Electronic Circuits Summary

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Constants (@300K)

\[ \varepsilon_0 = 8.854 \times 10^{-12} \text{F/m} \]
\[ m_0 = 9.11 \times 10^{-31} \text{kg} \]
\[ k = 1.38 \times 10^{-23} \text{J/K} = 8.617 \times 10^{-5} \text{eV/K} \]
\[ \frac{kT}{q} = 0.0259 \text{V} \quad \frac{q}{kT} = 38.61 \frac{1}{V} \quad kT = 25.9 \text{meV} \]
\[ 1 \text{eV} = 1.602 \times 10^{-19} \text{J} \quad q = 1.602 \times 10^{-19} \text{As} \]

1. Transistor Characteristics

Resistor: \[ V_R = R \times I_R \]
Capacitor: \[ I_C = C \frac{d}{dt} V_C \]
Inductor: \[ V_L = L \frac{d}{dt} I_L \]

Bipolar Junction Transistor (BJT)

\[ I_C = I_S e^\frac{V_{BE}}{V_T} \left( 1 + \frac{V_{CE}}{V_A} \right) , \quad V_T = \frac{kT}{q} \approx 26 \text{mV} \]
\[ I_B = \frac{I_C}{\beta}, \quad I_E = (1 + \beta) I_B, \quad V_A: \text{Early voltage} \]

Small Signal Equivalent Circuit BJT

Consider only small oscillations around operation point
\[ \rightarrow \text{Linearize as approximation, } V_{CC} = 0 = V_{BE} \text{ as const.} \]
\[ i_C = \frac{\beta}{\beta + 1} e^\frac{V_{BE}}{V_T} (v_{in} - v_E) = g_m \Delta V_{BE} \approx \frac{v_E}{R_E} \]
\[ v_{out} \approx -g_m \frac{R_L}{1 + g_m R_E} v_{in} \]

2. Single-Transistor Amplifiers

MOSFET

\[ I_D = \frac{K'}{2} \left( \frac{V_{GS} - V_t}{V_{DS}} \right)^2 (1 + \lambda V_{DS}) , \quad V_{DS} > V_{GS} - V_t \]
\[ K': \text{Intrinsic conductance coeff.} \]
\[ V_t: \text{Threshold voltage} \]
\[ W/L: \text{Gate width / Gate length} \]
\[ \lambda: \text{Characteristic length} \]

Millers theorem

\[ Z_{in} = \frac{Z}{1 + |A_V|} , \quad Z_{out} = \frac{Z}{1 + |A_I|} \]
**Common-Emitter / Source Amplifier**

\[ R_{\text{in}} = \frac{v_{\text{in}}}{i_{\text{in}}} = r_{\pi} \]

\[ R_{\text{out}} = \frac{v_{\text{out}}}{i_{L}} \bigg|_{v_{\text{in}}=0} = r_0 \]

\[ i_L = -\frac{r_0}{r_0 + R_L} g_m v_{\text{in}} \]

\[ A_V = \frac{v_{\text{out}}}{v_S} \approx -\frac{r_\pi}{r_\pi + r_0} g_m v_{\text{in}} \]

Inverting Amplifier → 180° phase shift

\[ A_I = \left| \frac{i_{\text{out}}}{i_S} \right|_{v_{\text{out}}=0} = \frac{R_S}{R_S + r_\pi} \beta, \quad i_{\text{in}} \bigg|_{v_{\text{out}}=0} = g_m r_\pi = \beta \]

**MOSFET:** instead of BJT, no current into the gate

R\text{out} \rightarrow \infty, v_{\text{in}} = v_S

\[ R_{\text{out}} = r_0, \quad A_V \approx -g_m R_L \]

**Common-Base / Gate Amplifier**

\[ R_{\text{in}} = \frac{v_{\text{in}}}{i_{\text{in}}} \approx \frac{1}{g_m} \]

\[ R_{\text{out}} = \frac{v_{\text{out}}}{i_{\text{out}}} \bigg|_{R_S \gg r_\pi} \approx (r_\pi | R_S | g_m + 1) r_0 \approx \beta r_0 \]

\[ A_V = \frac{v_{\text{out}}}{v_S} \approx \frac{g_m R_L}{1 + g_m R_S} : \text{non-inverting amp.} \]

\[ A_I = \left| \frac{i_{\text{out}}}{i_S} \right|_{v_{\text{out}}=0} = -\frac{g_m}{1 + R_S | R_\pi | + g_m} \approx -1 \text{ for } R_S \gg r_\pi \]

**MOSFET:** R\text{S} = 0: voltage source; R\text{S} = \infty: current source

\[ A_V \approx \frac{g_m R_L}{1 + g_m R_S} \]

\[ A_I = \left| \frac{i_{\text{out}}}{i_S} \right|_{v_{\text{out}}=0} = -1 \]

\[ R_{\text{in}} \approx \frac{1}{g_m} \]

\[ R_{\text{out}} \approx (1 + g_m R_S) r_0 \]

**Common-Collector / Drain Amplifier**

Also known as emitter / source follower

\[ v_{\text{out}} \approx g_m R_L v_{\text{in}} \]

\[ A_V = \frac{v_{\text{out}}}{v_S} \approx \frac{g_m R_L}{1 + g_m R_S} \]

\[ A_I = \left| \frac{i_{\text{out}}}{i_S} \right|_{v_{\text{out}}=0} = -1 \]

\[ R_{\text{in}} = \frac{1}{g_m} \]

\[ R_{\text{out}} \approx \frac{1}{g_m} \]

**MOSFET:** no current flowing into the gate (A\text{I} = \infty)

\[ A_V \approx \frac{1}{1 + g_m R_L} \]

\[ i_{\text{out}} = g_m v_{\text{out}} \]
Comparison of the three basic amplifiers

<table>
<thead>
<tr>
<th></th>
<th>CE/CS Amplifier</th>
<th>CC/CD Amplifier (Voltage Buffer)</th>
<th>CB/CG Amplifier (Current Buffer)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage Gain</td>
<td>High</td>
<td>Low</td>
<td>High</td>
</tr>
<tr>
<td>Current Gain</td>
<td>High</td>
<td>Moderate</td>
<td>-1</td>
</tr>
<tr>
<td>Input Resistance</td>
<td>High</td>
<td>High</td>
<td>Low</td>
</tr>
<tr>
<td>Output Resistance</td>
<td>High</td>
<td>Low</td>
<td>High</td>
</tr>
</tbody>
</table>

After voltage buffer: lower output resistance (better V-source)
After current buffer: larger output resistance (better C-source)

Impedance Matching

\[ P_L \text{ is maximized when } R_L = R_S \]

3. Frequency Response of Amplifiers
Change of charge vs. voltage across pn-junctions between BJTs can be represented by a parasitic capacitance

\[ C_C : \text{ capacitance between } B \text{ and } E, \quad C_C \approx C_{gs} \]
\[ C_C : \text{ capacitance between } B \text{ and } C, \quad C_C \approx C_{gd} \]

Stage 1:

\[ v_{out} = -g_{m2}v_1Z_2 \approx -\frac{g_{m2}R_2}{1 + sC_2R_2}v_1, \quad Z_2 = \frac{R_2}{1 + sC_2R_2} \]

\[ A_{V2}(s) = \frac{v_{out}(s)}{v_1(s)} = -\frac{g_{m2}R_2}{1 + sC_2R_2} \]

Cut-off frequencies: defined by poles

\[ A_V(s) = \frac{v_{out}(s)}{v_{in}(s)} = \frac{g_{m1}R_1 * g_{m2}R_2}{(1 + s R_1 C_{gs2})(1 + s R_2 C_2)} \]

\[ \omega_1 = |p_1| = \frac{1}{R_1C_2}, \quad \omega_2 = |p_2| = \frac{1}{R_2C_2} \]

Symmetry between left and right branch \( \rightarrow \) split circuit into two independent parts and analyze separately (once)

Differential amplifier: In order to filter out DC component before the amplification, we use a fixed tail current \( I_E \), which also enables DC coupling of stages (current splitted)

4. Differential Amplifiers

Transmitting information with two complementary signals

\[ V_{id} = V_i - \frac{V_{dd}}{2} \text{ Common-mode (DC component)} \]

\[ V_{id} = V_i - V_{cc} \text{ Differential component} \]

\[ V_E (\text{emitter-node potential}) \text{ remains constant } \rightarrow v_E = 0 \]

Differential amplification: \( A_{vd} = \frac{v_{od}}{v_{id}} = -g_m R_C \)

Common-mode amplification: \( A_{vcm} = \frac{v_{od}}{v_{id}} = \frac{v_{acm}}{v_{icm}} = -\frac{R_C}{2R_E} \)
Common Mode Rejection Ratio
Indicates how strong a common mode signal is attenuated compared to a differential signal

\[ G = \frac{A_{vd}}{A_{vcm}} = -\frac{g_m R_C}{R_C/2R_E} = 2g_m R_E \]

\[ CMRR = G_{dB} = 20 \times \log_{10} G \]

\[ GBP = A_0 \times \omega_C \]

**Operational amplifiers**

**Ideal:** \( Z_{in} \rightarrow \infty, Z_{out} \rightarrow 0, A_{vd} \rightarrow \infty, CMRR \rightarrow \infty \)

**Non-ideal Small Signal Equivalent**

Voltage Comparator

\[ V_{out} = \text{sign}(V_{in}) \times V_{CC} \]

Voltage follower (Buffer)

\[ V_{out} = V_{in} \]

Inverting amplifier

\[ V_{out} = -\frac{R_2}{R_1} V_{in} \]

Non-inverting amplifier

\[ V_{out} = \left(1 + \frac{R_2}{R_1}\right) V_{in} \]

Integrator

\[ V_{out}(s) = \frac{V_{in}(s)}{sRC} \]

\[ V_{out}(t) = V_{out}(0) - \frac{1}{RC} \int V_{in}(t) \, dt \]

Differentiator

\[ V_{out}(s) = -V_{in}(s)RCs \]

\[ V_{out}(t) = -RC \frac{d}{dt} V_{in}(t) \]

**5. Instrumentation Amplifier**

Precise amplification of weak, distorted sensor signals

High input impedance, internal feedback loop

**Basic Instrumentation Amplifier**

Amplifies voltage difference with a precise gain

Differential gain must be equal for both input branches

\[ V_0 = \frac{R_2}{R_1} \frac{1 + R_4/R_3}{1 + R_2/R_1} V_{i+} - \frac{R_4}{R_3} V_{i-} \]

Set \( R_1 = R_3, R_2 = R_4 \) to equally load both input branches:

\[ V_0 = G \times V_{i+} - G \times V_{i-}, \quad G = \frac{R_2}{R_3} \]

**Buffered Instrumental Amplifier**

Obtain ideally high input impedance by input buffering

\[ V_0 = V_{iCM} \left( \frac{R_2(R_3 + R_4)}{R_3(R_1 + R_2)} - \frac{R_4}{R_3} \right) + \frac{V_{id}}{2} \left( \frac{R_2(R_3 + R_4)}{R_3(R_1 + R_2)} + \frac{R_4}{R_3} \right) \]

\[ CMRR = \frac{A_d}{A_{vCM}} = \frac{V_0/V_{id}}{V_0/V_{iCM}} = \frac{(R_3 + R_4)R_2 + (R_1 + R_2)R_4}{2 \times (R_2R_3 - R_4R_1)} \]
6. Voltage Regulators, Logarithmic & Anti-Logarithmic Amplifiers

Input stage gain

Differential & common mode gain of input stage:

\[ A_B = \frac{V_{B+} - V_{B-}}{V_{id}} = 1 + \frac{R_5 + R_6}{R_7} \]

\[ A_{cm,B} = \frac{V_{B+} + V_{B-}}{V_{i+} + V_{i-}} = 1 \rightarrow \text{no current through } R_5, R_6, R_7 \]

Differential & common mode gain in total:

\[ A'_d = \frac{V_B}{V_{id}} = \frac{A_B}{2} \left( \frac{R_2(R_3 + R_4)}{R_3(R_1 + R_2)} + \frac{R_4}{R_3} \right) \]

\[ R_1 = R_3, R_2 = R_4 \rightarrow A'_d = \frac{R_2}{R_1} A_B \]

\[ A_{cm} = A_{cm,B} \left( \frac{R_2(R_3 + R_4)}{R_3(R_1 + R_2)} - \frac{R_4}{R_3} \right) \]

CMRR = \frac{A'_d}{A_{cm}} = \frac{A_B}{R_1} \rightarrow \text{increased by factor } A_B

Voltage offset: Offset voltage in combination with a small input signal is highly undesired. The output signal then reaches the saturation level even for small values of \( V_i \) and is therefore distorted.

If this is not appropriate, chopper amplifiers can be used.

Linear voltage regulators

Non-inverting topology

\[ I_L = \frac{V_{out}}{R_L} = \frac{R_{CTRL}}{R_L + R_{CTRL}} V_{sup} \]

Inverting topology

\[ I_L = \frac{V_{out}}{R_L} \approx I_E \]

\[ V_{out} = \left( 1 + \frac{R_{F1}}{R_{F2}} \right) V_{ref} \]

Choose \( R_{F1}, R_{F2} \gg R_L \rightarrow I_L \approx I_E \]

Logarithmic & Anti-Logarithmic Amplifiers

Non-linear circuit whose output voltage is proportional to the logarithm / exponential of the input voltage

Logarithmic Amplifier: Rely on logarithmic relationship of \( I_C \) & \( V_{BE} \)

\[ I_{in} = \frac{V_{in}}{R_1} = I_C e^{-\frac{V_{in}}{V_T}}, V_{out} = -V_T \ln \left( \frac{I_{in}}{I_S} \right) = -V_T \ln \left( \frac{V_{in}}{V_T R_1 I_S} \right) \]

Anti-Logarithmic Amplifier

\[ V_{BE} = -V_{in}, I_C = I_S e^{-\frac{V_{in}}{V_T}}, V_{out} = I_C R_1 = I_S R_1 e^{-\frac{V_{in}}{V_T}} \]
7. Active RC Filters

Filter is a frequency-selective circuit that passes a specified band of frequencies and blocks frequencies outside of it.

Passive Filters: based on passive elements such as R / L / C
Active Filters: based on op-amps in addition to R / L / C

Cutoff frequency
Poles define cut-off: $|T(j\omega_p)| = \frac{1}{\sqrt{2}} \max(|T(j\omega)|)$

$\omega_n = |p_n|$, $BW_{-3dB} = \omega_0/Q_0$

First order passive filters

Low-pass Filter
- Transfer function: $T(s) = \frac{V_o}{V_i} = \frac{1}{1 + sRC}$
- Amplitude response: $|T(j\omega)| = \frac{1}{\sqrt{1 + (\omega RC)^2}}$
- Phase response: $\angle T(j\omega) = -\tan^{-1}(\omega RC)$

High-pass Filter
- Transfer function: $T(s) = \frac{V_o}{V_i} = \frac{sRC}{sRC + 1}$
- Amplitude response: $|T(j\omega)| = \frac{\omega RC}{\sqrt{\omega^2 RC^2 + 1}}$
- Phase response: $\angle T(j\omega) = 90° - \tan^{-1}(\omega RC)$

Comparison of first order filters

<table>
<thead>
<tr>
<th>Passive Filters</th>
<th>Active Filters</th>
</tr>
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<tr>
<td>2 passive elements that determine pole</td>
<td>3 passive elements, pole determined by elements in op-amp feedback</td>
</tr>
<tr>
<td>Fixed gain of 1 in pass band</td>
<td>Variable pass band gain possible</td>
</tr>
<tr>
<td>No power consumption</td>
<td>Op-amp consumes power</td>
</tr>
<tr>
<td>Real filter transfer function close to ideal filter function</td>
<td>Real filter transfer function dependent on op-amp DC-gain and gain-bandwidth product</td>
</tr>
</tbody>
</table>

Second order passive filters

2nd order passive filters can be synthesized from 1st order

Low-pass filter
- $T(s) = \frac{1}{s^2 + \frac{\omega_0^2}{Q_0} s + \frac{1}{Q_0 \omega_0}}$
- $\omega_0 = \frac{1}{\sqrt{LC}}$, $Q_0 = \frac{1}{R \sqrt{LC}}$
- Denominator: $D(s) = s^2 + \frac{\omega_0^2}{Q_0} s + \frac{1}{Q_0 \omega_0}$
- Poles: $p_{1,2} = -\frac{\omega_0}{2Q_0} \pm \frac{\omega_0}{2} \frac{1}{\sqrt{4Q_0 - 1}}$

High-pass filter
- $T(s) = \frac{s^2 + \frac{\omega_0^2}{Q_0} s + \omega_0^2}{s^2 + \frac{\omega_0^2}{Q_0} s + \omega_0^2}$

Band-pass filter
- $T(s) = \frac{s^2 + \frac{\omega_0^2}{Q_0} s + \frac{1}{\omega_0^2}}{s^2 + \frac{R}{L} s + \frac{1}{LC}}$

Resonance frequency:

$\omega_0 = \frac{1}{\sqrt{LC}}$

Quality factor:

$Q_0 = \frac{1}{R \sqrt{LC}}$

Energy vs. freq:

The higher Q, the narrower and sharper the peak.
**Sallen-Key amplifier**
Allow sharp gains without using inductors (expensive)

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**Low-pass filter**

\[
T(s) = \frac{K}{s^2 + \left(\frac{1}{R_1C_1} + \frac{1}{R_2C_2} + \frac{1-K}{R_1C_1} \right)s + \frac{1}{R_1R_2C_1C_2}}
\]

\[
\omega_0 = \frac{1}{\sqrt{R_1R_2C_1C_2}}
\]

**High-pass filter**

\[
T(s) = \frac{Ks}{s^2 + \left(\frac{1}{R_2C_2} + \frac{1}{R_2C_1} + \frac{1-K}{R_1C_1} \right)s + \frac{1}{R_1R_2C_1C_2}}
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**Band-pass filter**

\[
T(s) = \frac{Ks}{s^2 + \left(\frac{1}{R_1C_1} + \frac{1}{R_2C_2} + \frac{1-K}{R_1C_1} \right)s + \frac{1}{R_1R_2C_1C_2}}
\]

\[
\omega_0 = \frac{1}{\sqrt{R_1R_2C_1C_2}}
\]

**Tow-Thomas Biquad filter**
Less sensitive to tolerance difference; combine Sallen-Keys

\[
T(s)_{\text{LP}} = \frac{V_2}{V_1} = \frac{1}{s^2 + \left(\frac{1}{R_1C_1} + \frac{1}{R_2C_2} + \frac{1-K}{R_1C_1} \right)s + \frac{1}{R_1R_2C_1C_2}}
\]

\[
T(s)_{\text{HP}} = \frac{V_2}{V_1} = \frac{1}{s^2 + \left(\frac{1}{R_2C_2} + \frac{1}{R_2C_1} + \frac{1-K}{R_1C_1} \right)s + \frac{1}{R_1R_2C_1C_2}}
\]

\[
\omega_0^2 = \frac{1}{R_2R_1C_1C_2}
\]

\[
Q_o = \omega_0 R_1 C_1
\]

---

**8. Switched capacitor filters**
Motivation: some systems require an active RC low-pass filter with very low \(f_{\text{cut-off}}\) → We need a large resistor in a highly integrated chip, whereby it is also inaccurate.

**Concept of switched capacitor devices**

Transfer of charge \(\Delta Q\) from potential \(V_1\) to potential \(V_2\) at a fixed rate \(f_c = \frac{1}{T_C}\)

Transferred charge per \(T_C\): \(\Delta Q = C \cdot (V_1 - V_2)\)

Average current: \(I_{2,\text{avg}} = \frac{\Delta Q}{T_C} = \frac{C \cdot (V_1 - V_2)}{T_C}\)

Equivalent resistor: \(R_{\text{eq}} = \frac{T_C}{C} = \frac{1}{f_c \cdot C}\)
Inverting Integrator using SC

**Phase 1:** \( \Phi_1 \) on, charge accumulates on \( C_1 \) and \( C_2 \)

\[
C_2 V_{out}(nT_C) = C_2 V_{out}(n-1)T_C - C_1 V_{in}(nT_C)
\]

\[ V_{out}(nT_C) = \frac{-C_2}{TC_2} \int_0^{nT_C} V_{in}(t) \, dt \]

\( \Rightarrow \) seems continuous for small enough \( T_C \)

Non-inverting Integrator using SC

**Phase 1:** \( \Phi_2 \) on, \( C_1 \) is discharged

\[
C_2 V_{out}(nT_C) = C_2 V_{out}(n-1)T_C + C_1 V_{in}(nT_C)
\]

**Phase 2:** Charge is transferred to \( C_2 \)

\[
C_2 V_{out}(nT_C) = C_2 V_{out}(n-1)T_C + C_1 V_{in}(nT_C)\]

\[
\frac{V_{out}(z)}{V_{in}(z)} = \frac{C_1}{C_2} \frac{1}{1 - z^{-1}}
\]

Switched capacitor Tow-Thomas biquad

\[ T(z) = \frac{V_{out}(z)}{V_{in}(z)} = -\frac{C_{R4} C_{R2} C_{R3}}{C_{R1}} \frac{C_{R2} C_{R1} C_{R3}}{C_{R2} C_{R1} + C_{R3}} \]

Design equations:

\[
\frac{C_{R4}}{C_{R3}} = -k, \quad \frac{C_{R2}}{C_{R1}} = \omega_0, \quad \frac{C_{R3}}{C_{R1}} = Q
\]

Z-transform extended

Time discrete equivalent of the Laplace function.

\[
X(z) = \sum_{k=-\infty}^{\infty} x[k]z^{-k}
\]

Delay by \( n \) samples:

\[ x[k] \rightarrow x[k-n] \Leftrightarrow X[z] \rightarrow z^{-n}X[z] \]

Transformation to frequency:

\[ z = e^{i\omega T_c} = e^{i2\pi f c} \]

Differentiation:

\[ Z \left( \frac{df(t)}{dt} \right) = F(z) \frac{1 - z^{-1}}{T_c} \]

Integration:

\[ Z \left( \int f(t) \, dt \right) = F(z) \frac{T_c}{1 - z^{-1}} \]

Forward Euler Transform: \( s = \frac{z^{-1}}{T_c} \)

Backward Euler Transform: \( s = \frac{1 - z^{-1}}{T_c} \)
9. Appendix

Transimpedance amplifiers
Sensing an input current & converting it to output voltage

\[ r_m = \frac{dV_0}{dI_{in}}, \quad Z_{in} \to 0, \quad Z_{out} \to 0 \]

Frequency Response of Transimpedance Amplifiers

Transimpedance experiences broadboding due to feedback:

\[ Z_T(s) = \frac{V_{out}}{I_f} = -A_v(s)Z_1 = \frac{R_fA_0}{1 + A_0} \frac{Z_s}{R_f + Z_s} \]

Bandwidth is limited by GBP of op-amp.

Loop gain:

\[ A_L(s) = A_v(s) \frac{Z_s}{R_f + Z_s} \]

Feedback-factor:

\[ \beta(s) = \frac{Z_s}{R_f + Z_s} \]

High bandwith trade-off: High transimpedance gain results in lower bandwidth

Due to the capacitance \( C'_i = C_i + C_{in} \) the transimpedance amplifier becomes a second-order system with DC transimpedance gain \( \approx -R_f \) and a loop gain

\[ A_L(s) = \frac{A_0}{1 + \frac{s}{\omega_p}} \]
Switched Capacitors Examples

Example:

At the end of $\Phi_1(n-1)$:

$Q_{C1} = C_1 \cdot V_i(n-1)$
$Q_{C2} = C_2 \cdot V_0(n-1)$

At the end of $\Phi_2(n-0.5)$:

$Q_{C1} = C_1 \cdot 0$
$Q_{C2} = C_2 \cdot V_0(n-0.5)$

Charge conservation:

At $(n-0.5)$:

$C_2 V_0(n-0.5) = C_2 V_0(n-1) + C_1 V_i(n-1)$

At $(n)$: (nothing happens to $Q_{C2}$)

$C_2 V_0(n) = C_2 V_0(n-0.5)$

Results in:

$C_2 V_0(n) = C_2 V_0(n-1) + C_1 V_i(n-1)$

$V_0(Z) = C_2 \cdot Z^{-1} \cdot V_i(Z)$

$V_i(Z) = C_1 \cdot [1 - Z^{-1}]$

Example 2:

$\Phi_1(n-1)$:

$Q_1 = C_1 V_i(n-1)$
$Q_2 = C_2 V_0(n-1)$

$\Phi_1(n-0.5)$:

$Q_1 = 0$
$Q_2 = C_2 V_0(n-0.5)$

Charge conservation:

Überlegen wo positive und negative Ladung hingehn.
Verstärker verstärkt solange bis die
eingangsspannungs differenz 0 ist.

@ $(n-0.5)$:

$C_2 V_0(n-0.5) = C_2 V_0(n-1) - C_1 V_i(n-1)$

@ $(n)$:

$C_2 V_0(n) = C_2 V_0(n-0.5) - C_1 V_i(n)$

$V_0(Z) = C_2 \cdot [1 + Z^{-1}]$

$V_i(Z) = -\frac{C_1 \cdot 1 + Z^{-1}}{C_2 \cdot 1 - Z^{-1}}$