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Classic Uses of Opamps

An Operational Amplifier (Opamp) is a high gain voltage amplifier with differential input. Classic applications are:

(a) 
(b) 
(c)

Figure 6.1  
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Two Stage CMOS Opamps

The classic way of getting high gain is a two stage solution, also providing high output swing (as opposed to e.g. cascode gain stages).

General principle:

The compensation capacitor $C_{cmp}$ in conjunction with the output resistance of the first stage limits the bandwidth, which can be handy to stabilize the circuit when employed in a feedback configuration.
Two Stage CMOS Opamp Example

A simple example:

DC Gain (in a first, mostly valid approximation):

\[ A = A_{v1}A_{v2} \]
First Order Approximation of Frequency Response

In mid range (only $C_C$ matters) simplified to:

\[ A_{v1} = -g_{m1}Z_{out1} \ (6.5) \]
\[ \approx -g_{m1} \left( r_{ds2} \parallel r_{ds4} \parallel \frac{1}{sC_C A_v2} \right) \]

\[ A_v(s) \approx A_{v2} \frac{g_{m1}}{sC_C A_v2} = \frac{g_{m1}}{sC_C} \]

\[ \omega_{ta} \lesssim \frac{g_{m1}}{C_C} \ (6.9) \]
\[ = \frac{l_{bias}}{V_{eff1C_C}} \ (6.10, \text{strong inv.}) \]
\[ = \frac{q I_{bias}}{2nkTC_C} \ (\text{weak inv.}) \]
Second order becomes necessary for analysis close to $\omega_{ta}$. Without $R_C$:

$$A_v(s) = \frac{g_{m1}g_{m7}R_1R_2 \left(1 - \frac{sC_C}{g_{m7}}\right)}{1 + sa + s^2b} \quad (6.15)$$
Interrupt: Second Order Approximation
Deduction 1

\[ v_1 \left( \frac{1}{R_1} + sC_1 + sC_C \right) + v_{\text{in}}g_{m1} = v_{\text{out}}sC_C \quad |1 \]

\[ v_{\text{out}} \left( \frac{1}{R_2} + sC_2 + sC_C \right) + v_1g_{m7} = v_1sC_C \quad |2 \]

\[ v_1 = \frac{v_{\text{out}}}{sC_C - g_{m7}} \left( \frac{1}{R_2} + sC_2 + sC_C \right) \quad | \Rightarrow 2 \rightarrow 3 \]
Interrupt: Second Order Approximation

Deduction 2

\[ v_{out} \frac{1}{R_2} + sC_2 + sC_C \left( \frac{1}{R_1} + sC_1 + sC_C \right) + v_{in}g_m \quad = \quad v_{out} sC_C \quad |3 \rightarrow 1 = 4 \]

\[ v_{out} \left[ \frac{1}{R_1 R_2} + s\left( \frac{C_2 + C_C}{R_1} + \frac{C_1 + C_C}{R_2} \right) + s^2 (C_1 + C_C)(C_2 + C_C) \right] \quad = \quad v_{in}g_m \quad | \equiv 4 = 5 \]

\[ v_{out} \left[ \frac{sC_C(sC_C - g_m)}{sC_C - g_m} - \frac{1}{R_1 R_2} - s\left( \frac{C_2 + C_C}{R_1} + \frac{C_1 + C_C}{R_2} \right) - s^2 (C_1 + C_C)(C_2 + C_C) \right] \quad = \quad v_{in}g_m \quad | \equiv 5 = 6 \]
Interrupt: Second Order Approximation

Deduction 3

\[
\frac{v_{\text{out}}}{R_1 R_2} - s \left( C_C g_m 7 + \frac{C_2 + C_C}{R_1} + \frac{C_1 + C_C}{R_2} \right) + s^2 \left[ \frac{C_C^2}{s C_C - g_m} - (C_1 + C_C)(C_2 + C_C) \right] = v_{\text{in}} g_m \left| \equiv 6 = 7 \right.
\]

\[
\frac{v_{\text{out}}}{v_{\text{in}}} = g_m \left( - \frac{1}{R_1 R_2} - s \left( C_C g_m 7 + \frac{C_2 + C_C}{R_1} + \frac{C_1 + C_C}{R_2} \right) + s^2 \left[ \frac{C_C^2}{s C_C - g_m} - (C_1 + C_C)(C_2 + C_C) \right] \right) \left| \equiv 7 = 8 \right.
\]
Interrupt: Second Order Approximation

Deduction 4

\[
\frac{v_{out}}{v_{in}} = \frac{R_1 R_2 g_m g_{m7} \left( \frac{s C_C}{g_{m7}} - 1 \right)}{-1 - s(C_C g_{m7} R_1 R_2 + R_2 (C_2 + C_C) + R_1 (C_1 + C_C)) + s^2 \left[ R_1 R_2 (C_C^2 - (C_1 + C_C)(C_2 + C_C)) \right]} \quad | \Leftrightarrow 7 = 8
\]

\[
z_1 = -\frac{g_{m7}}{C_C}
\]
Interrupt: Second Order Approximation

Deduction 5

\[ z_1 = -\frac{g_{m7}}{C_C} \]

with \( R_C \)

\[
\frac{sC_C}{g_{m7}} - 1 \rightarrow \frac{sC_C}{(1 + sC_CR_C)g_{m7}} - 1
\]

\[
\frac{sC_C - g_{m7} - sg_{m7}C_CR_C}{(1 + sC_CR_C)g_{m7}}
\]

Nominator:

\[
\frac{s(C_C - g_{m7}C_CR_C) - g_{m7}}{g_{m7}} \rightarrow \frac{s(C_C - R_C) - 1}{g_{m7}}
\]
Second Order Approximation of Frequency Response (2/2)

\[
|\omega_1| \approx \frac{1}{g_m R_1 R_2 C_C} \quad (6.19)
\]

\[
|\omega_2| \approx \frac{g_m}{C_1 + C_2} \quad (6.20)
\]

\[
Z_1 = \frac{g_m}{C_C}
\]

The problem with positive zeros is negative phase shift, here dependent on \(C_C\): Increasing \(C_C\) will reduce \(\omega_{ta}\) but also the frequency at which the phase shift becomes \(-180^\circ\), making a feedback system no more stable.
Compensation Tools

Dominant pole compensation: Moving (only) the dominant pole of the open loop gain to a lower frequency. (Shifting $\omega_t$ to a frequency smaller than the second most dominant pole)

Lead compensation: Introducing a ‘right hand side’ zero that shifts the $-180^\circ$ phase shift to higher frequencies.
Lead compensation (1/2)

With $R_C$ the zero becomes (without much influencing the poles!):

$$z_1 = \frac{-1}{C_C \left( \frac{1}{g_{m7}} - R_C \right)} \quad (6.43)$$

$R_C$ can now be chosen to eliminate the zero:

$$R_C = \frac{1}{g_{m7}} \quad (6.44)$$

or to negate the non-dominant pole $\omega_2$ (using (6.20)):

$$R_C = \frac{1}{g_{m7}} \left( 1 + \frac{C_1 + C_2}{C_C} \right) \quad (6.45)$$
Lead compensation (2/2)

Or to choose $R_C$ even higher to not cancel phase shift due to $\omega_{2/eq}$ to $-180^\circ$ entirely but to ’delay’ it (create a phase lead), e.g. (dependent on a $\beta$ in a closed loop application):

$$R_C \approx \frac{1}{1.7\beta g_{m1}}$$

In all of the above $R_C$ may conveniently be implemented as transistor $Q_9$ in triode region (this $\beta$ is the EKV notation $\beta = \mu C_{ox} \frac{W}{L}$):

$$R_C = \frac{1}{\beta V_{eff9}}$$
Slew Rate Concept

The ‘speed’ of an OpAmp output is not only limited by bandwidth but also by the bias current, as the output current cannot be bigger than the bias current. Thus, a big input step will get the transconductance out of its linear range and the output current saturates. Thus the maximum output gradient of an OpAmp is called slew rate (SR) in units [V/s].
Slew Rate Illustration

the step response that would be expected from a linear system

the actual response to large steps is at first slew-rate limited, with linear settling observed only at the end

a small step response exhibits exponential (linear) settling

\[ V_{step,max} < SR \dot{\tau} \]
Increasing the Slew Rate

The slew rate is dictated by the bias current and the compensation capacitor:

\[ SR = \frac{I_D}{C_C} \]

However, simply increasing the bias current or decreasing \( C_C \) will raise \( \omega_{ta} \), potentially making the circuit unstable. Thus, one needs also to increase \( \omega_2 \) and/or \( V_{eff1} \) (i.e. reduce \((W/L)_1\)) to maintain proper compensation, which the book says are the only ways to design for higher slew rate.
Systematic Offset

Basically the ‘zero output’ of stage one has to closely match the ‘zero input of stage two. (What happens otherwise?) ‘Zero input’ of stage two means the currents in $Q_6$ and $Q_7$ need to be equal. ‘Zero output’ from stage one means that $Q_4$ is sinking half the bias current (while $Q_5$ is sourcing the whole bias current). Thus, if for instance $Q_5$ and $Q_6$ have the same W/L, then $Q_7$ needs to have twice the W/L of $Q_4$. More generally:

\[
\frac{W/L_7}{W/L_4} = 2 \frac{W/L_6}{W/L_5} \quad (6.38)
\]
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The cascode current mirror in chapter 3 reduces the output headroom by $V_{out} > 2V_{eff} + V_{t0}$ (3.42). The problem is that the sources of the transistor closest to the output is at $V_{eff} + V_{t0}$. There are alternatives that provide equally high output resistance with less loss of headroom/output-swing.
Wide-Swing Current Mirrors

Think this circuit through for the case where $I_{bias} = I_{in}$. Then the current through all transistors is the same. For constant current in strong inversion (!) if you scale $W/L$ by $1/a^2$, $V_{eff}$ scales with $a$.

$$V_{s1} = V_{g5} - V_{gs1} = (V_{eff}(n + 1) + V_{t0}) - (V_{eff}n + V_{t0}) = V_{eff}$$

and thus $V_{out} > (n + 1)V_{eff}$ (6.78) for all transistors to be saturated. For instance for $n=1$ the optimum $V_{out} > 2V_{eff}$ (6.79) is obtained. For $I_{in} < I_{bias}$, the minimum $V_{out}$ will shrink in absolute terms, but will no longer be optimal in terms of $V_{eff}$. For $I_{in} > I_{bias}$ the output resistance drops dramatically as the transistors enter the triode region.
Enhanced Output Impedance Current Mirrors (1/2)

Similarly to the cascode current mirror \( V_{d2} \) (and thus the current through \( Q_2 \)) is attempted to be kept as constant as possible. While \( V_{g1} \) is constant and only in the cascode current mirror, here it is actively moved to compensate the influence of \( V_{out} \) on \( V_{d2} \).

So while the a circuit with constant \( V_{g1} \) would have \( R_{out} \approx g_{m1}r_{ds1}r_{ds2} \) (like a cascode current mirror), this circuit has:

\[
R_{out} \approx (A + 1)g_{m1}r_{ds1}r_{ds2} \quad (6.82)
\]
Enhanced Output Impedance Current Mirrors (2/2)

Note:

- $V_{bias}$ needs to be big enough to keep $Q_2$ in saturation!
- Stability of feedback loop needs to be verified!
- Parasitic resistance from drain to bulk may become the actual limiting factor!
The same technique can be used to enhance the output resistance, and thus the gain of a cascode gain stage. **Note:** The current source needs a similarly enhanced output resistance!

\[
A_V(s) = -g_{m2} \left( R_{out} \parallel \frac{1}{sC_L} \right) \quad (6.83)
\]

\[
R_{out}(s) = g_{m1}r_{ds1}r_{ds2}(1 + A(s)) \quad (6.84)
\]
The amplifier is a common source gain stage. **Note:** Again the output swing is quite limited by
\[ V_{out} > V_{eff3} + V_{tn} + V_{eff1} \] (one way of looking at this is that the amp’s \( V_{bias} = V_{eff3} + V_{tn} \))

\[ r_{out}(s) \approx g_{m1}r_{ds1}r_{ds2}\left(\frac{g_{m3}r_{ds3}}{2}\right) \] (6.93)
Wide Swing AND enhanced impedance

\[ I_{in} \approx 7I_{bias} \]

\[ I_{bias} \]

\[ I_{bias} \]

\[ I_{bias} \]

\[ I_{bias} \]

\[ I_{out} = I_{in} \]
Space and Power Conserving Variant

Quite equivalent with worse current matching but less power and layout space consumption. More modular with splitting $Q_2$ and presumably better stability.
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Operational Transconductance Amplifiers

These are operational amplifiers with high output impedance, limited in bandwidth by the output load (and not in internal nodes that are low impedance nodes). Thus, mainly suited for capacitive output loads only!
A simple concept boosting the output current resulting in good bandwidth and good slew rate assuming $C_L$ is dominant.

\[
A_V(s) = \frac{Kg_{m1}r_{out}}{1+s r_{out}C_L} \quad (6.119)
\]

\[
\omega_{ta} \approx \frac{Kg_{m1}}{C_L} = \frac{2K I_{D1}}{C_L V_{eff1}} \quad (6.121)
\]

\[
SR = \frac{K I_b}{C_L} \quad (6.124)
\]
Basic Concept

Ignore $Q_{12}$ and $Q_{13}$ for an initial analysis. Think of it as an extension of a differential pair: the cascodes simply increase the output resistance of the differential output current $\Rightarrow$ higher voltage gain given the same transconductance.
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Interrupt: Common Mode Rejection Ratio

On the white board...
Basic TransAmp with Diff Output

Power = \( IV_{DD} \)
Swing = \( V_{pp,\text{max}} \)
Slew Rate = \( 1/C_L \)

Figure 6.24
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Small Signal Considerations

(a) dc gain: $G_{ma}r_{out}$

$\omega_{-3dB} = 1/(r_{out}C_L)$

$\omega_{ta} = G_{ma}/C_L$

(b) dc gain: $\frac{G_{ma}}{2}r_{out} = G_{ma}r_{out}$

$\omega_{-3dB} = 1/(r_{out}C_L)$

$\omega_{ta} = G_{ma}/C_L$
Fully Differential Current Mirror Opamp

![Circuit Diagram]

Figure 6.29
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Dual Single Ended Structure

Actively pulling the output up and down. (Class AB amplifier as opposed to class A). Better symmetrical slew rate. CMFB needed!
Partially Dual Single Ended Structure

Actively pulling the output up and down. Also better (symmetrical) slew rate, but maybe worse bandwidth (due to more capacitance in current mirrors). CMFB needed!
Wide Input Fully Differential Cascode OpAmp

A problem with low supply voltage is the minimum requirement for the common mode voltage. Complementary input pairs help.
Two Stage Differential OpAmp

Another challenge with low supply voltage is the output swing. Common source output stages do comparatively well: just one $V_{eff}$ away from the rails.

![Two Stage Differential OpAmp Diagram](image)
Common Mode Feedback Principle

Carefull: A feedback loop that needs to be stable!
Continuous Common Mode Feedback Variant

Figure 6.34
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Continuous Common Mode Feedback Variant 2

Saturation of the diff-pairs is a problem as the outputs swing much wider as the input ⇒ reduce gain.
Continuous Common Mode Feedback Variant

\[ V_A = V_{CM} - (V_{eff1} + V_{t1}) \]

\[ V_{ref} = -(V_{eff1} + V_{t1}) \]
Switched Cap Common Mode Feedback

Figure 6.38
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DiffPair

At the core of almost all differential CMOS amplifiers is the diff-pair. The diff-pair invariably translates a differential input voltage into a differential output current around a small signal operating point. (Interestingly this is true for any large signal monotonic function of $I_D(V_{GS})$...) From here on it is best to think in currents for a while rather than voltages.
Differential Current Source

With complementary inputs \( (v_{+x} = -v_{-x}) \), \( v_x \) will be clamped to 0V, simplifying things considerably: two small signal current sources with a parallel output-resistance (common source gain stages, in fact). The simplification holds even for the large signal model where the output current is limited by \([0, I_B]\) and the DC \( V_{x\pm} \) is the average of both inputs. The large signal model is a sinking(!) current source for an nFET pair: so practically you can only connect anything at the top terminal.
Differential Transconductance Amplifier

Using a current mirror you can turn one of your large signal sinking current sources into a sourcing current source. Thus, you can connect the two output currents in a single node thereby subtracting them: you get a single ended current output if you connect it to a low (input) impedance node.
Differential Operational Transconductance Amplifier (1/2)

Or you get a single ended voltage output if you connect it to a high (input) impedance node.
Differential Operational Transconductance Amplifier (2/2)

Rearranging the circuit yields a very simple small signal (DC) model. If the output is connected to a significant capacitive load, this model is even good enough for AC.
Folded Casode OpAmp

The only difference from a small signal perspective of the folded cascode opamp is an increased output resistance. Simply see the cascode gain stage in chapter 3 if you want to understand how this is achieved starting from two Differential Current Sources, or more precisely from two differential common source gain stages.
Current Mirror OpAmp

And the current mirror opamp simply increases the transconductance.
Fully Differential OpAmps

Fully Differential Opamps are in a way simpler, as one can go back to only considering the diff-pair small signal model with complementary inputs. The tricky bit is the large signal point of operation, as one needs to provide exactly matched current sources of $\frac{I_b}{2}$ for each branch of the diff pair to ensure zero output for zero input. Thus, the common mode feedback circuits.
Cascode Principle

This whole section deals (again) with the marvels of a cascode transistor that hugely enhances a common source stage output resistance, bringing it closer to a ideal current source. It basically adds a series resistance of $g_m r_{ds}$:
Advanced Current Mirrors

The rest of this section in the book introduces various ways to a) deduce an optimal $V_{bias}$ to maximize the output swing and b) to make $V_{bias}$ dynamic to increase the output resistance even more. The basic principle of b) is illustrated in:

![Diagram of Current Mirror]

Cadence demonstrations live ...