Operational Amplifiers

Electronics:
Experimental techniques in the photon sciences

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October 2012

Outline

Basics
- Complex impedance & the low pass-filter
- The decibel scale and Bode plots

Introduction

General characteristics
- Basic operation
- The ideal op-amp

The Concept of Feedback
- Basic idea of feedback
- The noninverting and inverting amplifier

Circuit Examples
- Summing and differential amplifier
- Integrator and Differentiator
- Nonlinear applications

Real op-amps
- Frequency Response and Slew Rate
- Input Offset Voltage and Bias Current

Grounding problems

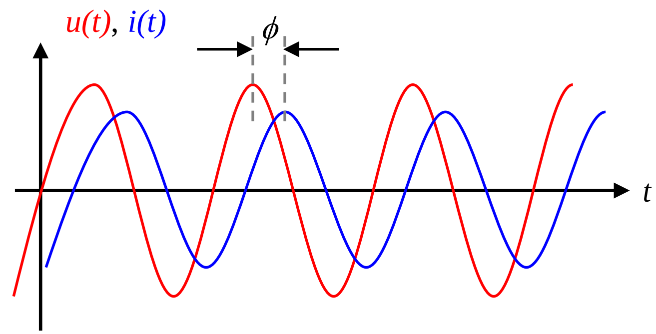
Complex impedance

Sinusoidal signals

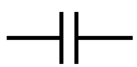
voltage $u(t) = \hat{u} \cdot \exp(j\omega t)$

current $i(t) = \hat{i} \cdot \exp(j\omega t + \phi)$

impedance $Z = \dfrac{u(t)}{i(t)} = |Z| \cdot \exp(i\phi)$

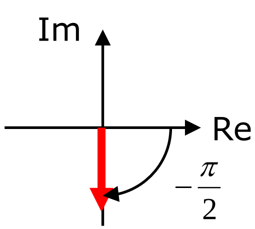


- **Capacities** have to be charged first to have a certain voltage

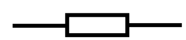


$$i = C\frac{du}{dt}$$

$$Z(\omega) = \frac{1}{j\omega C}$$

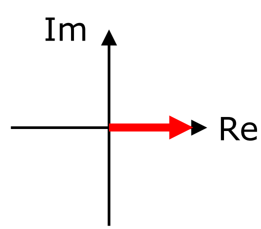


- Voltage drop over ohmic **resistor** is proportional to current



$$u = R \cdot i$$

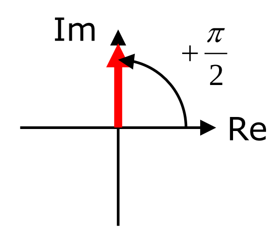
$$Z(\omega) = R$$



- Current change through **coil** induces voltage drop

$$u = L\frac{di}{dt}$$

$$Z(\omega) = j\omega L$$



The decibel scale

- decibels (dB) measure the power (intensity) ratio on a logarithmic scale
  → transfer functions of consecutive elements can be "added", e.g. gain of two amplifiers

$$G = 10 \cdot \log_{10}\left(\frac{P_E}{P_A}\right)$$

- In electronics, the power of a signal is usually proportional to the squared field amplitude
  (e.g. $P = u^2 / R$ for an ohmic resistor)

$$G = 10 \cdot \log_{10}\left(\frac{u_E^2}{u_A^2}\right) = 20 \cdot \log_{10}\left(\frac{u_E}{u_A}\right)$$

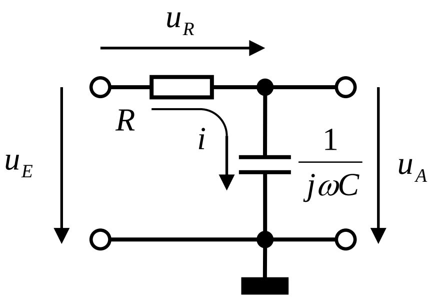
dB	**-20**	**-3**	**0**	**3**	**10**	**20**	**30**
power	1/100	1/2	1	2	10	100	1000
amplitude	1/10	1/√2	1	√2	√10	10	√1000

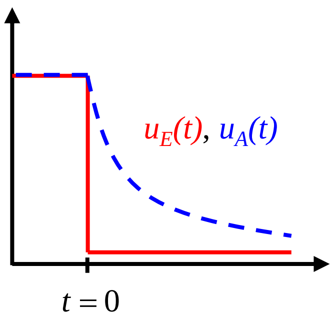
- Suffix indicates fixed reference → absolute quantity

  - dBV → relative to 1 V (0 dBV = 1 V, 20 dBV = 10 V)
  - dBm → relative to 1 mW (0 dBm = 1 mW, 30 dBm = 1W)
  - dBc → power ratio to carrier (e.g. for sideband modulation)

The low-pass filter

- Many physical systems show low-pass characteristic, i.e. they linearly transmit a signal with a certain time delay (PT1 element in controller theory)
- Exponential step response characteristic





input signal step $\quad u_E = \begin{cases} u_0 : t < 0 \\ 0 : \text{else} \end{cases}$

$\Rightarrow$ for $t > 0$ is $i_C + i_R = 0 \quad \Rightarrow \quad C\frac{du_A}{dt} + \frac{u_A}{R} = 0$

$\Rightarrow \quad u_A(t) = u_0 \cdot \exp(-t/\tau)$ with $\tau = RC$

- The complex transfer function $H(j\omega)$ relates the output $u_A(t)$ to the input $u_E(t)$ signal via a (magnitude) and a phase shift
- translate linear DE into frequency space via Laplace- / Fourier-transformation, transfer function is a fraction of polynomials
- poles and zero characterize transfer function (→ control theory)

ohm's law
$$u_E = i \cdot (Z_R + Z_C)$$
$$u_A = i \cdot Z_C$$

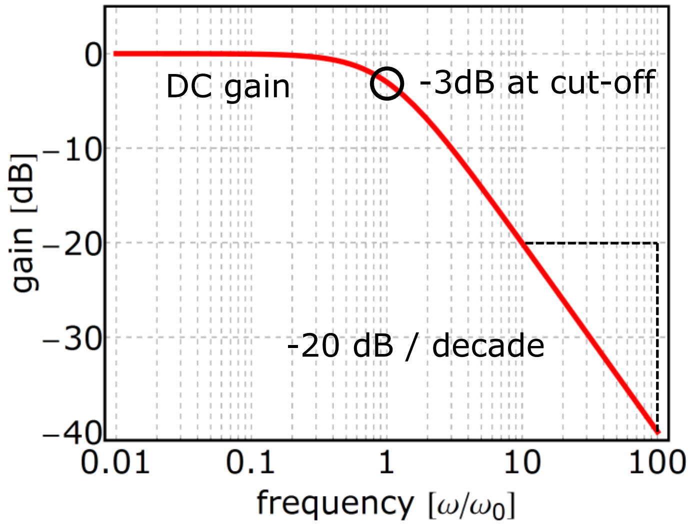
transfer function $\quad H(j\omega) = \frac{u_A}{u_E} = \frac{1/j\omega C}{R + 1/j\omega C} = \frac{1}{1 + j\omega RC}$

gain $\quad G(\omega) = |H(j\omega)| = \frac{1}{\sqrt{1 + (\omega/\omega_0)^2}}$ with $\omega_0 = \frac{1}{RC}$

phase $\quad \phi(\omega) = \arctan[H(j\omega)] = \arctan\left[-\frac{\omega}{\omega_0}\right]$

Bode plots

- Characterize transfer function by plotting magnitude (in dB) and phase on a log. frequency scale
- Polynomials appear as straight lines, poles and zeros can be seen as "kinks"



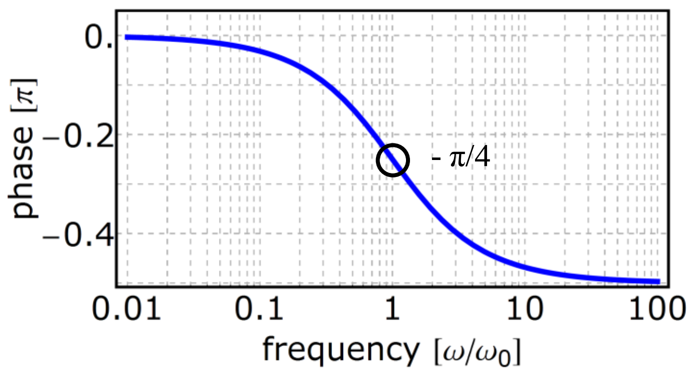
gain $$G(\omega) = \frac{1}{\sqrt{1+(\omega/\omega_0)^2}} \quad \text{with} \quad \omega_0 = \frac{1}{RC}$$

phase $$\phi(\omega) = \arctan\left[-\frac{\omega}{\omega_0}\right]$$

cut-off $$|Z_R| = |Z_C| \quad \Rightarrow \quad \omega = \frac{1}{RC} = \omega_0$$

output $$u_A = \frac{1}{\sqrt{2}} u_E$$

phase $$\phi = \arctan(-1) = -\pi/4$$

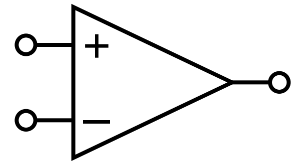


Pros and Cons of op-amps

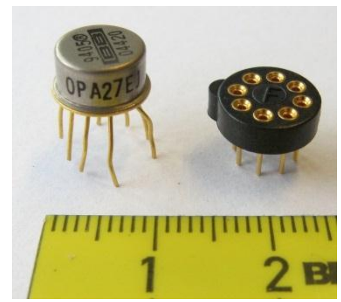
Basic idea:

Use integrated circuit (**black box** – internal realization unknown) as a modular device to manipulate signals, so that behavior of the circuit is purely characterized by external elements.

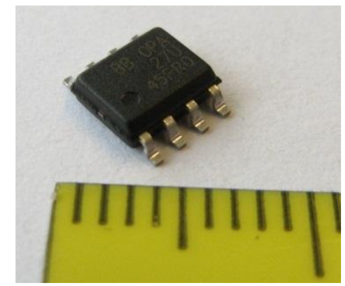
→ Why not use a transistor?



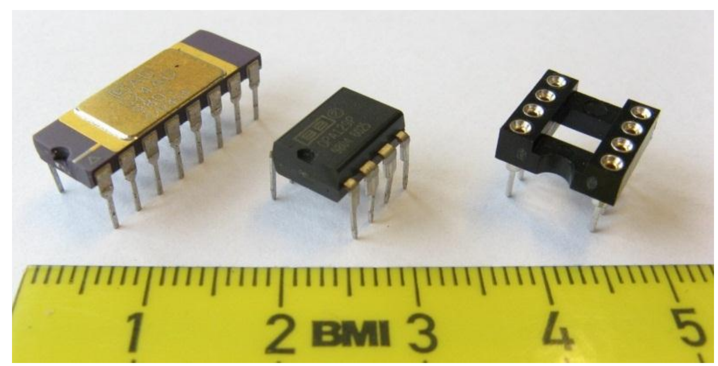
circuit symbol



TO-99 (transistor-single-outline)



SOIC-8 (small-outline-integrated-circuit)



CERDIP / PDIP (dual-inline-package)

Advantages of op-amps

- **Versatility** - operation only determined by external surrounding circuit
- No bias current needed to determine operation point
- High input impedance, low output impedance
- High intrinsic gain
- Very linear and precise amplification over broad voltage an frequency range

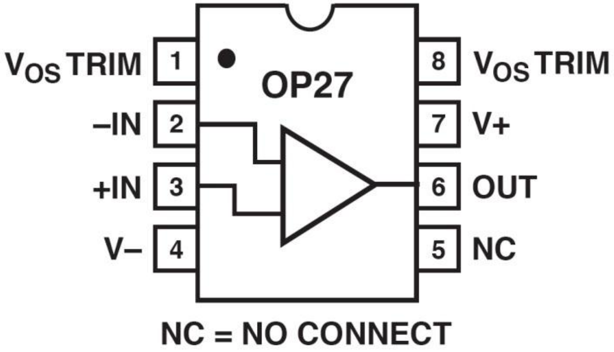
Disadvantages

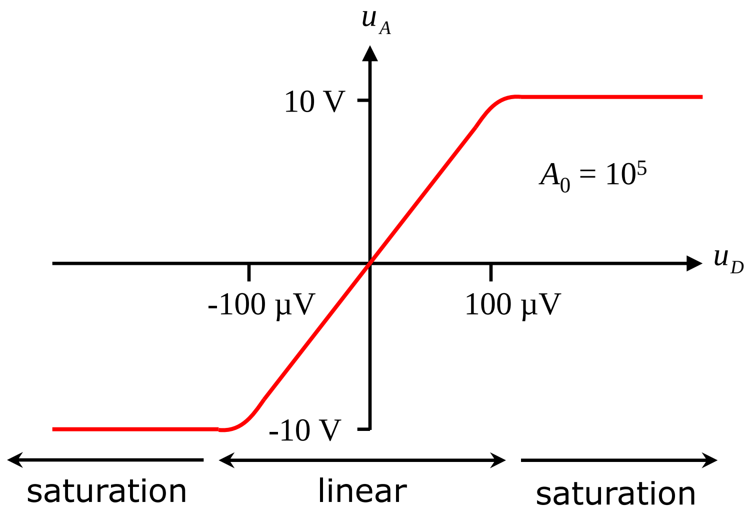
- higher noise due to multiple amplifier stages
- lower cut-off-frequency than single-stage transistor amplifiers
- Strong feedback over multiple amplifier stages causes non-linear distortions

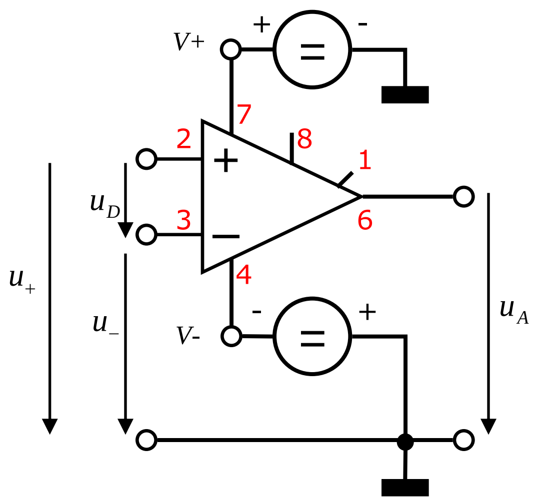
Basic operation and wiring

Typical pin assignment

- inverting (-) and noninverting (+) input (2,3)
- voltage difference $u_D = u_+ - u_-$ between inputs is amplified linearly by factor A_0 (open-loop-gain) to output $u_A = A_0 \cdot (u_+ - u_-)$ (6)
- every potential is measured relatively to ground potential (usually GND or 0V)
- two connections (4,7) for power supply $V\pm$ (usually ±15V)
- no separate ground connection, output ground reference by $u_A = 0$ for $u_+ - u_- = 0$ (for ideal op-amp)
- usually two extra pins for external offset compensation (8,1)

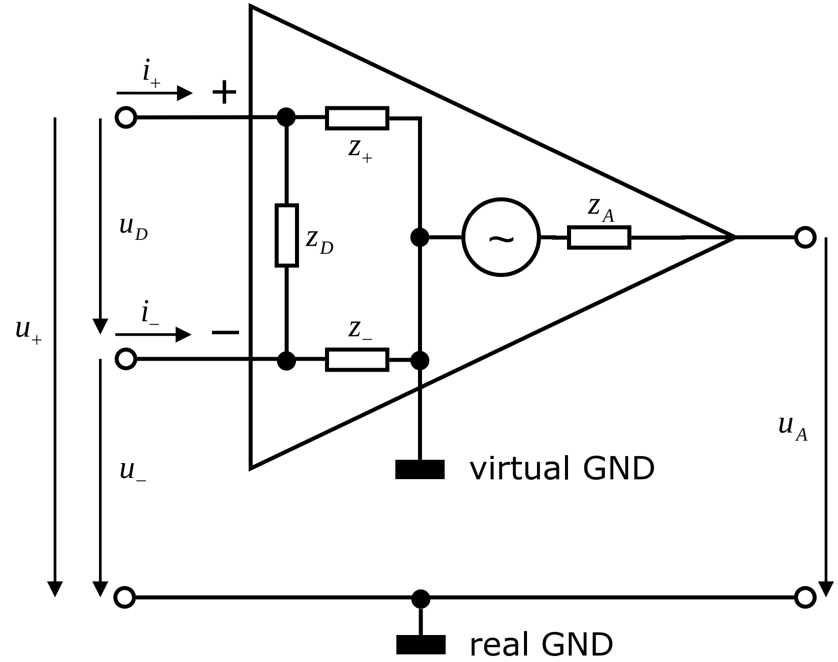






The ideal op-amp

Parameter	Symbol	Ideal	Real (OP27)
Input Impedance Differential-Mode	z_D	∞	4 MΩ || ≤ 1 pF
Input Impedance Common-Mode	z_+ , z_-	∞	2 GΩ || ≤ 1 pF
Input Bias Current	i_+ , i_-	0	±15 nA
Input Offset Voltage	u_0	0	30 μV
Output Impedance	z_A	0	70 Ω
Unity Gain Bandwidth	$f\,(A_0 = \text{-3 dB})$	∞	8 MHz
Open-Loop-Gain	$A_0 = du_A/du_D$	∞	10^6 = 120 dB
Common-Mode Rejection Ratio	$A_0 / du_A/d(u_+ + u_-)$	∞	10^6 = 120 dB
Slew rate	$\max(du_A/dt)$	∞	2.8 V/μs



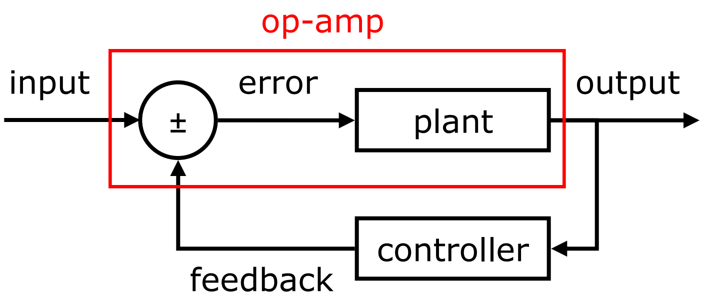
Basic idea of feedback

Problem

- open-loop-gain A_0 too high, input voltage range of ≤100 µV to small
- amplification A not adjustable
- gain A_0 determined by device (and hence differs between various devices), instabilities

Solution

- feed factor k_F of output signal back into input
- signal experiences certain amplification / attenuation and phase shift while passing the loop, effect determined by phase difference at inputs



Consequences

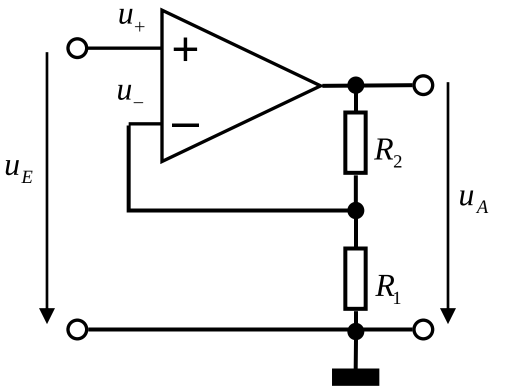
- reduced gain, but improved **linearity, frequency response, bandwidth and stability**
- more negative feedback results in less dependency on device parameters
- in general, feedback can be frequency-dependant (equalizers and filters) or amplitude-dependant (logarithmic amplifiers or multipliers)

The “golden rules” concerning op-amps

1. The output attempts to do whatever is necessary to make the voltage difference between the inputs zero (infinite open-loop-gain).
2. The inputs draws no current (infinite input impedance).

The noninverting amplifier

- realize controller by voltage divider R_2-R_1 , rising output voltage hinders input difference u_+-u_-



negative feedback $u_- = \frac{R_1}{R_1 + R_2} \cdot u_A$

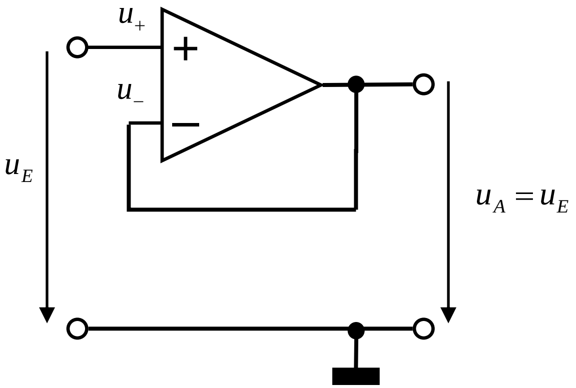
output signal $u_A = A_0(u_+ - u_-) = A_0(u_E - \frac{R_1}{R_1 + R_2} \cdot u_A)$

closed-loop-gain $A = \frac{u_A}{u_E} = \left(\frac{1}{A_0} + \frac{R_1}{R_1 + R_2} \right)^{-1}$

ideal OP-Amp $A \cong \frac{R_1 + R_2}{R_1} = 1 + \frac{R_2}{R_1}$

from feedback $u_A = A_0(u_+ - u_-) = \frac{R_1 + R_2}{R_1} \cdot u_-$

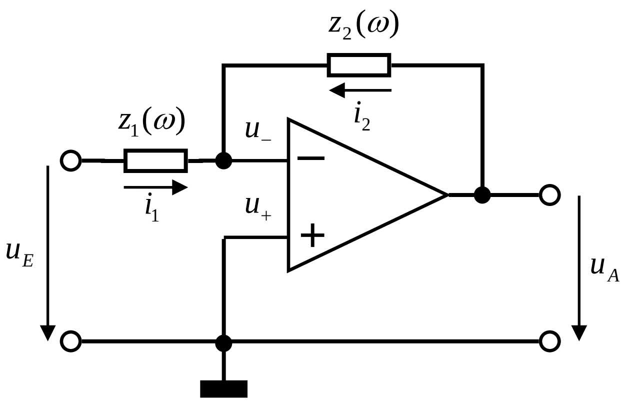
virtual short-circuit $u_+ = \underbrace{\left(\frac{1}{A_0} \cdot \frac{R_1 + R_2}{R_1} \right)}_{\approx 0} \cdot u_- + u_- \cong u_-$



- because generally $A_0 >> A$, the input voltage difference tends to zero **(virtual short-circuit)** (complies with **1st golden rule**)
- high input impedance ($10^{12}\Omega$), low output impedance ($10^{1}\Omega$)
- for $R_2 \rightarrow 0$ and $R_1 \rightarrow \infty$ the closed-loop-gain $A \rightarrow 1$, op-amp works as buffer (voltage follower)

The inverting amplifier

- connect input signal u_E to lower end of feedback voltage divider R_2-R_1
- assume input voltage $u_E = 1\text{V}$, but no output $u_A = 0\text{V}$
  - since noninverting input is connected to GND $u_+ = 0\text{V}$, op-amp sees high input unbalance
  - this forces the output u_A to go negative, until both input are on GND $u_+ = u_- = 0\text{V}$



kirchhoff's junction rule $\quad 0 = i_1 + i_2 + i_-$

open loop gain $\quad u_D = u_A / A_0$

high op-amp impedance $\quad 0 = i_-$

$$\Rightarrow \quad 0 = \frac{u_E - u_D}{z_1(\omega)} + \frac{u_A - u_D}{z_2(\omega)}$$

$$\Rightarrow \quad u_A \cdot z_1 + \frac{u_A \cdot z_2}{A_0} + \frac{u_A \cdot z_1}{A_0} = -u_E \cdot z_2$$

closed loop gain

$$A(\omega) = \frac{u_A}{u_E} = -\frac{z_2}{z_1 + z_2/A_0 + z_1/A_0} \approx -\frac{z_2(\omega)}{z_1(\omega)}$$

input impedance

$$z_E(\omega) = \frac{du_E}{di_E} = z_1(\omega)$$

- frequency-dependant (but linear) feedback provides active filters, integrators, differentiators
- non-linear feedback allows to build amplitude-dependant devices, i.e. exponential or logarithmic amplifiers

The summing amplifier

Problem

- summing currents is easy, but since all potentials are grounded, how to add potentials?

Solution

- take inverting amplifier and add currents in feedback loop
- assume that high input impedance $i_- = 0$ and virtual GND $u_+ \approx u_- = 0$

kirchhoff's junction rule

$$0 = i_1 + i_2 + \ldots + i_n + i_F$$

$$= \frac{u_1}{R_1} + \frac{u_2}{R_2} + \ldots + \frac{u_n}{R_n} + \frac{u_A}{R_F}$$

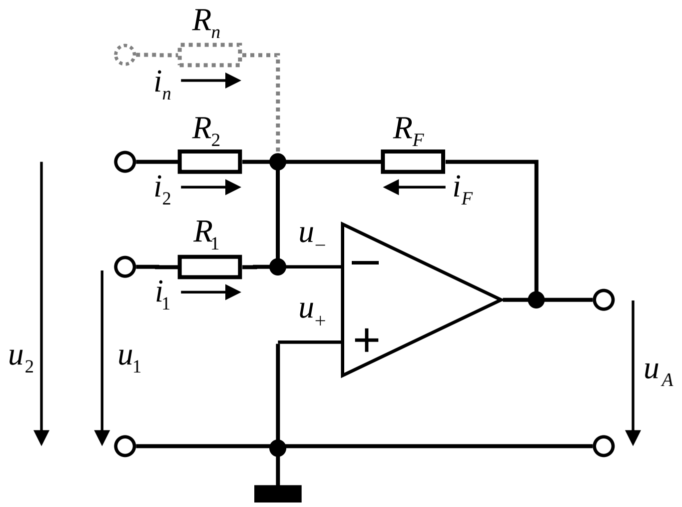
$$u_A = -R_F\left(\frac{u_1}{R_1} + \frac{u_2}{R_2} + \ldots + \frac{u_n}{R_n}\right)$$

assume equal inputs

$$R = R_1 = R_2 = \ldots = R_n \quad \Rightarrow$$

closed-loop-gain

$$u_A = -\frac{R_F}{R}\left(u_1 + u_2 + \ldots + u_n\right)$$



The differential amplifier

Problem

- measure voltage drop over some impedance, what if none of both connections has GND contact?

Solution

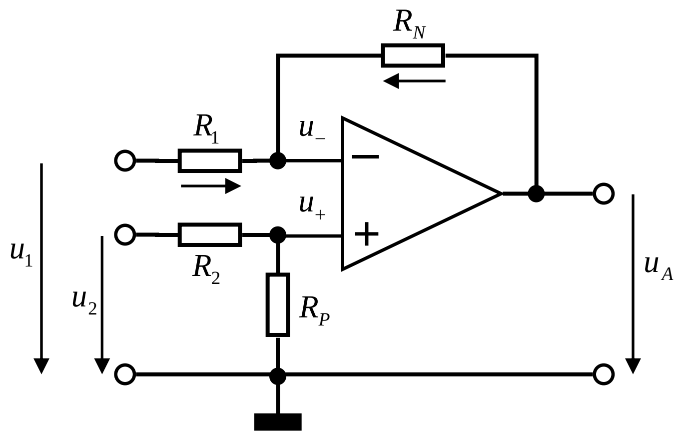
- use features from both inverting and noninverting amplifier

virtual short-circuit $u_- \cong u_+ = u_2 \cdot \frac{R_P}{R_2 + R_P}$

kirchhoff's junction rule $0 = i_1 + i_N = \frac{u_1 - u_-}{R_1} + \frac{u_A - u_-}{R_N}$

insert u_- $u_A = -\frac{R_N}{R_1} \cdot \left[u_1 - \frac{R_1/R_N + 1}{R_2/R_P + 1} \cdot u_2 \right]$

if $R_1 = R_2$ and $R_P = R_N$ $\boxed{u_A = -\frac{R_N}{R_1} \cdot (u_1 - u_2)}$



Caution

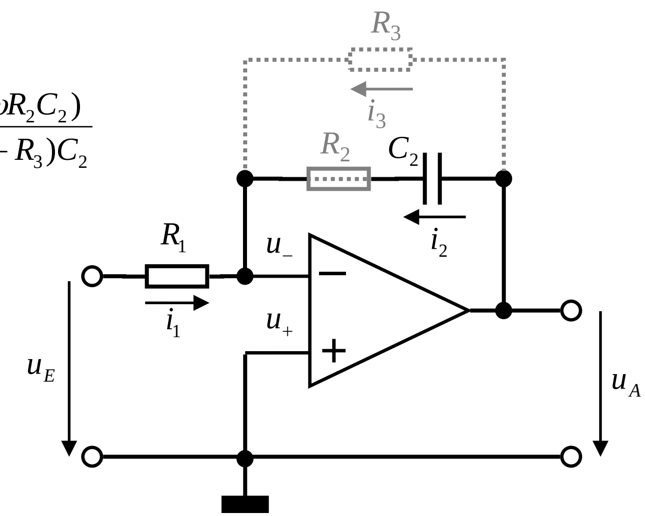
- assume $R_N = \alpha \cdot R_1$ and thereby the closed-loop-gain $A = -\alpha$, resistors have tolerance $\Delta\alpha$

$$\text{CMRR} = A \Big/ \frac{du_A}{d(u_+ + u_-)} \approx (1+\alpha)\frac{\alpha}{\Delta\alpha} \quad \Rightarrow \quad \text{error, if resistor values differ}$$

- different input impedances require source with low output impedance
- better results with instrumentation amplifier

The integrator – frequency response

- use inverting amplifier with RC-high-pass C_2-R_1 filter in feedback
- strong negative feedback for high frequencies (respectively weak feedback for low frequencies) results in **low-pass-characteristic**



use $z_1(\omega) = R_1$ and $z_2(\omega) = (C_2 + R_2) \| R_3 = \dfrac{R_3 \cdot (1 + j\omega R_2 C_2)}{1 + j\omega(R_2 + R_3)C_2}$

closed loop gain $$A(\omega) = -\frac{z_2(\omega)}{z_1(\omega)} = -\frac{R_3}{R_1} \cdot \frac{1 + j\omega R_2 C_2}{1 + j\omega(R_2 + R_3)C_2}$$

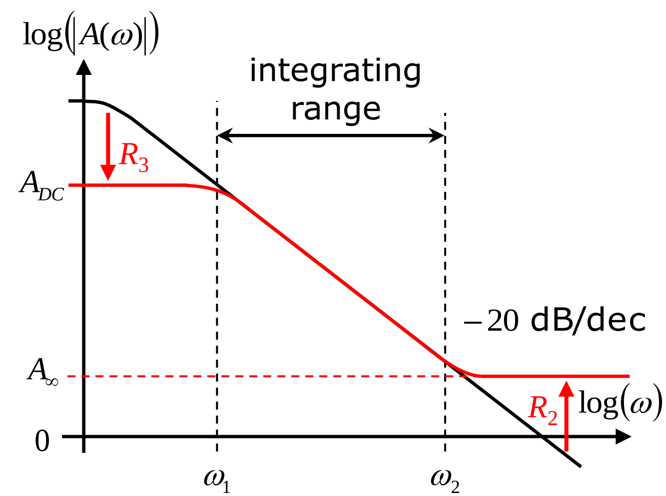
- missing feedback at $\omega = 0$ makes circuit instable, insertion of bypass R_3 limits feedback at low frequencies
- attenuation ($|A| < 1$) at high frequencies can be avoided by insertion of R_2

lower limit $$|A(\omega = 0)| = A_{DC} = -\frac{R_3}{R_1}$$

lower cut-off $$\left(|A(\omega_1)| = \frac{1}{\sqrt{2}} \cdot |A_{DC}|\right) \quad \omega_1 = \frac{1}{\sqrt{(R_3^{\,2} + 2R_3R_2 - R_2^{\,2}) \cdot C_2}}$$

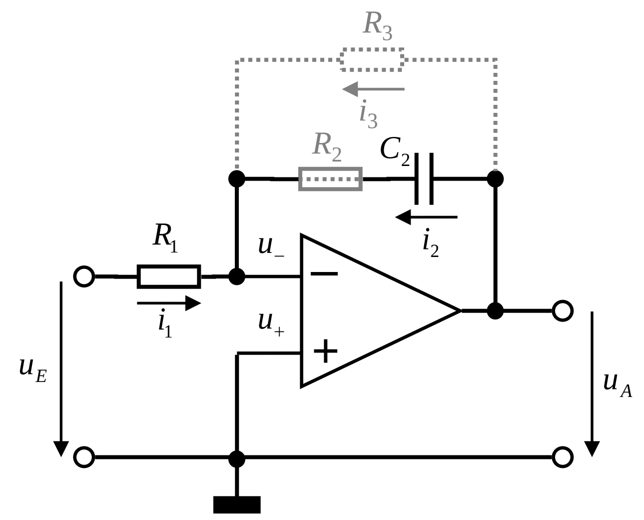
upper cut-off $$\left(|A(\omega_2)| = \sqrt{2} \cdot |A_\infty|\right) \quad \omega_2 = \frac{\sqrt{(R_3^{\,2} + 2R_3R_2 - R_2^{\,2})}}{R_2(R_2 + R_3) \cdot C_2}$$

upper limit $$|A(\omega = \infty)| = A_\infty = -\frac{R_2 \cdot R_3}{R_1 \cdot (R_2 + R_3)}$$



The integrator – operating method

closed loop gain $A(\omega) = -\frac{z_2(\omega)}{z_1(\omega)} = -\frac{R_3}{R_1} \cdot \frac{1 + j\omega R_2 C_2}{1 + j\omega (R_2 + R_3) C_2}$



kirchhoff's junction rule

$$0 = i_1 + i_2 + i_3$$
$$= \frac{u_E}{R_1} + C_2 \frac{d(u_A - u_{R2})}{dt} + \frac{u_A}{R_3}$$

Approximations

- for frequencies far above **lower limit** $\omega >> \omega_1$, current $i_3 = u_A / R_3$ can be neglected (shortened by C_2)
- for frequencies far beneath **upper limit** $\omega << \omega_2$, impedance of C_2-R_2 is dominated by condensator, so the voltage change du_{R2}/dt over R_2 can be neglected

$$0 = \frac{u_E}{R_1} + C_2 \frac{du_A}{dt} \quad \Rightarrow \quad \boxed{u_A(t) = -\frac{1}{R_1 C_2} \cdot \int_0^t u_E(t')\, dt' + u_A(0)}$$

- integration constant $u_A(0) = Q_0 / C$ is determined by initial condensator charge and capacity
- input bias current i_- and input offset voltage u_0 result in additional condensator current $u_0 / R + i_-$, which changes output voltage by

$$\frac{du_A}{dt} = \frac{1}{C_2} \cdot \left(\frac{u_0}{R} + i_- \right)$$

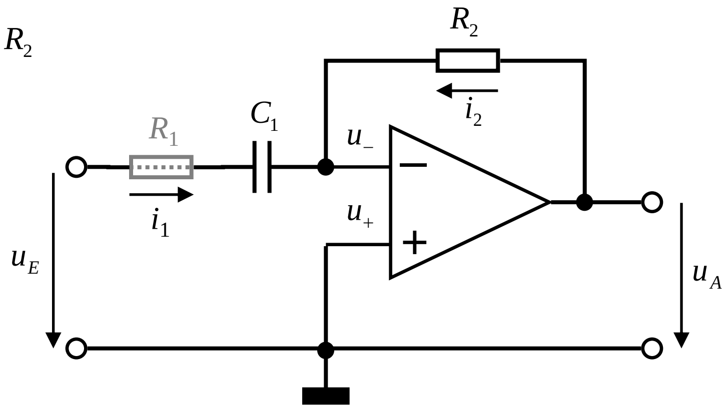
- $i_- = 1\ \mu A$ results with $C = 1\ \mu F$ in a change of $1\ V$ per second **(offset compensation)**

The differentiator – frequency response

- swap condensator and resistor in integrator
- RC-low-pass filter R_2-C_1 delivers in high negative feedback for low frequencies, resulting in a **high-pass characteristic**

use $z_1(\omega) = R_1 + \frac{1}{j\omega C_1} = \frac{1 + j\omega R_1 C_1}{j\omega C_1}$ and $z_2(\omega) = R_2$

closed loop gain $A(\omega) = -\frac{z_2(\omega)}{z_1(\omega)} = -\frac{j\omega R_2 C_1}{1 + j\omega R_1 C_1}$



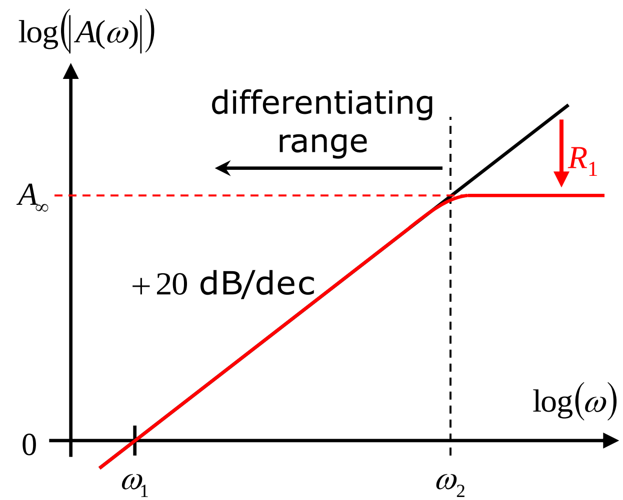
- at high frequencies ω, missing feedback makes circuit unstable
- HF noise is strongly amplified, circuit tends to oscillate due to phase delay of op-amp and feedback
- gain limiting at high frequencies can be achieved by insertion of R_1

lower limit $|A(\omega = 0)| = A_{DC} = 0$

lower cut-off $\left(|A(\omega_1)| = 1\right)$ $\omega_1 = \frac{1}{\sqrt{R_2^{\,2} - R_1^{\,2}} \cdot C_1}$

upper cut-off $\left(|A(\omega_2)| = \frac{1}{\sqrt{2}} \cdot |A_\infty|\right)$ $\omega_2 = \frac{1}{\sqrt{R_1 R_2 - R_1^{\,2}} \cdot C_1}$

upper limit $|A(\omega = \infty)| = A_\infty = -\frac{R_2}{R_1}$

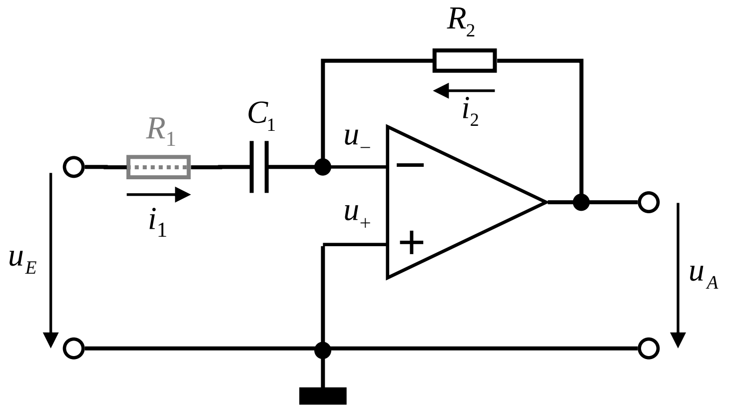


The differentiator – operating method

closed loop gain $A(\omega) = -\frac{z_2(\omega)}{z_1(\omega)} = -\frac{j\omega R_2 C_1}{1 + j\omega R_1 C_1}$

kirchhoff's $0 = i_1 + i_2$

junction rule $= C_1 \frac{d(u_E - u_{R1})}{dt} + \frac{u_A}{R_2}$



Approximation

- for frequencies far beneath **upper limit** $\omega << \omega_2$, impedance of R_1-C_1 is dominated by condensator, so the voltage change du_{R1}/dt over R_1 can be neglected

$$0 = C_1 \frac{du_E}{dt} + \frac{u_A}{R_2} \quad \Rightarrow \quad \boxed{u_A(t) = -R_2 C_1 \cdot \frac{du_E}{dt}}$$

- differentiator is bias-stable, optional roll-off-condensator in feedback can reduce bandwidth **(bandpass)**
- capacitive input impedance draws current from source, problems possible at high frequencies

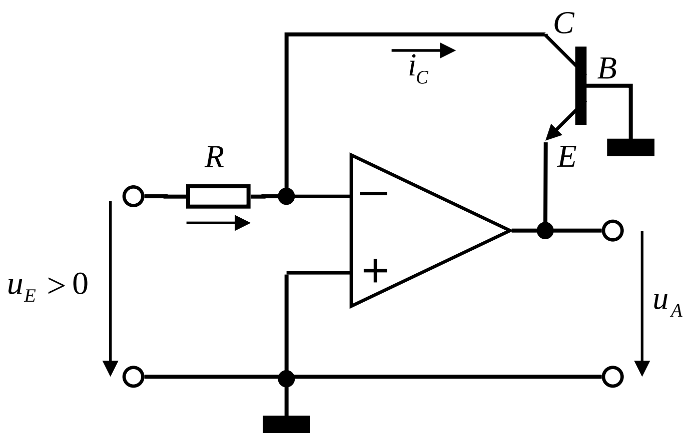
The logarithmic amplifier

Problem

- measured signal has large dynamic range

Idea

- instead of linear characteristic curve $i_R = u_R / R$ of a resistor, use exponential I-U-dependency of semiconductors
- use collector current i_C of a bipolar transistor



$$i_C = i_{CS}(T, u_{CE}) \cdot \exp(u_{BE}/u_T) \quad \Rightarrow \quad u_{BE} = u_T \cdot \ln(i_C / i_{CS})$$

i_{CS}	reverse leakage current
$u_T = kT/e$	thermal voltage

since $\quad u_A = -u_{BE} \quad , \quad i_C = u_E / R$

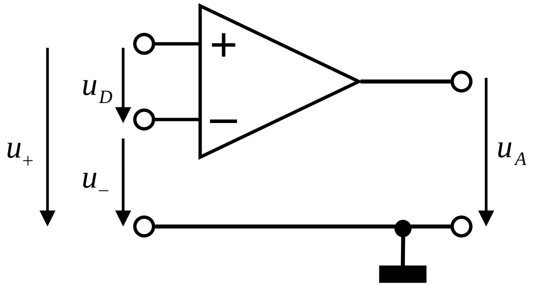
$$\Rightarrow \qquad u_A = -u_T \cdot \ln\left(\frac{u_E}{i_{CS} R}\right) \qquad \text{(for } u_E > 0\text{)}$$

- no error due to collector-base-current i_{CB}, since $u_{CB} = 0$, but **temperature drift**
- with appropriate transistor and op-amp with low bias-current, usually **nine decades** available
- swap resistor and transistor: **exponential amplifier**
- **multiply / divide signals** by taking logarithm, add / subtract, take exponential value

The comparator

Problem

- device for comparison of two signals, which decides if voltage is below / above some threshold

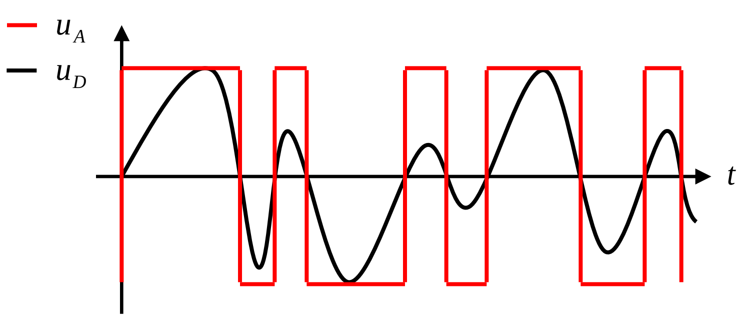


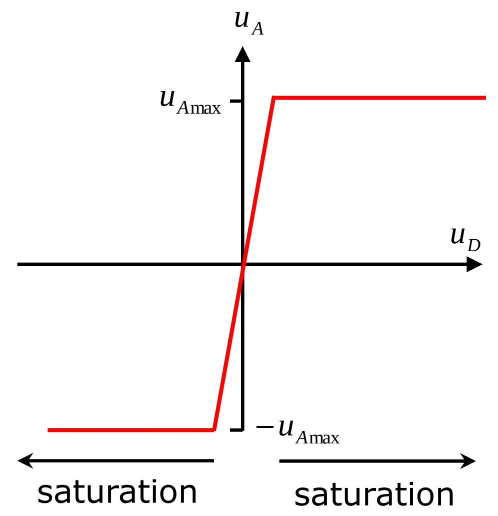
Solution

- use op-amp without feedback to compare voltages **(comparator)**
- because of high open-loop-gain, circuit is sensitive to small voltage differences u_D and switches between saturated values -u_{Amax} and u_{Amax}
- **in saturation no virtual short-circuit between inputs, i.e.** $u_+ \neq u_-$

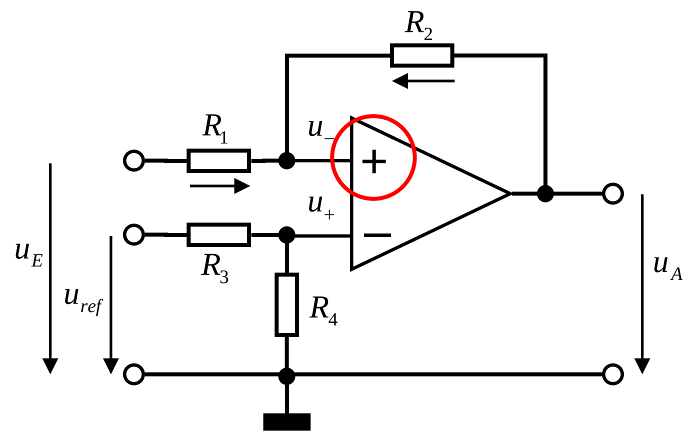
$$u_A = \begin{cases} +u_{A\max} & \text{for } u_+ > u_- \\ -u_{A\max} & \text{for } u_+ < u_- \end{cases}$$

- useful for regeneration of digital signals or as trigger
- special devices for fast applications





The noninverting Schmitt trigger



Problem

- **comparator** has no well-defined output for small input signals $u_D \approx 0$
- different switching values for high and low state desired

Solution

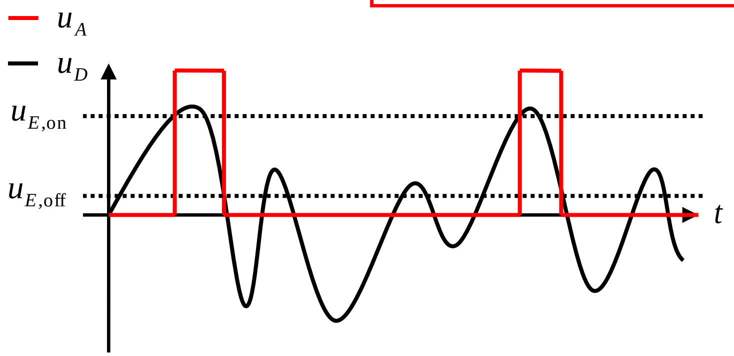
- use positive feedback to create hysteresis for switching

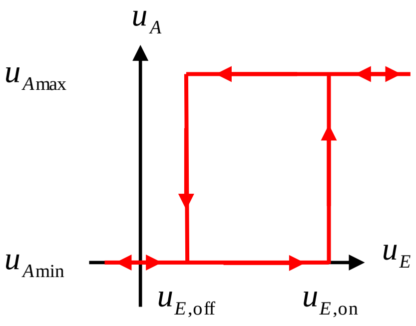
input $$u_- = \frac{R_4}{R_3 + R_4} \cdot u_{ref} \qquad u_+ = \frac{R_2}{R_1 + R_2} \cdot (u_E - u_A)$$

- assume high positive input u_E, so that output is $u_{A\max}$
- with falling input voltage u_E, output u_A remains unchanged until $u_+ = u_- = 0$

upper switching point $$u_{E,\text{on}} = \frac{R_4 \cdot (R_1 + R_2)}{R_2 \cdot (R_3 + R_4)} \cdot \left(u_{ref} + u_{A\max}\right)$$

lower switching point $$u_{E,off} = \frac{R_4 \cdot (R_1 + R_2)}{R_2 \cdot (R_3 + R_4)} \cdot \left(u_{ref} + u_{A\min}\right)$$





The real op-amp

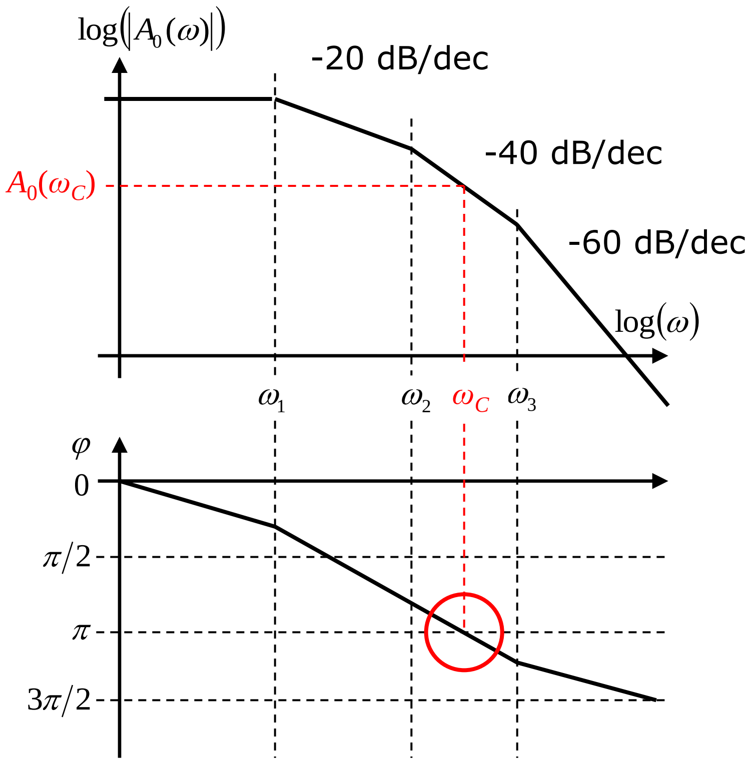
- differential **open-loop-gain** $A_0 = du_A/du_D$ is usually not infinite -> error in approx. $A_0 = \infty$
- even with shortened inputs $u_+ - u_- = 0$, op-amp amplifies common-mode voltage $u_+ + u_-$, usually expressed by ratio of differential open-loop-gain A_0 to $du_A/d(u_+ + u_-)$
  -> **common-mode rejection ratio**
- **finite input impedance** draws current from source, common-mode input impedance (input to GND) usually negligible, effect compensated if adjusted
- **nonzero output impedance** negligible, since decrease of output u_A by output load is compensated by feedback
- transistors as well as resistors produce **noise**, which is amplified towards the output

Parameter	Symbol	Ideal	Real (OP27)
Input Impedance Differential-Mode	z_D	∞	4 MΩ || ≤ 1 pF
Input Impedance Common-Mode	z_+ , z_-	∞	2 GΩ || ≤ 1 pF
Input Bias Current	i_+ , i_-	0	±15 nA
Input Offset Voltage	u_0	0	30 μV
Output Impedance	z_A	0	70 Ω
Unity Gain Bandwidth	$f\,(A_0$=-3 dB)	∞	8 MHz
Open-Loop-Gain	$A_0 = du_A/du_D$	∞	10^6 = 120 dB
Common-Mode Rejection Ratio	$A_0 / du_A/d(u_+ + u_-)$	∞	10^6 = 120 dB
Slew rate	$\max(du_A/dt)$	∞	2.8 V/μs

Real op-amps - frequency response

- real op-amps are multi-stage transistor-amplifiers
- every single stage represents a **low-pass filter** with a particular cut-off-frequency ω, that **reduces the gain by -20 db per decade** and adds a certain **phase shift of maximal $\varphi = \pi/2$** (subsequent stages usually have increasing bandwidths)
- if $A_0(\omega)$ is open-loop-gain, and $|k_F| \leq 1$ is feedback factor, requirement for oscillation is (negative input adds phase shift $\varphi = \pi$)

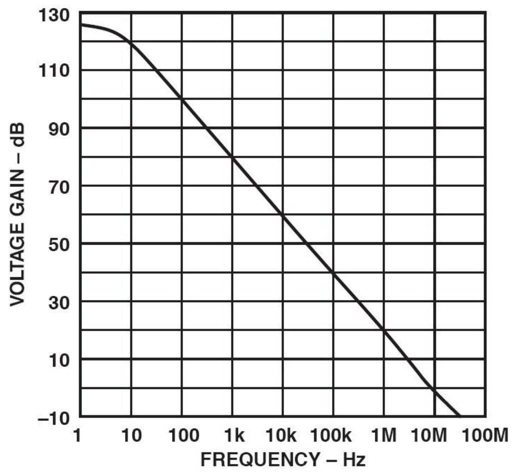
$$k_F(\omega) \cdot A_0(\omega) = 1 \quad \Rightarrow \quad \begin{cases} |k_F| \cdot |A_0| = 1 \\ \varphi = \arg(k_F A_0) = 0, 2\pi, \ldots \end{cases}$$



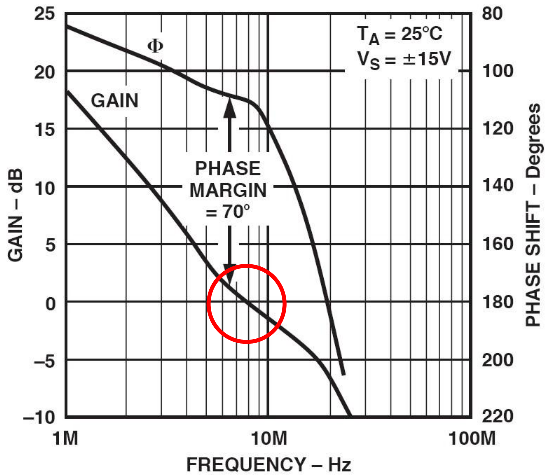
- to prevent oscillation at ω_C, feedback factor has to be maximal $k_F \leq 1 / A_0(\omega_C)$ (signal gain in one loop passage must be smaller than one)

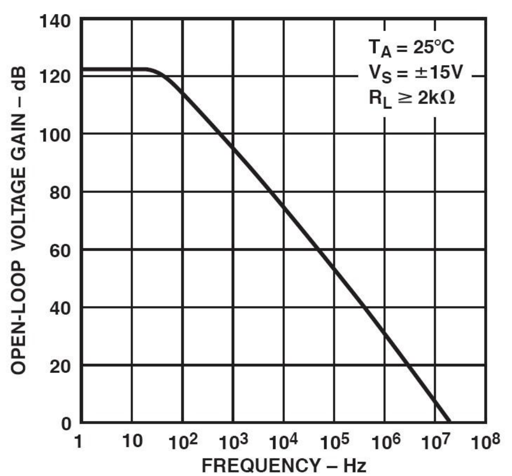
If an op-amp is **not internally frequency-compensated** (and by that **not unity-gain stable**), external frequency compensation or minimal closed-loop-gain is necessary.

Comparison OP27 – OP37

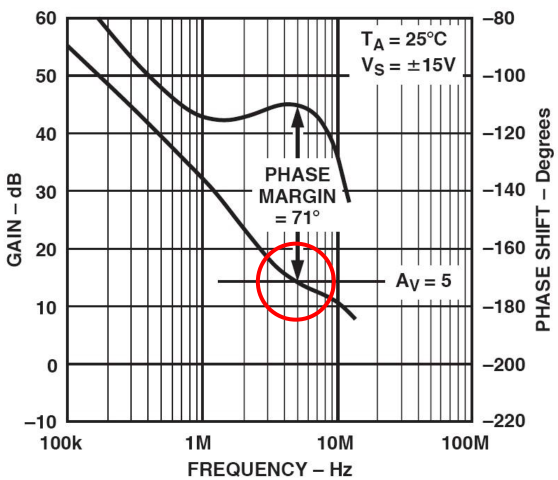


OP27 (unity-gain stable)





OP37 (not unity-gain stable)



Bandwidth

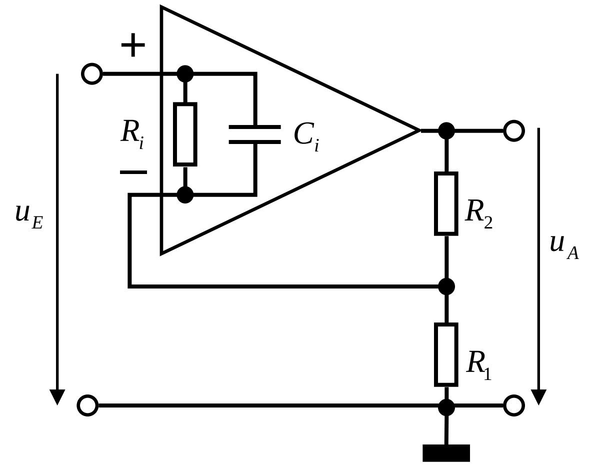
- frequency-compensated op amps can be regarded as 1st order low-pass

$$A_0(\omega) = \frac{A_{DC}}{1 + i \cdot (\omega/\omega_1)} \quad \Rightarrow \quad \omega_1 = R_i C_i$$

- for frequencies $\omega >> \omega_1$ above cut-off, open-loop gain is approximately

$$A_0(\omega) \cong -i(\omega_1 \cdot A_{DC}/\omega)$$

$\omega_1 \cdot A_{DC}$ ... gain-bandwidth-product (unity-gain bandwidth)



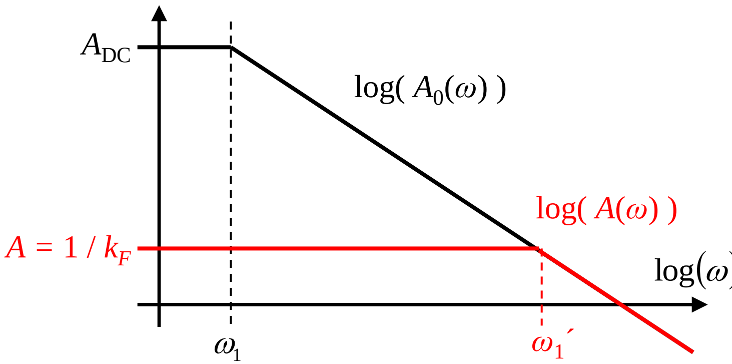
- usually A_{DC} is in the range of 120 dB, but cut-off frequency very low $\omega_1 / 2\pi = 10$ Hz
- use closed-loop gain from noninverting amplifier with feedback factor $k_F = R_1 / (R_1 + R_2)$ to calculate new frequency response

$$A(\omega) = \frac{u_A}{u_E} = \left(\frac{1}{A_0(\omega)} + \frac{R_1}{R_1 + R_2} \right)^{-1} = \frac{A_0(\omega)}{1 + k_F A_0(\omega)} \quad \text{insert } A_0(\omega)$$

$$A(\omega) = \frac{A}{1 + i \cdot (\omega/\omega_1')} = \begin{cases} A & \text{for } \omega << \omega_1' \\ -i(\omega_1'/\omega) \cdot A & \text{for } \omega >> \omega_1' \end{cases}$$

$A = \frac{1}{k_F} = \frac{R_1}{R_1 + R_2}$ (DC) closed-loop gain

$\omega_1' = k_F \omega_1 A_{DC}$ closed-loop cut-off frequency



Slew Rate

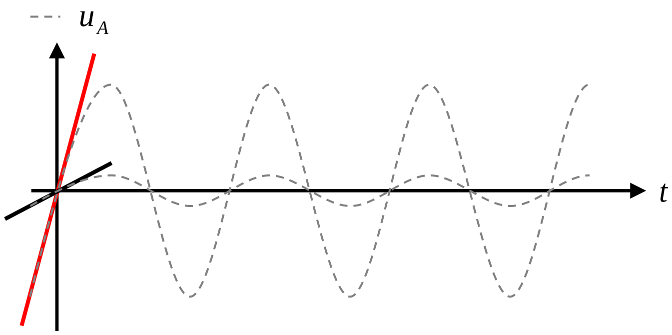
- internal capacities and frequency compensation act as a **low-pass**, whose output for a unit-step input is a (more or less fast) exponential function
- output transistors and capacities need driving current from (previous) driving stage, whose **output current is limited**
- output change rate is limited to the **slew rate** **SR = max(du_A/dt)**
- image sine-wave input signal

$$u_E = u_{E\max} \cdot \sin(\omega \cdot t) \quad \Rightarrow \quad \max\left(\frac{du_A}{dt}\right) = A(\omega) \cdot \omega \cdot u_{E\max}$$

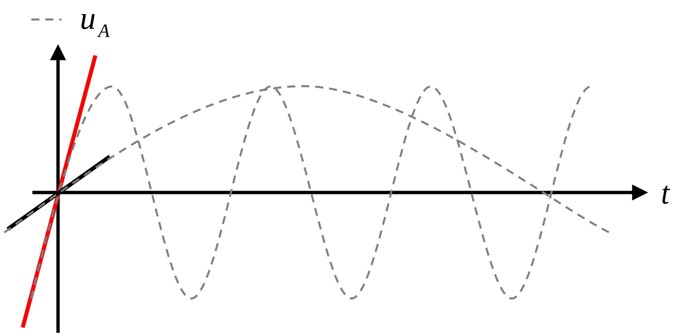
- **slew rate limits amplitude of undistorted sine-wave output swing above some critical frequency (power bandwidth)**

$$u_{E\max} = \frac{SR}{\omega}$$

amplitude-dependant distortion



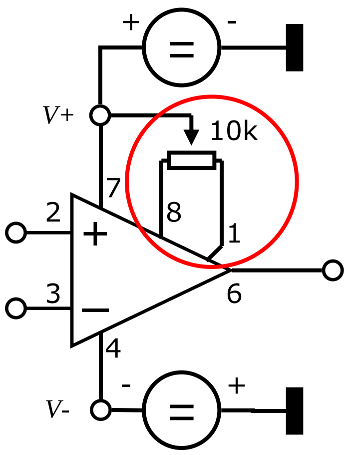
frequency-dependant distortion



Input offset voltage and bias current

Offset voltage

- output signal $u_A \neq 0$ even for no input signal $u_+ = u_- = 0$, due to unsymmetrical differential-amplifier at input (usually several µV)
- **amplified** towards output
- **can be compensated** by potentiometer at extra pins
- problem is **temperature drift** of usually $1\ \mu V/°C$



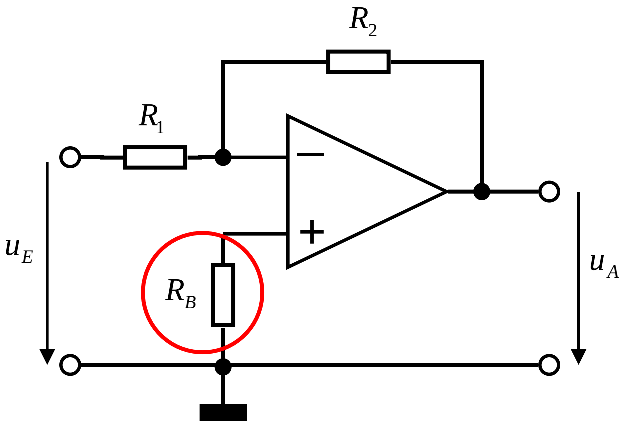
Bias current

- input transistors need **constant base- or gate-current** for operation, which is delivered by power supply

Bipolar	100 nA
Darlington	1 nA
FET	1 pA

- **can be compensated** by additional bias resistor R_B

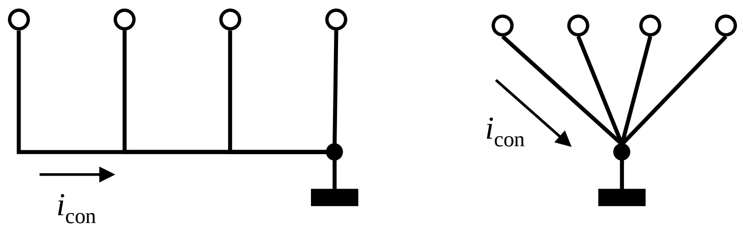
$$R_B = \frac{R_1 R_2}{R_1 + R_2}$$



Grounding problems

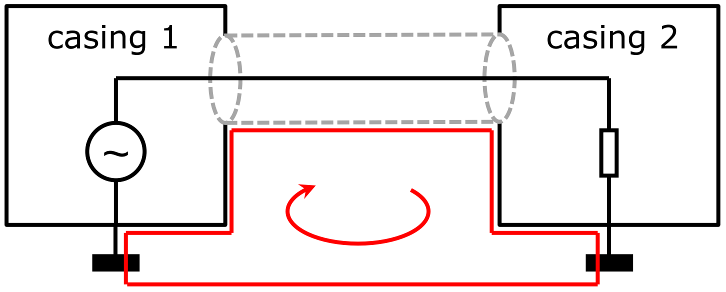
Ground connections

- Conductive tracks have **nonzero resistance**, so flowing currents produce a voltage drop (ground shift) -> **separate signal ground from consumer ground**
- If possible, connect every device/component **starlike** to one common ground reference

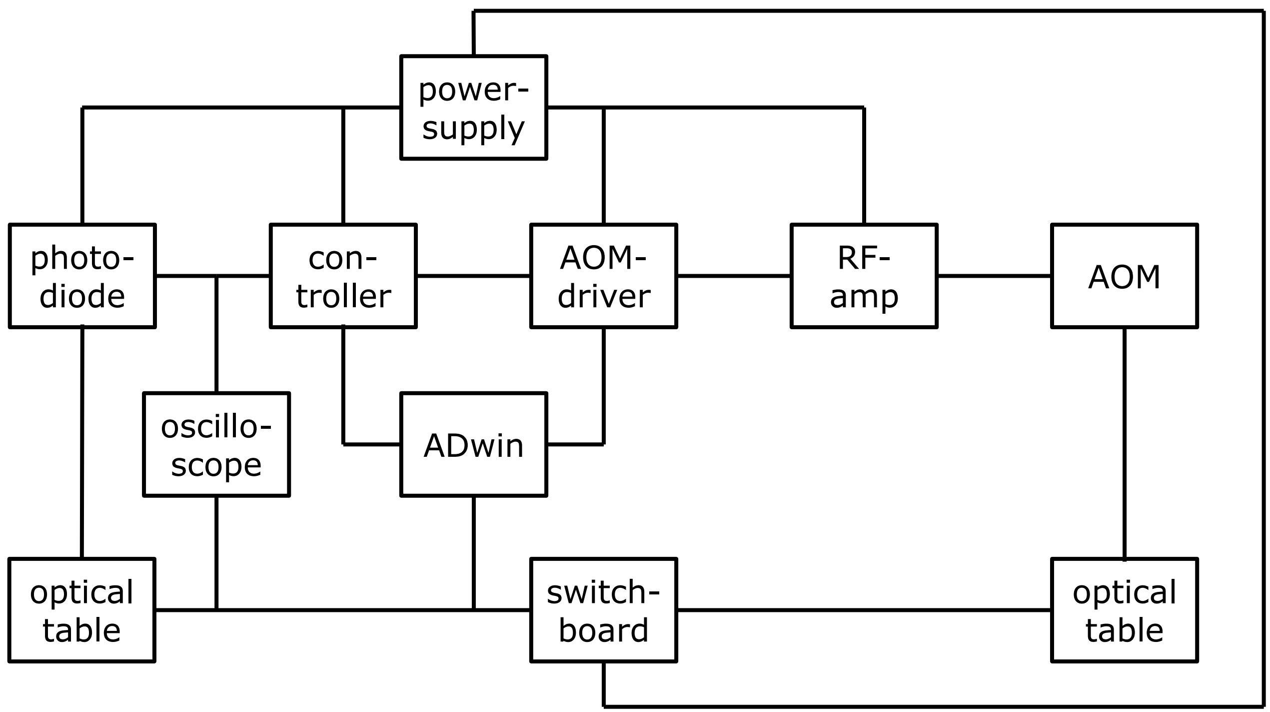


Ground loops

- RF-signal need shielding to avoid crosstalk → use coaxial cables as waveguides
- **Problem:** DC-signals need two wires, i.e. a clear voltage reference (usually GND)
  - → shielding provides conductor loop between two grounded casings
- Time-varying stray fields from **transformers** induce voltages / currents in loop
  - → a 10 µT stray field of a 50 Hz transformer induces **~100 mV** in a 10 m² conductor loop



Avoiding ground loops



1. Don’t use shielding as conductor, i.e. signal path
2. Isolate all casings from metal surfaces
3. Use separate power supplies
4. Power supplies don’t need a ground reference
5. Use digital / analog isolators (e.g. ADUM524x and AD210)
6. Measure signals differentially
7. No oscilloscopes at unbuffered / source signals

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