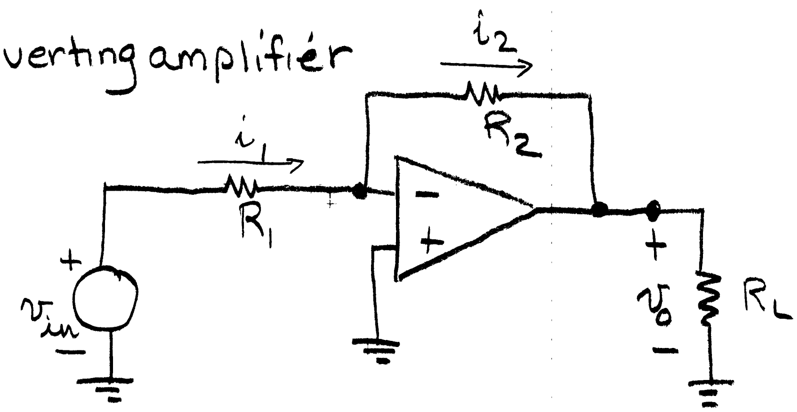
Chapter 2 Operational Amplifiers

Basic inverting amplifier



Negative feedback
output returned in opposition to input.

Assumptions: $I_{in} = 0$

$V_+ = V_-$ (virtual short)

summing point constraint

$$i_1 = \frac{v_{in}}{R_1}$$

since $I_{in} = 0$

$$i_2 = i_1$$

sign due to current direction

$$v_o = V_- - i_2 R_2$$

voltage at inverting input

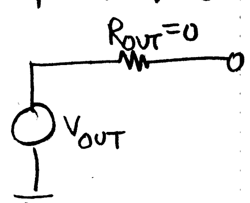
$$v_o = -i_2 R_2 = -i_1 R_2 = -\left(\frac{v_{in}}{R_1}\right) R_2$$

$$\frac{v_o}{v_{in}} = -\frac{R_2}{R_1}$$

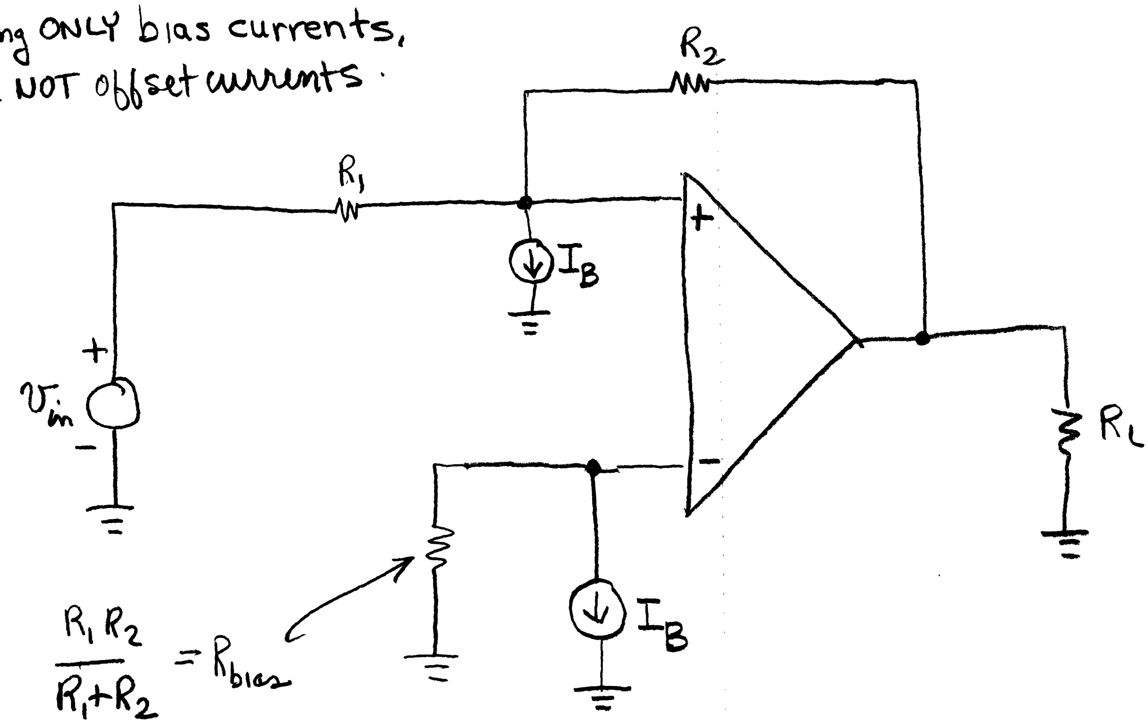
Input impedance $= R_1$

Since output independent of R_L we conclude output is a perfect voltage source with $R_{out} = 0$

i.e.

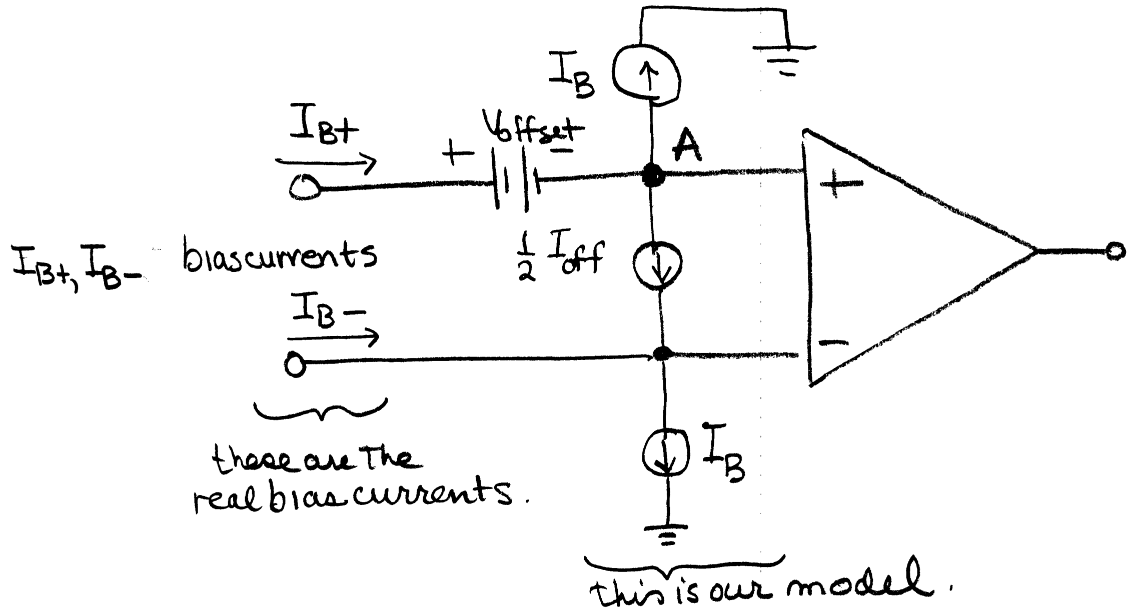


Modeling ONLY bias currents, and NOT offset currents.



Assumes we neglect v_{offset} and $\frac{1}{2} I_{off}$

2.8 DC Imperfections - Bias current cancelation



difference between biascurrents is called offset current

$$I_{off} = I_{B+} - I_{B-}$$

$$I_B = \frac{I_{B+} + I_{B-}}{2}$$

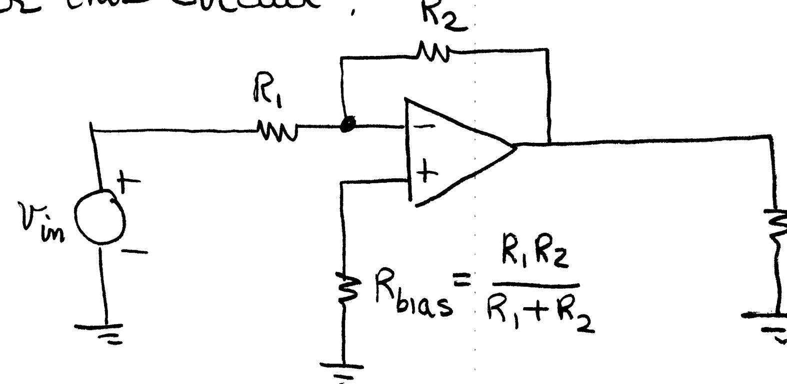
Sum currents @ A

to check this out. $I_{B+} = I_B + \frac{1}{2} I_{off} = \frac{1}{2} I_{B+} + \frac{1}{2} I_{B-} + \frac{1}{2} I_{B+} - \frac{1}{2} I_{B-}$

$$= \frac{1}{2} I_{B+} + \frac{1}{2} I_{B+} = I_{B+}$$

We can cause bias currents to cancel out with other currents.

Consider this circuit.



What does this look like in frequency domain?

$$A_{OL}(f) = \frac{A_{\phi OL}}{1 + j\, f/f_{BOL}}$$

$$A_{CL}(f) = \frac{A_{OL}(f)}{1 + \beta A_{OL}(f)} = \frac{\dfrac{A_{\phi OL}}{1 + j\, f/f_{BOL}}}{1 + \dfrac{\beta A_{\phi OL}}{1 + j\, f/f_{BOL}}}$$

$$= \frac{\dfrac{A_{\phi OL}}{1 + j f/f_{BOL}}}{\dfrac{1 + j\, f/f_{BOL} + \beta A_{\phi OL}}{1 + j\, f/f_{BOL}}} = \frac{A_{\phi OL}}{1 + \beta A_{\phi OL} + j\, f/f_{BOL}}$$

$$\frac{v_o}{v_s} = A_{CL}(f) = \frac{\dfrac{A_{\phi OL}}{1 + \beta A_{\phi OL}}}{1 + j \dfrac{f}{f_{BOL}} \dfrac{1}{(1 + \beta A_{\phi OL})}}$$

open loop dc gain.

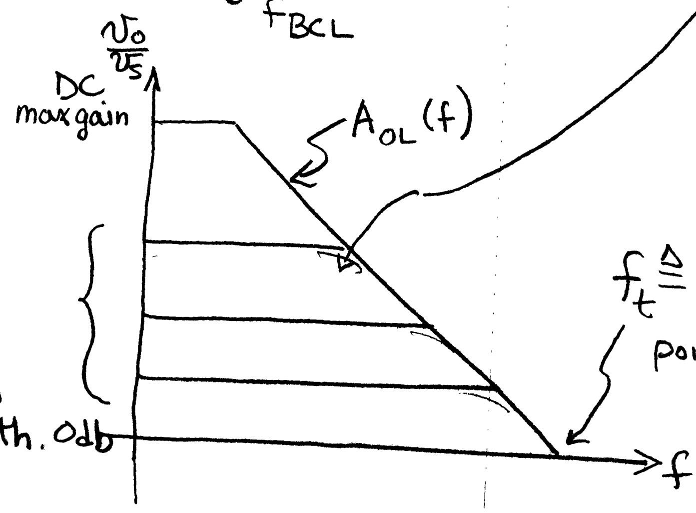
define $A_{\phi CL} = \dfrac{A_{\phi OL}}{1 + \beta A_{\phi OL}}$ closed loop dc gain less than open loop dc gain,

and $f_{BCL} = f_{BOL}(1 + \beta A_{\phi OL})$ closed loop breakpoint frequency higher

then $\dfrac{v_o}{v_s} = A_{CL}(f) = \dfrac{A_{\phi CL}}{1 + j \dfrac{f}{f_{BCL}}}$

β is function of external resistors.

$A_{\phi CL}$ decreases with β.



These are closed loop gains. As we increase gain we lose bandwidth.

This is what equation means.

$f_t \triangleq A_{\phi CL} f_{BCL}$

point at which $\frac{v_o}{v_s} = 1$ (0dB)

Why do we see this frequency response.

This is getting slightly ahead of ourselves <u>but</u>

$$A_{OL}(f) = \frac{A_{\phi OL}}{1 + j\left(\frac{f}{f_{BOL}}\right)}$$

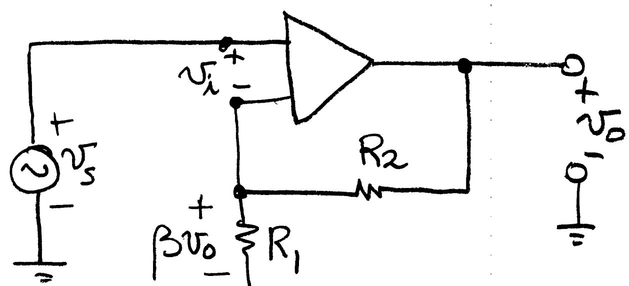
$A_{\phi OL}$ ← DC, i.e. zero frequency open loop gain, big <u>but</u> finite, typically $10^4 \le A_{\phi OL} \le 10^6$

looks like single-pole, i.e. a low-pass filter

down to 0.707 (−3dB) at f_{BOL} ← break frequency

called dominant pole assumption

Analyze non-inverting amplifier using above op-amp model



where $\beta = \frac{R_1}{R_1 + R_2}$

input loop $\quad -v_s + v_i + \beta v_o = 0$

$$v_o = A_{OL} v_i$$

eliminating v_i we get $\quad -v_s + \frac{v_o}{A_{OL}} + \beta v_o = 0$

$$v_s = \left(\frac{1}{A_{OL}} + \beta\right) v_o$$

define closed loop gain $\quad A_{CL} = \frac{v_o}{v_s} = \frac{1}{\frac{1}{A_{OL}} + \beta} = \frac{A_{OL}}{1 + A_{OL}\beta}$

Note that as $A_{OL} \to \infty \quad A_{CL} \to \frac{1}{\beta} = 1 + \frac{R_2}{R_1}$

2.9 Computer analysis of op-amp circuits

often do bandwidth calculations

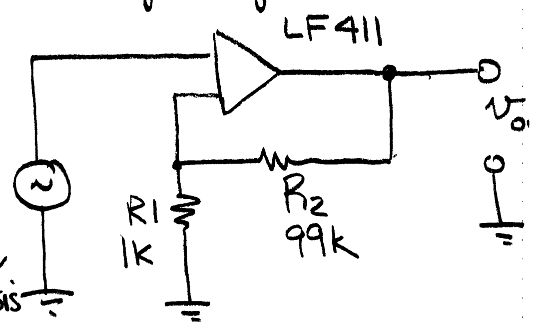
choose AC Sweep for this

use Decade sweep

specify points/decade

like a transient analysis plot (PROBE) would come up with frequency on the horizontal axis

Consider analysis of



1V = VAMPL
Not realistic
PSpice uses linear model for ac analysis

Not sure if LF411 in current PSpice library,

Other popular op amps.

UA741

LM324 (single voltage)

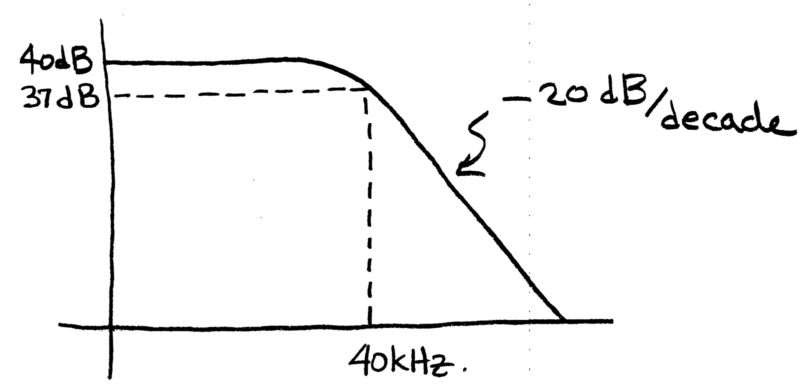
ideal gain would be

$$A_v = \frac{v_o}{v_{in}} = 1 + \frac{R_2}{R_1} = 1 + \frac{99k}{1k} = 100.$$

If we plot vertical axis in dB $20\log(100) = +40\text{dB}$

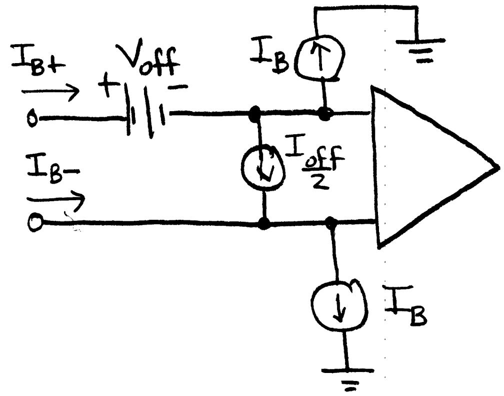
we get

−3dB =

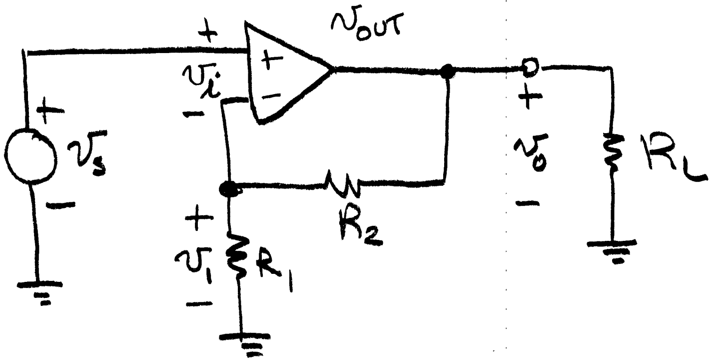


Op Amp non-ideal characteristics

1. Non-ideal R_{in}, R_{out}
2. finite, frequency dependent gain
3. output voltage limits, typically close to $\pm V_{cc}$
   ± 12 volts for 741
4. output current limits
   up to ± 25mA for 741
5. slew rate limitation $\left|\frac{dv_o}{dt}\right| \leq SR$
6. power bandwidth
   frequency range over which op amp can produce undistorted sine wave with Vpeak equal to guaranteed maximum voltage
7. input offset current and voltage model



Basic noninverting amplifier



positive output to inverting input → NEGATIVE feedback

$\Rightarrow v_i = 0$

$\Rightarrow v_1 = v_i$ by virtual short.

but $v_1 = \dfrac{R_1}{R_1 + R_2} v_o$

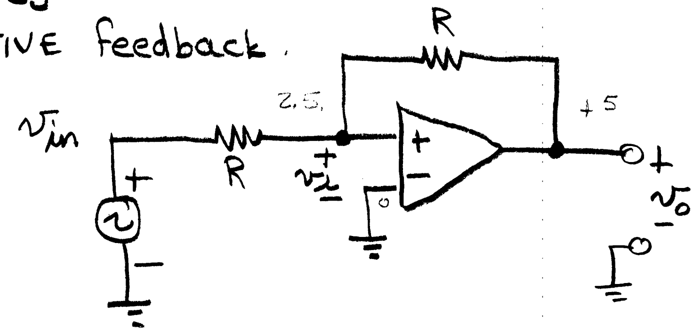
$\therefore \quad v_i = \dfrac{R_1}{R_1 + R_2} v_o \quad$ or $\quad \dfrac{v_o}{v_i} = \dfrac{R_1 + R_2}{R_1} = 1 + \dfrac{R_2}{R_1}$

$R_{in} \to \infty$

As we argued for inverting amplifier $R_{OUT} \to 0$ since v_{OUT} independent of R_L

Schmitt trigger

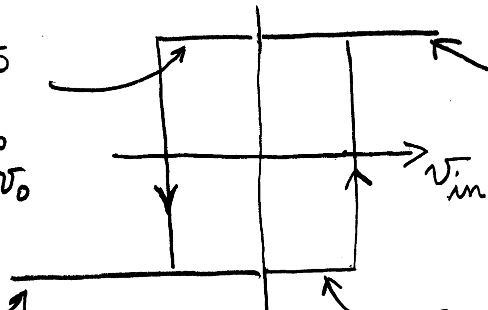
POSITIVE feedback.



system transfer function

remains +5 until input forces $+v_i$ to zero. This occurs at $v_{in} = -v_o$ or −5 volts.

$+V_{cc}$ for all positive voltages

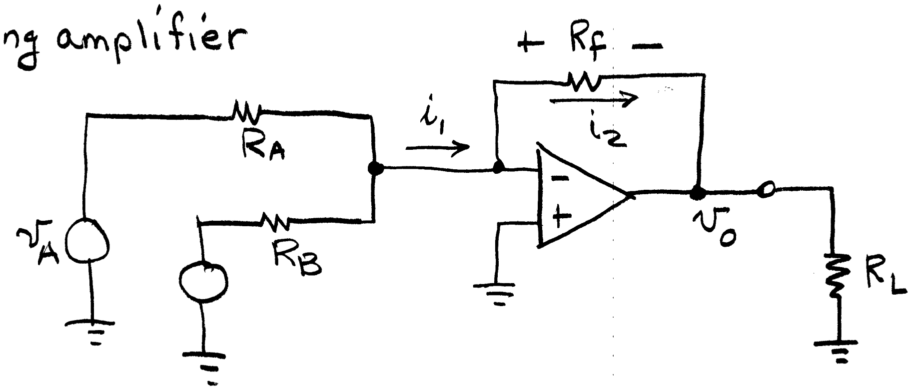


$-V_{cc}$ for all negative voltages.

remains at −5 until v_{in} forces $+v_i$ back to zero, i.e. at $v_{in} = -(-5) = +5$

Good circuit to square up waveforms.

Summing amplifier



Same as before $V_- = V_+ = 0$.

$$i_1 = \frac{v_A}{R_A} + \frac{v_B}{R_B}$$

$$i_2 = i_1$$

$$v_o = V_- - i_2 R_f = 0 - i_2 R_f = -\left(\frac{v_A}{R_A} + \frac{v_B}{R_B}\right) R_f$$

$$v_o = -\frac{R_f}{R_A} v_A - \frac{R_f}{R_B} v_B$$