- OPERATIONAL AMPLIFIERS

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- OPERATIONAL AMPLIFIERS (Introduction and Properties)
**Phase relationships:**

- Non-inverting input to output is 0°
- Inverting input to output is 180°

**OPERATIONAL AMPLIFIER**

- Differential-input, Single-Ended (or Differential) output,
- DC-coupled,
- High-Gain amplifier

<table>
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<th>IDEAL OP-AMP</th>
<th>REAL OP-AMP</th>
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<td>$A_{VO}$, Gain (Differential)</td>
<td>$\infty$</td>
<td>$10^4 - 10^6$ Bipolar input</td>
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<td>$R_{id}$</td>
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<td>$10^5 - 10^6$ MOS or JFET input</td>
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<td>$R_{out}$</td>
<td>0</td>
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<td>C.M.R.R.</td>
<td>$\infty$</td>
<td>90 - 120 dB</td>
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<td>GBW (Gain – Bandwidth Product)</td>
<td>$\infty$</td>
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<td>Offsets</td>
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**IBQ1 and IBQ2**

\[ I_{\text{input} - \text{bias}} \approx |I_{\text{BQ1}}| = |I_{\text{BQ2}}| \]

**Input Offset Current**

\[ I_{\text{OS}} = (I_{\text{BQ1}} - I_{\text{BQ2}}) \]

**Real Operational Amplifiers**

\[ V_{\text{OS}} = V_{\text{OS}}(T), \quad I_{\text{OS}} = I_{\text{OS}}(T), \quad A_{\text{VO}} = A_{\text{VO}}(T) \]

**Temperature Coefficient of Input Offset Voltage**

\[ \frac{dV_{\text{OS}}}{dT} \approx \frac{V_{\text{OS}}}{T} \]

where \( T \) is absolute temperature,

i.e. \(^\circ\text{K}\)
\[ V_{OS} \leq 5 \text{ mV (Bipolar)} \]
\[ (I_{OS}, I_{BIAS}) \leq 1 \mu A \text{ (Bipolar)} \]
\[ \sim \text{ pA (JFET)} \]
\[ \sim \text{ pA (MOSFET)} \]

**OPERATIONAL AMPLIFIER CIRCUITS (APPLICATIONS)**

In the following analyses, unless mentioned otherwise, the operational amplifiers are treated as if they are "ideal" or near ideal.

**I / V (CURRENT / VOLTAGE) CONVERTER**
\[
\left( \frac{I_{\text{Bias}+}}{I_{\text{Bias}-}} \right) \equiv 0 \quad V_{\text{OS}} \equiv 0
\]

Ideal Operational Amplifier

\[
v_{\text{OUT}} = A_{\text{VO}} (v_+ - v_-)
\]

If \( A_{\text{VO}} \rightarrow \infty \) \( (v_+ - v_-) = \frac{v_{\text{OUT}}}{A_{\text{VO}}} \rightarrow 0 \quad v_+ \equiv 0 \Rightarrow v_- = 0 \) (Virtual Ground)

For \( v_- = 0 \) \( (v_- - (v_{\text{OUT}})) = R_i R \Rightarrow R_i \rightarrow \quad v_{\text{OUT}} = -R_i \)

Therefore, this current-voltage converter circuit behaves just like a "Current-Controlled-Voltage-Source" with zero input resistance. (see Figure below)

- **INVERTING AMPLIFIER**
Since $v_- = 0$ (virtual ground) 

\[ i_{IN} = i_1 = \frac{v_1 - (v_-)}{R_1} = \frac{v_1}{R_1} \]

Therefore

\[ i_{IN} = \frac{v_1}{R_1} \]

flows into the I–V converter which creates

\[ v_{OUT} = -R_F i_{IN} \]

resulting in

\[ v_{OUT} = \frac{-R_F}{R_1} v_1 \]

or,

\[ A_v = \frac{R_F}{R_1} \quad R_{OUT} = 0 \]

\[ R_{IN} = \frac{v_1}{i_1} = \frac{v_1}{i_{IN}} = R_1 \]

for the "inverting" amplifier.

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**THE SUMMING AMPLIFIER**

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Diagram of a summing amplifier circuit.
Since $v_\text{in} = 0$ (virtual ground)  
\[ i_1 = \frac{v_1 - 0}{R_1} = \frac{v_1}{R_1} \]
\[ i_2 = \frac{v_2 - 0}{R_2} = \frac{v_2}{R_2} \]
\[ \ldots \]
\[ i_N = \frac{v_N - 0}{R_N} = \frac{v_N}{R_N} \]

\[ i_{\text{IN}} = \sum_{i=1}^{N} i_i \]

and  
\[ v_{\text{OUT}} = -RF \cdot i_{\text{IN}} \]

\[ v_{\text{OUT}} = -\frac{RF}{R_1} \cdot v_1 - \frac{RF}{R_2} \cdot v_2 - \ldots - \frac{RF}{R_N} \cdot v_N \]

where  
\[ a_i = -\frac{RF}{R_i} \]

Comment:
(1) This is equivalent to, mathematically, a weighted sum of inputs or simply an addition operation except for the (-) signs.
(2) To change the sign of the coefficient of an element of the sum simply employ a unity gain inverting amplifier at its input.
THE NON-INVERTING AMPLIFIER

From the resulting equation observe that:

1. The voltage gain is \( + \left( \frac{R_F}{R_1} + 1 \right) \) positive.
2. If \( \frac{R_F}{R_1} \to 0 \) \( (\text{R}_F \to 0 \text{ (short), R}_1 \to \infty \text{ (open)}) \)

then the gain becomes \((+1)\), i.e. **unity gain** (absolute unity). \( \text{v}_{\text{OUT}} = \text{v}_1 \)

UnityGainBuffer Amplifier

3. Since operational amplifier does not draw any current \( \text{i}_+ \approx 0 \) \( \Rightarrow \text{R}_{\text{IN}} \approx \infty \)
Effect of Finite $A_{VO}$ on Feedback Amplifier Gain

BETA $\equiv \beta$ is a transfer relationship. In general $\beta = \beta(j\omega)$

For finite $A_{VO}$, $v_{OUT} = A_{vo}(v_+ - v_-)$, $v_+ = v_1$, $v_- = \beta.v_{OUT}$

$v_{OUT} = A_{vo}(v_1 - \beta.v_{OUT}) \Rightarrow (1 + \beta A_{vo})v_{OUT} = A_{vo}v_1 \Rightarrow$

\[
\frac{v_{OUT}}{v_1} = \frac{A_{vo}}{1 + \beta A_{vo}} \Rightarrow v_{OUT} = \frac{1}{1 + \frac{1}{\beta A_{vo}}} v_1
\]

Therefore, as long as $\beta A_{vo} >> 1$

\[
v_{OUT} \approx \frac{1}{\beta} v_1
\]

Example:
Special Case: \[ \beta = \frac{V_{out}}{V_{in}} = \frac{R_1}{R_F + R_1} \]

\[ V_{OUT} = \frac{1}{R_2 + R_F} \cdot V_1 = \left(1 + \frac{R_F}{R_1}\right) V_1 \]

Non-inverting amplifier

- **FREQUENCY RESPONSE OF REAL OPERATIONAL AMPLIFIERS**

    Internally Frequency Compensated Op.Amp. (One pole / capacitor dominates)

\[ |A_vo(j\omega)|_{dB} = \begin{cases} 100 & \text{if single pole, -20dB/dec} \\ \frac{A_vo(0)}{1 + j \frac{\omega}{\omega_0}} & \text{GBW} \end{cases} \]

\[ A_vo(j\omega) \approx \frac{A_vo(0)}{1 + j \frac{\omega}{\omega_0}} \]

\[ A_vo(j\omega) \gg \omega_0 \approx \frac{A_vo(0)}{j \frac{\omega}{\omega_0}} \]

\[ \omega = \omega_1 = \text{GBW} \]

Notes/Definitions:

Unity Gain Frequency = Gain Bandwidth Product (if single pole, -20dB/dec)

Unity Gain Frequency ≠ GBW (if not -20dB/dec)

Definitions:

- \( A_vo \): The Open Loop Gain (unloaded, no feedback)
- \( \beta \): The Feedback Ratio
- \( \beta A_vo \): The Closed Loop Gain
Non-Inverting Amplifier's Gain BandWidth Product

In the feedback amplifier utilizing this "Real Operational Amplifier"

\[ V_{out}(j\omega) = \frac{1}{\beta(j\omega) \left( 1 + \frac{1}{\beta(j\omega) \cdot A_v(j\omega)} \right)} \cdot V_1(j\omega) \]

For the non-inverting amplifier

\[ \beta(j\omega) = \frac{R_1}{R_1 + R_F} = \text{constant} \]

\[ \frac{V_{out}}{V_1} = \frac{1}{\beta} \cdot \frac{1}{1 + \frac{1}{\beta \cdot A_v(\omega)}} = \frac{1}{\beta} \cdot \frac{1}{1 + \frac{1+j\omega}{\beta \cdot A_v(0)}} = \frac{1}{\beta} \cdot \frac{1}{1 + \frac{1}{\beta \cdot A_v(0)} + \frac{j\omega}{\beta \cdot A_v(0)}} \]

For \( GBW = \omega_0 \cdot A_v(0) \) and, as long as \( \beta \cdot A_v(0) >> 1 \)

\[ \frac{V_{out}}{V_1} = \frac{1}{\beta} \cdot \frac{1}{1 + \left( \frac{j\omega}{\omega_{-3\text{dB}}} \right)} = \frac{1}{\beta} \cdot \frac{1}{1 + \left( \frac{\omega}{\omega_{-3\text{dB}}} \right)} \]

where \( (\omega_{-3\text{dB}}) \) of the feedback amp = \( \frac{\text{GBW}}{\left( \frac{1}{\beta} \right)} \)

and \( \left( \frac{1}{\beta} \right) \) is low frequency gain of the feedback amplifier

\( (\text{Gain} \cdot \text{Bandwidth})_{\text{feedback amplifier}} = (\text{GBW})_{\text{operational amplifier}} \)
FREQUENCY RESPONSE OF AN OPAMP AMPLIFIER WITH FREQUENCY DEPENDENT FEEDBACK

Definitions:

$A_{vo}$ The Open Loop Gain (of the opamp, unloaded, no feedback)

BETA ≡ $\beta$ The Feedback Ratio

$\beta A_{vo}$ The Closed Loop Gain

$A_f$ The Feedback Gain (of the overall amplifier)

What if $A_{vo} = A_{vo}(j\omega)$ and $\beta(j\omega)$ the frequency response of the operational amplifier is more complicated?

Answer: As long as $\beta A_{vo} \gg 1$ the response will be $1/\beta(j\omega)$ as shown with the example below.
THE DIFFERENCE AMPLIFIER
11. Operational Amplifiers

\[ v_1' = v + \]

\[ v_2' \]

\[ i_{B+} \]

\[ i_{B-} \]

\[ R_A \]

\[ R_B \]

\[ R_1 \]

\[ R_F \]

\[ v_{OUT} \]
Use the idea of superposition for 1 & 2:

1. \( v_1 \equiv 0 \), \( v_2 \neq 0 \)

\[
\text{v}_{\text{OUT}} = -\frac{R_F}{R_1} \cdot v_2
\]

2. \( v_1 \neq 0 \), \( v_2 \equiv 0 \)

\[
v_1' = \frac{R_B}{R_A + R_B} \cdot v_1
\]

\[
v_{\text{OUT}} = \left( \frac{R_F}{R_1} + 1 \right) v_1' = \left( \frac{R_F}{R_1} + 1 \right) \frac{R_B}{R_A + R_B} \cdot v_1
\]

\[
v_{\text{OUT}} = -\frac{R_F}{R_1} v_2 + \left( \frac{R_F}{R_1} + 1 \right) \frac{R_B / R_A}{(R_B / R_A) + 1} \cdot v_1
\]

If \( \frac{R_B}{R_A} = \frac{R_F}{R_1} \)

\[
v_{\text{OUT}} = \frac{R_F}{R_1} (v_1 - v_2)
\]

Pure Differential Response

---

A Different Approach:

\( (A_v)_{\text{Op.Amp}} \rightarrow \infty \Rightarrow v_+ \approx v_- \)

\[
v_+ = \frac{R_B}{R_A + R_B} v_1 \quad \quad v_- = \frac{R_F}{R_1 + R_F} v_2 + \frac{R_1}{R_1 + R_F} v_{\text{OUT}}
\]

\[
\frac{R_B}{R_A + R_B} v_1 \approx \frac{R_F}{R_1 + R_F} v_2 + \frac{R_1}{R_1 + R_F} v_{\text{OUT}}
\]

\[
v_{\text{OUT}} = \left( \frac{R_1 + R_F}{R_1} \right) \left( \frac{R_B}{R_A + R_B} v_1 - \frac{R_F}{R_1 + R_F} v_2 \right)
\]

If \( R_B = R_F \) and \( R_A = R_1 \)

\[
\frac{R_B}{R_A} = \frac{R_F}{R_1}
\]

Then

\[
v_{\text{OUT}} = \frac{R_1 + R_F}{R_1} \left( \frac{R_F}{R_1 + R_F} v_1 - \frac{R_F}{R_1 + R_F} v_2 \right)
\]

Choose \( R_B = R_F \) and \( R_A = R_1 \) to make the resistances \( (R_A // R_B) \) and \( (R_1 // R_F) \) equal so that equal \( i_{B+} \) and \( i_{B-} \) bias currents (i.e. zero \( I_{O S} \)) create balanced shifts in \( v_+ \) and \( v_- \). Otherwise, the difference will get amplified and result in a large offset at the output.
**LOGARITHMIC AMPLIFIERS**

If \( v_- \) is virtual ground then

\[
i_1 \approx \frac{v_1 - v_-}{R_1} \Rightarrow i_1 \approx \frac{v_1}{R_1}
\]

\[
i_1 = i_C = I_S e^{\frac{v_{BE}}{kT}} \quad \text{for} \quad v_{BE} = -v_{OUT}
\]

Then for \( v_1 > 0 \) output becomes

\[
v_{OUT} = -\frac{(kT)}{q} \ln\left(\frac{v_1}{R_I I_S}\right) < 0
\]
For $v_1 < 0$, $v_{BE} = v_{OUT}$, $i_C = I_S e^{v_{BE}/kT} = I_S e^{v_{OUT}/kT}$. Then:

$$v_{OUT} = + \frac{(kT)}{q} \ln \left( \frac{-v_1}{R_F I_S} \right) > 0$$

- **ANTI-LOGARITHMIC AMPLIFIERS**

\[ i_C = I_S e^{v_{BE}/kT}, \quad v_{OUT} = -R_F i_C = -R_F I_S e^{v_{BE}/kT} \]

For $v_{EB} = v_1$:

$$v_{OUT} = -R_F I_S e^{v_1/kT}$$

Works for $v_1 > 0$.
Amp - Log Amplifier for $v_1 < 0$

$V_{OUT} = +RF.I_S e^{-\frac{v_1}{kT/q}}$ (works if $v_1 < 0$)

**Applications of Log / Anti-Log Amplifiers:**

\[
\begin{align*}
\log(A \cdot B) &= \log A + \log B \\
\log(A / B) &= \log A - \log B \\
\log^a A + b \log^b B &= \log^a A \cdot \log^b B = A_a B_b
\end{align*}
\]

Therefore, all of the following mathematical functions, multiplication, division, logarithms, exponentiation, power including roots can be implemented using the circuit shown above.
DIFFERENTIATING AMPLIFIER (The OpAmp Differentiator)

Since \( v_- \rightarrow 0 \) (virtual ground) then,
\[
 v_{\text{OUT}} = -R_1 i_R + v_- \quad \rightarrow \quad v_{\text{OUT}} = -R_1 i_R
\]
\[
 i_C = C_1 \frac{dv}{dt} = C_1 \frac{d(v_1 - v_-)}{dt} \quad \rightarrow \quad i_C = C_1 \frac{dv_1}{dt}
\]
\[
 v_{\text{OUT}} = -R C_1 \frac{dv_1}{dt}
\]

To change the sign use unity gain inverter at the input (or the output).

Frequency Response:
\[ v_{\text{OUT}}(+) = -RC \frac{d}{dt} v_1(t) \text{ in } (s \text{ or } jw \text{ domain}) \rightarrow V_{\text{OUT}}(s) = -RC \frac{d}{dt} V_1 e^{st} \]

\[ V_{\text{OUT}}(s) = -sRC \cdot V_1 e^{st} \quad V_1 e^{st} = V_1(s) \rightarrow V_{\text{OUT}}(s) = -sRC \cdot V_1(s) \]

\[ V_{\text{OUT}}(jw) = -jwRC \cdot V_1(jw) \Rightarrow H(jw) = -j \left( \frac{w}{1/RC} \right) \]

### INTEGRATING AMPLIFIER (The OpAmp Integrator)

For \( v_- = \) virtual ground and \( i_R = i_C \)

\[ i_R = \frac{v_1 - v_-}{R} = \frac{v_1}{R} \quad \text{and} \quad i_C = C \frac{d}{dt} (v_- - v_{\text{OUT}}) = -C \frac{dv_{\text{OUT}}}{dt} \]

\[ \frac{v_1}{R} = -C \frac{dv_{\text{OUT}}}{dt} \Rightarrow v_{\text{OUT}} = -\frac{1}{RC} \int_0^t v_-(t') dt' + v_{\text{OUT}}(0) \]

By introducing a switch the initial value, \( v_{\text{OUT}}(0) \) set at zero.
The switch, S keeps shorting the terminals of the capacitor for $t < 0$.
When it opens at $t = 0$ the capacitor's initial voltage is set to be zero.

Frequency Response:

$$V_{OUT}(s) = -\frac{1}{sRC}.V_1(s)$$
$$V(j\omega) = -\frac{1}{j\omega RC} V_1(j\omega)$$
POWER OPERATIONAL AMPLIFIERS

Complementary pair $Q_1$ and $Q_2$ have similar $V_{BE}$ - I_C chs. and identical $\beta$'s but one is NPN, the other is PNP.

$I_{OUT} = (\beta + 1).I_{OA}$
$(I_{OUT})_{MAX} \leq (\beta + 1).(I_{OA})_{MAX}$
$(I_{OA})_{MAX} \sim 10 \text{ mA }$ then $(I_{OUT})_{MAX} \sim 1 \text{ A}$

For higher currents you need to use a Darlington transistor.

Complementary Emitter Follower Chs.
Effect of Negative Feedback on Cross-Over Distortion

If $A_{VO}$ of Op Amp. is very high $v_+ = v_-$

$v_1 = \frac{R_1}{R_1 + R_F} \cdot v_{OUT}$ \quad $v_{OUT} = \left(1 + \frac{R_F}{R_1}\right) v_1$

Note that the voltage $v_1$ will need to be $\frac{\pm 0.7 \, V}{A_{VO}} \approx \pm 7 \, \mu V$ to bring the output out of the cross-over distortion.

With an output swing of $\pm 10 \, V$, Gain $\approx 100$ input swing is $\frac{\pm 10}{100} \approx \pm 0.1 \, V$.

Therefore the ratio of (peak input / input referred cross-over voltage) $\approx \frac{100 \, 000 \, \mu V}{7 \, \mu V} \approx 10^3$. 
POWER AND OUTPUT RANGE LIMITATIONS

\[ I_{OA} \leq I_{OAMAX} \sim 10 \text{ mA}, \quad V_{OA} \leq V_{OAMAX} \]

1. \[ I_{LMAX} = I_{OUTMAX} \approx (\beta + 1)I_{OAMAX} \]

Typically \[ V_{OAMAX} \approx V_{CC} - 1 \text{ V} \]

2. \[ V_{LMAX} \leq V_{OAMAX} - V_{BE1} \]

3. \[ P_{DQ1} = I_L(V_{CC} - V_L) \]

\[ P_{DQA} = I_L(V_{CC} - V_L) \]

\[ P_{MAXQ1} \geq I_{LMAX}V_{CC} \quad \text{worst case: } V_L \approx 0 \]

For a resistive load

\[ V_L = R_LI_L \]

\[ P_Q = \frac{V_L}{R_L}V_{CC} - \frac{V_L^2}{R_L} \]

For non-resistive loads it will be different.
i. OP AMP. DC VOLTAGE REGULATOR

Disadvantages to overcome:
1. Fixed voltage
2. It cannot handle large $I_L$ because it needs high $P_{DMAX}$ zeners
3. Output voltage varies because of finite $r_Z \sim 1-10 \, \Omega$
4. Output resistance $\approx r_Z$, cannot be smaller.
\[
V_+ \approx V_- \
k_1 V_Z = k_2 V_L \quad \Rightarrow \quad V_L = \frac{k_1}{k_2} V_Z
\]
where \(0 \leq k_1 \leq 1\) and \(k_2 = \frac{R_1}{R_1 + R_F}\)

Power is supplied by \(V_{CC}\). To make \(V_{UR}\) the source that supplies power to \(R_L\) simply eliminate \(V_{CC}\) use \(V_{UR}\) for it.

**Limitations:**

- \(I_{L_{MAX}} < (\beta + 1) I_{O_{MAX}}\)
- For \(V_{CC} = V_{UR_{MAX}}\) \(P_{D_{MAX}} > I_{L_{MAX}}(V_{CC} - V_{L_{MIN}})\)
- \(V_{L_{MAX}} < V_{UR_{MIN}} - V_{BE1} - V_{O_{A DROP}}\)
**OP AMP. VOLTAGE REGULATOR WITH ADJUSTABLE CURRENT LIMIT**

Current limit is turned on when \( I_{RCL} \cdot R_{CL} \approx V_{BE \, Q2} \approx 0.7 \, V \)

\[
I_{RCL} = I_L + I_{FEEDBACK} \quad \Rightarrow \quad I_{L_{MAX}} \cdot R_{CL} \approx 0.7 \, V
\]

Comment: Design of zener diode regulator is much simpler since its load is a constant (does not vary) and is also a high resistance load (does not need high current to operate).

**PRECISION RECTIFIERS**

Input: \( v_{IN} \) (t)

Output: \( v_{OUT} \) (t) = \(
\begin{cases} 
\mp K v_{IN} (t) & \text{for } v_{IN} > 0 \\
0 & \text{for } v_{IN} < 0 
\end{cases}
\)

Variations:

(A) Responds to: \( v_{IN} > 0 \) and \( v_{IN} > 0 \) \( v_{IN} > 0 \)

1. inverts or,
2. non – inverting

(B) Responds to: \( v_{IN} < 0 \) and \( v_{IN} < 0 \)

1. inverts or
2. non – inverting
Circuit for A1

\[ v_{OUT} = \frac{R_2}{R_1} \cdot v_{IN} \]

When \( v_{IN} > 0 \) \( v_{OUT} = (v_-) - i_{IN} \cdot R_2 \), \( i_{IN} = \frac{v_{IN} - (v_-)}{R_1} \) and \( (v_-) \to 0 \)

\[ v_{OUT} = \frac{R_2}{R_1} \cdot v_{IN} \]

When \( v_{IN} > 0 \) \( v_{OUT} = (v_-) - i_{R_2} \cdot R_2 = 0 \) for \( (v_-) \to 0 \) and \( i_{R_2} = 0 \)

The "Precision Half-Wave Rectifier Circuit A1" given above responds to positive \( v_{in} \), amplifies it and inverts.

Circuit for B1

\[ v_{OUT} = \begin{cases} \frac{R_2}{R_1} v_{IN} & \text{if } v_{IN} < 0 \\ 0 & \text{if } v_{IN} > 0 \end{cases} \]

The "Precision Half-Wave Rectifier Circuit B1" given above responds to negative \( v_{in} \), amplifies it and inverts.