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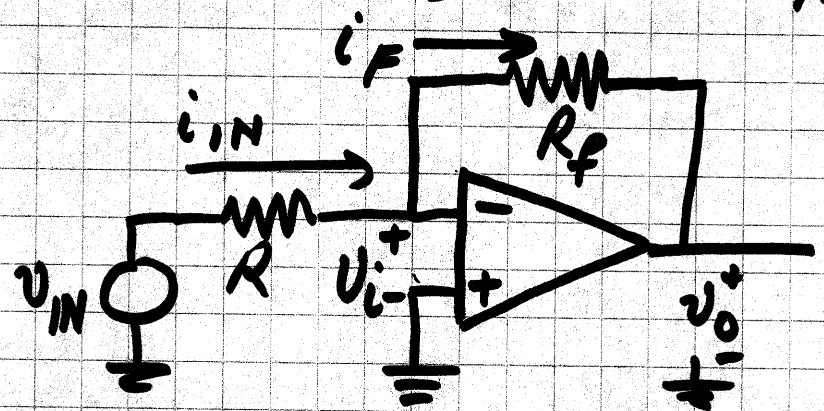
NEGATIVE FEEDBACK & the VIRTUAL S.C.



Virtual SC $\longrightarrow$ $v^- = v^+ = 0$

If $v^- \to +ve$, $\quad v_i \to +ve$ & $v_o \to -Av_i \longrightarrow (-\infty)$

$\therefore \; i_F = i_{IN} \longrightarrow \dfrac{v_i - (-Av_i)}{R_f} \approx \dfrac{Av_i}{R_f} \longrightarrow (\infty)$

$\therefore \; v_i = v_{IN} - i_{IN} R \longrightarrow -\infty$

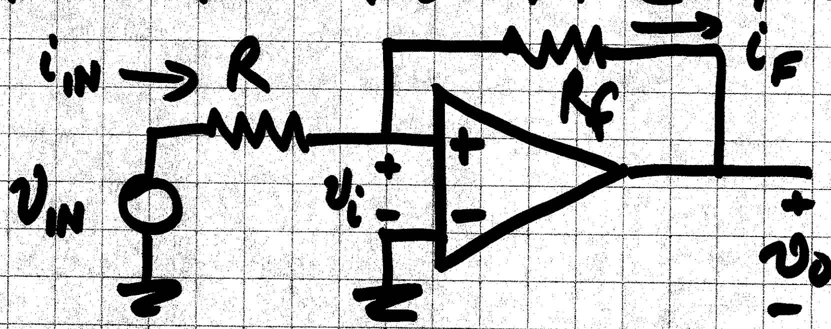
i.e. VERY large op amp gain has effect of "over-reacting" to any changes in v_i, i.e. v^-

Small +ve v^- $\longrightarrow$ large tendency to negative v^-

$\|^y$ " −ve " $\longrightarrow$ " " " positive "

i.e. v_i held to zero $\longrightarrow$ virtual short circuit.

Note for POSITIVE FEEDBACK



If $v_i > 0$ $\quad v_o = A v_i \longrightarrow +\infty$ Limited at $+V_S$

$i_F < 0 \qquad i_F = i_{IN} = \dfrac{V_S - v_{IN}}{R + R_F}$

$$\therefore\ v_i = v_{IN} + R\,\frac{(V_S - v_{IN})}{R + R_F} = \underbrace{\frac{R_F}{R + R_F}\, v_{IN} + \frac{R}{R + R_F}\, V_S}_{>0}$$

$$\therefore\ v_o \Rightarrow A v_i \longrightarrow V_S$$

(consistent)

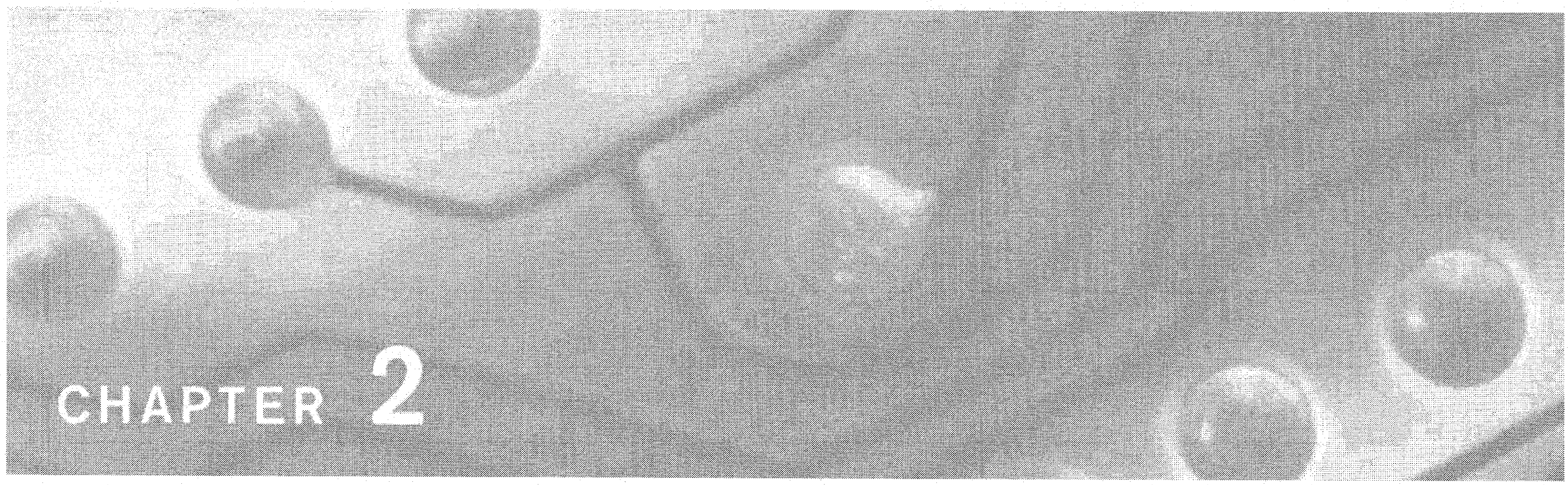
Positive feedback $\longrightarrow$ large A drives v_o to $\pm V_S$

ECE321 ELECTRONICS I
FALL 2006

PROFESSOR JAMES E. MORRIS

Lecture 3

3rd October, 2006



Operational Amplifiers (Real)

2.4 Difference Amplifier

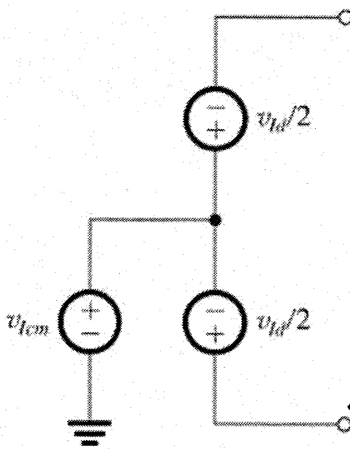
2.5 Finite Open-Loop Gain & Bandwidth

2.6 Large Signal (Saturation & Slew Rate)

Remember Common Mode & Difference Amp!

$$v_O = A_d\, v_{Id} + A_{cm} v_{Icm}$$

A_{cm} — Ideally 0



$v_{I1} = v_{Icm} - v_{Id}/2$

& $v_O = A_d (v^+ - v^-)$

$v_{Id} = v_{I2} - v_{I1}$

$v_{Icm} = \frac{1}{2}(v_{I2} + v_{I1})$

$v_{I2} = v_{Icm} + v_{Id}/2$

Define $\text{CMRR} = 20 \log_{10} \dfrac{|A_d|}{|A_{cm}|}$

Figure 2.15 Representing the input signals to a differential amplifier in terms of their differential and common-mode components.

Common Mode Rejection Ratio

Difference Amplifier : Assume $A=\infty$

We are interested in non-ideal common-mode effects here.

Solve from first principles: $v^- = v^+$

$$v_{I1} + \frac{R_1}{R_1+R_2}(v_O - v_{I1}) = \frac{R_4}{R_3+R_4} v_{I2}$$

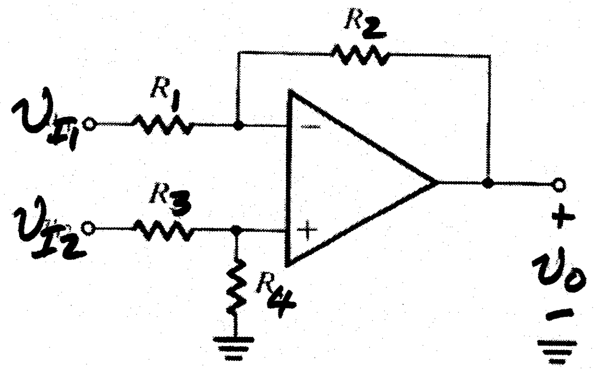


Figure 2.16 A difference amplifier.

& solve $\Longrightarrow$ $v_O = \frac{R_4}{R_1} \cdot \frac{R_1+R_2}{R_3+R_4} v_{I2} - \frac{R_2}{R_1} v_{I1}$

To work as a difference amplifier, we need

$$\frac{R_4}{R_1} \cdot \frac{R_1+R_2}{R_3+R_4} = \frac{R_2}{R_1}, \text{ ie. } \frac{R_3+R_4}{R_4} = \frac{R_1+R_2}{R_2}$$

ie. $R_3/R_4 = R_1/R_2$ to get $v_O = \frac{R_2}{R_1}(v_{I2} - v_{I1})$

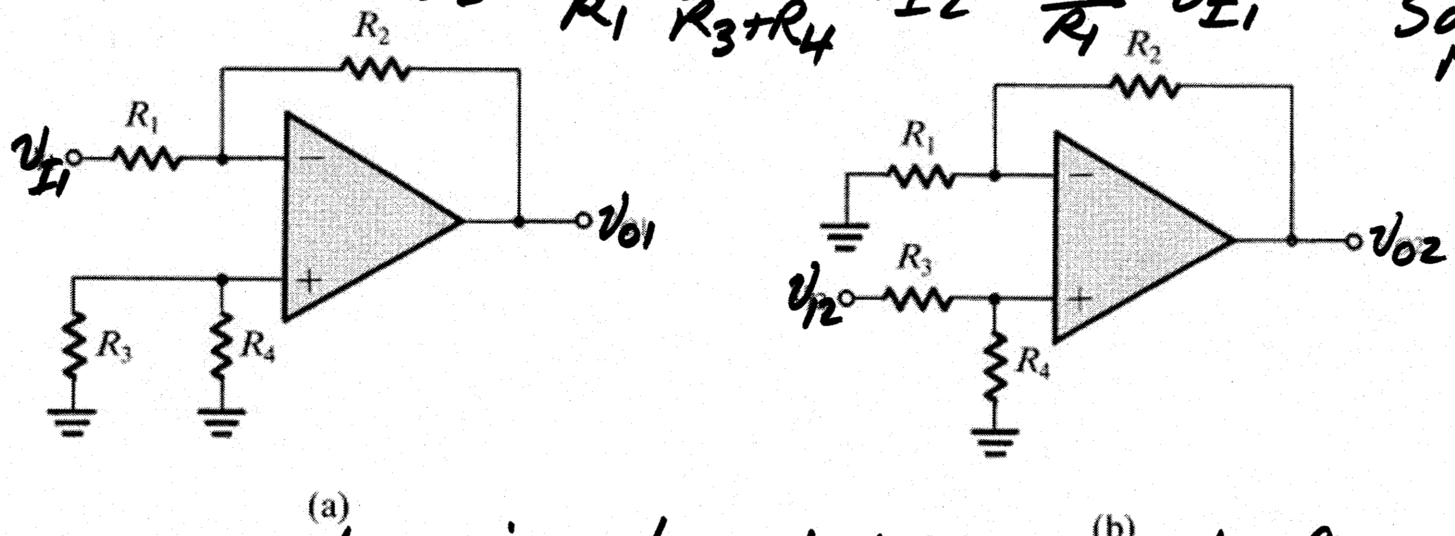
OR do this by superposition (skipped some algebra)

$$v_{O1} = -\frac{R_2}{R_1} v_{I1} \qquad v_{O2} = \left(1 + \frac{R_2}{R_1}\right) \frac{R_4}{R_3+R_4} v_{I2}$$

$$= \frac{R_4}{R_1} \frac{R_1+R_2}{R_3+R_4} v_{I2}$$

$$\therefore v_O = v_{O1} + v_{O2} = \frac{R_4}{R_1} \frac{R_1+R_2}{R_3+R_4} v_{I2} - \frac{R_2}{R_1} v_{I1}$$

Same result!!



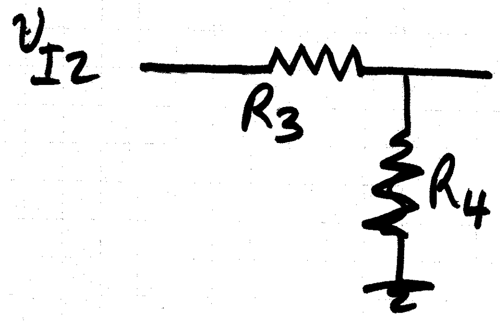
So superposition is not as obvious as brute force, but probably easier, and <u>definitely</u> less error-prone.

Note the concept here — combine inv & non-inv amp's

Figure 2.17 Application of superposition to the analysis of the circuit of Fig. 2.16.

But non-inv gain $(1 + R_2/R_1) >$ inv gain R_2/R_1

$\therefore$ Reduce non-inv gain by voltage divider & make:



$$\frac{R_4}{R_3+R_4}\, v_{I2} \quad \times \left(1+\frac{R_2}{R_1}\right)$$

equal to $\dfrac{R_2}{R_1}$

$$\frac{R_4}{R_3+R_4}\;\frac{R_1+R_2}{R_1} = \frac{R_2}{R_1}$$

$\longrightarrow$ same result $\dfrac{R_3}{R_4} = \dfrac{R_1}{R_2}$

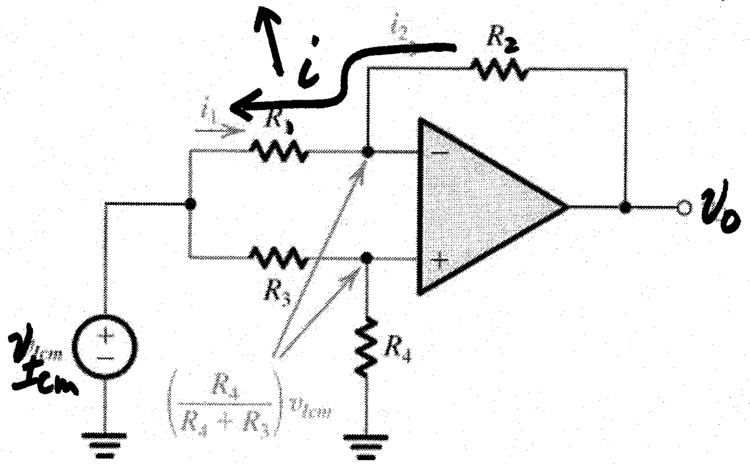
Often make $R_1 = R_3$ and $R_2 = R_4$

Back to CMRR — need A_{cm}

Voltage divider: $v^+ = \frac{R_4}{R_3+R_4} v_{Icm} = v^-$

$\therefore v_{Icm} + R_1\left(\frac{v_O - v_{Icm}}{R_1+R_2}\right) = \frac{R_4}{R_3+R_4} v_{Icm}$

$\uparrow i$



Rearrange to get

$$\frac{v_O}{v_{Icm}} = \frac{R_4}{R_1}\cdot\frac{R_1+R_2}{R_3+R_4} - \frac{R_2}{R_1} \xRightarrow{\frac{R_4}{R_3}=\frac{R_2}{R_1}} 0$$

i.e. CMRR $\rightarrow \infty$ if diff amp condition met

Figure 2.18 Analysis of the difference amplifier to determine its common-mode gain A_{cm} ; v_O / v_{Icm}.

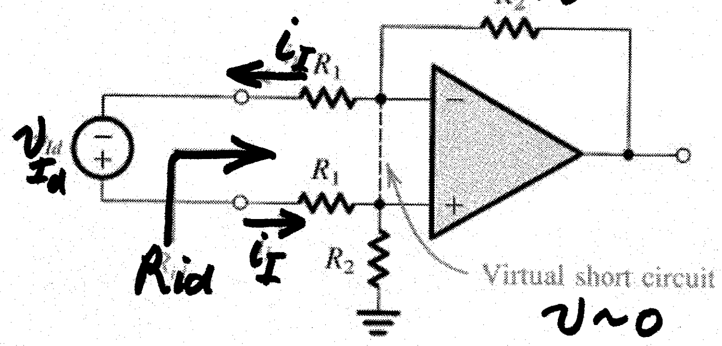
Also want high R_{in} for difference signal

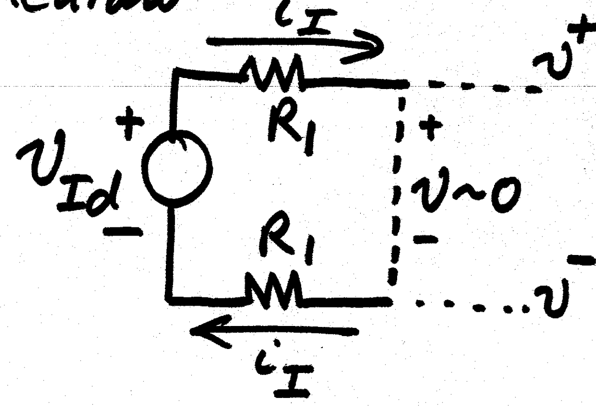
$R_{id} = v_{Id}/i_I$ as shown

Redraw

& use virtual S.C. again

($R_3 = R_1$, $R_4 = R_2$ for simplicity)





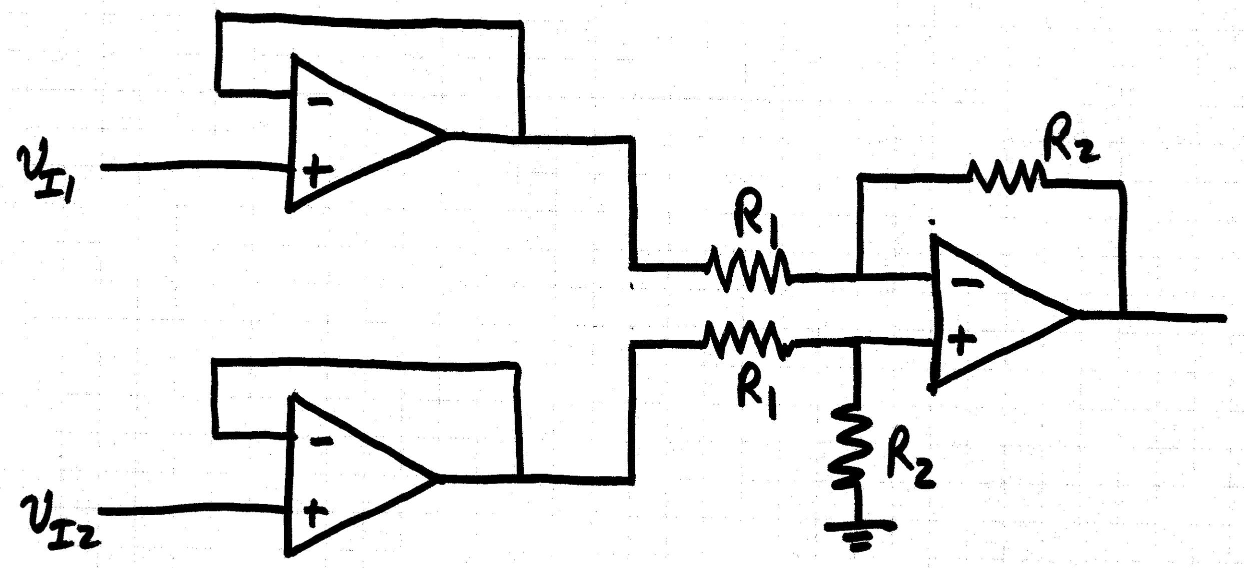
$$\therefore R_{id} = \frac{v_{Id}}{i_I} = 2R_1$$

R_{id} not as high as we would like!

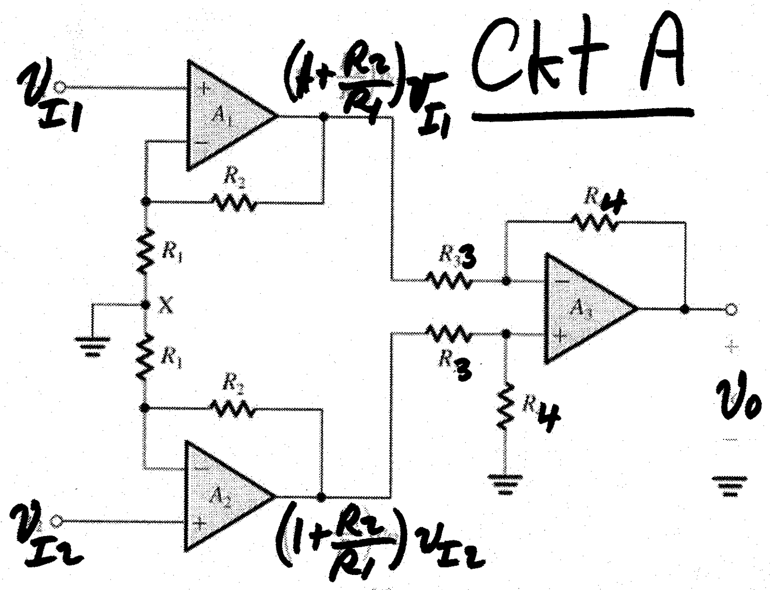
Figure 2.19 Finding the input resistance of the difference amplifier for the case $R_3 = R_1$ and $R_4 = R_2$.

Obvious solution:

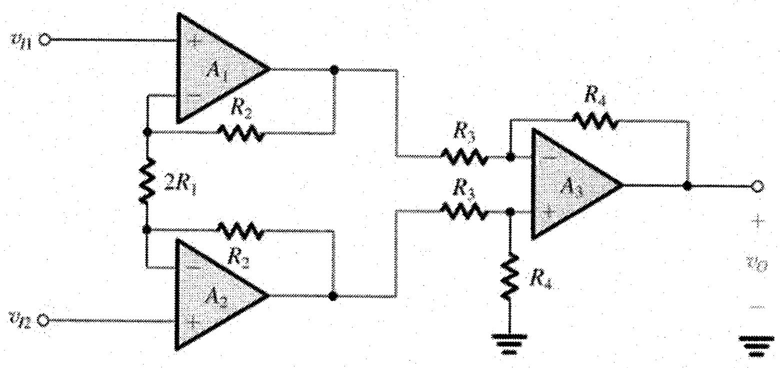
Voltage follower buffers



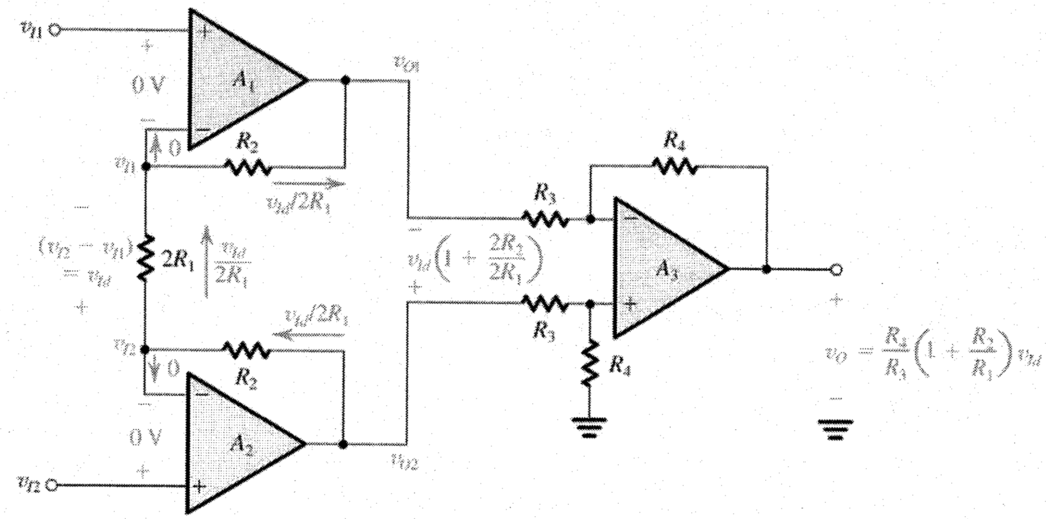
But here all the gain from one stage
Better to use cascaded lower gain stages



(a)



(b)



(c)

$$v_o = \frac{R_4}{R_3}\left(1+\frac{R_2}{R_1}\right)v_{Id}$$

But A_1, A_2 amplify CM signals $\times(1+R_2/R_1)$

Possible problems: A_1, A_2 saturation

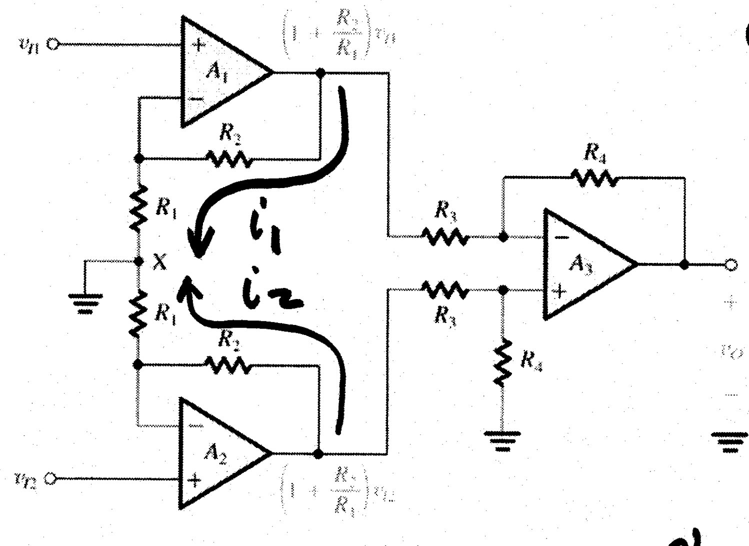
Non-ideal 3rd stage transmit CM signal

Figure 2.20 A popular circuit for an instrumentation amplifier: **(a)** Initial approach to the circuit; **(b)** The circuit in (a) with the connection between node X and ground removed and the two resistors R_1 and R_1 lumped together. This simple wiring change dramatically improves performance; **(c)** Analysis of the circuit in' **(b)** assuming ideal op amps.

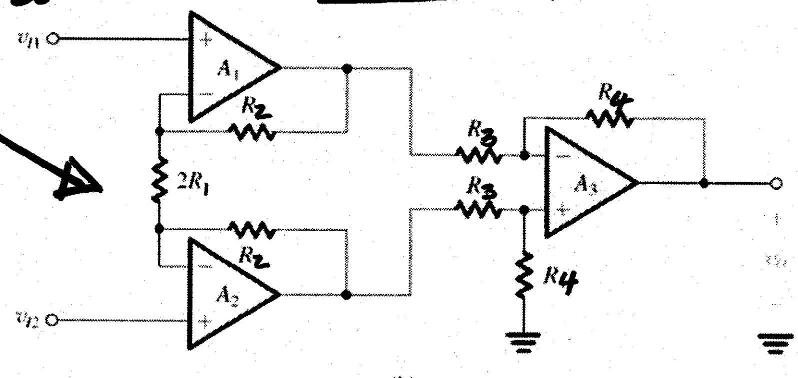
Also: Must simultaneously vary 2 R's to change gain

Drop ground connection

Ckt B



(a)

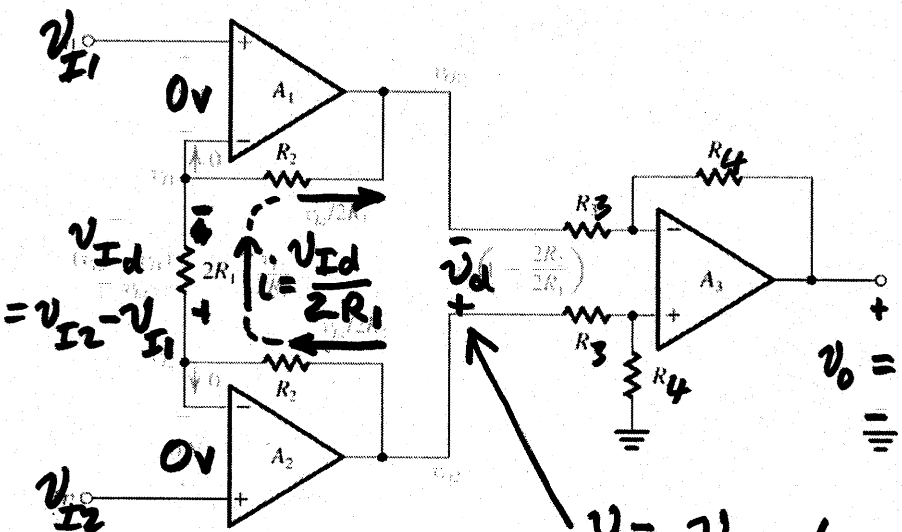


(b)

Balanced: $i_1 = i_2$

$\therefore$ No ground current

$\therefore$ No purpose



$$v_o = \frac{R_4}{R_3}\left(1 + \frac{R_2}{R_1}\right) v_{Id}$$

(c)

$$v_d = \frac{v_{Id}}{2R_1}(2R_1 + 2R_2) = \left(1 + \frac{R_2}{R_1}\right) v_{Id}$$

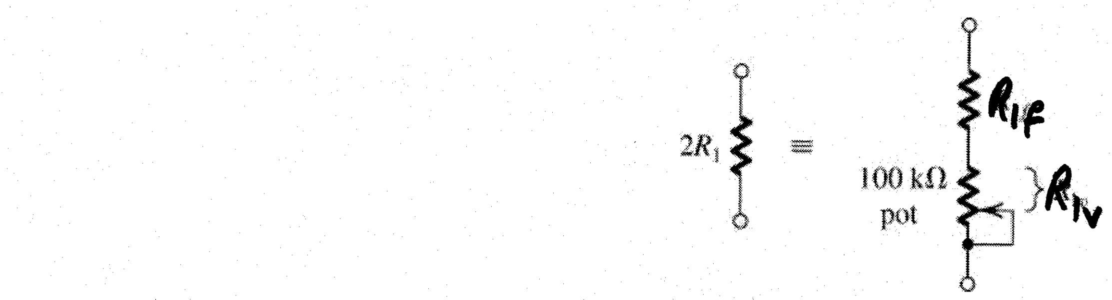
Figure 2.20 A popular circuit for an instrumentation amplifier: **(a)** Initial approach to the circuit; **(b)** The circuit in **(a)** with the connection between node X and ground removed and the two resistors R_1 and R_1 lumped together. This simple wiring change dramatically improves performance; **(c)** Analysis of the circuit in' **(b)** assuming ideal op amps.

Note: i completely defined by v_{Id} — no CM amplification

Vary gain with $2R_1$ — now single resistor

Note Example 2.3

Vary gain, bias points etc by "pot" (potentiometer)



Here $R_{1f} + R_{1v}$

$R_{1f} \longrightarrow R_{1f} + 100K\Omega$

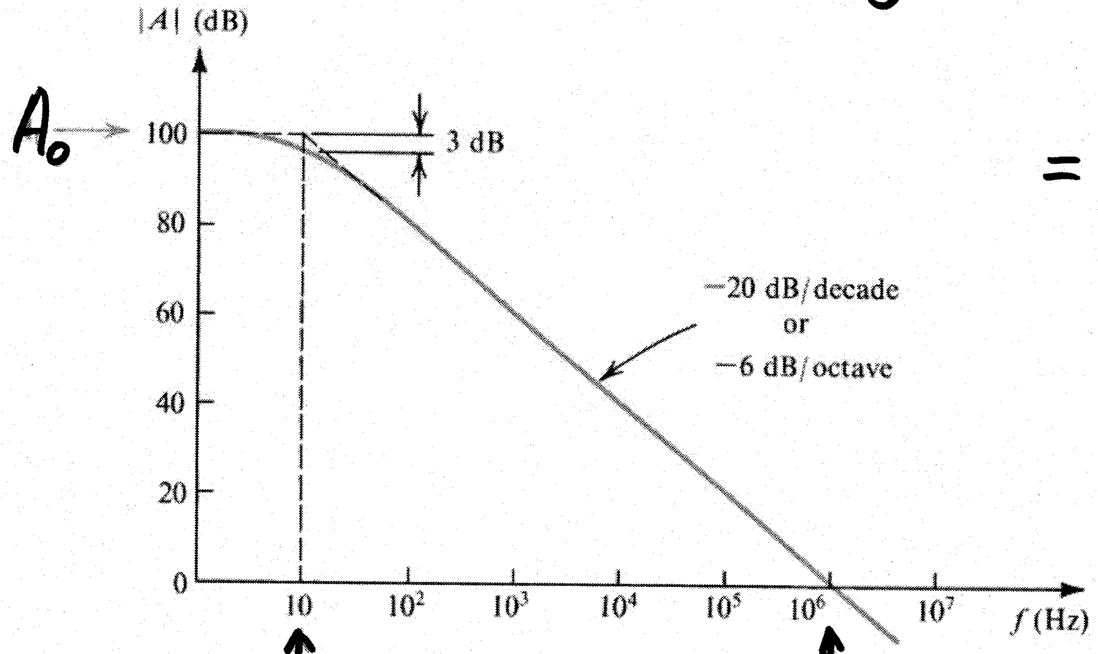
Figure 2.21 To make the gain of the circuit in Fig. 2.20(b) variable, $2R_1$ is implemented as the series combination of a fixed resistor R_{1f} and a variable resistor R_{1v}. Resistor R_{1f} ensures that the maximum available gain is limited.

Frequency-dependent gain: adapt RC concepts

A not ∞, and decreases with frequency

$$A(j\omega) = \frac{A_0}{1 + j\,\omega/\omega_b}$$

$$= \frac{A_0}{1 + j\,f/f_b}$$



f_b (3dB point)

f_t ← unity gain bandwidth if $f_t \gg f_b$

Figure 2.22 Open-loop gain of a typical general-purpose internally compensated op amp.

(Note: $G \times BW$ is Gain-Bandwidth Product)

So $1 \approx \dfrac{A_0}{\omega_t/\omega_b}$

$\omega_t \approx A_0 \omega_b$ or $f_t \approx A_0 f_b = G \times BW$

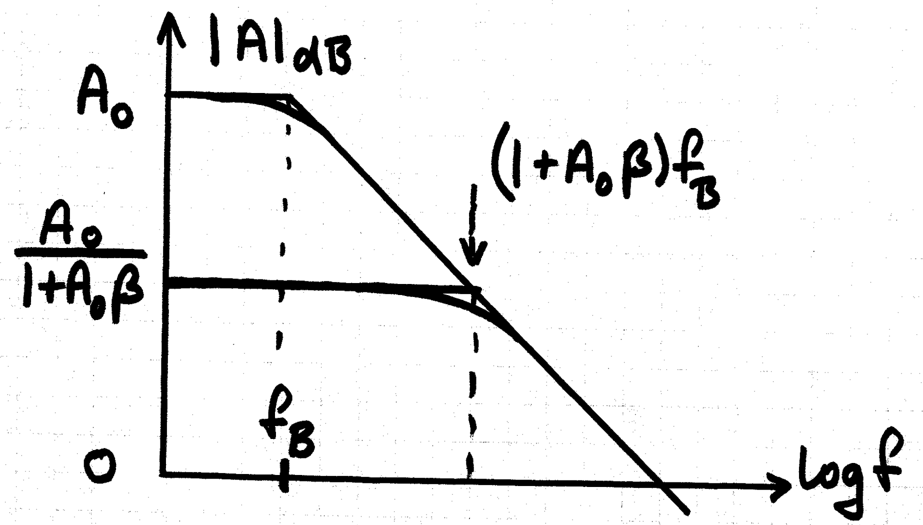
Non-Inverting Amplifier

Previous result $\dfrac{v_O}{v_I} = \dfrac{A}{1+A\beta}$ where $\beta = \dfrac{R_1}{R_1+R_2}$

$\approx 1/\beta$ for $A \to \infty$

$$\frac{v_O}{v_I} = \frac{A(j\omega)}{1+A(j\omega)\beta} = \frac{A_o/[1+j(f/f_B)]}{1+\beta A_o/[1+j(f/f_B)]} = \frac{A_o}{A_o\beta + [1+j(f/f_B)]}$$

$$= \frac{A_o}{(1+A_o\beta)+j(f/f_B)} = \frac{A_o}{(1+A_o\beta)\left(1+\dfrac{jf}{(1+A_o\beta)f_B}\right)} \xrightarrow[A_o\to\infty]{} \frac{1}{\beta}\,\frac{1}{1+j\dfrac{f}{(1+A_o\beta)f_B}}$$



Note: $G \times BW$ product

$$\frac{A_o}{1+A_o\beta}\cdot(1+A_o\beta)f_B = A_o f_B = \text{constant}$$

Independent of β

Inverting Amplifier

Previous result: $\dfrac{v_O}{v_I} = -\dfrac{R_2}{R_1+R_2}\,\dfrac{A}{1+A\beta}$ where $\beta = \dfrac{R_1}{R_1+R_2}$

$$\therefore \frac{v_O}{v_I} \longrightarrow -\frac{R_2}{R_1+R_2}\,\frac{A_0/(1+jf/f_B)}{1+\beta A_0/(1+jf/f_B)}$$

$$= -\frac{R_2}{R_1+R_2}\,\frac{A_0}{1+A_0\beta}\cdot\frac{1}{1+j\dfrac{f}{(1+A_0\beta)f_B}} \xrightarrow{A_0\to\infty} -\frac{R_2}{\cancel{(R_1+R_2)}}\,\frac{\cancel{(R_1+R_2)}}{R_1}\,\frac{1}{1+j\dfrac{f}{(1+A_0\beta)f_B}}$$

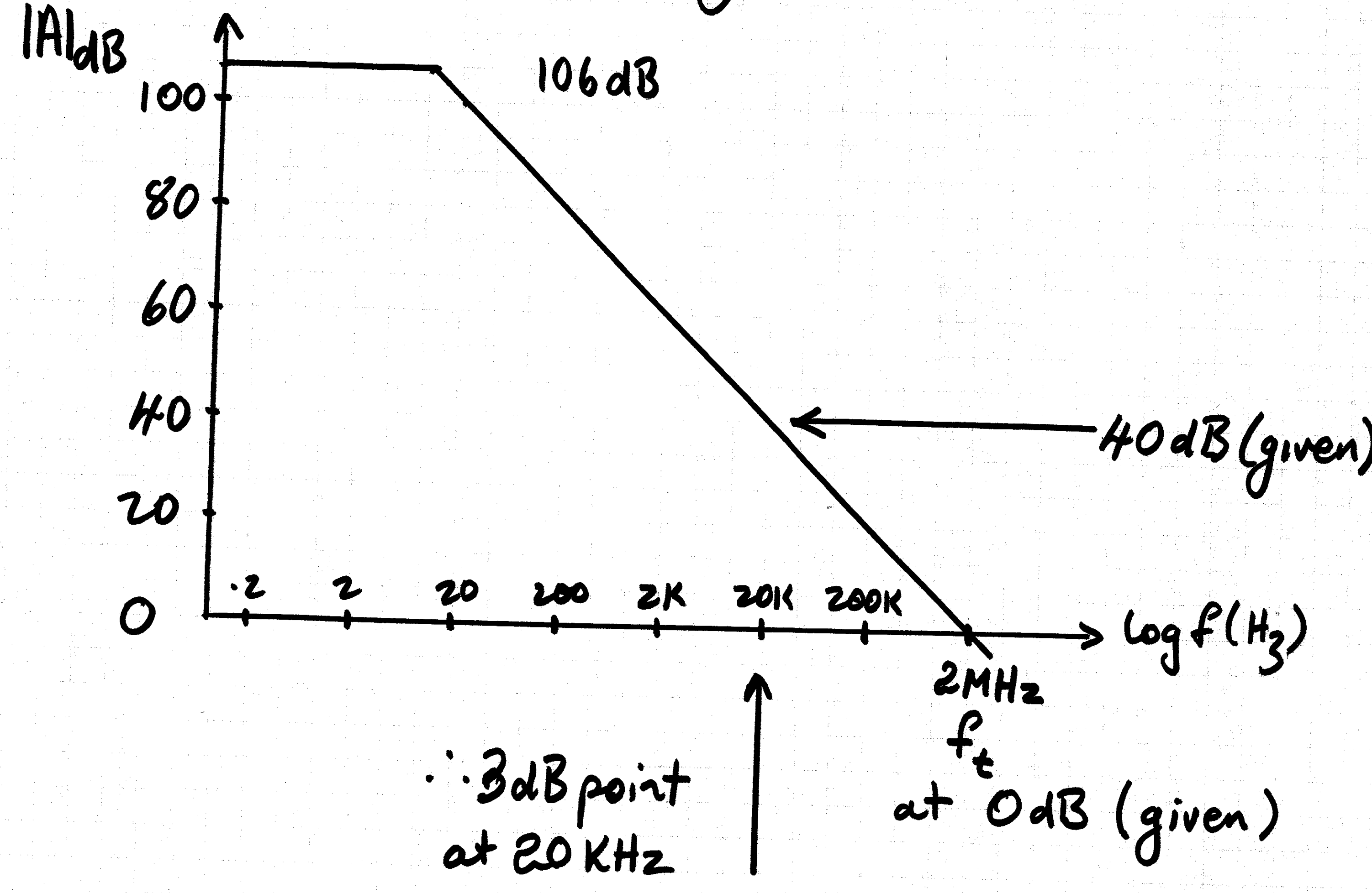
Note: $$g\times BW = \frac{R_2}{R_1+R_2}\,\frac{A_0}{1+A_0\beta}\times(1+A_0\beta)f_B$$

$$= A_0 f_B\,\frac{R_2}{R_1+R_2} = A_0 f_B\,\frac{(R_2/R_1)}{1+(R_2/R_1)}$$

Ex. 2.20 Opamp $A_o = 106\,dB$ $f_t = 2MHz$

Non-Inv amp: dc gain 100, (ie. 40dB)

Find 3dB point of CL gain



Large Signal: Saturation

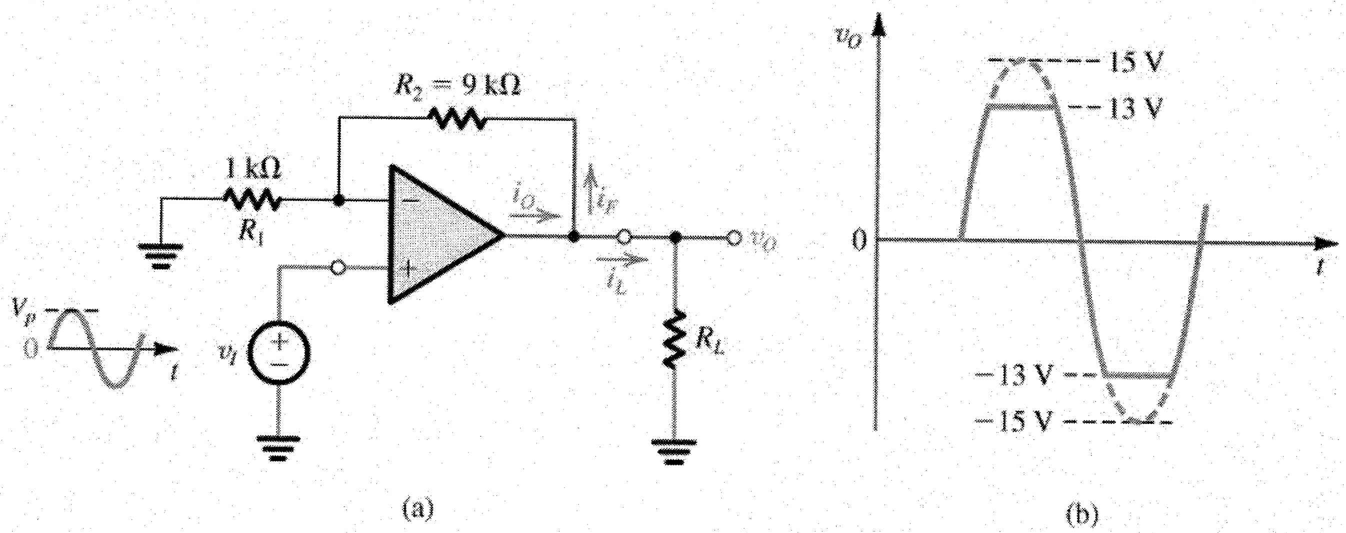
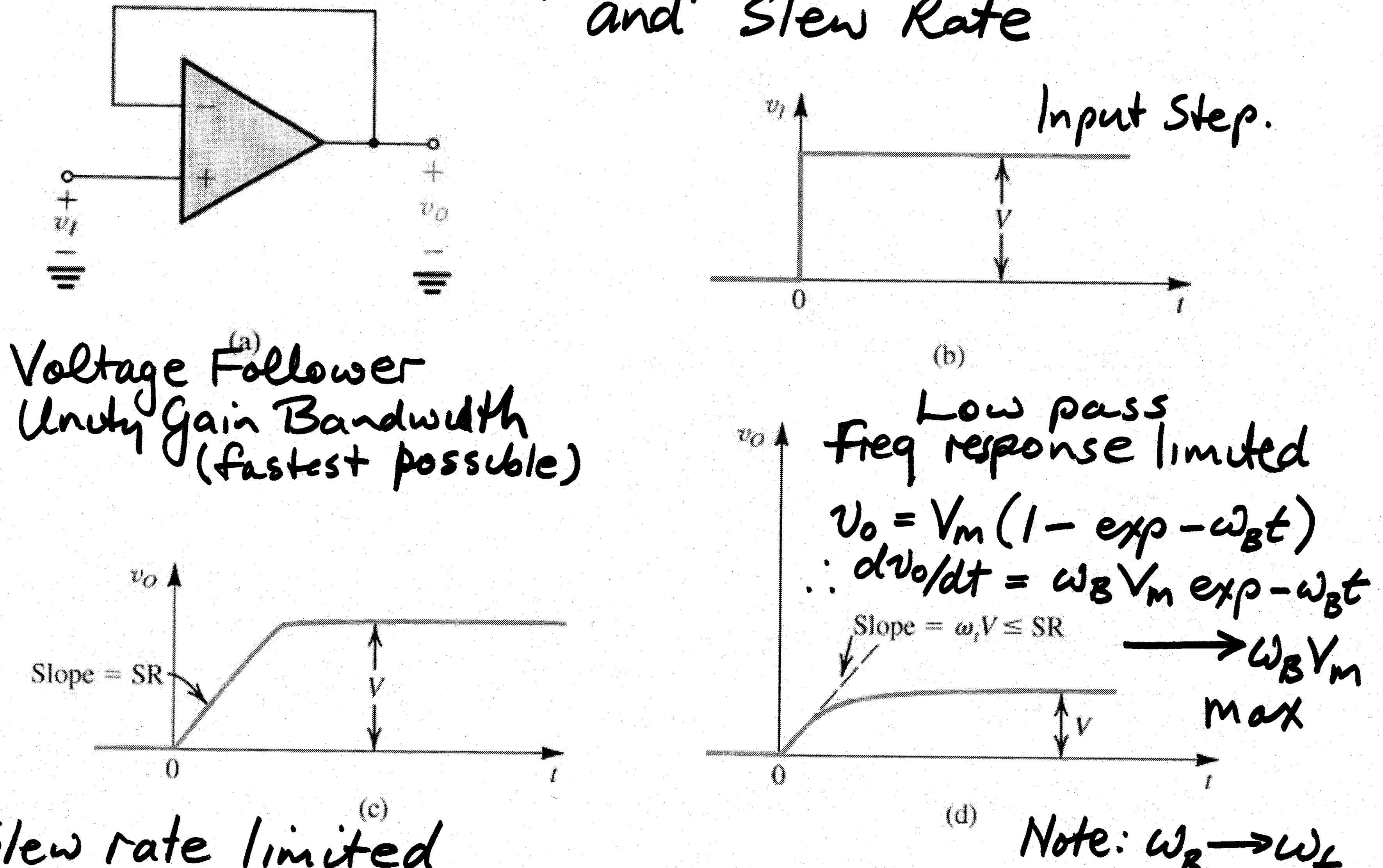


Figure 2.25 **(a)** A noninverting amplifier with a nominal gain of 10 V/V designed using an op amp that saturates at ±13-V output voltage and has ±20-mA output current limits. **(b)** When the input sine wave has a peak of 1.5 V, the output is clipped off at ±13 V.

Frequency Response Limited and Slew Rate

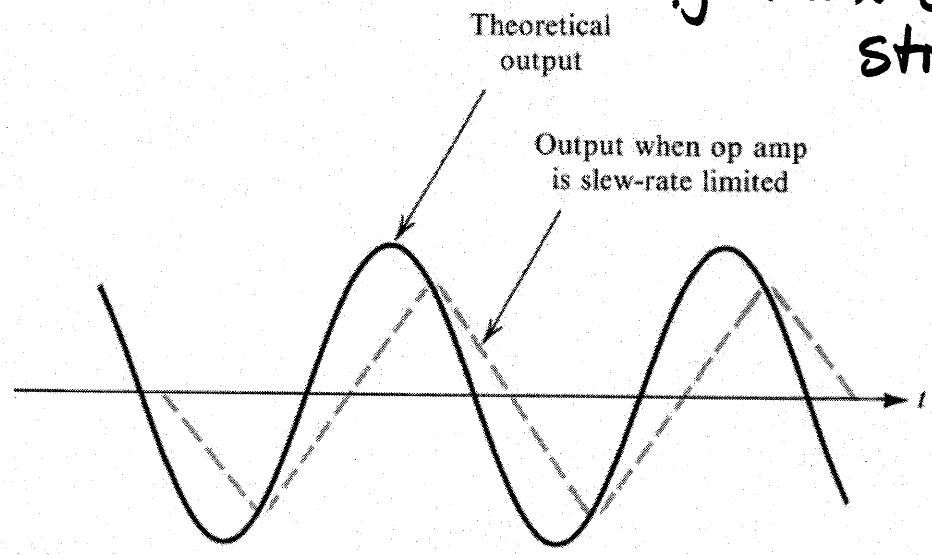


Note: $\omega_B \longrightarrow \omega_t$ for Voltage Follower

Figure 2.26 **(a)** Unity-gain follower. **(b)** Input step waveform. **(c)** Linearly rising output waveform obtained when the amplifier is slew-rate limited. **(d)** Exponentially rising output waveform obtained when V is sufficiently small so that the initial slope ($v_t V$) is smaller than or equal to SR.

Slew Rate (SR) $= \left.\frac{dv_o}{dt}\right)_{max}$

Due to internal limits e.g. max current to charge stray capacitance.



Ideal $v_o = V_m \sin \omega t$

For no SR limiting, need $\left.\frac{dv_o}{dt}\right)_{max}$ at $t=0$

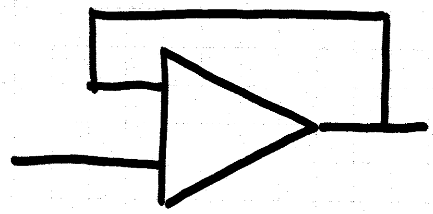
Figure 2.27 Effect of slew-rate limiting on output sinusoidal waveforms.

$= \omega V_m \cos \omega t \longrightarrow \omega V_m < SR$.

∴ Limit frequency <u>or</u> V_m

Full power BW $f_m = \frac{SR}{2\pi V_m}$

Ex. 2.21



$f_t = f_m = 1MHz$, $SR = 1v/\mu s$

Find max step input for no SR limiting

For this voltage, what is 10-90% risetime?

Find 10-90% risetime for 10x this input

Output & Input:

$V_{max}(1 - \exp -\omega_t t)$

Initial slope $\omega_t V_{max} \leq SR$

$\therefore V_{max} \leq \dfrac{1v/\mu s}{2\pi f_t} = 10^6/2\pi 10^6 \approx 0.16$ volts

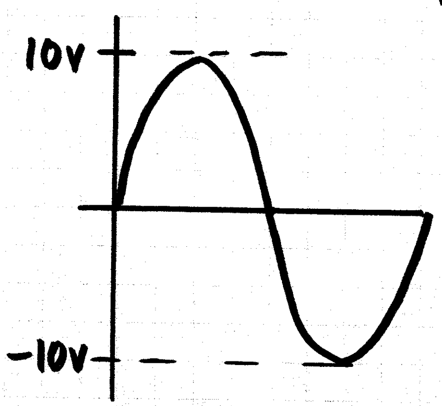
For 10%, 90% $1 - \exp -2\pi 10^6 t = 0.1, 0.9$ gives $t = 0.017, 0.37 \mu s$

$\therefore$ 10-90% risetime $= 0.37 - 0.02 \approx 0.35 \mu s$

For 1.6 volts; assume nearly fully SR limited (check)

$\therefore$ 90% = 1.44v, 10% = 0.16v $\therefore$ 1.28v linear rise at slew rate $= 1.28 \mu s$

Ex. 2.22 Opamp rated output ±10v, SR = 1v/μs. Full power BW f_M?
For input sinusoid at $5f_M$ to voltage follower, find max V_m for no SR limit.



$10v \sin\omega t \longrightarrow \omega_m 10_v \leq 1v/\mu s$ i.e. $\omega_m \leq 10^6/10 = 10^5$

$f_m = 10^5/2\pi \approx 16$ KHz

For $5f_m$ & unity gain ωV_m still 10^6

$\underset{5\omega_m}{\omega} \; V_m \quad \therefore 10v/5 = 2v$

Summary

- Difference (instrumentation) amplifier
  - A_{cm} & A_d definitions; CMRR
  - Difference amplifier gain (formula or derive)
  - CMRR → 0 for ideal difference amp condition
  - High R_{id} ; (R_{id} and virtual SC)
- Finite A_0 & bandwidth
  - $A(j\omega) = A_0 /(1+j\omega/\omega_b)$
  - Derive $A_{CL}(j\omega)$ for inv & non-inv amplifiers
  - Gain x bandwidth product; unity gain BW
- Large signal saturation & slew rate
  - Full power BW amplitude dependent