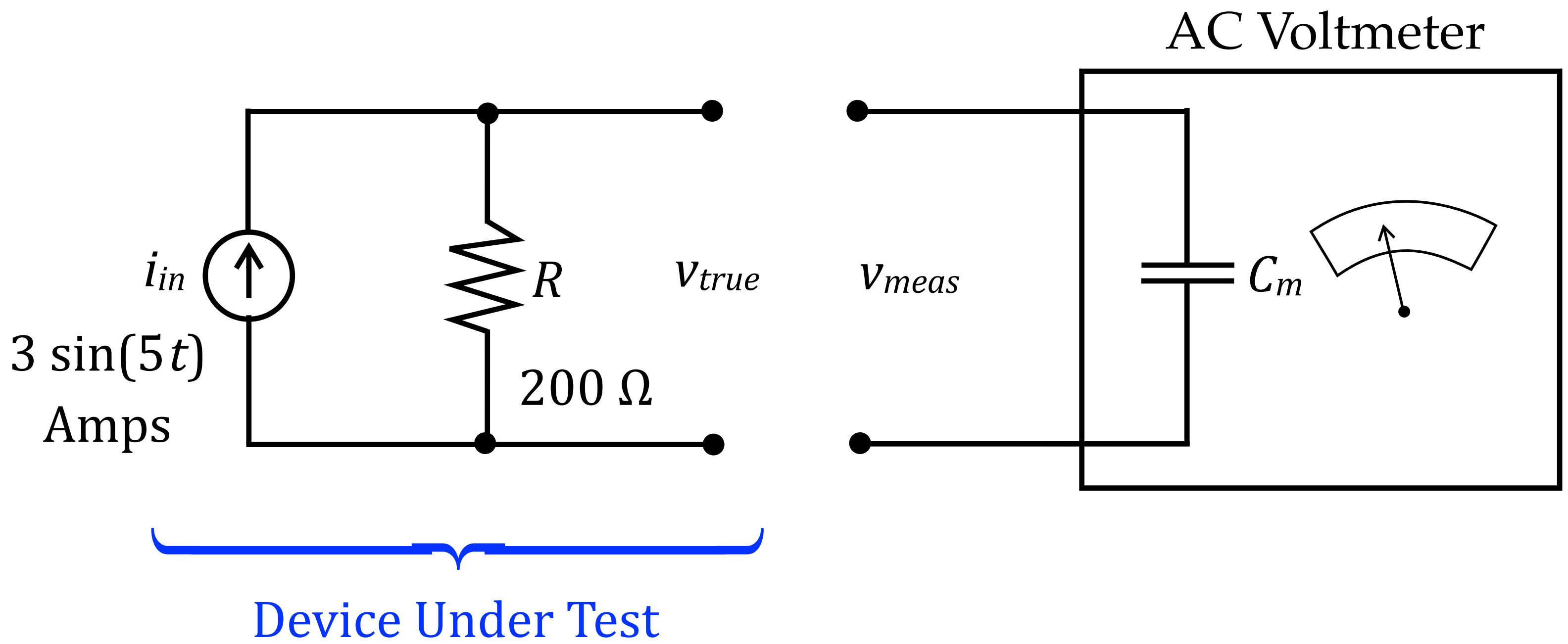
Voltage divider formula yields: $$v_{meas} = v_{eq} \frac{R_m}{R_m + R_{eq}} = v_{true} \frac{R_m}{R_m + R_{eq}} = 600\text{ V} \frac{R_m}{R_m + 200\,\Omega}$$

Required: $599\text{ V} \leq v_{meas} \leq 601\text{ V}$

The smallest satisfactory R_m satisfies

$$599\text{ V} = 600\text{ V} \frac{R_m}{R_m + 200\,\Omega} \Rightarrow R_m = 119{,}800\,\Omega$$

Loading: Example #2

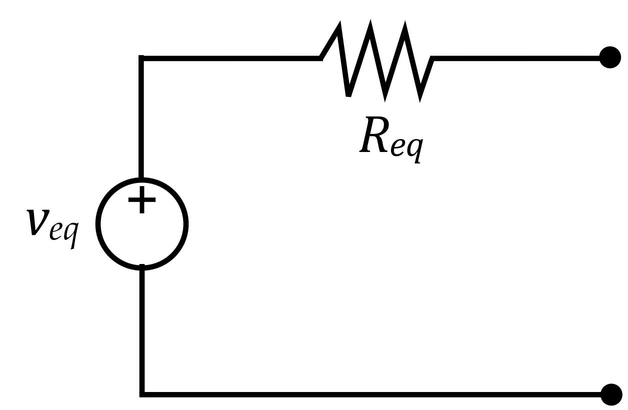


AC voltmeter model: purely capacitive

$Z_m(s) = 1/(C_m s)$

What is the largest C_m such that, in the steady state, the amplitude displayed on the voltmeter will be within 1 percent of the amplitude of v_{true}?

Thevenin-equivalent model of Device Under Test:



$$
\begin{aligned}
v_{eq} &= i_{in} R \\
&= v_{true} \\
&= 600 \sin(5t) \text{ V}
\end{aligned}
$$

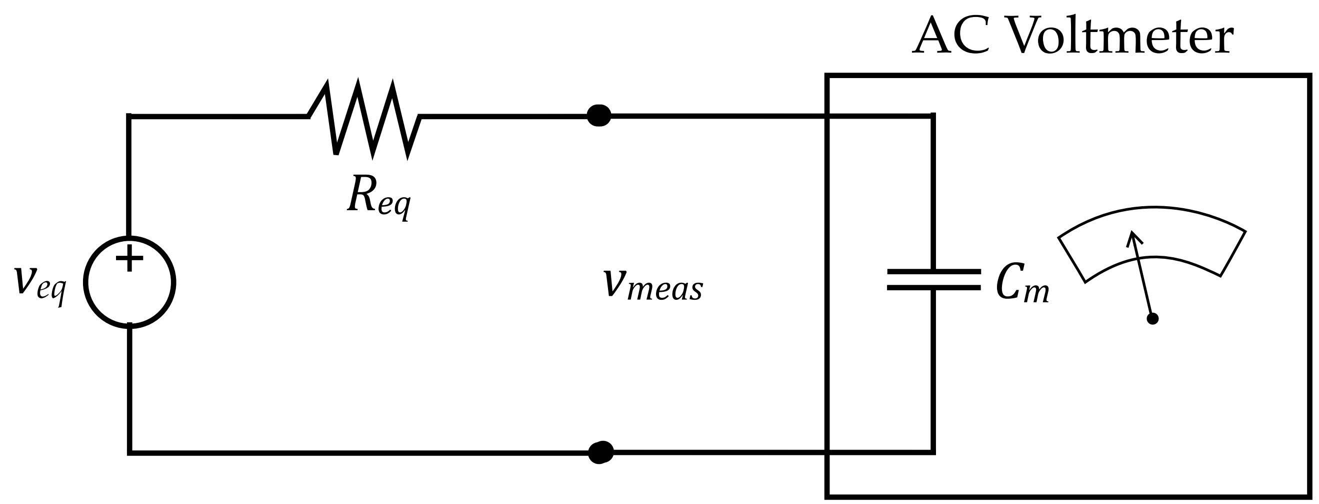
$$
\begin{aligned}
R_{eq} &= R \\
&= 200 \text{ Ohms}
\end{aligned}
$$

Loading: Example #2

Thevenin-equivalent-based model of complete system:

$$v_{eq} = i_{in} R = v_{true} = 600 \sin(5t) \text{ V}$$

$$R_{eq} = R = 200 \text{ Ohms}$$



What is the largest C_m such that, in the steady state, the amplitude displayed on the voltmeter will be within 1 percent of the amplitude of v_{true}?

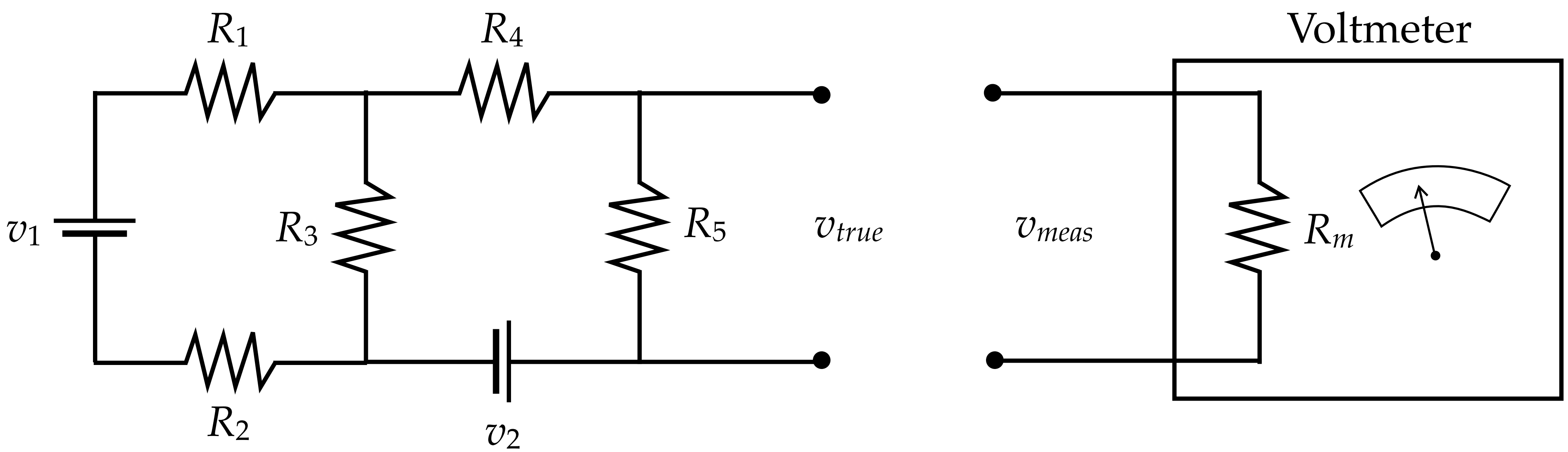
Voltage divider formula yields: $$V_{meas}(s) = V_{eq}(s) \frac{Z_m(s)}{Z_m(s) + R_{eq}} = V_{true}(s) \left(\frac{\frac{1}{C_m s}}{\frac{1}{C_m s} + R} \right) = V_{true}(s) \left(\frac{1}{RC_m s + 1} \right)$$

Required: $$\frac{99}{100} \leq \left| \frac{1}{RC_m s + 1} \right| \Bigg|_{\substack{R = 200\,\Omega \\ s = j5 \text{ sec}^{-1}}} \leq \frac{101}{100}$$

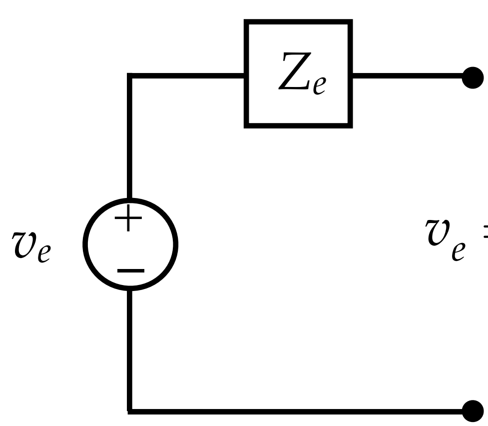
The largest satisfactory C_m value satisfies

$$\frac{99}{100} = \frac{1}{\sqrt{[(200) C_m (5)]^2 + 1}} \quad \Rightarrow \quad C_m = \frac{1}{1000} \sqrt{\left(\frac{100}{99} \right)^2 - 1} \approx 0.1425 \times 10^{-3} \text{ F}$$

Loading: Example #3



When we measure v_{true} with the voltmeter, how much does the measured voltage, v_{meas}, differ from v_{true}?



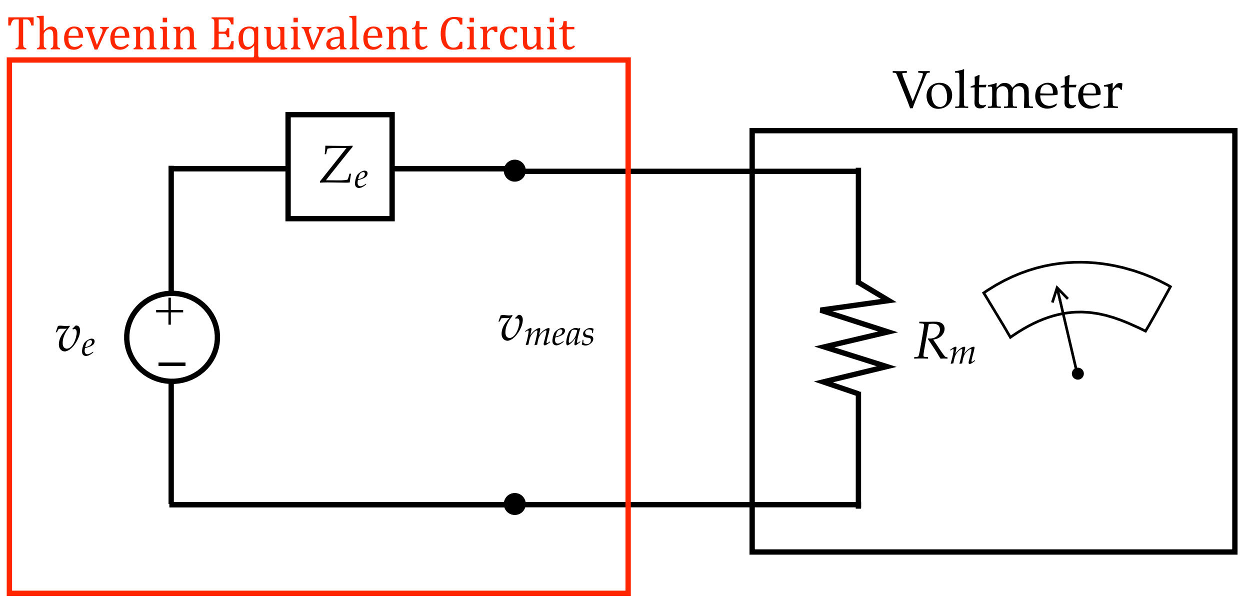
Thevenin Equivalent

$$v_e = v_{true} = \frac{\left[v_1 R_3 - v_2(R_1+R_2+R_3)\right]R_5}{(R_1+R_2)R_3+(R_1+R_2+R_3)(R_4+R_5)}$$

hard to get

$$Z_e = \frac{R_5\left[R_4 + \dfrac{R_3(R_1+R_2)}{(R_3+R_1+R_2)}\right]}{R_5 + \left[R_4 + \dfrac{R_3(R_1+R_2)}{(R_3+R_1+R_2)}\right]}$$

easy to get



When we measure v_{true} with the voltmeter, how much does the measured voltage, v_{meas}, differ from v_{true}?

$$\left.\begin{aligned} v_{meas} &= v_e \frac{R_m}{R_m + Z_e} = v_e \frac{1}{1 + \dfrac{Z_e}{R_m}} \\ v_{true} &= v_e \end{aligned}\right\} \Rightarrow \boxed{v_{meas} = v_{true} \frac{1}{1 + \dfrac{Z_e}{R_m}}} \Rightarrow$$

For $v_{meas} \approx v_{true}$ must have $R_m \gg Z_e$

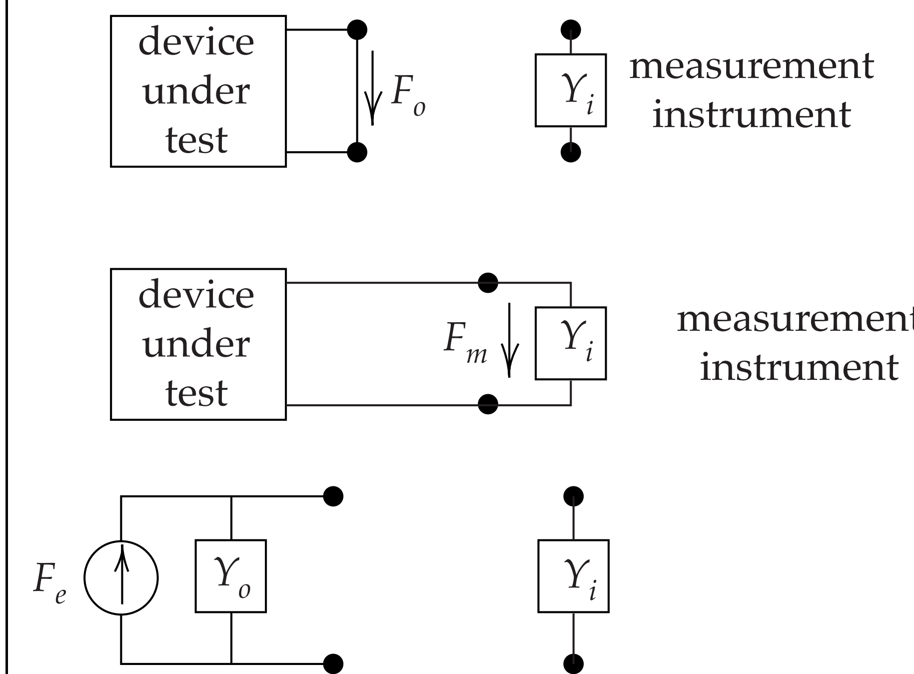
Knowledge that $v_e = \dfrac{[v_1 R_3 - v_2(R_1 + R_2 + R_3)]R_5}{(R_1 + R_2)R_3 + (R_1 + R_2 + R_3)(R_4 + R_5)}$ not required!

Through Variable Measurements

Alternately, suppose that we wish to measure a through variable in a device under test with a measurement instrument. In this case, the variable of interest flows through the measurement instrument. We desired that the measured variable be undisturbed by the connection of the instrument. That is, we want F_m to be as nearly equal to F_o as possible.

The *output admittance* of the device under test is the equivalent admittance defined by its Norton's model $Y_o = 1/Z_e$ for the unloaded output terminals.

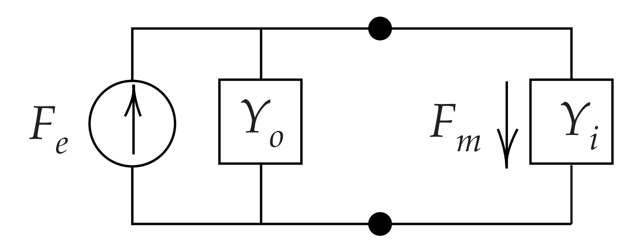
Similarly, the *input admittance* Y_i of the measurement instrument is the Norton equivalent admittance defined for its input terminals.



Connecting the Norton model for the device under test to the input admittance of the measurement instrument we have the network at the right.

The Norton equivalent through variable source is by definition equal to F_o, the value that we wish to measure. Applying the through variable divider rule: $\dfrac{F_m(s)}{F_o(s)} = \dfrac{1}{1+Y_o/Y_i}$.

Since we desire that the ratio approach unity, the input admittance of the measurement instrument must be large in comparison with the output admittance of the device under test: $Y_i >> Y_o$

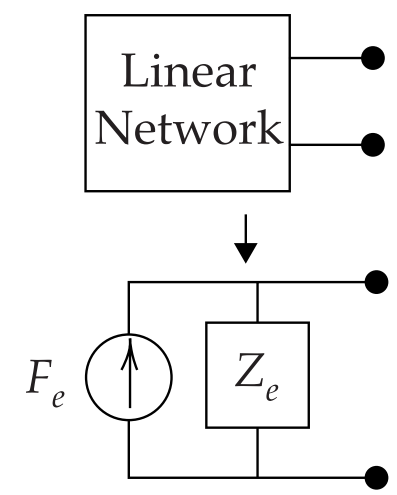


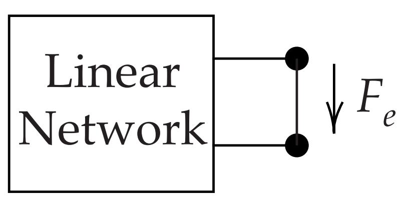
Norton's Theorem

A linear two-terminal network is equivalent to a through variable source F_e in *parallel* with an equivalent impedance Z_e, where

$Z_e =$ the impedance of the network with all sources set equal to zero, and

$F_e =$ a *through variable source* equal to the through variable that would flow through the *short* circuited terminals of the network.





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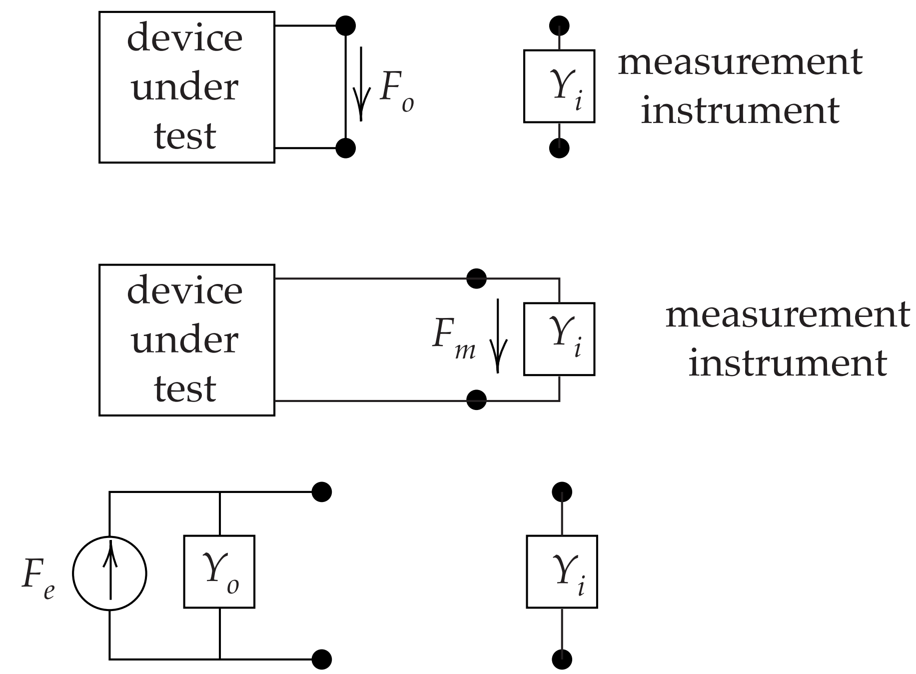

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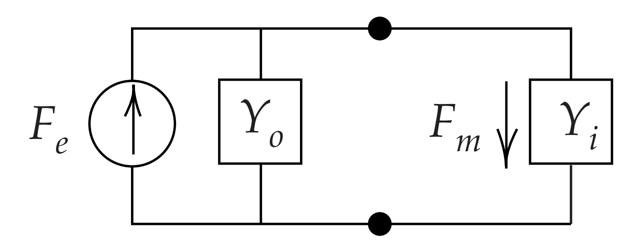
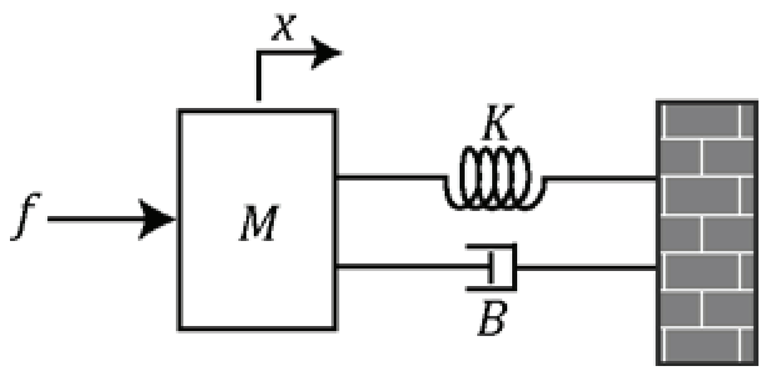
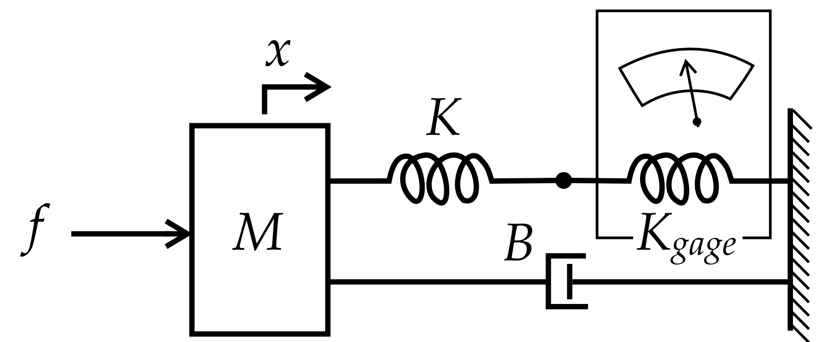


Example 4





Consider the mass-spring-damper system shown above. A spring-based force gage, with spring constant K_{gage}, is to be inserted between the spring K and the wall, to measure the force in K in response to the applied force f. With f_{true} representing the force in K without the gage present, and f_{gage} representing the force in K with the gage present, f_{true} and f_{gage} satisfy

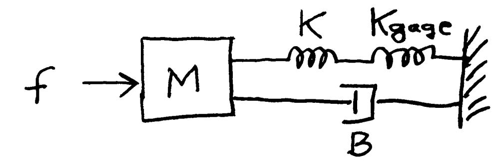
$$\frac{F_{gage}(s)}{F_{true}(s)} = \frac{n(s)}{d(s)}$$

where $n(s)$ and $d(s)$ are polynomials in the Laplace variable s.

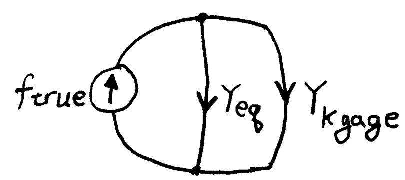
(a) Determine $n(s)$ and $d(s)$.

(b) When f (the applied force) is constant, what is the relationship, in the steady state, between f_{gage} and f_{true}?

Solution using Norton Equivalent of Device Under Test



Norton-equivalent-based model of complete system:



Mass: $F_M(s) = Ms\, V_M(s)$

$\Rightarrow Y_M(s) = Ms$

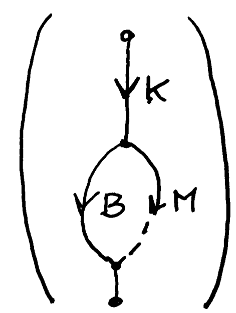
Spring: $F_K(s) = K \dfrac{V_K(s)}{s}$

$\Rightarrow Y_K(s) = \dfrac{K}{s}$

Damper: $F_B(s) = B\, V_B(s)$

$\Rightarrow Y_B(s) = B$

Y_{eq} = admittance of



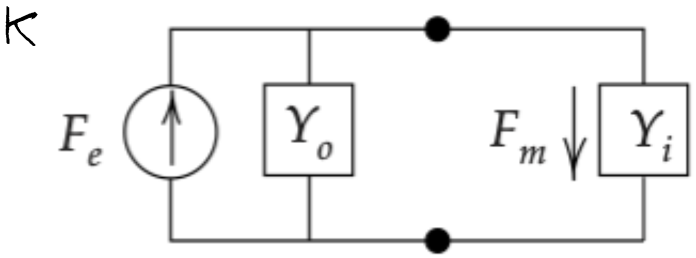
Admittances "sum in parallel", so

$$Y_{eq} = \frac{(Y_B + Y_M)\,Y_K}{(Y_B + Y_M) + Y_K} = \frac{(B + Ms)\,\frac{K}{s}}{(B + Ms) + \frac{K}{s}} = \frac{(Ms+B)K}{Ms^2 + Bs + K}$$

Through variable divider principal then gives

$$\frac{F_{gage}(s)}{F_{true}(s)} = \frac{Y_{gage}}{Y_{eq} + Y_{gage}} = \frac{\frac{K_{gage}}{s}}{\frac{(Ms+B)K}{Ms^2+Bs+K} + \frac{K_{gage}}{s}}$$

$$= \boxed{\frac{(Ms^2+Bs+K)\,K_{gage}}{s(Ms+B)K + (Ms^2+Bs+K)\,K_{gage}}}$$



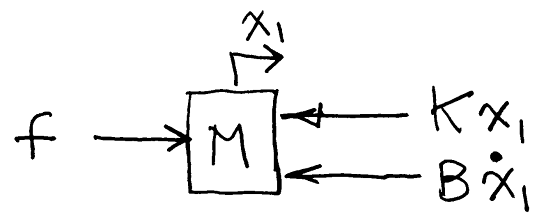
$$\frac{F_m(s)}{F_e(s)} = \frac{1}{1+Y_o/Y_i}.$$

If f is constant then, in the steady state,

$$\frac{f_{gage}}{f_{true}} = \frac{F_{gage}(0)}{F_{true}(0)} = 1$$

Solution not using Norton Equivalent of Device Under Test

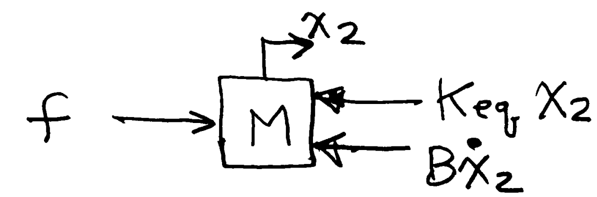
Without Gage



$$f - B\dot{x}_1 - Kx_1 = M\ddot{x}_1$$

$$\Rightarrow F(s) = (Ms^2 + Bs + K)\,X_1(s)$$

With Gage



$$K_{eq} = \frac{K K_{gage}}{K + K_{gage}}$$

$$f - B\dot{x}_2 - K_{eq}x_2 = M\ddot{x}_2$$

$$\Rightarrow F(s) = (Ms^2 + Bs + K_{eq})\,X_2(s)$$

When the same $f(t)$ is applied to the two systems, the resulting $x_1(t)$ and $x_2(t)$ will satisfy

$$(Ms^2 + Bs + K)\,X_1(s) = (Ms^2 + Bs + K_{eq})\,X_2(s) \qquad (1)$$

Furthermore,

$$f_{true} = Kx_1 \Rightarrow F_{true}(s) = K\,X_1(s) \qquad (2)$$

$$f_{gage} = K_{eq}x_2 \Rightarrow F_{gage}(s) = K_{eq}\,X_2(s) \qquad (3)$$

Solution not using Norton Equivalent of Device Under Test

From (1), using (2) and (3),

$$\frac{1}{K}\left(Ms^2+Bs+K\right)\underbrace{K\,X_1(s)}_{F_{true}(s)} = \frac{1}{K_{eq}}\left(Ms^2+Bs+K_{eq}\right)\underbrace{K_{eq}\,X_2(s)}_{F_{gage}(s)}$$

$$\Rightarrow \quad \frac{F_{gage}(s)}{F_{true}(s)} = \frac{K_{eq}}{K}\;\frac{Ms^2+Bs+K}{Ms^2+Bs+K_{eq}}$$

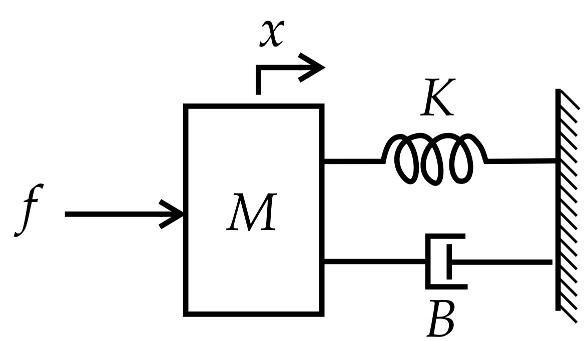
$$= \frac{\dfrac{\cancel{K}K_{gage}}{\cancel{K+K_{gage}}}}{\cancel{K}}\;\frac{Ms^2+Bs+K}{Ms^2+Bs+\dfrac{KK_{gage}}{K+K_{gage}}}\;\frac{\dfrac{\cancel{K+K_{gage}}}{1}}{\dfrac{K+K_{gage}}{1}}$$

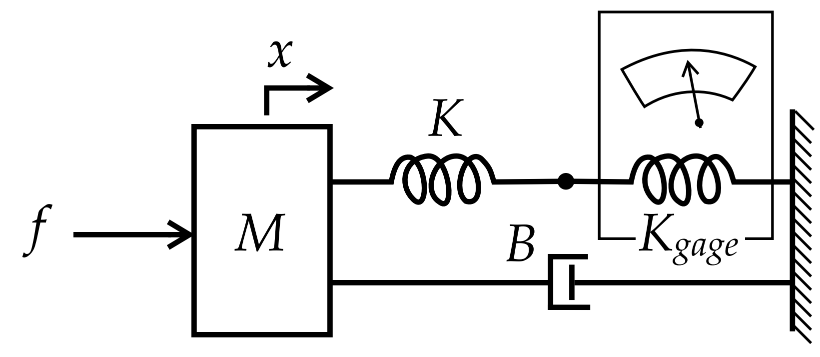
$$= \frac{K_{gage}}{1}\;\frac{Ms^2+Bs+K}{\left(Ms^2+Bs\right)\left(K+K_{gage}\right)+KK_{gage}}$$

$$= \frac{K_{gage}}{1}\;\frac{Ms^2+Bs+K}{s(Ms+B)K+\left(Ms^2+Bs+K\right)K_{gage}}$$

just as derived above, much more simply, using the Norton-equivalent-based approach.

Same Example: Closer Look at the *Frequency Dependence* of the Loading Effect





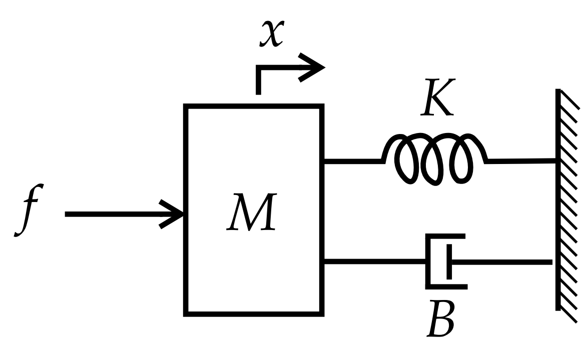
When the force in the spring is measured with a spring-based force gage, how much does the measured force, f_{gage}, differ from the true force, f_{true}, that would appear in the spring with no force gage present?

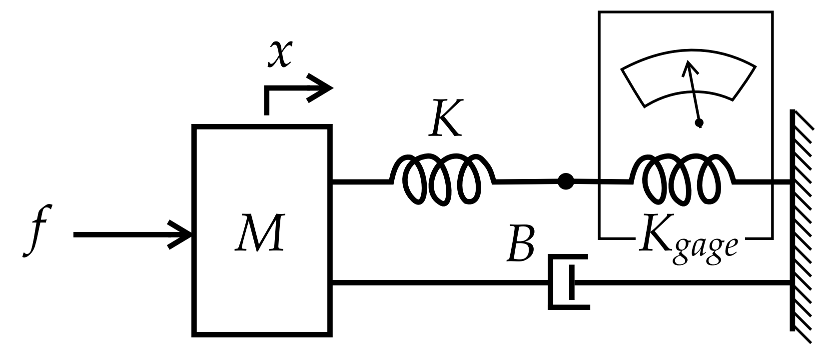
We have shown that:

$$\frac{F_{gage}(s)}{F_{true}(s)} = \frac{\left(Ms^2 + Bs + K\right)K_{gage}}{s(Ms+B)K + (Ms^2+Bs+K)K_{gage}}$$

A transfer function?

Same Example: Closer Look at the *Frequency Dependence* of the Loading Effect





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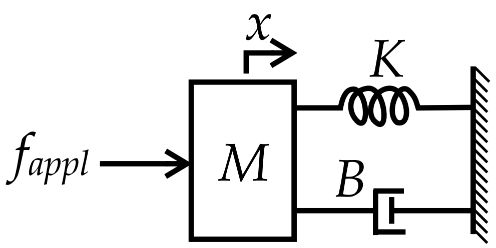
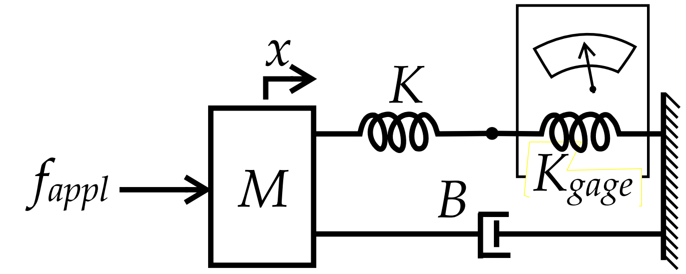
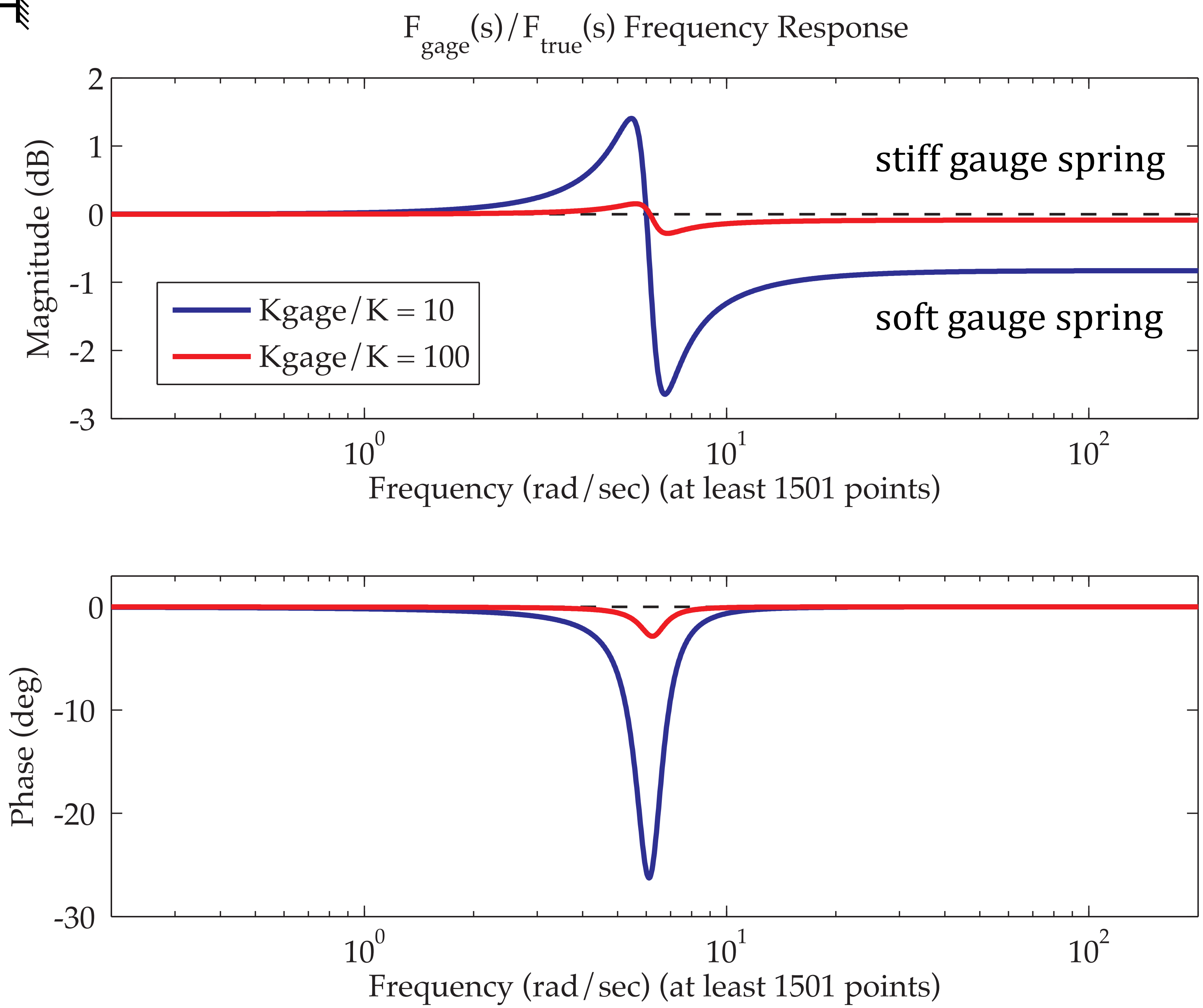






Zero
loading effect
at
low frequencies

Physical explanation
At steady-state, the force is supported by both springs: it passes "through" them

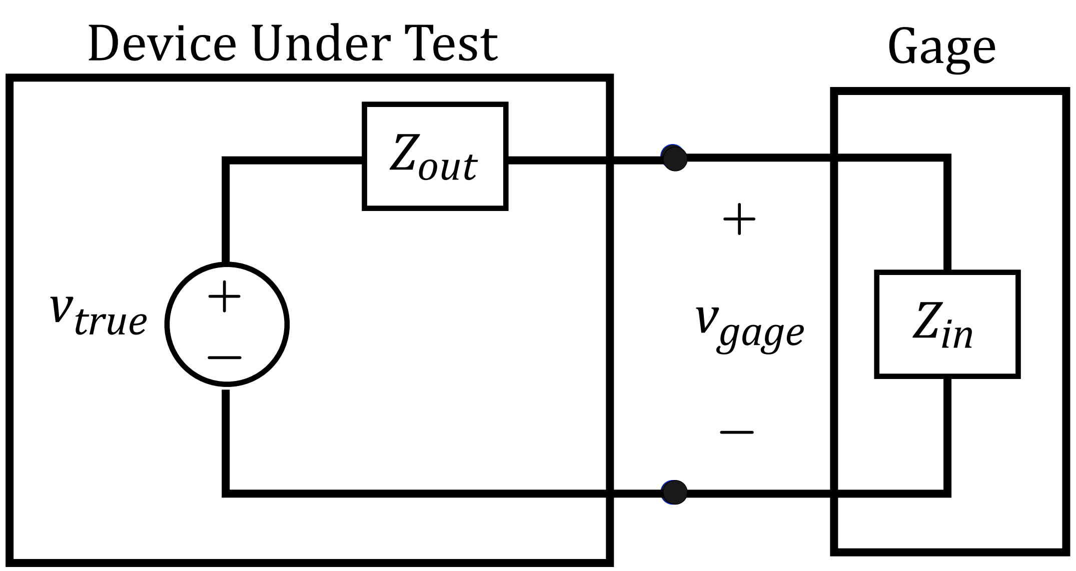
Loading effect
at
high frequencies

$M = 1 \qquad K = (2\pi)^2 \qquad B = 2\,(0.1)\,(2\pi)$

$$\frac{F_{gage}(s)}{F_{true}(s)} = \frac{\left(Ms^2+Bs+K\right)K_{gage}}{s(Ms+B)K+(Ms^2+Bs+K)K_{gage}} \quad \Rightarrow \quad \lim_{s\to j\infty}\frac{F_{gage}(s)}{F_{true}(s)} = \lim_{s\to j\infty}\frac{\left(Ms^2\right)K_{gage}}{s(Ms)K+(Ms^2)K_{gage}} = \frac{K_{gage}}{K+K_{gage}}$$

The Loading Effect: The Big Picture

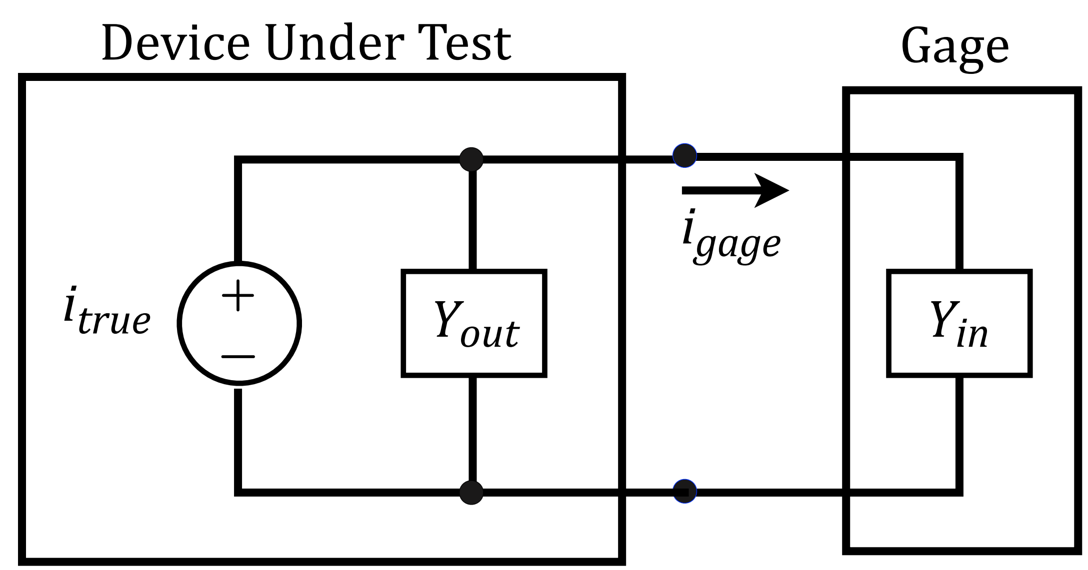
Across Variable Measurement Case



$$\frac{v_{gage}}{v_{true}} = \frac{Z_{in}}{Z_{in} + Z_{out}} = \frac{1}{1 + \frac{Z_{out}}{Z_{in}}}$$

Reduce loading effect by
increasing Z_{in} or *decreasing Z_{out}* or both

Through Variable Measurement Case

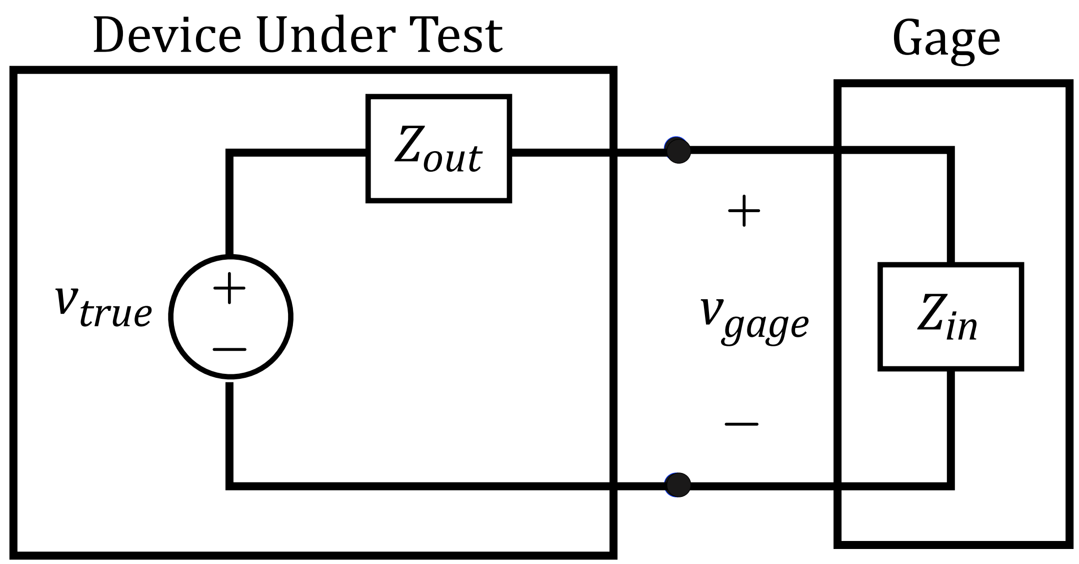


$$\frac{i_{gage}}{i_{true}} = \frac{Y_{in}}{Y_{in} + Y_{out}} = \frac{1}{1 + \frac{Y_{out}}{Y_{in}}}$$

Reduce loading effect by
increasing Y_{in} or *decreasing Y_{out}* or both

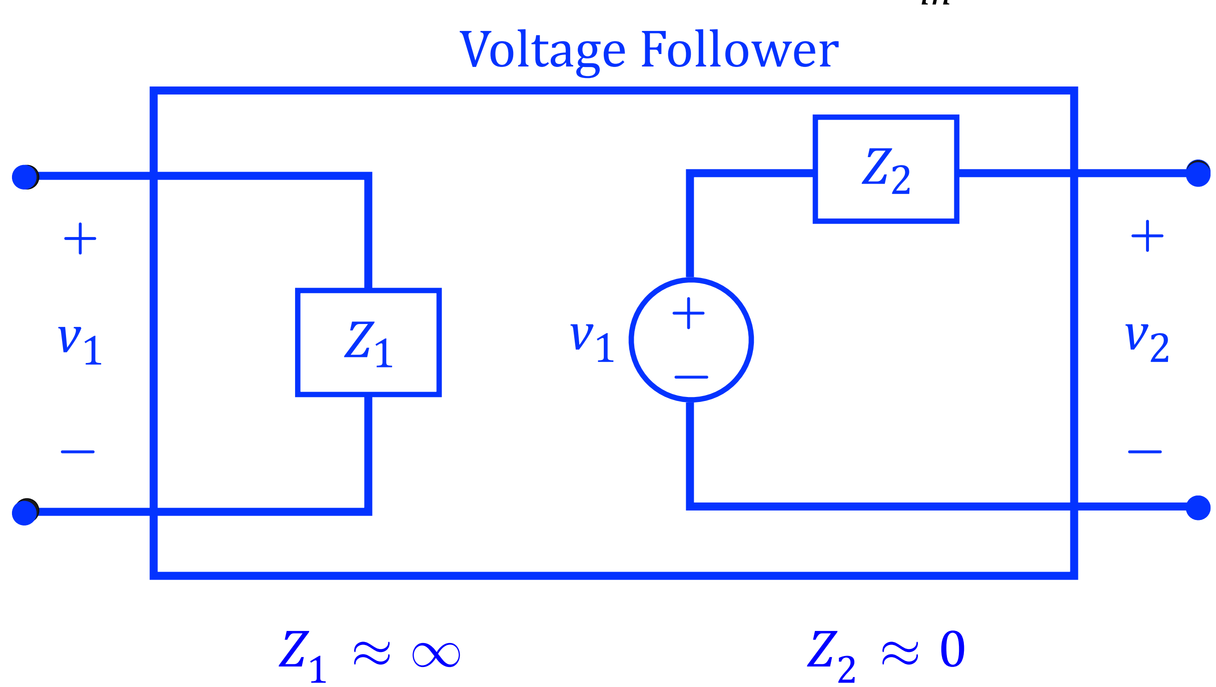
A Motivation for the Voltage Follower

Across Variable Measurement Case



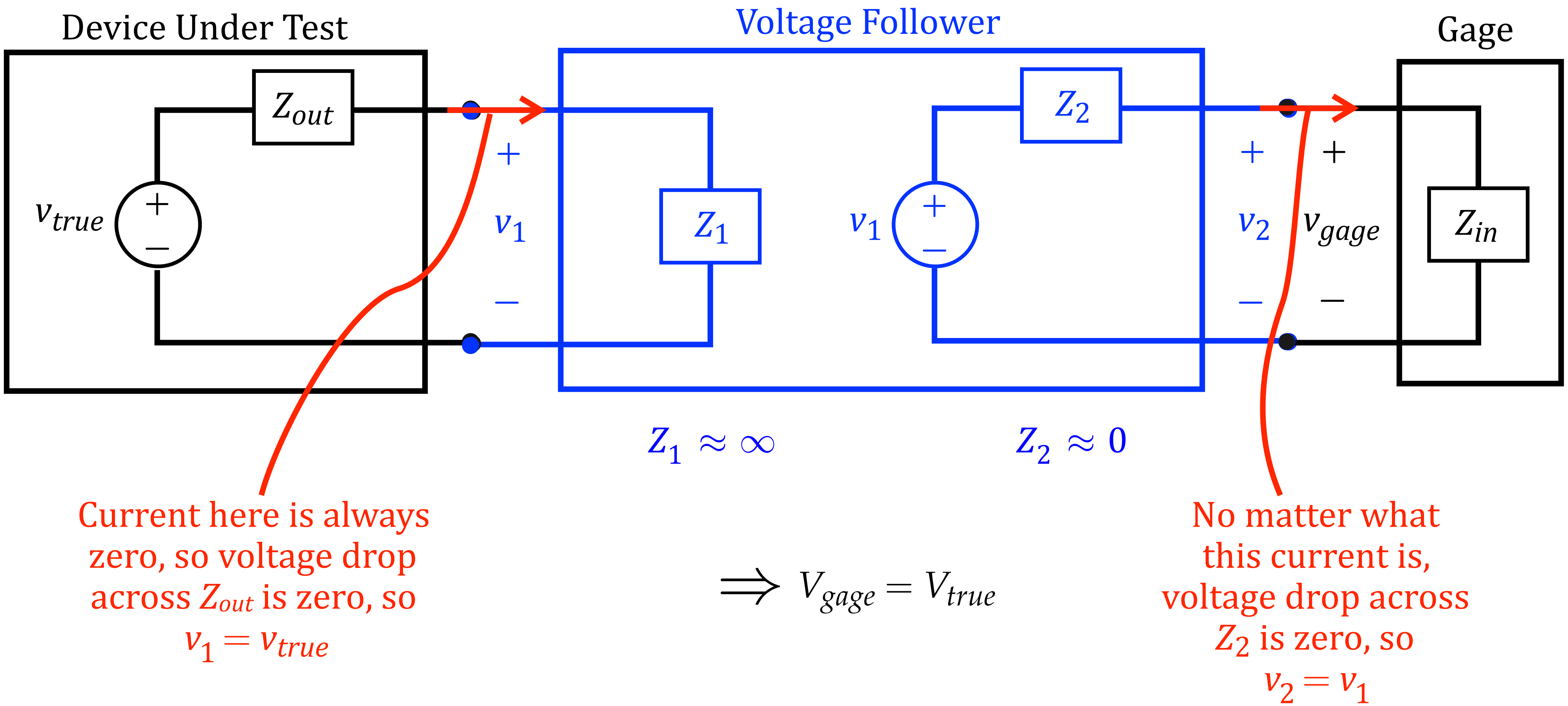
Reduce loading effect by
increasing Z_{in}
or *decreasing Z_{out}*
or both

$$\frac{v_{gage}}{v_{true}} = \frac{Z_{in}}{Z_{in} + Z_{out}} = \frac{1}{1 + \dfrac{Z_{out}}{Z_{in}}}$$



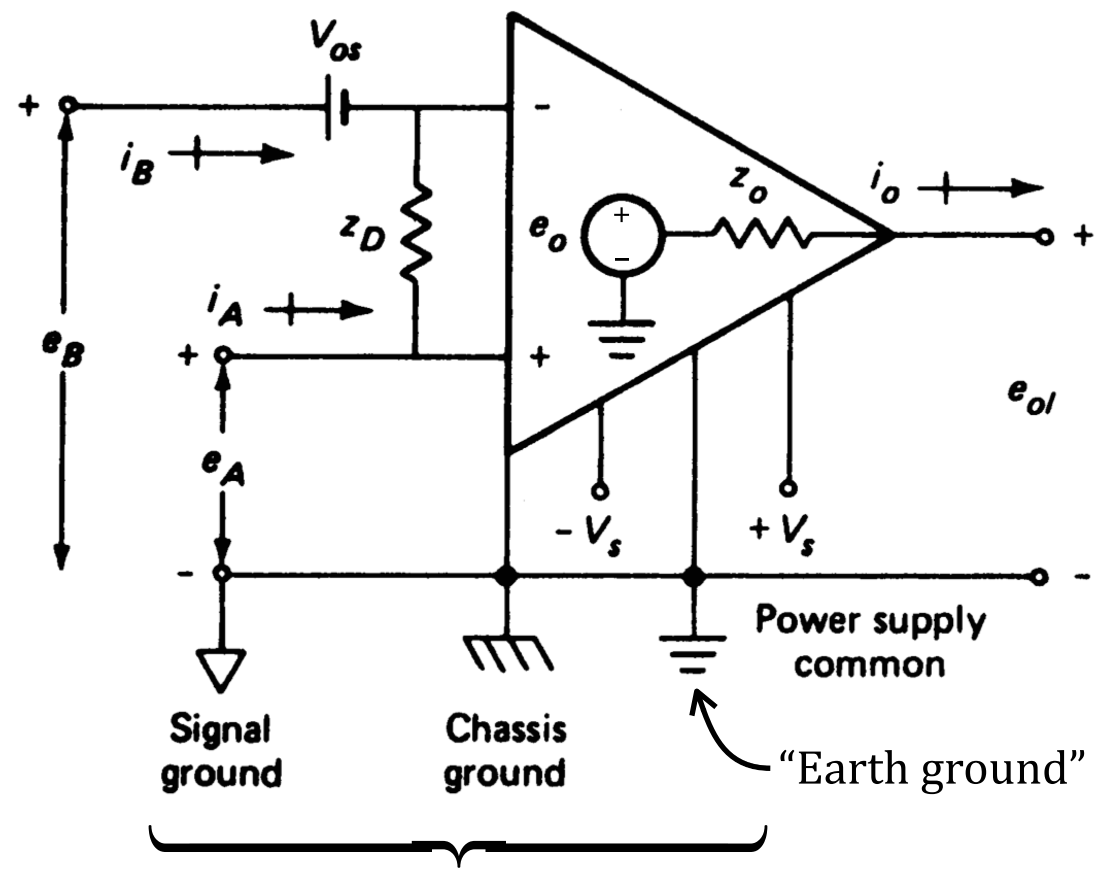
Can fix an
impedance mismatch!

An Impedance Mismatch Fix



$$\Longrightarrow V_{gage} = V_{true}$$

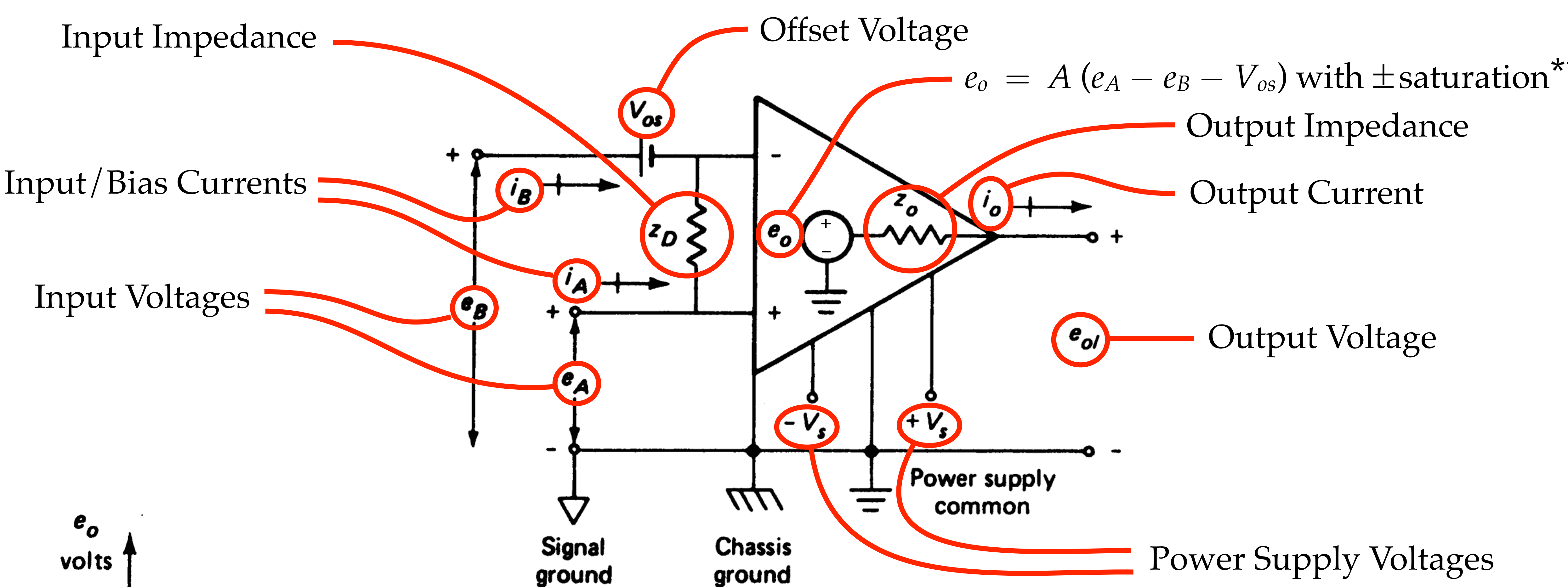
Operational Amplifier Model



These three grounds ideally at same potential.

From *Measurement Systems Application and Design*, 5th Edition, by Earnest O. Doebelin.

Operational Amplifier Model



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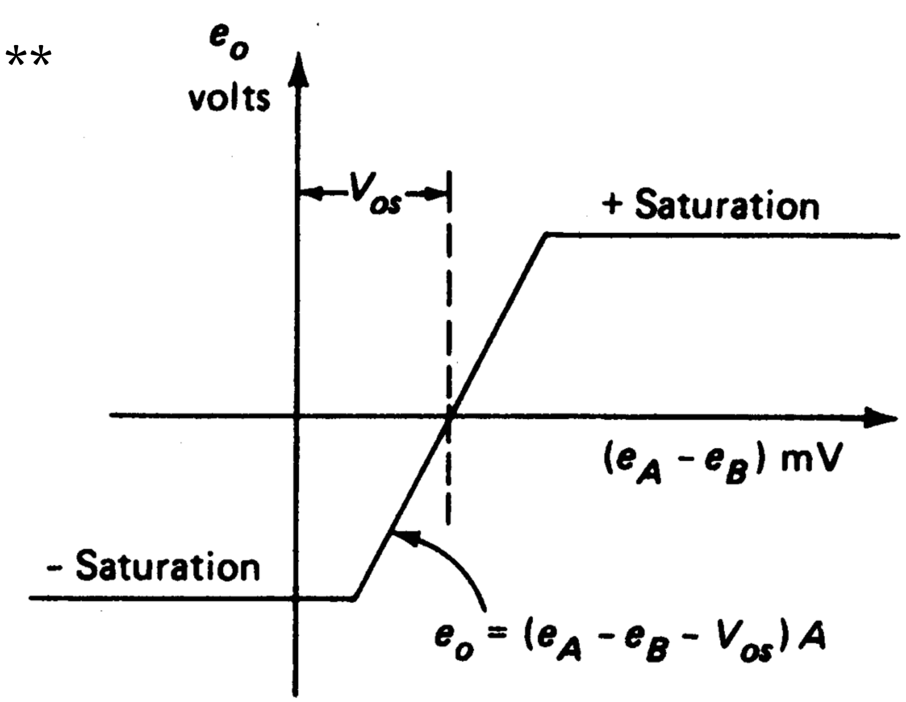
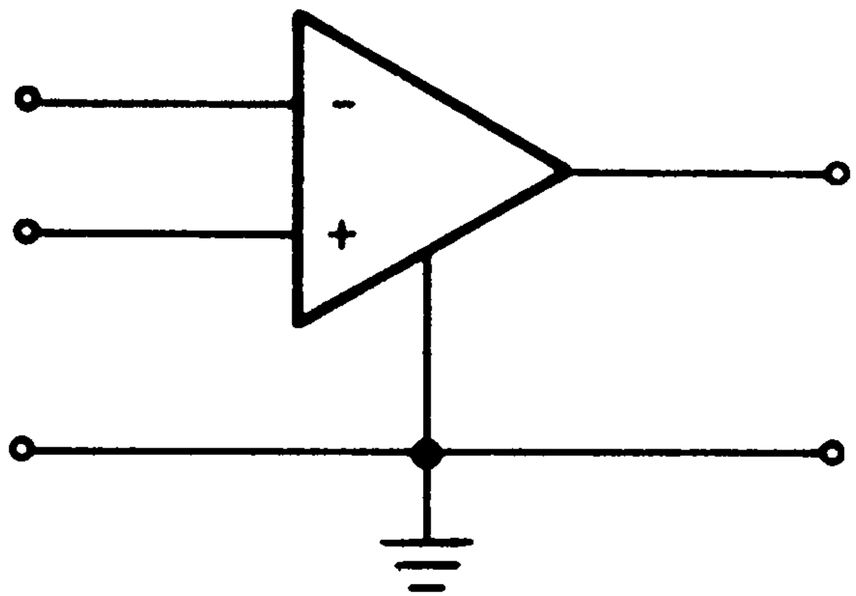


Characteristic	Ideal value	Typical real-world value
Open-loop gain A	∞	100,000 V/V
Offset voltage V_{os}	0	± 1 mV @ 25°C
Bias currents i_A, i_B	0	10^{-6} to 10^{-14} A
Input impedance Z_D	∞	10^5 to 10^{11} Ω
Output impedance Z_o	0	1 to 10 Ω

Ideal Operational Amplifier



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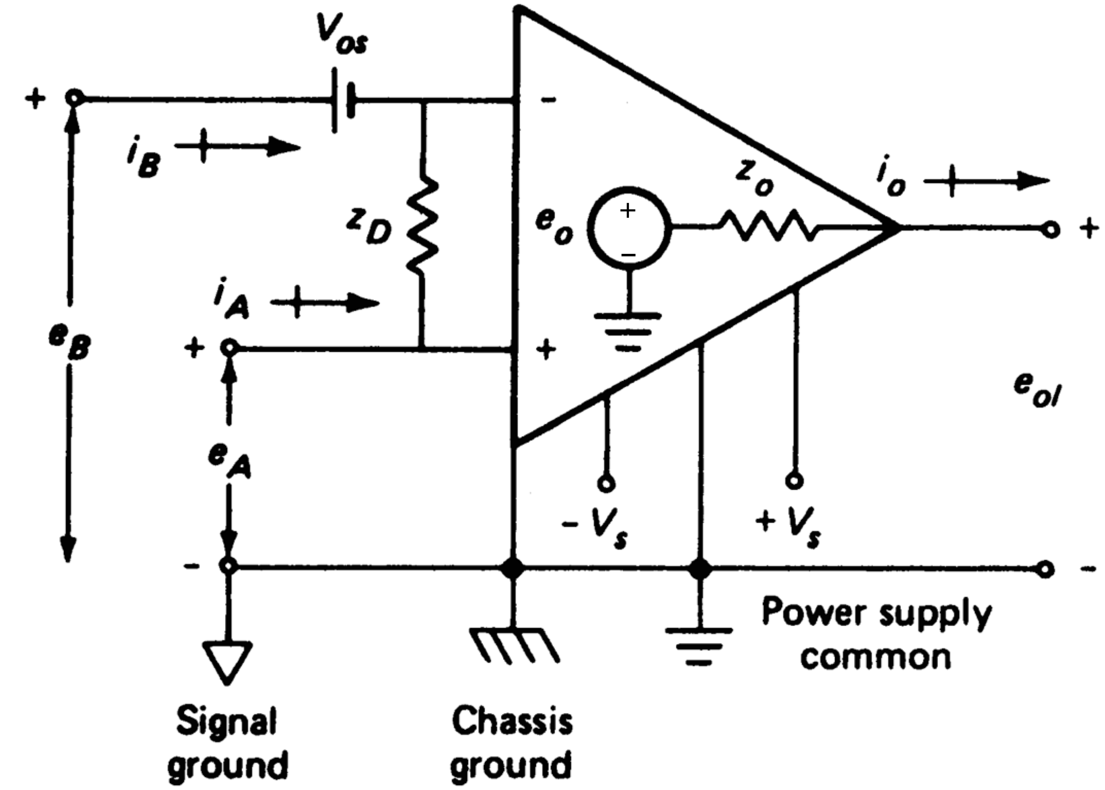

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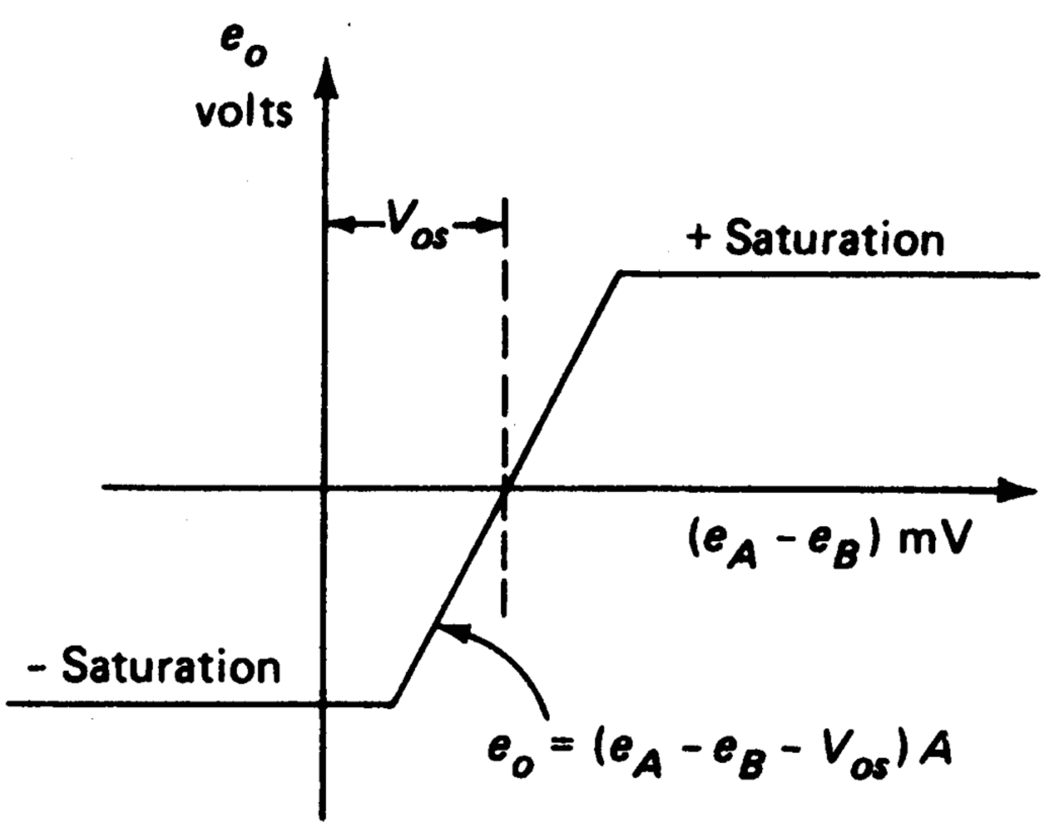

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Characteristic	Ideal value	Typical real-world value
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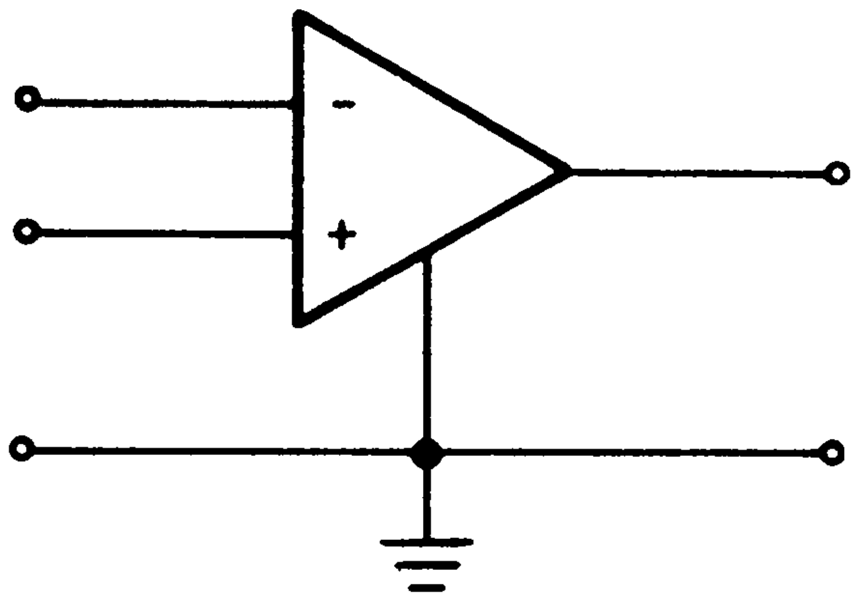
Offset voltage V_{os}	0	± 1 mV @ 25°C
Bias currents i_A, i_B	0	10^{-6} to 10^{-14} A
Input impedance Z_D	∞	10^5 to 10^{11} Ω
Output impedance Z_o	0	1 to 10 Ω

What could such a device possibly be useful for?

Open-loop (i.e., without feedback): Almost nothing.

Closed-loop (i.e., with feedback): Plenty. See below.



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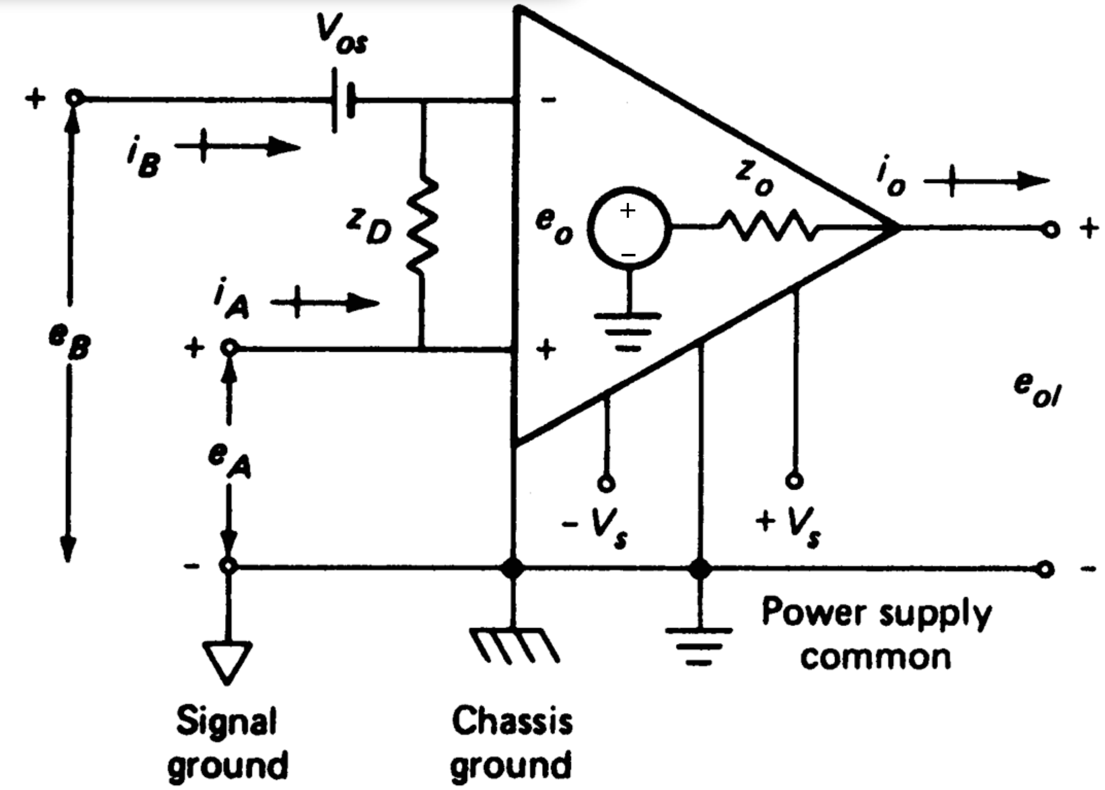

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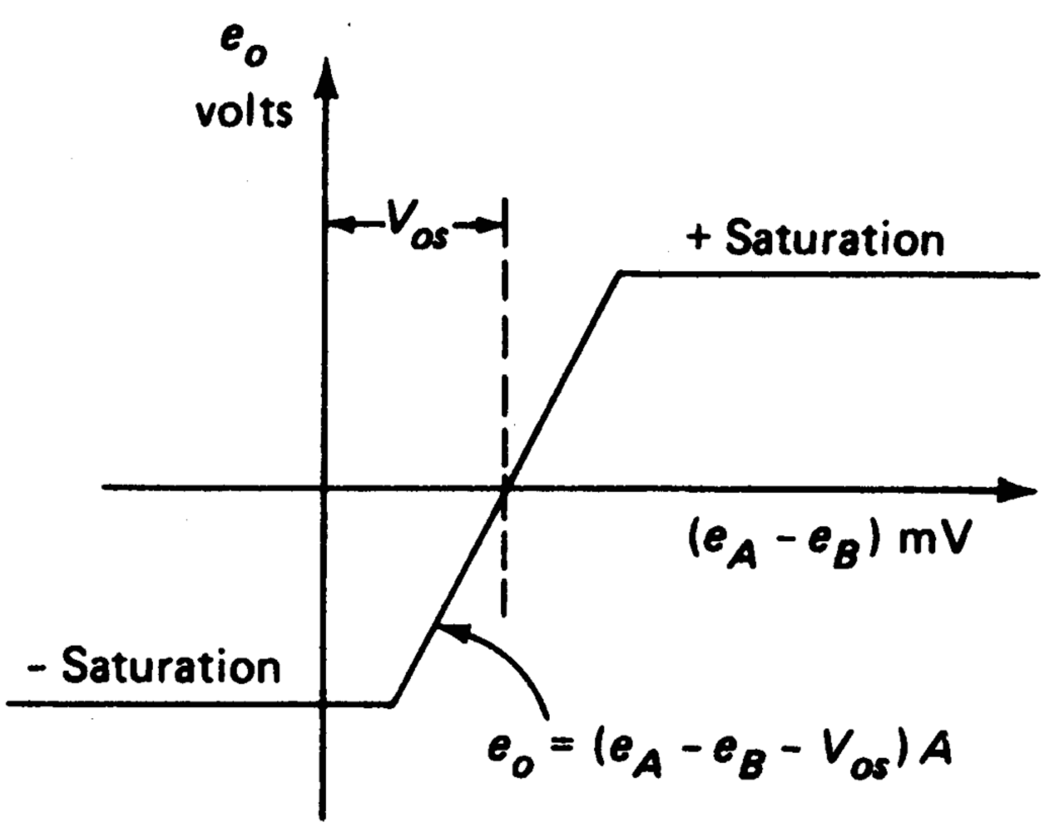

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Characteristic	Ideal value	Typical real-world value
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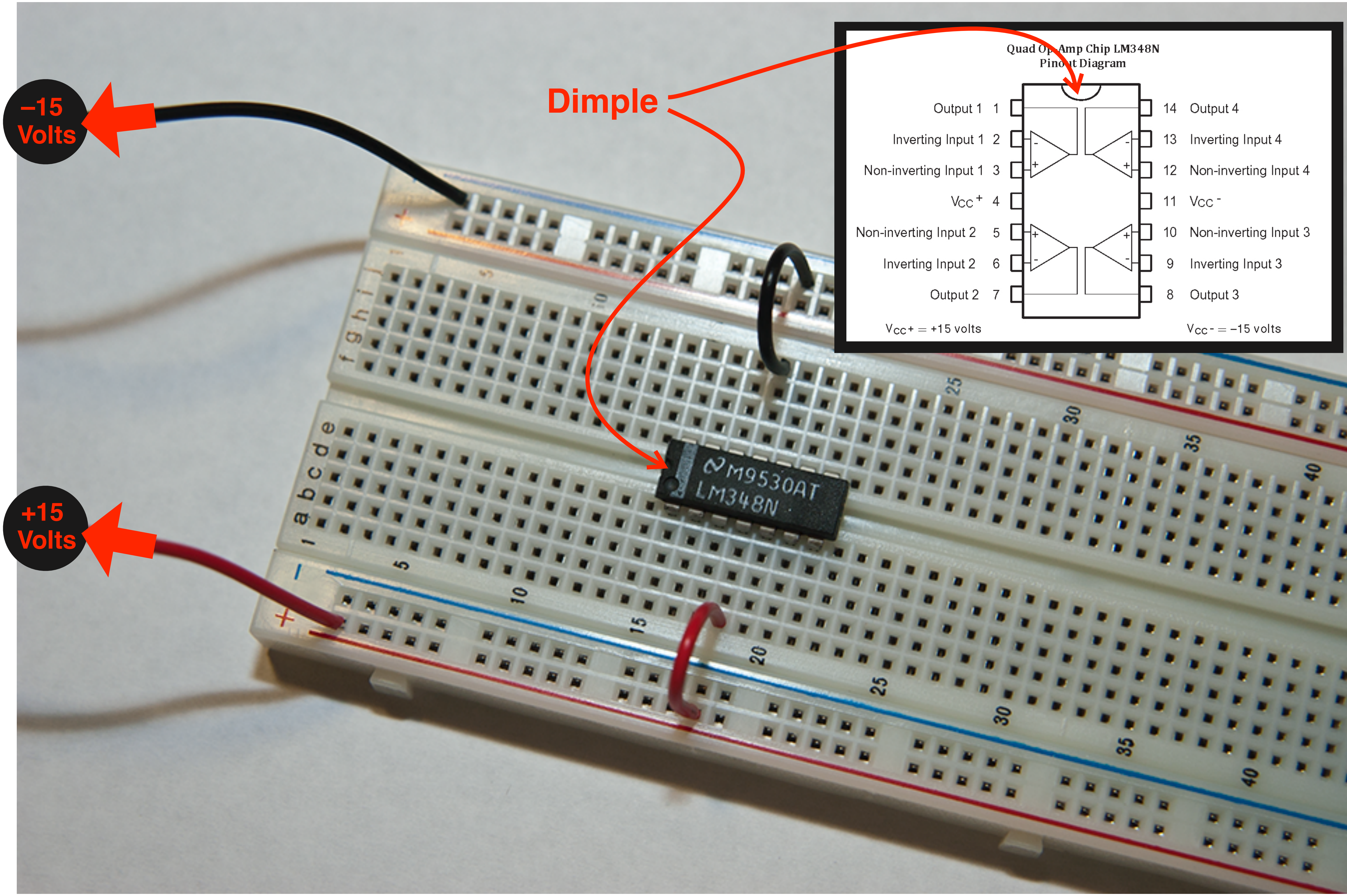
Offset voltage V_{os}	0	± 1 mV @ 25°C
Bias currents i_A, i_B	0	10^{-6} to 10^{-14} A
Input impedance Z_D	∞	10^5 to 10^{11} Ω
Output impedance Z_o	0	1 to 10 Ω

Quad Op-Amp Chip LM348N



LM148, LM248, LM348 **
QUADRUPLE OPERATIONAL AMPLIFIERS

SLOS058C – OCTOBER 1979 – REVISED FEBRUARY 2002

electrical characteristics at specified free-air temperature, $V_{CC\pm} = \pm 15$ V (unless otherwise noted)

PARAMETER		TEST CONDITIONS†		LM148 MIN	LM148 TYP	LM148 MAX	LM248 MIN	LM248 TYP	LM248 MAX	LM348 MIN	LM348 TYP	LM348 MAX	UNIT
V_{IO}	Input offset voltage	$V_O = 0$	25°C		1	5		1	6		1	6	mV
			Full range			6			7.5			7.5	
I_{IO}	Input offset current	$V_O = 0$	25°C		4	25		4	50		4	50	nA
			Full range			75			125			100	
I_{IB}	Input bias current	$V_O = 0$	25°C		30	100		30	200		30	200	nA
			Full range			325			500			400	
V_{ICR}	Common-mode input voltage range		Full range	±12			±12			±12			V
V_{OM}	Maximum peak output voltage swing	$R_L = 10\ k\Omega$	25°C	±12	±13		±12	±13		±12	±13		V
		$R_L \geq 10\ k\Omega$	Full range	±12			±12			±12			
		$R_L = 2\ k\Omega$	25°C	±10	±12		±10	±12		±10	±12		
		$R_L \geq 2\ k\Omega$	Full range	±10			±10			±10			
A_{VD}	Large-signal differential voltage amplification	$V_O = \pm 10$ V, $R_L = \geq 2\ k\Omega$	25°C	50	160		25	160		25	160		V/mV
			Full range	25			15			15			
r_i	Input resistance‡		25°C	0.8	2.5		0.8	2.5		0.8	2.5		MΩ
B_1	Unity-gain bandwidth	$A_{VD} = 1$	25°C		1			1			1		MHz
ϕ_m	Phase margin	$A_{VD} = 1$	25°C		60°			60°			60°		
CMRR	Common-mode rejection ratio	$V_{IC} = V_{ICR}$min, $V_O = 0$	25°C	70	90		70	90		70	90		dB
			Full range	70			70			70			
k_{SVR}	Supply-voltage rejection ratio ($\Delta V_{CC\pm}/\Delta V_{IO}$)	$V_{CC\pm} = \pm 9$ V to ±15 V, $V_O = 0$	25°C	77	96		77	96		77	96		dB
			Full range	77			77			77			
I_{OS}	Short-circuit output current		25°C		±25			±25			±25		mA
I_{CC}	Supply current (four amplifiers)	No load, $V_O = 0$	25°C					2.4	4.5		2.4	4.5	mA
		No load, $V_O = V_{OM}$			2.4	3.6							
V_{O1}/V_{O2}	Crosstalk attenuation	f = 1 Hz to 20 kHz	25°C		120			120			120		dB

† All characteristics are measured under open-loop conditions with zero common-mode input voltage, unless otherwise specified. Full range for T_A is −55°C to 125°C for LM148, −25°C to 85°C for LM248, and 0°C to 70°C for LM348.
‡ This parameter is not production tested.

**Posted on course web page.

Box 2.1 Op-Amps **

Ideal Op-Amp Properties

- Infinite open-loop differential gain
- Infinite input impedance
- Zero output impedance
- Infinite bandwidth
- Zero output for zero differential input

Ideal Analysis Assumptions

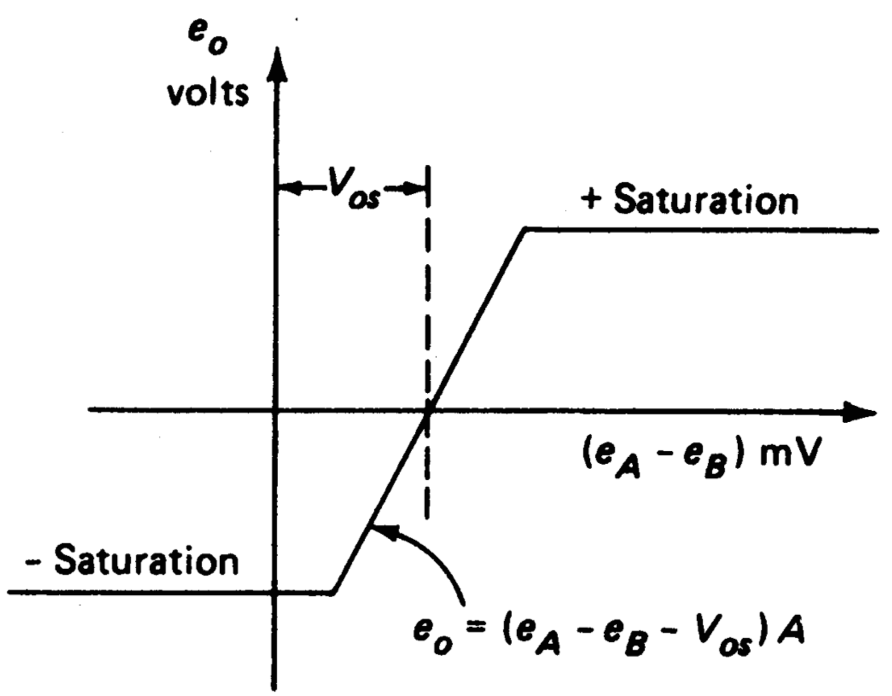
- Voltages at the two input leads are equal.
- Current through either input lead is zero.

Definitions

- Open-loop gain $= \left|\dfrac{\text{Output voltage}}{\text{Voltage difference at input leads}}\right|$, with no feedback.
- Input impedance $= \dfrac{\text{Voltage between an input lead and ground}}{\text{Current through that lead}}$, with other input lead grounded and the output in open circuit.
- Output impedance $= \dfrac{\text{Voltage between output lead and ground in open circuit}}{\text{Current through that lead}}$, with normal input conditions.
- Bandwidth is the frequency range in which the frequency response is flat (gain is constant).
- GBP = Open-loop gain × Bandwidth at that gain
- Input bias current is the average DC current through one input lead.
- Input offset current is the difference in the two input bias currents.
- Differential input voltage is the voltage at one input lead with the other grounded when the output voltage is zero.
- Common-mode gain $= \dfrac{\text{Output voltage when input leads are at the same voltage}}{\text{Common input voltage}}$
- Common-mode rejection ratio (CMRR) $= \dfrac{\text{Open loop differential gain}}{\text{Common-mode gain}}$
- Slew rate is the rate of change of output of a unity-gain op-amp, for a step input.

**From our textbook

Recall this model of op-amp's gain characteristic:



It assumes that e_A and e_B affect e_o only via their difference, $e_A - e_B$.

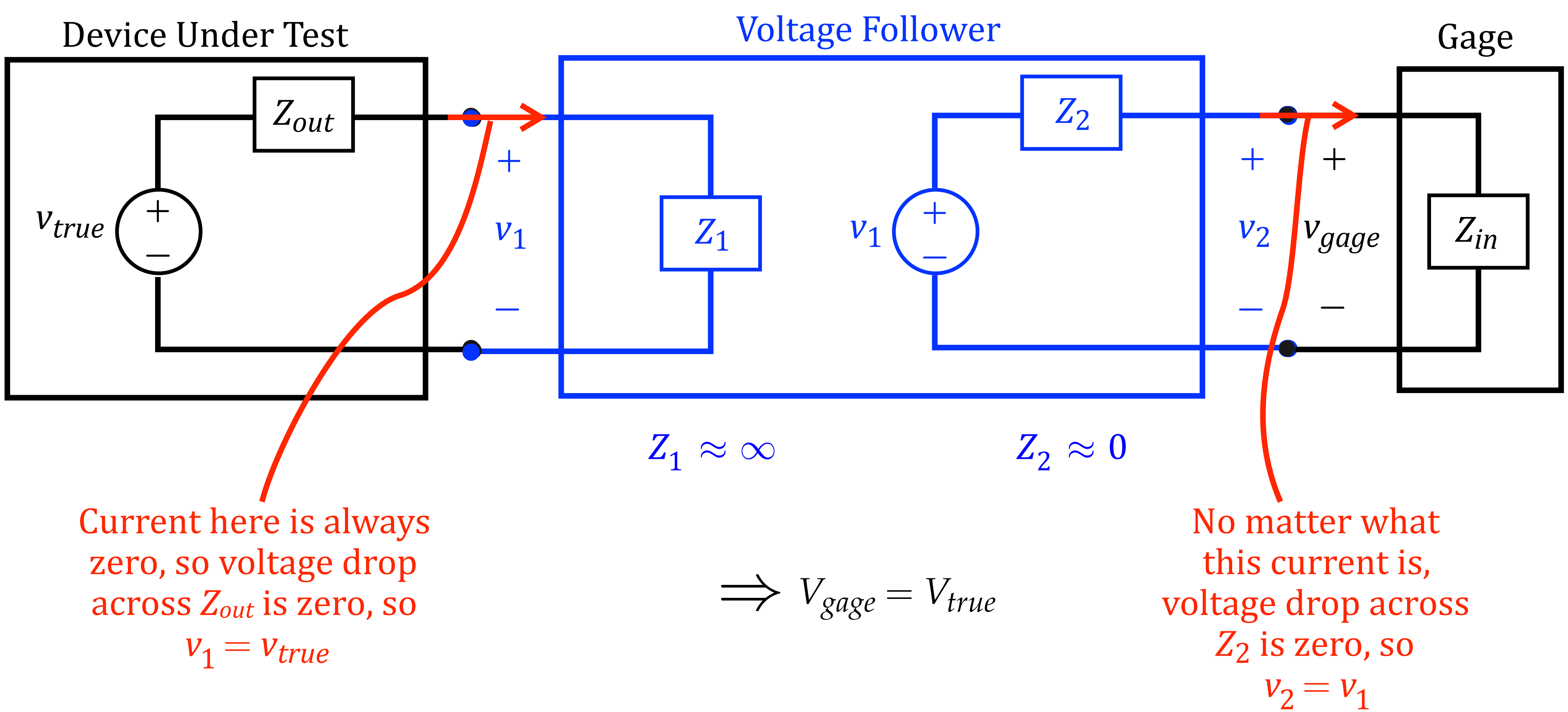
In practice, the *average* of e_A and e_B also affects e_o.

That is, physical op-amps have nonzero *common-mode gain*.

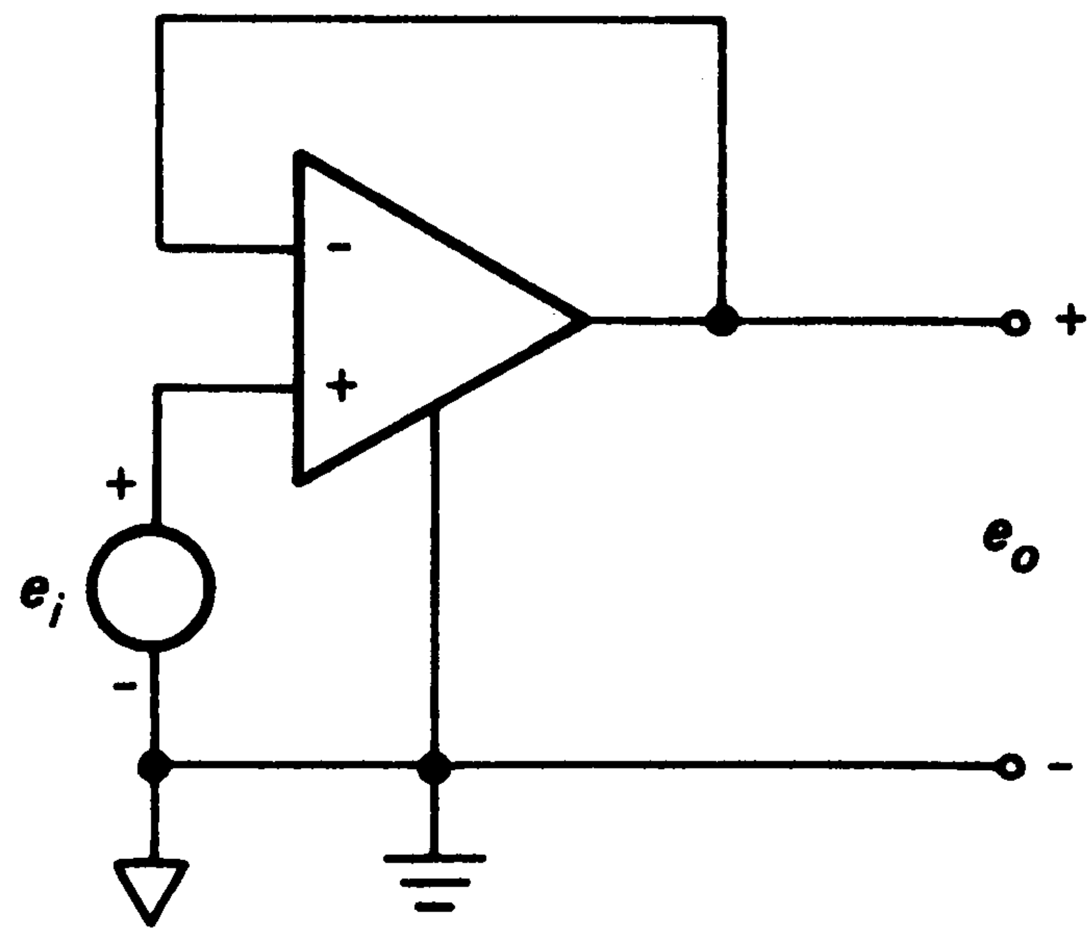
See Section 2.4.2.1 of de Silva for details.

(Complication: Doebelin and de Silva employ different definitions of *common-mode gain*!)

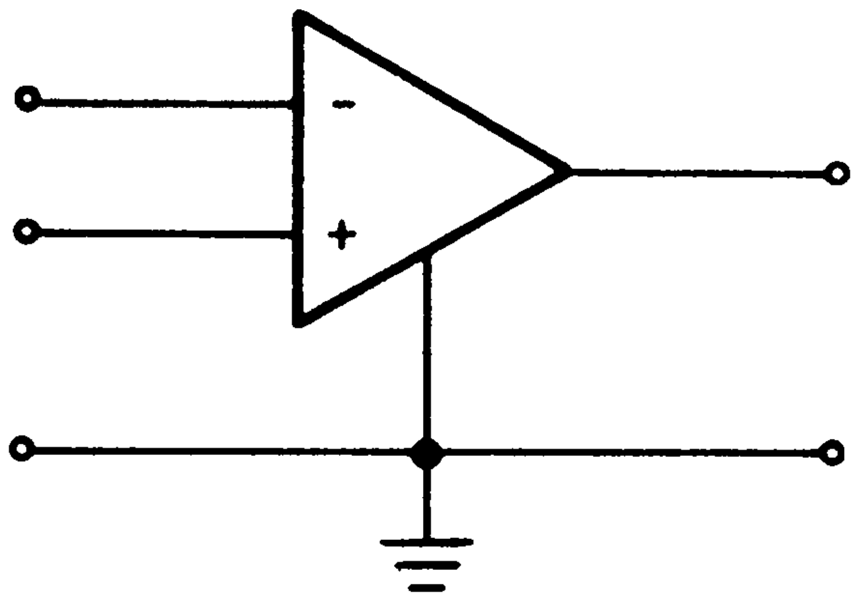
Example Application: Voltage Follower



Example Application: Voltage Follower



Ideal Operational Amplifier



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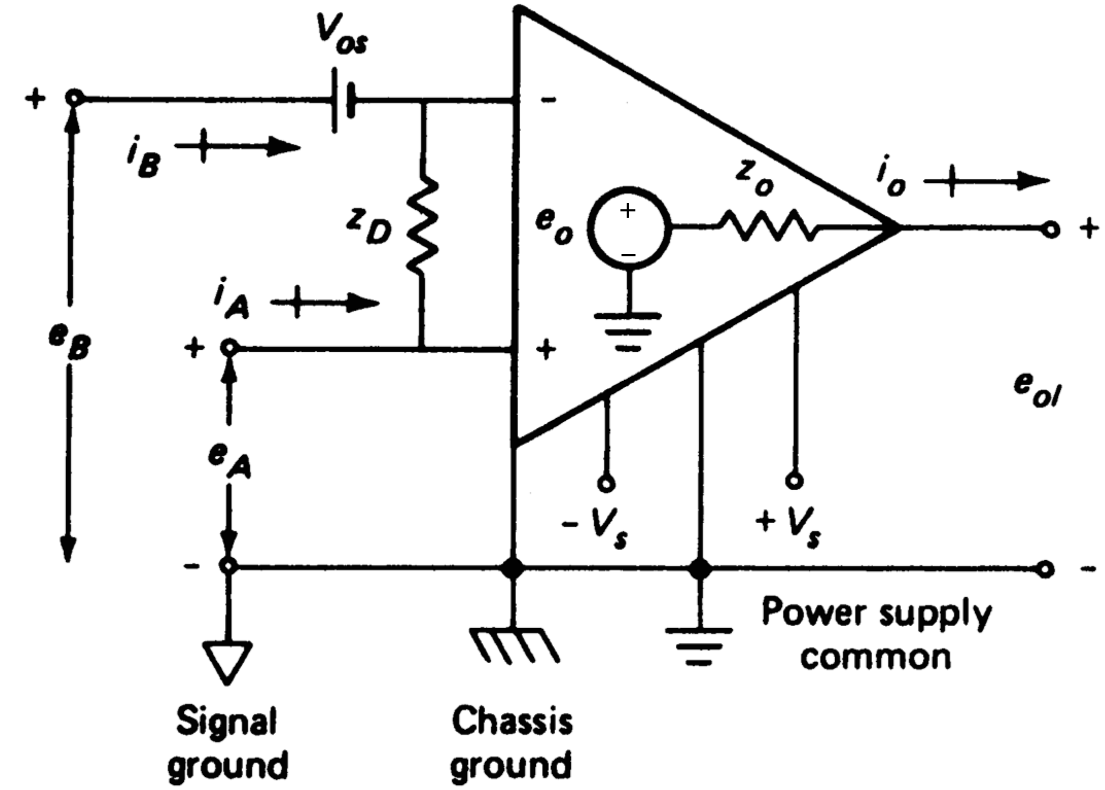

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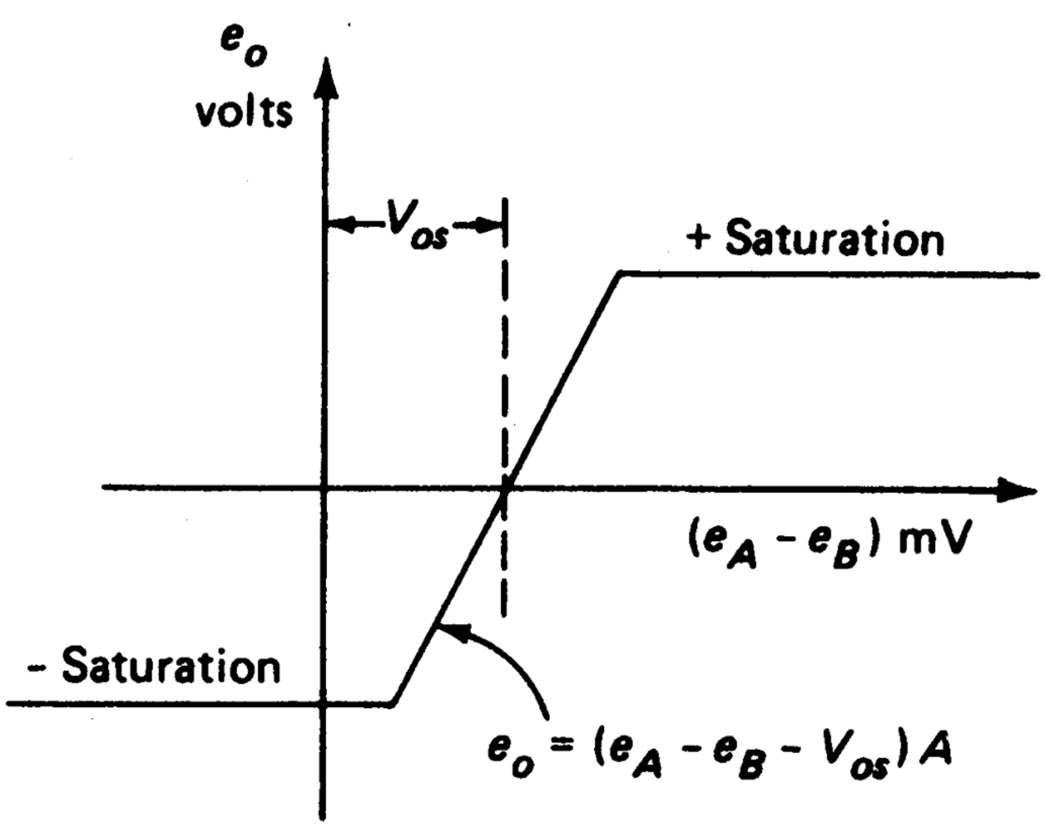

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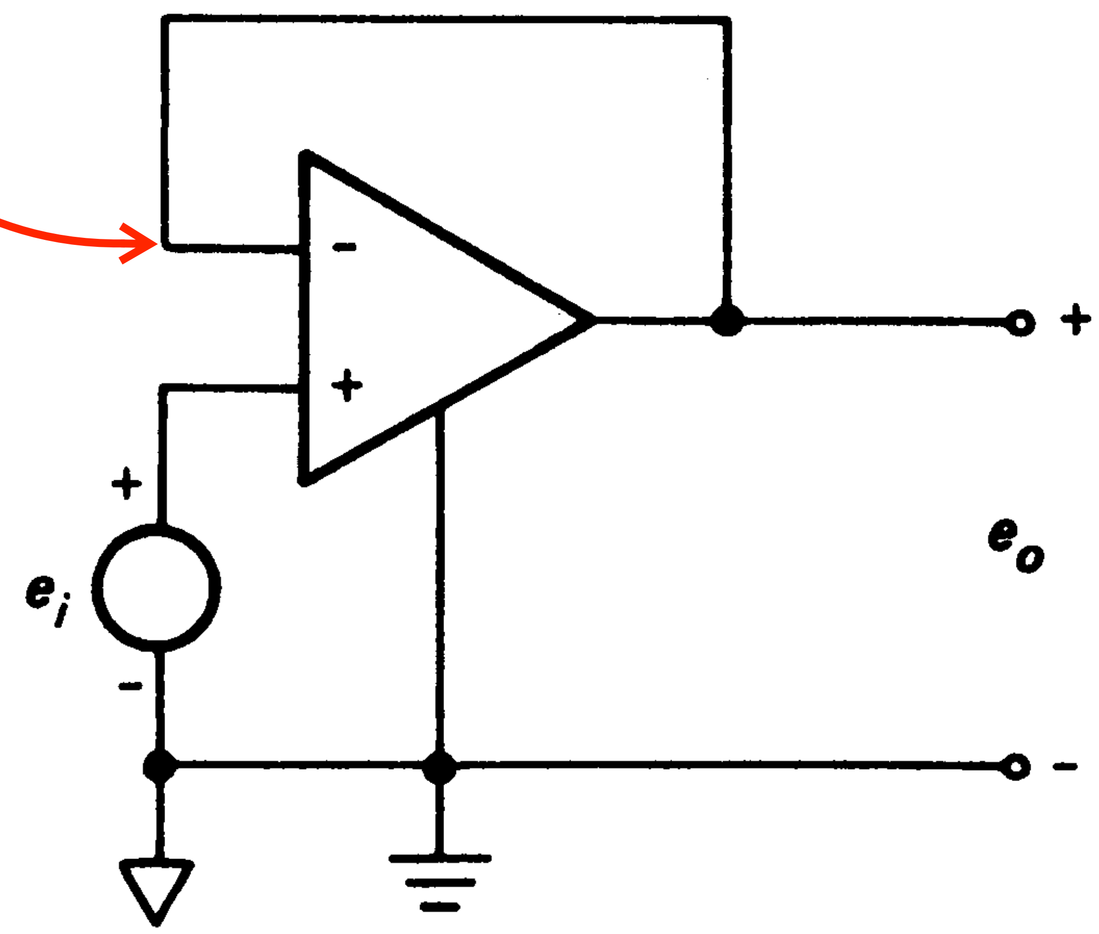
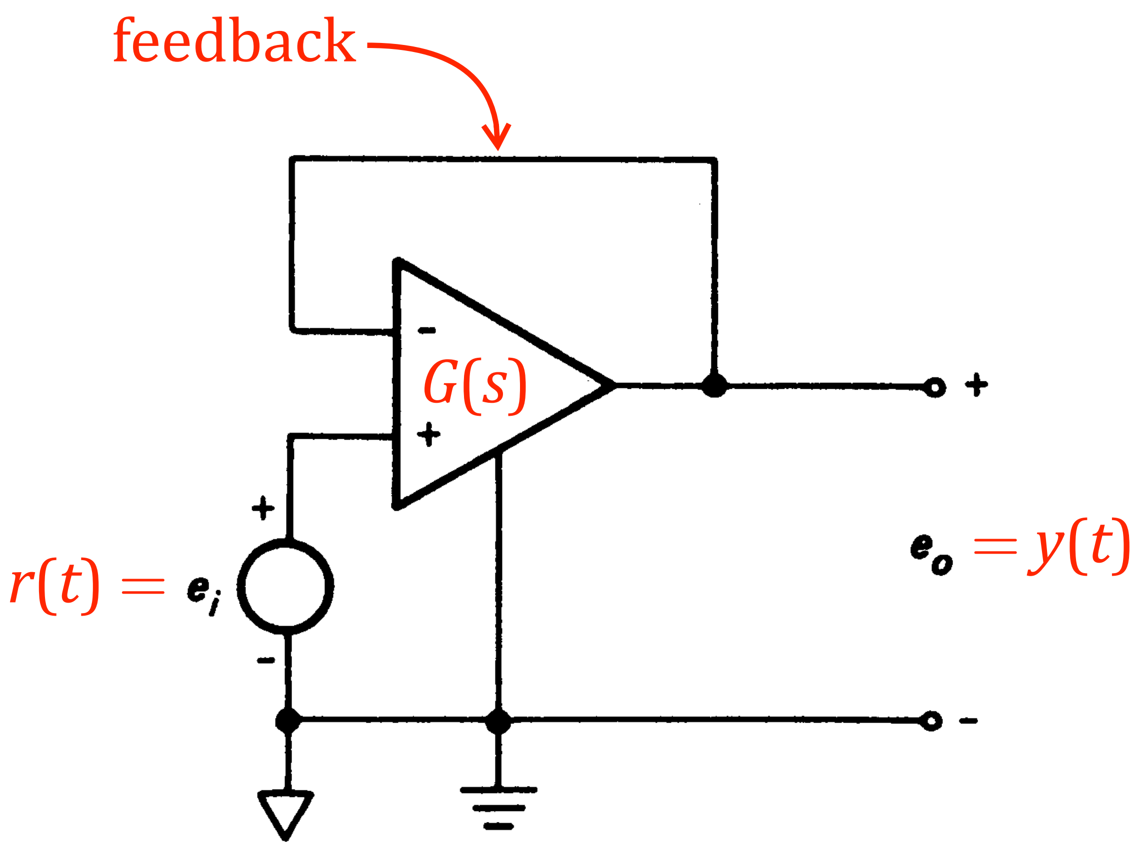
Characteristic	Ideal value	Typical real-world value
Open-loop gain A	∞	~~100,000 V/V~~
Offset voltage V_{os}	0	~~± 1 mV @ 25°C~~
Bias currents i_A, i_B	0	~~10^{-6} to 10^{-14} A~~
Input impedance Z_D	∞	~~10^5 to 10^{11} Ω~~
Output impedance Z_o	0	~~1 to 10 Ω~~

Example Application: Voltage Follower

For e_o to not saturate, voltage here must be e_i. But, by inspection, voltage here is also e_o, so it must also be that $e_o = e_i$.

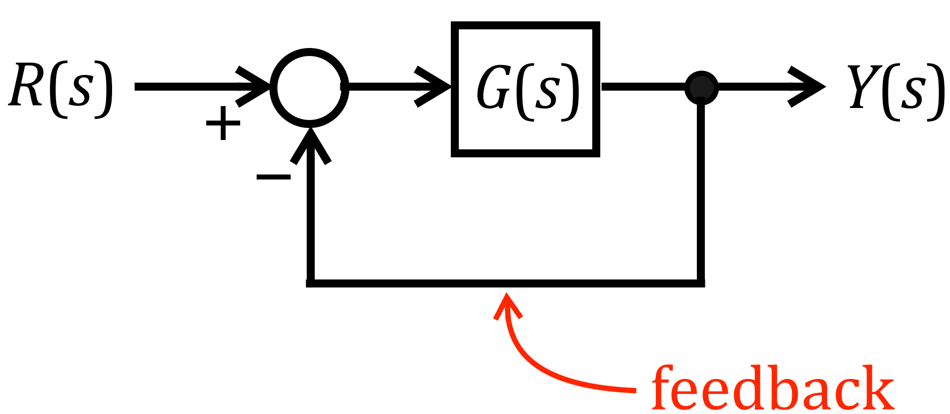


What's going on?



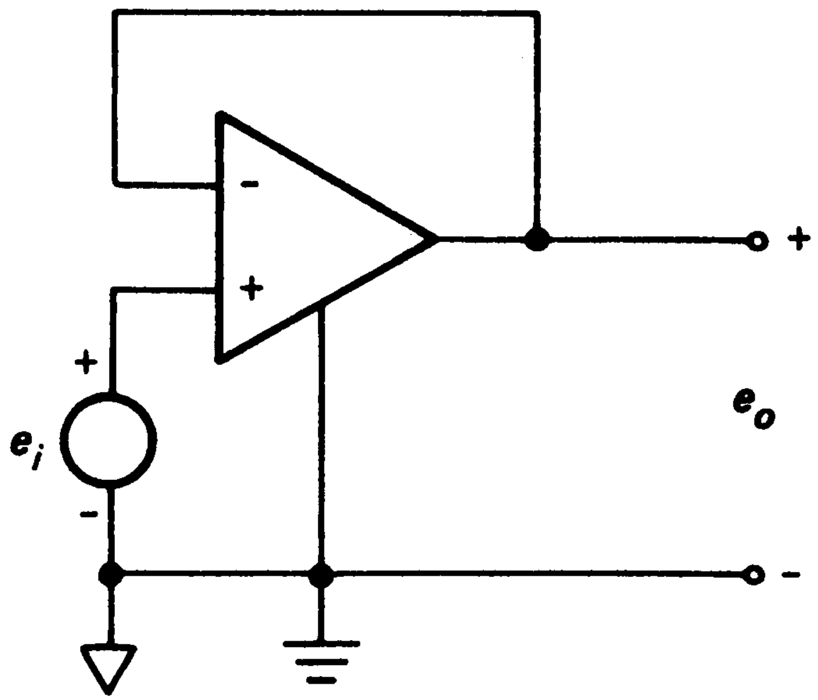
$$\Rightarrow \qquad \frac{e_o}{e_i} = 1 \quad ?$$

Consider:



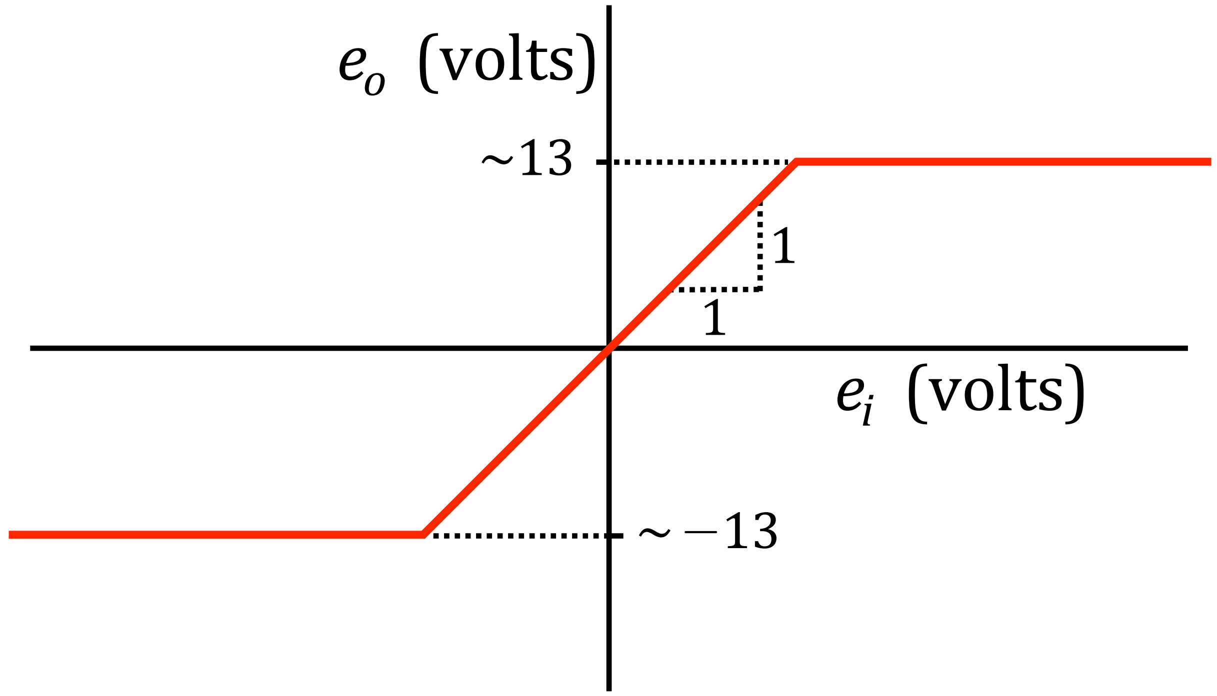
$$\Rightarrow \quad \frac{Y(s)}{R(s)} = \frac{G(s)}{1+G(s)} \quad \Rightarrow \quad \lim_{|G(s)|\to\infty} \frac{Y(s)}{R(s)} = 1$$

"Saturation" limits voltage range



$$\Rightarrow \qquad \frac{e_o}{e_i} = 1$$

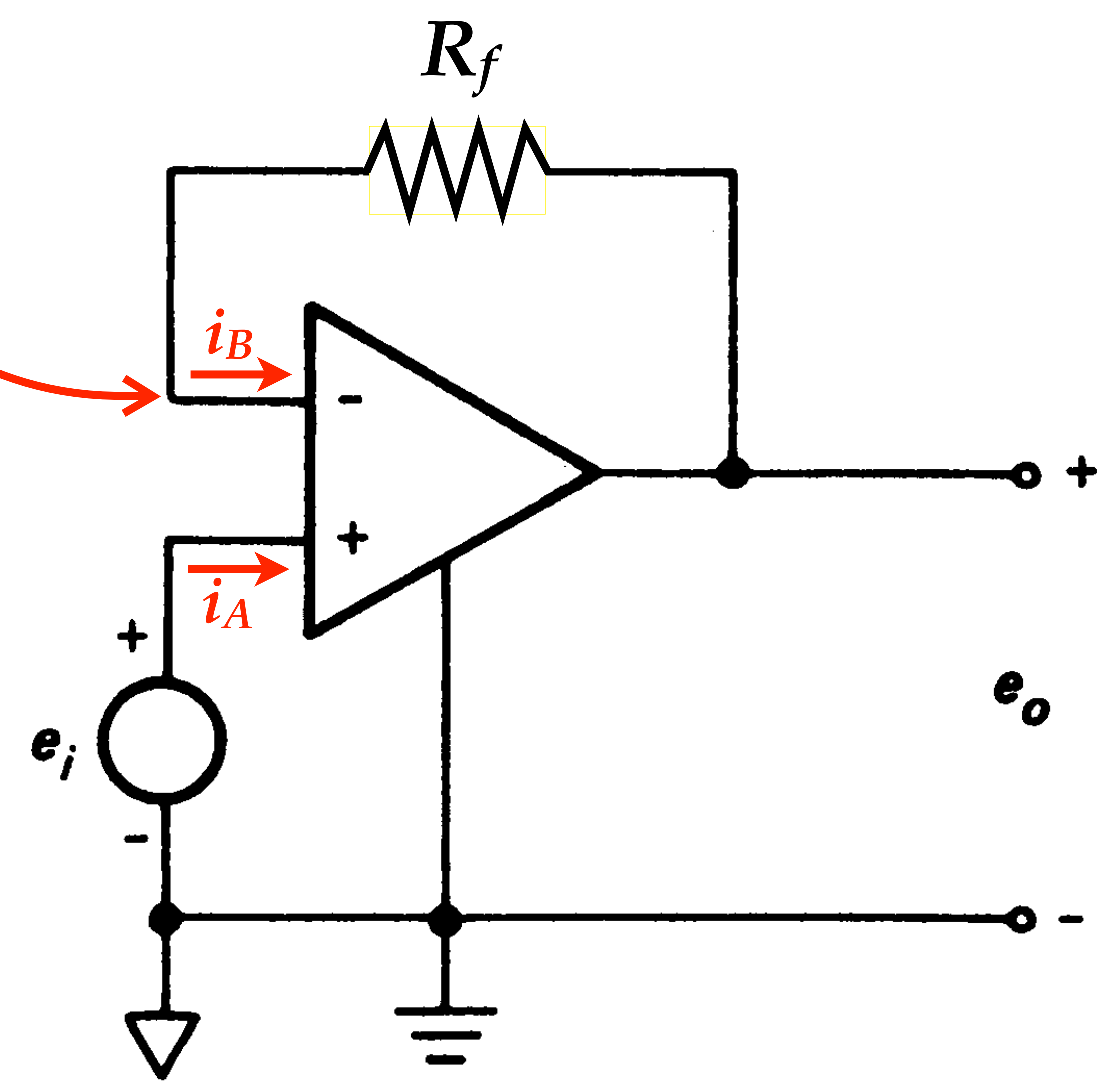
With saturation:



Another Voltage Follower

For e_o to not saturate, voltage here must be e_i.

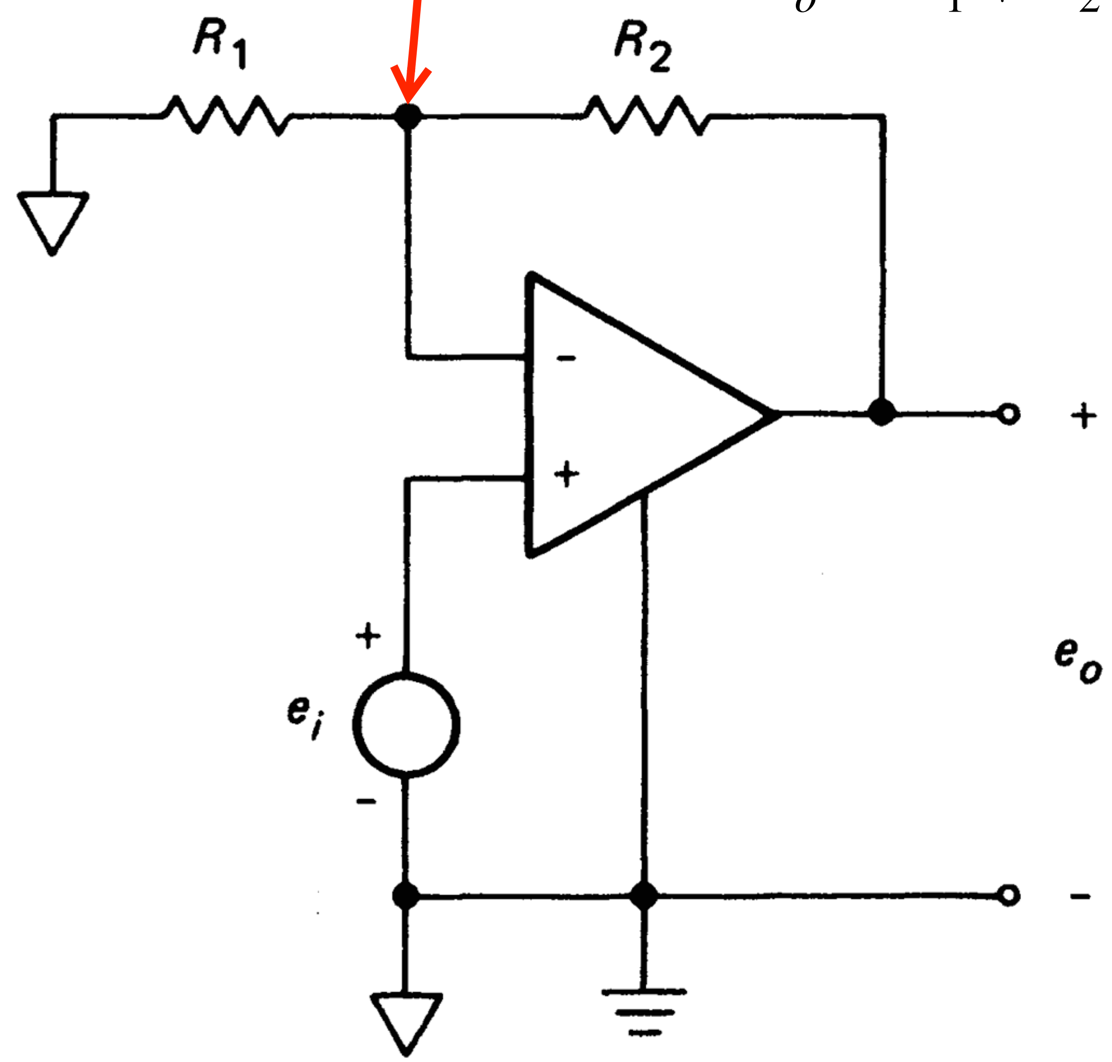
Furthermore, $i_B = 0$, so it must also be that $e_o = e_i$.



For e_0 to not saturate, voltage here must be e_i.

The voltage divider formula then yields

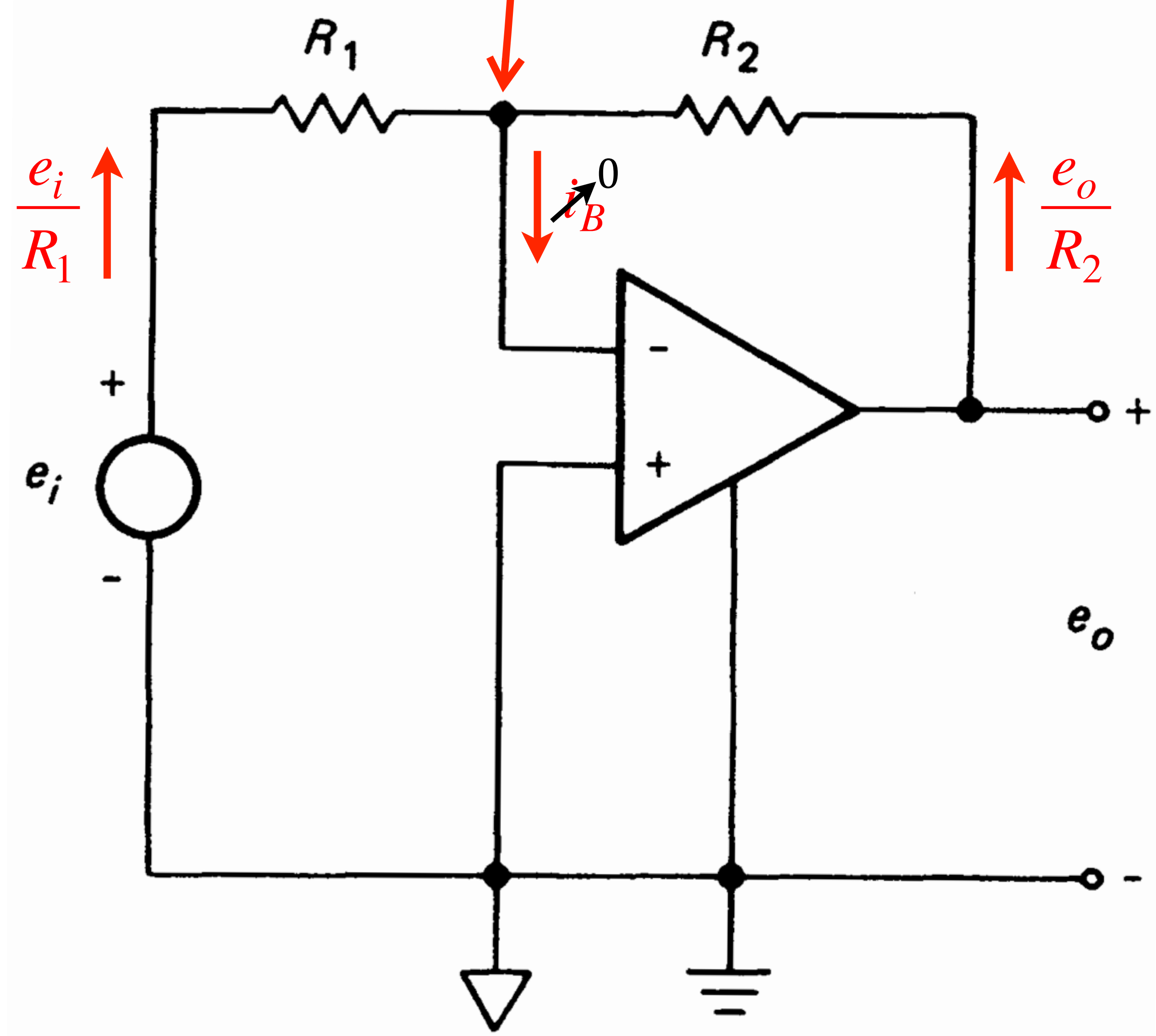
$$\frac{e_i}{e_o} = \frac{R_1}{R_1 + R_2} \qquad \Rightarrow \qquad \frac{e_o}{e_i} = \frac{R_1 + R_2}{R_1}$$



Noninverting amplifier

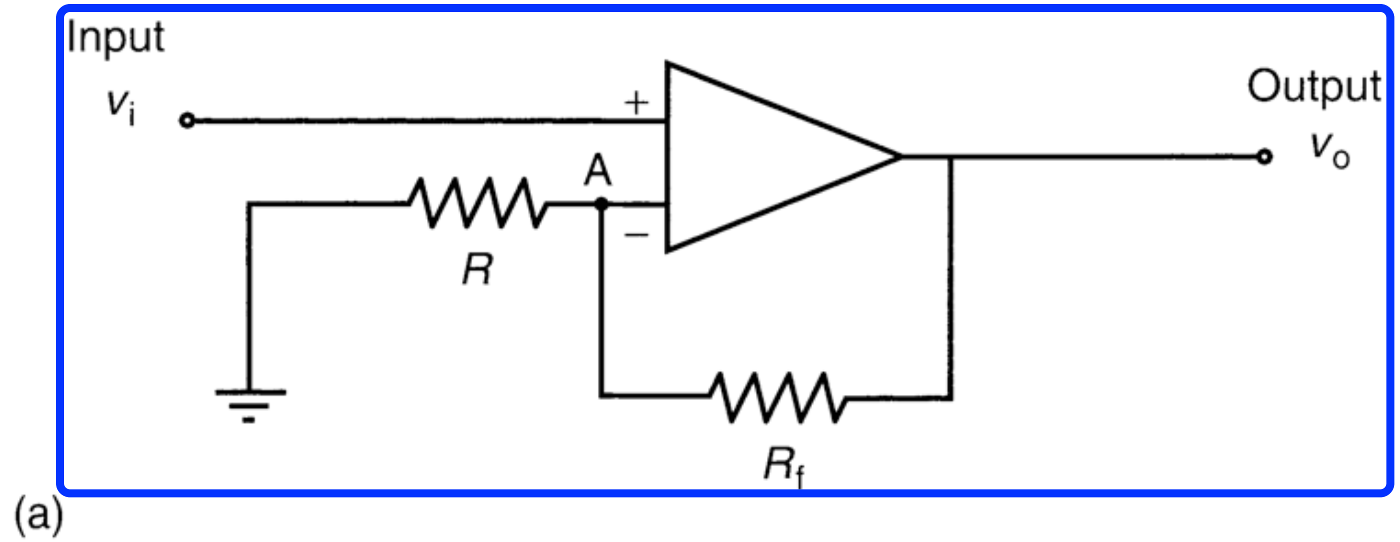
For e_0 to not saturate, voltage here must be 0. Furthermore $i_B = 0$, from which it follows that

$$\frac{e_i}{R_1} = -\frac{e_o}{R_2} \quad \Rightarrow \quad \frac{e_o}{e_i} = -\frac{R_2}{R_1}$$

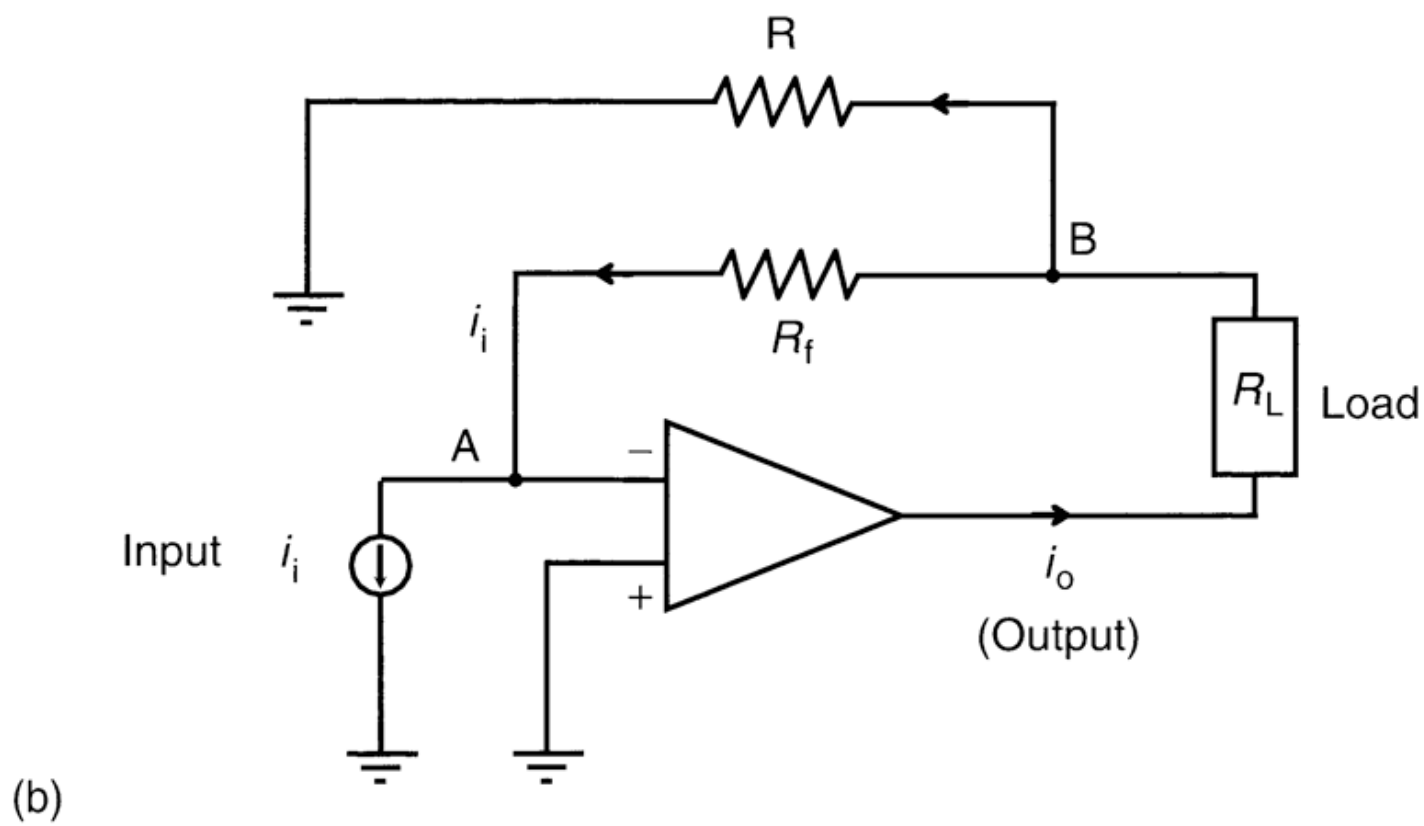


Inverting amplifier

Same noninverting amplifier as above, drawn differently



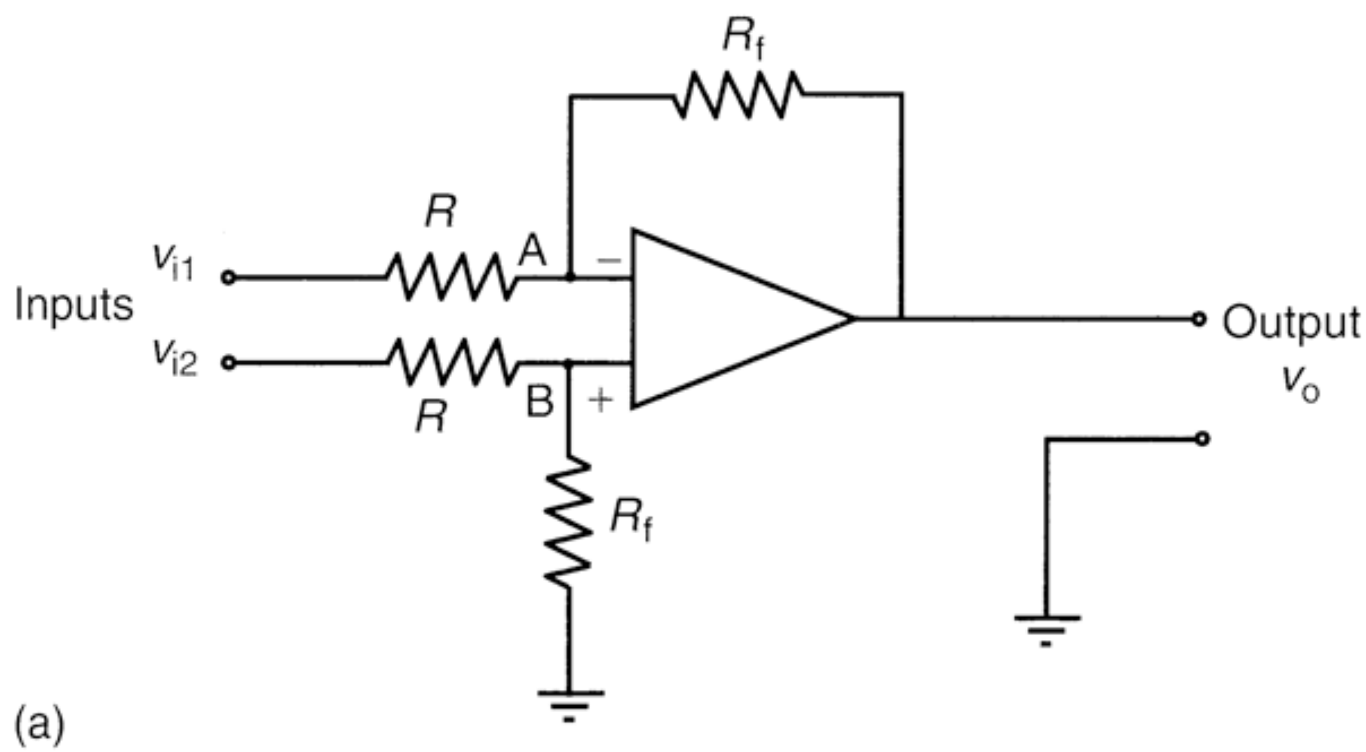
$$\Rightarrow \quad v_o = \frac{R+R_f}{R} v_i = \left(1+\frac{R_f}{R}\right) v_i$$



$$\Rightarrow \quad i_o = \frac{R+R_f}{R} i_i = \left(1+\frac{R_f}{R}\right) i_i$$

(see de Silva for derivation)

Figure 2.15
(a) A voltage amplifier.
(b) A current amplifier.



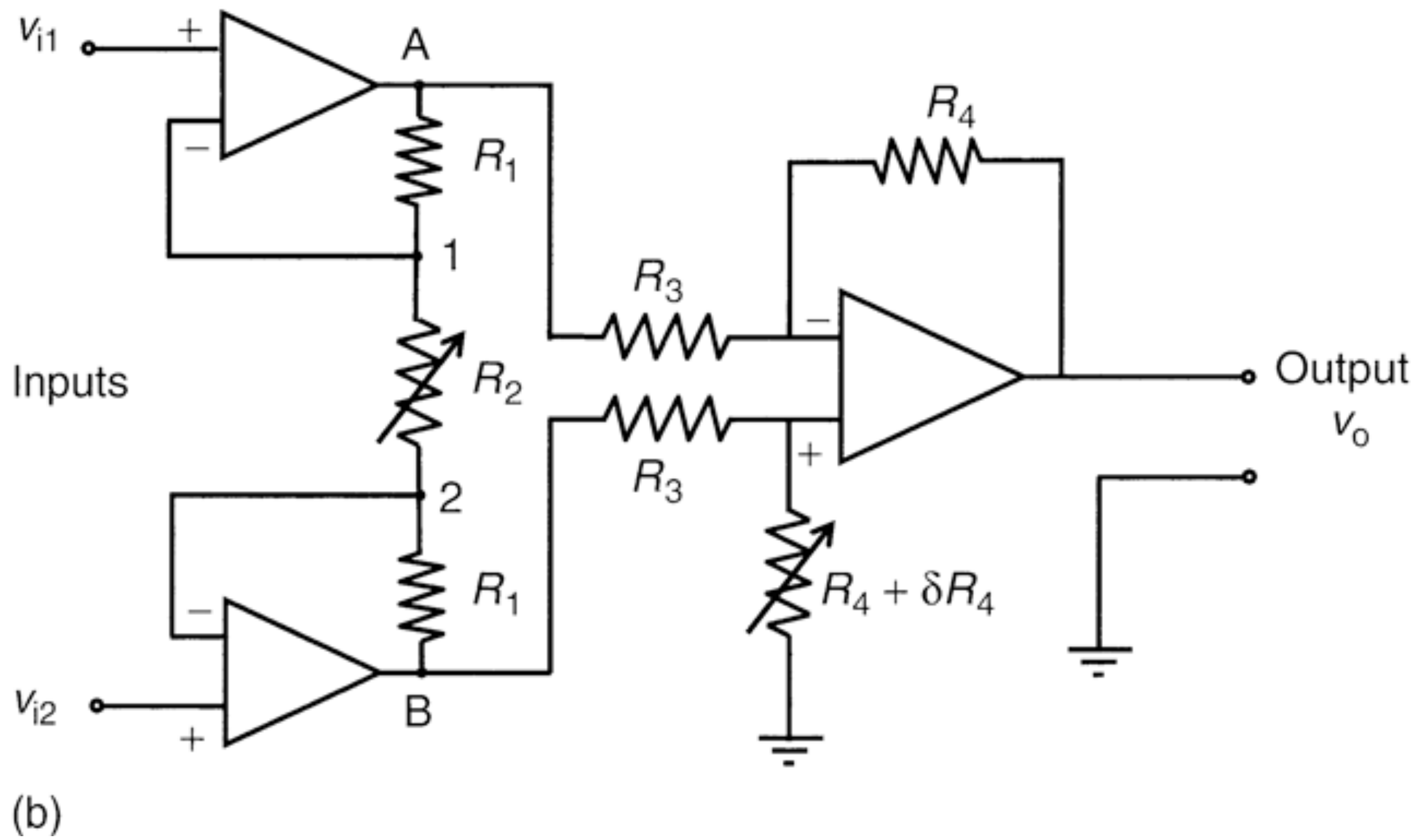
$$\Rightarrow \quad v_o = \frac{R_f}{R}(v_{i2} - v_{i1})$$

(see de Silva for derivation)

$\Uparrow$

Differential Amplifiers

$\Downarrow$



$$\Rightarrow \quad v_o = \frac{R_4}{R_3}\left(1 + \frac{2R_1}{R_2}\right)(v_{i2} - v_{i1})$$

(see de Silva for derivation)

Figure 2.16
(a) A basic differential amplifier.
(b) A basic instrumentation amplifier

$$v_o = \frac{R_f}{R}(v_{i2} - v_{i1})$$

$$v_o = \frac{R_4}{R_3}\left(1 + \frac{2R_1}{R_2}\right)(v_{i2} - v_{i1})$$

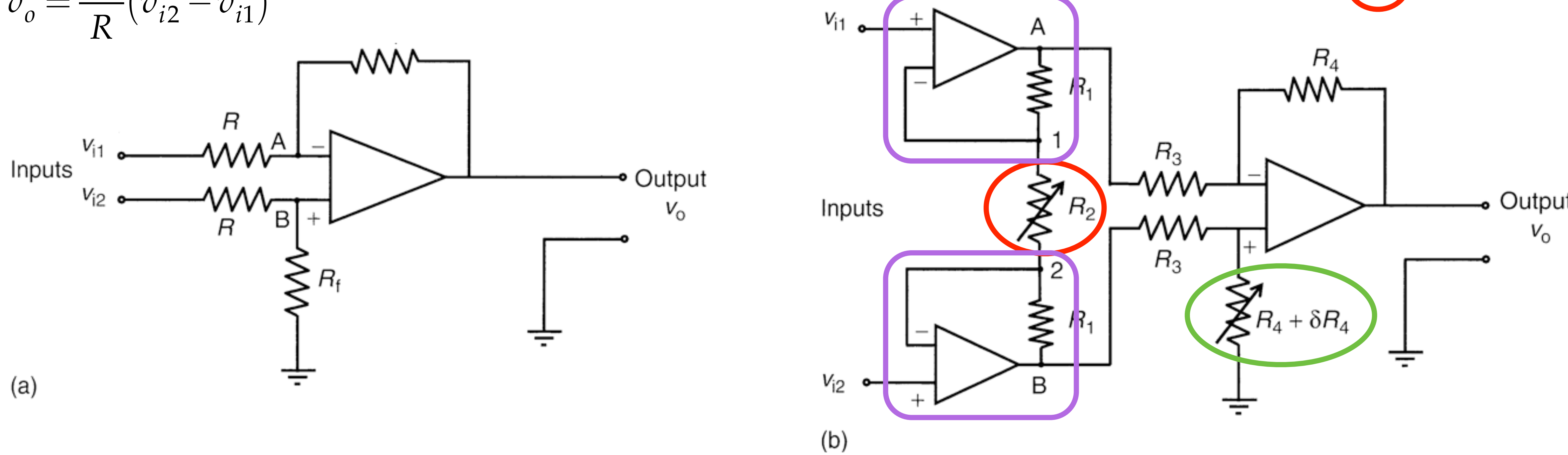


Figure 2.16a

2.4.4.2 Instrumentation Amplifier

The basic differential amplifier, shown in Figure 2.16a and discussed earlier, is an important component of an instrumentation amplifier. In addition, an instrumentation amplifier should possess the capability of adjustable gain. Furthermore, it is desirable to have a very high input impedance and very low output impedance at each input lead. It is desirable for an instrumentation amplifier to possess a higher and more stable gain, and also a higher input impedance than a basic differential amplifier. An instrumentation amplifier that possesses these basic requirements may be fabricated in the monolithic IC form as a single package. Alternatively, it may be built using three differential amplifiers and high-precision resistors, as shown in Figure 2.16b. The amplifier gain can be adjusted using the fine-tunable resistor R_2. Impedance requirements are provided by two voltage-follower type amplifiers, one for each input, as shown. The variable resistance δR_4 is necessary to compensate for errors due to unequal common-mode gain. Let us first consider this aspect and then obtain an equation for the instrumentation amplifier.