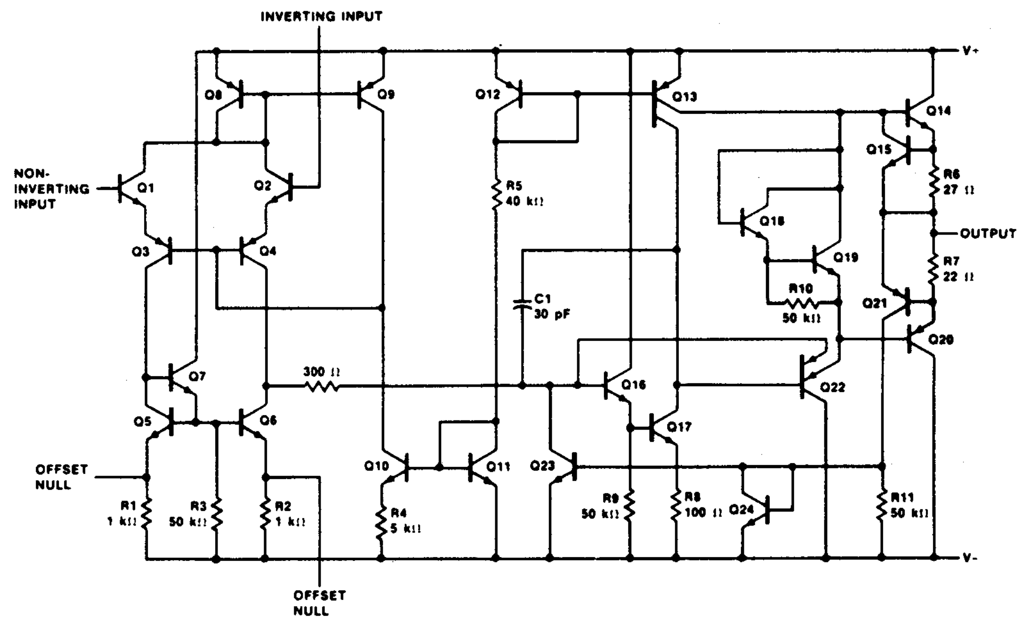
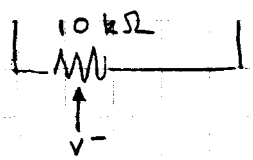
2.5 DC Problems – Offset Voltage and Input Bias Current

Offset voltage

- Even when the two op amp inputs are shorted together, there is a non zero voltage at the output. It's due to component mismatches in the differential stage at the input.

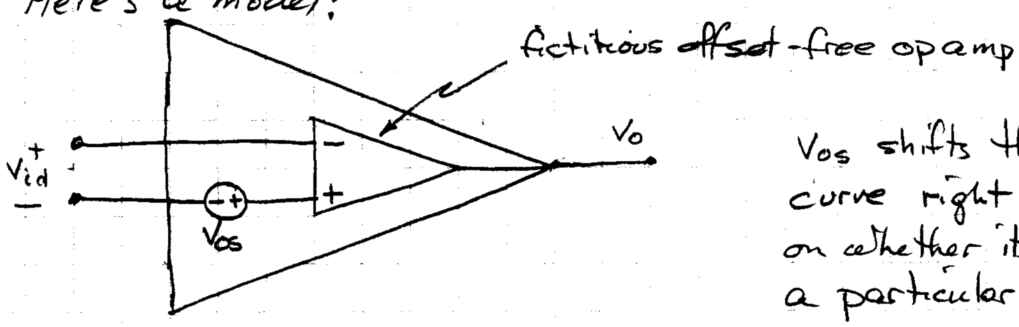
Equivalent Circuit 741





Some chips have offset nulling pins, but there can be drift.

- Here's a model:

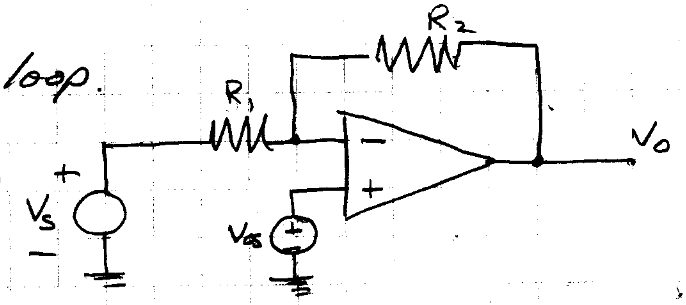


Vos shifts the Vo vs Vid curve right or left, depending on whether it's +ve or −ve for a particular chip.

For a 741, $-3\,mV \le V_{os} \le 3\,mV$, typically $\pm 0.8\,mV$

- *Effect on closed loop.*

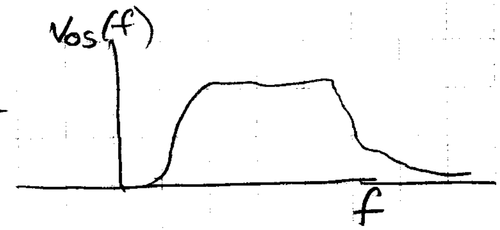
Inverting amp



By superposition, $V_o = -\frac{R_2}{R_1} V_s + \left(1 + \frac{R_2}{R_1}\right) V_{os}$

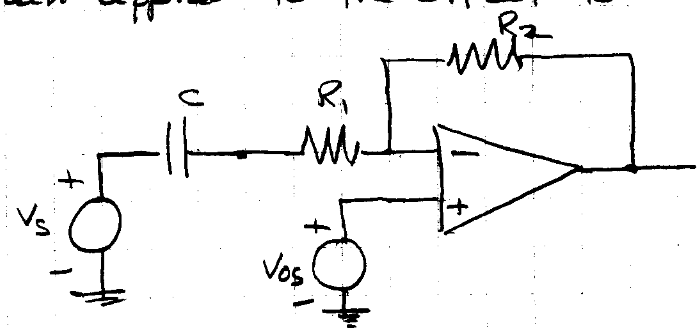
Note if gain is large, the 1 mV or so offset produces large output. Equivalent to having V_s almost as small as V_{os}

- Do we care?



- Not if the signal has no dc component

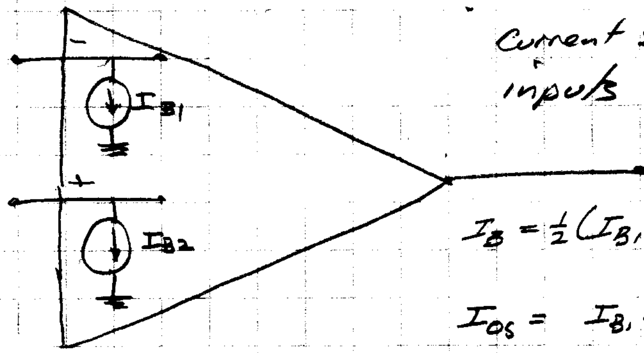
We can simply capacatively couple.
Not only do we separate signal and offset, but the gain applied to the offset is cut drastically



Signal gain is $-R_2/R_1$ as before. Cap is open circuit at dc, so offset sees a unity gain follower,

- Input bias currents are needed to make Q1 and Q2 work. They're very small (typ 30 nA for 741, down to pA for FET op amps) but still affect behaviour because of large gains.

- Model



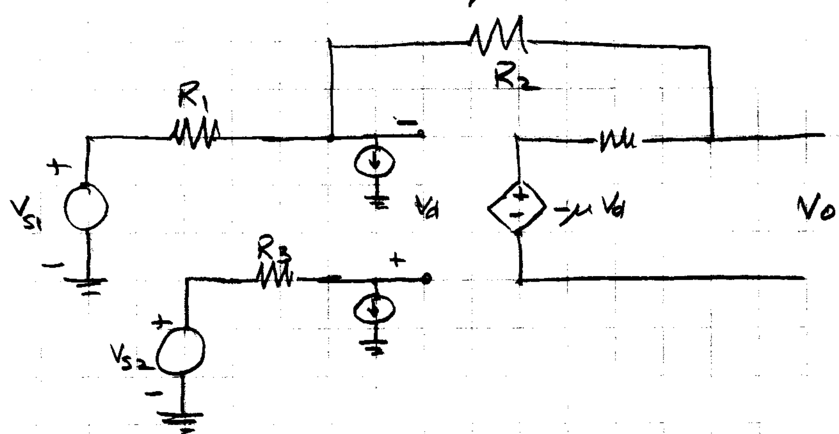
Current sources, independent of inputs

$I_B = \frac{1}{2}(I_{B1} + I_{B2})$ input bias current

$I_{OS} = I_{B1} - I_{B2}$ input offset current

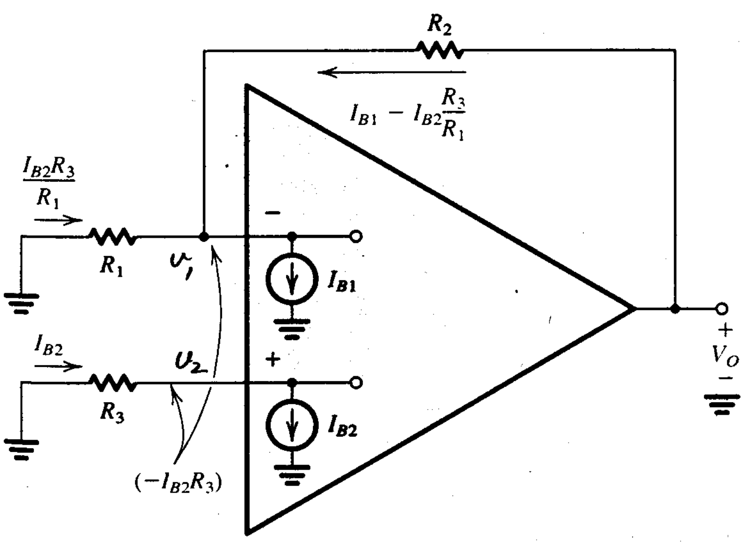
$I_{OS} \ll I_B$ (factor of 10 or so)

- Effect on closed loop



a general configuration for differential amp

- by superposition, ignore V_{S1}, V_{S2}.



- voltages at two inputs are equal, by neg feedback

- so $v_1 = v_2 = -I_{B2} R_3$

- so feedback current is
$I_{B1} - \frac{I_{B2} R_3}{R_1}$

- so output voltage is

$$V_o = -I_{B2}R_3 + R_2\left(I_{B1} - \frac{I_{B2} R_3}{R_1}\right)$$
$$= R_2 I_{B1} - R_3\left(1 + \frac{R_2}{R_1}\right) I_{B2}$$

- so $V_0 = R_2 I_{B1} - R_3(1+\frac{R_2}{R_1}) I_{B2}$ makes output voltage significant (though a smaller problem than offset voltage).

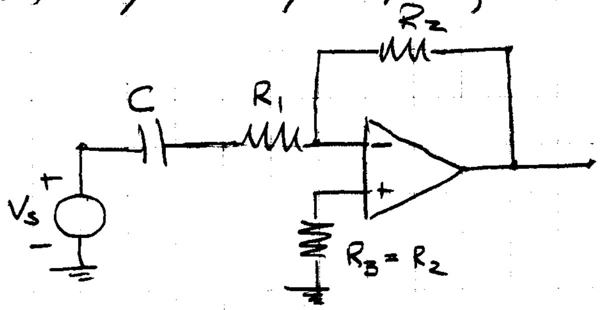
  if 100 nA, then $R_2 I_{B1} \sim 10\,mV$ (for $R_3 = 0$, $R_2 = 100\,k\Omega$)

- Since I_{B1}, I_{B2} within about 10% of each other, try to cancel effects:

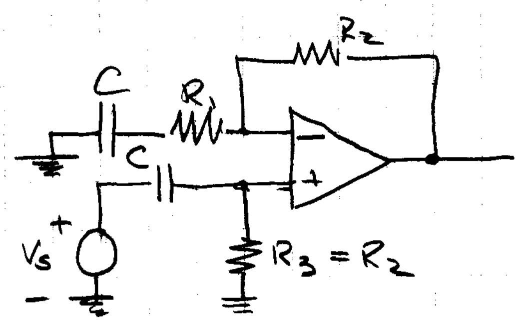
$$R_2 = R_3\left(1+\frac{R_2}{R_1}\right) \Rightarrow R_3 = \frac{R_1 R_2}{R_1+R_2} = R_1 \| R_2$$

  This is the guideline given earlier, p2.3. Cuts error by about order of magnitude.

- If capacitively coupled, note cap is short for signal, open for dc



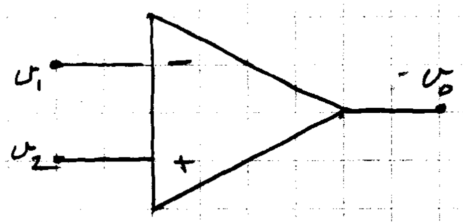
inverting



non inverting

- resistance, (at dc) seen by -ve terminal is R_2
- need R_3 in non inverting to provide dc path for I_{B2}.
- if R_3 were absent in the noninverting configuration, the bias current I_{B2} would slowly charge the cap, adding an increasing dc voltage to v_2; by negative feedback, v_1 would track v_2, and both voltages would together slide out of a useful range and creep toward the supply voltage.

2.6 Common Mode Rejection



Our basic model is

$$v_o = A(v_2 - v_1)$$

but the subtraction isn't perfect

Suppose it's really $\quad v_o = A\left((1+\epsilon_2)v_2 - (1+\epsilon_1)v_1\right) \qquad |\epsilon_1|, |\epsilon_2| \ll 1$

Since it's not a pure difference anymore, convenient to change variables:

$$v_{id} = v_2 - v_1, \qquad v_{Icm} = \tfrac{1}{2}(v_1 + v_2)$$

$$v_1 = v_{Icm} - \tfrac{1}{2}v_{id}, \qquad v_2 = v_{Icm} + \tfrac{1}{2}v_{id}$$

Substitute to get

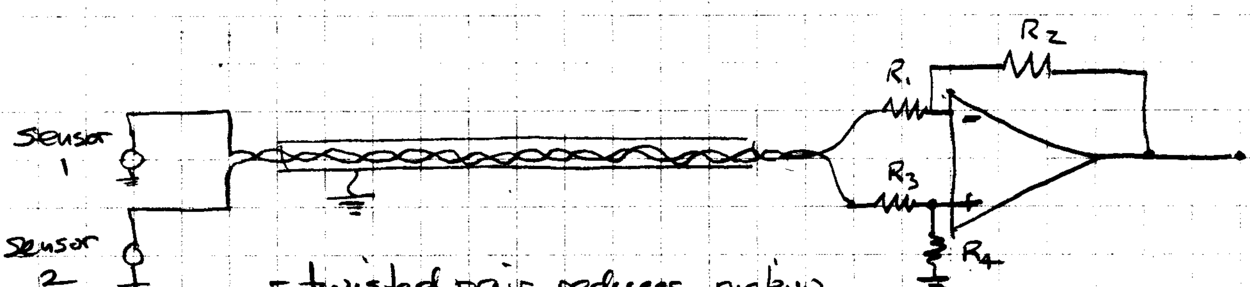
$$v_o = A\left(1 + \tfrac{\epsilon_2}{2} - \tfrac{\epsilon_1}{2}\right)v_{id} + A(\epsilon_2 - \epsilon_1)\,v_{Icm}$$

$$\approx A\,v_{id} + A_{cm}\,v_{Icm}$$

Common mode rejection ratio $\quad CMRR = \dfrac{A}{A_{cm}}$

typically 80 to 100 dB; ie, $\quad A_{cm} = 10^{-4}$ to $10^{-5}\,A$

Text shows that the CMRR also applies to closed loop.

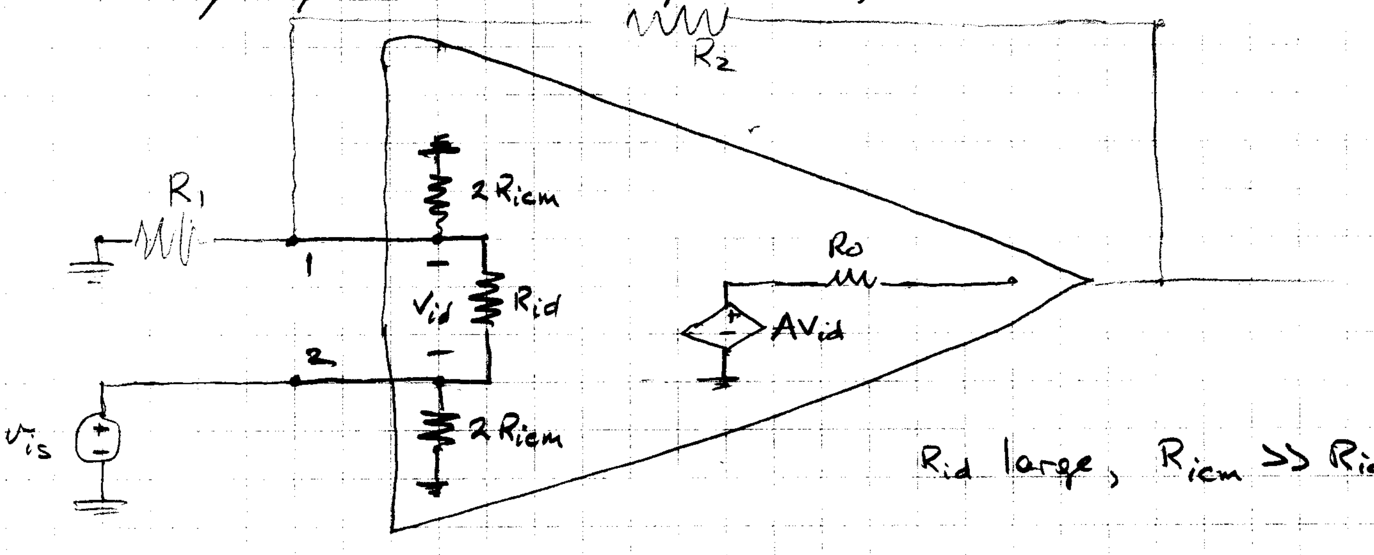


- twisted pair reduces pickup
- grounded shielding reduces pickup

If most of the pickup appears as common mode, it will be suppressed relative to difference by CMRR. Since not exact, some pickup will be differential and get through.

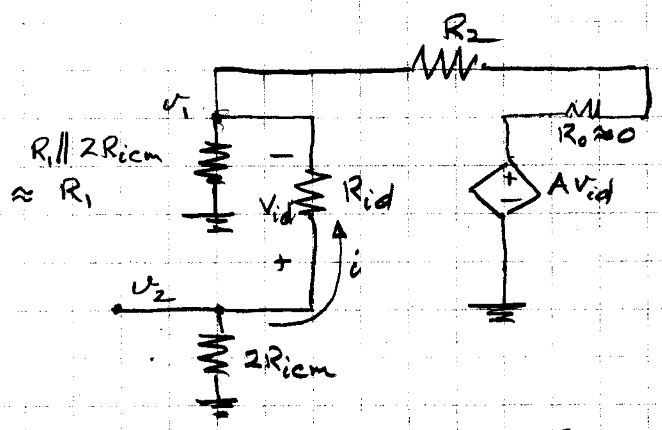
2.7 Input and Output Resistances

- Input Resistance. Text provides model and results of analysis, but not the analysis itself



R_{id} large, $R_{icm} \gg R_{id}$

- R_{cm} split into two resistors in parallel to ground, each $2R_{icm}$
- For inverting, input resistance is R_1, no significant change
- For noninverting, equivalent circuit is



$R_1 \ll 2R_{icm}$

$R_o \approx 0$

$R_{in} = 2R_{icm} \| \;?$ — determine this

do current balance node 1

$$\frac{v_2 - iR_{id}}{R_1} = i + \frac{AiR_{id} - (v_2 - iR_{id})}{R_2}$$

$$v_2 = i \cdot \frac{R_2R_{id} + R_2R_1 + AR_1R_{id} + R_1R_{id}}{R_1 + R_2}$$

$R_2 \ll AR_{id}$

$$= iR_{id}\left(\frac{R_1 + R_2 + AR_1}{R_1 + R_2}\right)$$

$$= iR_{id}(1 + A\beta)$$

so $R_{in} = 2R_{icm} \,\|\, R_{id}(1+A\beta)$

$\approx 2R_{icm}$ for low freq, low gain β^{-1}

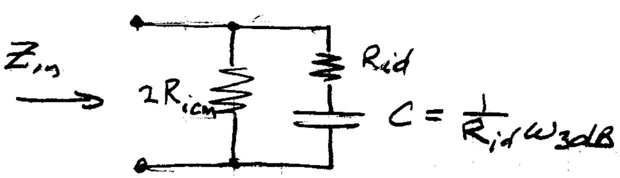
$\approx R_{id}(1+A\beta)$ for high freq, high gain β^{-1}

- If we take into account the drop in gain A with frequency, $A \approx \frac{\omega_t}{j\omega}$

so $R_{id}(1+A\beta) = R_{id}\left(1+\frac{\beta\omega_t}{j\omega}\right) = R_{id} + \frac{R_{id}\,\omega_{3dB}}{j\omega}$

The second term behaves like a capacitor

$$\frac{1}{j\omega C} \longleftrightarrow \frac{R_{id}\,\omega_{3dB}}{j\omega} \qquad C = \frac{1}{R_{id}\,\omega_{3dB}}$$

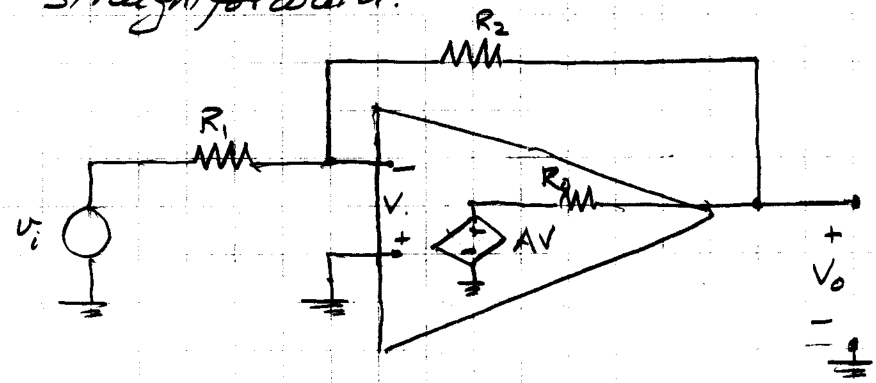


for $\omega \ll \omega_{3dB}$ it's ________

for $\omega \gg \omega_{3dB}$ it's ________

- The point is that
  - we can model the input simply
  - we can (with some effort) find that feedback changes the appearance of the input
  - the input resistance becomes a frequency-dependent impedance as a result of the frequency-dependent open loop gain

- Output Resistance.
  - This also depends on feedback. Text derivation is straightforward.



- We want a Thevenin equivalent. So set sources to zero and either
  - apply a voltage at the output and calculate current
  - apply current source at output, calculate voltage

- $R_{out} = (R_1+R_2) \,\|\, \frac{R_o}{1+A\beta} \approx \frac{R_o}{1+A\beta}$

  In our first cut at op amps, we concluded that the output resistance was zero. It's close:

  $R_o \approx 75\,\Omega \qquad A\beta \gg 1$

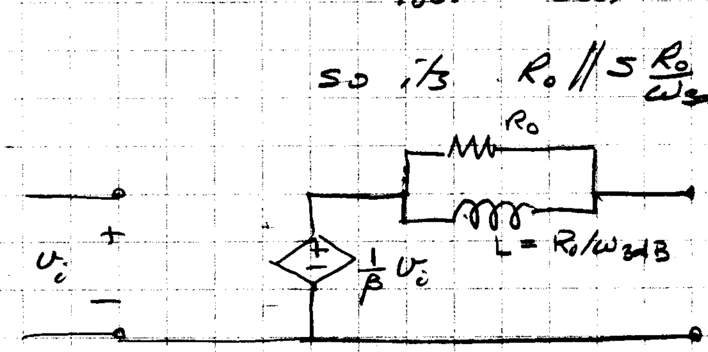
- Again, at high gain, high frequency, $A\beta$ drops. Consequences?

  $A \cong \frac{\omega_t}{j\omega} = \frac{\omega_t}{s}$

  $R_{out} \rightarrow Z_{out} = \frac{R_o}{1+\frac{\beta\omega_t}{s}} = \frac{sR_o}{s+\omega_{3dB}}$

  easier: $Y_{out} = Z_{out}^{-1} = \frac{1}{R_o} + \frac{\omega_{3dB}}{R_o s}$

  so it's $R_o \,\|\, s\frac{R_o}{\omega_{3dB}}$



equivalent model of closed loop

– We have the equivalent of an inductor ?!?
This will be important shortly.

2.7 Applications

- We've looked at imperfections in op amps and consequently
  - developed design guidelines
  - developed some feel for range of validity of the ideal model

- Now briefly revisit ideal model for other applications.

- Electromagnetic sensors

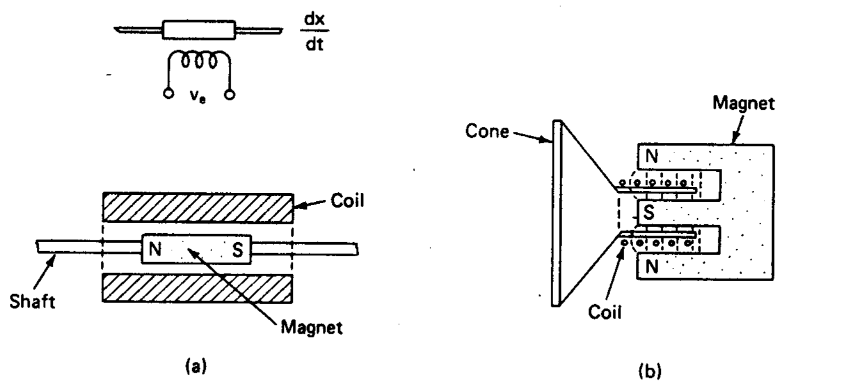


Fig. 8-4 Electromagnetic sensors: (*a*) linear velocity transducer; (*b*) loudspeaker.

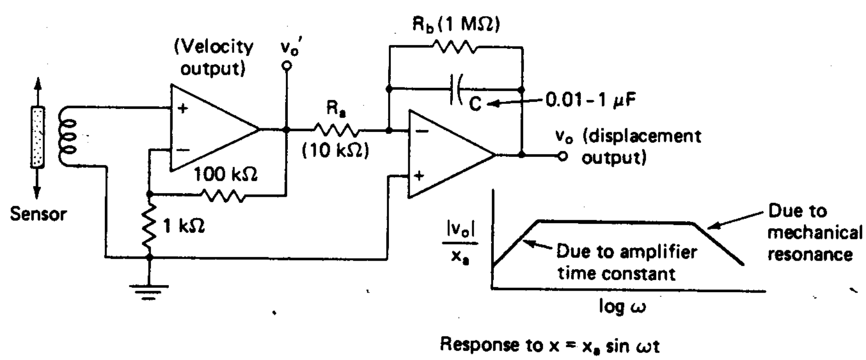


Fig. 8-5 Electromagnetic sensor amplifier with velocity and displacement outputs.

from Wobschall

- Piezoelectric sensors,
  - typically quartz, voltage generated by strain
  - quartz does not conduct electricity so coupled through inherent capacitance.
  - remember op amp terminals need dc path to ground
  - so, . .

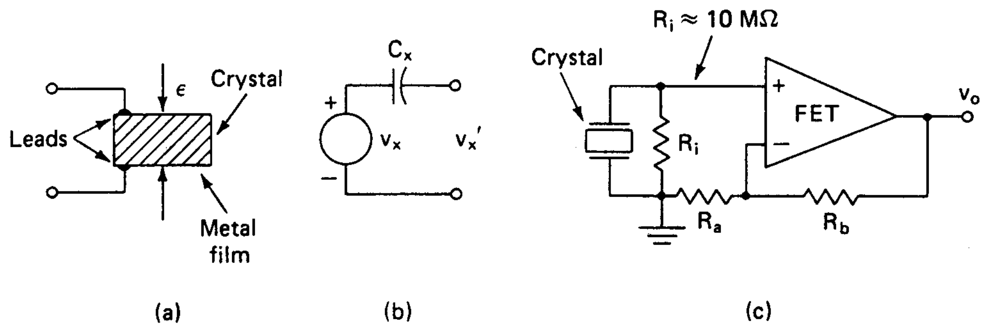
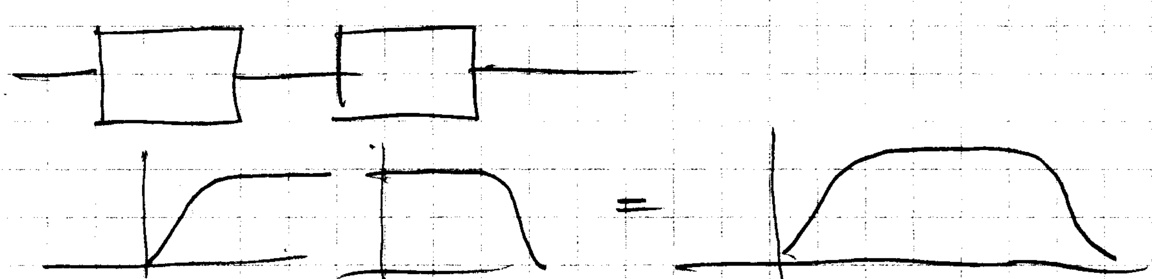


Fig. 8-15 Piezoelectric transducers: (*a*) configuration; (*b*) equivalent circuit; and (*c*) amplifier.

- Active filters
  - we can get STC lowpass, highpass from op amps with capacitors

$$T_{lp}(s) = \frac{a}{s+b} \qquad T_{hp}(s) = \frac{as}{s+b}$$

  - Cascading stages gives any transfer function we like (sort of)



as long as real roots (no resonances)

- For complex roots, we need inductors, too. Or do we? Recall
  - op amp with feedback had inductive R_{out}
  - gyrator makes C look like L
- Active filter handbooks show how to make arbitrary biquadratic stages from op amps with the right connections

$$T(s) = \frac{a s^2 + bs + c}{ds^2 + es + f}$$

more complex filters by cascading

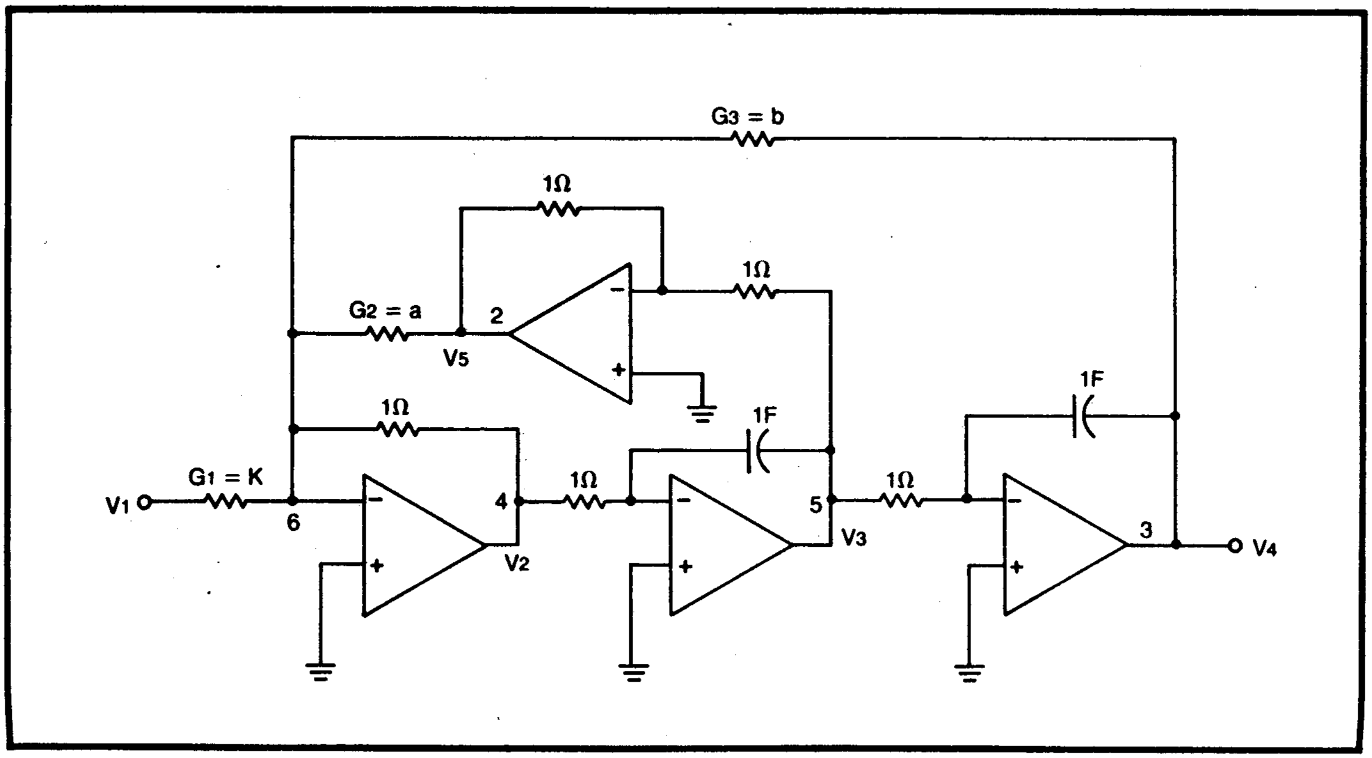


Fig. 4-10. A biquad filter circuit that can obtain a low-pass, high-pass, and band-pass filter characteristic.

from F P Tedeschi "The Active Filter Handbook" TAB Books, 1979

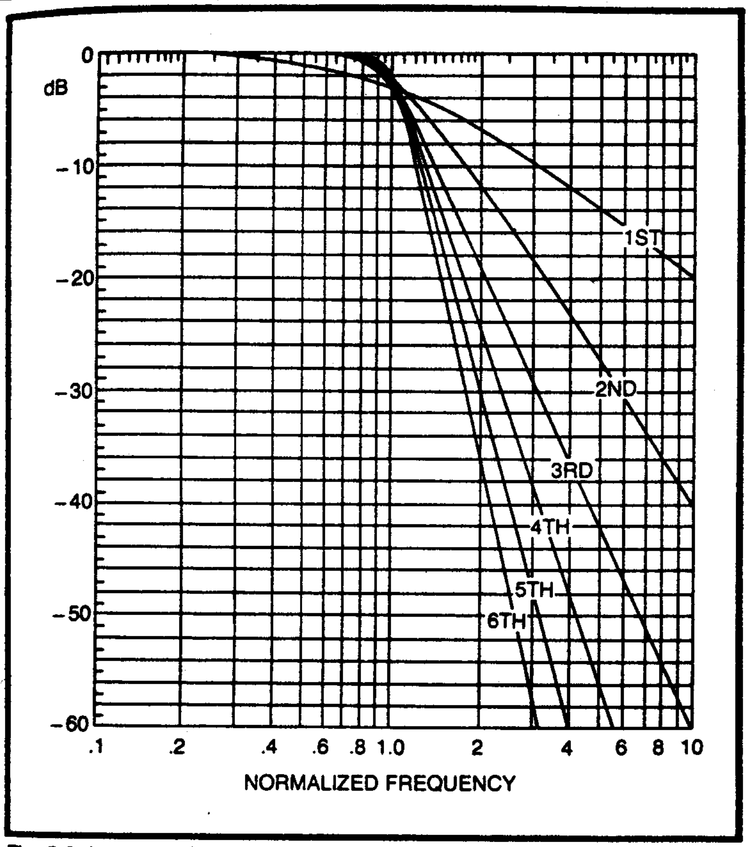


Fig. 5-9. Low-pass frequency response for a Butterworth filter with order n = 1,2,3,4,5, and 6.

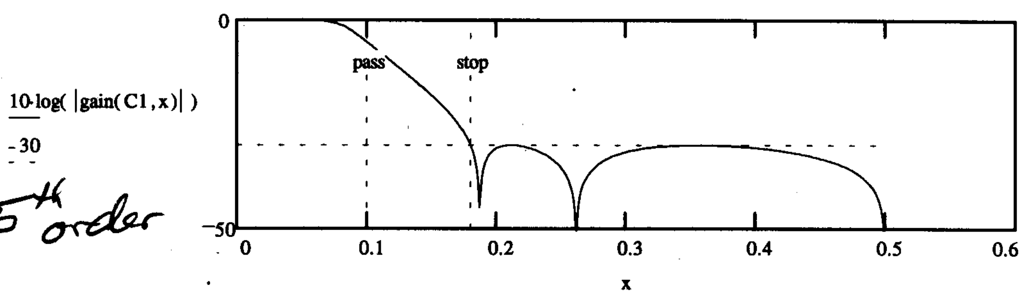
Butterworth filters are very common, very flat in the passband, reasonable drop to stopband.

← from Tedeschi

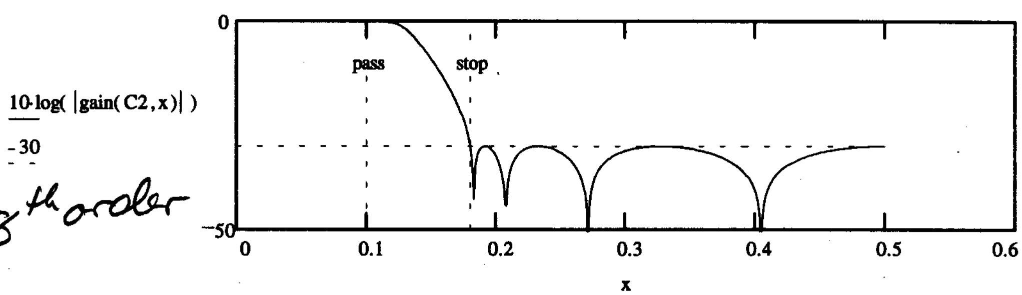
For a larger attenuation, say 1000, a higher order would be needed to get the transition band down to the required width. We'll plot the gains in dB, marking the desired 30 dB attenuation:

C1 := iirlow(cheby2(5, scale, 1000), pass) C2 := iirlow(cheby2(8, scale, 1000), pass)

from Mathcad Signal Processing Function Pack



5th order



8th order

Chebyshev II — very fast transition to stop band