AGH University of Science and Technology
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ELECTRONIC DEVICES

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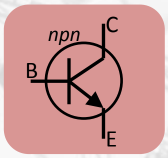
BIPOLAR JUNCTION TRANSISTOR

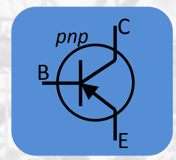
BIPOLAR JUNCTION TRANSISTOR
INTRODUCTION

Do you know that ....:

.... until recently, bipolar transistor has been the most widely used semiconductor device. By saying the word "transistor" it was meant a bipolar transistor.



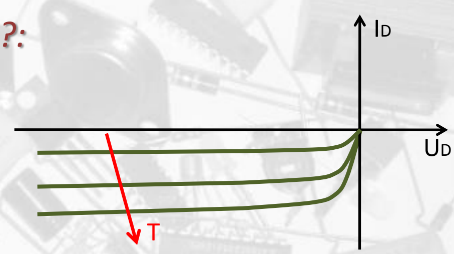
.... the current flowing between the two terminals of the bipolar transistor is controlled by a relatively small current flowing through the third terminal.

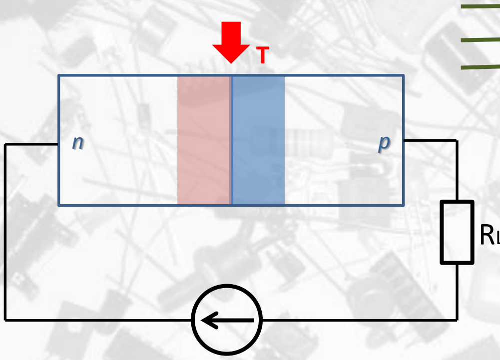


.... in bipolar transistor, both electrons and holes are involved in current flow.

BIPOLAR JUNCTION TRANSISTOR
INTRODUCTION

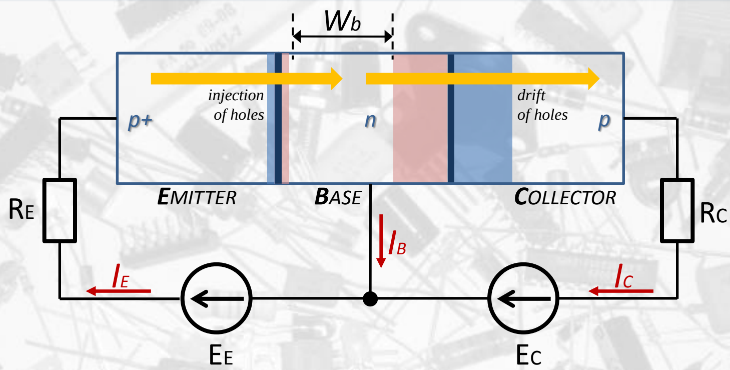
How was it with a diode ......?:





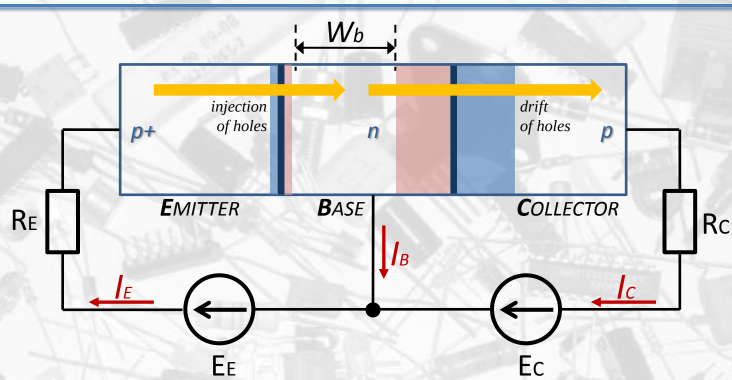
Other ways of increasing the current ......???

BIPOLAR JUNCTION TRANSISTOR
INTRODUCTION



In a good pnp transistor almost all the holes injected from the emitter to the base are „collected" in the collector. This assumption is fulfilled in case of the narrow base ($W_b \ll L_p$) and the long lifetime of holes τ_p.

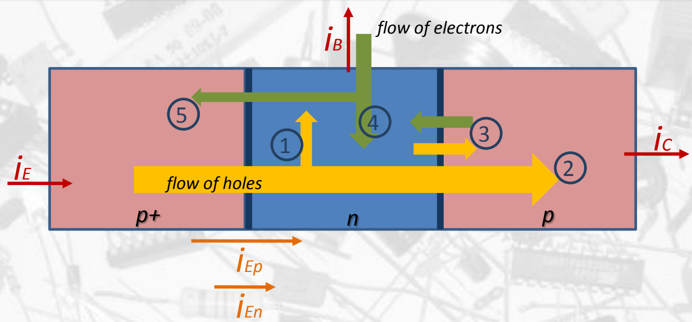
BIPOLAR JUNCTION TRANSISTOR
INTRODUCTION



The base current flow consist of:

1. Current of electrons recombining with holes in the base.
2. Current of electrons injected into the emitter despite the emitter is heavily doped than the base.
3. A small current of electrons (resulting from thermal generation) flowing into the base from the reverse polarized collector junction.

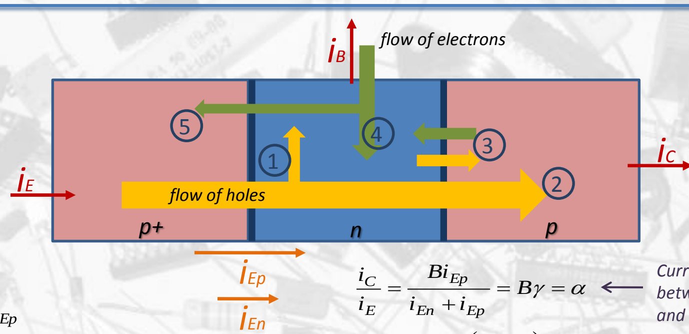
BIPOLAR JUNCTION TRANSISTOR
BALANCE OF ELECTRONS AND HOLES FLOW



1. injected holes that are lost during recombination in the base
2. holes reaching the reverse biased collector junction
3. thermal generation of electrons and holes forming the saturation current of the reverse biased collector
4. electrons supplied by the base contact and recombining with holes
5. electrons injected into the emitter through the junction

from: „Przyrządy półprzewodnikowe", Ben G. Streetman

BIPOLAR JUNCTION TRANSISTOR
CURRENT GAIN



$$i_C = Bi_{Ep}$$

Base transport factor (what part of the injected holes reached the collector via the base)

$$\gamma = \frac{i_{Ep}}{i_{En} + i_{Ep}}$$

The emitter injection efficiency

$$\frac{i_C}{i_E} = \frac{Bi_{Ep}}{i_{En} + i_{Ep}} = B\gamma = \alpha \leftarrow$$

Current gain between collector and emitter

$$i_B = i_{En} + (1 - B)i_{Ep}$$

$$\frac{i_C}{i_B} = \frac{Bi_{Ep}}{i_{En} + (1 - B)i_{Ep}} = \frac{B\left[i_{Ep}/\left(i_{En} + i_{Ep}\right)\right]}{1 - B\left(i_{Ep}/\left(i_{En} + i_{Ep}\right)\right)}$$

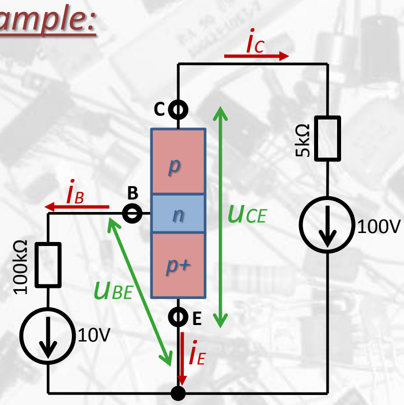
$$\frac{i_C}{i_B} = \frac{B\gamma}{1 - B\gamma} = \frac{\alpha}{1 - \alpha} = \beta = \frac{\tau_p}{\tau_t}$$

from: „Przyrządy półprzewodnikowe", Ben G. Streetman

BIPOLAR JUNCTION TRANSISTOR

COMMON EMITTER AMPLIFIER – QUALITATIVE DESCRIPTION

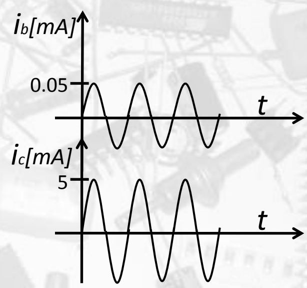
Example:



$$\frac{i_C}{i_B} = \beta = \frac{\tau_p}{\tau_t} = 100$$

$$I_B = \frac{10V}{100k\Omega} = 0.1mA$$

$$I_C = \beta I_B = 10mA$$



$$\tau_p = 10\mu s$$

$$\tau_t = 0.1\mu s$$

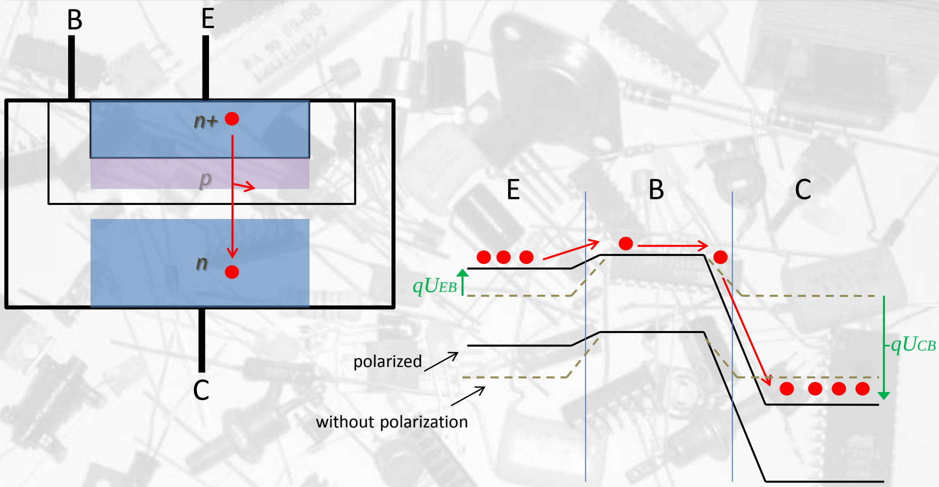
BIPOLAR TRANSISTOR

STRUCTURES

E- emitter B – base C - collector

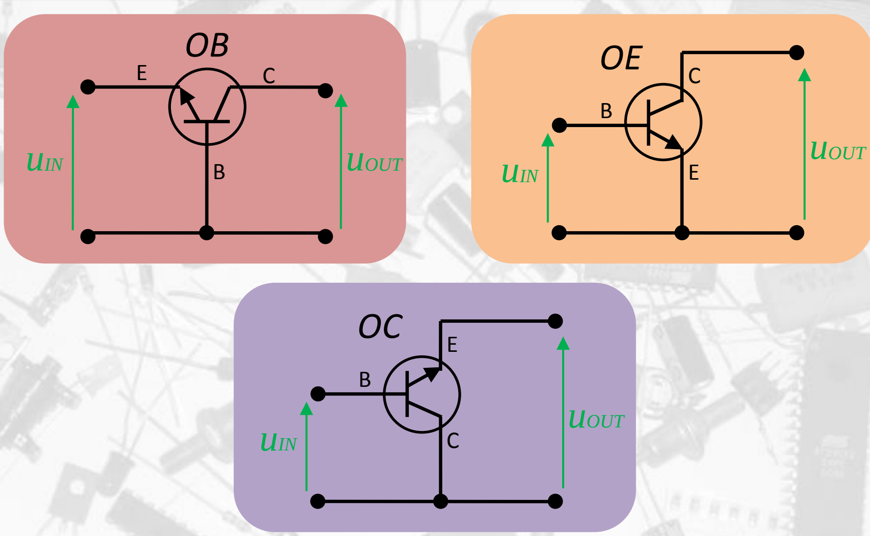
BIPOLAR TRANSISTOR

ENERGY BAND MODEL



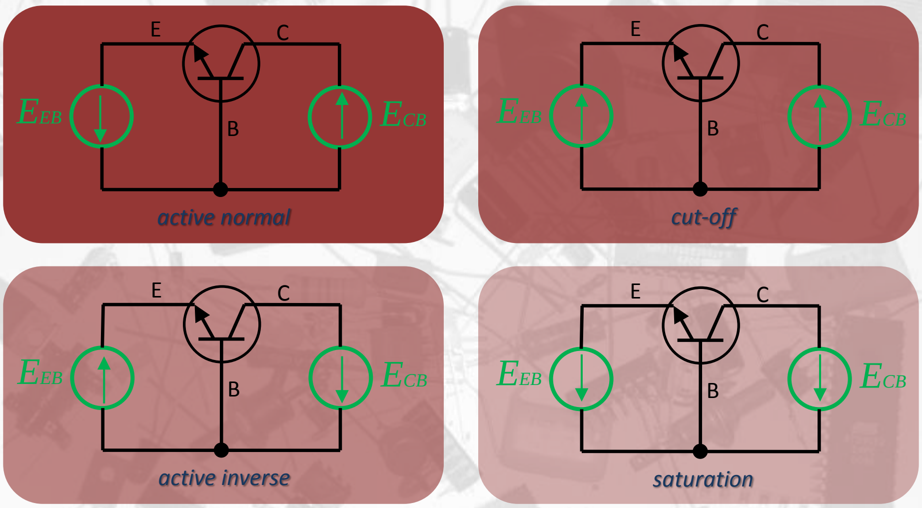
BIPOLAR TRANSISTOR

OPERATION CONFIGURATIONS OF THE BIPOLAR TRANSISTOR



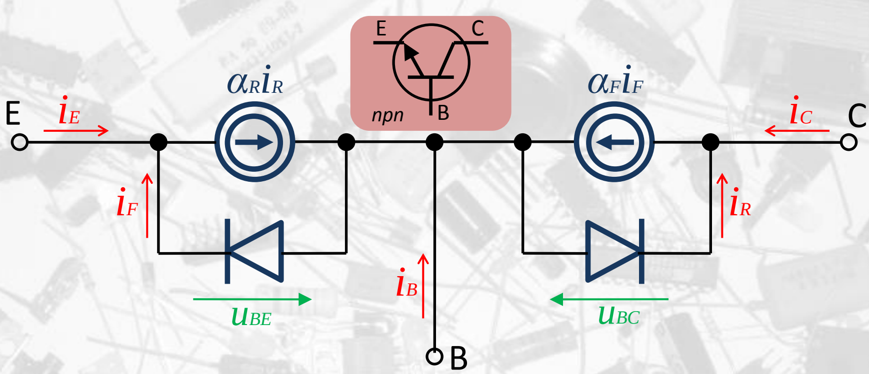
BIPOLAR TRANSISTOR

OPERATION STATES OF THE BIPOLAR TRANSISTOR



BIPOLAR JUNCTION TRANSISTOR

EBERS-MOLL MODEL



$$i_E = -I_{ES}\left(e^{\frac{u_{BE}}{n_E U_T}} - 1\right) + \alpha_R I_{CS}\left(e^{\frac{u_{BC}}{n_C U_T}} - 1\right)$$

$$i_C = \alpha_F I_{ES}\left(e^{\frac{u_{BE}}{n_E U_T}} - 1\right) - I_{CS}\left(e^{\frac{u_{BC}}{n_C U_T}} - 1\right)$$

BIPOLAR JUNCTION TRANSISTOR
EBERS-MOLL MODEL

I_{ES} – reverse saturation emitter current at shorted collector junction

$$I_{ES} = \frac{I_{E0}}{1-\alpha_F \alpha_R}$$

I_{CS} – reverse saturation collector current at shorted emitter junction

$$I_{CS} = \frac{I_{C0}}{1-\alpha_F \alpha_R}$$

n_E, n_C – non-ideality factors of emitter and collector junctions

α_F – DC current gain of the transistor working in active normal configuration in OB mode

$$I_C = -\alpha_F I_E + I_{C0} \qquad \alpha_F = \frac{I_C - I_{C0}}{I_E}$$

α_R - DC current gain of the transistor working in active reverse configuration in OB mode

BIPOLAR JUNCTION TRANSISTOR
EBERS-MOLL MODEL

$\alpha_F I_{ES} = \alpha_R I_{CS} \equiv I_S$ Onsager's identity

I_S – transport saturation current

$$i_E = -\frac{I_S}{\alpha_F}\left(e^{\frac{u_{BE}}{n_E U_T}} - 1\right) + I_S\left(e^{\frac{u_{BC}}{n_C U_T}} - 1\right)$$

$$i_C = I_S\left(e^{\frac{u_{BE}}{n_E U_T}} - 1\right) - \frac{I_S}{\alpha_R}\left(e^{\frac{u_{BC}}{n_C U_T}} - 1\right)$$

E-M equations dependent only on three parameters

BIPOLAR JUNCTION TRANSISTOR
EBERS-MOLL MODEL

If we define as i_F the forward current of a emitter diode in active normal mode and the i_R as collector diode current in active inverse mode, then:

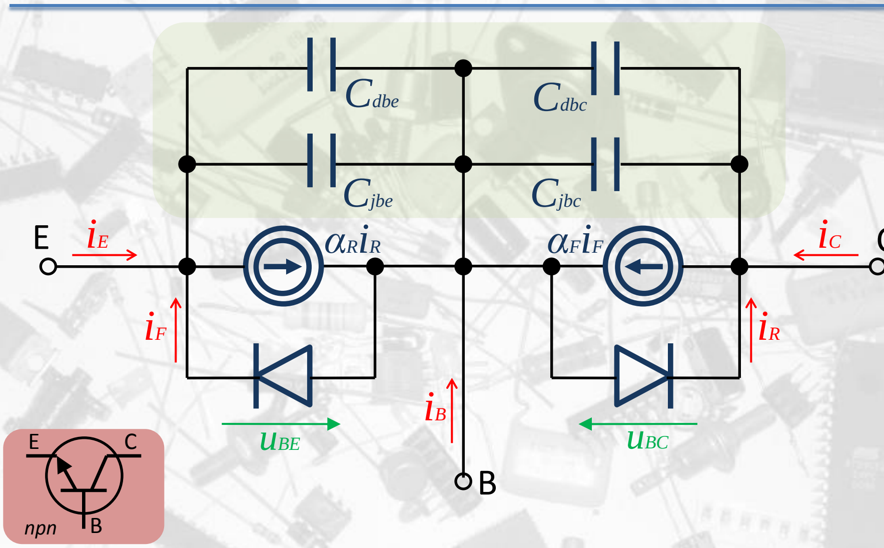
$$i_F = I_{ES}\left(e^{\frac{u_{BE}}{n_E U_T}} - 1\right) \approx I_{ES}\left(e^{\frac{u_{BE}}{n_E U_T}}\right)$$

$$i_R = I_{CS}\left(e^{\frac{u_{BC}}{n_E U_T}} - 1\right)$$

we get the E-M equations in the form of:

$$i_E = -i_F + \alpha_R i_R$$
$$i_C = \alpha_F i_F - i_R$$

BIPOLAR JUNCTION TRANSISTOR
EBERS-MOLL MODEL



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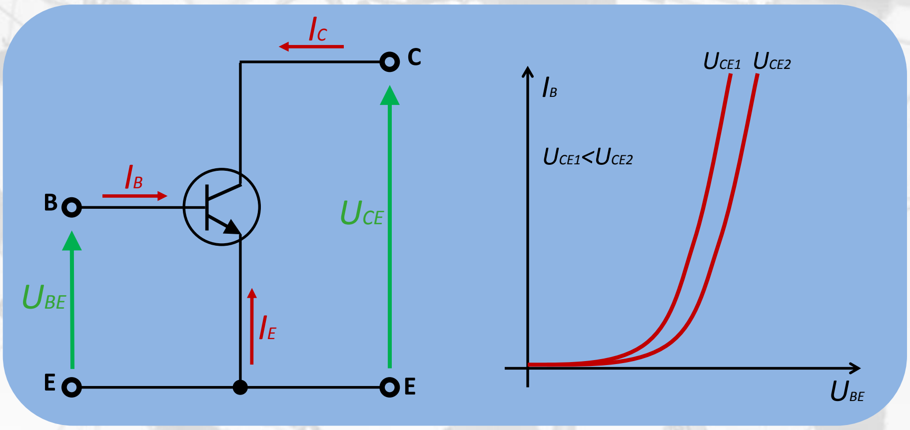
BIPOLAR JUNCTION TRANSISTOR
OE MODE - CHARACTERISTICS

Input characteristics

$$I_B = f(U_{BE})\big|_{U_{CE}=const.}$$



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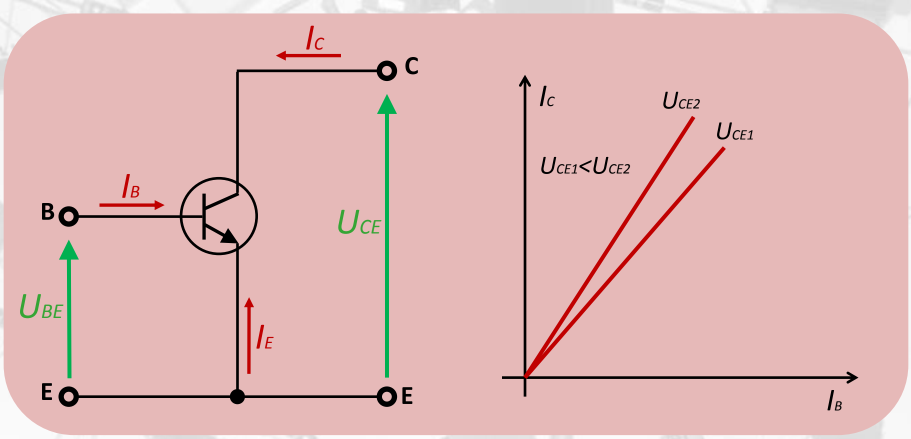
BIPOLAR JUNCTION TRANSISTOR
OE MODE - CHARACTERISTICS

Transitional characteristics

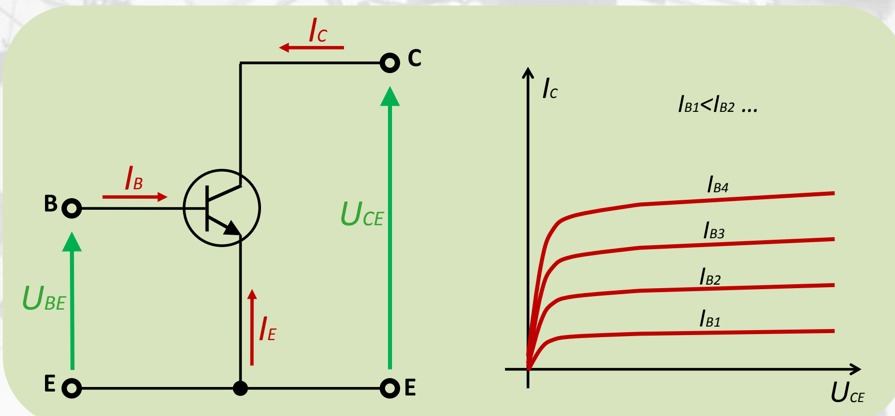
$$I_C = f(I_B)\big|_{U_{CE}=const.}$$



BIPOLAR JUNCTION TRANSISTOR
OE MODE - CHARACTERISTICS

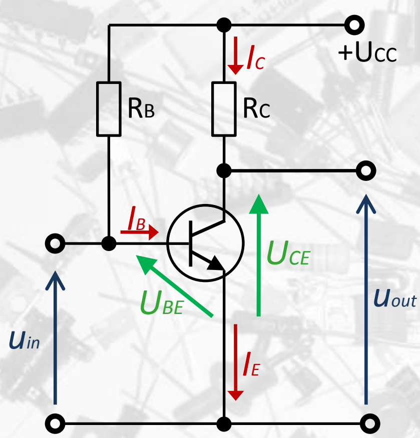
Output characteristics

$$I_C = f(U_{CE})\big|_{I_B=const.}$$



BIPOLAR JUNCTION TRANSISTOR
OE MODE - ANALYSIS

Determination of operation point Q



$$U_{CC} = I_B \cdot R_B + U_{BE}$$

$$I_B = \frac{U_{CC} - U_{BE}}{R_B}$$

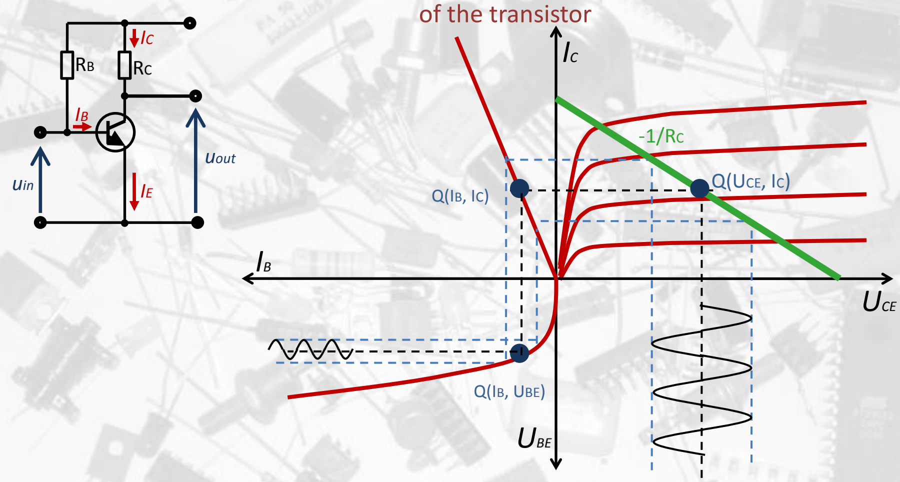
$$I_C = \beta \cdot I_B$$

$$U_{CC} = I_C \cdot R_C + U_{CE}$$

$$U_{CE} = U_{CC} - I_C \cdot R_C$$

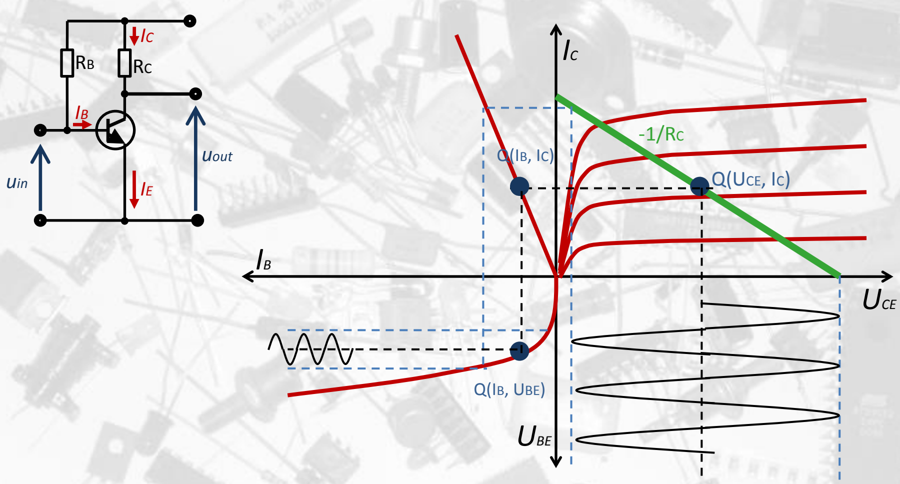
BIPOLAR JUNCTION TRANSISTOR
OE MODE - ANALYSIS

The impact of the choice of operation point on the amplifying properties of the transistor



BIPOLAR JUNCTION TRANSISTOR
OE MODE - ANALYSIS

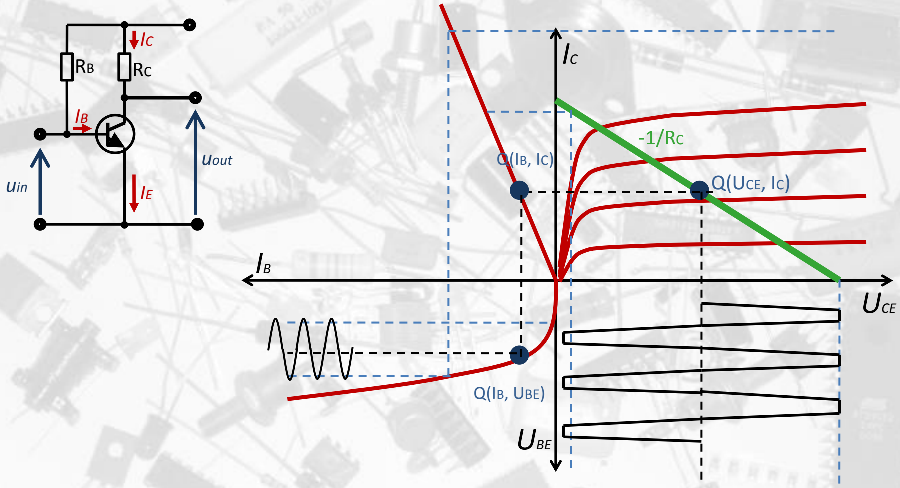
The operating point for maximum dynamics of the output voltage



BIPOLAR JUNCTION TRANSISTOR

OE MODE - ANALYSIS

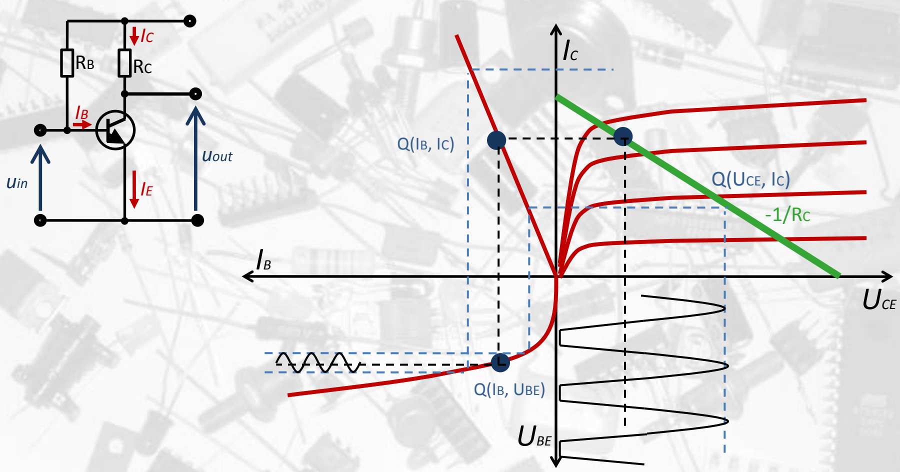
Output distortions due to too high input amplitude



BIPOLAR JUNCTION TRANSISTOR

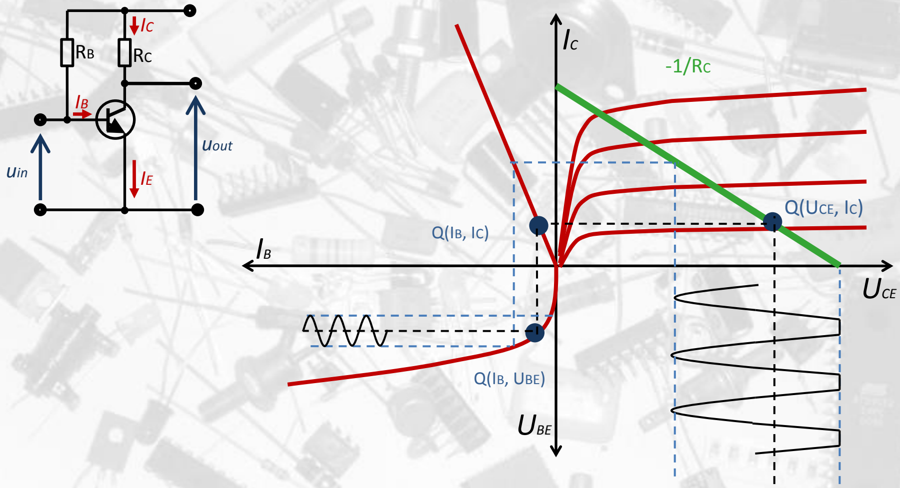
OE MODE - ANALYSIS

Operation point too close to the „saturation" area



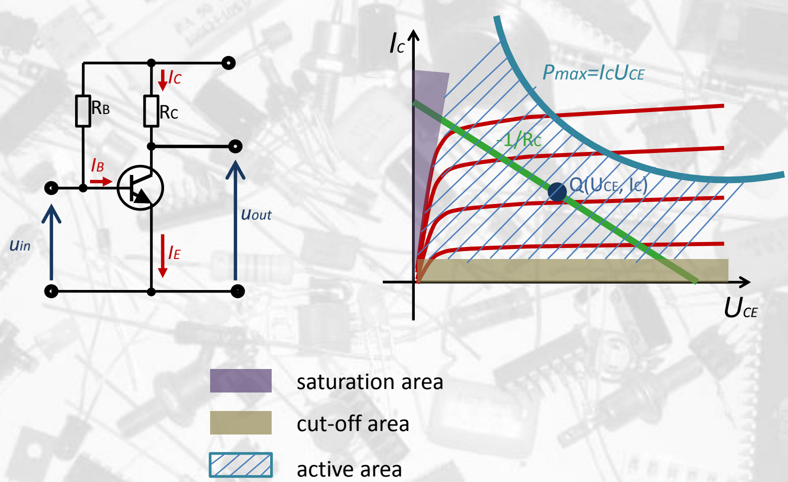
BIPOLAR JUNCTION TRANSISTOR
OE MODE - ANALYSIS

Operation point too close to the „cut-off" area



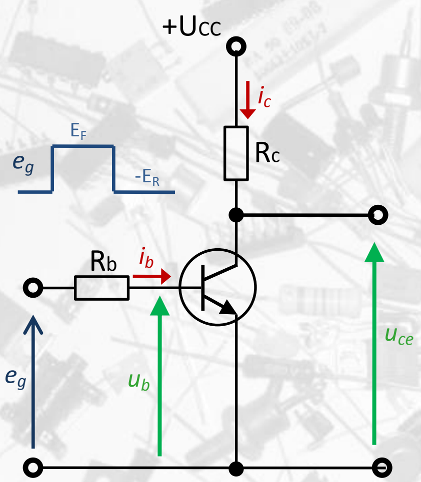
BIPOLAR JUNCTION TRANSISTOR
OE MODE - ANALYSIS

Operation area of the transistor at the output characteristics



SWITCHING BIPOLAR JUNCTION TRANSISTOR

SWITCHING BIPOLAR JUNCTION TRANSISTOR



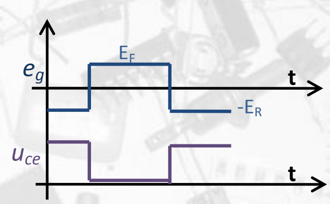
$$E_F = I_B \cdot R_b + U_{BEP}$$

$$I_B = \frac{E_F - U_{BEP}}{R_b}$$

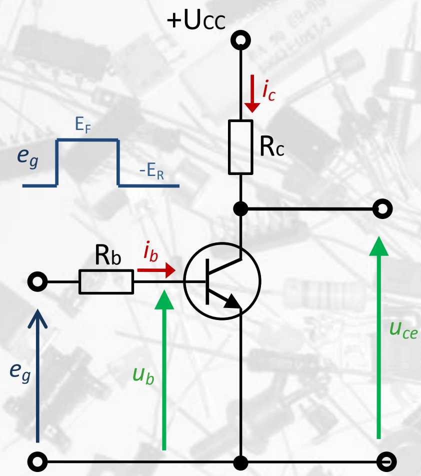
$$I_C = \beta \cdot I_B$$

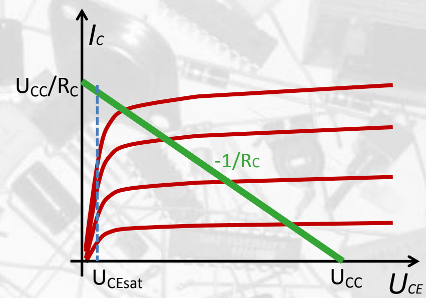
$$U_{CC} = I_C \cdot R_c + U_{CE}$$

$$U_{CE} = U_{CC} - I_C \cdot R_c$$



SWITCHING BIPOLAR JUNCTION TRANSISTOR





$$I_{BF}=\frac{E_F-U_{BEP}}{R_b}$$

$$I_{CM}=\frac{U_{CC}-U_{CEsat}}{R_c}\approx\frac{U_{CC}}{R_c}$$

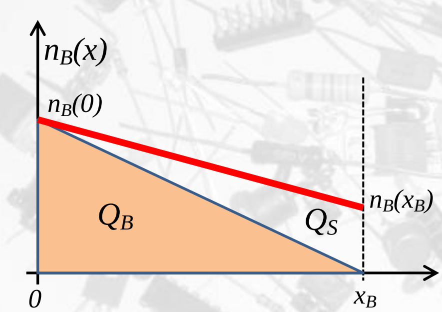
SWITCHING BIPOLAR JUNCTION TRANSISTOR

$$U_{CEsat}=U_T\ln\left[\frac{\frac{1}{\alpha_r}+\frac{I_C}{I_B}\left(\frac{1-\alpha_R}{\alpha_R}\right)}{1-\frac{I_C}{I_B}\left(\frac{1-\alpha_F}{\alpha_F}\right)}\right]=U_T\ln\left(\frac{1}{\alpha_R}\right)$$

$$I_C=\frac{Q_B}{\tau_t}$$

$$Q_B=\tau_{BF}I_B$$

$$\frac{\tau_t}{\tau_{BF}}=\frac{I_C}{I_B}=\beta$$



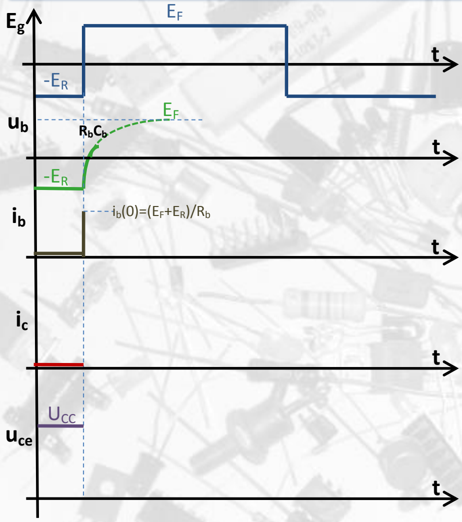
$$i_B(t)=\frac{Q_B}{\tau_{BF}}+\frac{dQ_B}{dt}$$

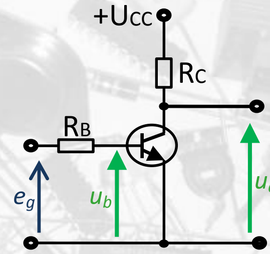
control equation of base charge in active normal mode

$$i_B(t)=\frac{Q_B}{\tau_{BF}}+\frac{Q_S}{\tau_S}+\frac{dQ_B}{dt}+\frac{dQ_S}{dt}$$

control equation of base charge in saturation mode

SWITCHING BIPOLAR JUNCTION TRANSISTOR





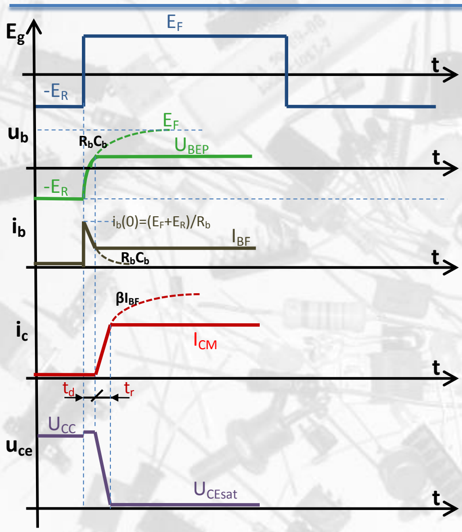
$$K_F = \frac{I_{BF}}{I_{BG}} = \frac{I_{BF}}{I_{CM}}\beta \geq 1$$

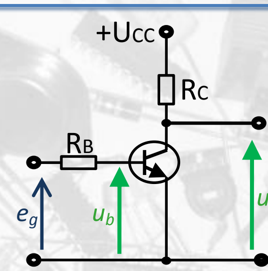
$$C_b = C_{jc} \parallel C_{je}$$

$$I_{BG} = \frac{I_{CM}}{\beta} \qquad i_b(0) = \frac{E_F + E_R}{R_b}$$

from: „Układy elektroniczne cz. II Układy analogowe nieliniowe i impulsowe", J. Baranowski, G. Czajkowski

SWITCHING BIPOLAR JUNCTION TRANSISTOR





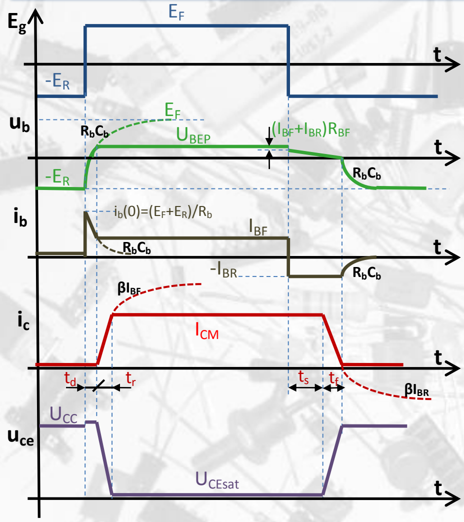
delay time

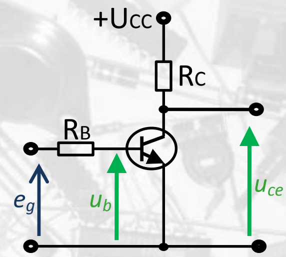
$$t_d = R_b C_b \ln\frac{E_F + E_R}{E_F - U_{BEP}}$$

rising time

$$t_r = \beta\left(t_t + R_c C_{jc}\right)\ln\frac{K_F}{K_F - 1}$$

SWITCHING BIPOLAR JUNCTION TRANSISTOR

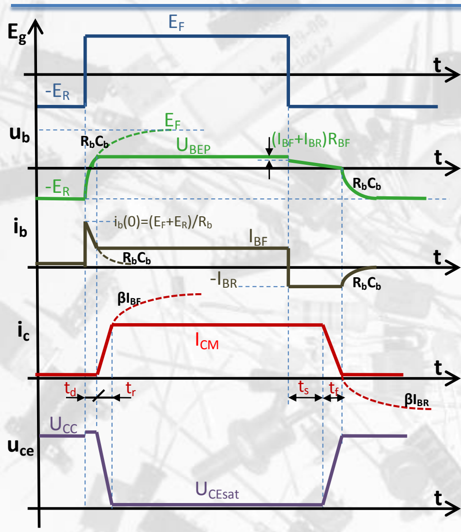


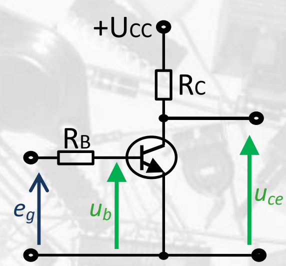


$$t_s = \tau_S \ln \frac{I_{BF} - I_{BR}}{\frac{I_{CM}}{\beta} - I_{BR}}$$

storage time

SWITCHING BIPOLAR JUNCTION TRANSISTOR





$$t_f = \beta\left(t_t + R_c C_{jc}\right)\ln \frac{I_{BR} + I_{BG}}{I_{BR}}$$

falling time

t_t – transfer time (through the base)

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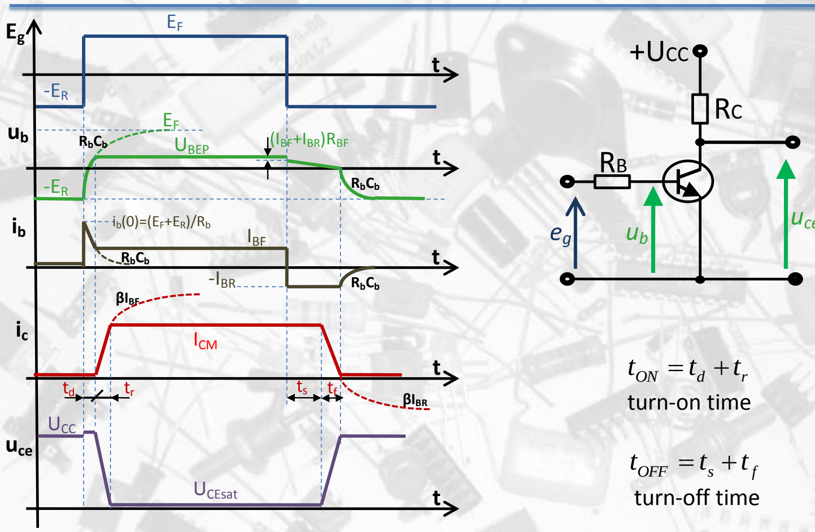
SWITCHING BIPOLAR JUNCTION TRANSISTOR



$t_{ON} = t_d + t_r$
turn-on time

$t_{OFF} = t_s + t_f$
turn-off time

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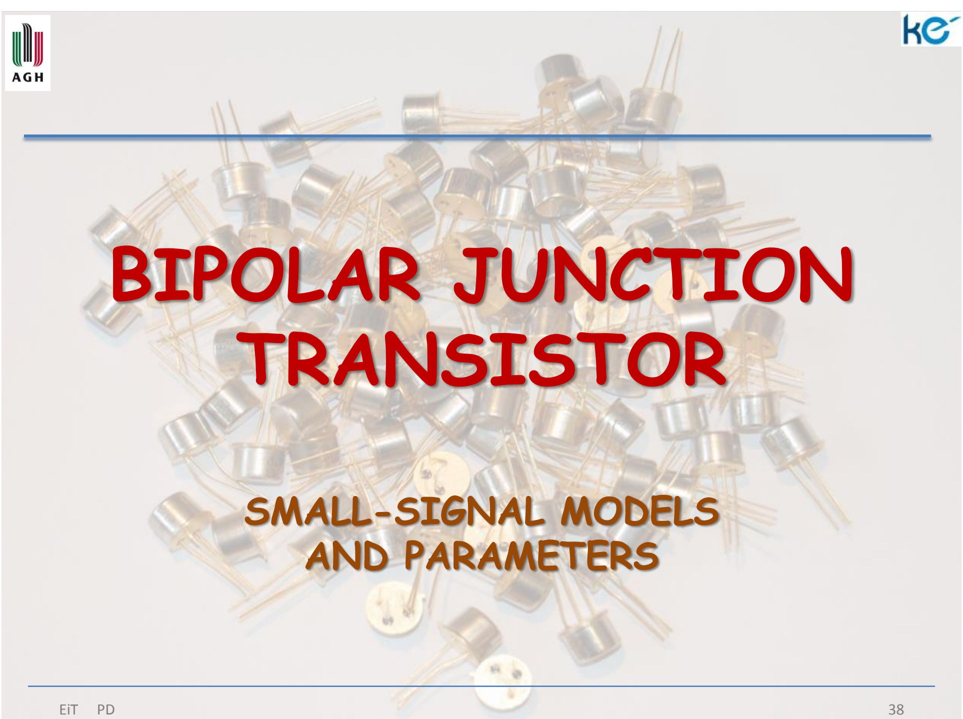


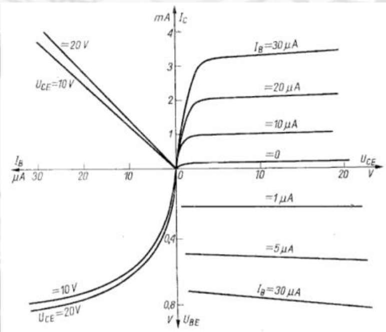
BIPOLAR JUNCTION TRANSISTOR

SMALL-SIGNAL MODELS AND PARAMETERS

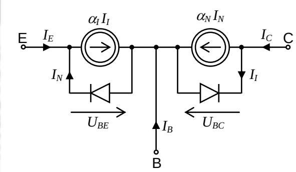
SMALL-SIGNAL MODEL – – OBJECTIVES

Transistor – a non linear device

Non linear characteristics → *Non linear model*



The characteristics of a bipolar transistor working in the common emitter configuration



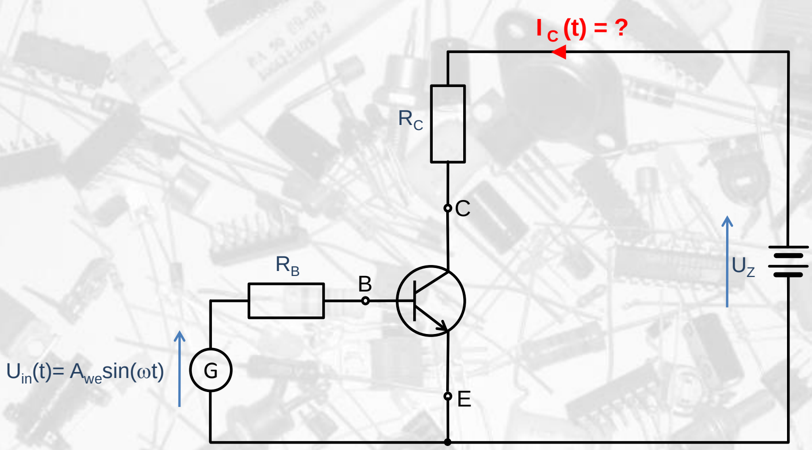
Ebers-Moll Model
of a bipolar npn transistor

$$i_E = -I_{ES}(e^{\frac{u_{BE}}{nU_T}}-1)+\alpha_I I_{CS}(e^{\frac{u_{BC}}{mU_T}}-1)$$

$$i_C = \alpha_N I_{ES}(e^{\frac{u_{BE}}{nU_T}}-1)-I_{CS}(e^{\frac{u_{BC}}{mU_T}}-1)$$

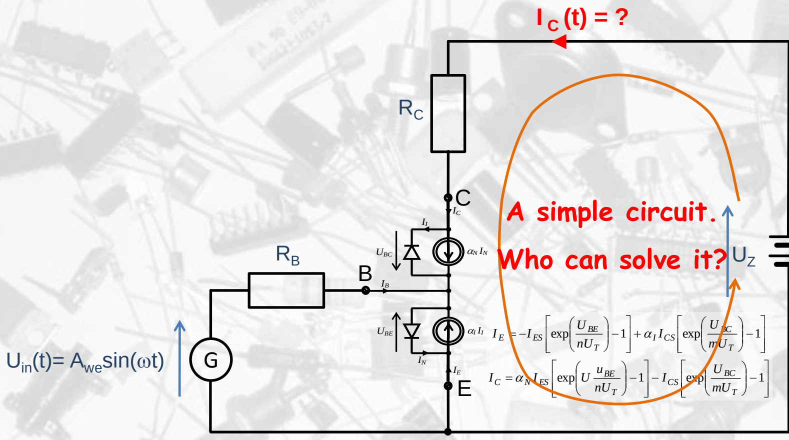
Picture from: W. Marciniak „Przyrządy półprzewodnikowe i układy scalone", WNT 1979

SMALL-SIGNAL MODEL – OBJECTIVES
transistor in a circuit



Non linear Model (eg.: Ebers-Moll) is inconvenient for the analysis of the transistor in larger electronic circuits

SMALL-SIGNAL MODEL – OBJECTIVES
transistor in a circuit

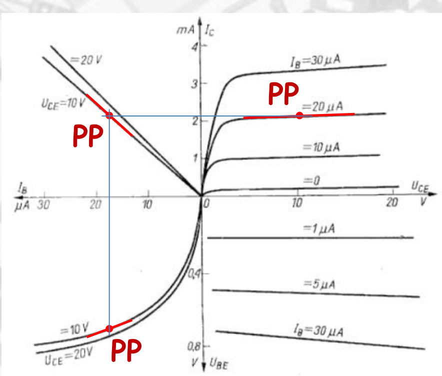


Non linear Model (eg.: Ebers-Moll) is inconvenient for the analysis of the transistor in larger electronic circuits

SMALL-SIGNAL MODEL – how to …?

Transistor – a non linear device → Linear model

How to do it ?



Characteristics of a bipolar transistor working in a the common emitter mode

Around operation point **PP**

↓

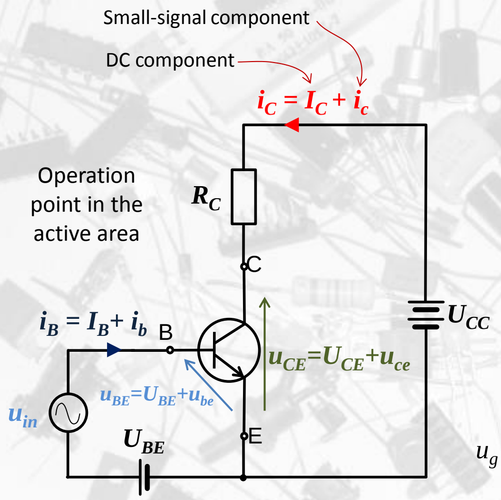
Linearization of characteristics

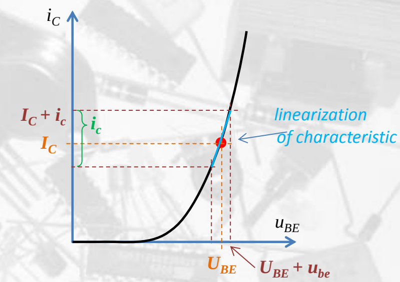
↓

Model built of linear elements (with some minor limitations)

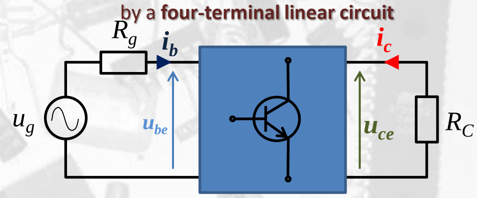
Picture from: W. Marciniak „Przyrządy półprzewodnikowe i układy scalone", WNT 1979

SMALL-SIGNAL MODEL
transistor as an active four-terminal unit



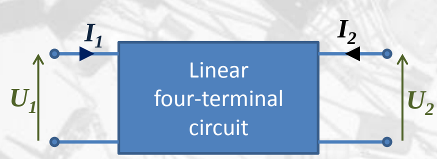


For small alternating signals (low amplitude). transistor can be replaced by a **four-terminal linear circuit**



LINEAR FOUR-TERMINAL MODELS
(repetition – circuit theory)

In a general case:



Impedance equations:

$$U_1 = Z_{11} I_1 + Z_{12} I_2$$
$$U_2 = Z_{21} I_1 + Z_{22} I_2$$

Admittance equations:

$$I_1 = Y_{11} U_1 + Y_{12} U_2$$
$$I_2 = Y_{21} U_1 + Y_{22} U_2$$

Mixed equations (hybrid):

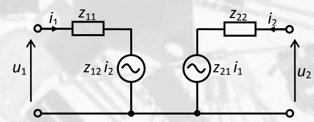
$$U_1 = H_{11} I_1 + H_{12} U_2$$
$$I_2 = H_{21} I_1 + H_{22} U_2$$

FOUR-TERMINAL MODELS (small-signal models)

Small signals – we denote as: small letters small indexes

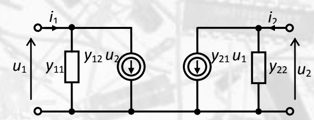
Impedance equations:

$$u_1 = z_{11}\, i_1 + z_{12}\, i_2$$
$$u_2 = z_{21}\, i_1 + z_{22}\, i_2$$



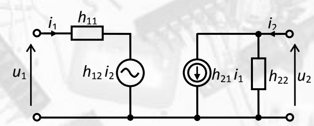
Admittance equations:

$$i_1 = y_{11}\, u_1 + y_{12}\, u_2$$
$$i_2 = y_{21}\, u_1 + y_{22}\, u_2$$

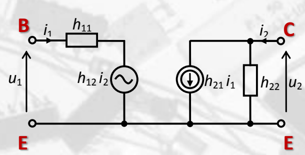


Hybrid equations:

$$u_1 = h_{11}\, i_1 + h_{12}\, u_2$$
$$i_2 = h_{21}\, i_1 + h_{22}\, u_2$$



HYBRID MODEL parameters for OE



- Input impedance
  at shorted output (for small-signal voltage component at the output)

$$h_{11} = \left.\frac{u_1}{i_1}\right|_{u_2=0} = \left.\frac{\Delta u_{BE}}{\Delta i_B}\right|_{U_{CE}=const} = \left.\frac{u_{be}}{i_b}\right|_{u_{ce}=0} = h_{11e}$$

- Reverse voltage transfer function
  at open input (open small-signal current source at input)

$$h_{12} = \left.\frac{u_1}{u_2}\right|_{i_2=0} = \left.\frac{\Delta u_{BE}}{\Delta u_{CE}}\right|_{I_B=const} = \left.\frac{u_{be}}{u_{ce}}\right|_{i_b=0} = h_{12e}$$

- Current transmittance - current gain
  at shorted output (for small-signal voltage component at the output)

$$h_{21} = \left.\frac{i_2}{i_1}\right|_{u_2=0} = \left.\frac{\Delta i_C}{\Delta i_B}\right|_{U_{CE}=const} = \left.\frac{i_c}{i_b}\right|_{u_{ce}=0} = h_{21e}$$

- Output admittance
  at open input (open small-signal current source at input)

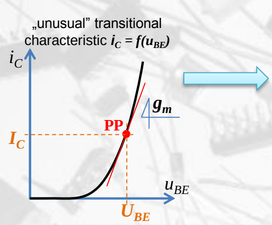
$$h_{22} = \left.\frac{i_2}{u_2}\right|_{i_1=0} = \left.\frac{\Delta i_C}{\Delta u_{CE}}\right|_{I_B=const} = \left.\frac{i_c}{u_{ce}}\right|_{i_b=0} = h_{22e}$$

HYBRID MODEL
parameters for different configurations

	OE	OB	OC
h_{11}	h_{11e}	$h_{11b} = \dfrac{h_{11e}}{1+h_{21e}}$	$h_{11c} = h_{11e}$
h_{12}	h_{12e}	$h_{12b} = \dfrac{h_{11e}h_{22e}}{1+h_{21e}} - h_{12e}$	$h_{12c} = 1 - h_{12c}$
h_{21}	h_{21e}	$h_{21b} = \dfrac{h_{21e}}{1+h_{21e}}$	$h_{21c} = 1 + h_{21c}$
h_{22}	h_{22e}	$h_{22b} = \dfrac{h_{22e}}{1+h_{21e}}$	$h_{22c} = h_{22c}$

SMALL-SIGNAL MODEL of BJT

Representation of physical phenomena occurring in the transistor – equivalent circuit



- Transconductance – the impact of input to output

$$g_m \equiv \left.\frac{\partial i_C}{\partial u_{BE}}\right|_{U_{BE},U_{CE}=const}$$ transitional characteristic

- Feedback transconductance – the impact of the output voltage to the input

$$g_r \equiv \left.\frac{\partial i_B}{\partial u_{CE}}\right|_{U_{BE},U_{CE}=const}$$ feedback characteristic

- Input conductance – input characteristic (transistor seen „at the input side")

$$g_\pi \equiv \left.\frac{\partial i_B}{\partial u_{BE}}\right|_{U_{BE},U_{CE}=const}$$ input characteristic

- Output conductance – output characteristic (transistor seen „at the output side")

$$g_o \equiv \left.\frac{\partial i_C}{\partial u_{CE}}\right|_{U_{BE},U_{CE}=const}$$ output characteristic

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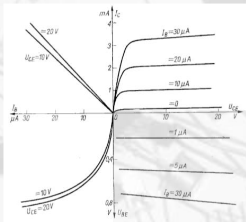
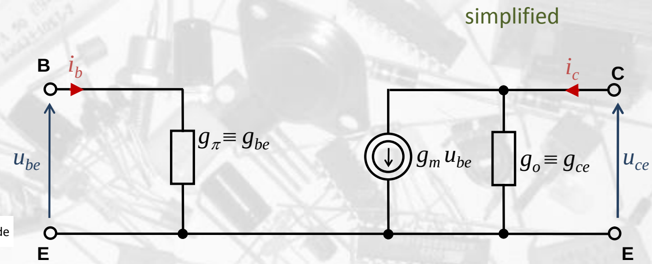
HYBRID-π MODEL for OE MODE



Characteristics of bipolar transistor in the OE mode



- at the output side: current source controlled from the input: $g_m u_{be}$
- at the input side: input conductance: $g_\pi \equiv g_{be}$
- at the output side : output conductance: $g_o \equiv g_{ce}$`

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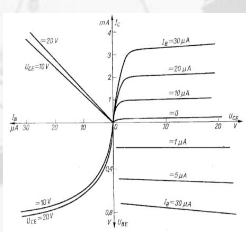
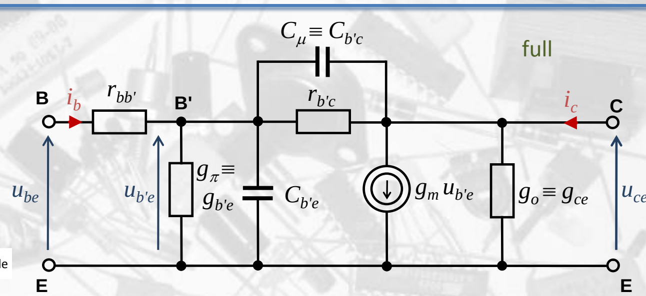
HYBRID-π MODEL for OE MODE



Characteristics of bipolar transistor in the OE mode



- at the output side: current source controlled from the input: $g_m u_{be}$
- at the input side: input conductance: $g_\pi \equiv g_{be}$
- at the output side : output conductance: $g_o \equiv g_{ce}$`
- at the input side : base resistance: $r_{bb'}$
- from the input to the output a direct: resistive feedback base-collector: $r_{b'c}$
- capacitance of emitter junction $C_{b'e}$ and capacitance of collector junction $C_{b'c}$

HYBRID-π MODEL for OE MODE
estimation of parameters (1)

Transconductance g_m

- From a definition: $g_m = \frac{\partial I_C}{\partial U_{BE}}$ the slope of „unusual" transitional characteristic $I_C = f(U_{BE})$
- From a operation point:
  differentiating emitter diode current from Ebers-Moll model:

$$g_m = \frac{\partial(\alpha I_E)}{\partial U_{BE}} = \alpha \frac{I_E}{n_E U_T}$$

simplified relationship

$$I_E = I_{ES}(e^{\frac{U_{BE}}{n_E U_T}} - 1)$$

and taking into account collector current ($I_C = \alpha I_E$):

$$g_m = \frac{I_C}{n_E U_T}$$

In practice $n_E = 1$

$$g_m = \frac{I_C}{U_T}$$

α – coefficient of current gain for OB mode
n_E – coefficient of non ideality of emitter junction
U_T – electrothermal potential
I_C – constant collector current

HYBRID-π MODEL for OE MODE
estimation of parameters (2)

Input conductance $g_{b'e}$

- From a definition: $g_{b'e} = \frac{\partial I_B}{\partial U_{BE}}$ slope of the input characteristic $I_B = f(U_{BE})$
- From the operation point: for OE mode, it is:

$$I_E = I_C + I_B$$
$$I_B = I_E - I_C, \quad I_C = \alpha I_E$$
$$I_B = I_E(1-\alpha)$$

therefore from Ebers-Moll model, base current for OE is: $I_B \approx (1-\alpha) I_{ES} \exp\left(\frac{U_{BE}}{n_E U_T}\right)$

Taking into account the definition: $g_{b'e} = \frac{1}{n_E U_T}(1-\alpha) I_{ES} \exp\left(\frac{U_{BE}}{n_E U_T}\right)$

$$g_{b'e} = \frac{I_B}{n_E U_T}$$

and: $I_B = \frac{I_C}{\beta_0}$ we have: $g_{b'e} = \frac{I_C}{\beta_0 n_E U_T} = \frac{g_m}{\beta_0}$

α – coefficient of current gain for OB
β_0 – coefficient of current gain for OE
n_E – coefficient of non ideality of emitter junction
U_T – electrothermal potential
I_C – constant collector current

HYBRID-π MODEL for OE MODE
estimation of parameters (3)

Distributed resistance of the base $r_{bb'}$

- From the comparison of hybrid-π and hybrid models :

$$r_{bb'} = h_{11e} - r_{b'e}$$

Output conductance g_{ce}

- From a definition: $g_{ce} = \frac{\partial I_C}{\partial U_{CE}}$ $\quad g_{ce} \equiv h_{22e}$
- Taking into account Early effect: $I_C = \beta_0 I_B \left(1 + \frac{U_{CE}}{U_A}\right)$

U_A – Early voltage

- And after differentiation: $g_{ce} = \beta_0 I_B \frac{1}{U_A} = \frac{I_C}{U_A + U_{CE}}$

Resistive feedback $r_{b'c}$

- From a definition: $r_{b'c} = \frac{\partial U_{CB}}{\partial I_B}$
- but $U_{CB} >> U_{BE}$, then: $r_{b'c} \approx \frac{\partial U_{CE}}{\frac{\partial I_C}{\beta_0}} = \frac{\beta_0}{g_{ce}} = \beta_0 \frac{U_A + U_{CE}}{I_C}$ but if: $U_A >> U_{CE}$, then:

$$r_{b'c} \approx \beta_0 \frac{U_A}{g_m U_T}$$

HYBRID-π MODEL for OE MODE
estimation of parameters (4)

Input capacitance C_π – emitter junction $C_{b'e}$

$$C_\pi \equiv C_{b'e} = C_{de} + C_{je}$$

junction capacitance

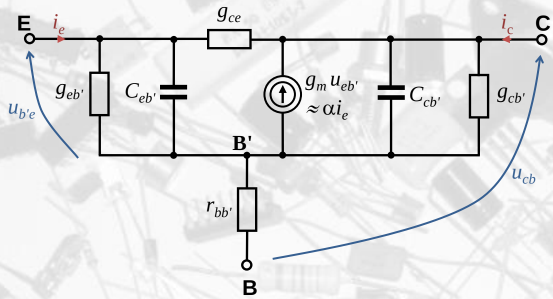
diffusion capacitance

$$C_{b'e} = C_{de} = \tau_F \frac{I_E}{U_T} = \tau_F g_{b'e} \frac{1}{1-\alpha}$$

Coupled capacitance C_μ – collector junction $C_{b'c}$

Junction capacitance of reverse biased base-collector junction

HYBRID-π MODEL for OB MODE



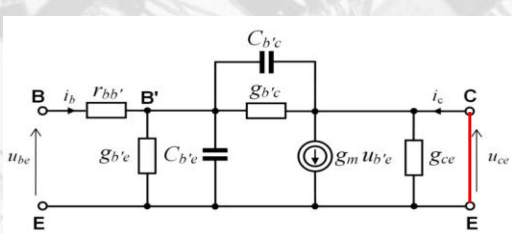
$$\alpha \equiv \left.\frac{\partial i_C}{\partial i_E}\right|_{U_{BC}=const}$$

$$g_{eb'} \equiv \left.\frac{\partial i_E}{\partial u_{BE}}\right|_{U_{BE},U_{BC}=const} \approx \frac{I_E}{U_T}$$

$$g_{b'e} = \frac{g_{eb'}}{\beta_0 + 1} \qquad g_{b'e} \approx g_m$$

FREQUENCY LIMITS

When, for which frequency does transistor stop to fulfil its basic function, that is amplification of signals?



applicability : $\tau_F << T$

small-signal current gain β for shorted output:

$$\beta(j\omega) \equiv \left.\frac{i_c}{i_b}\right|_{u_{ce}=0}$$

$$\beta(j\omega) = \frac{g_m u_{b'e}(j\omega)}{i_b(j\omega)}$$

$$\beta(j\omega) = \frac{\dfrac{g_m}{g_{b'e}}}{1 + j\omega\left(\dfrac{C_{b'e} + C_{b'c}}{g_{b'e}}\right)}$$

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FREQUENCY LIMITS

- Cut-off frequency f_β
  - when $\beta(f)$ is reduced by 3dB:

$$|\beta(f_\beta)| \equiv \frac{\beta_0}{\sqrt{2}}$$

$$\beta(j\omega) = \frac{\dfrac{g_m}{g_{b'e}}}{1 + j\omega\left(\dfrac{C_{b'e} + C_{b'c}}{g_{b'e}}\right)}$$

Denoting : $G = \dfrac{g_m}{g_{b'e}}, \quad X = \dfrac{C_{b'e} + C_{b'c}}{g_{b'e}}$

$$\beta(j\omega) = \frac{G}{1 + j\omega X} = \frac{G}{1 + \omega^2 X^2} + j\frac{\omega G X}{1 + \omega^2 X^2}$$

magnitude

$$|\beta(j\omega)| = \sqrt{\left(\frac{G}{1 + \omega^2 X^2}\right)^2 + \left(\frac{\omega G X}{1 + \omega^2 X^2}\right)^2} \qquad |\beta(f_\beta)| \equiv \frac{\beta_0}{\sqrt{2}}$$

$$g_{b'e} = \frac{g_m}{\beta_0}$$

$$\sqrt{\left(\frac{G}{1 + \omega_\beta^2 X^2}\right)^2 + \left(\frac{\omega_\beta G X}{1 + \omega_\beta^2 X^2}\right)^2} = \frac{\beta_0}{\sqrt{2}}$$

$$\left(\frac{G^2 + \omega_\beta^2 X^2 G^2}{1 + \omega_\beta^2 X^2}\right) = \frac{\beta_0}{\sqrt{2}}$$

$$1 + \omega_\beta^2 X^2 = 2\frac{G^2}{\beta_0^2} \qquad \omega_\beta = \frac{1}{X} \qquad \varpi_\beta = \frac{g_{b'e}}{C_{b'e} + C_{b'c}} \qquad f_\beta = \frac{g_{b'e}}{2\pi(C_{b'e} + C_{b'c})}$$

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FREQUENCY LIMITS

- Cut-off frequency f_α
  - when $\alpha(f)$ is reduced by 3dB:

$$|\alpha(f_\alpha)| \equiv \frac{\alpha_0}{\sqrt{2}}$$

Following the analogous procedure as for f_β:

$$f_\alpha = \frac{g_{eb'}}{2\pi C_{eb'}}$$

FREQUENCY LIMITS

- Threshold frequency f_T
  - when magnitude of $\beta(f)$ =1

$$\beta(j\omega)=\frac{\frac{g_m}{g_{b'e}}}{1+j\omega\left(\frac{C_{b'e}+C_{b'c}}{g_{b'e}}\right)}$$

$$f_\beta=\frac{g_{b'e}}{2\pi(C_{b'e}+C_{b'c})}$$

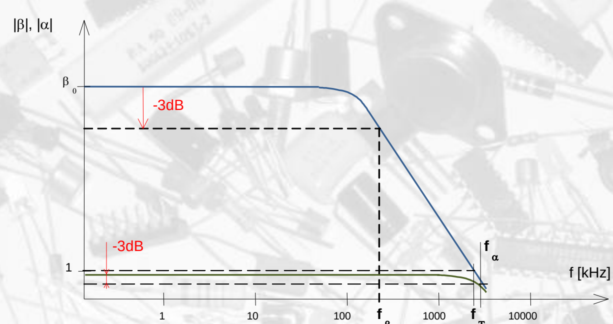
$$\beta(f)=\frac{\beta_0}{1+j\frac{f}{f_\beta}} \Rightarrow \frac{\beta(f)}{\beta_0}=\frac{1}{1+j\frac{f}{f_\beta}}\approx -j\frac{f_\beta}{f}$$ when: $f > f_\beta$

for: $f = f_T$ it is: $|\beta(f)|=1$

$$\frac{1}{\beta_0}=\frac{f_\beta}{f_T}$$

$$f_T=\beta_0 f_\beta$$

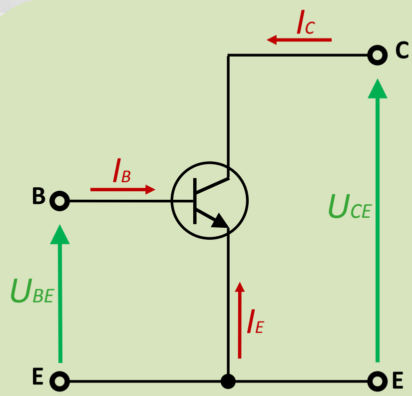
FREQUENCY LIMITS

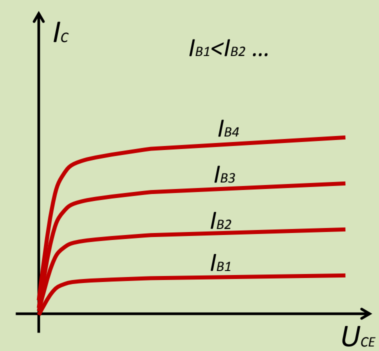


Bipolar transistor operating in a common base mode has β-times higher cut-off frequency

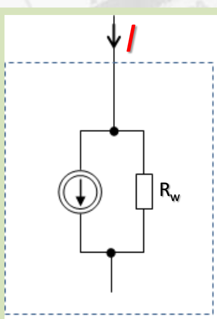
$$f_\alpha \approx \beta_0 f_\beta$$

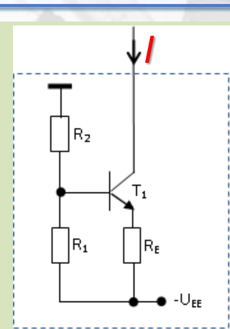
CURRENT SOURCES





CURRENT SOURCES





$$U_{R1} = U_{RE} + U_{BE}$$

$$U_{EE}\frac{R_1}{R_1 + R_2} = IR_E + U_{BE}$$

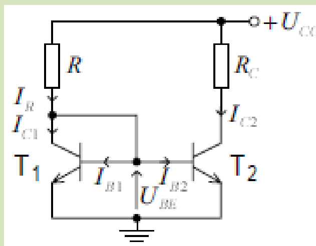
$$R_w = R_E II (r_{be} + R_B) + r_{ce}\left(1 + g_m \frac{R_E + r_{be}}{R_E + r_{be} + R_B}\right)$$

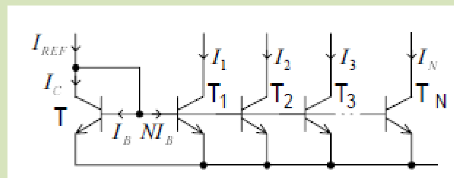
$R_B = R_1 II R_2$

$$I = \frac{U_{EE}\frac{R_1}{R_1 + R_2} - U_{BE}}{R_E}$$

CURRENT SOURCES

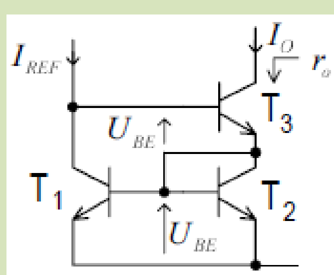
Current mirrors





CURRENT SOURCES

Wilson's current mirror



$$r_o \approx \beta_0 \frac{U_A}{2I_O}$$