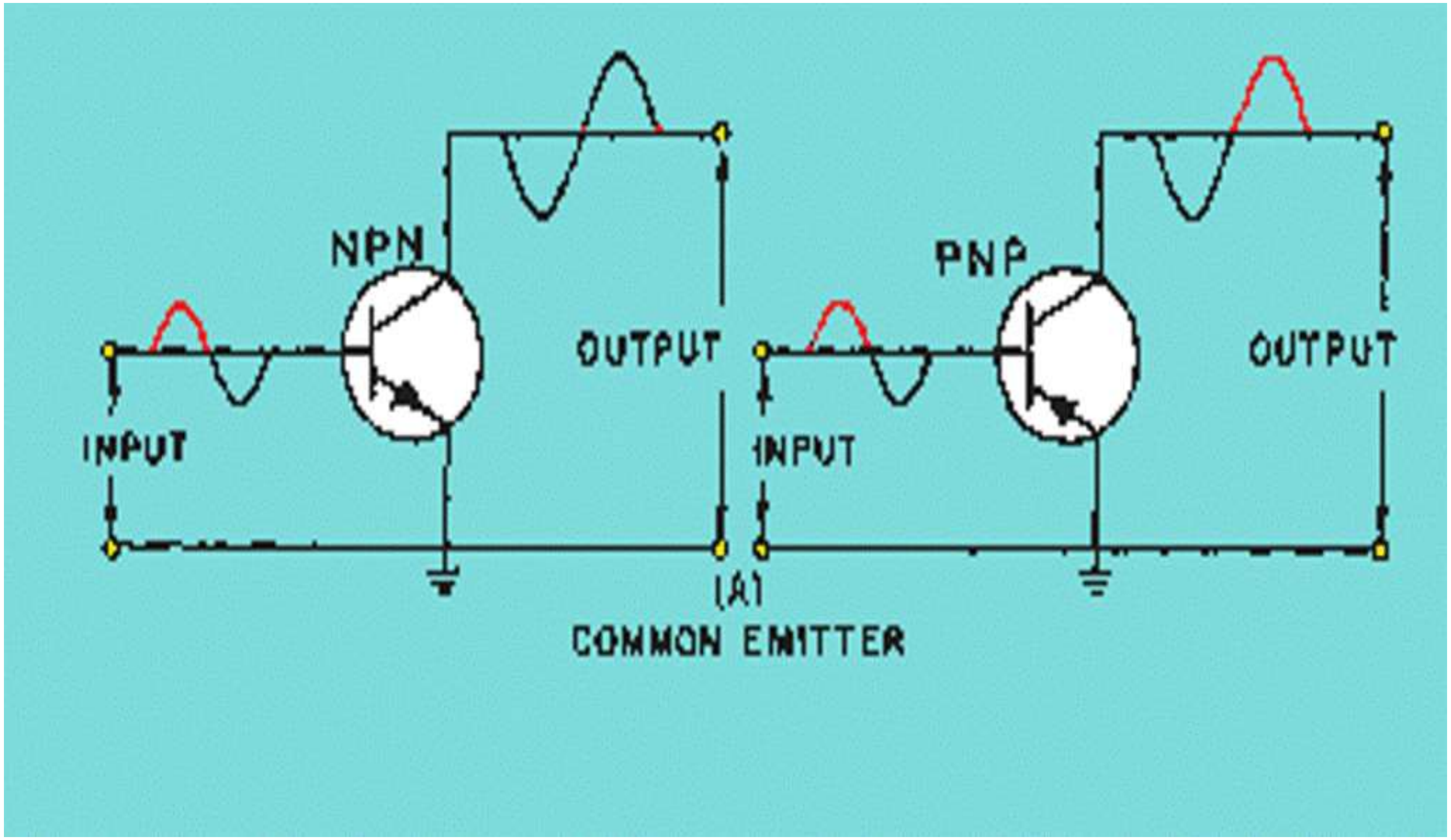
Unit-1
The transistor at low frequency

Graphical Analysis of the CE Configuration

CE configuration

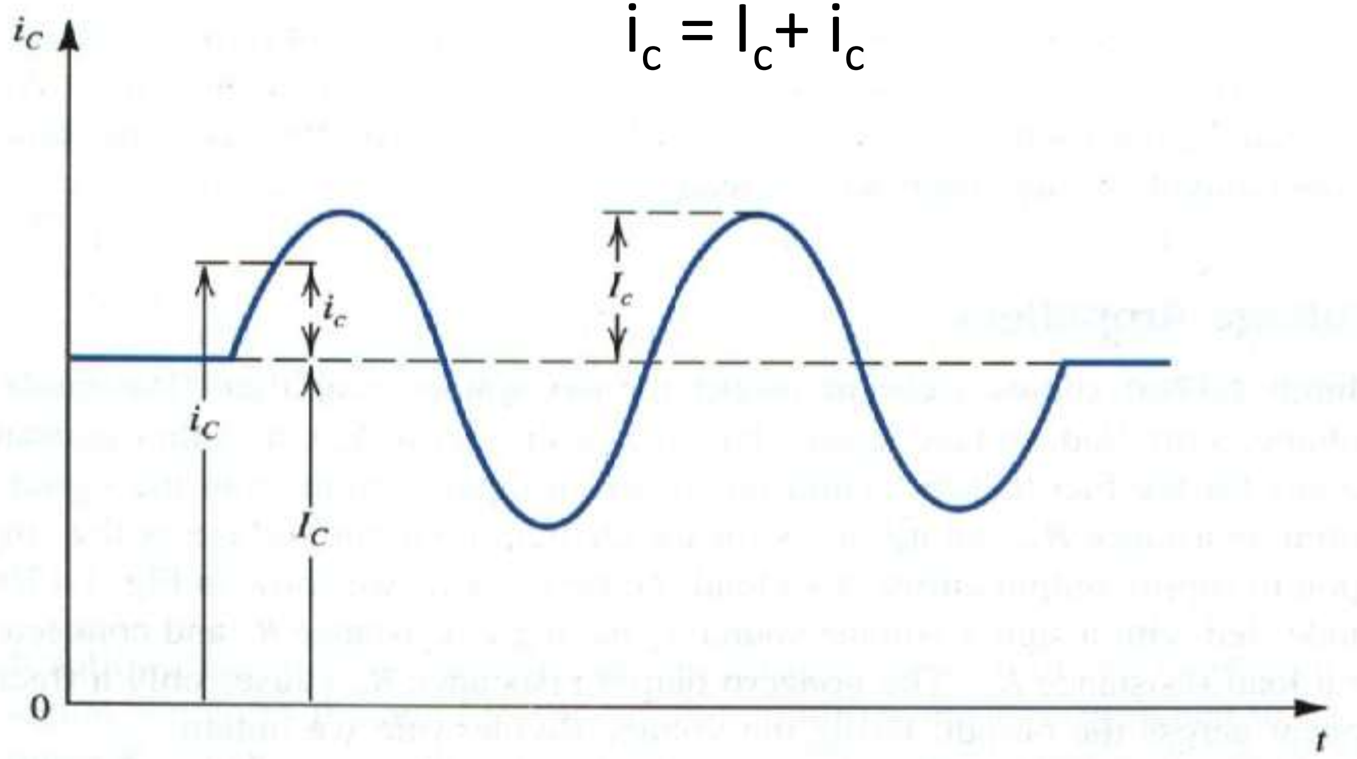


Notation

S. No.		Base (Collector) voltage with respect to emitter	Base (Collector) current toward electrode from external circuit
1	Instantaneous total value	v_B (v_C)	i_B (i_C)
2	Quiescent value	V_B (V_C)	I_B (I_C)
3	Instantaneous value of varying component	v_b (v_c)	i_b (i_c)
4	Effective value of varying component	V_b (V_c)	I_b (I_c)
5	Supply voltage	V_{BB} (V_{CC})	--------

Total signal = DC component (Quiescent value) + time varying signal

$i_c = I_c + i_c$



Distortions

- Out put nonlinear distortion
- Input nonlinear distortion

How to reduce the distortion?

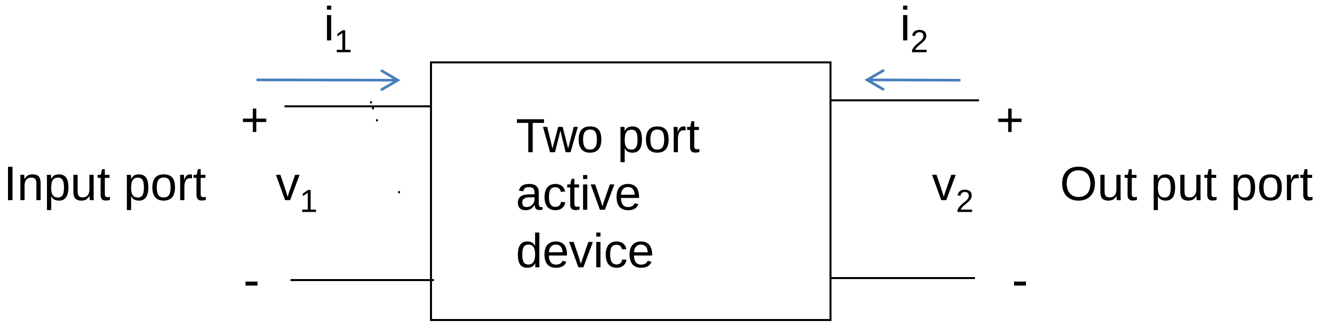
- If the amplifier is biased so that Q is near the center of the i_C-v_{CE} plane there will be less distortion if the excitation is a sinusoidal base voltage than if it is a sinusoidal base current.
- The dynamic load curve can be approximated by a straight line over a sufficiently small line segment, and hence, if the input signal is small, there will be negligible input distortion under any condition of operation(current-source or voltage-source driver).

TWO PORT DEVICE AND THE HYBRID MODEL

Two port network

$$v_1 = h_{11}\ i_1 + h_{12}\ v_2$$

$$i_2 = h_{21}\ i_1 + h_{22}\ v_2$$



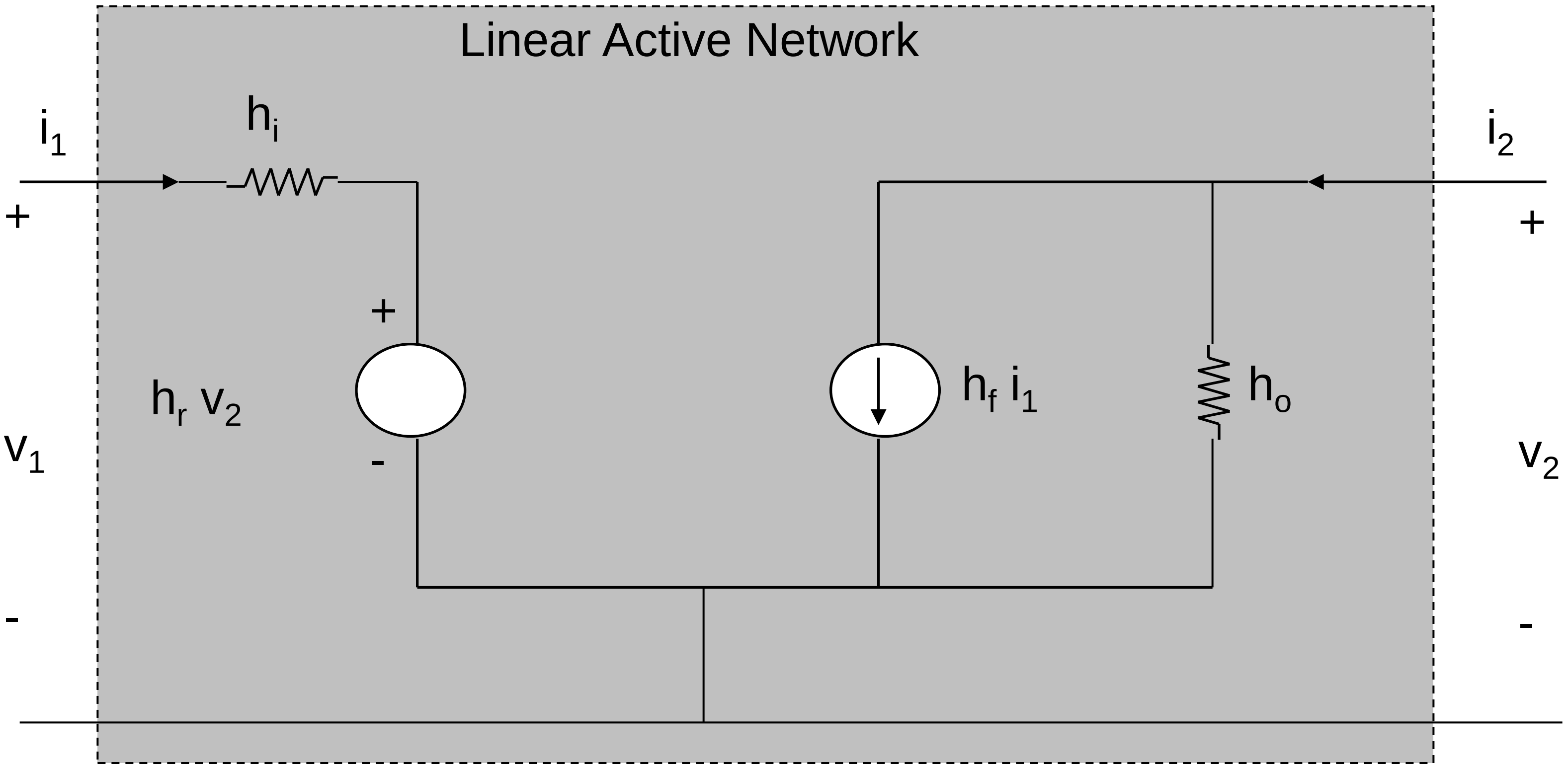
- h_{11}, h_{12}, h_{21} and h_{22} are called the h or hybrid parameter.

1. $h_i = h_{11} = v_1/i_1$ for $v_2 = 0$; input resistance with out put short-circuit(in ohms)
2. $h_r = h_{12} = v_1/v_2$ for $i_1 = 0$; reverse open circuit voltage amplification
3. $h_f = h_{21} = i_2/i_1$ for $v_2 = 0$; forward current transfer ratio or short-circuit current gain
4. $h_o = h_{22} = i_2/v_2$ for $i_1 = 0$; output conductance with input short circuit

h-Parameter Models (small signal model)

$$v_1 = h_i \; i_1 + h_r \; v_2$$

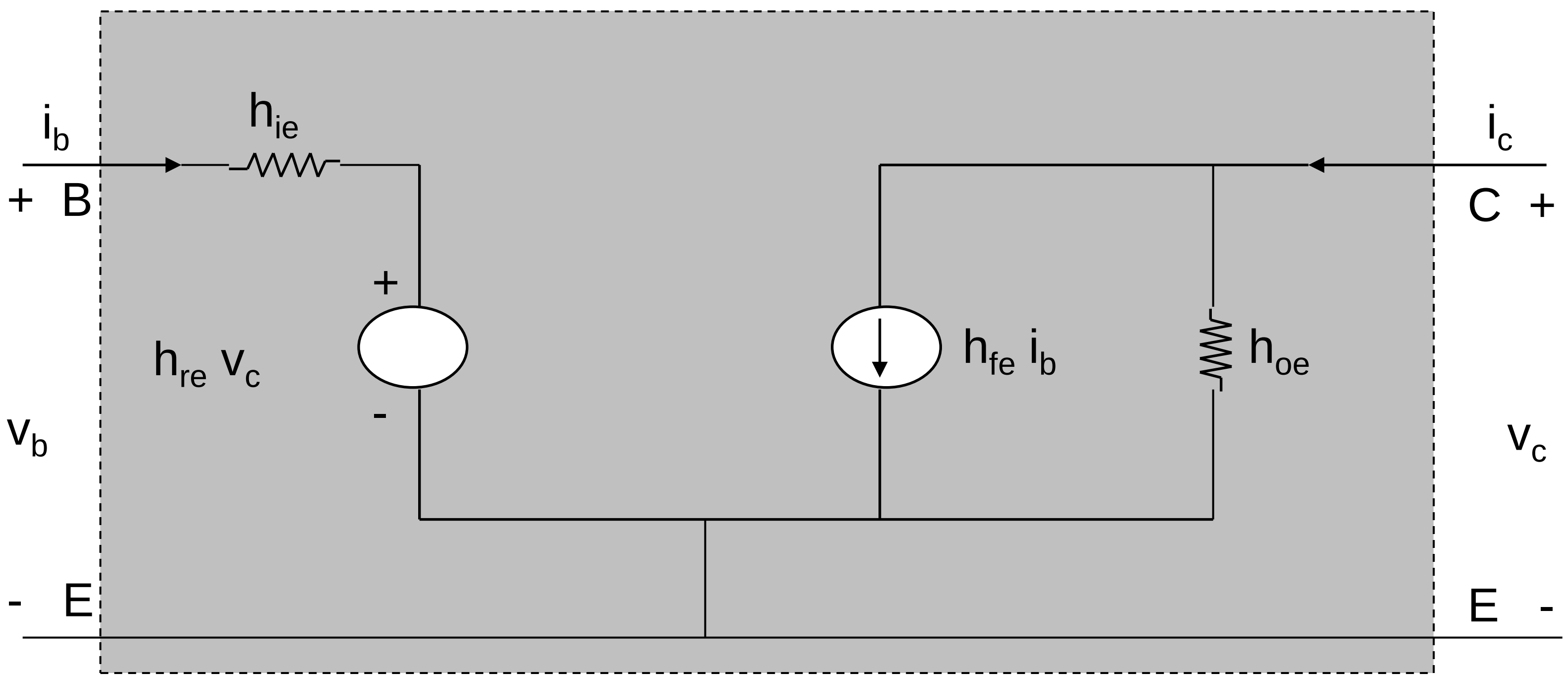
$$i_2 = h_f \; i_1 + h_o \; v_2$$



The hybrid small signal models for common-emitter configuration

$v_b = h_{ie} \; i_b + h_{re} \; v_c$

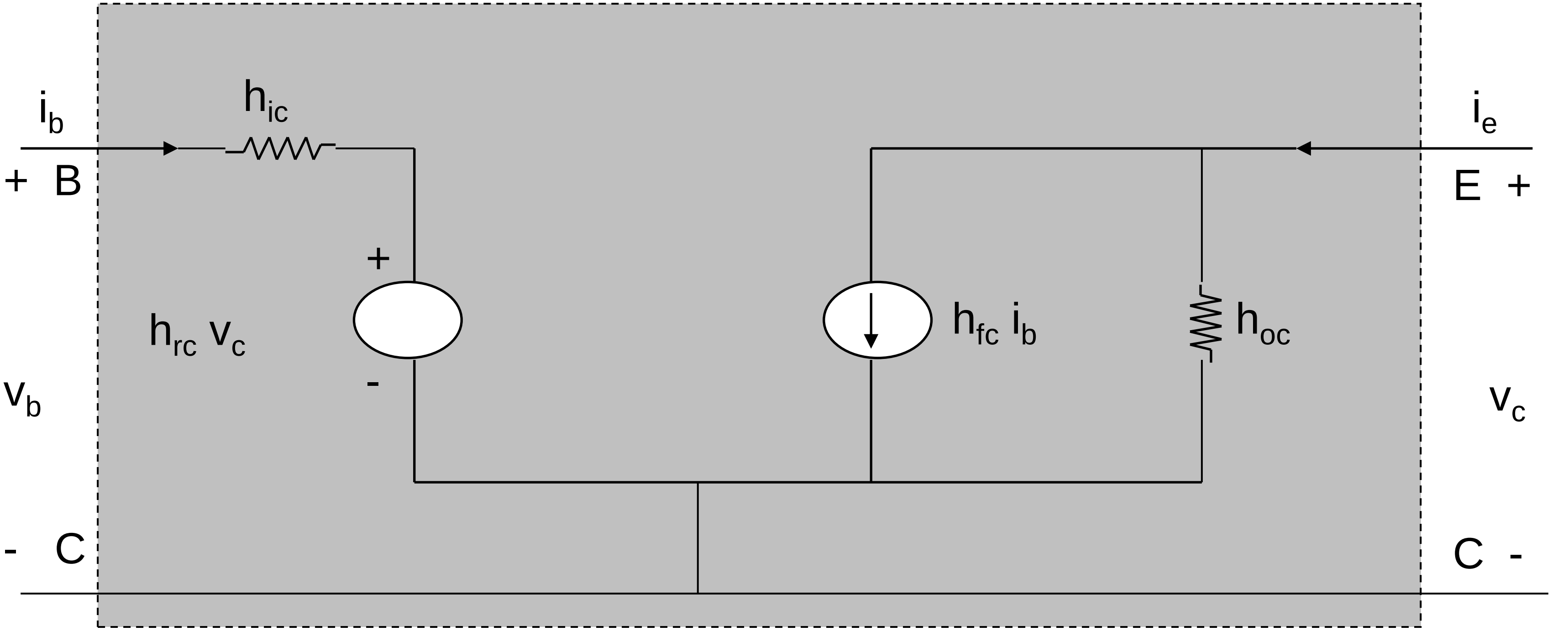
$i_c = h_{fe} \; i_b + h_{oe} \; v_c$



The hybrid small signal models for common-collector configuration

$$v_b = h_{ic}\, i_b + h_{rc}\, v_e$$

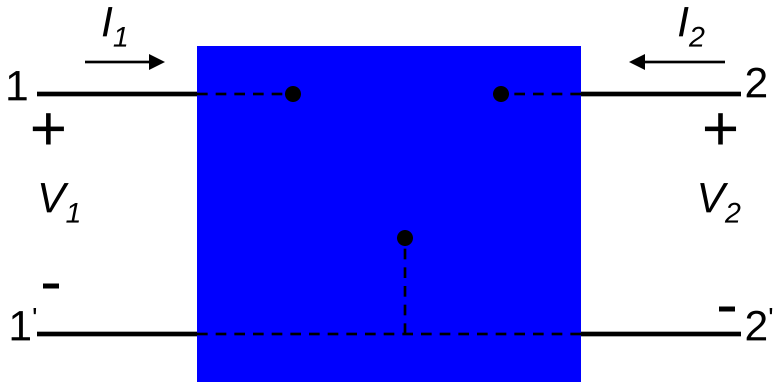
$$i_e = h_{fc}\, i_b + h_{oc}\, v_e$$



The hybrid model: *h*- parameter

Let us consider transistor as two port system having two voltages and two current values at the input and output ports. We may select two out of four as independent variables and express the remaining two in terms of the chosen independent variables.

If I_1 and V_2 are independent and if two port is linear:



$$V_1 = h_{11} I_1 + h_{12} V_2 \qquad .....1$$

$$I_2 = h_{21} I_1 + h_{22} V_2 \qquad .....2$$

$$h_{11} = \left. \frac{V_1}{I_1} \right|_{V_2=0}$$ (ohms) → input resistance with output short-circuited

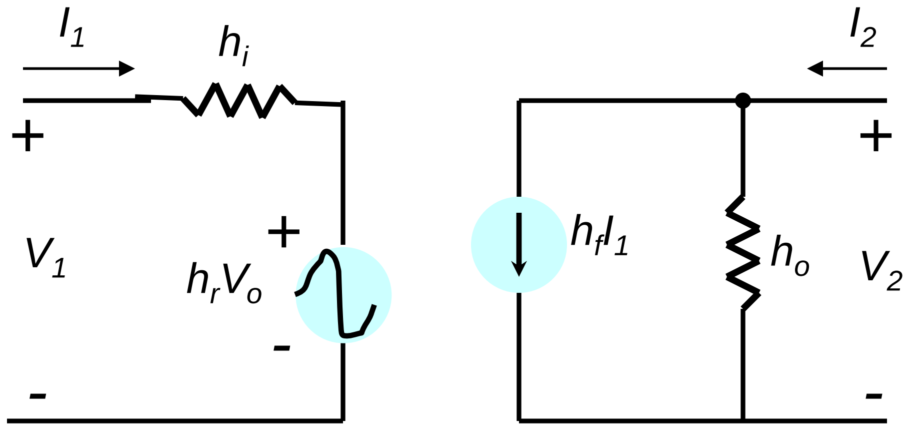
$$h_{12} = \left. \frac{V_1}{V_2} \right|_{I_1=0}$$ (unit less) → fraction of output voltage at input with input open-circuited

$$h_{21} = \left.\frac{I_2}{I_1}\right|_{V_2=0}$$ (unit less) → current transfer ratio with output short-circuited

$$h_{22} = \left.\frac{I_2}{V_2}\right|_{I_1=0}$$ (siemens) → output conductance with input open-circuited

The alternative subscript notation is : i = 11(input); o = 22(output); f = 21(forward transfer) r = 12 (reverse transfer), the subscript b,e,c is added to define the type of configuration.

Now applying Kirchhoff's laws of voltage and current for eqn. 1 and eqn. 2, respectively in reverse to find out the circuit.

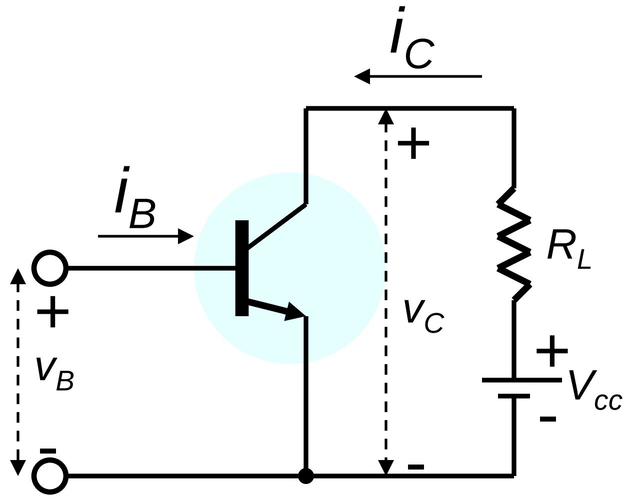


hybrid equivalent model

Determination of *h*- parameters from characteristics

Let us derive a hybrid model for CE transistor.
We may select i_B and v_C as independent variables.

$$v_B = f_1(i_B, v_C)$$
$$i_C = f_2(i_B, v_C)$$



Making a Taylor's series expansion of both the eqns. around the Q-point I_B and V_C and neglecting the higher order terms.

$$\Delta v_B = \left.\frac{\partial f_1}{\partial i_B}\right|_{V_C} \Delta i_B + \left.\frac{\partial f_1}{\partial v_C}\right|_{I_B} \Delta v_C$$

$$\Delta i_C = \left.\frac{\partial f_2}{\partial i_B}\right|_{V_C} \Delta i_B + \left.\frac{\partial f_2}{\partial v_C}\right|_{I_B} \Delta v_C$$

The partial derivatives are taken keeping collector voltage and base current constant.

Writing eqn. 1 and 2 for CE transistor:

$$v_B = h_{ie} i_B + h_{re} v_C$$
$$i_C = h_{fe} i_B + h_{oe} v_C$$

Therefore:

$$h_{ie} \equiv \frac{\partial f_1}{\partial i_B} = \left.\frac{\partial v_B}{\partial i_B}\right|_{V_C}$$

$$h_{re} \equiv \frac{\partial f_1}{\partial v_C} = \left.\frac{\partial v_B}{\partial v_C}\right|_{I_B}$$

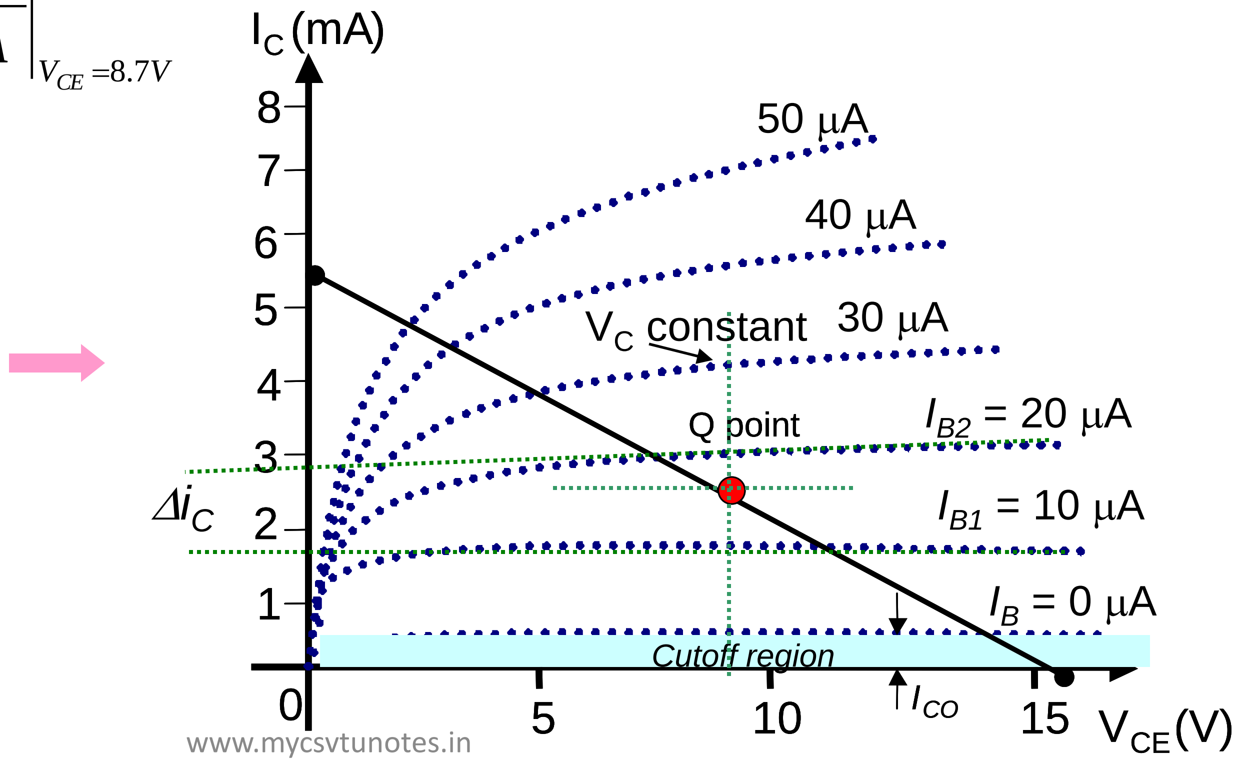
$$h_{fe} \equiv \frac{\partial f_2}{\partial i_B} = \left.\frac{\partial i_C}{\partial i_B}\right|_{V_C} \approx \frac{i_{C2} - i_{C1}}{i_{B2} - i_{B1}}$$

$$h_{oe} \equiv \frac{\partial f_2}{\partial v_C} = \left.\frac{\partial i_C}{\partial v_C}\right|_{I_B}$$

$$h_{fe} = \frac{\Delta i_C}{\Delta i_B} = \left.\frac{(2.7 - 1.7)mA}{(20 - 10)\mu A}\right|_{V_{CE}=8.7V}$$

$$\Rightarrow h_{fe} = 100$$

Transistor
output characteristics



h_{fe}

- Most important small signal parameter of the transistor.
- This common-emitter current transfer ratio, or CE alpha, is also written α_e or β' and called the small-signal beta of transistor.

$$h_{fe} \equiv \frac{\partial f_2}{\partial i_B} = \left.\frac{\partial i_C}{\partial i_B}\right|_{V_C} \approx \frac{i_{C2} - i_{C1}}{i_{B2} - i_{B1}}$$

Hybrid parameter variations

- The values of the h parameters depend upon the position of the quiescent point on the output and input characteristics curves.
- The characteristics curves depend on the junction temperature. Hence the h parameters depend up on Temperature.

Hybrid parameter variations

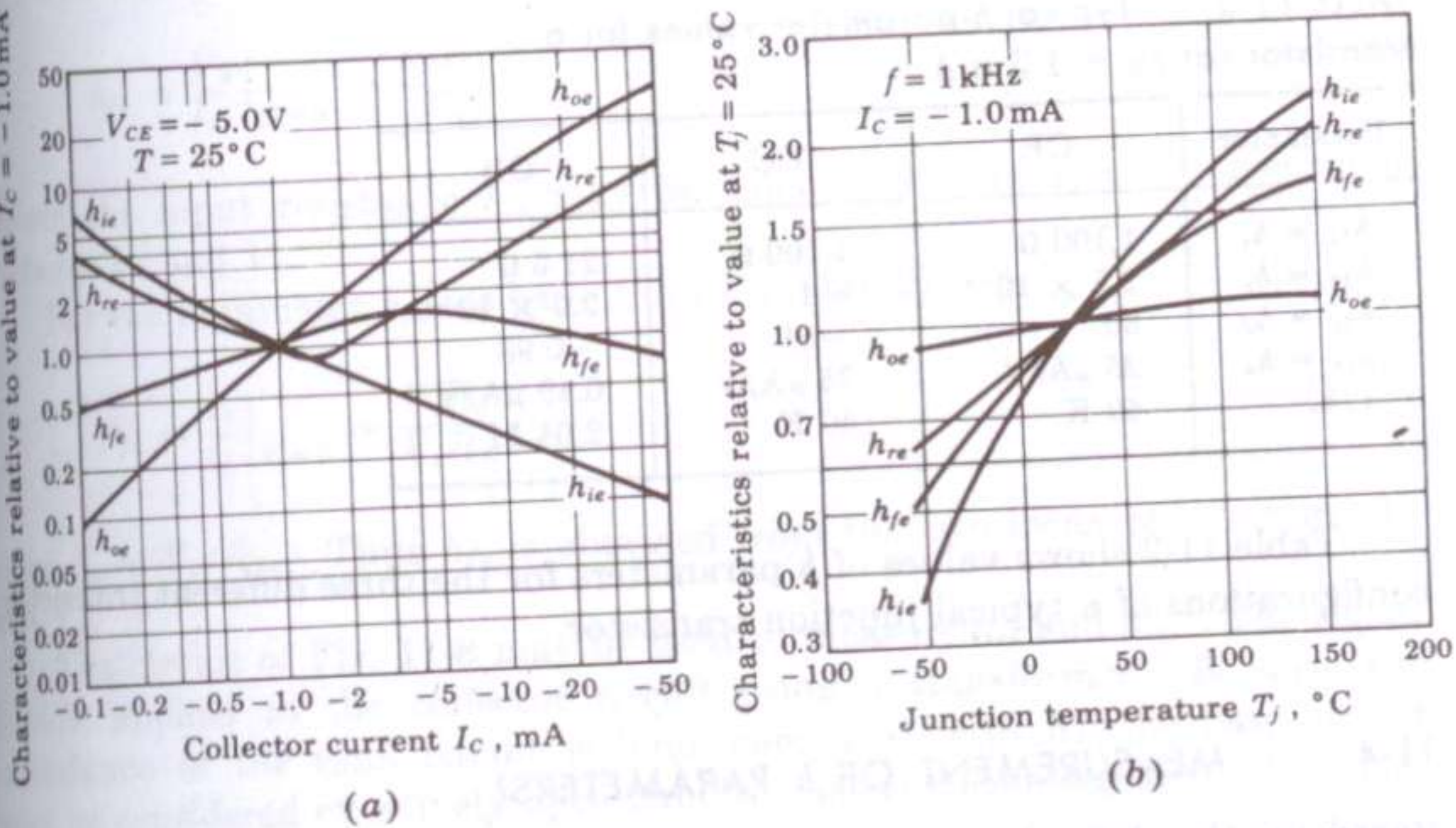


Fig. 11-5 Variation of common-emitter h parameters (a) with collector current normalized to unity at $V_{CE} = -5.0$ V and $I_C = -1.0$ mA for the type 2N996 diffused-silicon planar epitaxial transistor; (b) with junction temperature, normalized to unity at $T_j = 25°$C. (Courtesy of Fairchild Semiconductor.)

h-parameters of 2N3904

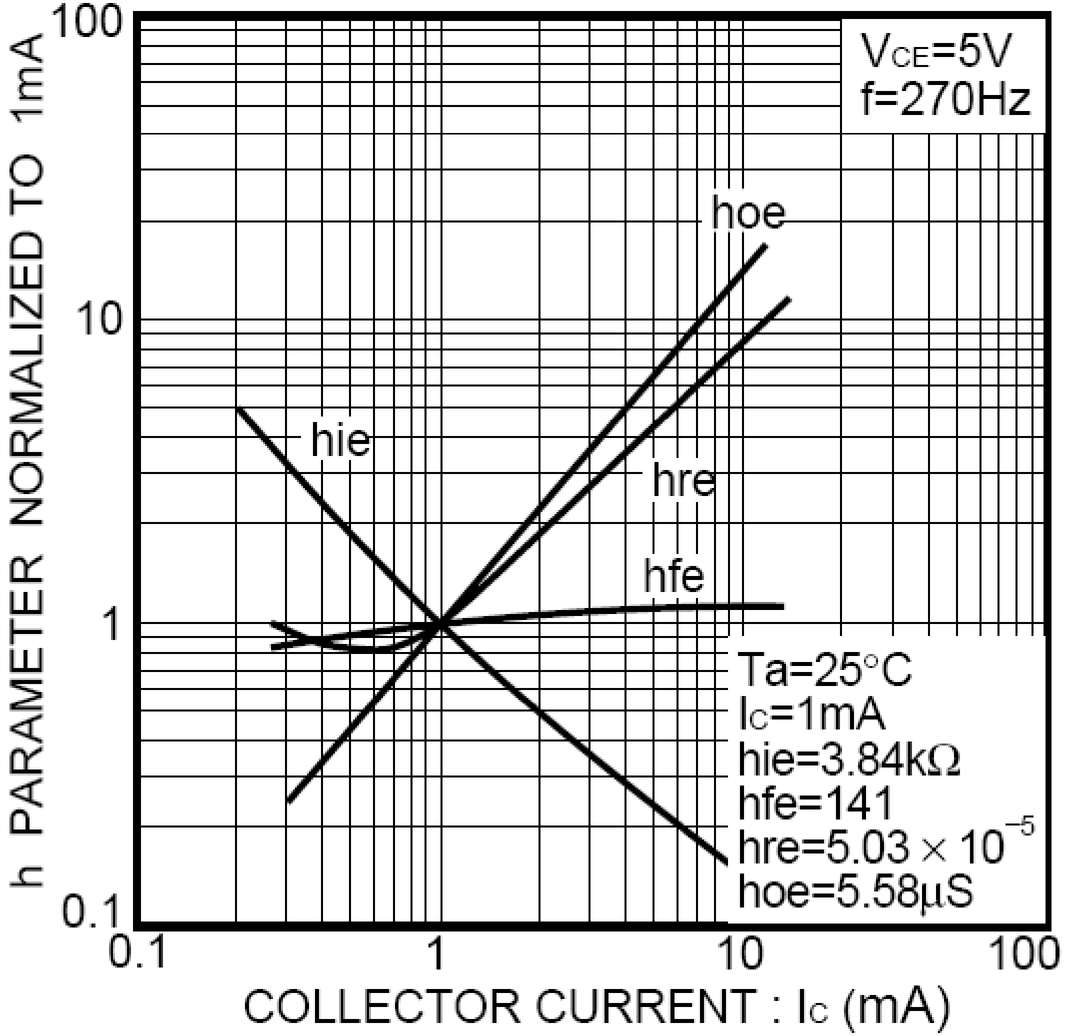


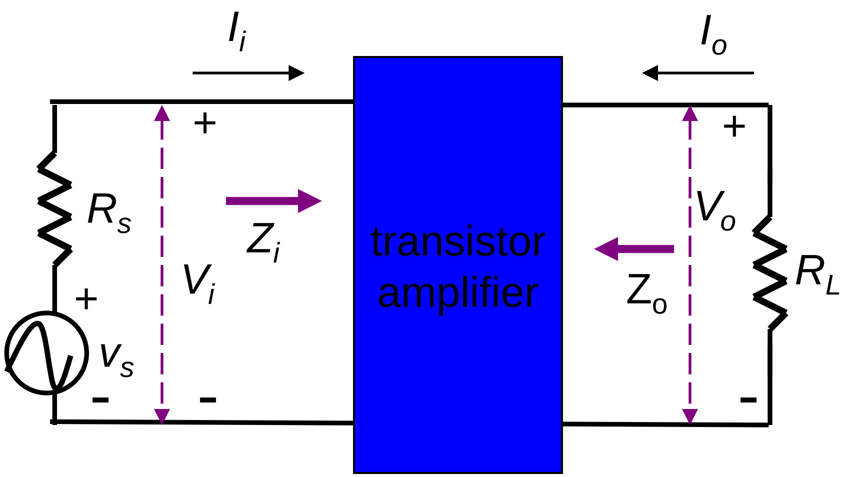
Fig.15 h parameter vs. collector current

Typical h-parameter values for a transistor(at I_E = 1.3 mA)

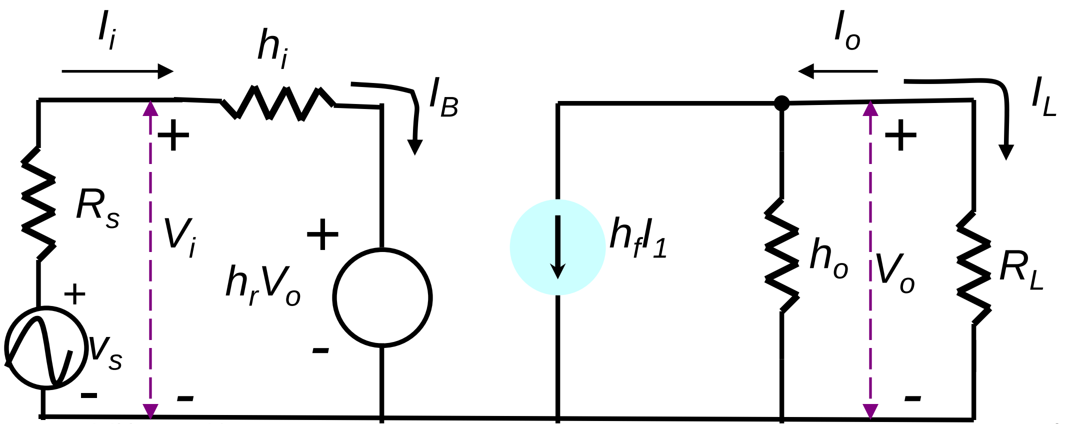
S.No	Parameter	CE	CC	CB
01	$h_i = h_{11}$	1100 Ω	1100 Ω	21.6 Ω
02	$h_r = h_{12}$	$2.5*10^{-4}$	~1	$2.9*10^{-4}$
03	$h_f = h_{21}$	50	-51	-0.98
04	$h_o = h_{22}$	24 μA/V	25 μA/V	0.49 μA/V
05	$1/h_o$	40K	40K	2.04 M

Analysis of a single stage amplifier circuit using hybrid parameter

For amplifying action we connect an external load and signal source to the transistor and properly bias it. Let us determine some important quantities.



A basic amplifier circuit



hybrid equivalent model

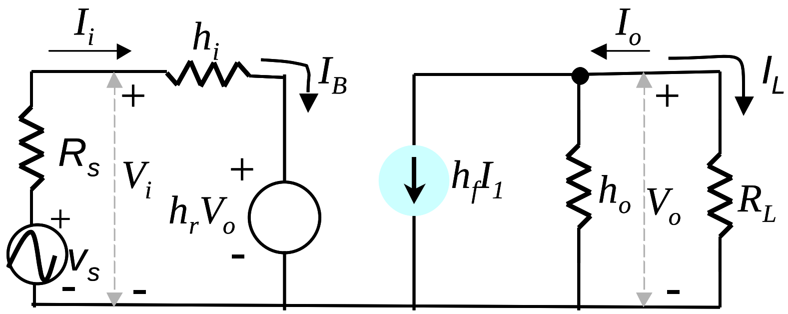
CURRENT GAIN A_I : $A_i = \frac{I_L}{I_i} = -\frac{I_o}{I_i}$

applying Kirchhoff's law to the output circuit yields

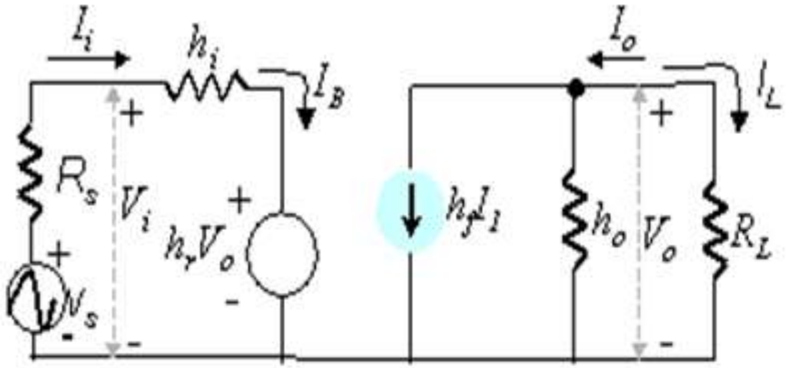
$$I_o = h_f I_i + h_o V_o$$; $V_o = - I_o R_L$

$$I_o = h_f I_i - h_o I_o R_L$$

$$\Rightarrow A_i = -\frac{h_f}{1 + h_o R_L} \cong -h_f$$ for $h_o R_L << 1$



INPUT IMPEDANCE $Z_i = \frac{v_i}{I_i}$



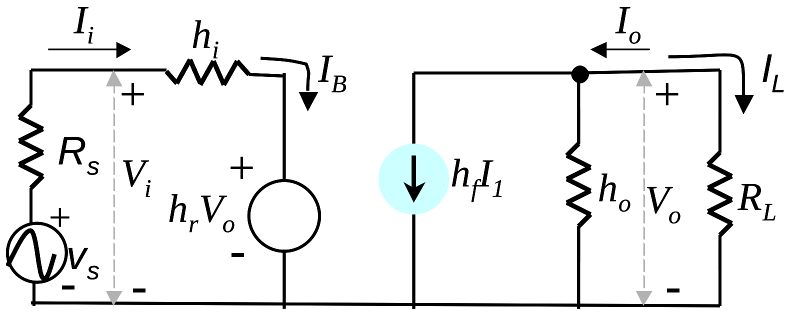
for input circuit

$$v_i = h_i I_i + h_r v_o$$ and $v_o = - I_o R_L$

$$Z_i = \frac{v_i}{I_i} = \frac{h_i I_i - h_r I_o R_L}{I_i} = h_i + h_r A_i R_L$$

$$\Rightarrow Z_i = h_i - \frac{h_f h_r R_L}{1 + h_o R_L}$$

VOLTAGE GAIN A_V : $Av = \frac{v_o}{v_i}$



$$Av = \frac{Vo}{V_i}$$

$$V_O = -I_O R_L = A_I I_i R_L$$

$$A_V = \frac{A_I I_i R_L}{V_i} = \frac{A_I R_L}{R_i}$$

IF USING IMPEDANCE

$$Av = \frac{Vo}{V_i}$$

$$V_O = -I_O Z_L = A_I I_i Z_L$$

$$A_V = \frac{A_I I_i Z_L}{V_i} = \frac{A_I Z_L}{Z_i}$$

OUTPUT IMPEDANCE Z_O : $Z_o = \left.\frac{v_o}{I_o}\right|_{v_S=0, Z_L=\infty}$

$$I_i R_s + h_i I_i + h_r v_o = 0$$ for input circuit with $v_s = 0$

$$I_o = h_f I_i + h_o v_o$$ for output circuit

putting value of I_i in above eqn. and rearranging, we get :

$$Z_o = \left.\frac{v_o}{I_o}\right|_{v_S=0} = \frac{1}{h_o - [h_f h_r / (h_i + R_s)]}$$

for input circuit

$$v_i = h_i I_i + h_r v_o$$

and $v_o = - I_o R_L$

$$\frac{V_i}{V_i} = h_i \frac{I_i}{V_i} + h_r \frac{V_O}{V_i}$$

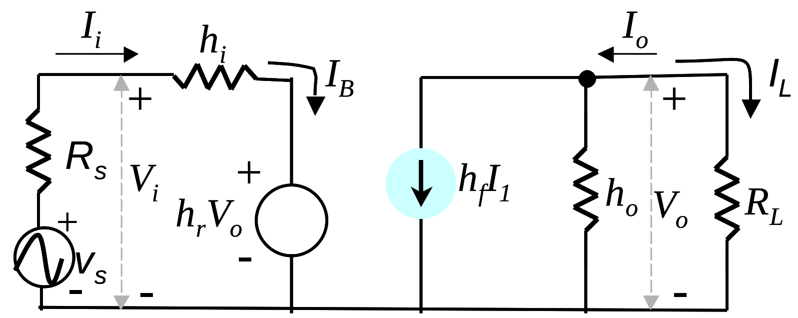
$$\Rightarrow Av = \frac{1}{h_r}[1 - \frac{h_i}{R_i}]$$

voltage amplification (considering the source resistance R_s) A_{vs} :

$$A_{vs} = \frac{v_o}{v_s} = \frac{v_o}{v_i}\frac{v_i}{v_s} = A_v\frac{v_i}{v_s}$$

$$v_i = \frac{v_s Z_i}{Z_i + R_s}$$

$$A_{vs} = \frac{v_o}{v_i} = \frac{A_v Z_i}{Z_i + R_s}$$



see the importance of A_{vs}

current amplification (considering the source resistance R_s) A_{is} :

$$A_{IS} = -\frac{I_o}{I_s} = -\frac{I_o}{I_i}\frac{I_i}{I_s} = A_I\frac{I_i}{I_s}$$

Draw norton equivalent circuit

$$I_i = \frac{I_s R_s}{Z_i + R_s} \quad \Rightarrow A_{IS} = A_I\frac{I_i}{I_s} = \frac{A_I R_s}{Z_i + R_s}$$

$$A_{IS} = A_I$$ if $R_s = \infty$

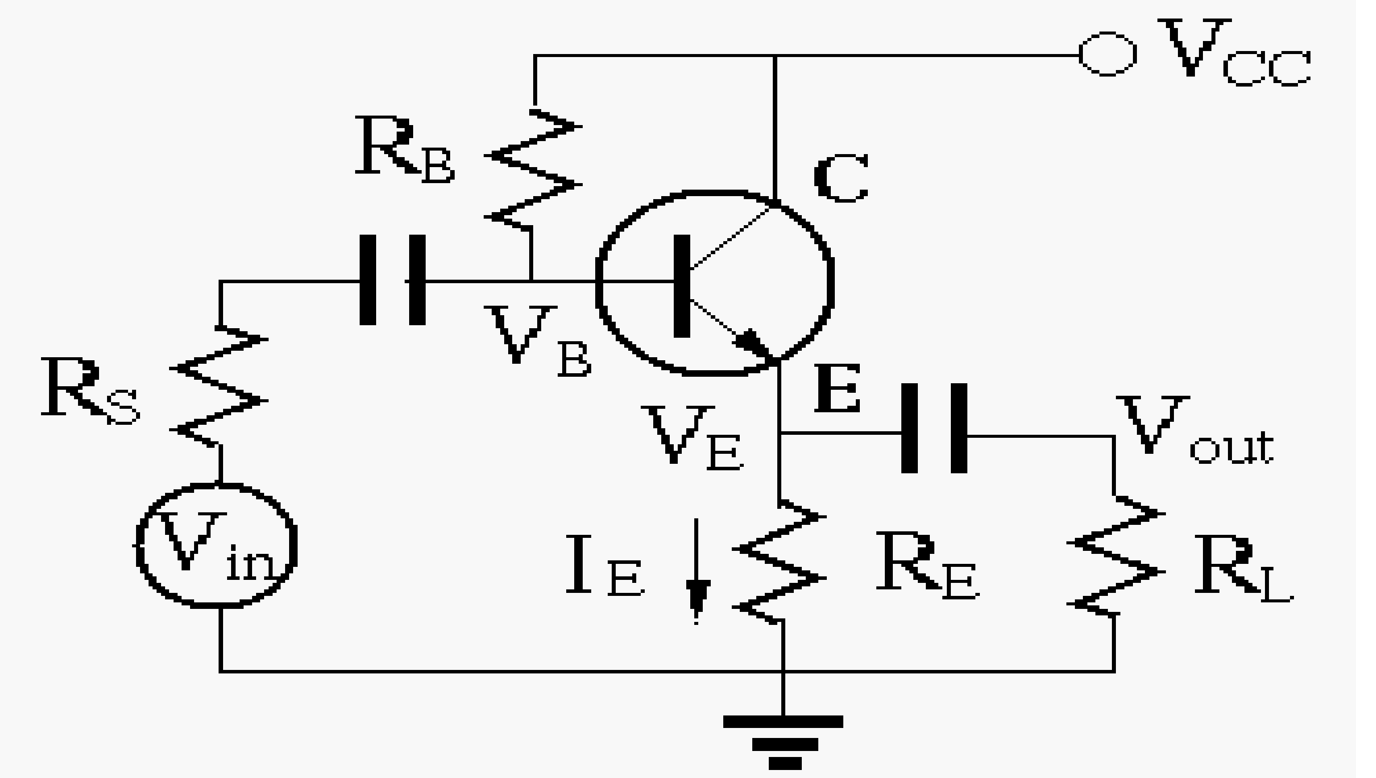
power gain A_p :

$$A_p = \frac{P_2}{P_1} = -\frac{v_o I_o}{v_s I_i} = A_V A_I$$

Small-signal analysis of a transistor amplifier

S No	*Quantities*	*Equation*
1	*Current gain*	$A_I = -h_f/(1+h_o Z_L)$
2	*Voltage gain*	$A_v = A_I Z_L/Z_i$
3	*Input impedance*	$Z_i = -h_i + h_r A_I Z_L$
4	*Output admittance*	$Y_o = h_o - h_f h_r/(h_i + R_s)$ $= 1/Z_o$
5	*Over all voltage gain*	$A_{vs} = A_v Z_i/(Z_i + R_s)$ $= A_I Z_L/(Z_i + R_s)$
6	*Over all current gain gain*	$A_{IS} = A_I R_s/(Z_i + R_s)$ $= A_{Vs} R_s/Z_L$

Emitter Follower Circuit



Emitter follower circuit

Characteristics of emitter follower circuit

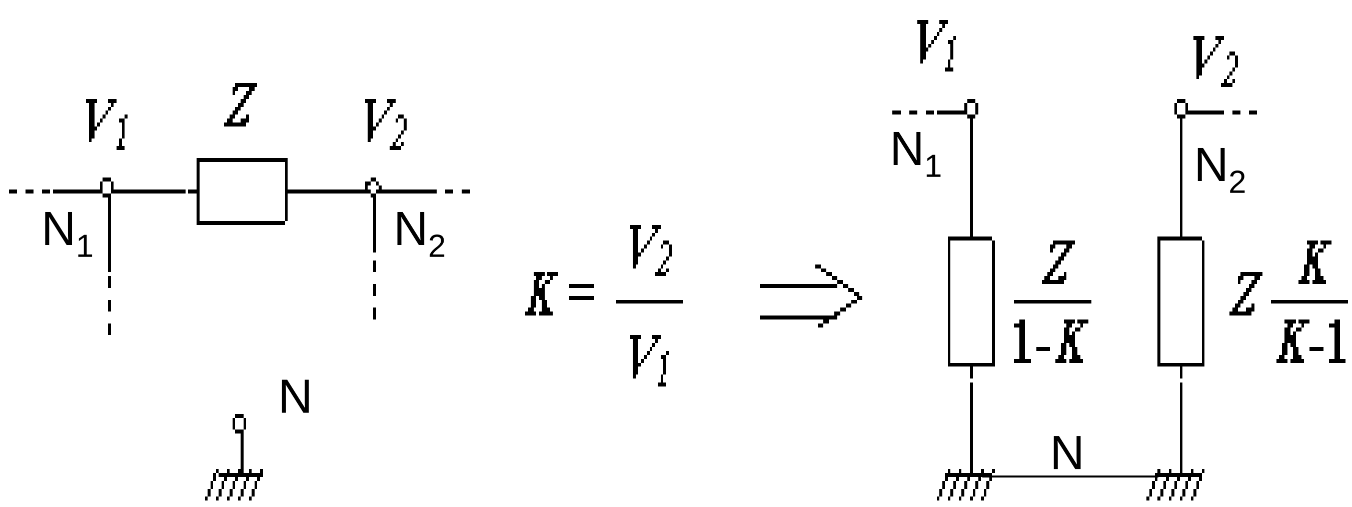
- Voltage gain is close to unity. (A_V = 0.997)
- Input resistance is very high (hundreds of kilo ohms). (R_i = 409K)
- Output resistance is very low (tens of ohms). (R_o = 41.2 ohms)
- It is use as a buffer stage which performs the resistance transformation over a wide range of frequencies, with voltage gain close to unity
- Emitter follower increases the power level of the signal.
- There is no phase shift between output and input in either voltage or current.

Comparison of transistor configurations (R_L=3KΩ, R_s=3KΩ)

S No	Quantity	CE	CC	CB
01	A_I	High(-46.5)	High(47.5)	Low(0.98)
02	A_V	High(-131)	Low(0.99)	High(131)
03	R_i	Medium(1065Ω)	High(144K)	Low(22.5Ω)
04	R_o	Medium(45.5K)	Low(80.5Ω)	High(1.72M)

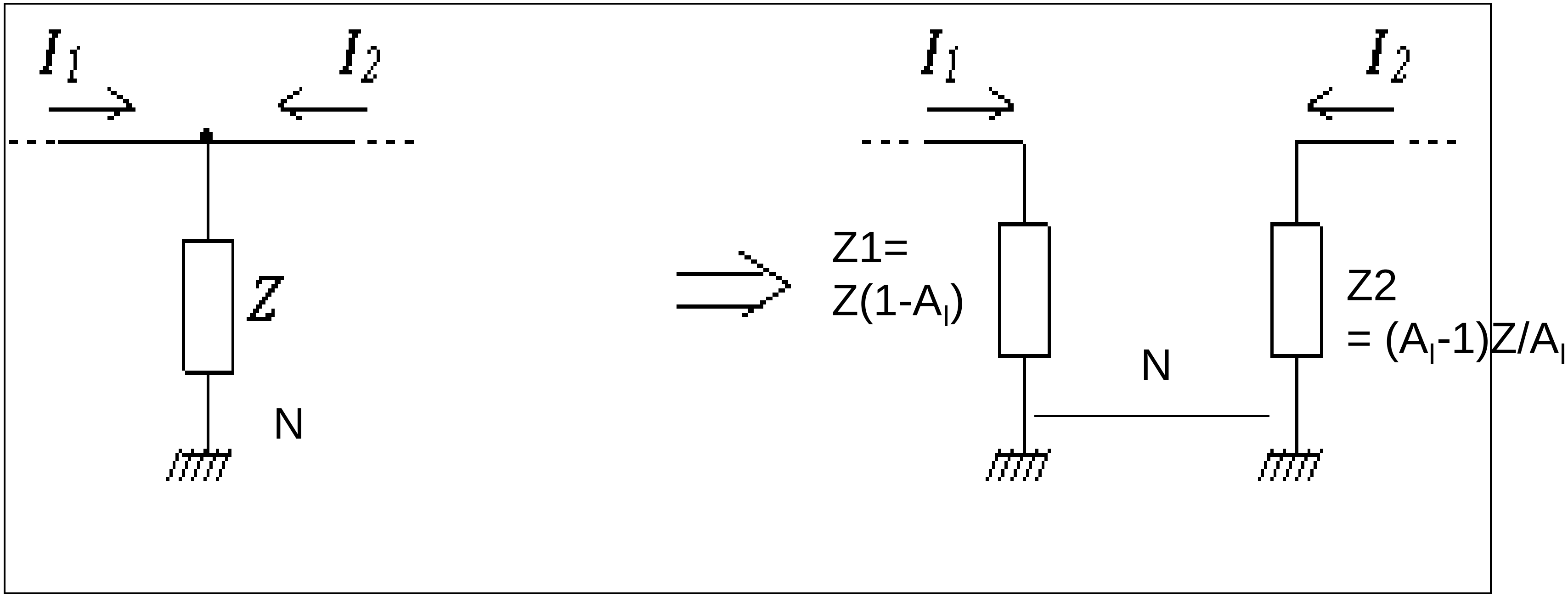
Miller's theorem

- ***The Miller's theorem establishes that in a linear circuit, if there exists a branch with impedance Z, connecting two nodes with nodal voltages V1and V2, we can replace this branch by two branches connecting the corresponding nodes to ground by impedances respectively Z/ (1-K) and KZ/ (K-1), where K = V2 / V1.***

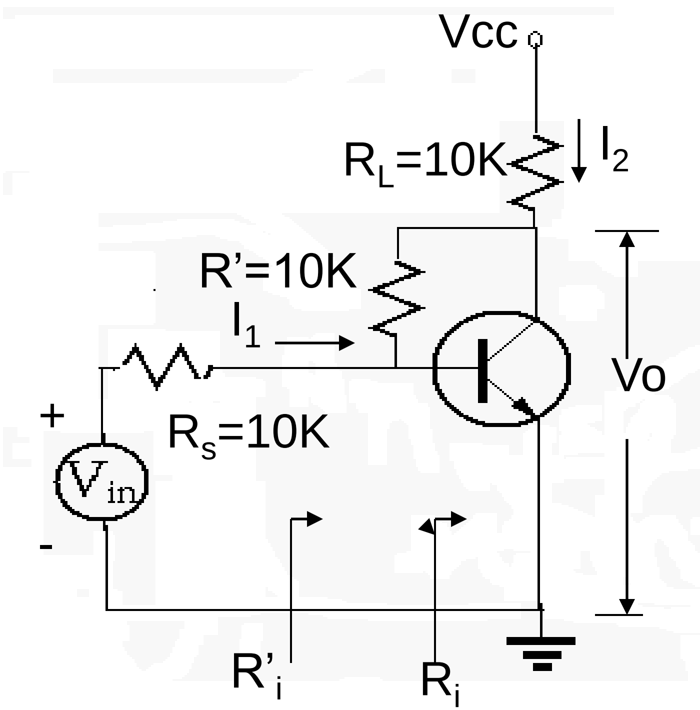


Miller's dual theorem

- If there is a branch in a circuit with impedance *Z* connecting a node, where two currents I_1 and I_2 converge, to ground, we can replace this branch by two conducting the referred currents, with impedances respectively equal to $(1- A_I)\,Z$ and $(A_I-1)\,Z / A_I$, where $A_I = -I_2 / I_1$.



Q.1 For the amplifier shown in fig calculate R_i, R'_i, A_V, A_{Vs} and $A'_I=-I_2/I_1$.



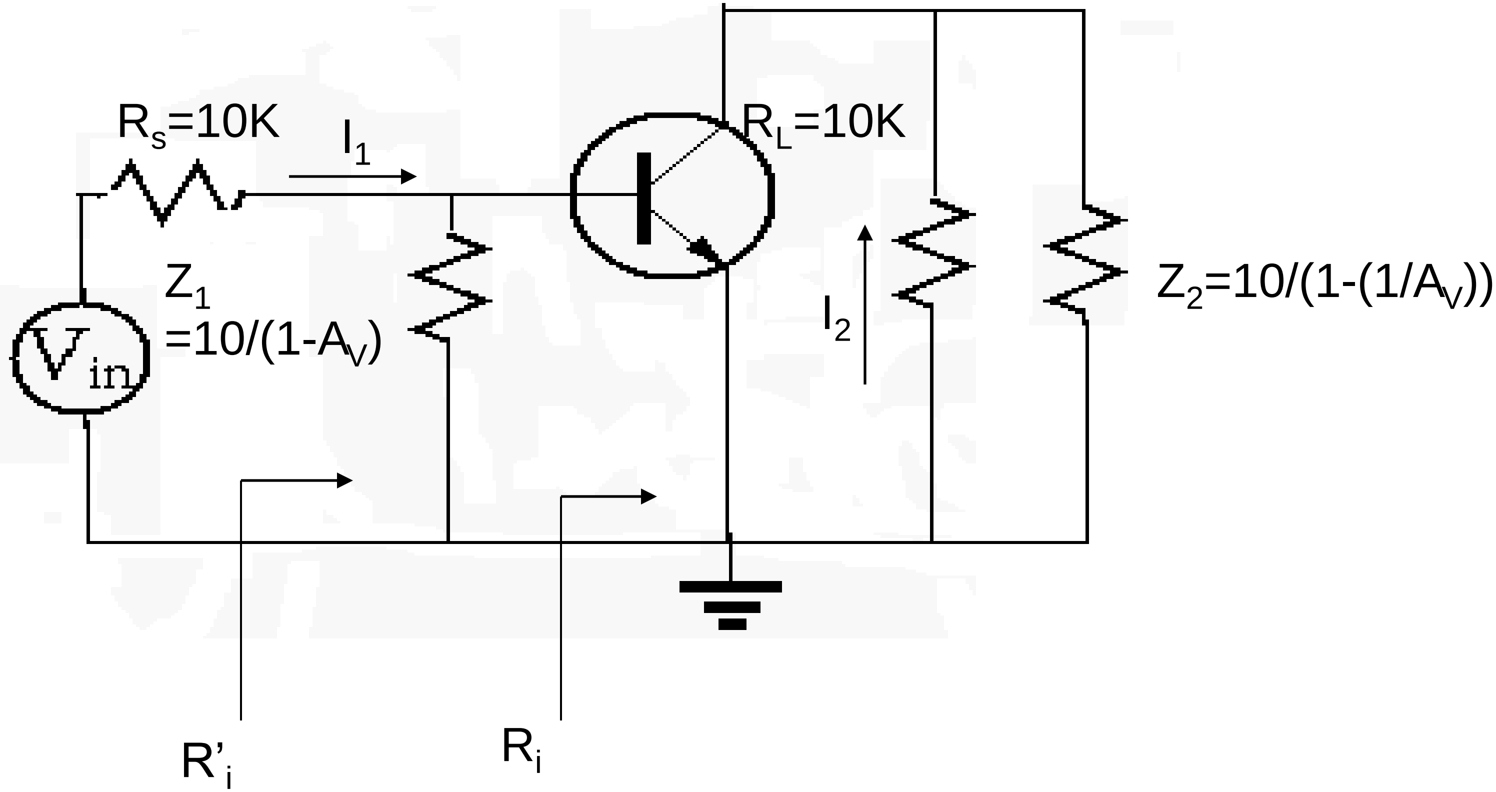
Steps

- Miller's theorem is applied to the 10K ohm resistance R'.
- $K=A_V$ is the voltage gain from base to collector.
- Assuming that gain is much larger than unity ($-A_V >> 1$), then

  $10/(1-(1/A_V)) \approx 10$

- Independent DC voltage source is replaced by a short circuit.

On applying Miller's Theorem

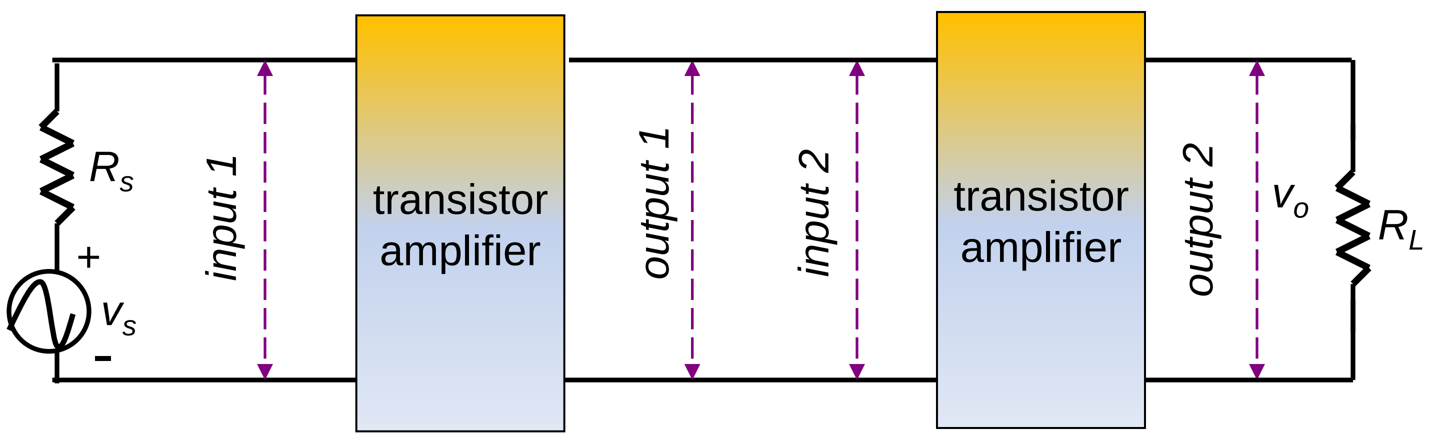


Multistage amplifier: cascading

- *Why do we require multistage cascading?*

1. *To increase the amplification*
2. *Impedance matching*

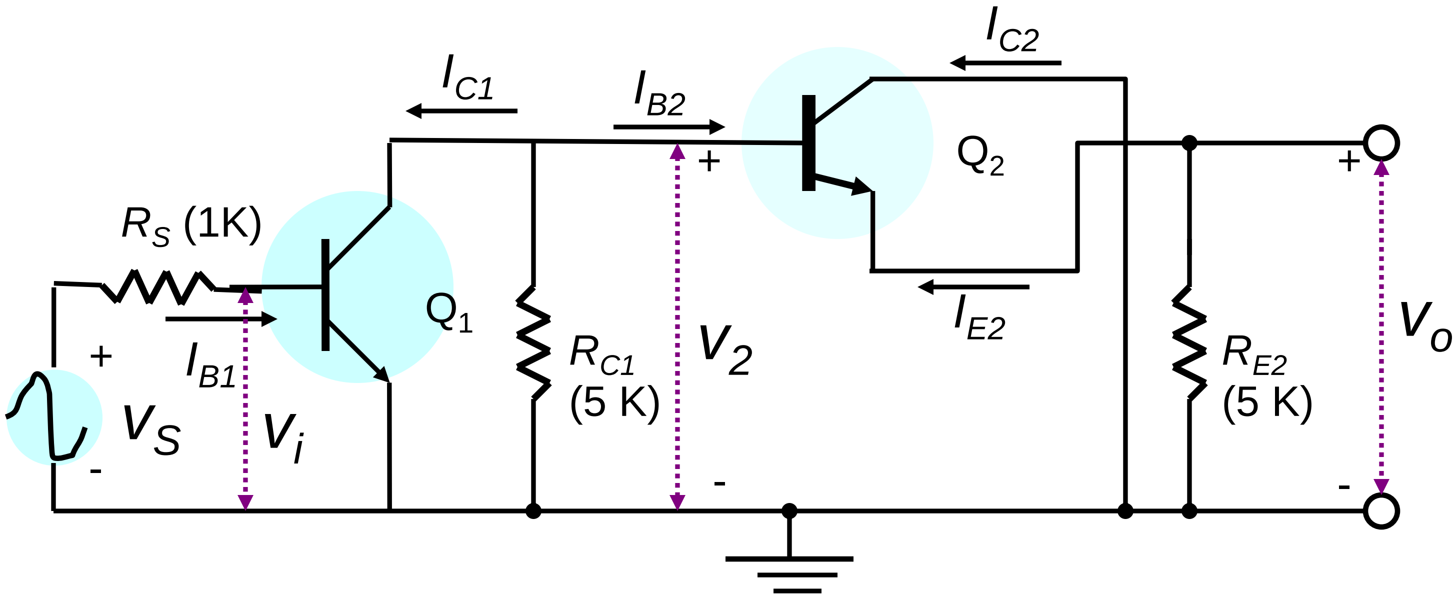
A cascade is a series connection with the output of one stage is applied to input of the second stage. ***The magnitude of the voltage gain equals the product of stage gains.*** *The resultant phase shift of a multistage amplifier equals the sum of the phase shift introduced by each stage.*



Two stage amplifier

One example: multistage amplifier

Let us find out the input, output impedances, voltage, current gains of Individual as well as overall for a CE-CC configuration.



An ac equivalent circuit for CE-CC cascade amplifier.

Q_1 CE	h_{ie} = 2K	h_{fe} = 50	h_{re} = 0.0006	h_{oe} = 25 μV/A
Q_2 CC	h_{ic} = 2K	h_{fc} = -51	h_{rc} = 1	h_{oc} = 25 μV/A

for Q_2 second stage CC configuration

$$A_{i2} = -\frac{h_{fc}}{1+h_{oc}R_{e2}} = -\frac{51}{1+25\times10^{-6}\times5\times10^{3}} = 45.3$$

$$Z_{i2} = h_{ic} - \frac{h_{fc}h_{rc}R_{e2}}{1+h_{oc}R_{e2}} = 2000 - \frac{-51\times1\times5\times10^{3}}{1+25\times10^{-6}\times5\times10^{3}} = 228.6\times10^{3}\Omega$$

$$A_{v2} = -\frac{h_{fc}R_{e2}}{h_{ic}+(h_{ic}h_{oc}-h_{fc}h_{rc})R_{e2}} = \frac{51\times5\times10^{3}}{2\times10^{3}+(2\times10^{3}\times25\times10^{-6}+51\times1)5000}$$

$$\Rightarrow A_{v2} \approx 1$$

for Q_1 first stage CE configuration the net load resistance $R_{L1}=R_{C1}||Z_{i2}$

$$R_{L1} = \frac{R_{C1}Z_{i2}}{R_{C1}+Z_{i2}} = \frac{5\times228.6}{2+228.6} = 4.95\times10^{3}\Omega$$

$$A_{i1} = -\frac{h_{fe}}{1+h_{oe}R_{L1}} = -\frac{50}{1+25\times10^{-6}\times4.95\times10^{3}} = -44.5$$

$$A_{v1} = -\frac{h_{fe}R_{L1}}{h_{ie} + (h_{ie}h_{ec} - h_{fe}h_{re})R_{L1}} = -\frac{50\times 4.95\times 10^3}{2\times 10^3 + (2\times 10^3 \times 25\times 10^{-6} - 50\times 6\times 10^{-4})4.95\times 10^3}$$

$$\Rightarrow A_{v2} \approx \boxed{116}$$

$$Z_{i1} = h_{ie} - \frac{h_{fe}h_{re}R_{L1}}{1+h_{oe}R_{L1}} = 2000 - \frac{50\times 6\times 10^{-4}\times 4.95\times 10^3}{1+25\times 10^{-6}\times 4.95\times 10^3} = \boxed{1.87\times 10^3\,\Omega}$$

$$Z_{o1} = \frac{1}{h_{oe} - [h_{fe}h_{re}/(h_{ie}+R_s)]} = 66.7\times 10^3\,\Omega$$

The above is open-circuit output impedance, so output impedance of first stage $Z_{o1`}=Z_{o1}||R_{C1}$= 4.65 K

The output impedance of the first stage will act as the effective source resistance for the second stage Q_2; so $R_{S`} = Z_{o1`}$= 4.65 K

$$Z_{o2} = \frac{1}{h_{oe} - [h_{fe}h_{re}/(h_{ie}+R_{s`})]} = 130\Omega$$

The above is open-circuit output impedance, so output impedance of second stage $Z_{o2`}=Z_{o2}||R_{e2}= 127\Omega$

overall current gain $A_I = -\frac{I_{e2}}{I_{b1}} = -\frac{I_{e2}}{I_{b2}}\frac{I_{b2}}{I_{c1}}\frac{I_{c1}}{I_{b1}} = -A_{i2}\frac{I_{b2}}{I_{c1}}A_{i1}$

$$\therefore I_{b2} = -\frac{R_{c1}}{Z_{i2}+R_{C1}} I_{C1} \Rightarrow A_I = \frac{R_{C1}}{Z_{i2}+R_{C1}} A_{i1}A_{i2} = \frac{5}{228.6+5} \times 45.3 \times (-44.5)$$

$$\Rightarrow A_I = -43.2$$

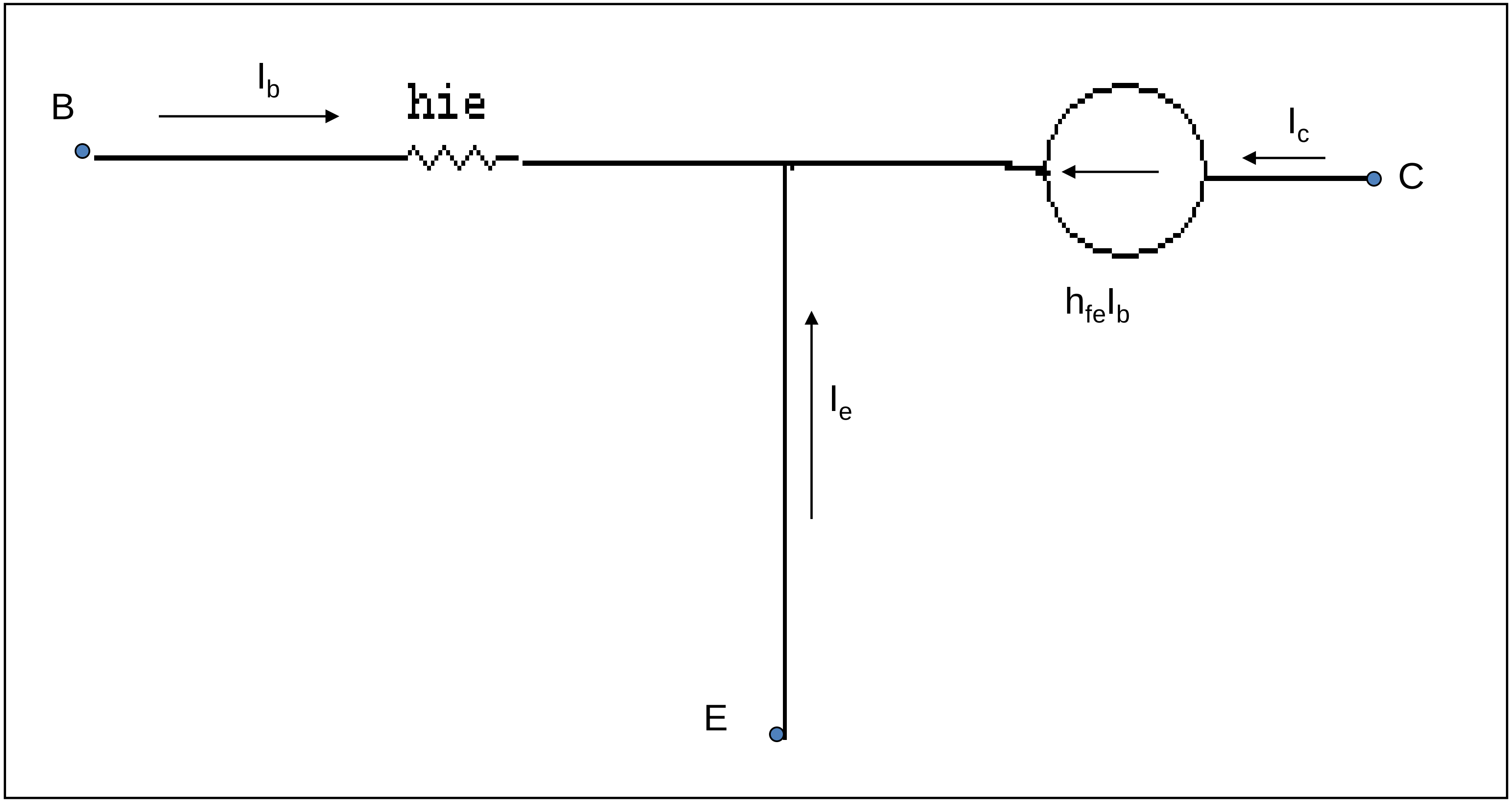
overall voltage gain $A_V = \frac{v_o}{v_i} = \frac{v_o}{v_2}\frac{v_2}{v_i} = A_{v1}A_{v2} \cong 1 \times (-116) = -116$

overall voltage gain considering source impedance R_s

$$A_{vs} = \frac{A_v Z_{i1}}{Z_{i1}+R_s} = -116 \times \frac{1.87}{1.87+1} = -75.3$$

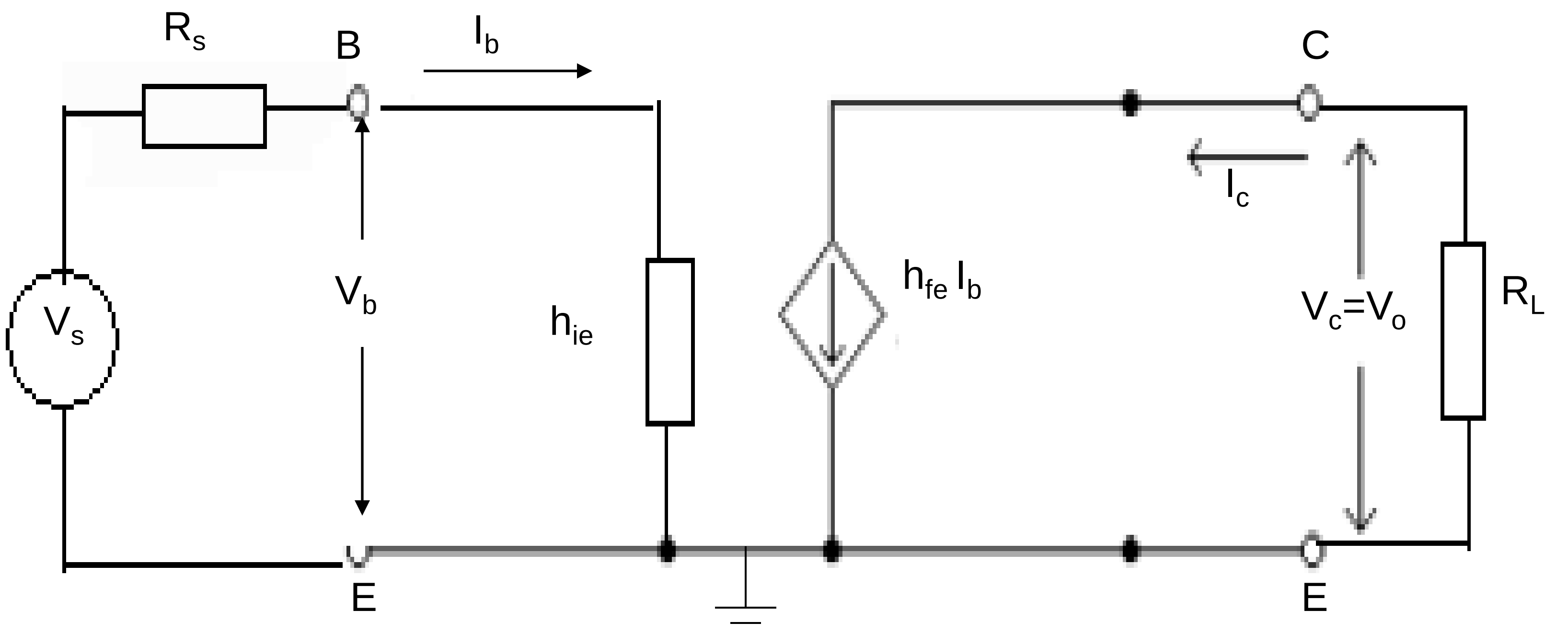
	Q_1 CE	Q_2 CC	overall CC-CE
A_i	-44.6 ×	45.3 ≠	-43.2
A_v	-116 ×	1 =	-116
Z_i	1.87 KΩ	228.6 KΩ	1.87 KΩ
Z_o`	4.65 KΩ	127 Ω	127 Ω

Simplified common-emitter hybrid model

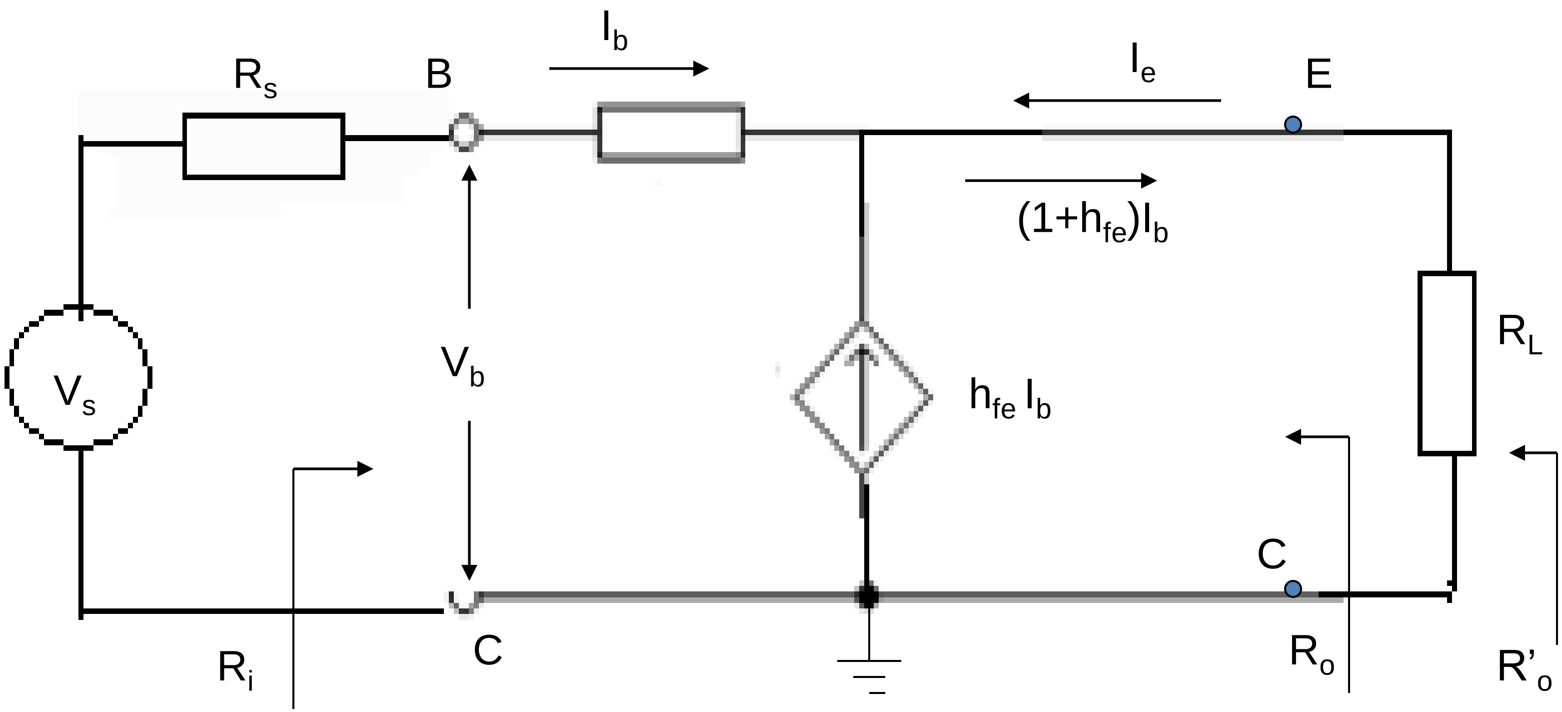


Approximate CE model

- It is valid if $h_{oe}R_L < 0.1$

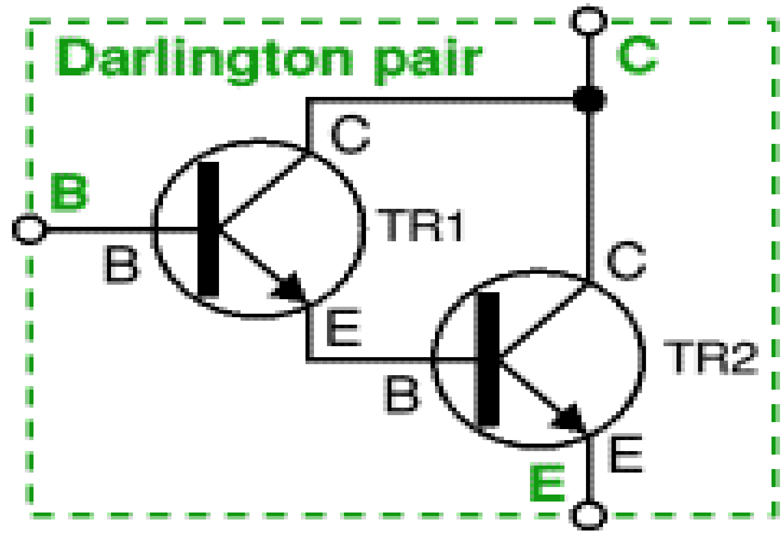


Simplified hybrid model for the common-collector circuit



Darlington Pair

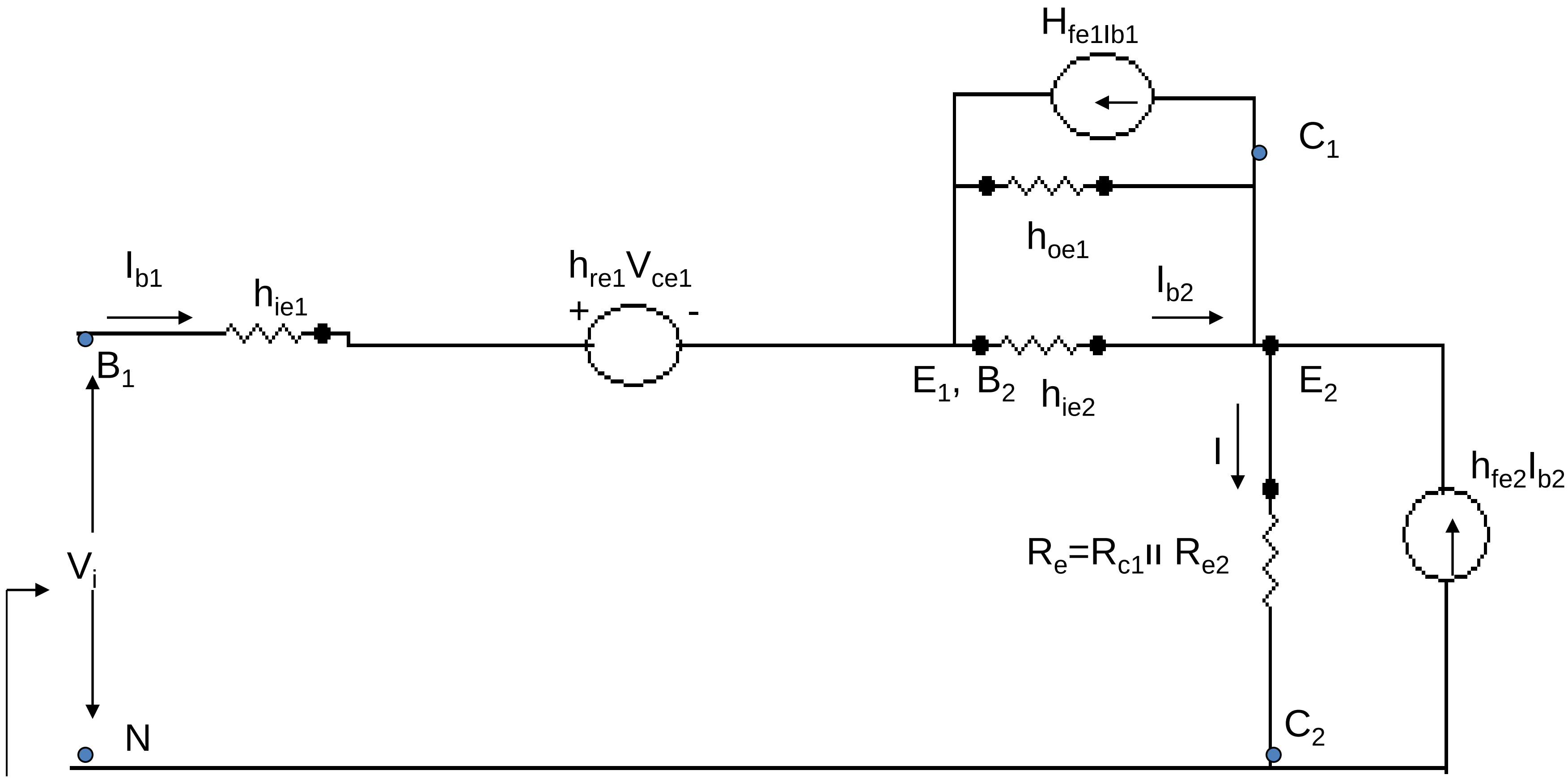
- ***What is a Darlington Pair?***
- A Darlington pair is two transistors that act as a single transistor but with a much higher current gain.



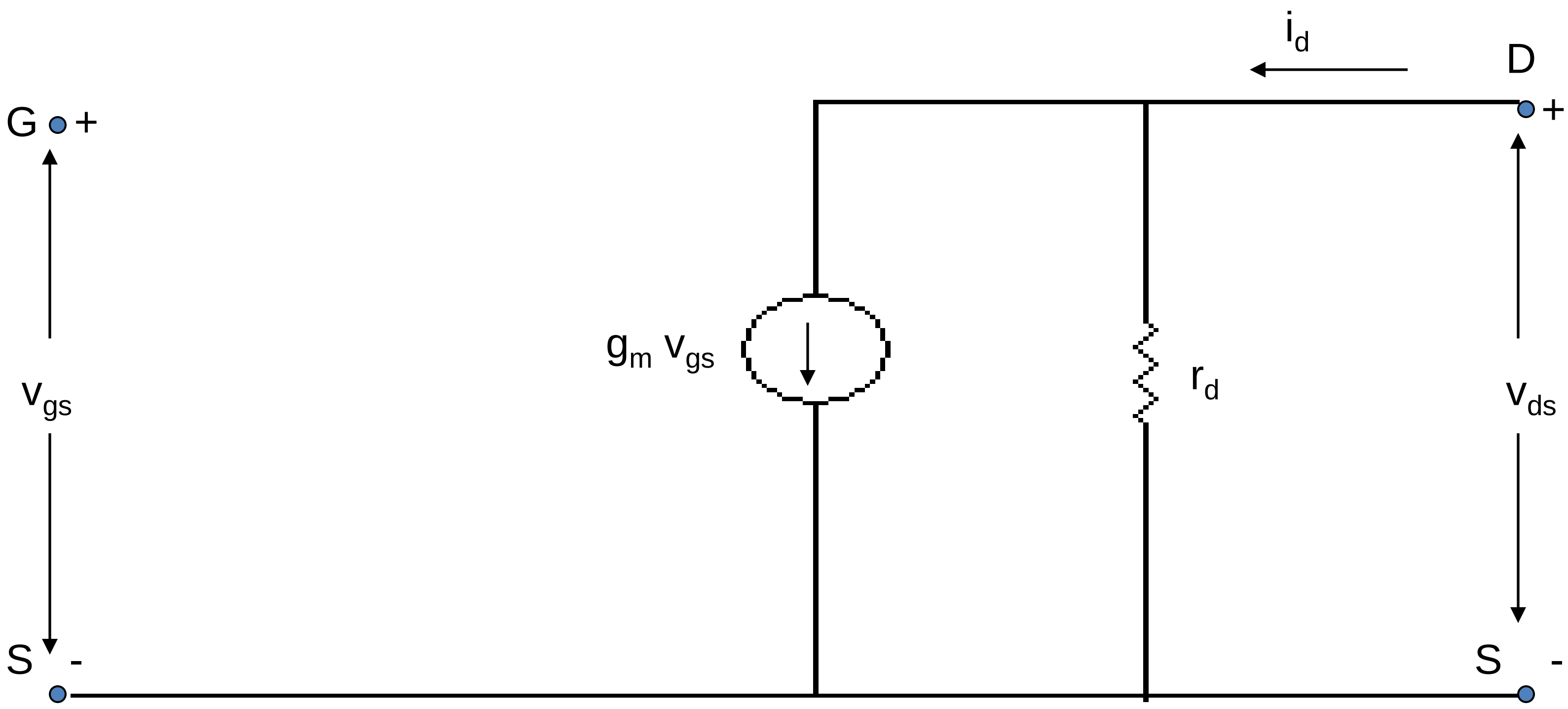
DRABACK OF DARLINGTON PAIR

- Leakage current of 1st transistor is amplified by 2nd .

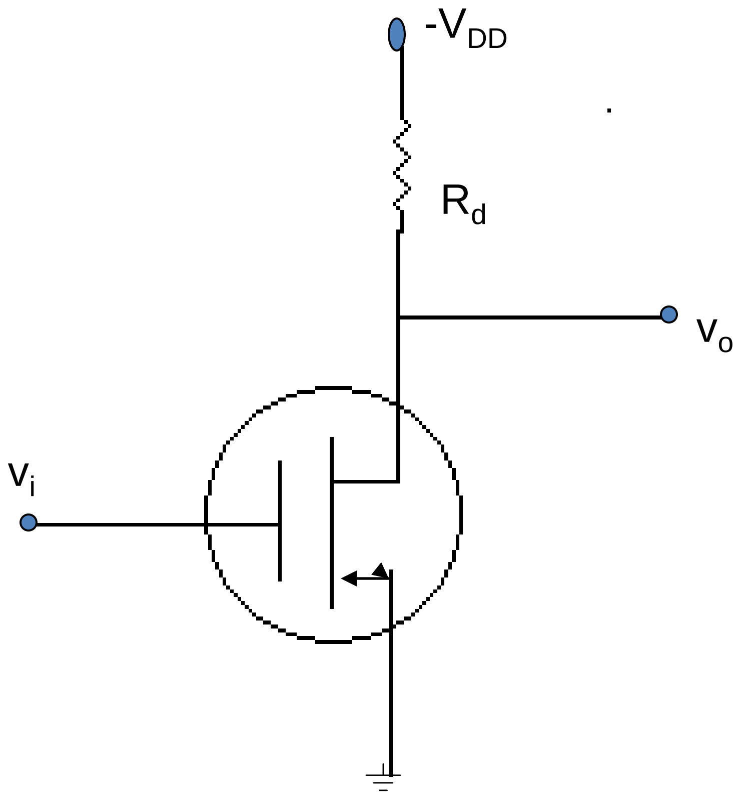
Bootstrapped Darlington equivalent circuit

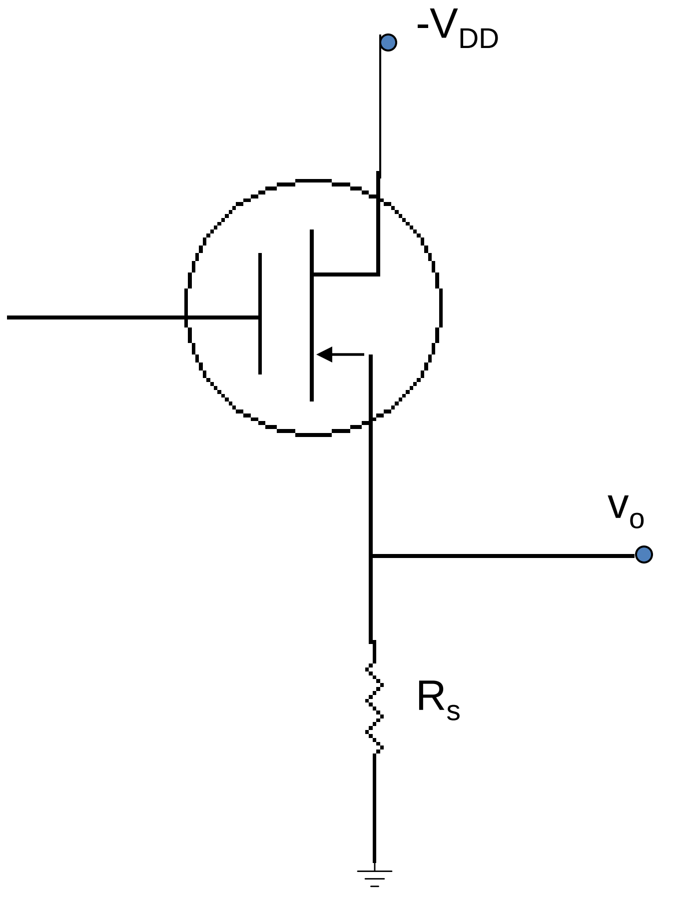


The low-frequency small-signal FET model



The CS and the CD configuration

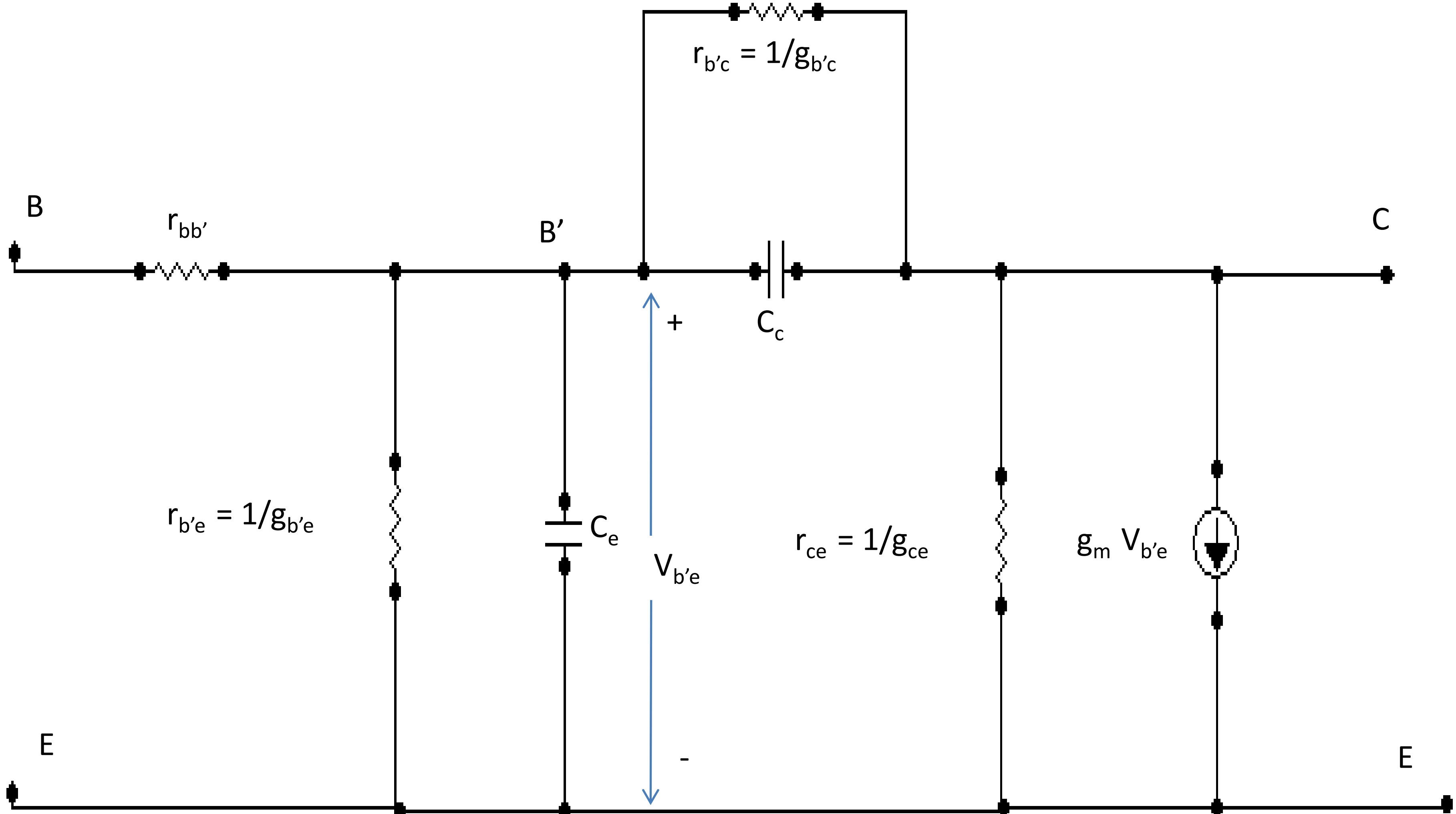




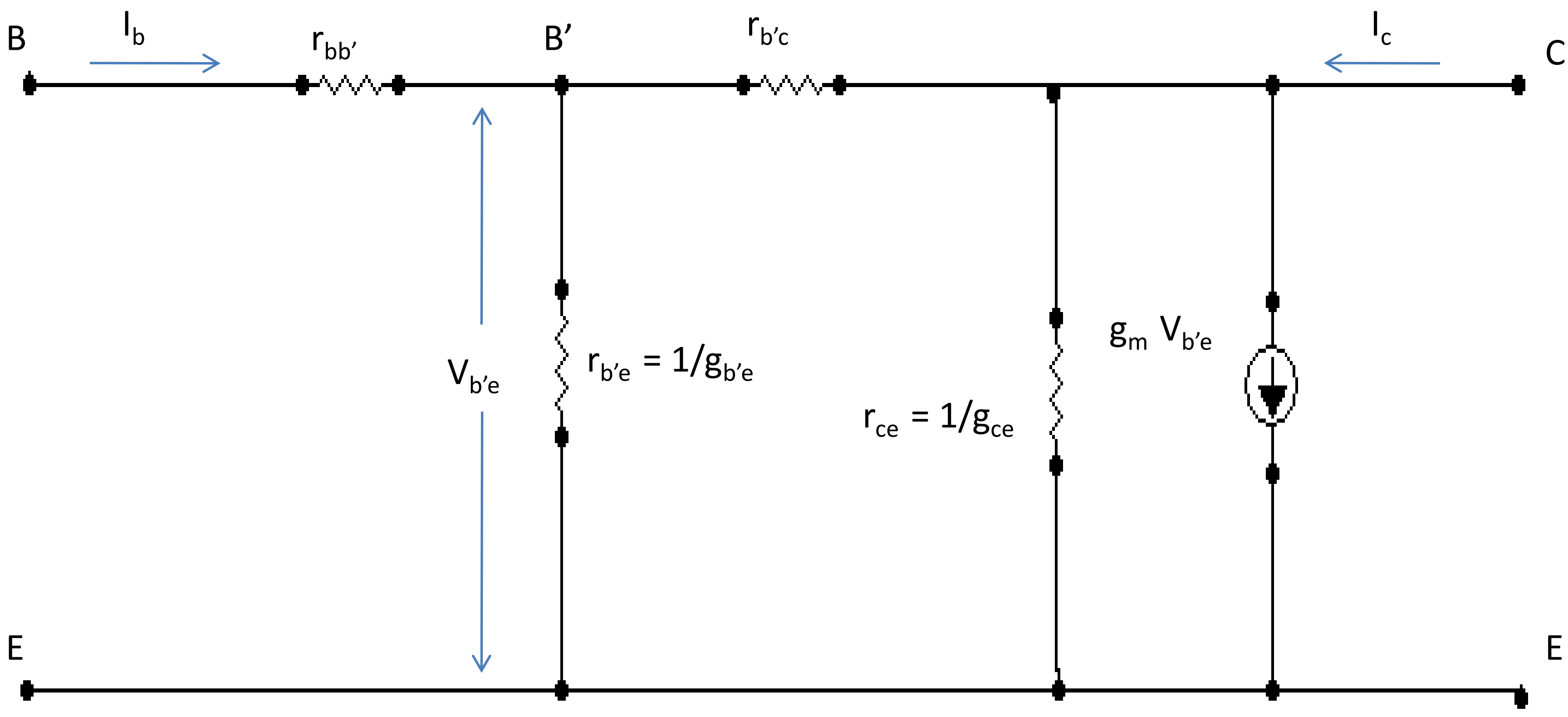
UNIT-2

TRANSISTOR AT HIGH FREQUENCIES

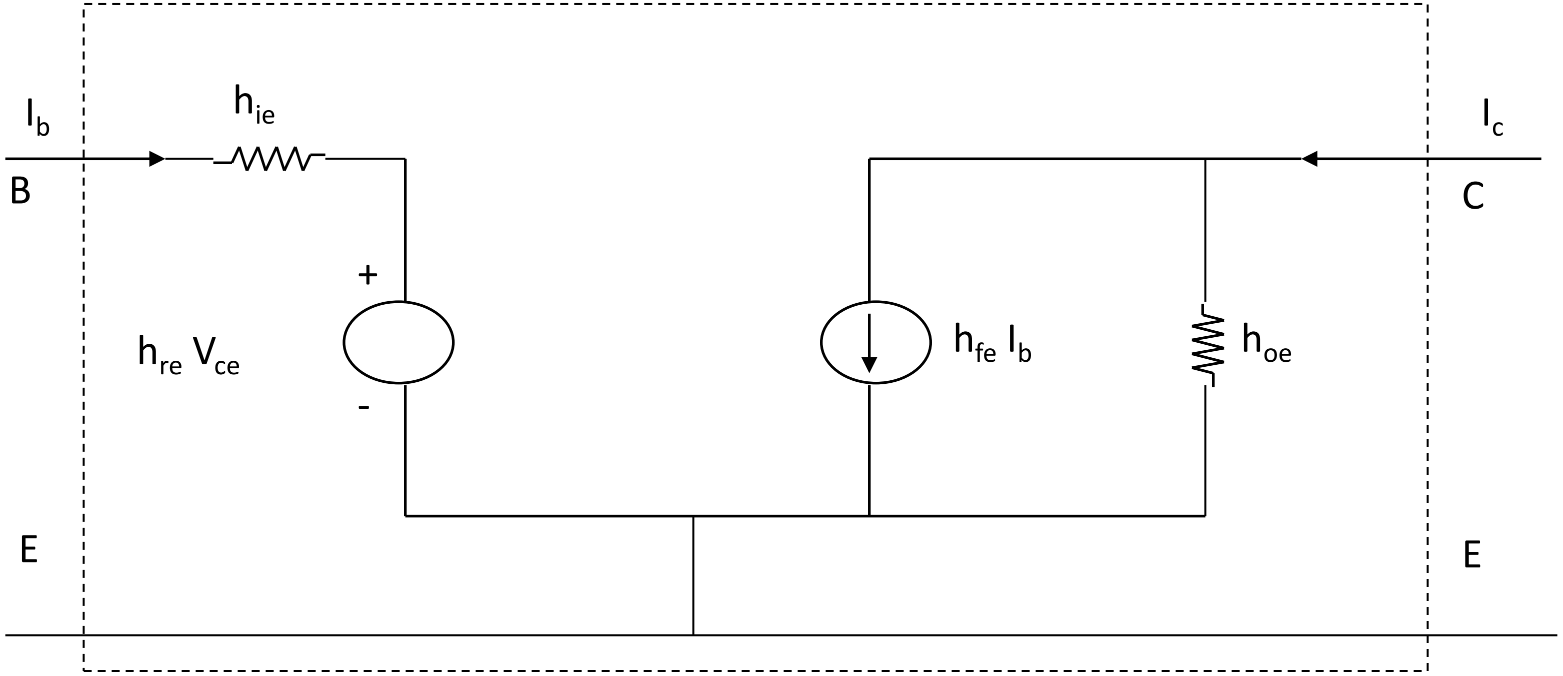
The hybrid-∏ model for a transistor in the CE configuration



The hybrid-∏ model for a transistor in the CE configuration at low frequency



The hybrid small signal models for common-emitter configuration

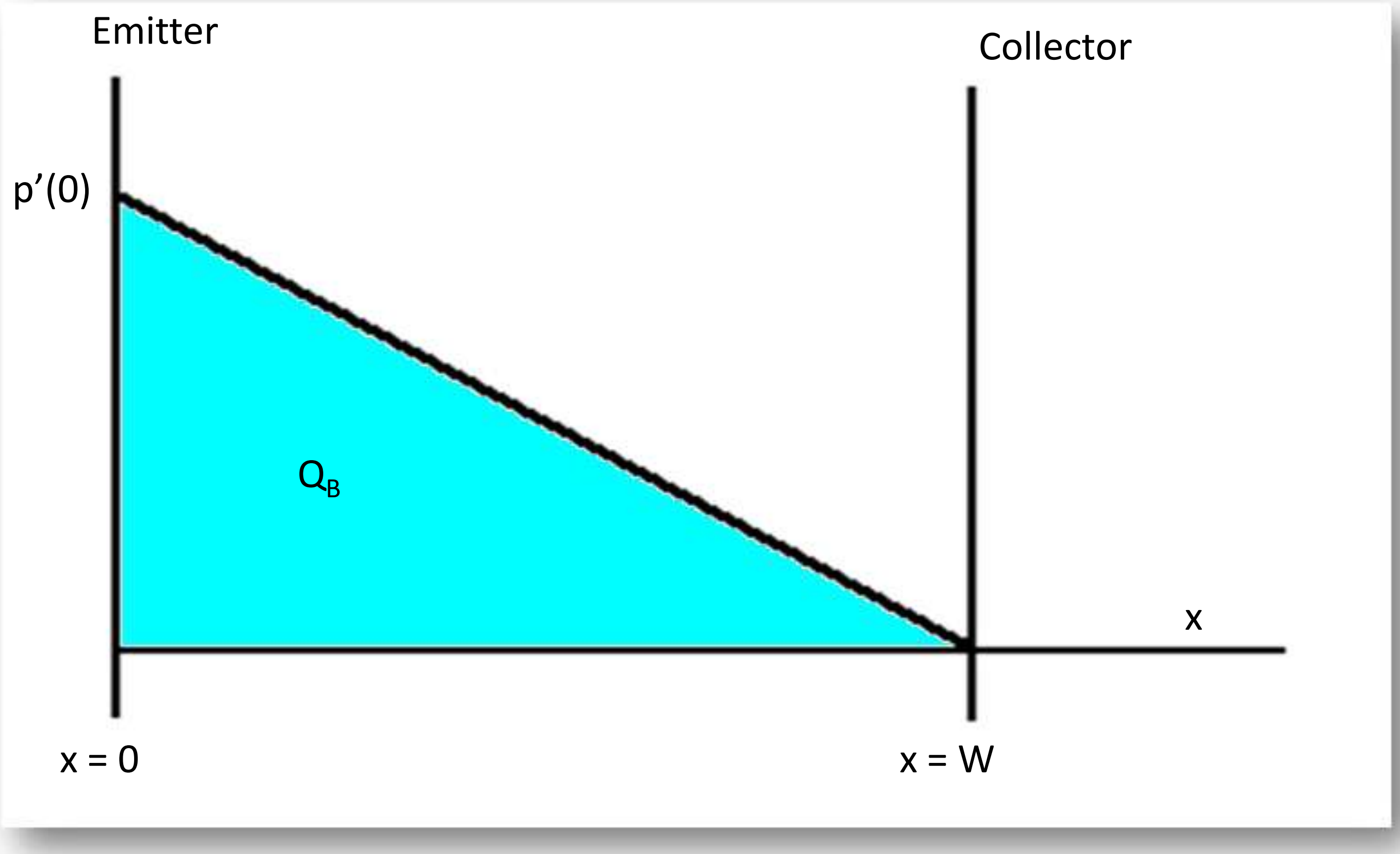


HYBRID-∏ CONDUCTANCES

S.NO	HYBRID-∏ CONDUCTANCES	EQUATION
1	Transistor Transconductance	$g_m = \|I_c\|/V_T$
2	The Input Conductance	$r_{b'e} = h_{fe}/g_m = h_{fe}V_T/\|I_c\|$ Or $g_{b'e} = g_m/h_{fe}$
3	The Feedback Conductance	$r_{b'c} = r_{b'e}/h_{re}$ Or $g_{b'c} = h_{re}/r_{b'e}$
4	The Base Spreading Resistance	$r_{bb'} = h_{ie} - r_{b'e}$
5	The Output Conductance	$g_{ce} = h_{oe} - (1+h_{fe})g_{b'c} = 1/r_{ce}$

HYBRID-∏ CAPACITANCE

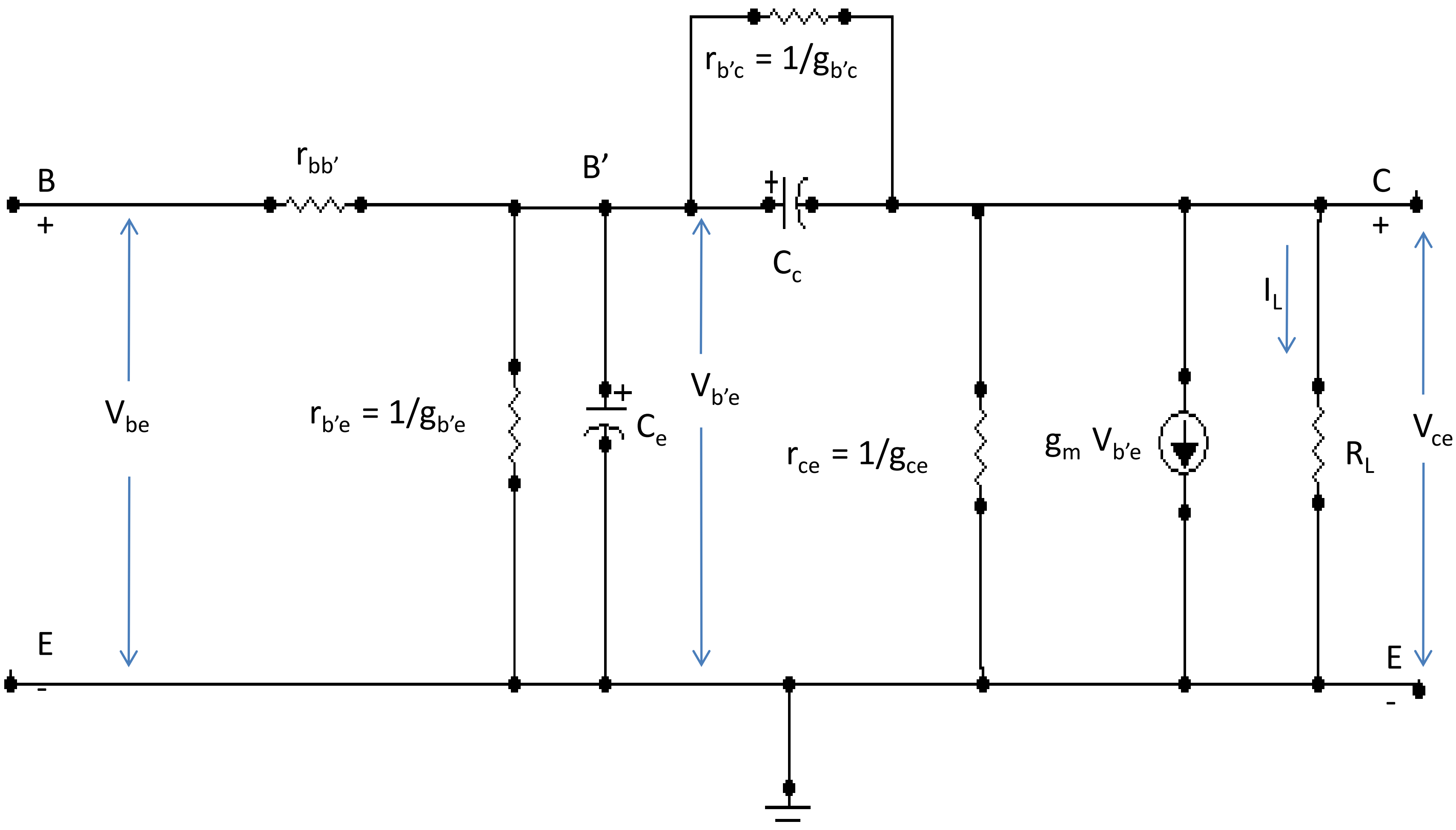
Minority-carrier charge distribution in the base region



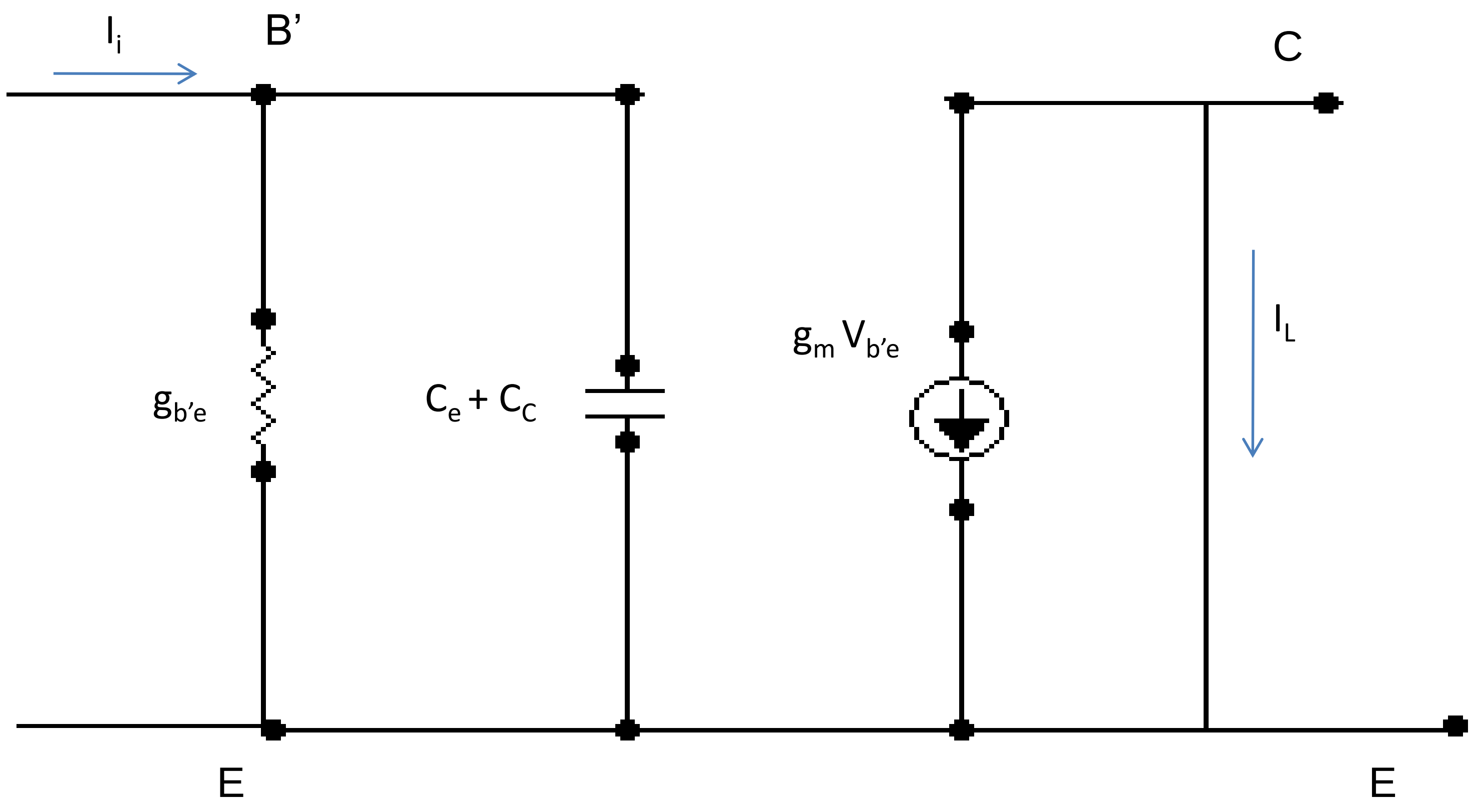
CE Short-Circuit Current gain

- The load R_L on this stage is the collector-circuit so that $R_c = R_L$.
- For short circuit current gain $R_L = 0$.

The hybrid-∏ circuit for a single transistor with a resistive load R_L



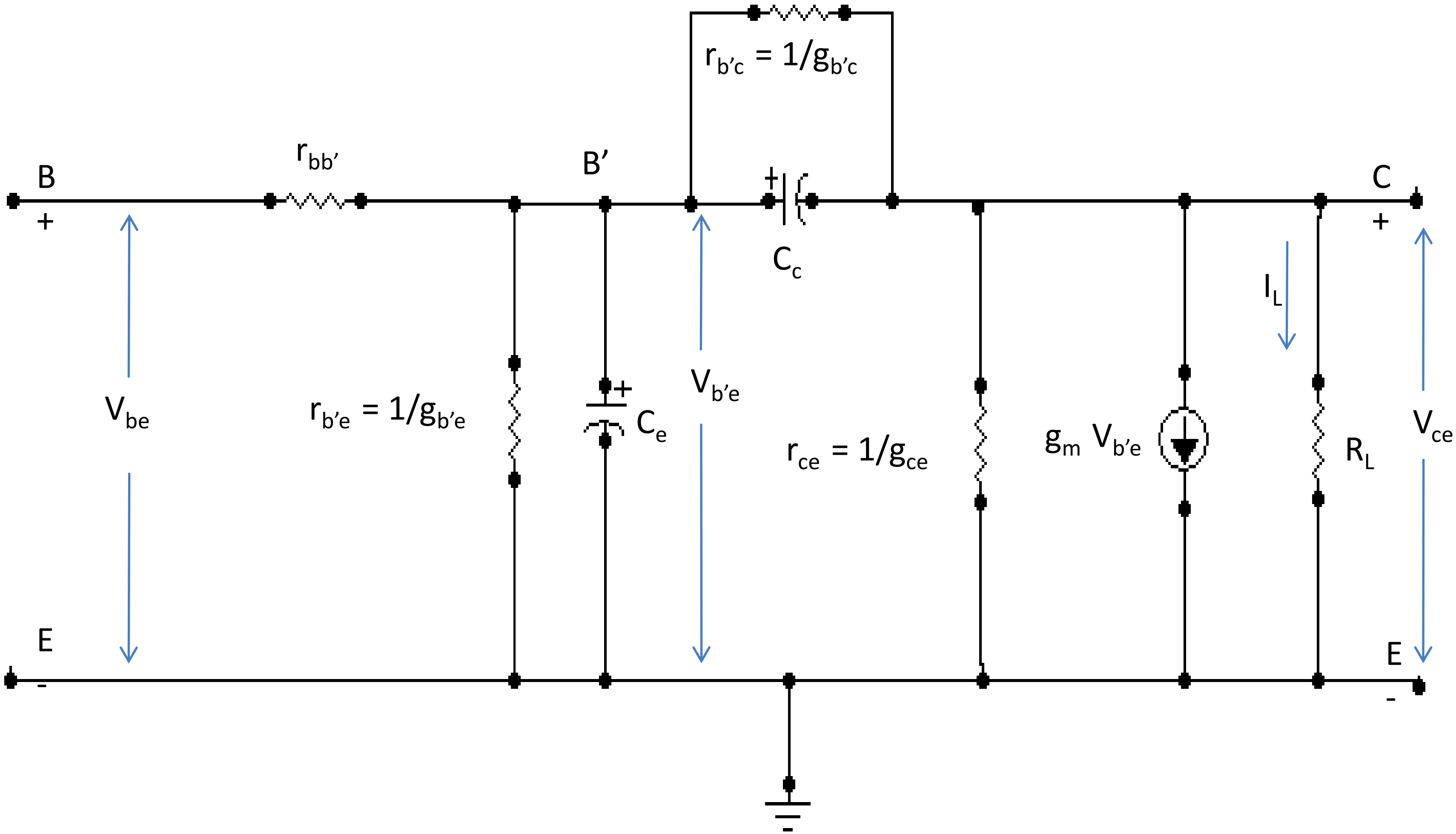
Approximate equivalent circuit for the calculation of the short-circuit CE current gain



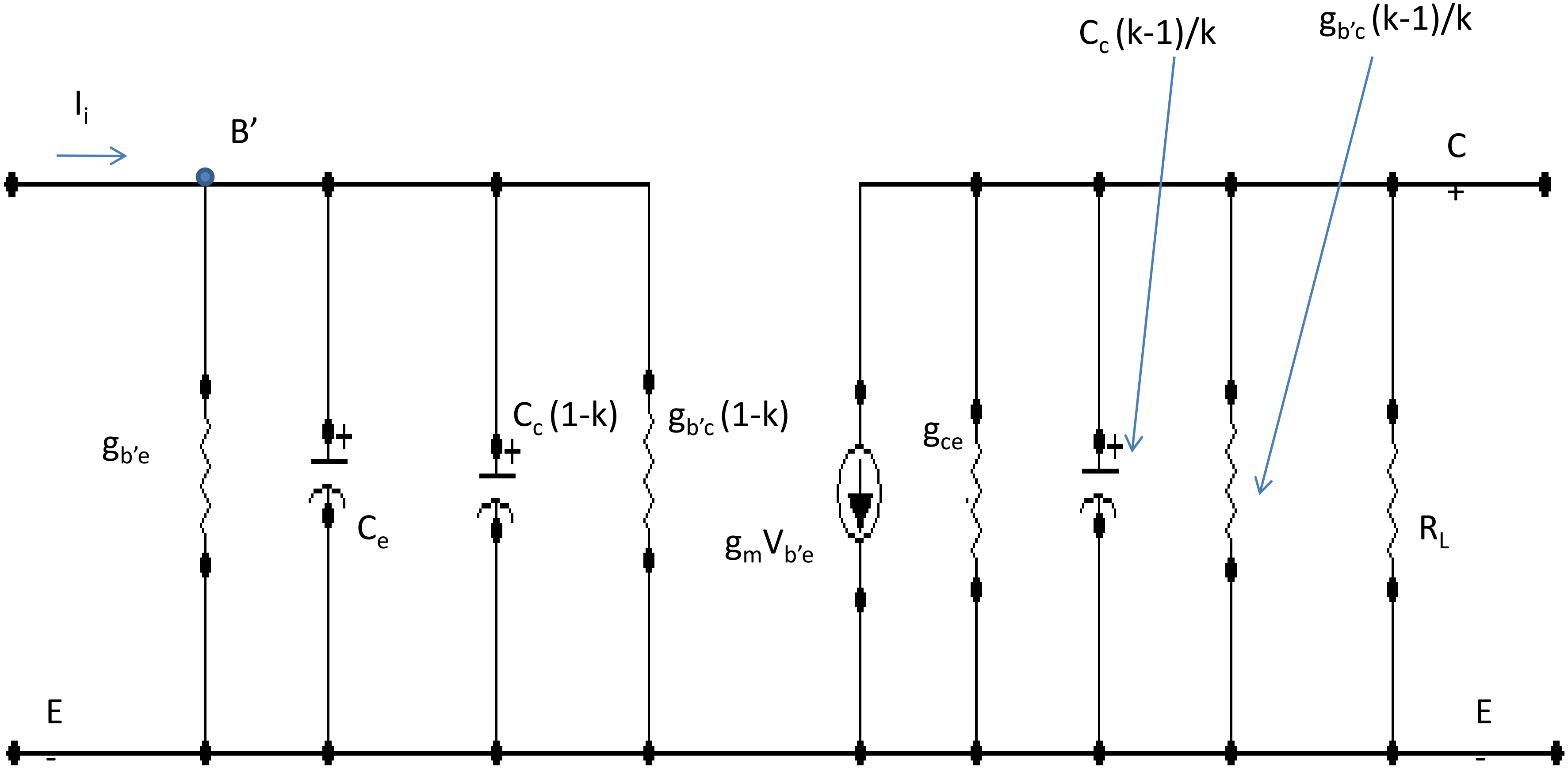
Current gain with resistive load

- For current gain, $R_L \neq 0$.

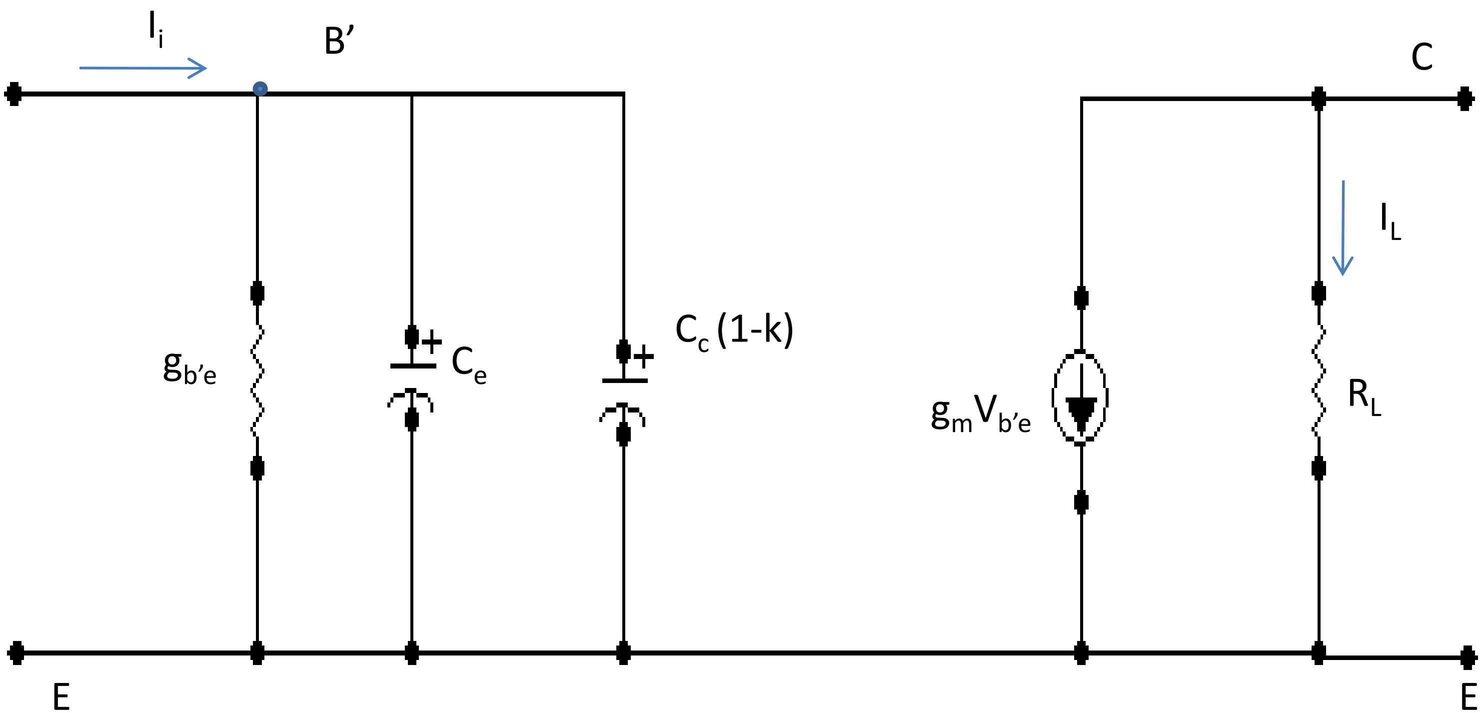
The hybrid-∏ circuit for a single transistor with a resistive load R_L



Apply miller's theorem at node B' and C.



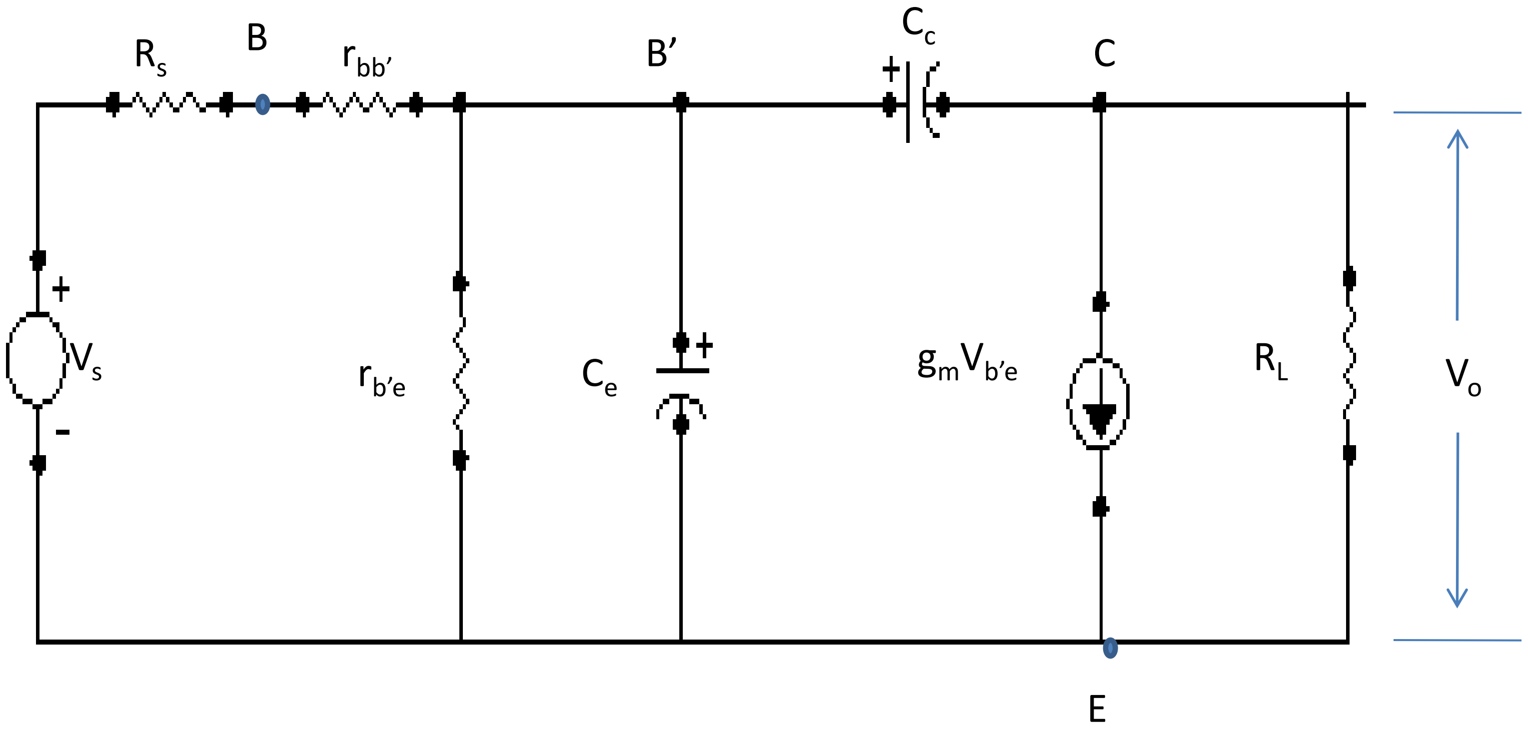
Further simplification of the equivalent circuit



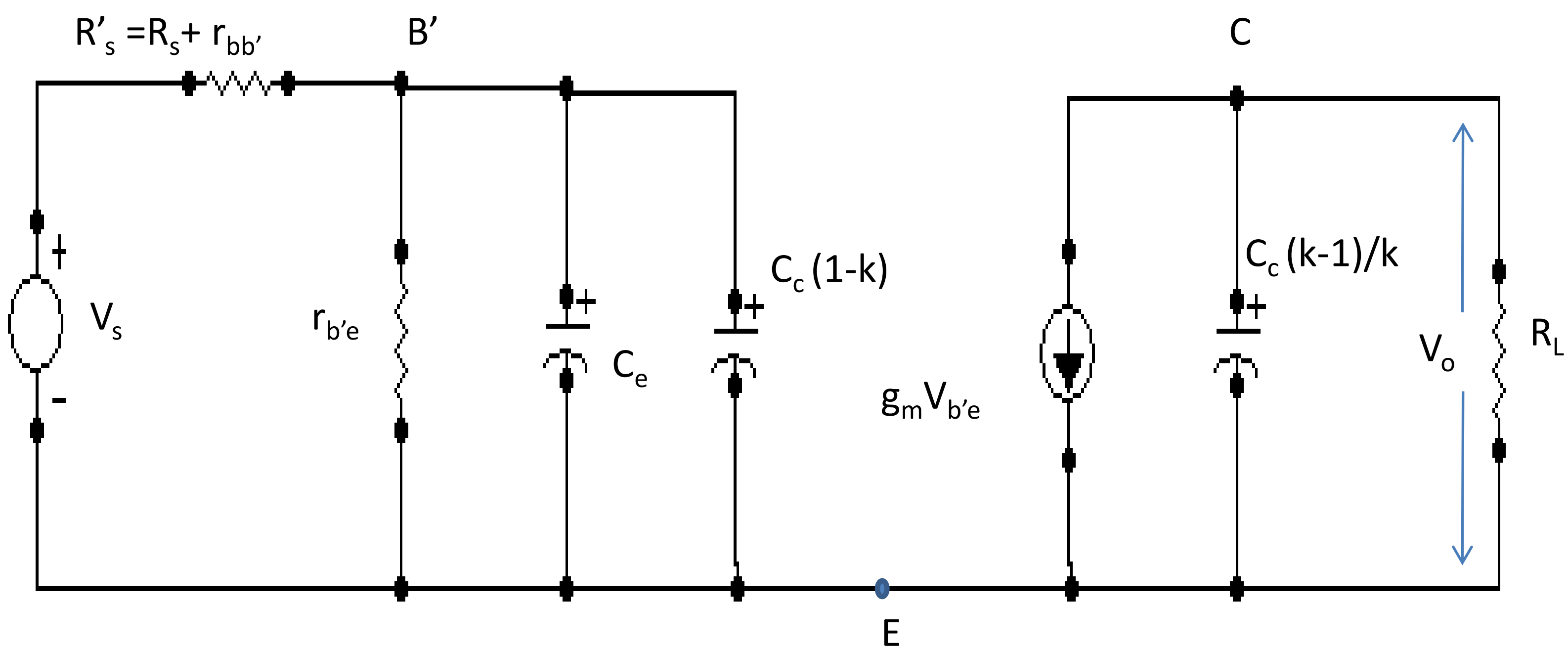
Single-stage CE transistor amplifier response

In the preceding section the Transistor is driven from an ideal current source which had infinite resistance. Now we use a voltage source V_s has a finite resistance R_s.

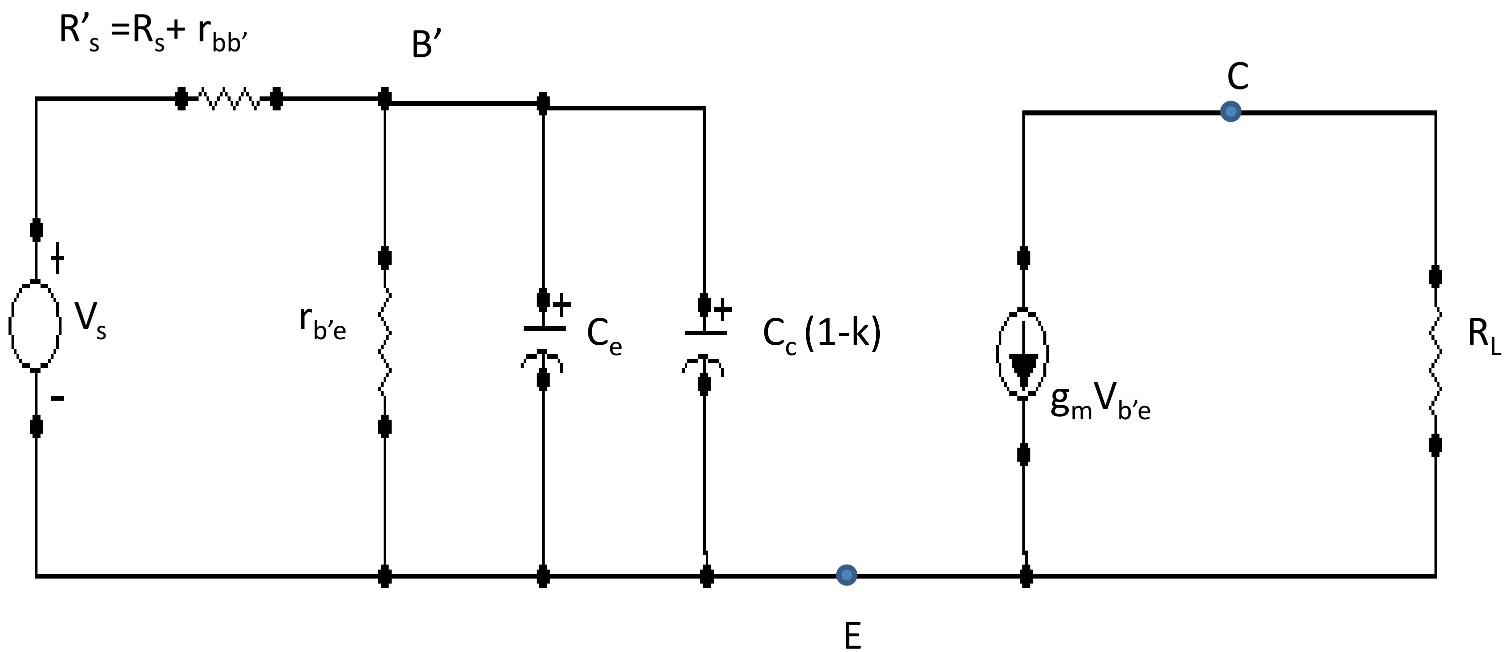
Equivalent circuit for frequency analysis of CE amplifier stage driven from a voltage source



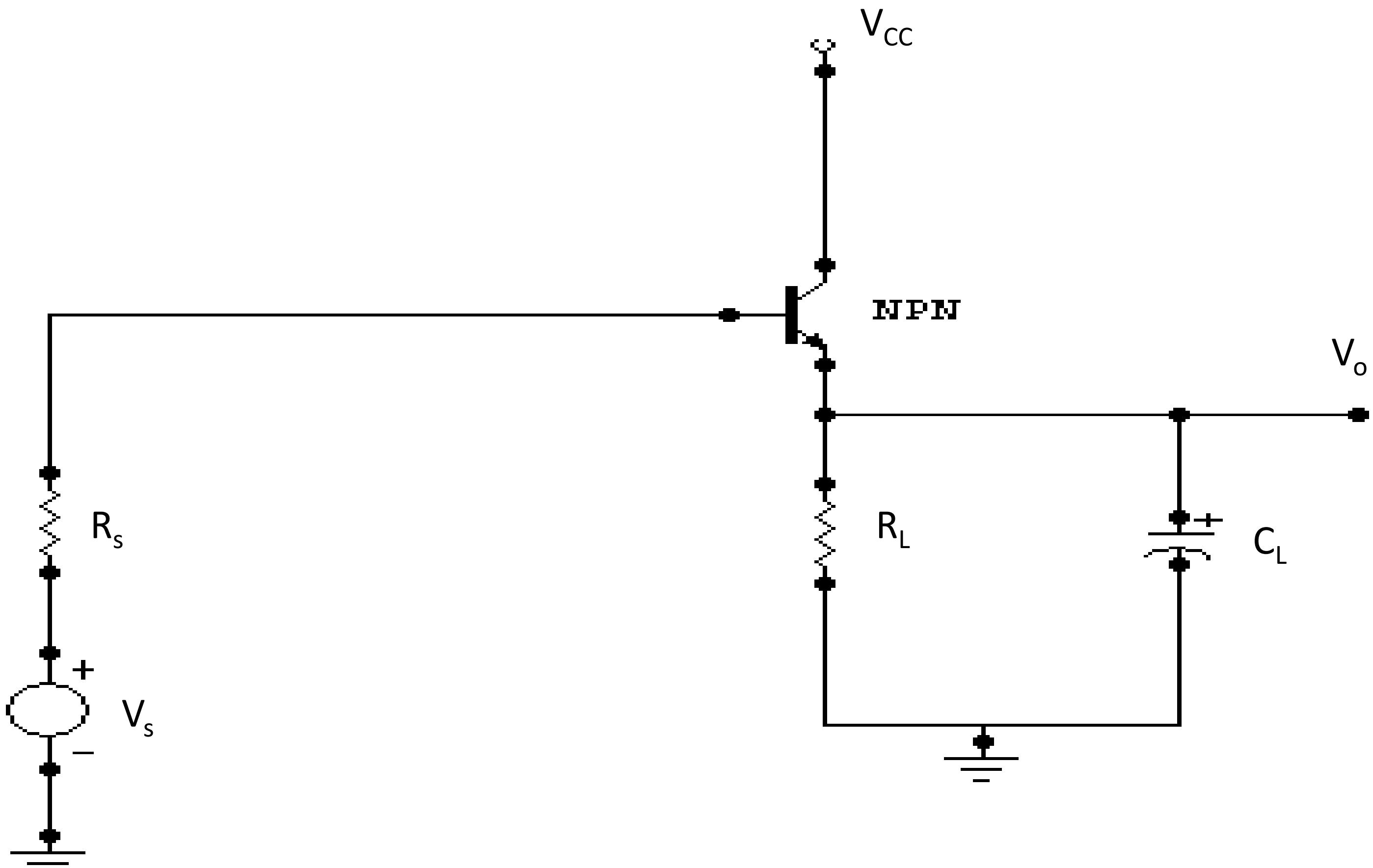
Equivalent circuit for frequency analysis of CE amplifier stage using the miller effect



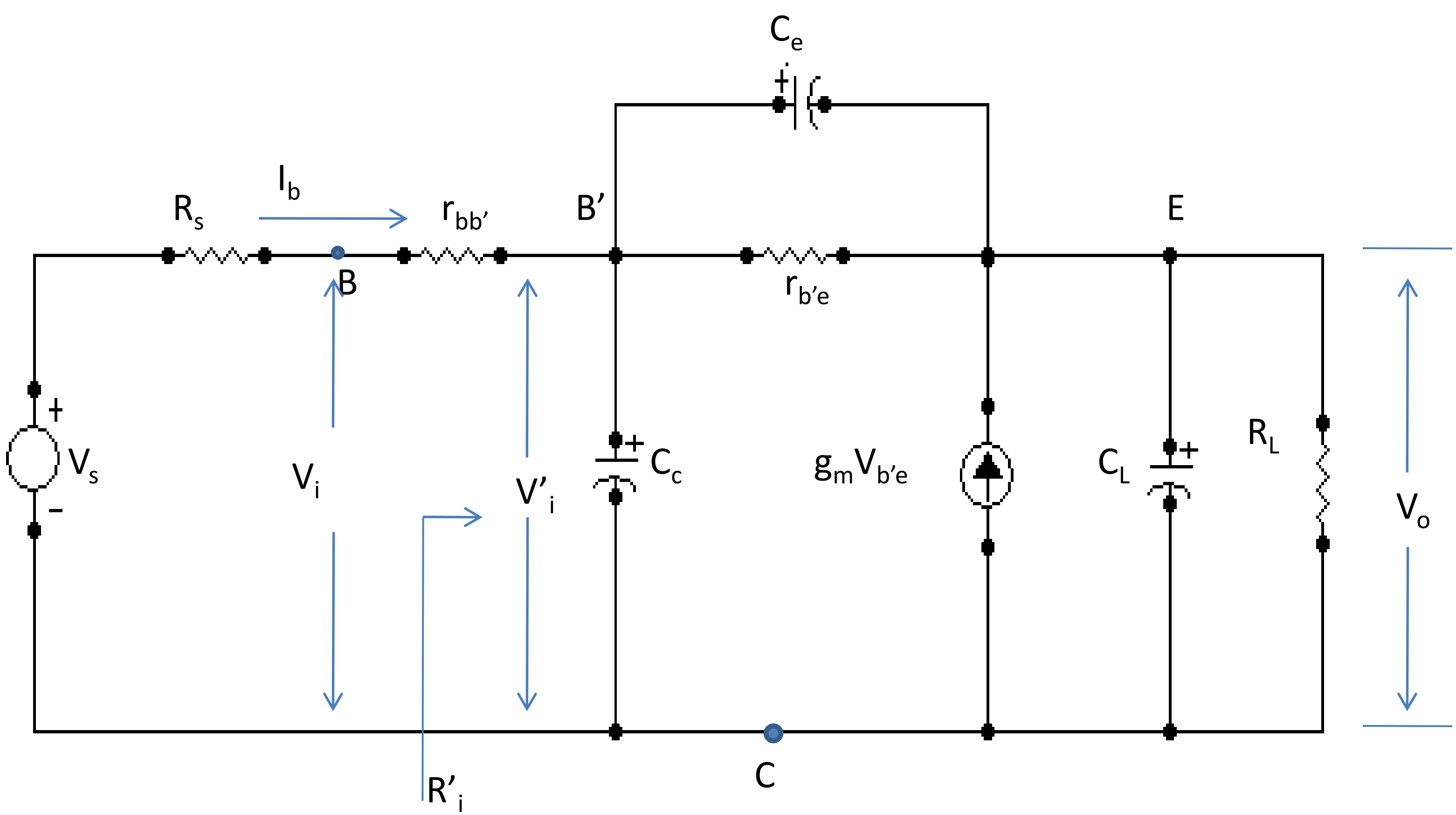
Neglecting the out put time constant



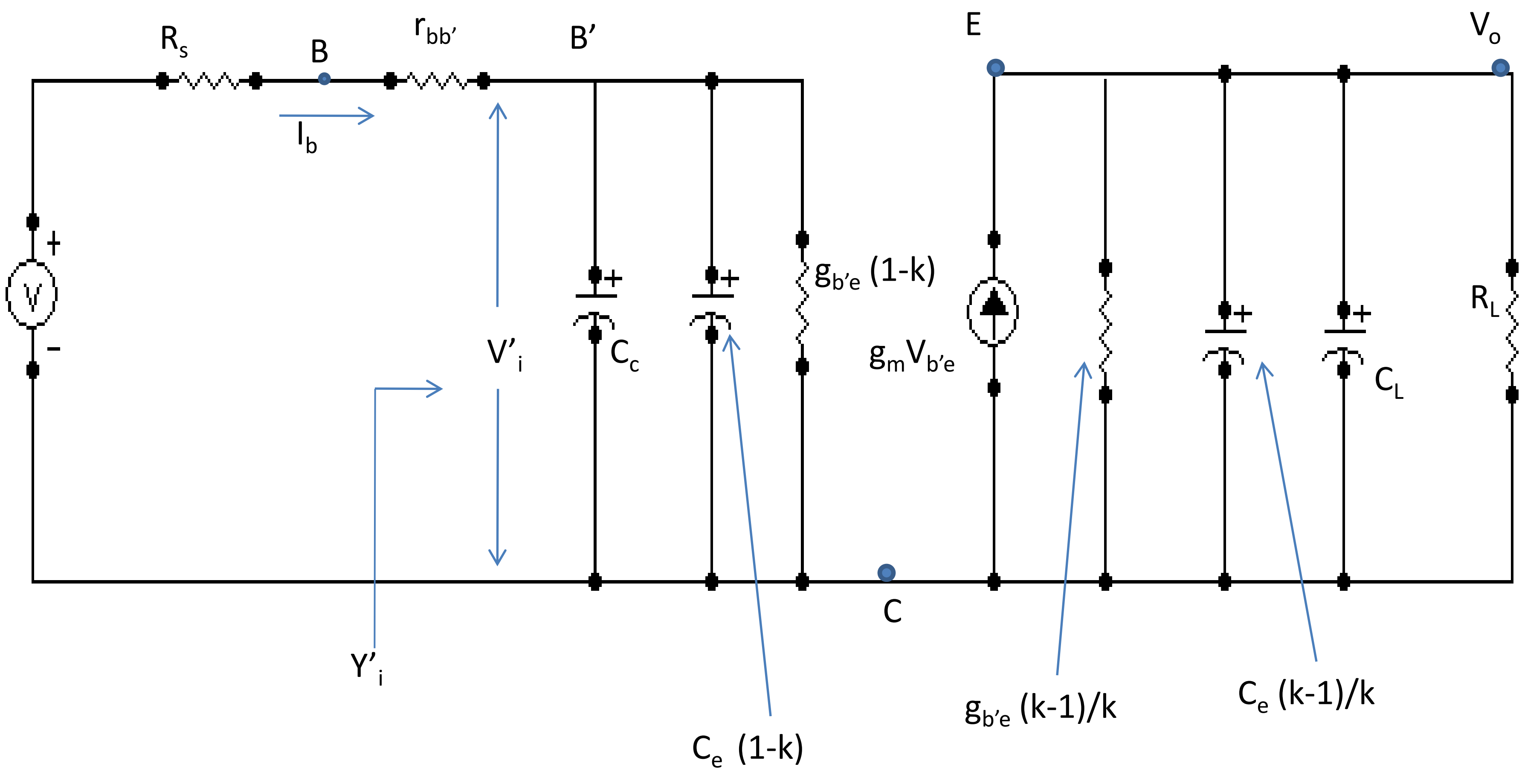
Emitter follower at high frequency



High frequency equivalent circuit of emitter follower



The equivalent circuit of emitter follower using miller's theorem



UNIT-3

MULTISTAGE AMPLIFIER

CLASSIFICATION OF AMPLIFIER

Amplifier are described in many ways, according to their

- Frequency range
- The method of operation
- The ultimate use
- The type of load
- The method of interstage coupling etc.....

CLASSIFICATION OF AMPLIFIERS BASED ON THE FREQUENCY RANGE

- DC Amplifier(from zero frequency)
- Audio Amplifier(20 Hz to 20kHz)
- Video or Pulse Amplifier(up to few mega Hz)
- Radio frequency Amplifier(few kilo Hz to Hundreds of mega Hz)
- Ultra high-frequency Amplifier(hundreds or thousands of mega Hz)

CLASSIFICATION OF AMPLIFIER BASED ON METHOD OF OPERATION

METHODS OF OPERATION ARE DEPEND UPON ***Q-POINT*** AND ***THE EXTENT OF THE CHARACHTERISTICS***

- CLASS A Amplifier
- CLASS B Amplifier
- CLASS AB Amplifier
- CLASS C Amplifier

Class A amplifier

Output Stages And Power Amplifiers

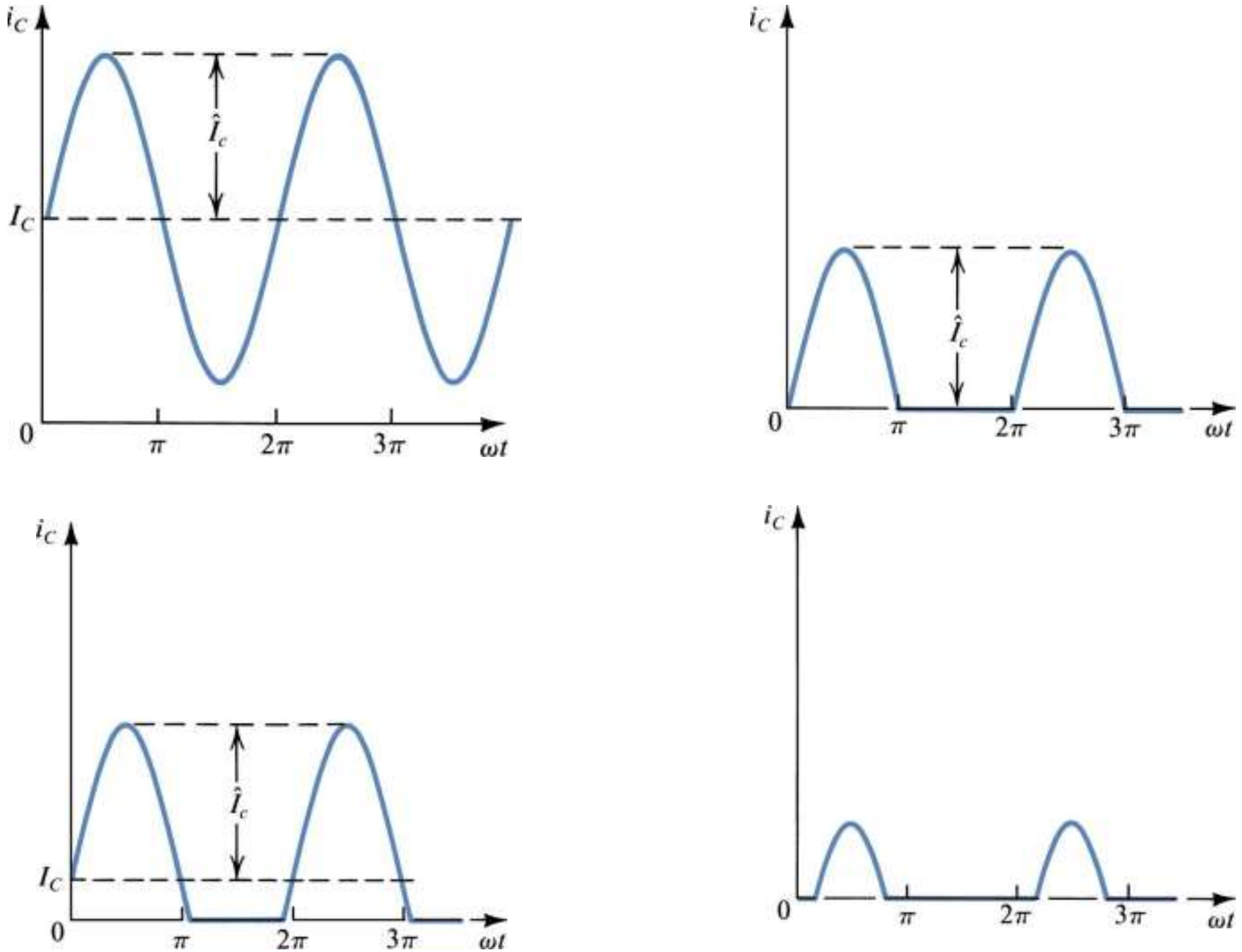
Low Output Resistance – no loss of gain

Small-Signal Not applicable

Total-Harmonic Distortion (fraction of %)

Efficiency

Temperature Requirements

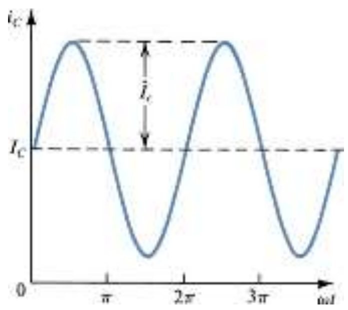


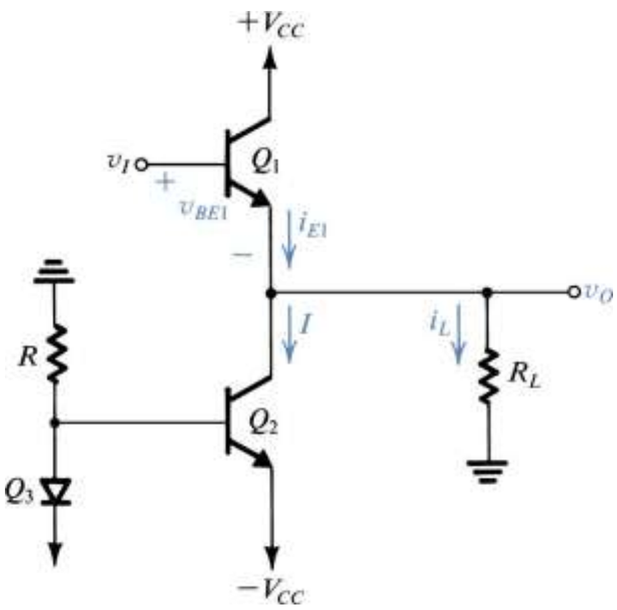
Collector current waveforms for transistors operating in (a) class A, (b) class B, (c) class AB, and (d) class C amplifier stages.

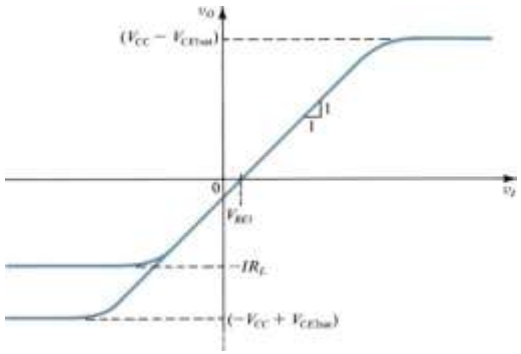
Class A

An emitter follower (Q_1) biased with a constant current I supplied by transistor Q_2.

Transfer Characteristics



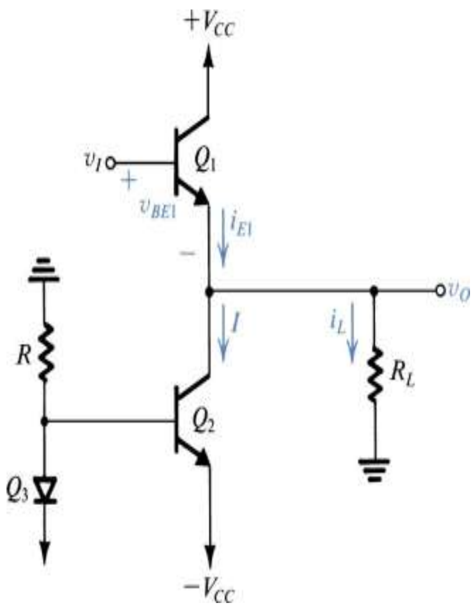




Transfer characteristic of the emitter follower. This linear characteristic is obtained by neglecting the change in v_{BE1} with i_L. The maximum positive output is determined by the saturation of Q_1. In the negative direction, the limit of the linear region is determined either by Q_1 turning off or by Q_2 saturating, depending on the values of I and R_L.

Class A

Transfer Characteristics



Crossover distortion can be eliminated by biasing the transistors at a small, non-zero current.
A bias Voltage VBB is applied between Qn and Qp.
For vi = 0, vo = 0, and a voltage VBB/2 appears across the base-emitter junction of each transistor.

$$i_N = i_P = I_Q = I_S \cdot e^{\frac{V_{BB}}{2 \cdot V_T}}$$

VBB is selected to result the required quiscent current IQ

$$v_o = v_i + \frac{V_{BB}}{2} - v_{BEN}$$

$$i_N = i_P + i_L$$

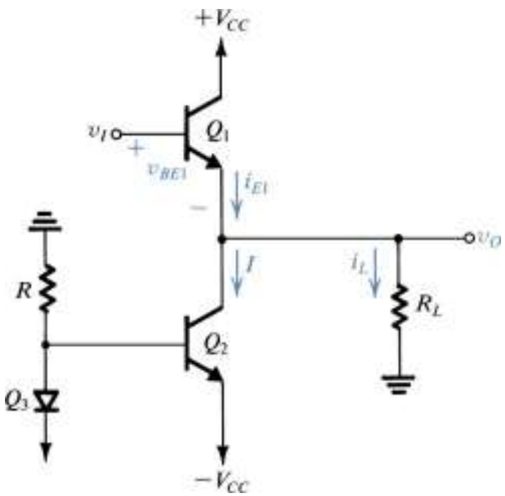
$$v_{BEN} + v_{EBP} = V_{BB} \qquad V_T \cdot \ln\left(\frac{i_N}{I_S}\right) + V_T \cdot \ln\left(\frac{i_P}{I_S}\right) = 2 \cdot V_T \cdot \ln\left(\frac{i_Q}{I_S}\right)$$

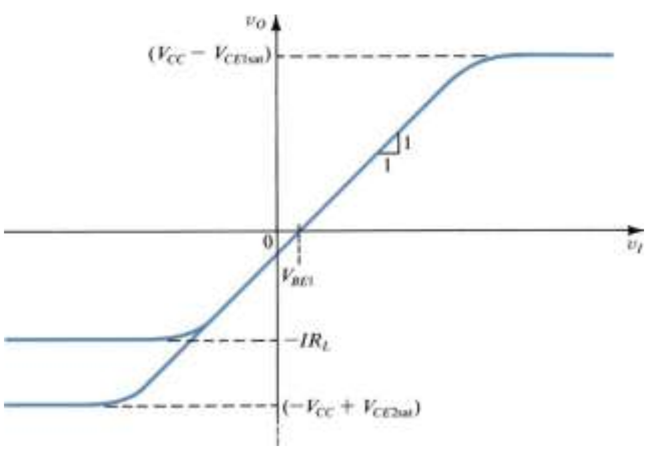
$$i_N^{\ 2} = I_Q^{\ 2}$$

$$i_N^{\ 2} - i_L \cdot i_N - I_Q^{\ 2} = 0$$

Class A

Transfer Characteristics





From figure 9.3 we can see that

$$v_{omax} = V_{CC} - V_{CE1sat}$$

In the negative direction, the limite of the linear region is determined either by Q1 turning off

$$v_{Omin} = -I \cdot R_L$$

or by Q2 saturating

$$v_{Omin} = -V_{CC} + V_{CE2sat}$$

Depending on the values of I and RL. The absolutely lowest output voltage is that given by the previous equation and is achieved provided that the bias current I is greater than the magnitude of the corresponding load current

$$I \geq \frac{\left|-V_{CC} + V_{CE2sat}\right|}{R_L}$$

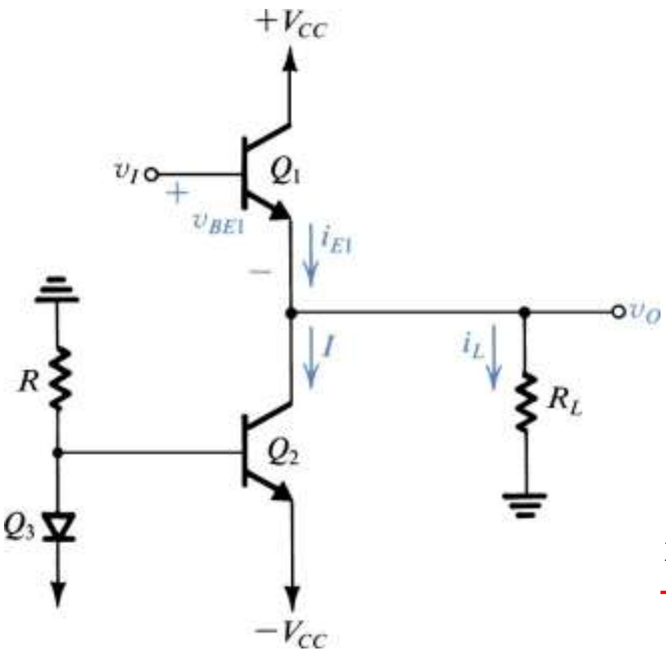
Class A

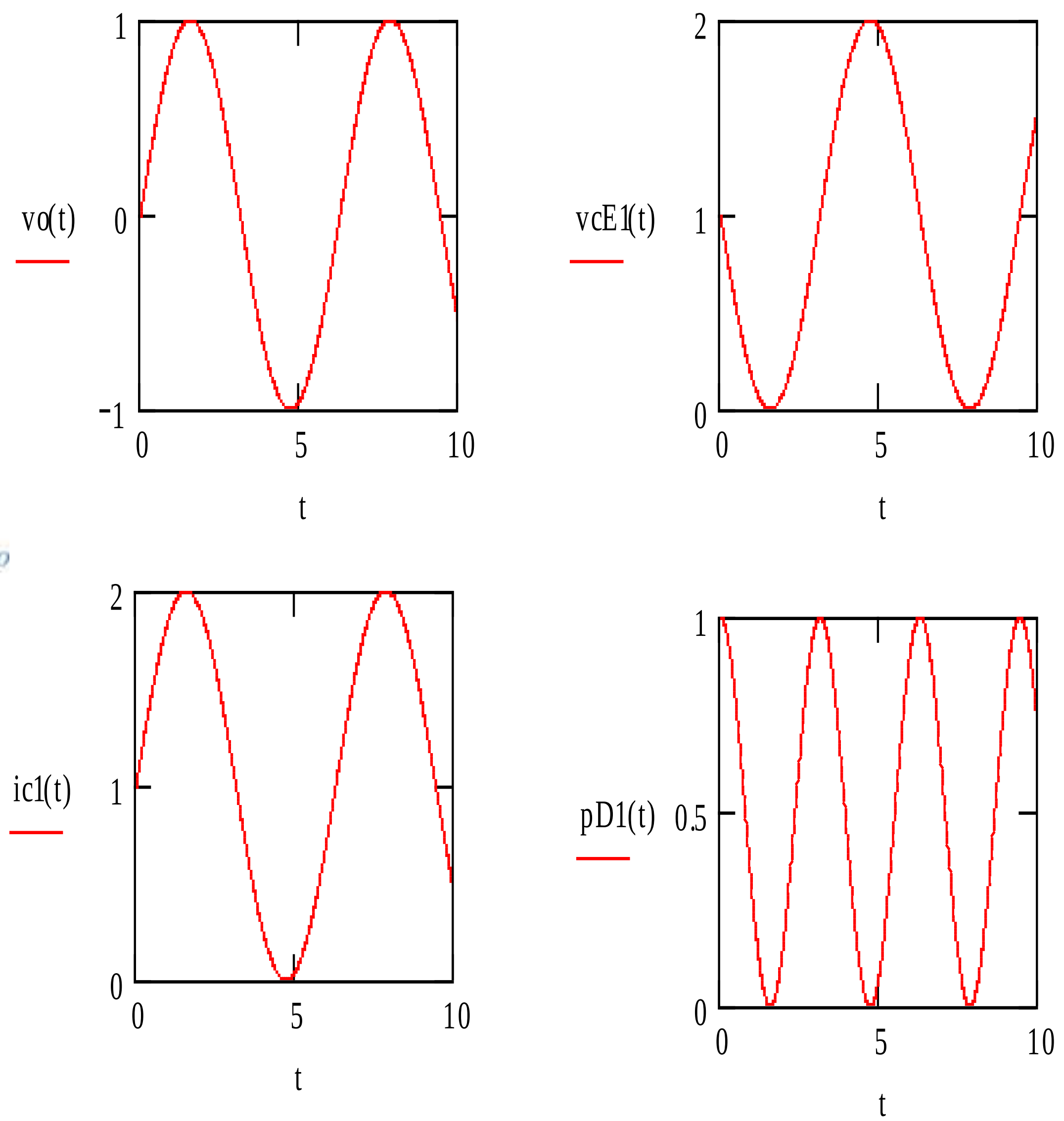
Transfer Characteristics

Exercises D9.1 and D9.2

Class A

Signal Waveforms





Class A

Power Dissipation

$$P = V_{CC} \cdot I$$

Largest Power Dissipation When vo = 0
Q1 must be able to withsatnd a continuous dissipation of VCC*I

The power dissipation of Q1 depends on the value of RL.
If RL is infinite, iC1 = I and the dissipation in Q1 depends on vo.
Maximum power dissipation will occur when vo = -VCC since vCE1 will be 2VCC
pD1 = 2VCC*I. This condition would not normally persist for a prolonged interva
the design need not be that conservative. The average pD1 = VCC*I

When RL is zero a positive voltage would result in a theoretically infinite current
practical value) would flow through Q1. Short-circuit protection is necessary.

Class A

Power Conversion Efficiency

$$\eta = \frac{\text{load\_power}(P_L)}{\text{supply\_power}(P_S)}$$

$$P_L = \frac{1}{2} \cdot \frac{V_o^{\,2}}{R_L}$$

V_o average voltag

$$P_S = 2 \cdot V_{CC} \cdot I$$

$$\eta = \frac{1}{4} \cdot \frac{V_o^{\,2}}{I \cdot R_L \cdot V_{CC}} = \frac{1}{4} \cdot \left(\frac{V_o}{I \cdot R_L}\right) \cdot \left(\frac{V_o}{V_{CC}}\right)$$

$V_o \le V_{CC}$ $\qquad$ $V_o \le I \cdot R_L$

maximum efficiency is obtained when

$$V_o = V_{CC} = I \cdot R_L$$

Class A

Exercise 9.4

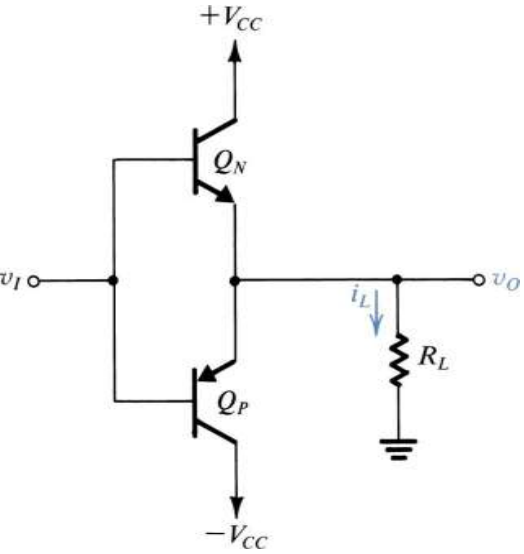
$Vopeak := 8$ $\quad I := 100 \cdot 10^{-3}$ $\quad R_L := 100$ $\quad V_{CC} := 10$

$$P_L := \frac{\left(\frac{Vopeak}{\sqrt{2}}\right)^2}{100} \qquad P_L = 0.32$$

$$Pplus := V_{CC} \cdot I \qquad Pplus = 1$$

$$Pminus := V_{CC} \cdot I \qquad Pminus = 1 \qquad P_S := Pplus + Pminus$$

$$\eta := \frac{P_L}{P_S} \qquad \eta = 0.16$$



Biasing the Class B Output

- No DC current is used to bias this configuration.
- Activated when the input voltage is greater than the Vbe for the transistors.
- npn Transistor operates when positive, pnp when negative.
- At a zero input voltage, we get no output voltage.

Class A

Power Conversion Efficiency

CLASS A

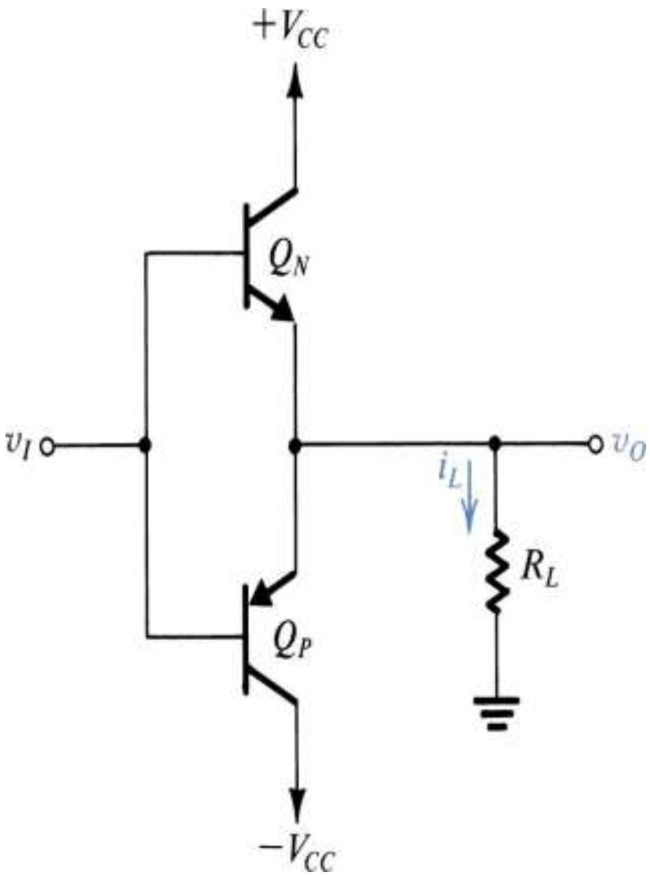
Many class A amplifiers use the same transistor(s) for both halves of the audio waveform. In this configuration, the output transistor(s) always has current flowing through it, even if it has no audio signal (the output transistors never 'turn off'). The current flowing through it is D.C.

A pure class 'A' amplifier is very inefficient and generally runs very hot even when there is no audio output. The current flowing through the output transistor(s) (with no audio signal) may be as much as the current which will be driven through the speaker load at FULL audio output power. Many people believe class 'A' amps to sound better than other configurations (and this may have been true at some point in time) but a well designed amplifier won't have any 'sound' and even the most critical 'ear' would be hard-pressed to tell one design from another.

NOTE: Some class A amplifiers use complimentary (separate transistors for positive and negative halves of the waveform) transistors for their output stage.

Class B

Circuit Operation



Class B output stage.

CLASS 'B'

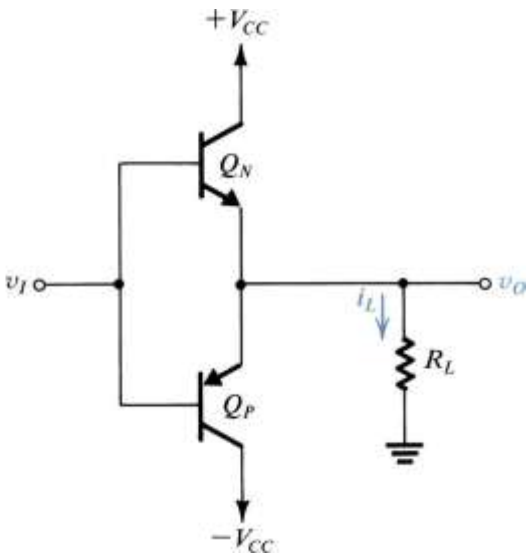
A class 'B' amplifier uses complimentary transistors for each half of the waveform.

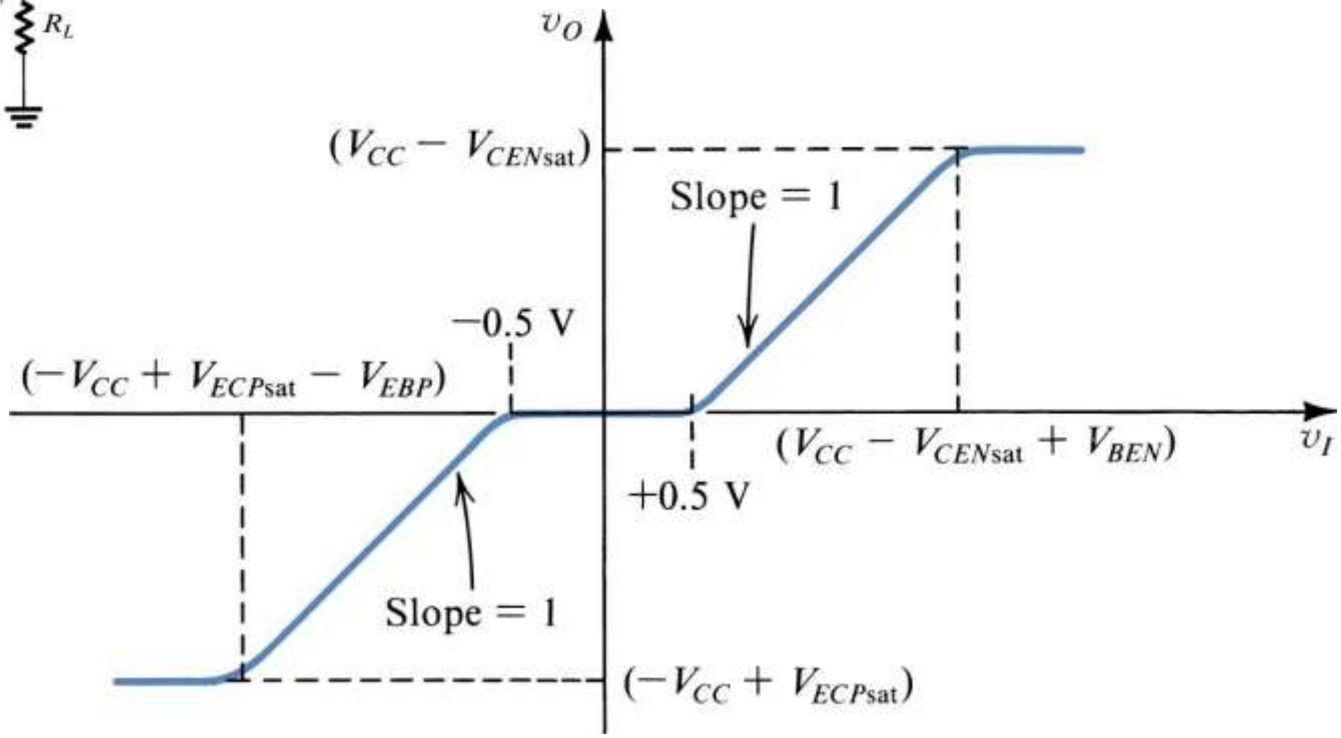
A true class 'B' amplifier is NOT generally used for audio. In a class 'B' amplifier, there is a small part of the waveform which will be distorted. You should remember that it takes approximately .6 volts (measured from base to emitter) to get a bipolar transistor to start conducting. In a pure class 'B' amplifier, the output transistors are not "biased" to an 'on' state of operation. This means that the the part of the waveform which falls within this .6 volt window will not be reproduced accurately.

The output transistors for each half of the waveform (positive and negative) will each have a .6 volt area in which they will not be conducting. The distorted part of the waveform is called 'crossover' or 'notch' distortion. Remember that distortion is any unwanted variation in a signal (compared to the original signal). The diagram below shows what crossover distortion looks like.

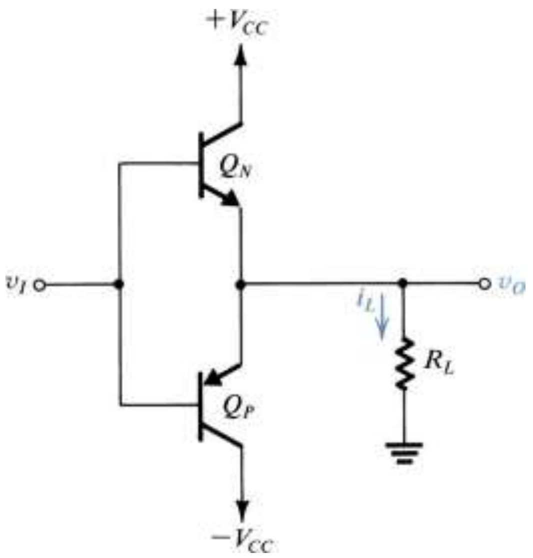
Class B

Circuit Operation





Transfer characteristic for the class B output stage in Fig. 9.5.



Operation

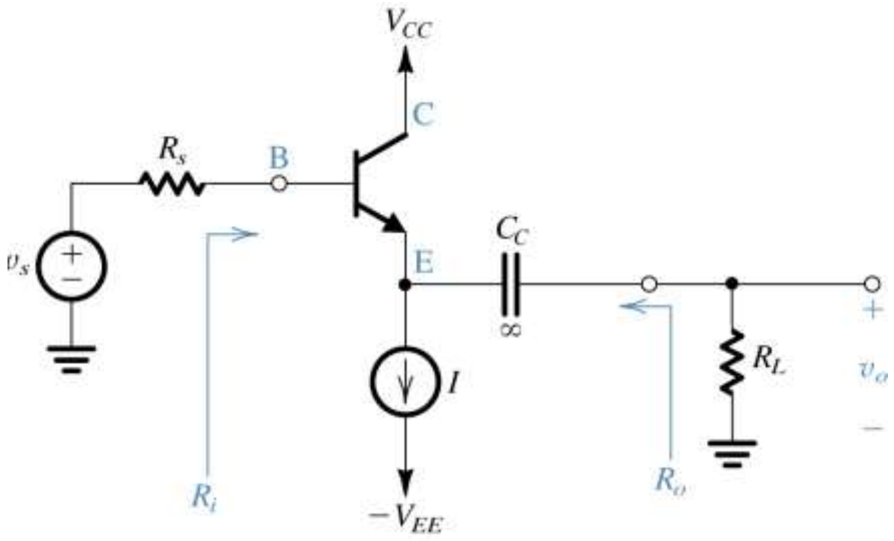
When the input voltage rises to be large enough to overcome the Vbe, it will begin to cause an output voltage to appear. This occurs because Qn begins to act like an emitter follower and Qp shuts off. The input will be followed on the emitter until the transistor reaches saturation. The maximum input voltage is equal to the following:

$$v_{imax} = V_{CC} - V_{CENsat}$$

The same thing will begin to happen if the input voltage is negative by more than the Veb of the transistor. This causes the Qp to act like an emitter follower and Qn turns off. This will continue to behave this way until saturation occurs at a minimum input voltage of:

$$v_{imin} = -V_{cc} + V_{ECPsat}$$

Emitter Follower Configuration (Chapter 4)



(a)

$$\frac{v_b}{v_s} = \frac{(\beta + 1)(r_e + par(R_L, r_o))}{R_S + (\beta + 1)(r_e + par(R_L, r_o))}$$

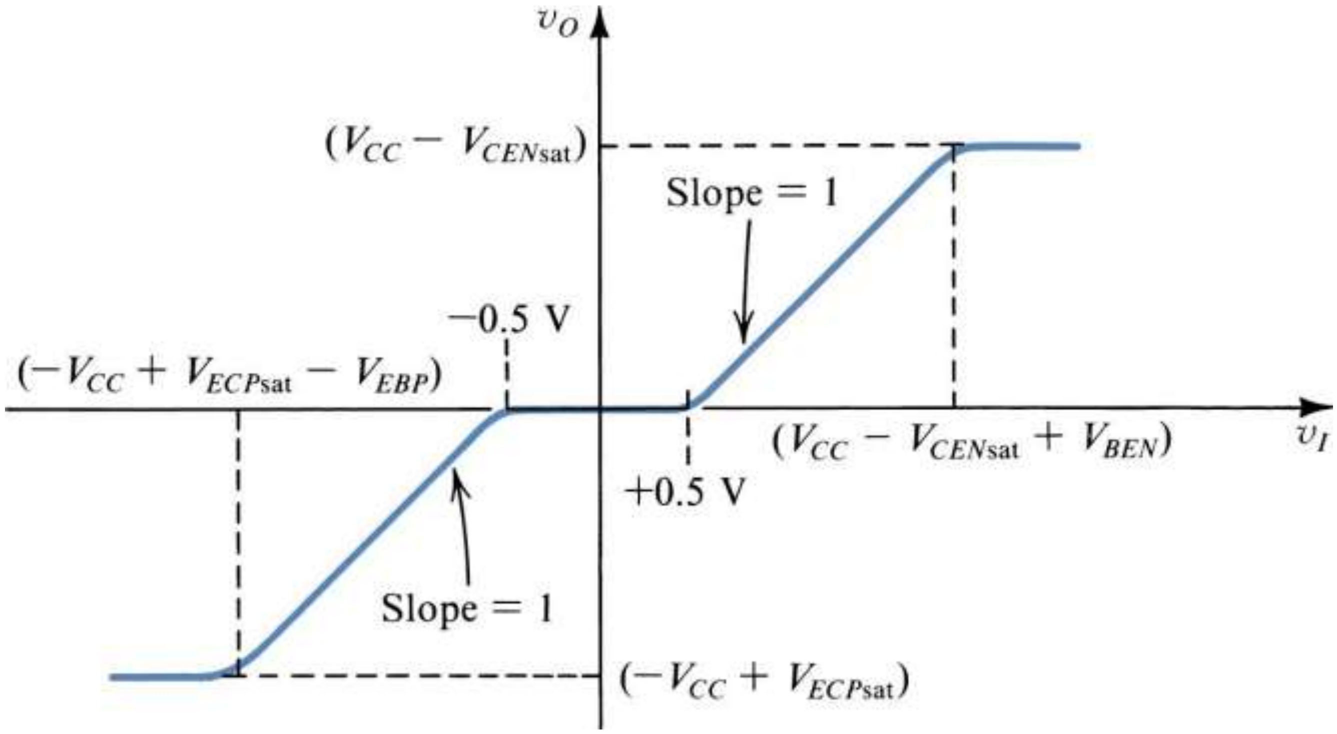
$$\frac{v_o}{v_b} = \frac{par(r_o, R_L)}{r_e + par(r_o, R_L)}$$

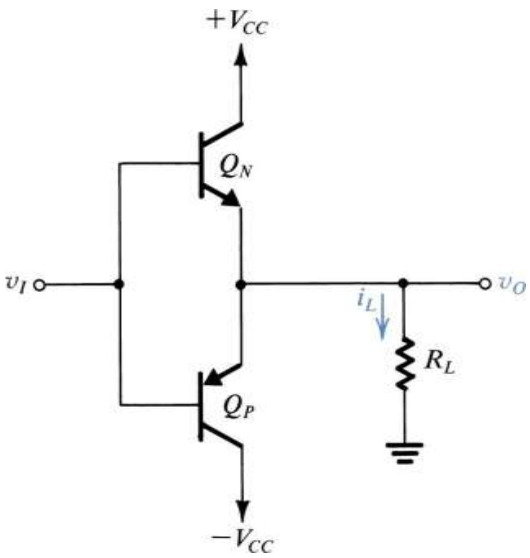
Rs will be small for most configurations, so the vb/vs will be a little less than unity. The same is true for re, so vo/vb will be a little less than unity making our vo/vs a little less than unity.

Characteristics of the Emitter Follower:

- High Input Resistance
- Low Output Resistance
- Near Unity Gain

Transfer Characteristic





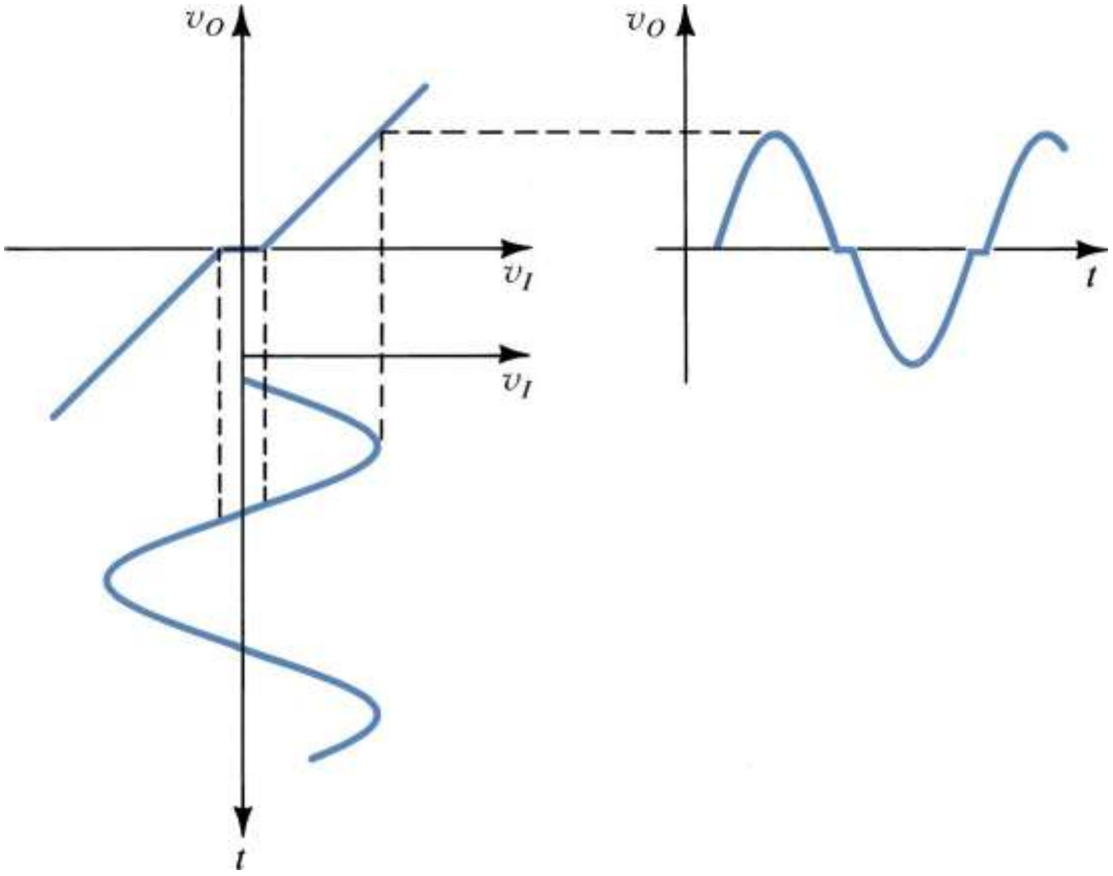
Push-Pull Nature of Class B

- Push: The npn transistor will push the current to ground when the input is positive.
- Pull: The pnp transistor will pull the current from the ground when the input is negative.

Crossover Distortion

The Crossover Distortion is due to the dead band of input voltages from -.5V to .5V. This causes the Class B output stage to be a bad audio amplifier. For large input signals, the crossover distortion is limited, but at small input signals, it is most pronounced.

Graph of Crossover Distortion



Illustrating how the dead band in the class B transfer characteristic results in crossover distortion.

Power Efficiency

Load Power:

$$P_L = \frac{1}{2} \cdot \frac{V_{op}^2}{R_L}$$

Since each transistor is only conducting for one-half of the time, the power drawn from each source will be the same.

$$P_s = \frac{1}{\pi} \cdot \frac{V_{op}}{R_L} \cdot V_{CC}$$

$$\eta = \frac{P_L}{2 \cdot P_s} = \frac{\frac{1}{2} \cdot \frac{V_{op}^2}{R_L}}{2 \frac{1}{\pi} \cdot \frac{V_{op}}{R_L} \cdot V_{CC}}$$

This efficiency will be at a max when Vop is at a max. Since Vop cannot exceed Vcc, the maximum efficiency will occur at pi/4.

$$\eta = \frac{\pi}{4} \cdot \frac{V_{op}}{V_{CC}}$$

$$\eta_{max} = \frac{\pi}{4}$$

This will be approximately 78.5%, much greater than the 25% for Class A.

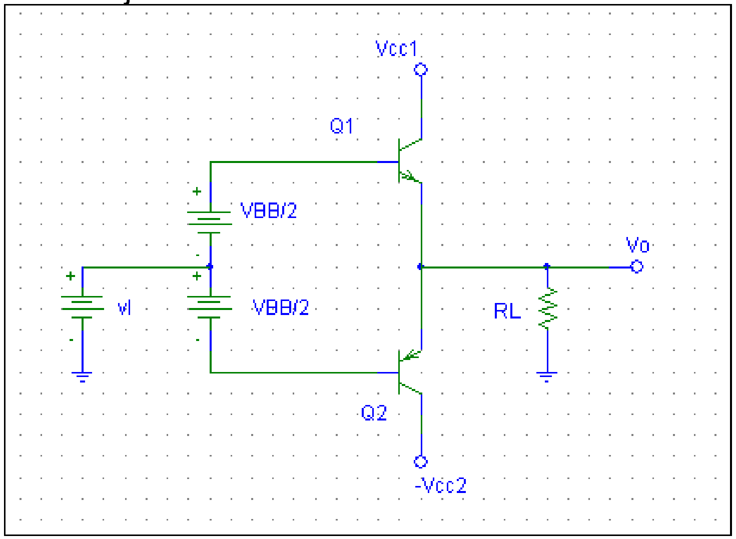
Class AB

Circuit Operation

Crossover distortion can be eliminated by biasing the transistors at a small, non-zero current.

A bias Voltage VBB is applied between Qn and Qp.

For vi = 0, vo = 0, and a voltage VBB/2 appears across the base-emitter junction of each transistor.



$$i_N = i_P = I_Q = I_S \cdot e^{\frac{V_{BB}}{2 \cdot V_T}}$$

VBB is selected to result the required quiscent current IQ

$$v_o = v_i + \frac{V_{BB}}{2} - v_{BEN}$$

$$i_N = i_P + i_L$$

$$v_{BEN} + v_{EBP} = V_{BB}$$

$$V_T \cdot \ln\left(\frac{i_N}{I_S}\right) + V_T \cdot \ln\left(\frac{i_P}{I_S}\right) = 2 \cdot V_T \cdot \ln\left(\frac{i_Q}{I_S}\right)$$

$$i_N^{2} = I_Q^{2}$$

$$i_N^{2} - i_L \cdot i_N - I_Q^{2} = 0$$

Class AB

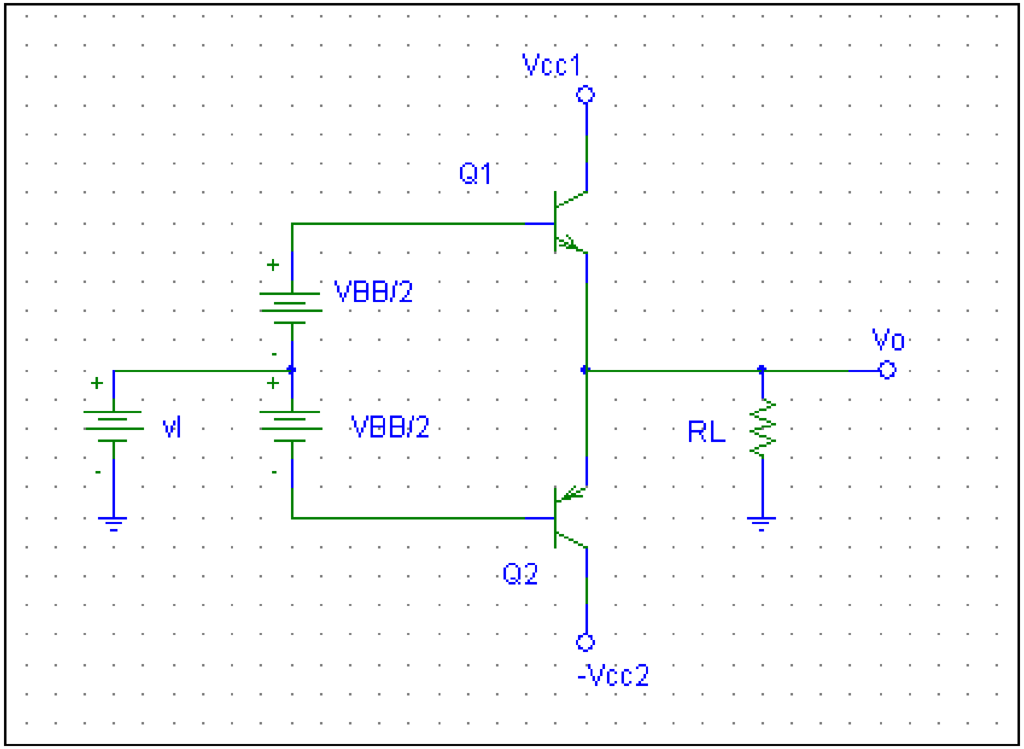
Output Resistance

Class AB

Exercise 9.6

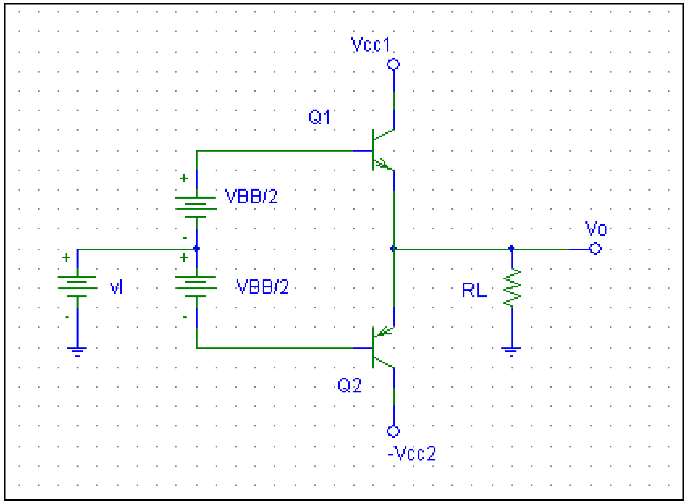
Calvin College - ENGR 332
Class AB Output Stage Amplifier

Consider the class AB circuit (illustrated below) with Vcc=15 V, IQ=2 mA, RL=100 ohms.
Determine VBB. Determine the values of iL, iN, iP, vBEN, vEBP, vI, vO/vI, Rout, and vo/vi v
vO for vO varying from -10 to 10V.
Note that vO/vI is the large signal voltage gain and vo/vi is the incremental gain obtained a
RL/(RL+Rout). The incremental gain is equal to the slope of the transfer curve.
Assume QN and QP to be matched, with IS=10E-13.



Class AB

Exercise 9.6



under quiescent conditions iN=iP=IQ   vO=vI=0

Solving for VBB

$VBB := 1$ $IS := 10^{-13}$ $VT := 0.025$ $IQ := 2 \cdot 10^{-3}$ $RL := 10($

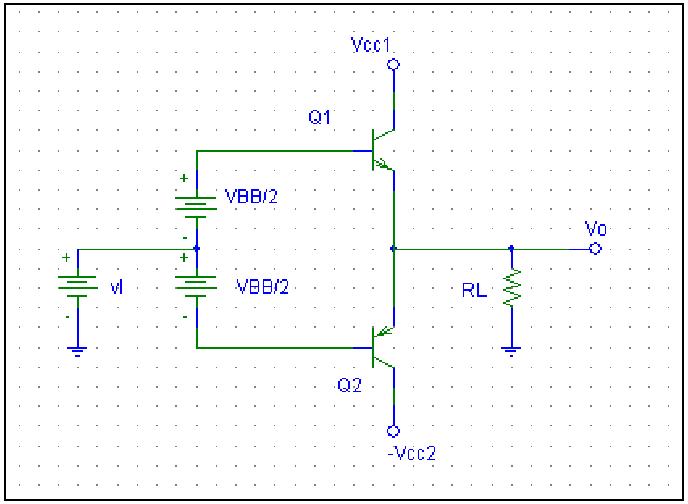
Given

$$IQ = IS \cdot e^{\frac{\frac{|VBB|}{2}}{VT}}$$

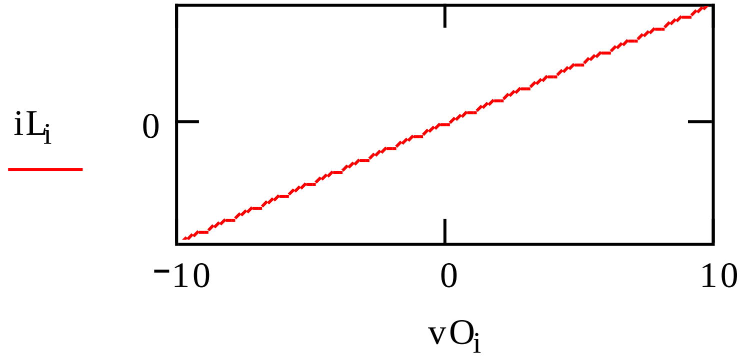
$VBB := Find(VBB)$ $i := 0..100$ $VBB = 1.186$

Class AB

Exercise 9.6

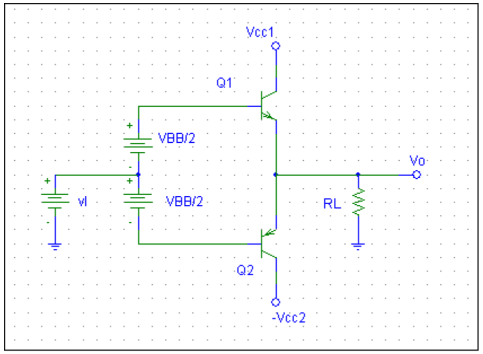


$$vO_i := -10 + \frac{i}{5} \qquad iL_i := \frac{vO_i}{RL}$$



Class AB

Exercise 9.6



Solving for iN

initial guesses $iN := 0.02$ $iLD := 0.02$ $IQ := 0.002$

Given

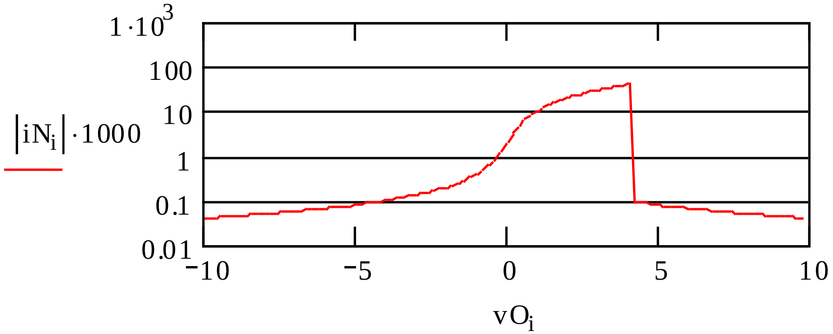
$$iN^2 - iLD \cdot iN - IQ^2 = 0$$

$$iNN(IQ, iLD) := \text{Find}(iN)$$

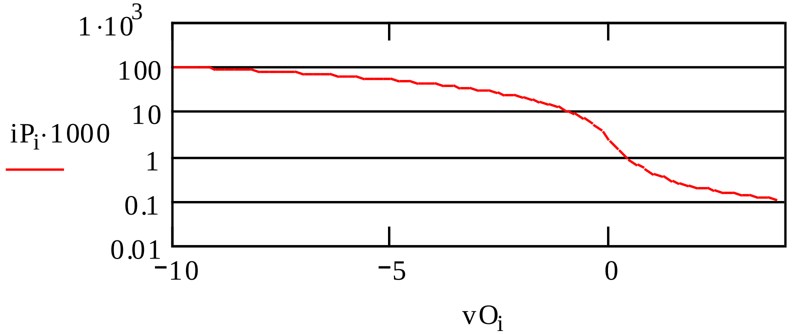
$i := 0..100$ $IQ_i := 0.002$ $iLD_i := iL_i$

$$iN_i := iNN\left(IQ_i, iL_i\right)$$

$$iN_{10} = 4.997 \times 10^{-5}$$



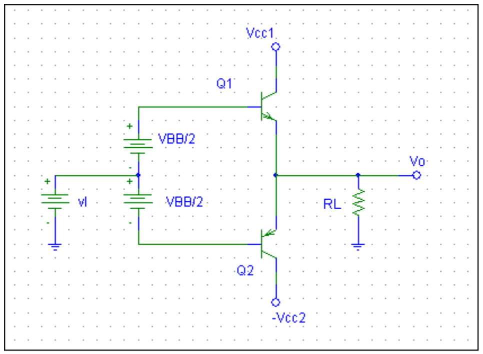
$$iP_i := iN_i - iLD_i$$



$\left(iN_i\right)$

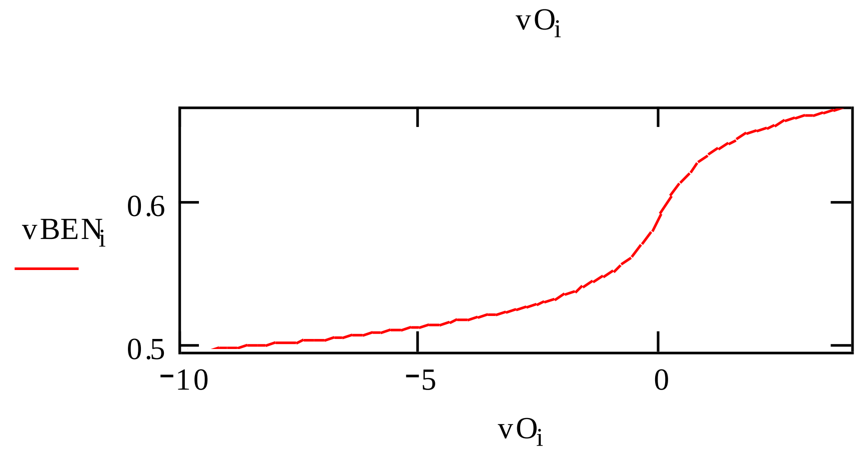
Class AB

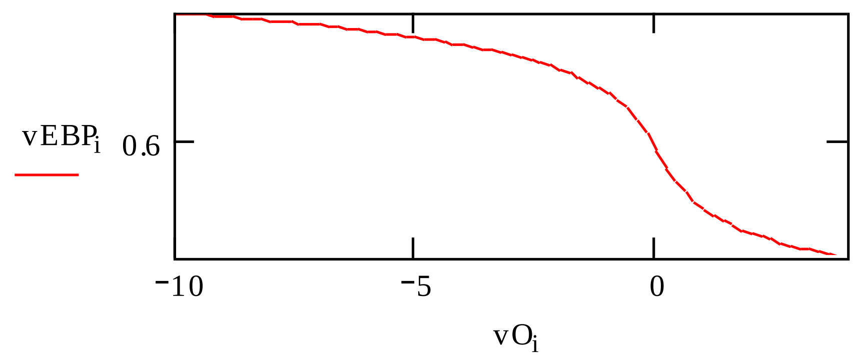
Exercise 9.6



$$vBEN_i := VT \cdot \ln\left(\frac{iN_i}{IS}\right)$$

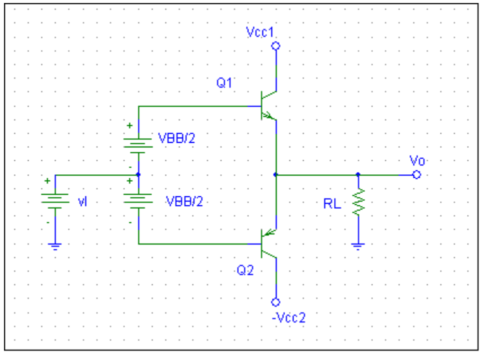
$$vEBP_i := VT \cdot \ln\left(\frac{iP_i}{IS}\right)$$



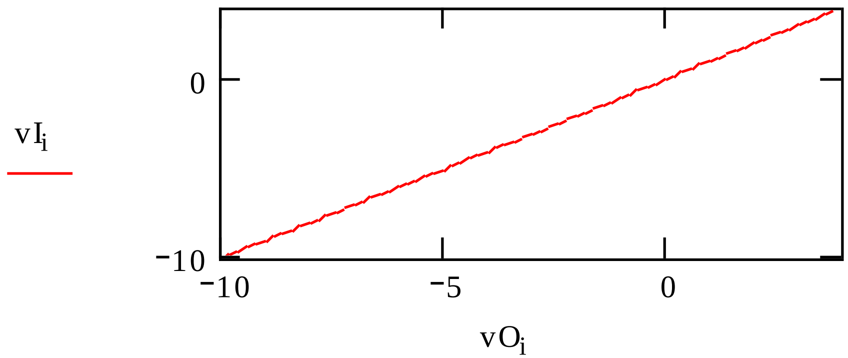


Class AB

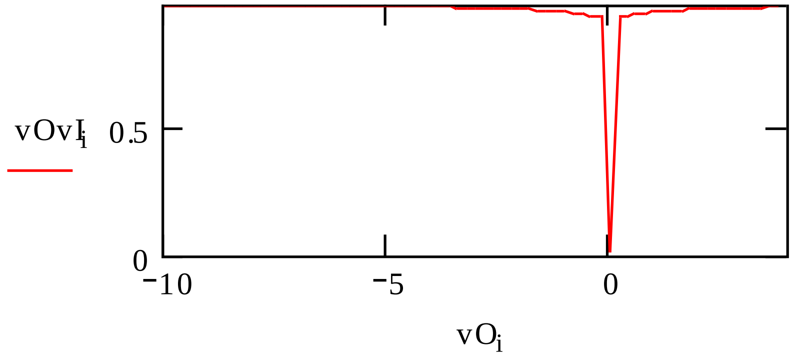
Exercise 9.6



$$vI_i := vO_i + vBEN_i - \frac{VBB}{2}$$

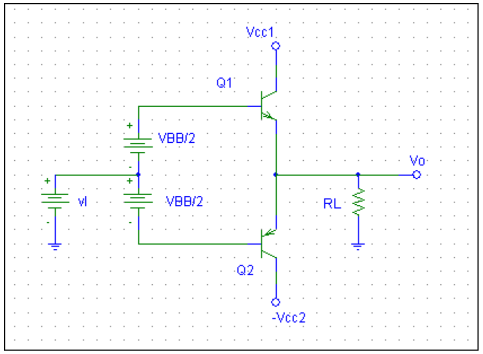


$$vOvI_i := \frac{vO_i}{vI_i}$$

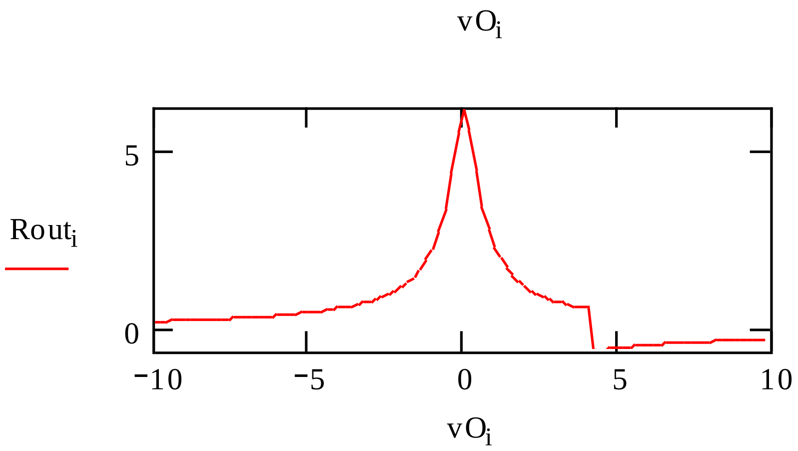


Class AB

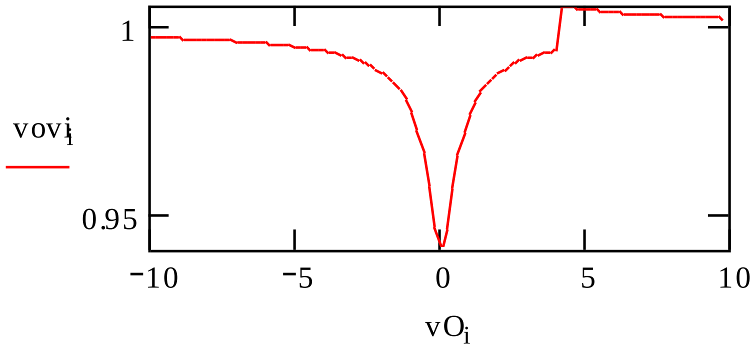
Exercise 9.6



$$Rout_i := \frac{VT}{iP_i + iN_i}$$



$$vovi_i := \frac{RL}{RL + Rout_i}$$



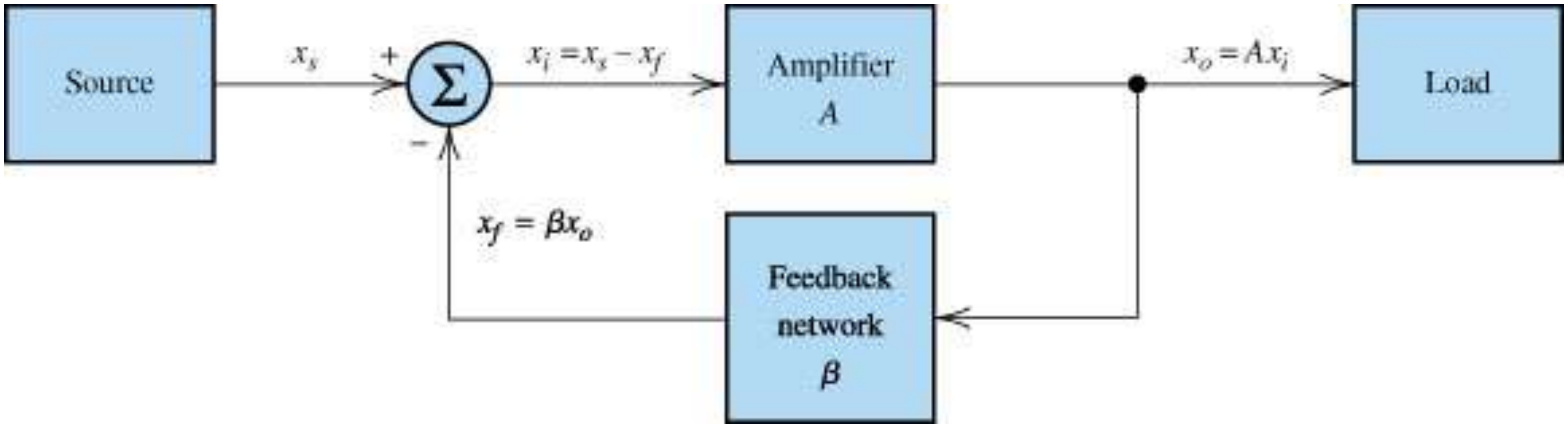
UNIT 4

FEEDBACK AMPLIFIER

Feedback of Amplifier Circuits I

- Feedback is to return part of the output to the input for a circuit/system (amplifiers in our context)
- Feedback is very useful in Control Theory and Systems and is well researched
- Amplifier circuit can have negative feedback and positive feedback. Negative feedback returns part of the output to oppose the input, whereas in positive feedback the feedback signal aids the input signal.
- Both negative feedback and positive feedback are used in amplifier circuits
- Negative feedback can reduce the gain of the amplifier, but it has many advantages, such as stabilization of gain, reduction of nonlinear distortion and noise, control of input and output impedances, and extension of bandwidth

Concept of amplifier feedback



$$A_f = \frac{x_o}{x_s} = \frac{A}{1 + A\beta}, \text{if } A_f < A, \text{then negative feedback}$$

A_f : the closed – loop gain of the amplifier

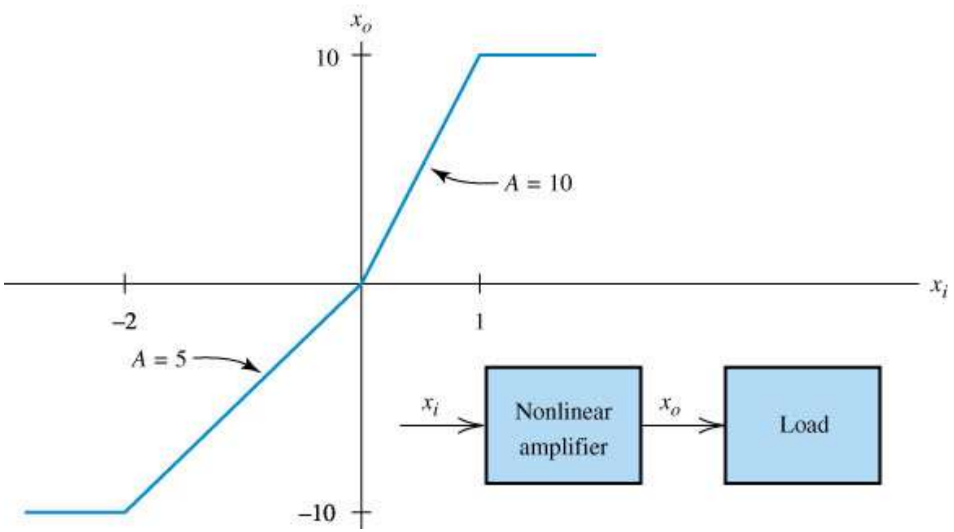
A : the open – loop gain of the amplifier

β : feedback coefficient

$A\beta$: loop gain

- If $A\beta >> 1$, then $A_f \approx 1/\beta$. Thus, the closed-loop gain would be much more stable and is nearly independent of changes of open-loop gain
- If $A\beta >> 1, x_f = x_s \frac{A\beta}{1 + A\beta} \approx x_s$, so $x_i = x_s - x_f \approx 0$. Thus, in a negative feedback amplifier, the output takes the value to drive the amplifier input to almost 0 (this is summing point constraints).

Amplifier negative feedback: reduce nonlinear distortion

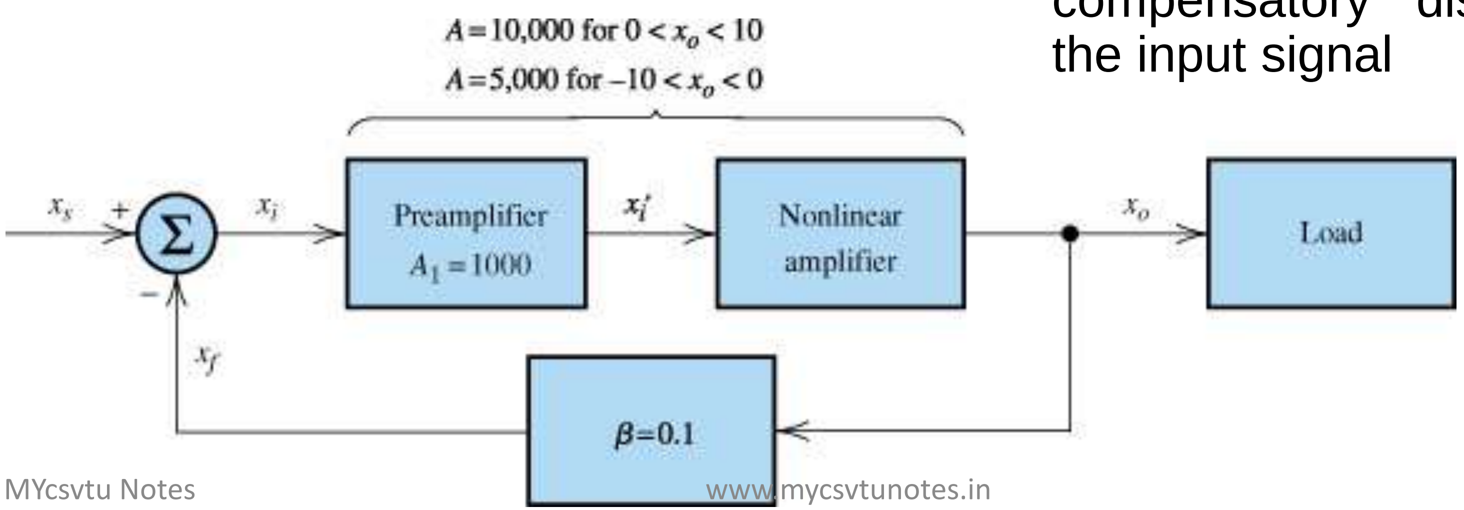


- If a pre-amplifier with gain 1000 is placed before the nonlinear one so that and the whole amplifier is used with negative feedback, $A\beta >> 1$ and the gain for whole amplifier becomes:

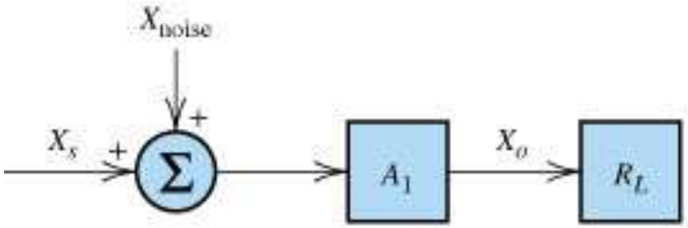
  $$A_f = 9.99 \text{ for } 0 < x_o < 10$$

  $$A_f = 9.98 \text{ for } -10 < x_o < 0$$

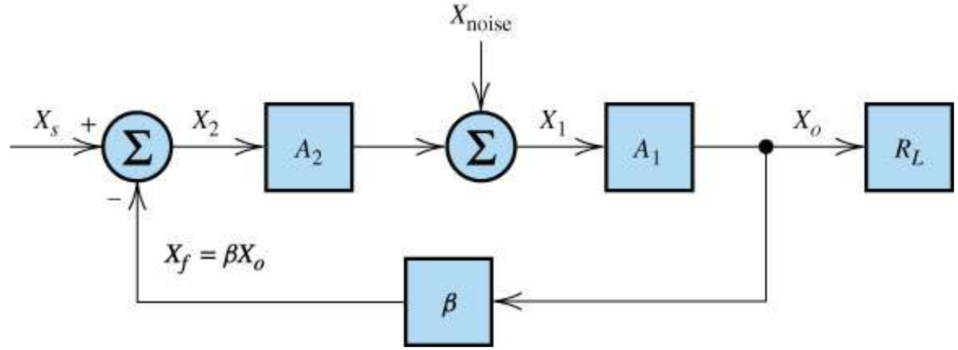
  which greatly reduce the nonlinear distortion.
- This is achieved through compensatory distortion of the input signal



Amplifier negative feedback: noise reduction



(a) Noise signal referred to the input



$$x_o(t) = x_s(t)A_1 + x_{noise}(t)A_1$$

$$SNR = \frac{(x_s)^2}{(x_{noise})^2}$$

- If an amplifier (assumed to be noise free or very low noise) is placed before the noisy amplifier, then the Signal-to-Noise (SNR) ratio is greatly enhanced (by a factor equal to the preceding amplifier gain)

$$x_2(t) = x_s(t) - \beta x_o(t)$$

$$x_1(t) = A_2 x_2(t) + x_{noise}(t)$$

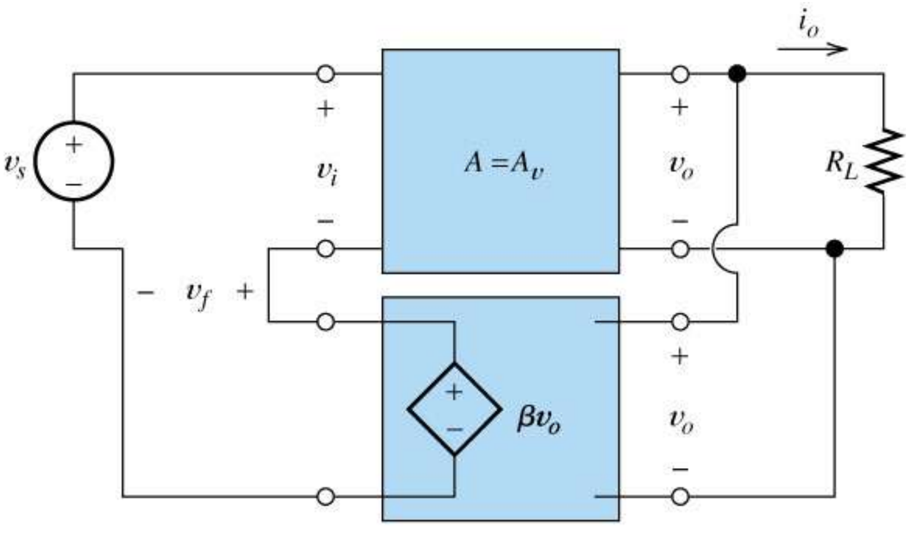
$$x_o(t) = A_1 x_1(t)$$

$$x_o(t) = x_s(t)\frac{A_1 A_2}{1+\beta A_1 A_2} + x_{noise}(t)\frac{A_1}{1+\beta A_1 A_2}$$

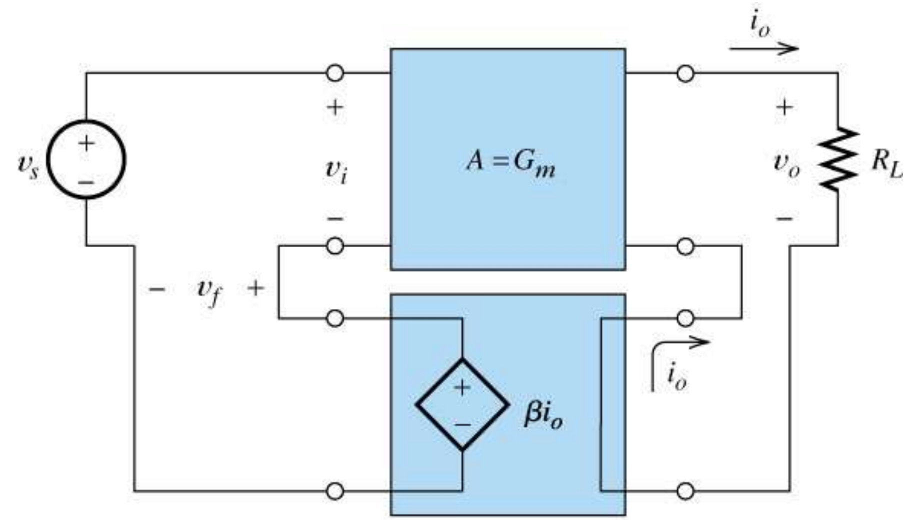
$$SNR = \frac{(x_s)^2}{(x_{noise})^2}(A_2)^2$$

- As a summary, negative feedback is very useful in amplifier circuits. It can help stabilize the gain, reduce nonlinear distortion and reduce noise.
- Also, as will be shown later, negative feedback in amplifiers can also control input and output impedance.

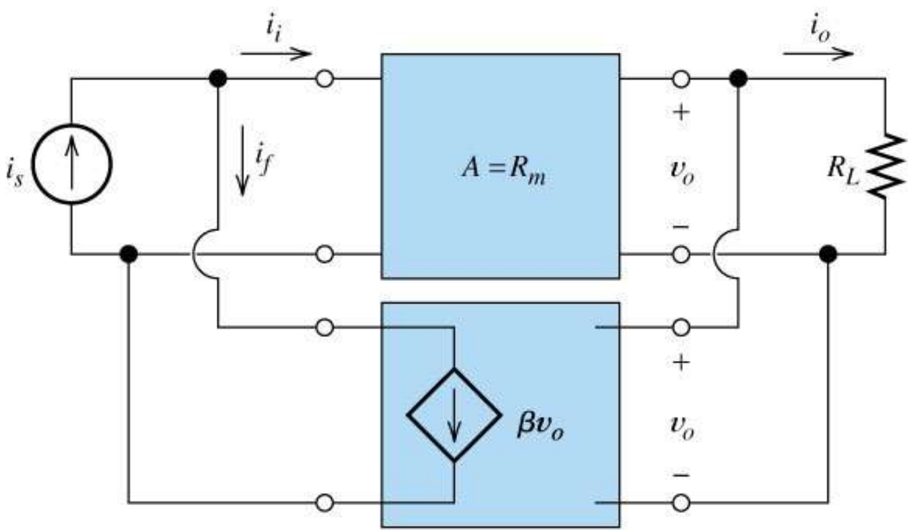
Amplifier negative feedback types



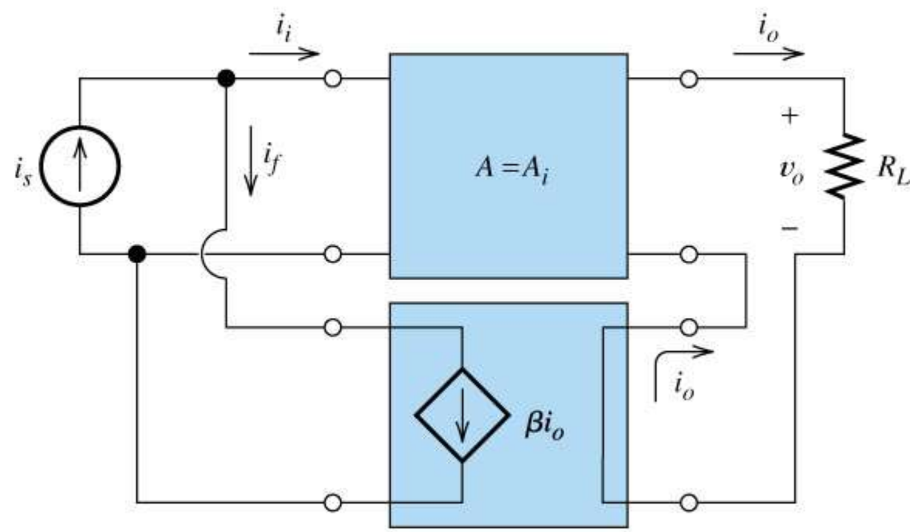
(a) Series voltage feedback



(b) Series current feedback



(c) Parallel voltage feedback



(d) Parallel current feedback

Amplifier negative feedback types

- If the feedback network samples the output voltage, it is *voltage feedback*. If it samples the output current, it is *current feedback*.
- The feedback signal can be connected in series or in parallel with the signal source and the amplifier input terminals, so called *series feedback* and *parallel feedback*.
- So, there are four types of negative feedback in amplifier circuits:
  - ➢ Series voltage feedback (corresponding to (a) in previous slide)
  - ➢ Series current feedback (corresponding to (b) in previous slide)
  - ➢ Parallel voltage feedback (corresponding to (c) in previous slide)
  - ➢ Parallel current feedback (corresponding to (d) in previous slide)
    - ✓ In voltage feedback, the input terminals of the feedback network are in parallel with the load, and the output voltage appears at the input terminals of the feedback block.
    - ✓ Whereas in current feedback, the input terminals of the feedback network are in series with the load, and the load current flows through the input of the feedback block.
    - ✓ As a result, a simple test on the feedback type is to open-circuit or short-circuit the load. If the feedback signal vanishes for an open-circuit load, then it is current feedback. If the feedback signal vanishes for a short-circuit load, it is voltage feedback.

Effect of negative feedback on gain

- In series voltage feedback, input signal is voltage and output voltage is sampled, so it is natural to model the amplifier as a voltage amplifier.

$$\text{In general}, A_f = \frac{A}{1+A\beta}, \text{so } A_{vf} = \frac{A_v}{1+A_v\beta}$$

- Amplifier employing series current feedback is modeled as a transconductance amplifier.

$$\text{In general } A_f = \frac{A}{1+A\beta}, \text{this is modeled by a transconductance amplifier, so } G_{mf} = \frac{G_m}{1+G_m\beta}$$

- Amplifier employing parallel voltage feedback is modeled as a transresistance amplifier.

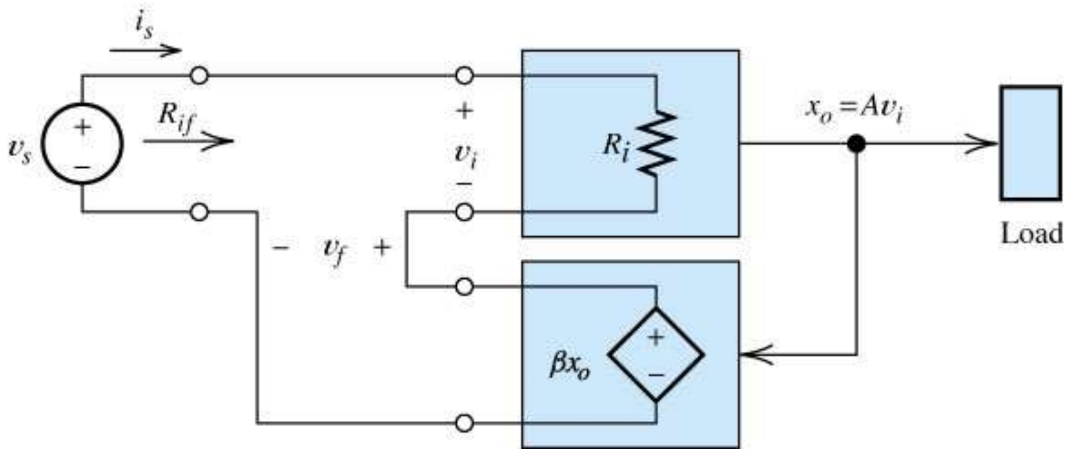
$$\text{In general } A_f = \frac{A}{1+A\beta}, \text{this is modeled by a transresistance amplifier, so } R_{mf} = \frac{R_m}{1+R_m\beta}$$

- Amplifier employing parallel current feedback is modeled as a current amplifier.

$$\text{In general } A_f = \frac{A}{1+A\beta}, \text{this is modeled by a current amplifier, so } A_{if} = \frac{A_i}{1+A_i\beta}$$

Negative feedback on input impedance

- For series feedback, the following model can be used for analysis of input impedance (the output x could be either voltage or current)



If the input impedance of the open-loop amplifier is Ri, then the closed-loop impedance is

$$R_{if} = R_i(1 + A\beta), \text{notice } A\beta >> 1 \text{ for negative feedback}$$

so, series feedback (either current or voltage) increase the input impedance

- Similarly, the effect of parallel feedback on input impedance can be analyzed using a similar model, the closed-loop input impedance would then be $R_{if} = R_i / (1 + A\beta)$

so, parallel feedback decrease the input impedance

Negative feedback on output impedance

- For voltage feedback, (it could be either series or parallel feedback), the closed-loop impedance is

$$R_{of} = R_o/(1 + A\beta)$$

so, voltage feedback decrease the output impedance

- Similarly, for current feedback (either series or parallel feedback), the closed-loop impedance is

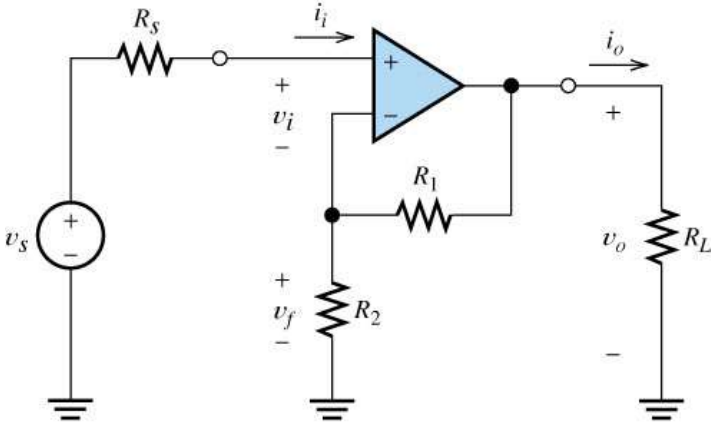
$$R_{of} = R_o(1 + A\beta)$$

so, current feedback increase the output impedance

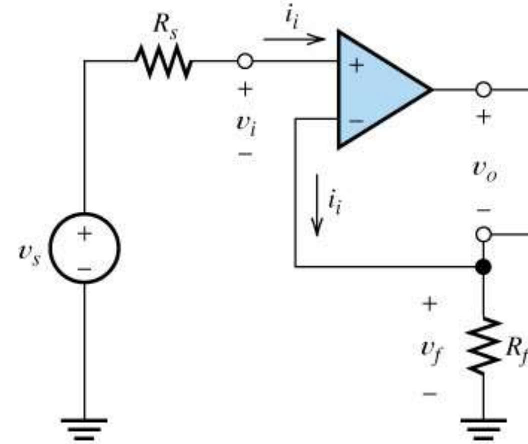
- As a summary, negative feedback tends to stabilize and linearize gain, which are desired effects.
- For a certain type of amplifier, negative feedback tends to produce an ideal amplifier of that type.
- For example, series voltage feedback increases input impedance, reduces output impedance, which gets closer to an ideal voltage amplifier.
- So, negative feedback should be used in amplifiers circuits.

Some practical feedback network in amplifiers

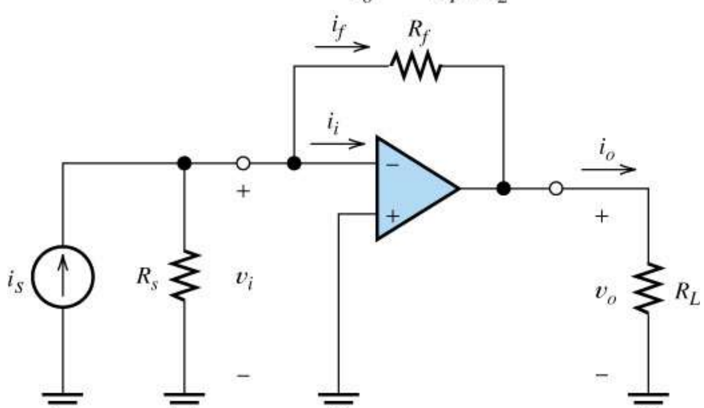
- In practice, negative feedback network consists of resistor or capacitors , whose value is much more precise and stable than active devices (such as transistors). Then amplifier characteristics mainly depends on feedback network, thereby achieving precision and stability.



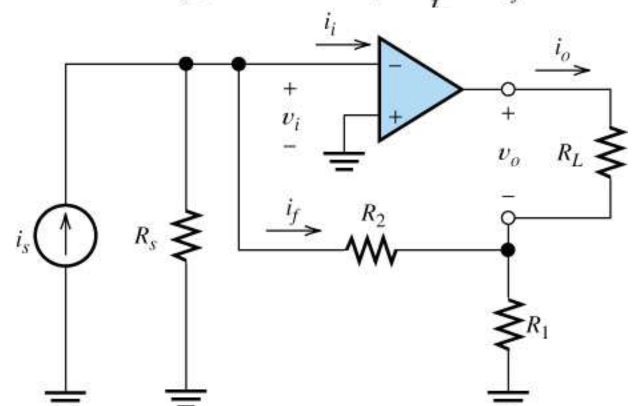
(a) Series voltage $\beta = \frac{v_f}{v_o} = \frac{R_2}{R_1 + R_2}$



(b) Series current $\beta = \frac{v_f}{i} = R_f$



(c) Parallel voltage $\beta = \frac{i_f}{v_o} = -\frac{1}{R_f}$



(d) Parallel current $\beta = \frac{i_f}{i_o} = -\frac{R_1}{R_1 + R_2}$

Design of negative feedback amplifiers

- A few steps to design negative feedback amplifiers:
  - Select the feedback type and determine feedback ratio β
  - Select an appropriate circuit configuration for the feedback network (adjustable resistor can be used so that feedback ratio can be set precisely)
  - Select appropriate values for resistance in the feedback network (this could be a difficult step due to various tradeoffs)

    E.g., in series voltage feedback (like the non-inverting amplifier), we do not want the feedback resistance too small because it loads the output of the amplifier, on the other hand, we do not want feedback resistance too large because it would cause part of the source signal to be lost).
  - Verify the design using Computer Simulations (real circuits could be very different from the ideal case)

Unit 5
Oscillators

An oscillator is any measurable quantity capable of repetition.

Examples:

Volume of a loudspeaker

Brightness of a bulb

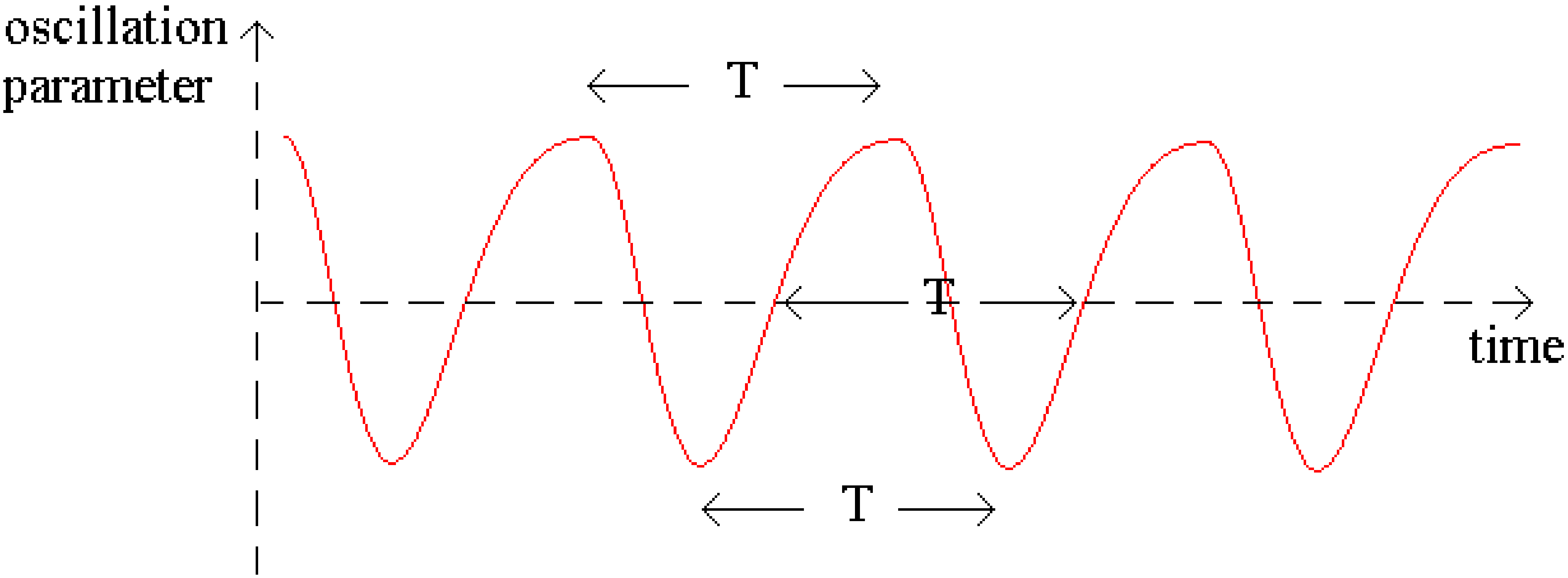
Amount of money in a bank account

The air pressure near your eardrum

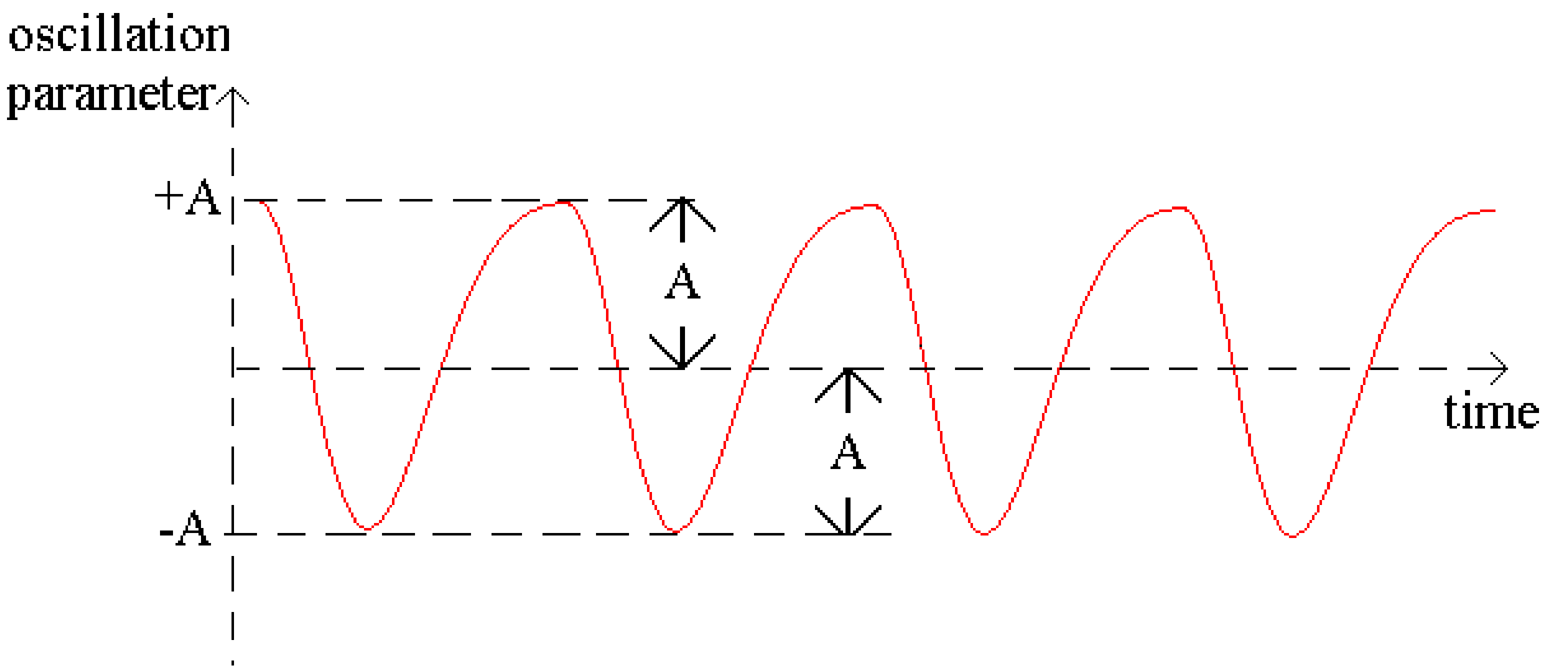
The *oscillator parameter* is the quantity that repeats. In our examples, the oscillator parameter has units of sound intensity, light intensity, dollars, and pressure, respectively.

Question: Is the x-component of the electric field, at a particular location in space, an oscillator? If not, why not? If so, what are the units of the oscillation parameter?

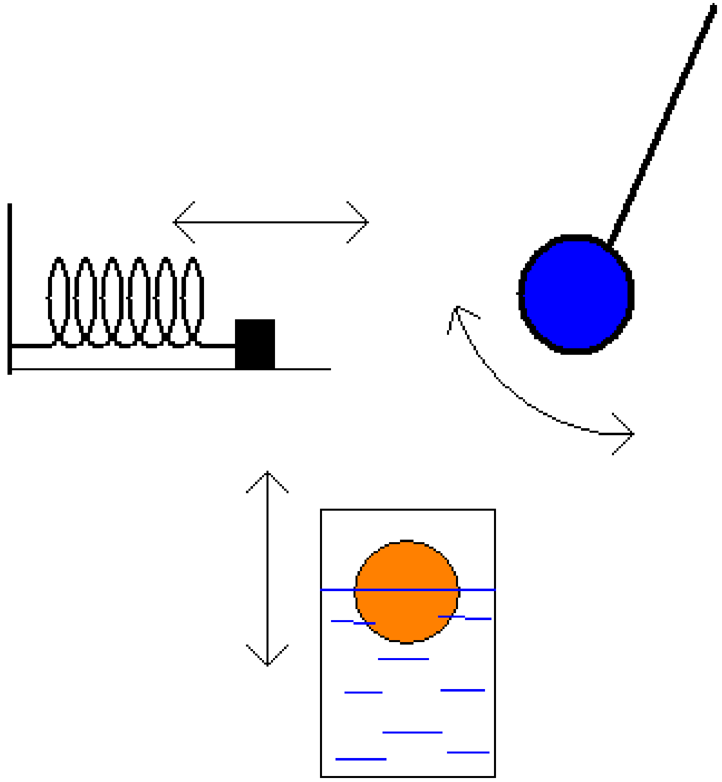
The period (denoted T, measured in seconds) is the time required for one cycle.



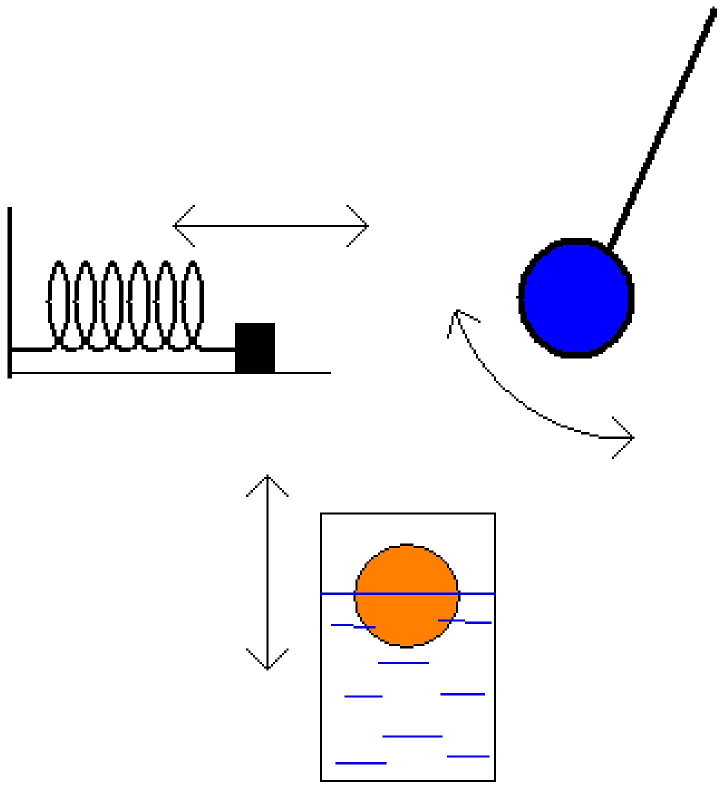
The *amplitude* (A) of the oscillation is the maximum value of the oscillation parameter, measured relative to the average value of that parameter.

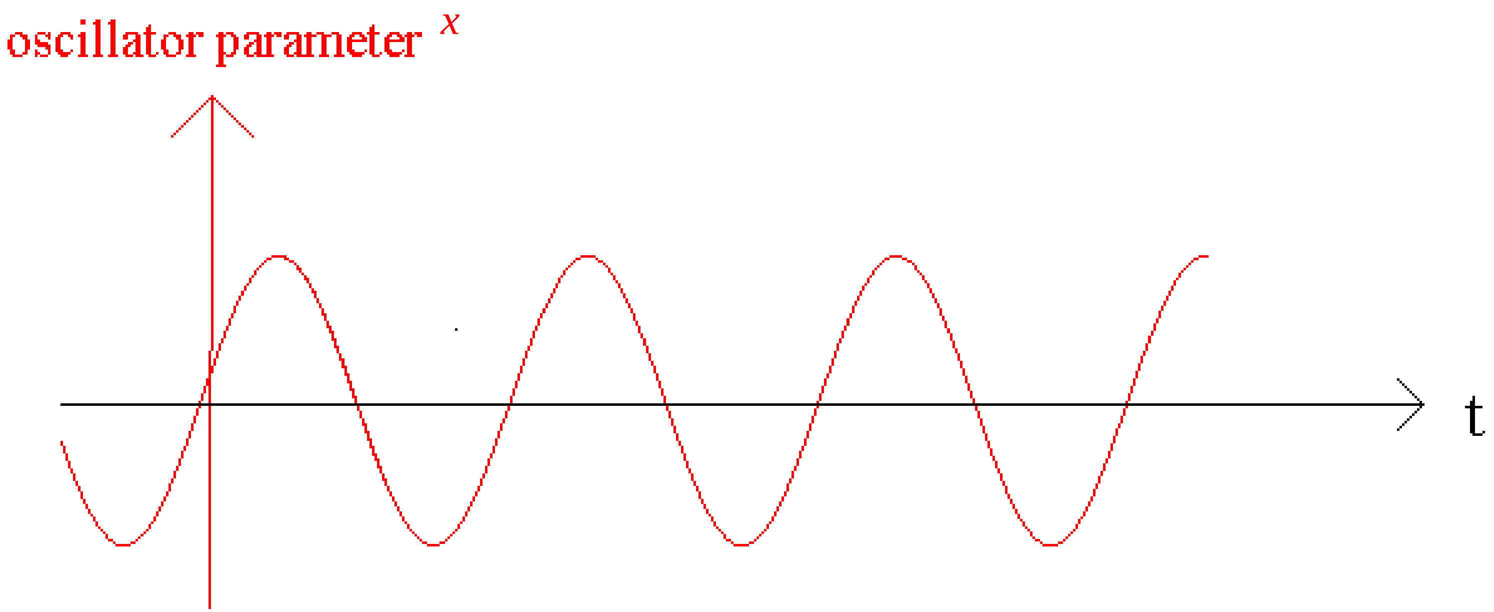


Many naturally occurring oscillators are of the class called *resonators*. They oscillate freely at one particular frequency. This kind of oscillation is called *simple harmonic motion*.



identify the oscillator parameter for the below resonators





For a free resonator, the oscillator parameter varies *sinusoidally* and can be expressed as

$$x(t) = x_0 \cos(\omega t + \phi)$$

Some definitions for simple harmonic motion.

$$x(t) = x_0 \cos(\omega t + \phi) \equiv \text{amplitude function}$$

$$x_0 \equiv \text{amplitude}$$

$$\omega t + \phi \equiv \text{phase}$$

$$\phi \equiv \text{phase angle } \textit{or} \text{ phase constant}$$

The oscillator parameter repeats itself

every time the phase $\left(\omega t+\phi\right)$ changes by 2π.

This happens every time t changes by one period (T)

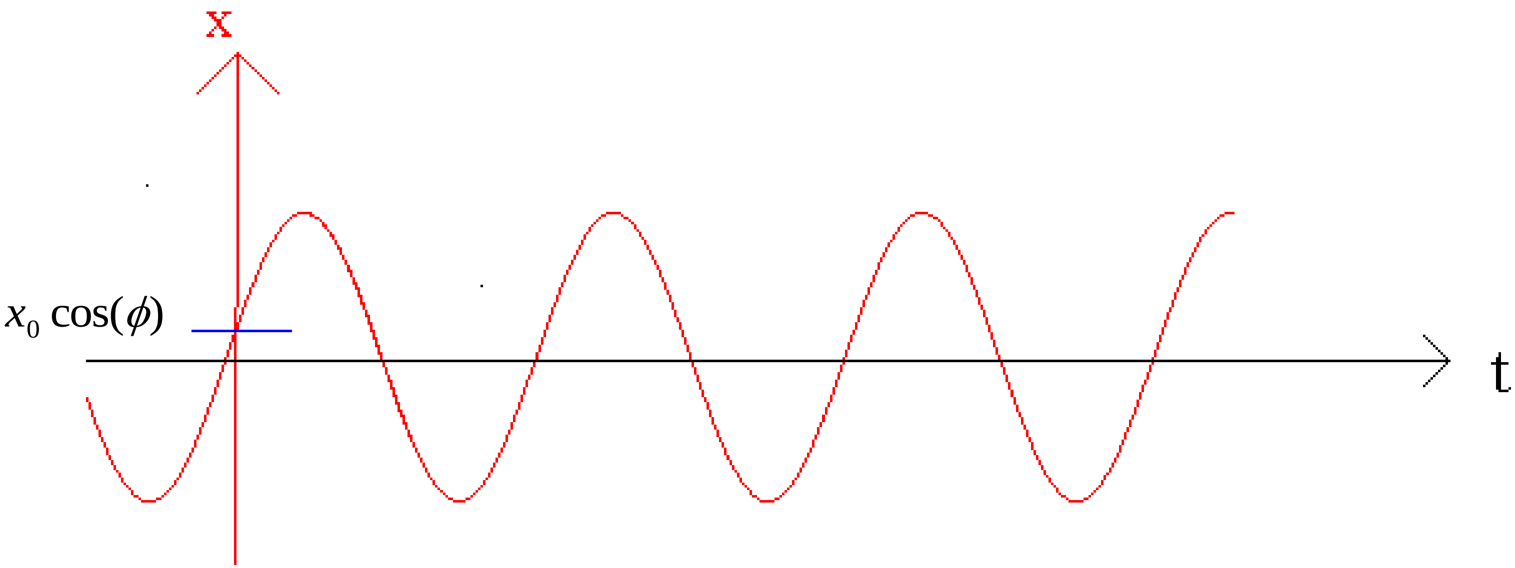
$$\Delta t = \frac{2\pi}{\omega} = T$$

The frequency of the motion is $f = \dfrac{1}{T} = \dfrac{\omega}{2\pi}$

expressed in units of Hertz = oscillations per s

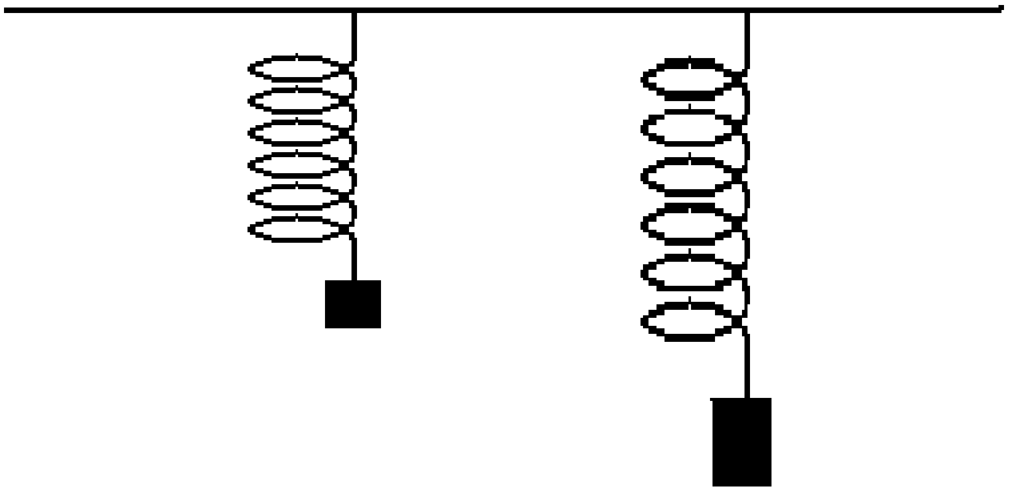
The phase changes at the rate ω radians per s.

This is called the circular frequency or angular frequency.

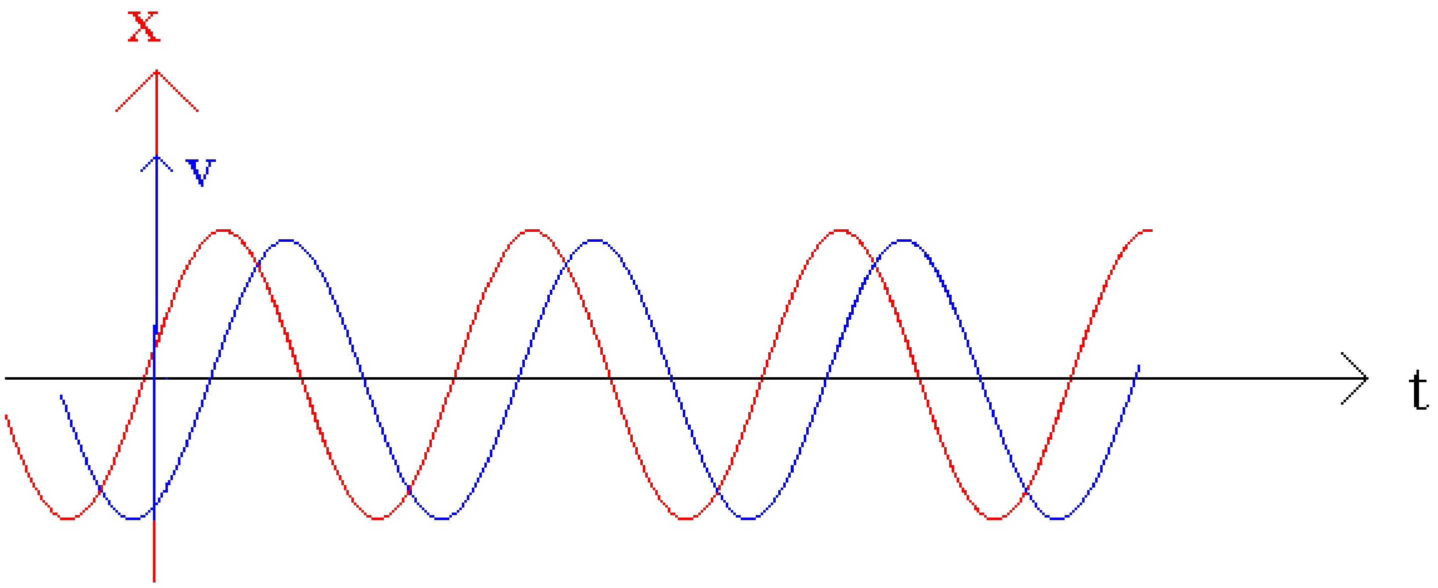


The phase constant ☠ specifies the oscillator parameter at t = 0.

$$x(0) = x_0 \cos(\omega 0 + \phi) = x_0 \cos(\phi)$$

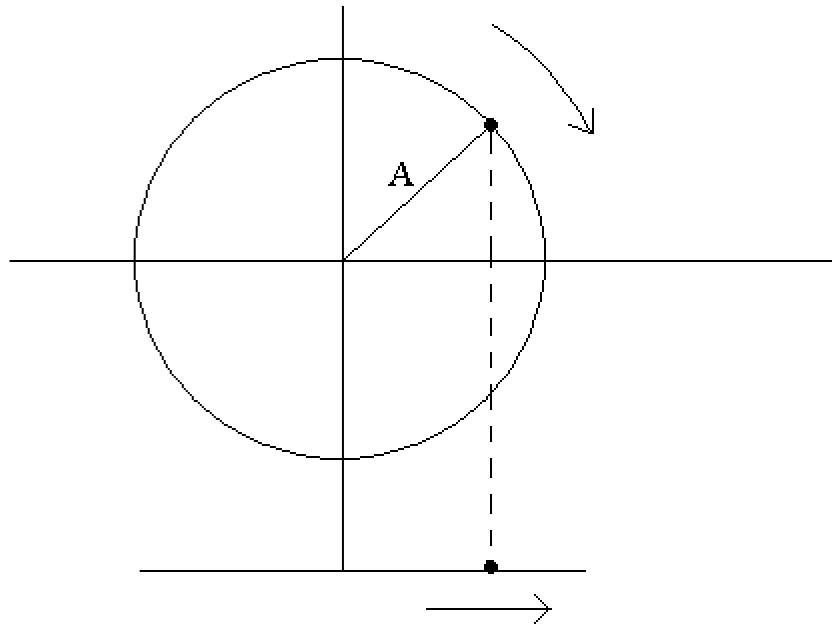


demo



$$v = \frac{dx}{dt} = -\omega x_0 \cos(\omega t + \phi)$$

v varies from $-\omega x_0$ to $+\omega x_0$



Here’s a way of thinking about simple harmonic motion that will be very useful later. Suppose a particle is in uniform circular motion on a circle of radius *A* and with angular speed T. The projection of that motion onto any axis is a point moving in simple harmonic motion with amplitude *A* and circular frequency T.

The total energy of a mass-spring system is

$$\text{total energy} = \tfrac{1}{2}mv^2 + \tfrac{1}{2}kx^2 = \tfrac{1}{2}m\left(\frac{dx}{dt}\right)^2 + \tfrac{1}{2}k\left[x_0\cos(\omega t+\phi)\right]^2$$

$$= \tfrac{1}{2}m\left[-\omega x_0\sin(\omega t+\phi)\right]^2 + \tfrac{1}{2}k\left[x_0\cos(\omega t+\phi)\right]^2$$

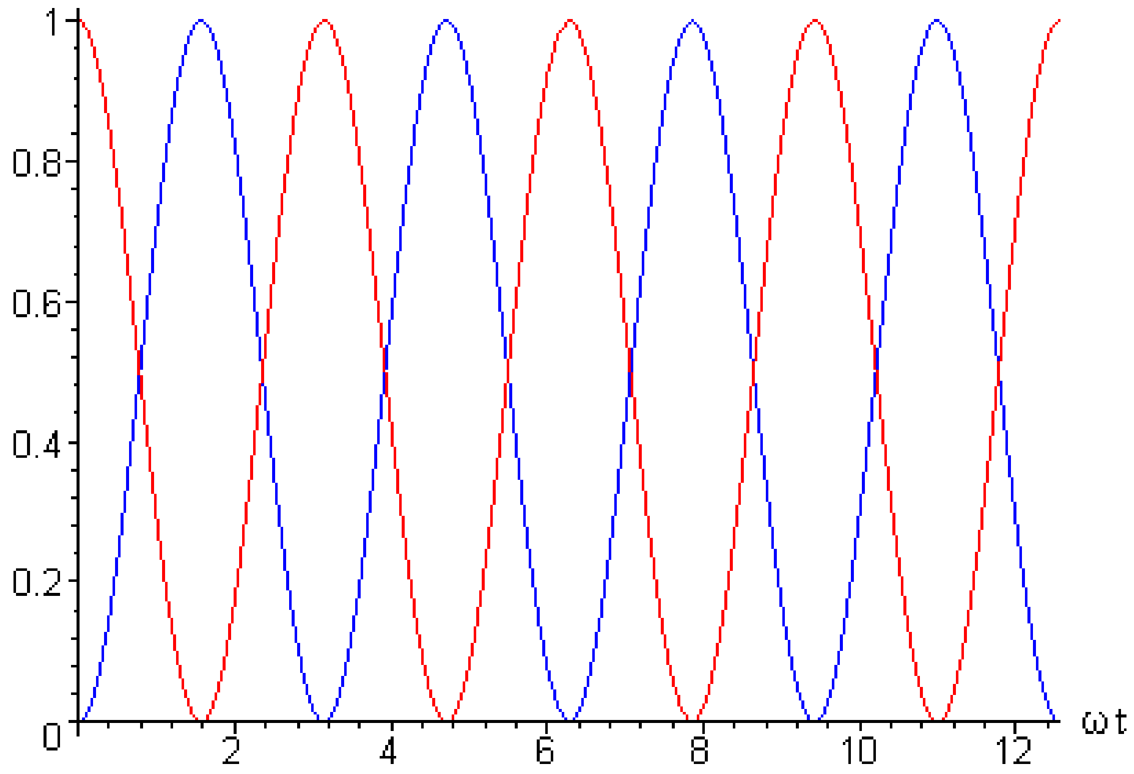
$$= \frac{m\omega^2 x_0^2}{2}\sin^2(\omega t+\phi) + \frac{k\,x_0^2}{2}\cos^2(\omega t+\phi)$$

Now, $\omega^2 = \dfrac{k}{m}$ for a spring, so that

$$\text{total energy} = \frac{m\omega^2 x_0^2}{2}\left[\sin^2(\omega t+\phi)+\cos^2(\omega t+\phi)\right] = \frac{m\omega^2 x_0^2}{2}$$

The total energy of a simple harmonic oscillator is constant in time and proportional to the square of the amplitude kinetic and potential energies (as fraction of total energy) for $x = x_0 \cos(Tt)$

U / (K+U), K/(K+U)



The frequency of a resonator is determined by the way energy shifts back and forth between kinetic and potential terms.

$$E = K + U = \tfrac{1}{2}mv^2 + \tfrac{1}{2}k\,x^2$$

$$0 = \frac{dE}{dt} = mv\frac{dv}{dt} + k\,x\frac{dx}{dt} = v\left(m\frac{dv}{dt} + k\,x\right)$$

since $v = \dfrac{dx}{dt}$. Using $\dfrac{dv}{dt} = \dfrac{d^2x}{dt^2}$ gives

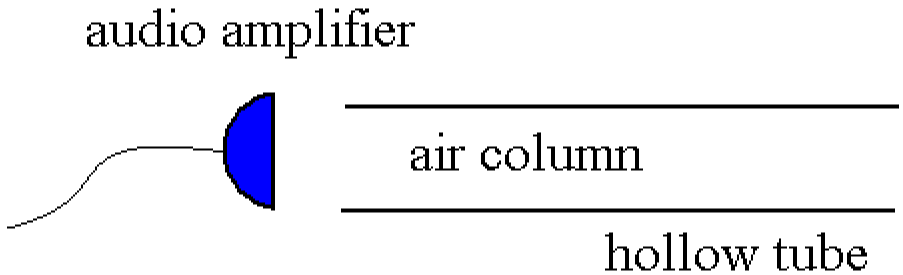
$$m\frac{d^2x}{dt^2} + k\,x = 0 \quad \Rightarrow \quad \frac{d^2x}{dt^2} + \frac{k}{m}x = 0$$

The solution to the equation $\frac{d^2x}{dt^2} + \frac{k}{m}x = 0$ is

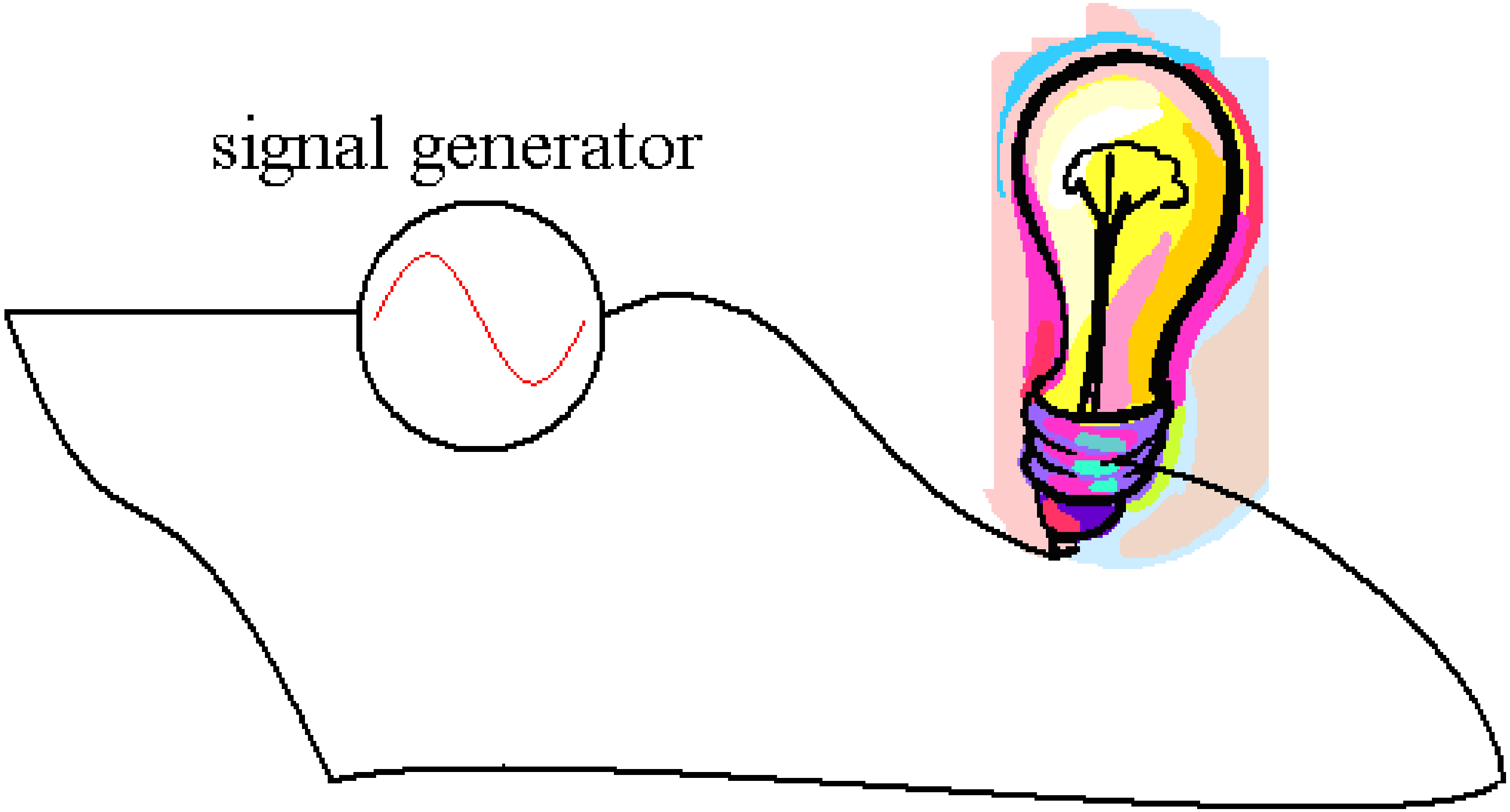
$x(t) = x_0 \cos\left(\sqrt{\frac{k}{m}}t + \phi\right)$ where x_0 and ϕ are constants that depend on the position and velocity of the mass at time t = 0.

These are called *sinusoidal* oscillations, or *simple harmonic motion*.

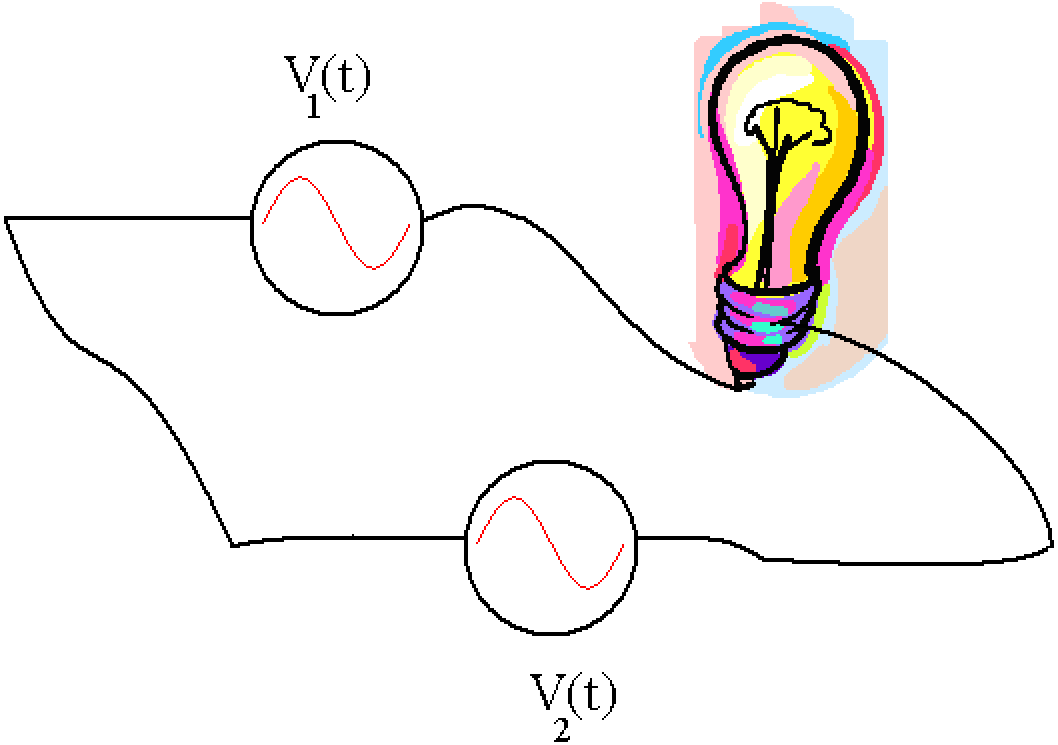
All oscillators can exhibit simple harmonic motion if they are driven harmonically.



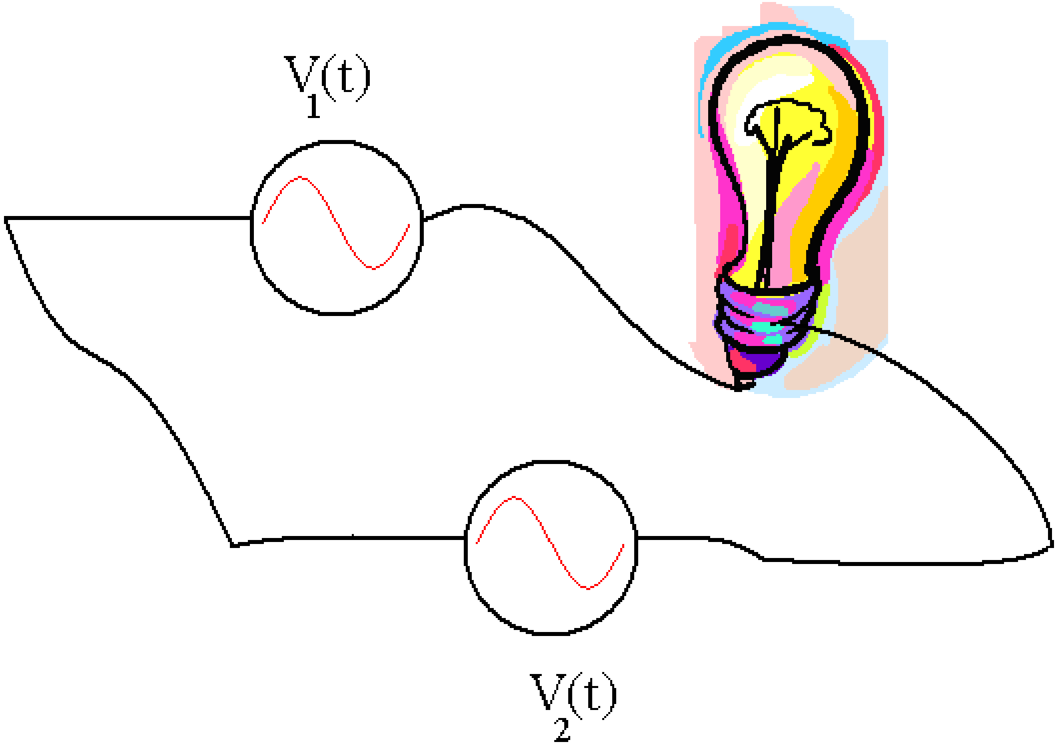
When a speaker oscillates sinusoidally, it causes the pressure at any point in a nearby column of air to vary sinusoidally


signal generator

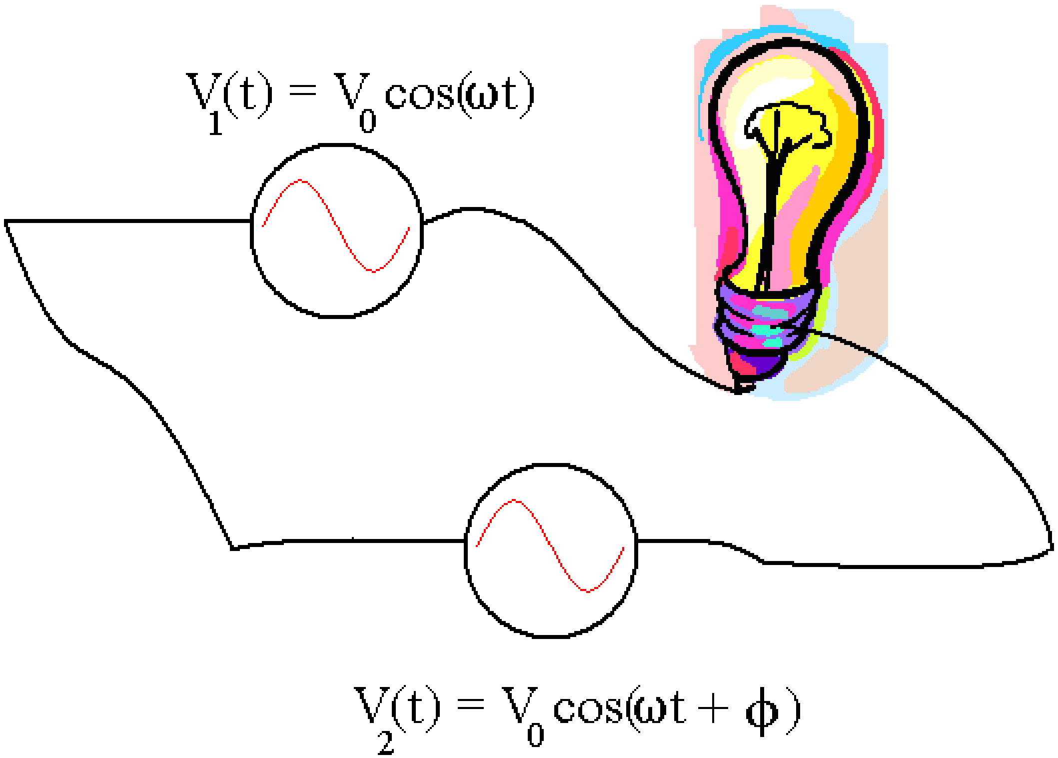
What happens if more than one oscillatory signal is sent to the light?

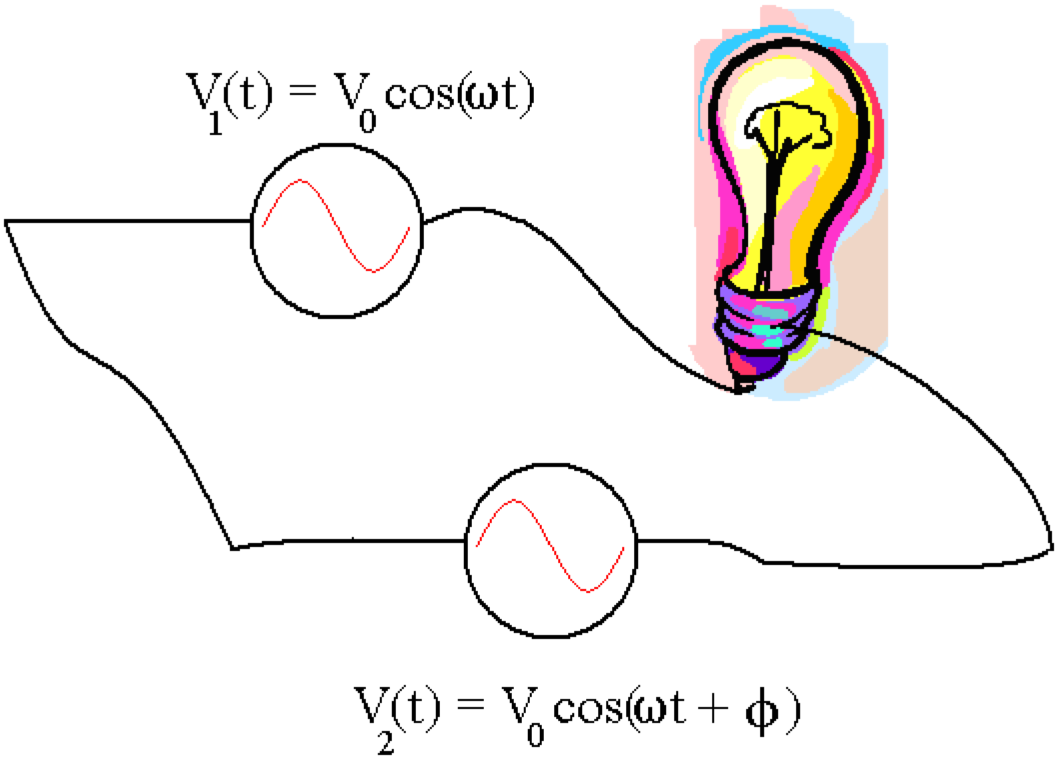


$$V(t) = V_1(t) + V_2(t)$$



Suppose the two signals have the same amplitude and frequency but differ in phase.



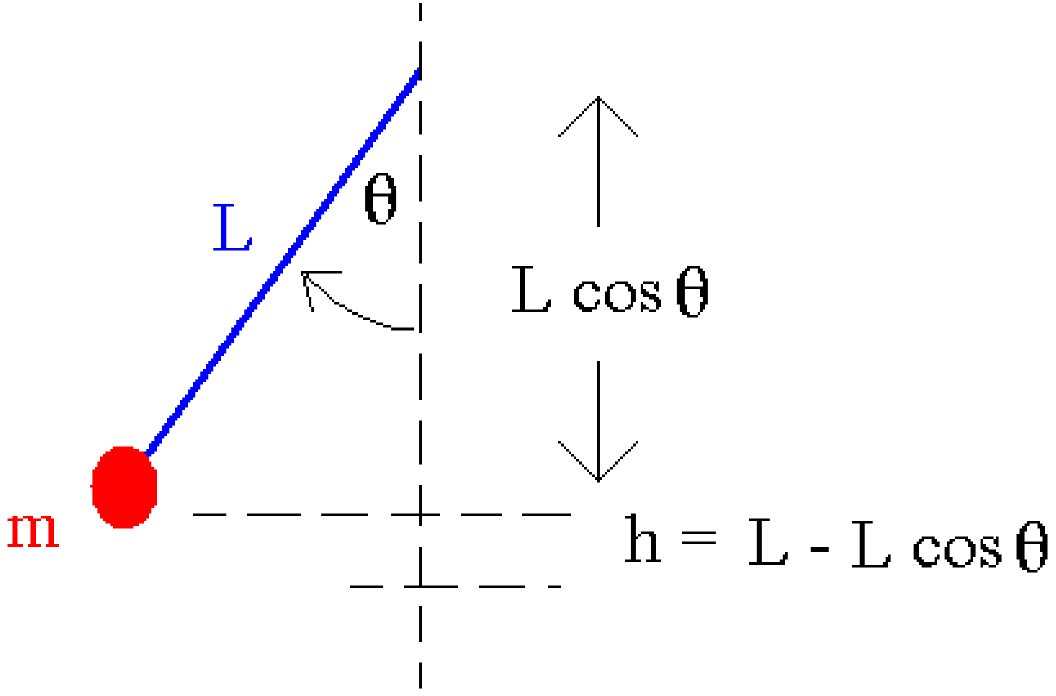

$V_1(t) = V_0 \cos(\omega t)$
$V_2(t) = V_0 \cos(\omega t + \phi)$

For a mass-and-spring system, the motion is simple harmonic at an angular frequency of

$$\sqrt{k/m}$$

The motion is always simple harmonic whenever the restoring force is a linear function of the oscillation parameter. Another way to say this is that the motion is simple harmonic whenever the potential energy is a quadratic function of the oscillation parameter.

For example, a pendulum has a kinetic energy of $K = ½ mv^2$ and a potential energy of $U = mgh$.

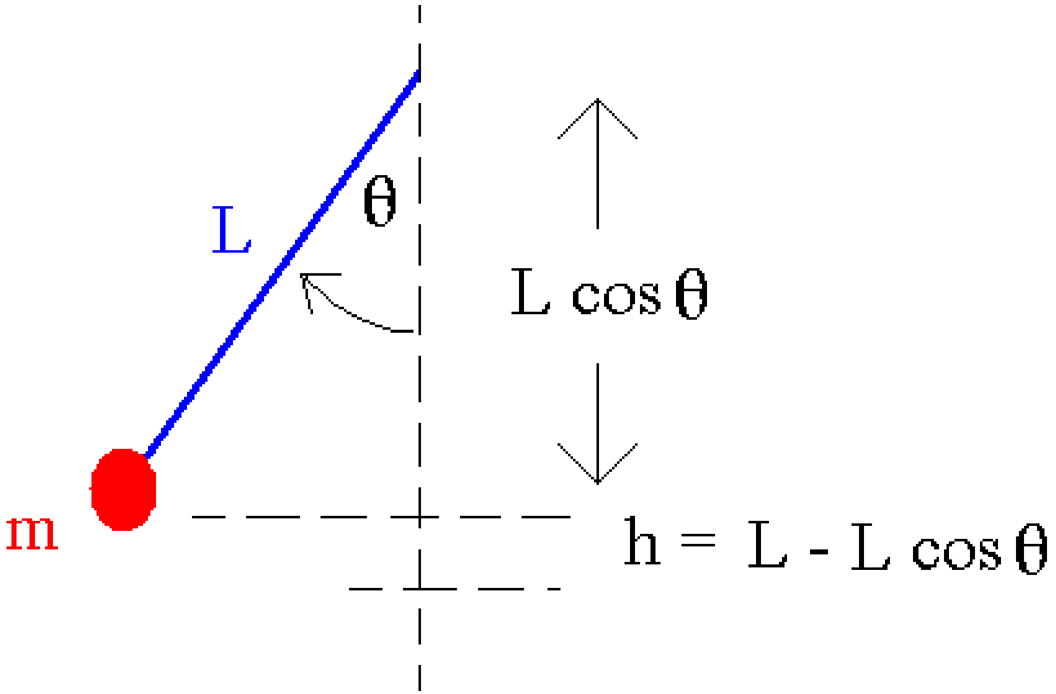


$$x = L\sin\theta \quad \Rightarrow \quad v_x = \frac{dx}{dt} = \left(L\cos\theta\right)\frac{d\theta}{dt}$$

$$y = L\cos\theta \quad \Rightarrow \quad v_x = \frac{dx}{dt} = \left(-L\sin\theta\right)\frac{d\theta}{dt}$$

$$K = \tfrac{1}{2}mv^2 = \tfrac{1}{2}m\left(v_x{}^2 + v_y{}^2\right)$$

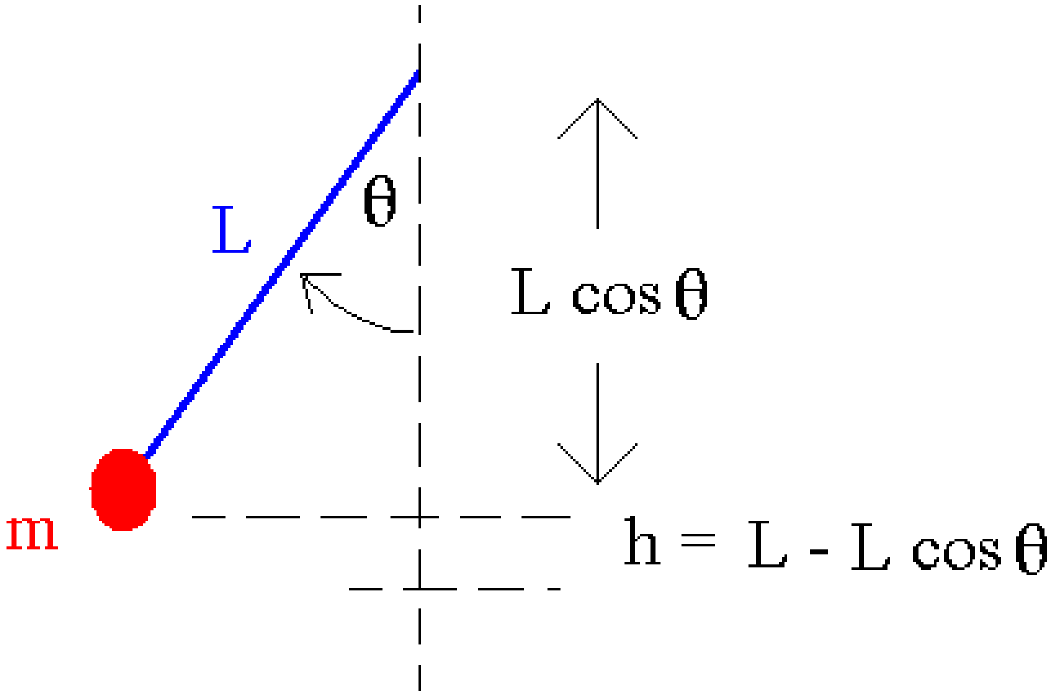
$$= \tfrac{1}{2}m\left(L^2\cos^2\theta + L^2\sin^2\theta\right)\left(\frac{d\theta}{dt}\right)^2 = \tfrac{1}{2}mL^2\left(\frac{d\theta}{dt}\right)^2$$



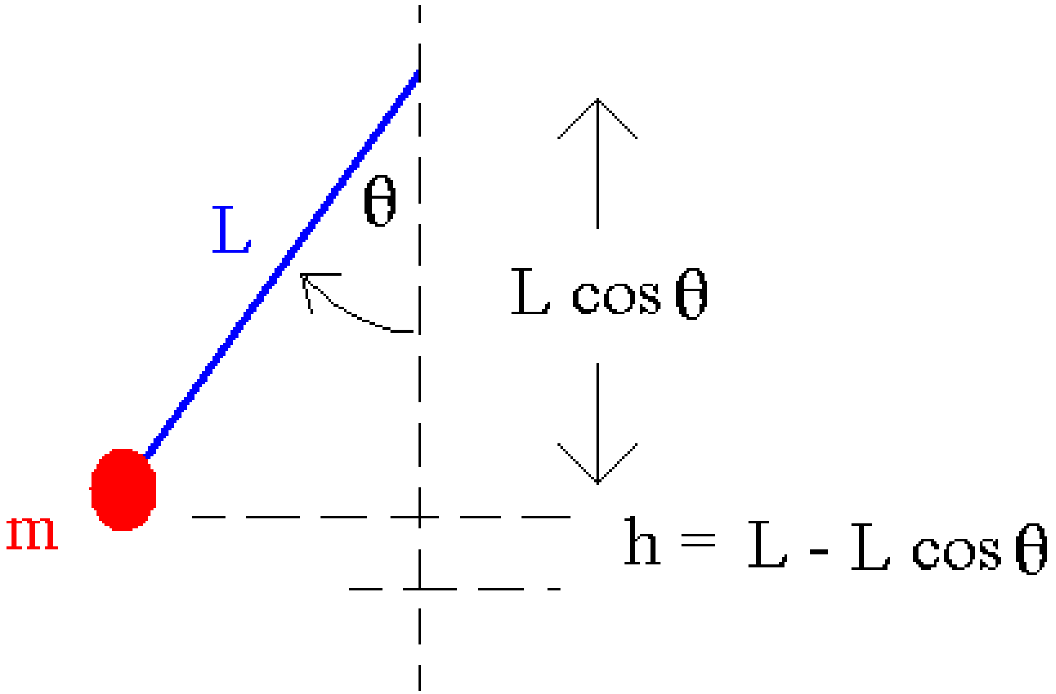
$$U = mgh = mgL(1 - \cos\theta)$$

$$\text{for small angles, } \cos\theta \approx 1 - \tfrac{1}{2}\theta^2$$

$$\text{so that } U \approx \tfrac{1}{2} mgL\theta^2$$



$$K = \frac{1}{2} m L^2 \left(\frac{d\theta}{dt} \right)^2 , \quad U \approx \frac{1}{2} m g L \theta^2$$



$K = ½\, m\, L^2\, (d\theta/dt)^2$, $U = ½\, mgL\, \theta^2$

Compare to

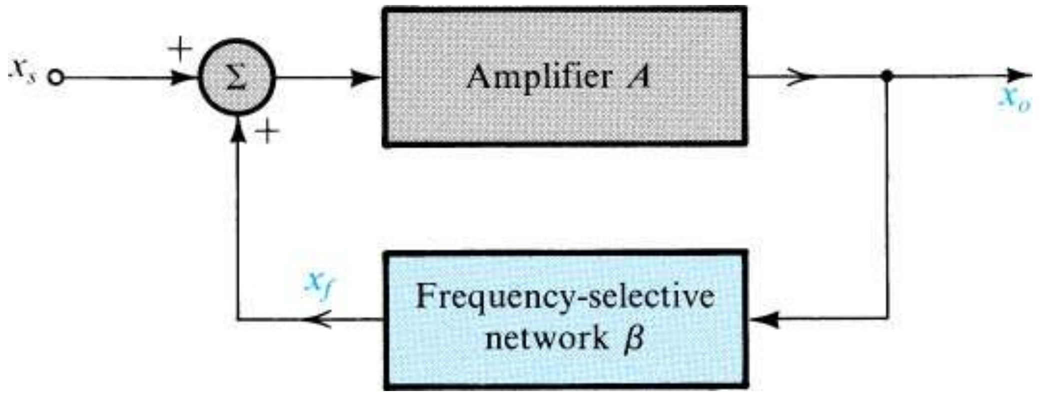
$K = ½\, m\, (dx/dt)^2$, $U = ½\, kx^2$

The circular frequency of the mass-spring system is $(½\, k\, /\, ½\, m)^{1/2} = (\, k\, /\, m)^{1/2}$

By analogy, the circular frequency of the pendulum is $[½\, mgL\, /\, ½\, m\, L^2]^{1/2} = [g\, /\, L]^{1/2}$

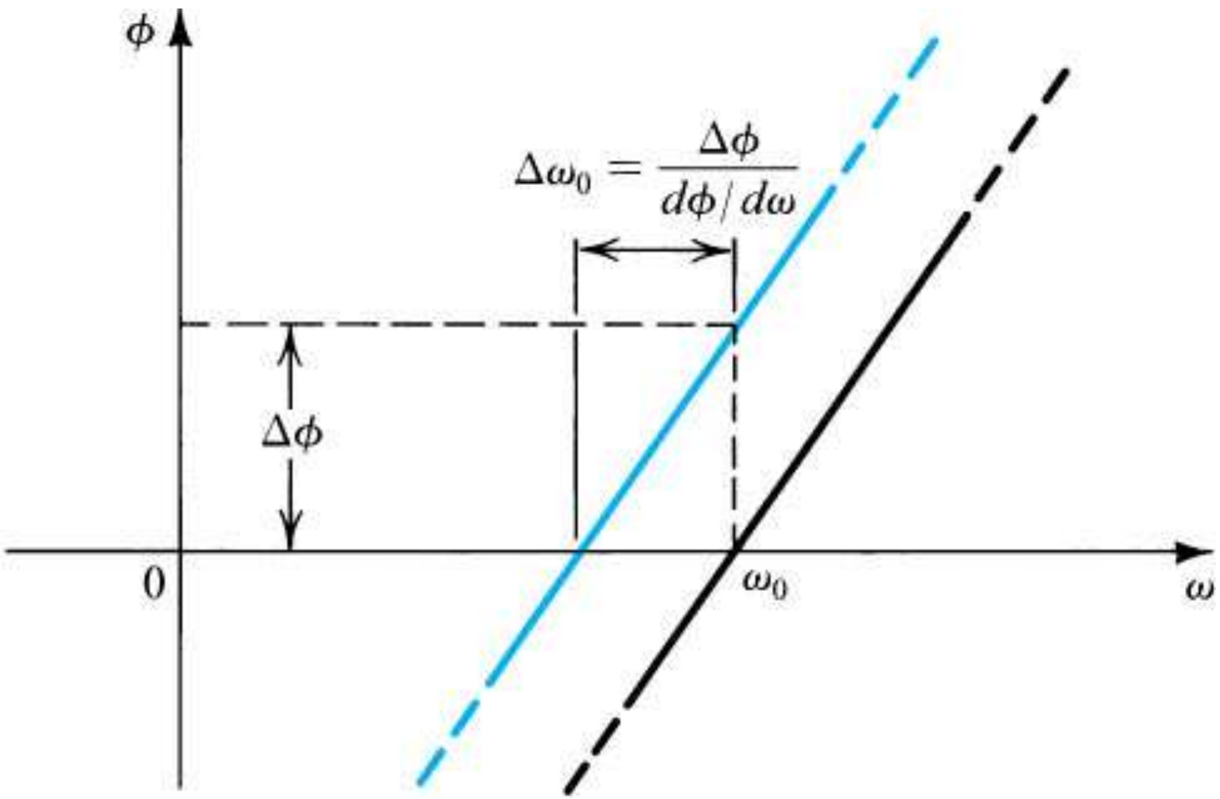
SIGNAL GENERATORS /OSCILLATORS

A positive-feedback loop is formed by an amplifier and a frequency-selective network

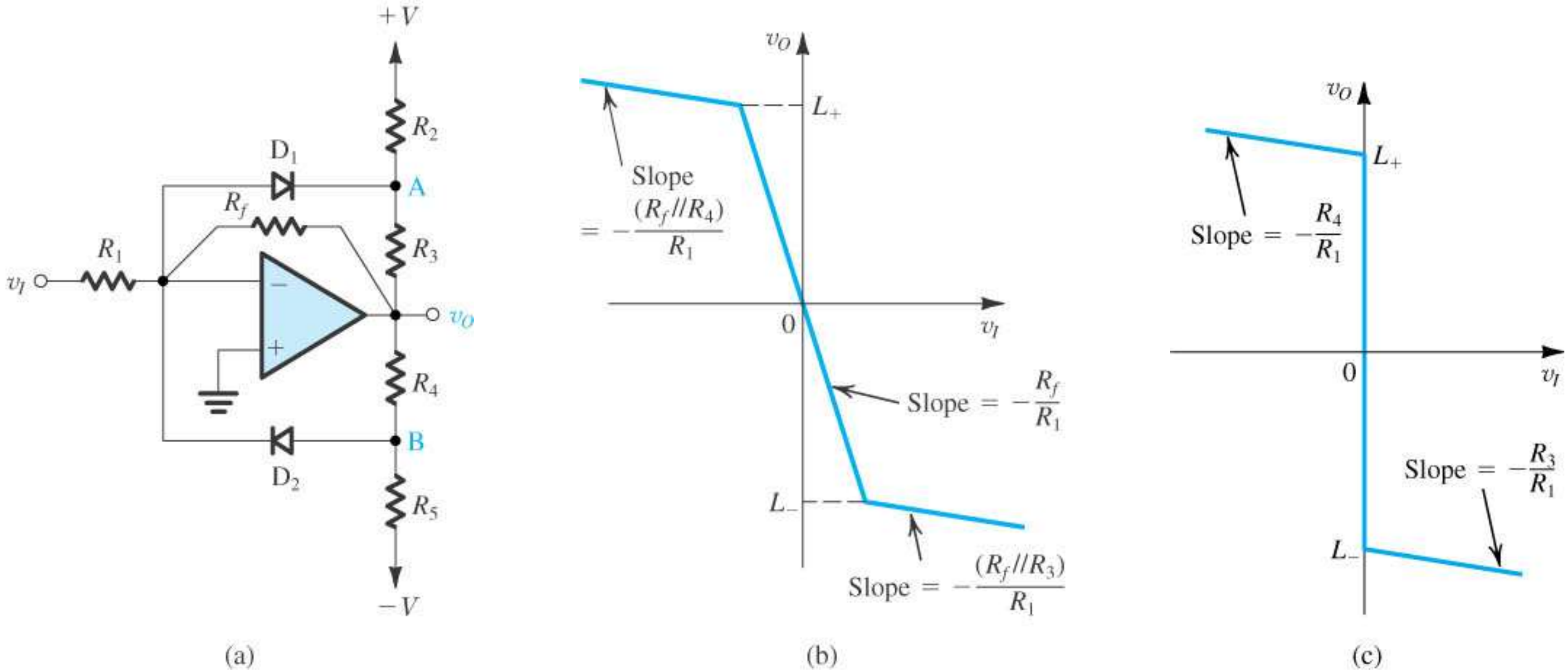


In an actual oscillator circuit, no input signal will be present

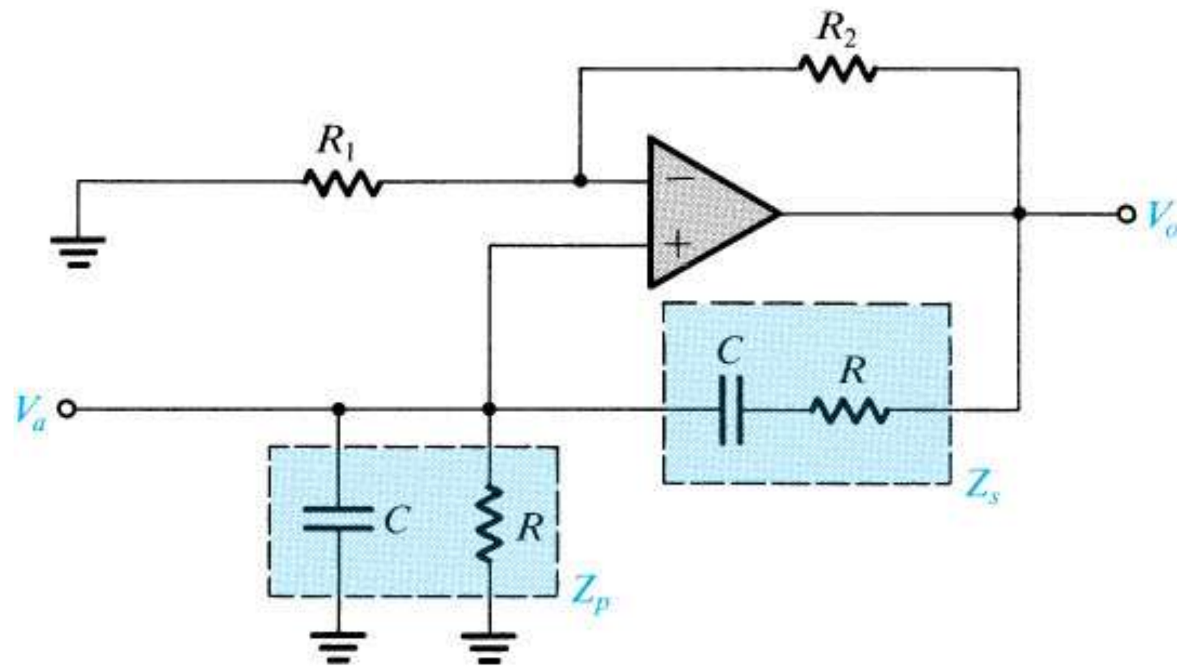
Oscillator-frequency stability



Limiter Ckt→ Comparator



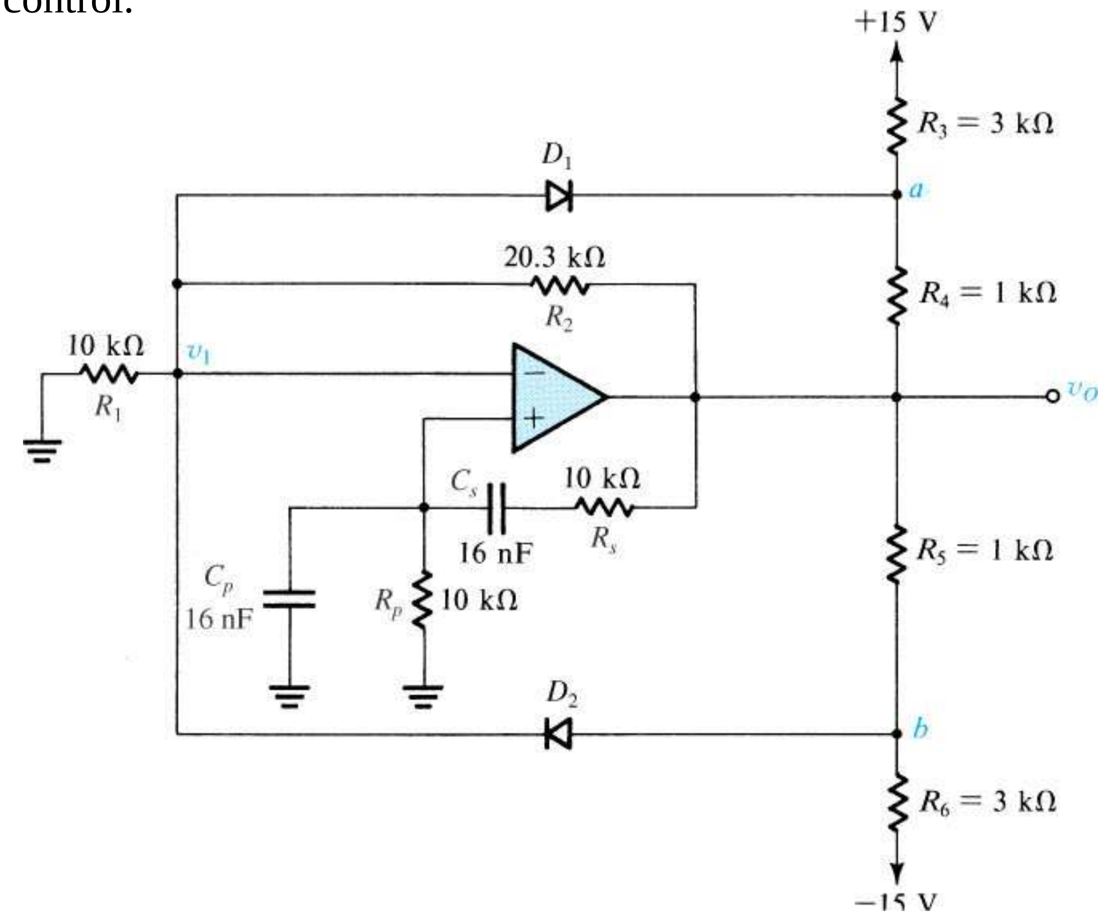
Wien-bridge oscillator



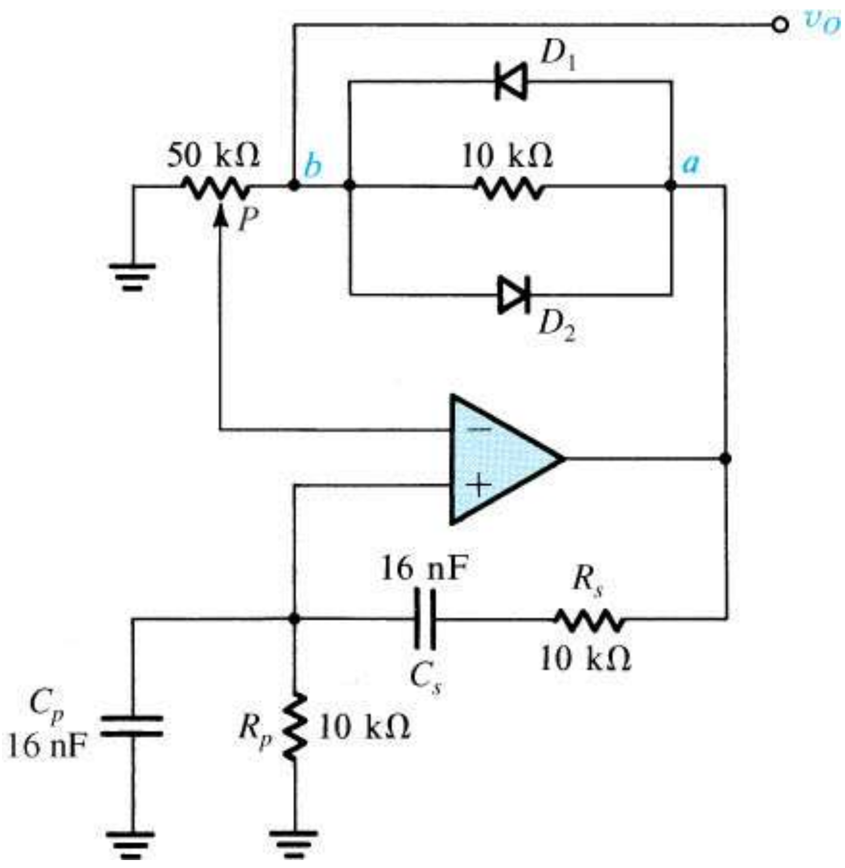
without amplitude stabilization.

Wien bridge w/ Amp. Stabil.

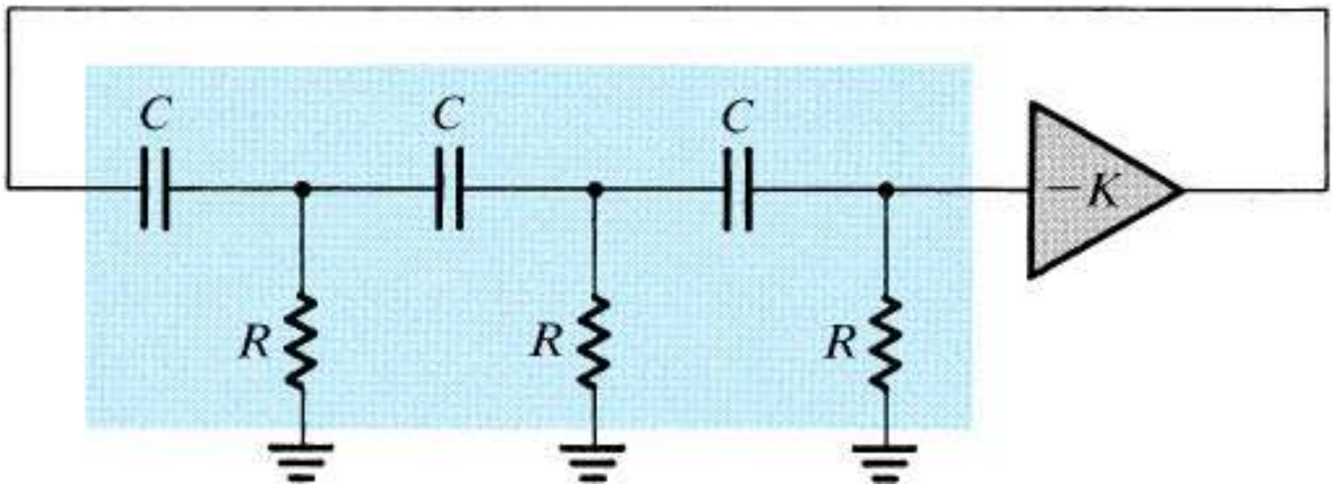
limiter used for amplitude control.



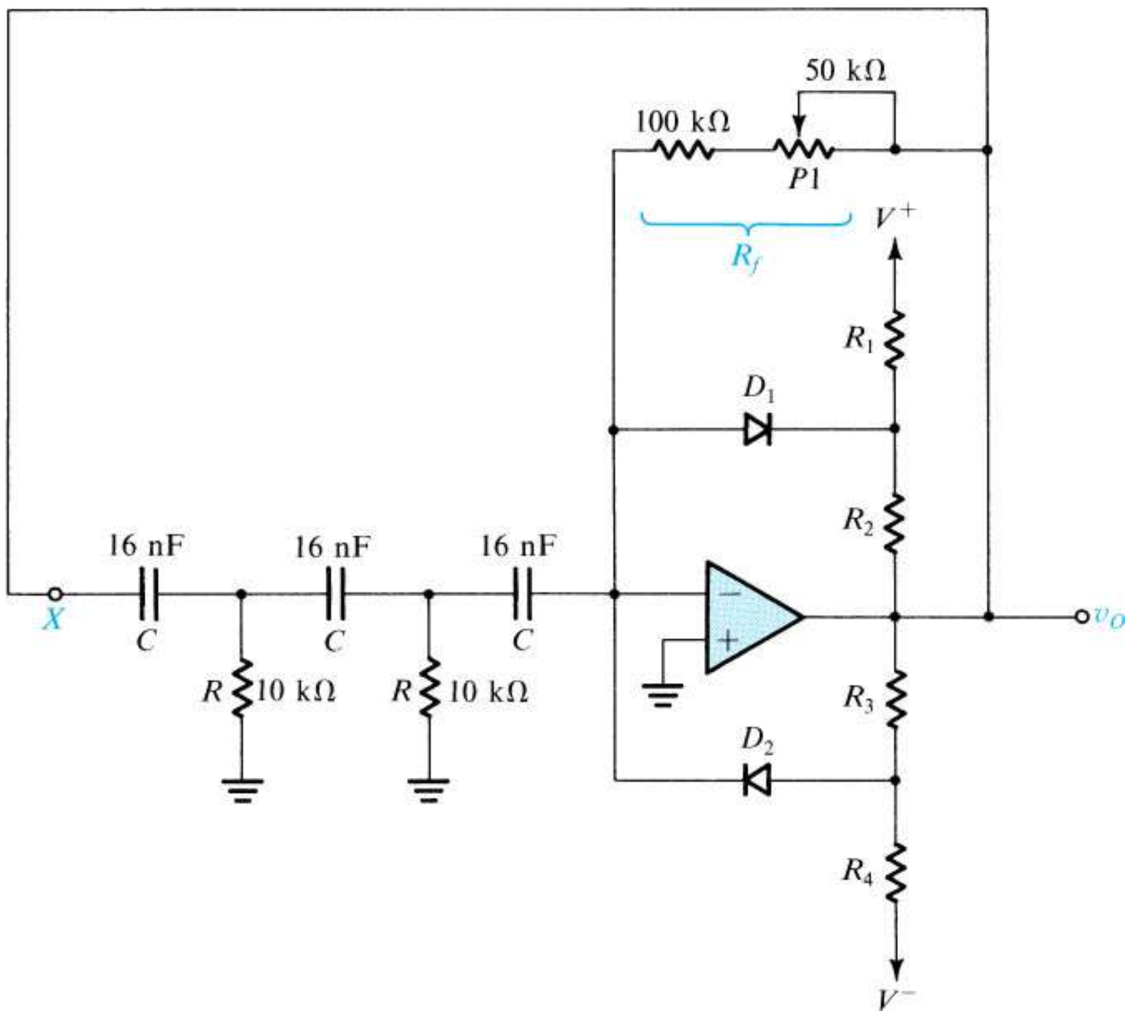
Alternate Wien bridge stabil.



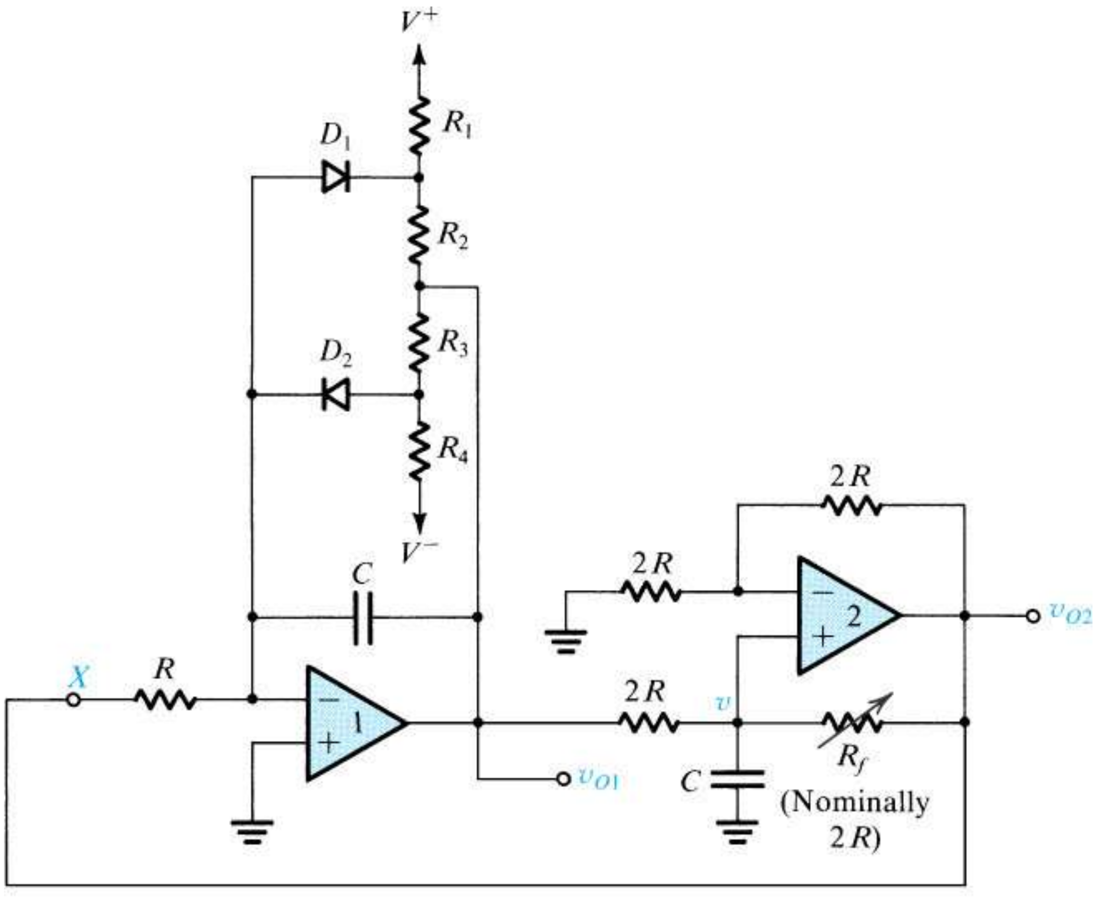
Phase Shift Oscillator



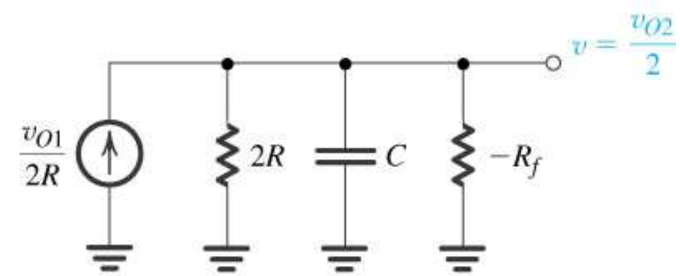
Phase Shift. Osc. W/ Stabil.



Quad Osc. Circuit

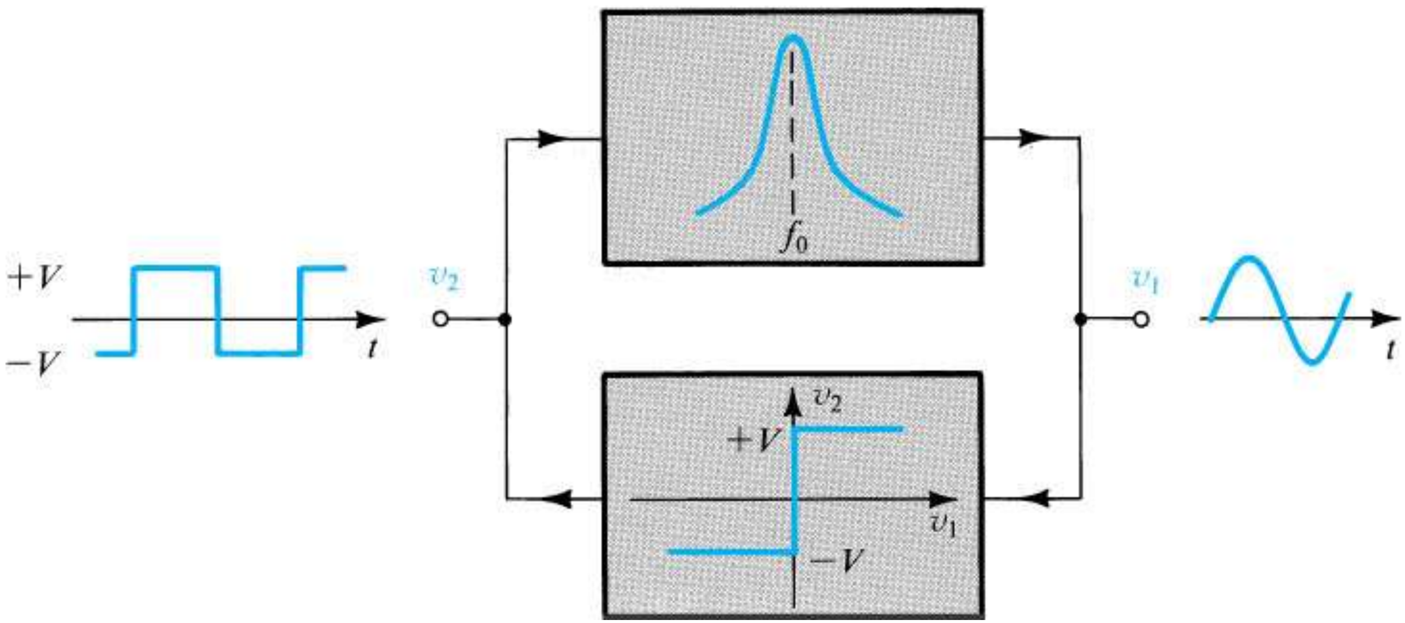


(a)

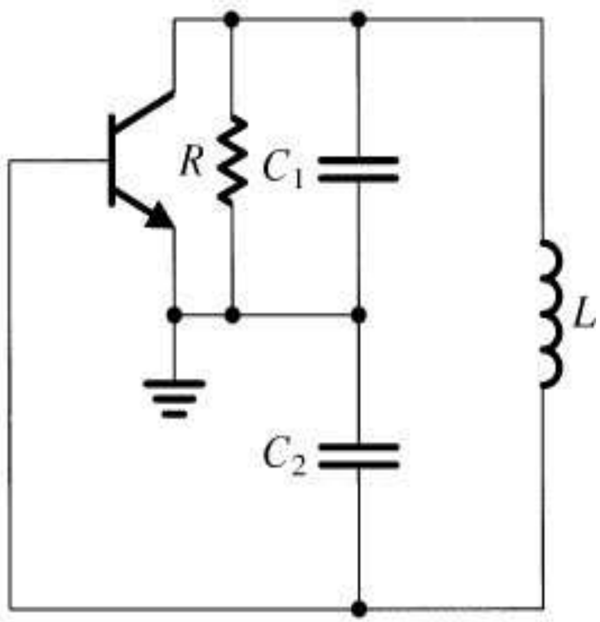


(b)

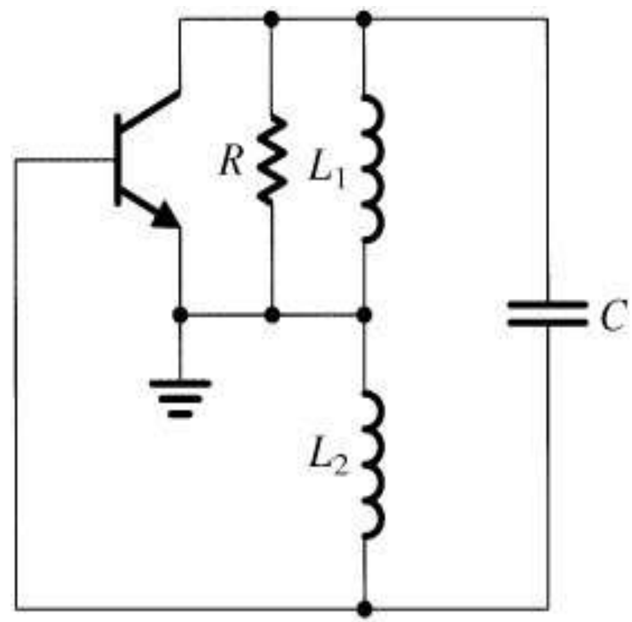
Active Tuned Osc.



Colpitts and Hartley Oscillators



(a)

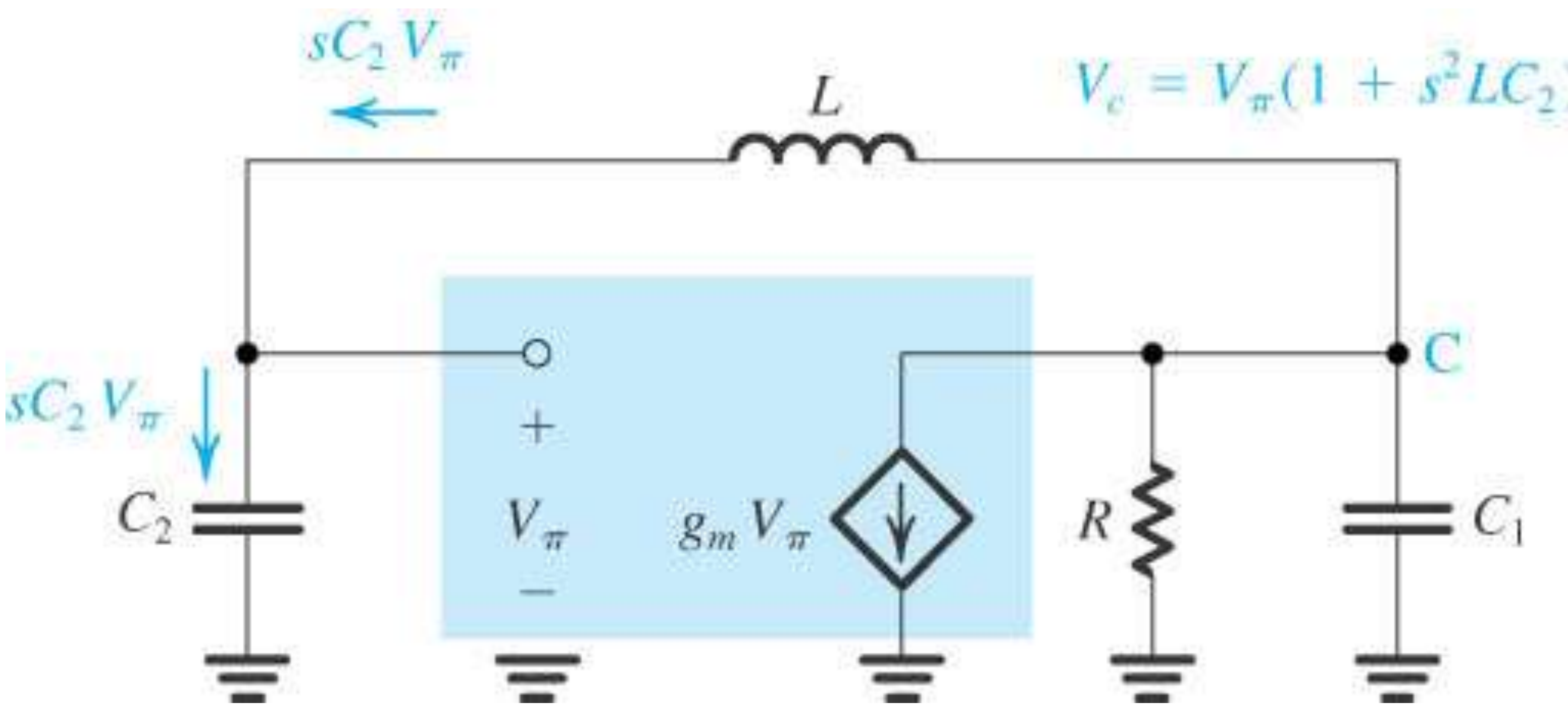


(b)

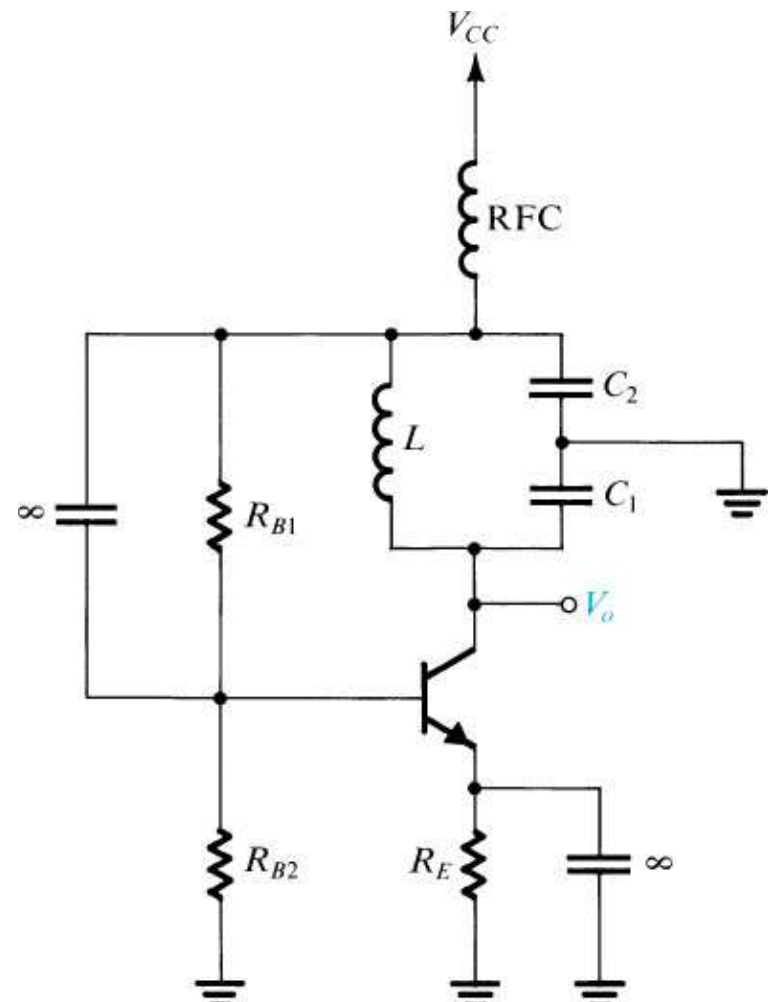
Equiv. Ckt

To simplify the analysis, negtlect *Cm* and *rp*

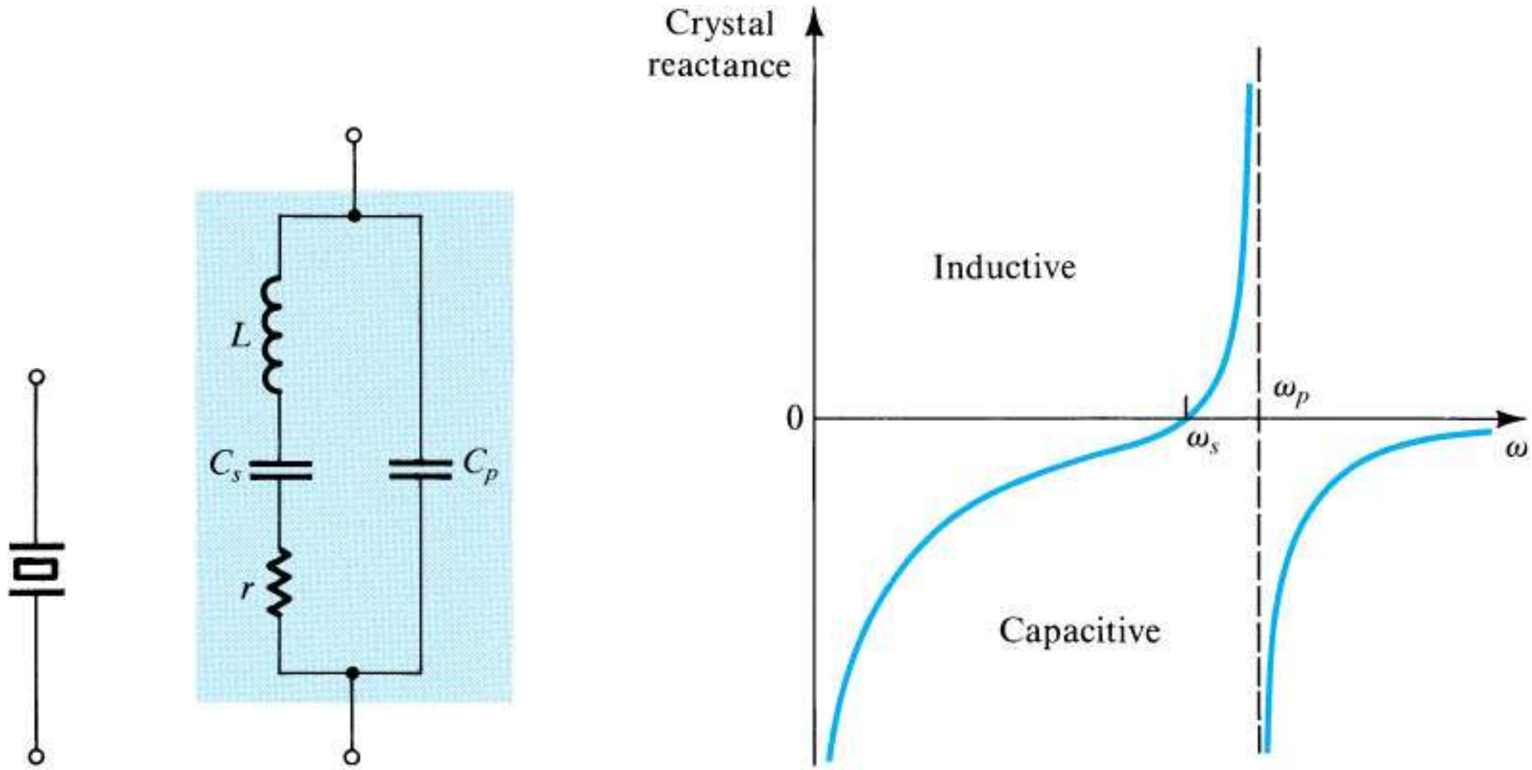
Consider *Cp* to be part of *C*2, and include *ro* in *R*.



Collpits Oscillator



Piezzoelectric Crystal



(a) (b) (c)