UNIT-1. SINGLE STAGE AMPLIFIERS.

An Electronic Amplifier circuit is one, which modifies the characteristics of input signal, when delivered the output side.

The modification of in characteristics of input signal can be with respect to voltage, current, power or phase. Any one or all the characteristics can be changed by amplifier circuit.

Classification of Amplifiers:

The amplifiers are classified in different ways as indicated below:

a) Based on Frequency range
b) Based on Type of coupling.
c) Based on power delivered / conduction angle.
d) Based on type of load.
e) Application.

a) Frequency range:

DC Amplifier	–	
Audio Frequency	–	40Hz – 15/20 KHz
Radio Frequency	–	>20KHz (30 - 300KHz)
Video Frequency	–	5- 8 MHz.
Very low Frequency	–	10 - 30KHz
low Frequency	–	30 - 300KHz
medium Frequency	–	300 - 3000KHz
High frequency	–	3- 30 MHz
Very High Frequency	–	30 - 300 MHz
Ultra High Frequency	–	300 - 3000, MHz
Super High Frequency	–	3,000 - 30,000 MHz.

Types of coupling:

* Direct coupled: Output of first stage is directly connected to the input of next stage. It does not block DC.
* RC coupled: Resistors and capacitors are used as coupling components. They block DC and gives
* Transformer coupled: Transformer is used as coupling component. It blocks DC and gives impedance matching.
* LC Tuned amplifiers, series fed

Output power delivered / conduction angle.

1. low power (ten of mW or less).
2. medium power (Hundreds of mW)
3. High power (Watts)

Class A (360°): An amplifier is said to be class A amplifier if the Q point and the input signal are selected such that the O/P signal is obtained for full input cycle.

Class B (180°): An amplifier is said to be class B amplifier if the Q-point and input signal are selected such that the O/P signal is obtained only for one half cycle for a full input cycle.

Class C (<180°): An amplifier is said to be class C amplifier if the Q-point and input signal are selected such that the O/P signal is obtained for less than half cycle for a full input cycle.

Class AB (180-360°): An amplifier is said to be class AB amplifier if the Q point and input signal are selected such that the O/P signal is obtained for more than 180° but less than 360° for a full input cycle.

type of load:

- Amplifier with resistive load
- Amplifier with inductive load

Based on Application:

> Voltage Amplifier – amplifies Voltage
  Current Amplifier – amplifies Current
  power Amplifier – amplifies both Voltage and Current
  Tuned Amplifier – Used for impedance matching.

Classification of Amplifiers

- Frequency Range:
  - * DC
  - * Audio
  - * Radio
  - * Video
  - * VLF
    - LF
    - MF
    - HF
    - VHF
    - UHF
    - SHF
- Coupling
  - * Direct
  - * RC
  - * Transformer
- O/p power delivered/ conduction angle
  - * class A
  - * class B
  - * class C
  - * class AB
- type of load.
  - * Resistive
  - * Inductive
- Application
  - * Voltage
  - * current
  - * power
  - * Tuned.

Distortion in Amplifiers

The i/p signals applied to amplifiers is alternating in nature. The basic features of any alternating signal are Amplitude, Frequency and phase.

The amplifier o/p should be reproduced faithfully i.e., there should not be a change or distortion in Amp, Freq. & phase. Hence the possible distortions in any amplifier are:

* Amplitude Distortion (or) non linear Dis.
* Frequency Distortion
* phase Distortion (or) Delay Dis.

Amplitude Distortion: This is also called Non linear or Harmonic distortion. It is due to the non linearity of characteristic of the device. This is due to the presence of frequency components in o/p, which are not present in input signal.

The component with same frequency as input signal is called fundemental frequency component. The additional freq components in o/p signal are having freq components which are integer multiples of fundemental freq comp. These components are called Harmonic Components or Harmonics. f - fundamental 2f - 2nd Harmonic, 3f - 3rd Harmonic. Out of all harmonics 2nd has largest amplitude.

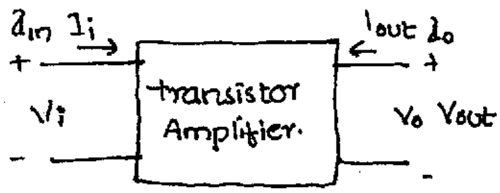
Frequency Distortion: This type of distortion exists when the signal components of different frequencies are amplified differently. This may be caused because of internal device capacitances or because the associated circuit is reactive.

Phase shift Delay Distortion: There will be phase shift between input and output signals & this phase shift will not be same for all frequencies. It also varies with frequency of input signal.

In o/p signal, all these distortions or any one may be present because of which amplifier response will not be good.

Hybrid model

let us consider a transistor amplifier as shown:



Here I_i = input current to amplifier
V_i = input voltage to amplifier
I_o = output current of amplifier
V_o = output voltage of amplifier.

Transistor is a current controlled device, input current is independent variable.

Input voltage and output current are dependent variables whereas output voltage and input current are independent variables.

Thus we can write.

$V_i = f_1(I_i, V_o)$
$I_o = f_2(I_i, V_o)$

These can be written in equation form as follows

$$V_i = h_{11} I_i + h_{12} V_o$$
$$I_o = h_{21} I_i + h_{22} V_o$$

The above equations can also be written using alphabetic notations.

$$V_i = h_i I_i + h_r V_o$$
$$I_o = h_f I_i + h_o V_o.$$

H- parameter Definitions:

$h_{11} = \left.\frac{V_i}{I_i}\right|_{V_o=0}$ Input resistance with output short circuited, in Ohms.

$h_{12} = \left.\frac{V_i}{V_o}\right|_{I_i=0}$ Reverse voltage transfer ratio with i/p open circuited.

$h_{21} = \left.\frac{I_o}{I_i}\right|_{V_o=0}$ Forward current transfer ratio or current gain with output short circuited.

$h_{22} = \left.\frac{I_o}{V_o}\right|_{I_i=0}$ output admittance with input open circuited in mhos.

All these four parameters are not same. they have different units. Otherwords, they are mixture of different units & hence referred to as hybrid parameters. For ac analysis h-parameters.

with o/p short circuited

$h_{11} = h_i$
$h_{21} = h_f$

with i/p open circuited

$h_{12} = h_r$
$h_{22} = h_o$

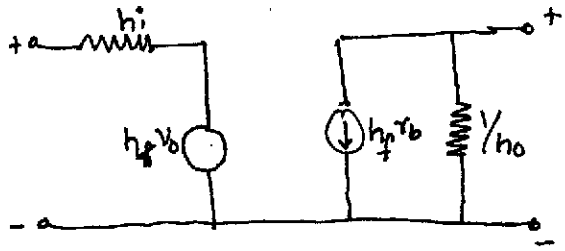
Transistor Hybrid model:

In order to analyze amplifier, calculate its input impedance, output impedance, current gain and voltage gain, it is necessary to replace the transistor circuit with its equivalent.

The equivalent circuit can be drawn with help of two equations,

$$V_i = h_i I_i + h_r V_o$$

$$I_o = h_f I_i + h_o V_o$$



H- Parameter equivalent circuit for CE configuration

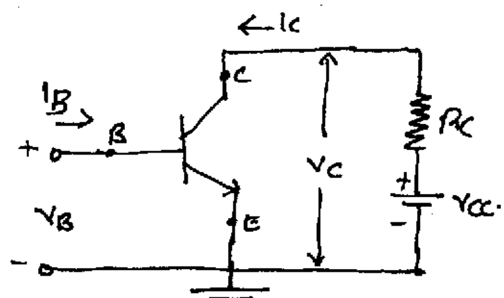
Common emitter configuration is shown in fig.

I_B = input current

I_C = output current

V_{BE} = input ~~voltage~~ voltage

V_{CE} = output voltage.

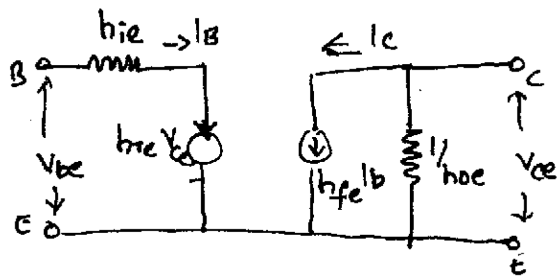


Simple CE Configuration.

The h- parameter equivalent circuit for CE configuration is shown in fig

$$V_{be} = h_{ie} I_b + h_{re} V_{ce}$$

$$I_c = h_{fe} I_b + h_{oe} V_{ce}$$



where

$$h_{ie} = \left.\frac{\Delta V_{be}}{\Delta I_b}\right|_{V_{CE} = \text{Const.}}$$

$$h_{re} = \left.\frac{\Delta V_{BE}}{\Delta V_{CE}}\right|_{I_B = \text{Const.}}$$

$$h_{fe} = \left.\frac{\Delta I_C}{\Delta I_B}\right|_{V_{CE} = \text{Const.}}$$

$$h_{oe} = \left.\frac{\Delta I_C}{\Delta V_{CE}}\right|_{I_B = \text{Const.}}$$

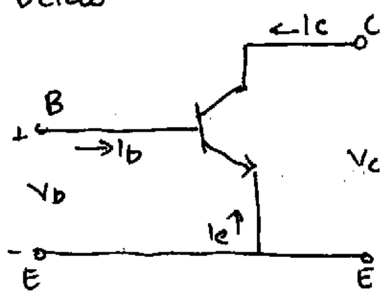
ΔV_{BE}, ΔV_{CE}, ΔI_B, ΔI_C represent the small change in base & collector voltages and currents.

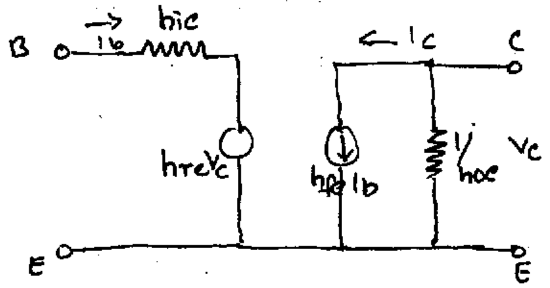
h- parameters for all three configurations.

Parameter	CB	CE	CC	CE	CC	CB
Input resistance	h_{ib}	h_{ie}	h_{ic}	1100	1100	21.6
Reverse voltage gain	h_{rb}	h_{re}	h_{rc}	2.5×10^{-4}	~1	2.9×10^{-4}
Forward current gain	h_{fb}	h_{fe}	h_{fc}	50	−51	−0.98
Output Admittance	h_{ob}	h_{oe}	h_{oc}	25 μA/V	25 μA/V	0.49 μA/V

The basic hybrid model for all the three configurations is same, only parameters are different.

The circuits and equations for three configurations is shown below.

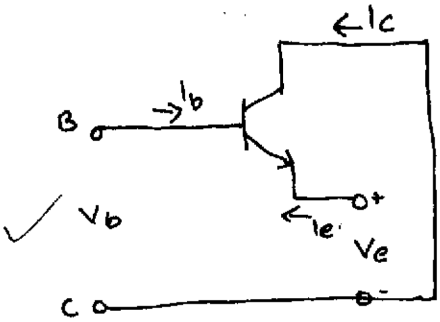


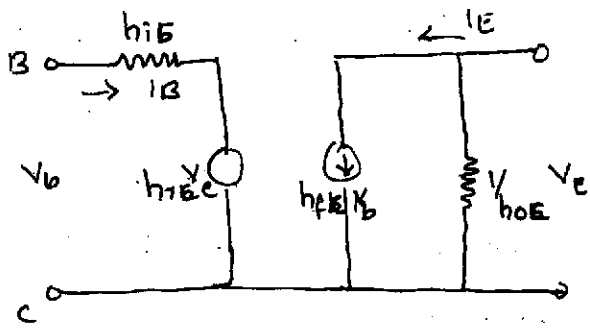


CE

$$V_b = h_{ie} I_b + h_{re} V_c$$

$$I_c = h_{fe} I_b + h_{oe} V_c$$

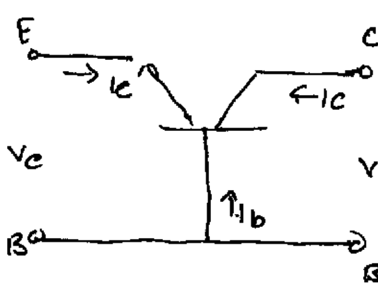


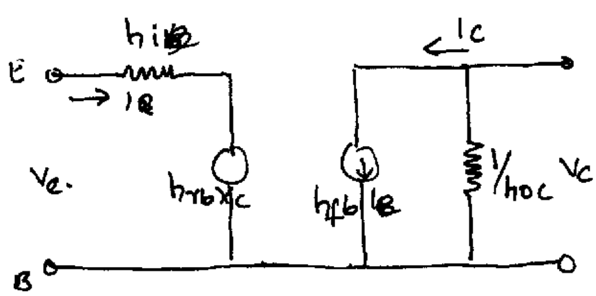


CC

$$V_b = h_{ic} I_b + h_{rc} V_e$$

$$I_c = h_{fc} I_b + h_{oc} V_e$$





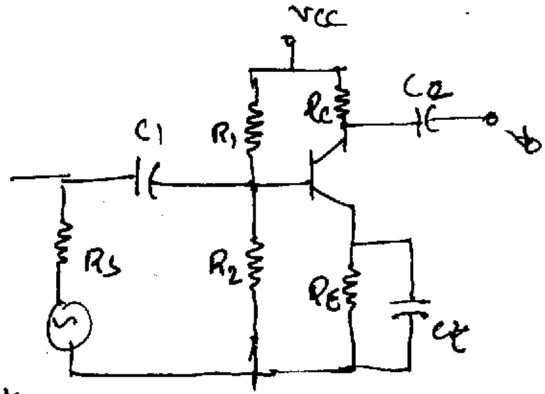
CB

$$V_e = h_{ib} I_e + h_{rb} V_c$$

$$I_c = h_{fb} I_e + h_{ob} V_c$$

small signal analysis of BJT Amplifiers:

In active devices, transistor is a non-linear one. To analyse this non linear characteristics is a difficult task. To make transistor in linear region by considering its o/p & analyze through small signal analysis.



Simple CE Amp. Circuit

The resistors R_1 R_2 R_E forms voltage divider bias for CE amp.

The capacitors C_1, C_C acts as coupling capacitors. C_1 is coupled to input, C_C is coupled to output voltage.

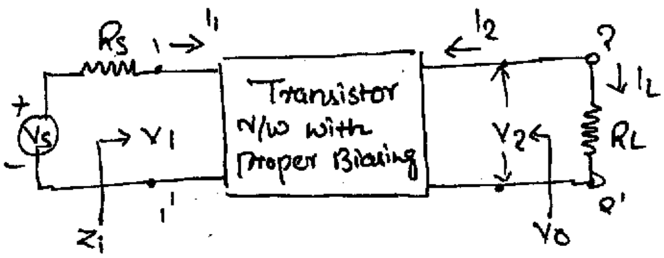
The purpose of these capacitors is to block any dc component present in signal & passes only ac signal for amplification.

The capacitor C_E acts as bypass capacitor. It is connected in parallel with R_E to provide low reactance path to o/p signal. If it is not inserted, the signal passing through R_E causes voltage drop across it which reduces the output voltage.

If AC signal is not present, only dc signal should pass through transistor, such condition is called zero bias condition or zero signal condition.

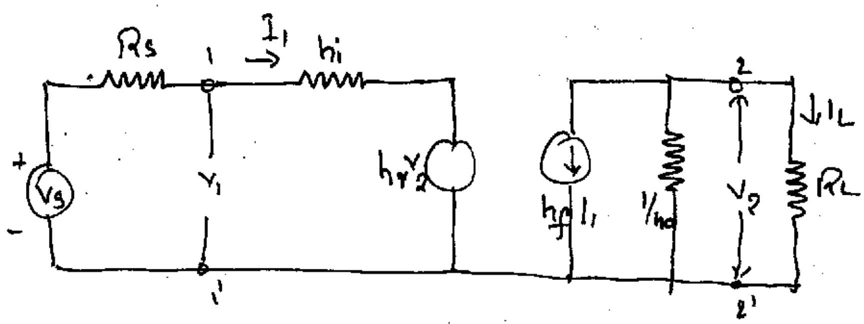
Small Signal Analysis of Transistor using h-parameters

Consider an Amplifier circuit as shown



This fig represents a transistor in any one of the configurations.

Replace the transistor circuit with its small signal Hybrid model



Current gain A_i

A_i is defined as ratio of output to input currents.

$$A_i = \frac{I_L}{I_1} = \frac{-I_2}{I_1}$$

Here I_L & I_2 are equal in magnitude but opposite in Directions.

$$I_2 = h_f I_1 + h_o V_2 \quad \text{—①}$$

Sub $V_2 = -I_2 R_L$ in eqn ①

$$I_2 = h_f I_1 + h_o(-I_2 R_L)$$

$$I_2 + I_2 h_o R_L = h_f I_1$$

$$I_2 (1 + h_o R_L) = h_f I_1$$

$$\therefore \frac{I_2}{I_1} = \frac{h_f}{1 + h_o R_L}$$

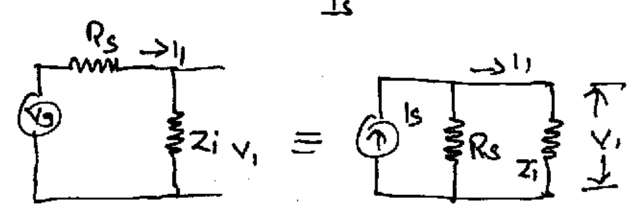
$$A_i = \frac{-I_2}{I_1} = \frac{-h_f}{1 + h_o R_L}$$

Current gain A_{is}

A_{is} takes into account the source resistance, R_s & the model is driven by current source.

$$A_{is} = \frac{-I_2}{I_s} = \frac{-I_2}{I_1} \cdot \frac{I_1}{I_s}$$

$$A_{is} = A_i \frac{I_1}{I_s}$$



Using current divider equation.

$$I_1 = \frac{I_s R_s}{Z_i + R_s}$$

$$\frac{I_1}{I_s} = \frac{R_s}{Z_i + R_s}$$

$$\therefore A_{is} = \frac{A_i \cdot R_s}{Z_i + R_s}$$

Input Impedance Z_i

$$Z_i = \frac{V_1}{I_1}$$

From circuit:

$$V_1 = h_i I_1 + h_r V_2.$$

$$Z_i = \frac{h_i I_1 + h_r V_2}{I_1}$$

$$Z_i = h_i + h_r \frac{V_2}{I_1}$$

$$Z_i = \frac{h_i + h_r (A_i I_1 R_L)}{I_1} \quad \left\{ \because V_2 = -I_2 R_L = A_i I_1 R_L. \right\}$$

$$Z_i = h_i + h_r A_i R_L$$

$$Z_i = h_i - \frac{h_r h_f R_L}{1 + h_o R_L} \quad \{ \because A_i = \}$$

Dividing Numerator & Denominator by R_L

$$Z_i = h_i - \frac{h_r h_f}{Y_L + h_o} \quad \text{where } Y_L = \frac{1}{R_L}$$

Output Admittance: Y_o

$$Y_o = \frac{I_2}{V_2}. \quad V_s = 0$$

$$I_2 = h_f I_1 + h_o V_2$$

$$\frac{I_2}{V_2} = \frac{h_f I_1}{V_2} + h_o$$

$$Y_o = h_f \frac{I_1}{V_2} + h_o \,//$$

$$R_s I_1 + h_i I_1 + h_r V_2 = 0 \ (V_s)$$

$$(R_s + h_i) I_1 = -h_r V_2.$$

$$\frac{I_1}{V_2} = \frac{-h_r}{R_s + h_i}$$

$$\therefore Y_o = \frac{-h_f h_r}{R_s + h_i} + h_o$$

$$Y_o = h_o - \frac{h_r h_f}{R_s + h_i} \,//$$

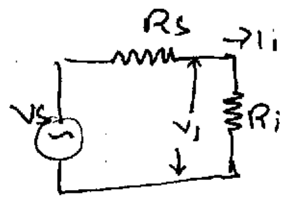
Voltage gain A_v:

$$A_v = \frac{V_2}{V_1}$$

$$A_v = \frac{A_i I_1 R_L}{V_1} = \frac{A_i R_L}{Z_i}$$

$$A_{vs} = \frac{V_2}{V_s} = \frac{V_2}{V_1} \times \frac{V_1}{V_s}$$

$$= A_v \frac{V_1}{V_s}.$$



Applying voltage divider theorem

$$V_1 = \frac{Z_i}{R_s + Z_i} V_s$$

$$\frac{V_1}{V_s} = \frac{Z_i}{R_s + Z_i}$$

$$\therefore A_{vs} = \frac{A_v Z_i}{R_s + Z_i} \,//$$

Power gain A_p

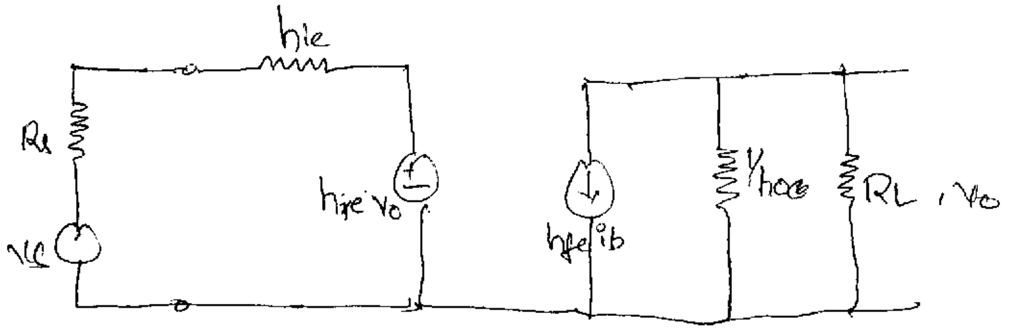
$$A_p = A_v \cdot A_I$$

$$= \frac{A_I I_1 R_L}{V_1} \cdot A_I$$

$$A_p = \frac{A_i^2 R_L}{Z_i} \,//$$

$$A_{vs} = \frac{A_{is} R_L}{R_s}$$

pblm.



$R_L = 10K$

$R_s = 1K$

$$A_i = \frac{-h_{fe}}{1 + h_{oe} R_L} = -40$$

$$R_i = h_{ie} + h_{re} A_i R_L \qquad 1K$$

$$A_{is} = \frac{A_i R_s}{R_i + R_s} \qquad -\cancel{400}\ 20$$

$$A_V = \frac{A_i R_L}{R_i} \qquad -400$$

$$A_{vs} = \frac{A_V R_i}{R_i + R_s} \qquad -200$$

$$R_o = \frac{1}{Y_o} = 52.6\,K\Omega$$

$$Y_o = h_{oe} - \frac{h_{fe} h_{re}}{h_{ie} + R_s} \qquad 19.0\,\mu A/V$$

x = a randomly or pseudorandomly generated integer with $0 < x < q$

USER'S PUBLIC KEY:

$y = g^x \bmod p$

USER'S PER-MESSAGE SECRET NUMBER:

k = a randomly or pseudorandomly generated integer with $0 < k < q$

The integers p, q, and g can be public and can be common to a group of users. A user's private and public keys are x and y, respectively. They are normally fixed for a period of time. Parameters x and k are used for signature generation only, and must be kept secret. Parameter k must be regenerated for each signature.

DIGITAL SIGNATURE GENERATION:

To begin with the process, a check (message) must be created. In order to create a digital signature with the check, a process known as "hash function must occur. The hash function is a mathematical algorithm that creates a digital representation or fingerprint in the form of a hash result or message digest. The hash function generally has a standard length that is usually much smaller than the message but nevertheless substantially unique to it. Hash functions ensure that there have been no modifications to the check since it was digitally signed.

The next step is to encrypt the check and signature. The sender' signature software transforms the result into a digital signature using the sender private key. The resulting signature is thus unique to both the message and the private key used to create it. Typically, a digital-signature is a appended to its message and stored or transmitted with the message. However, it may also be sent or stored as a separate data element, so long as it maintains a reliable association with its message. Since a digital signature is unique to its message, it is useless when wholly disassociated from the message.

Now the question arises how do one get a private and a public key? The answer is: You need to obtain something called a digital certificate. For that, you go to a certificate issuer, which will give you a digital certificate that says, in effect, "Here is Mike, and here is his public key. Anything he signs with his corresponding private key is valid." When you buy something online and digitally sign the transaction, you provide the merchant with your digital certificate. If the merchant trusts the issuer of the certificate, he uses the certificate to verify your signature. Often the authority that provides you with a digital certificate will also provide you with a private key. Certain computer systems will let you generate your own private key, but be careful! That is where the potential for fraud comes in. It's considered impossible to forge a digital signature the way one can forge a paper signature, but if you are careless with your private key—leaving it unprotected on your desktop, for instance—it's possible for you to compromise its integrity.

Mathematically

The signature of a message M is the pair of numbers r and s computed according to the equations below:

$r = (g^k \bmod p) \bmod q$ and

$s = (k^{-1}(SHA(M) + xr)) \bmod q$.

In the above, k^{-1} is the multiplicative inverse of k, mod q; i.e., $(k^{-1} k) \bmod q = 1$ and $0 < k^{-1} < q$. M is a message to be signed and the value of SHA(M) is a 160-bit string output by the Secure Hash Algorithm specified in FIPS 180. For use in computing s, this string must be converted to an integer.

As an option, one may wish to check if $r = 0$ or $s = 0$. If either $r = 0$ or $s = 0$, a new value of k should be generated and the signature should be recalculated (it is extremely unlikely that $r = 0$ or $s = 0$ if signatures are generated properly). The signature is transmitted along with the message to the verifier.

SIGNATURE VERIFICATION:

Prior to verifying the signature in a signed message, p, q and g plus the sender's public key and identity are made available to the verifier in an authenticated manner.

Let M', r' and s' be the received versions of M, r, and s, respectively, and let y be the public key of the signatory. To verifier first checks to see that $0 < r' < q$ and $0 < s' < q$; if either condition is violated the signature shall be rejected. If

Relation b/w A_{VS} & A_{IS}

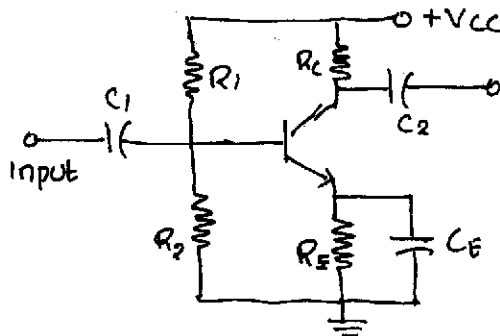
$$A_{VS} = \frac{A_I R_L}{Z_i + R_s} \quad \& \quad A_{IS} = \frac{A_i R_s}{Z_i + R_s}$$

$$\therefore \frac{A_{VS}}{A_{IS}} = \frac{R_L}{R_s} \Rightarrow A_{VS} = \frac{A_{IS} R_L}{R_s}$$

All these parameters are applicable to all configurations by changing subscript.

Guidelines for Analysis of Transistor circuit:

1. Draw the actual circuit diagram

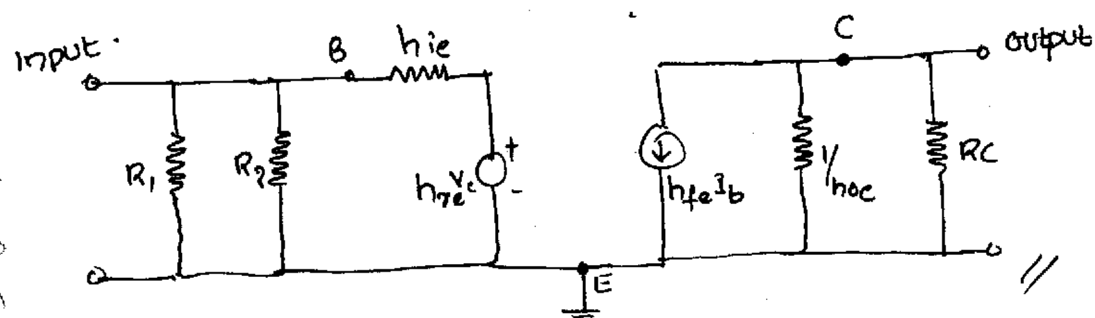


(1)

2. Replace coupling capacitors & emitter bypass capacitors by short circuit

3. & 4. Replace DC source by SC. in other words short Vcc and ground. mark B, C, E points on circuit diagram & locate these points as start of equivalent circuit

5. Replace the transistor by h-parameter model.



//

	CE	CC	
h_i	1100 Ω	1100	
h_r	2.5×10^{-4}	~1	
h_f	50	−51	21.6
h_o	25 μA/V	25 μA/V	2.4×10^{-4}
			−0.98
It			0.49 μA/V

∴
$$A_i = \frac{-h_f}{1 + h_o R_L}$$

$$A_{is} = \frac{A_i R_s}{Z_i + R_s}$$

$$Z_i = h_i + h_r A_i R_L = h_i - \frac{h_f h_r}{h_o + Y_L}$$

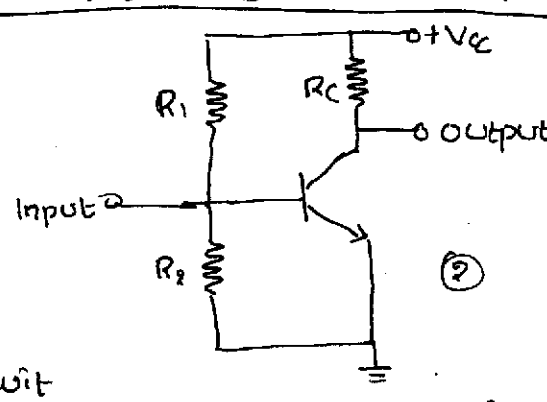
$$A_V = \frac{A_i R_L}{Z_i}$$

$$A_{VS} = \frac{A_{is} R_L}{R_s}$$

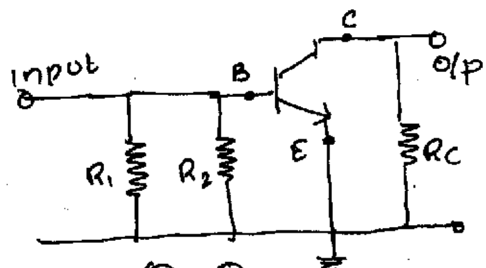
$$Y_o = h_o - \frac{h_f h_r}{h_i + R_s}$$

$$A_p = A_i^2 \frac{R_L}{Z_i}$$

A: If R_L = 10K R_s = 1K. A_i, A_v ?



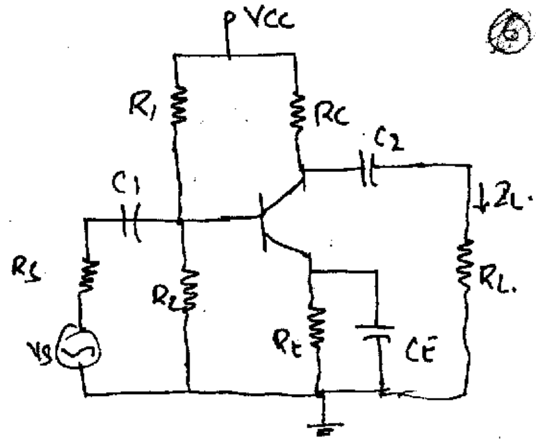
(2)

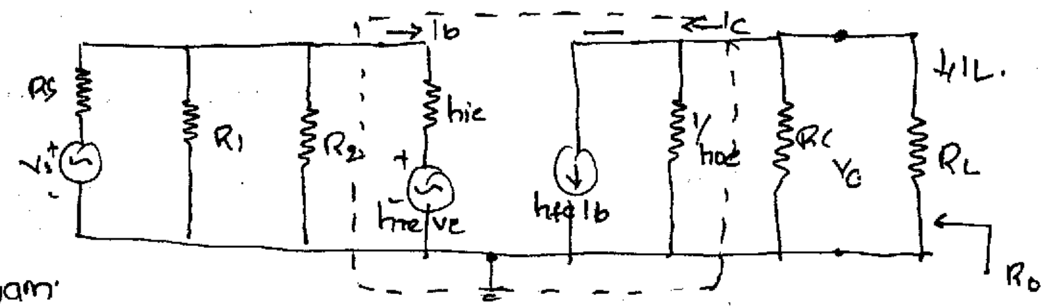


(3) & (4)

1 Consider a Single stage CE with $R_S = 1\,k\Omega$, $R_1 = 50\,k\Omega$, $R_2 = 2\,k\Omega$, $R_C = 1\,k\Omega$, $R_L = 1.2\,k\Omega$, $h_{fe} = 50$, $h_{ie} = 1.1\,K$, $h_{oe} = 25\,\mu A/V$, $h_{re} = 2.5\times10^{-4}$.

Calculate $A_I, R_i, A_V, A_{VS}, R_o$.





current gain

$$A_i = \frac{I_L}{I_b} = -\frac{I_c}{I_b} = \frac{-h_{fe}}{1+h_{oe}R_L'} \qquad R_L' = R_C \| R_L$$

$$= 1k \| 1.2k = 545.45\,\Omega.$$

$$A_i = \frac{-50}{1+25\mu A/V\ 545.45} = -49.32$$

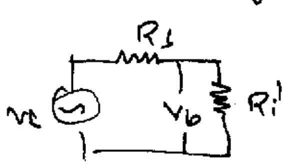
input impedance $R_i = h_{ie} + h_{re} A_i R_L'$

$$= 1098\,\Omega$$

Voltage gain $A_V = \frac{V_c}{V_b} = \frac{A_i R_L'}{R_i} = \frac{-49.32\times545.45}{1093} = -24.61$

Overall input resistance $R_i' = R_i \| R_1 \| R_2 = 696.9\,\Omega.$

Overall voltage gain $A_{VS} = \frac{V_c}{V_s} = \frac{V_c}{V_b}\times\frac{V_b}{V_s}$

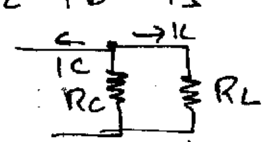


$$V_b = \frac{V_s R_i'}{R_s+R_i'} \qquad = \frac{V_b}{V_s} = \frac{R_i'}{R_s+R_i'}$$

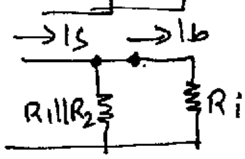
$$A_{VS} = A_V \cdot \frac{R_i'}{R_s+R_i'} = 10.1$$

$$A_{is} = \frac{I_L}{I_s} = \frac{I_L}{I_c}\times\frac{I_c}{I_b}\times\frac{I_b}{I_s}$$

$$I_L = \frac{-I_c R_c}{R_c+R_L}$$



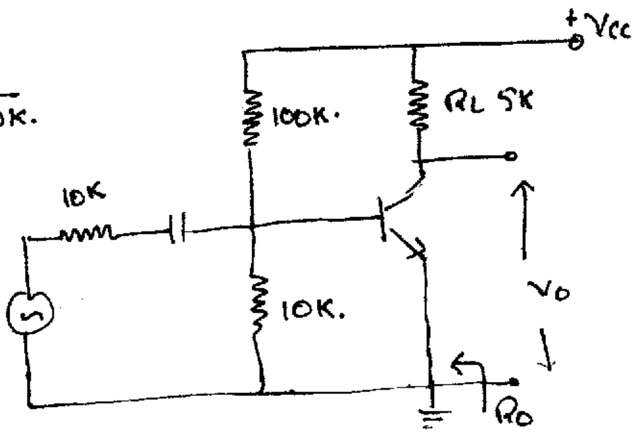
$$I_b = \frac{I_s R_B}{R_B+R_i}$$



$\therefore\ A_i = -14.29.$

$R_B = R_1 \| R_2.$

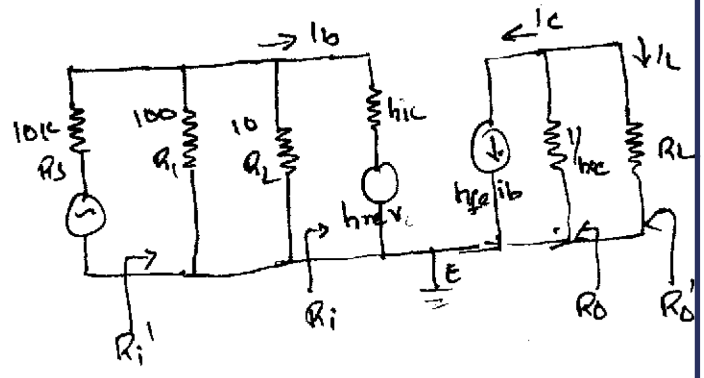
$h_{fe} = 50 \quad h_{re} = 2.5\times10^{-4} \quad h_{ie} = 1100\,\Omega \quad h_{oe} = \frac{1}{40K}.$



Current gain $A_i = \frac{-I_c}{I_b}$

$$A_i = \frac{-h_{fe}}{1+h_{oe}R_L} = \frac{-50}{1+\frac{1}{40000}\times 5000}$$

$$A_i = -44.44$$



Input Impedance.

$$R_i' = R_i \| R_B$$

$$R_i = h_{ie} + h_{re} A_i R_L$$

$$= 1100 + 25\times10^{-4}(-44.44)\cdot 5000$$

$$R_i = 1044.4\,\Omega$$

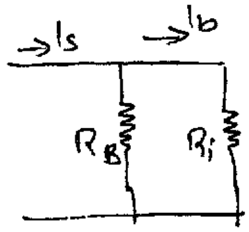
$$R_i' = R_i \| R_B$$

$$R_B = R_1 \| R_2 = 9.09K.$$

$$R_i' = 1044.4 \| 9.09K = 0.9367\,K\Omega$$

$$A_{is} = \frac{I_L}{I_s} = \frac{-I_c}{I_s} = \frac{-I_c}{I_b}\times\frac{I_b}{I_s}$$

$$\frac{-I_c}{I_b} = (-44.44)$$



Using current divider formula.

$$I_b = \frac{I_s R_B}{R_B + R_i'} \qquad R_B = R_1 \| R_2$$

$$\frac{I_b}{I_s} = \frac{R_B}{R_B + R_i'} = 0.8968$$

$$A_{is} = 0.8968 \times (-44.44) = -39.856.$$

$\underline{A_{vs}}$:

$$A_v = \frac{V_o}{V_b} = \frac{A_i R_L}{R_i} = \frac{-44.44 \times 5000}{1044.44} = -212.8.$$

$$A_{vs} = \frac{V_o}{V_s} = \frac{V_o}{V_b} \times \frac{V_b}{V_s}$$

$$= -212.8 \times 0.08564$$

$$= -18.22$$

voltage divider formula

$$V_b = \frac{V_s R_i'}{R_i' + R_s}$$

$$\frac{V_i}{V_s} = 0.08564$$

$$Y_o = h_o - \frac{h_r h_f}{h_i + R_s'}$$

$R_s' = R_1 || R_2 || R_s + h_{ie}$

$$= 22.87 \times 10^{-6}$$

$$R_o = \frac{1}{Y_o} = 43.73 K\Omega$$

$$R_o' = R_o || R_L = 4487\Omega$$

$R_s = 2K$

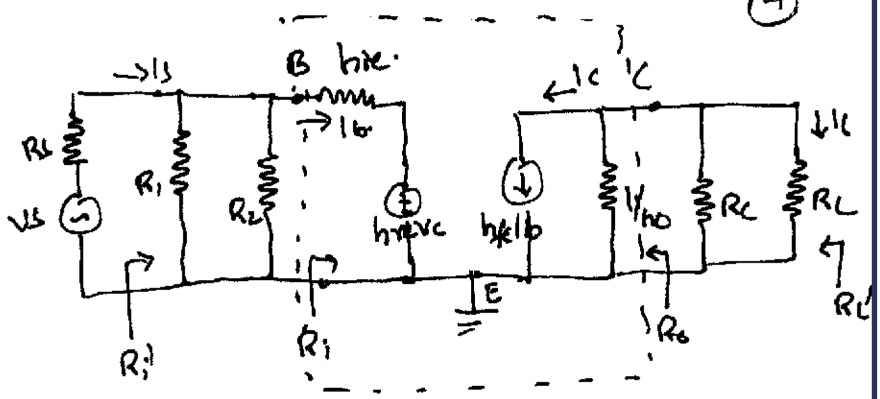
$R_1 = 100K$

$R_2 = 4K$

$R_C = 2K$

$R_L = 2.2K$.

H/W



$$A_i = \frac{-h_{fe}}{1 + h_{oe} R_L'}$$

$R_L' = R_C \| R_L$

$$R_L' = \frac{2.2K \times 2K}{4.2K} = 1.04 K\Omega$$

$$A_i = \frac{-50}{1 + 25\times10^{-6} \times 1040} = \frac{-50}{\cancel{\phi}.0548}$$

$A_i = -47.402$ //

$R_i = h_{ie} + h_{re} A_i R_L'$

$= 1100 + 2.5\times10^{-4}(-47.\cancel{40})\times 1.04\times10^{3}$

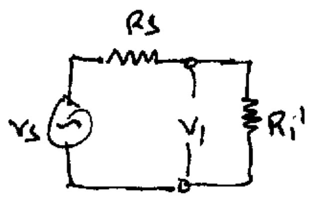
$= 1100 - 12.324 = 1.087 K\Omega$ //

$$A_V = \frac{A_i R_L'}{R_i} = \frac{(-47.402)(1.04\times10\cancel{00})}{1.087\times10\cancel{00}}$$

$A_V = -45.352$ //

$$A_{Vs} = \frac{V_2}{V_s} = \frac{V_2}{V_1} \times \frac{V_1}{V_s}$$

$$A_V \times \frac{V_1}{V_s}$$



$$\frac{V_1}{V_s} = \frac{V_s R_i'}{R_i' + R_s}$$

$$\frac{V_1}{V_s} = \frac{R_i'}{R_i' + R_s} \quad R_i' = R_1 \| R_2 \| R_i$$

$\therefore A_{Vs} = (-45.352)$ $\quad R_1 \| R_2 = 3.846K$.

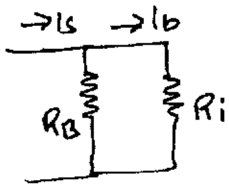
$R_i' = 3.846K \| 1.087K = 847.47\,\Omega$

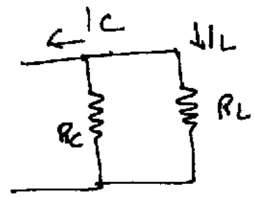
$$\therefore A_{Vs} = -45.352 \times \frac{847.47}{847.47 + 2000\,\Omega}$$

$A_{Vs} = -13.49$ //

$$A_{is} = \frac{I_L}{I_s} = \cancel{\frac{-I_C}{I_S}} = \cancel{\frac{-I_C}{I_b}} \times \frac{I_b}{I_s}$$

$$= \frac{I_L}{I_C} \times \frac{I_C}{I_b} \times \frac{I_b}{I_s}$$





$$I_b = \frac{I_s R_B}{R_i + R_B}$$

$$\frac{I_b}{I_s} = \frac{R_B}{R_i + R_B}$$

$$= \frac{3.846K}{1.087K + 3.846K}$$

$= 0.7796$.

$$I_L = \frac{-I_C R_C}{R_L + R_C}$$

$$\frac{I_L}{I_C} = \frac{R_C}{R_L + R_C}$$

$$= \frac{2K}{4.2K}$$

$= 0.476$

$$\therefore \frac{I_L}{I_C} \times \frac{I_C}{I_b} \times \frac{I_b}{I_s}$$

$(0.476) \times (-47.402)(0.7796)$

$= -17.59$ //

group of 23 or more people the probability that two or more people share the same birthday is better than 50%.

When software (code) is associated with publisher's unique signature, distributing software on the Internet is no longer an anonymous activity. Digital signatures ensure accountability, just as a manufacturer's brand name does on packaged software. If an organization or individual wants to use the Internet to distribute software, they should be willing to take responsibility for that software. This is based on the premise that accountability is a deterrent to the distribution of harmful code.

APPROACHES

A variety of approaches have been proposed for digital signature function. These approaches fall into two categories:

- Direct approach
- Arbitrated approach

Direct digital signature:

A direct digital signature involves only the communication parties (source and destination). It is assumed that the destination knows the public key of the source. A digital signature may be formed by encrypting the entire message with the sender's private key or by encrypting the hash code of the message with the sender's private key.

Confidentiality can be provided by further encrypting the entire message plus signature with either the receiver's public key or a shared secret key. It is important to perform the signature function first and then an outer confidentiality function. In case of dispute some third party must view the message and signature. If the signature is calculated on an encrypted message, the third party also needs access to the decryption key to read the original message.

All direct schemes described so far have a common flaw

The validity of the scheme depends on the security of the sender's private key. If a sender later wishes to deny sending a particular message , he can claim that the private key was lost or stolen and that someone else forged his signature.

Administrative controls relating to the security of private keys can be employed to thwart or at least weaken this ploy. One example is to require every signed message to include a timestamp (date and time) and to require prompt reporting to compromised keys by a central authority. Another threat is that the private key might be stolen from sender X at time T. The opponent can then send a message signed with X's signature and stamped with a time before or equal to T.

Arbitrated digital signature:

The problems associated with direct digital signatures can be addressed by using an arbiter. As with direct signature schemes, there are a variety of arbitrated signature schemes. In general terms, these all operate as follows: every signed message from sender X to the receiver Y goes first to the arbiter A, who subjects the message and its signature to the number of tests to check its origin and content. The message is then dated and send to Y with an indication that it has been verified to the satisfaction of the arbiter. With the presence of arbiter A, there are no chances of a sender X to disowning the message, as is the case with the direct digital signatures.

The arbiter plays a crucial role in arbitrated digital signatures and all parties must have a great deal of trust that the arbitration mechanism working properly. The use of a trusted system might satisfy this requirement.

HOW THE TECHNOLOGY WORKS :

Digital signatures require the use of public-key cryptography .If you are going to sign something , digitally, you need to obtain both a public key and a private key. The private key is something you keep entirely to yourself. You sign the document using your private key- which is really just a kind of code-then you give the person (the merchant of the website where you bought something or the bank lending your money to buy a house) who needs to verify your signature your corresponding public key. He uses your public key to make sure you are who you say you are. The public key and private key are related, but only mathematically , so knowing your private key, in fact, it's nearly impossible to figure out your private key from your public key

Symbol	Common Emitter	Common collector	Common base.
h_{ie}	$1{,}100\ \Omega$	h_{ic}^{*}	$\frac{h_{ib}}{1+h_{fb}}$
h_{re}	25×10^{-4}	$1-h_{rc}^{*}$	$\frac{h_{ib}h_{ob}}{1+h_{fb}}-h_{rb}$
h_{fe}	50	$-(1+h_{fc})^{*}$	$-\frac{h_{fb}}{1+h_{fb}}$
h_{oe}	$25\ \mu A/V$	h_{oc}^{*}	$\frac{h_{ob}}{1+h_{fb}}$
h_{ib}	$\frac{h_{ie}}{1+h_{fe}}$	$\frac{-h_{ic}}{h_{fc}}$	$21.6\ \Omega$
h_{rb}	$\frac{h_{ie}h_{oe}}{1+h_{fe}}-h_{re}$	$h_{fc}-\frac{h_{ic}h_{oc}}{h_{fc}}-1$	29×10^{-4}
h_{fb}	$-\frac{h_{fe}}{1+h_{fe}}$	$-\frac{1+h_{fc}}{h_{fc}}$	-0.98
h_{ob}	$\frac{h_{oe}}{1+h_{fe}}$	$\frac{-h_{oc}}{h_{fc}}$	$0.49\ \mu A/V$
h_{ic}	h_{ie}^{*}	$1{,}100\ \Omega$	$\frac{h_{ib}}{1+h_{fb}}$
h_{rc}	$1-h_{re}\approx1^{*}$	1	1
h_{fc}	$-(1+h_{fe})^{*}$	-51	$-\frac{1}{1+h_{fb}}$
h_{oc}	h_{oe}^{*}	$25\ \mu A/V$	$\frac{h_{ob}}{1+h_{fb}}$

conversion formulas:

CE		CC	CB
h_{ie}	$1100\,\Omega$	$h_{ic} = h_{ie}$	$h_{ib} = \frac{h_{ie}}{1+h_{fe}}$
h_{re}	2.5×10^{-4}	$h_{rc} = 1$	$h_{rb} = \frac{h_{ie}h_{oe}}{1+h_{fe}} - h_{re}$
h_{fe}	$50\,\Omega$	$h_{fc} = -(1+h_{fe})$	$h_{fb} = \frac{-h_{fe}}{1+h_{fe}}$
h_{oe}	$25\,\mu A/V$	$h_{oc} = h_{oe}$	$h_{ob} = \frac{h_{oe}}{1+h_{fe}}$

CB circuit determine A_i, R_i with typical values of CE h parameters.

h_{ib}

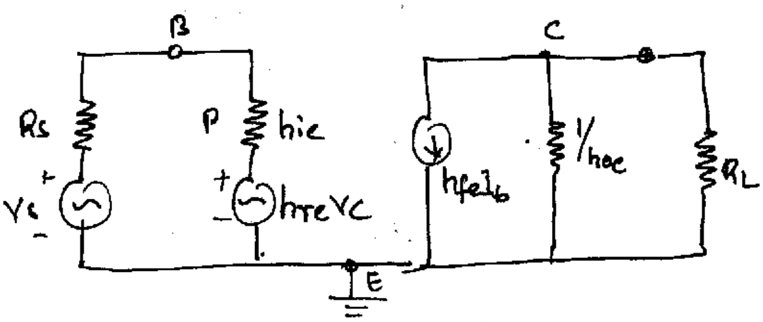
Analysis of CE, CB, CC configurations with simplified Hybrid model

Till now, we have seen the exact calculations of A_i, A_v, R_i, R_o of transistor circuits. But in most practical cases it is appropriate values of A_i A_v R_o R_i rather than to carry out more lengthy exact calculations.

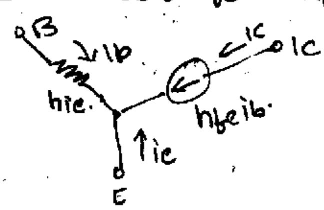
When to use approximate analysis?

There is a generalized rule that says if $h_{oe} \cdot R_L < 0.1$ then we can proceed for approximate analysis, otherwise do exact analysis.

Simplified CE Hybrid model



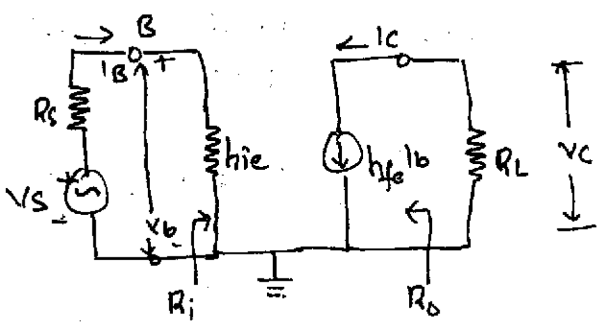
let us consider h-parameter equivalent circuit for amplifier.



Since $1/h_{oe}$ is in parallel with R_L & R_C if $1/h_{oe} >> R_L \| R_C$ we can neglect h_{oe}.

(Since h_{oe}, h_{re} are very small values, we neglect them for simplicity)

The approximate CE model is shown below



Current gain: $A_i = \frac{-I_c}{I_b} = \frac{-h_{fe}}{1+h_{oe}R_L}$

By neglecting h_{oe} $\boxed{A_i = -h_{fe}}$

Input impedance: $R_i = h_{ie} + h_{re} A_i R_L$

By neglecting h_{re} $R_i = h_{ie}$.

Voltage gain:

$$A_v = \frac{A_i R_L}{R_i} = \frac{A_i R_L}{h_{ie}}$$

Output impedance:

$$Y_o = h_{oe} - \frac{h_{fe} h_{re}}{h_{ie}+R_s}$$

neglecting h_{oe} & h_{re}

$Y_o = 0$

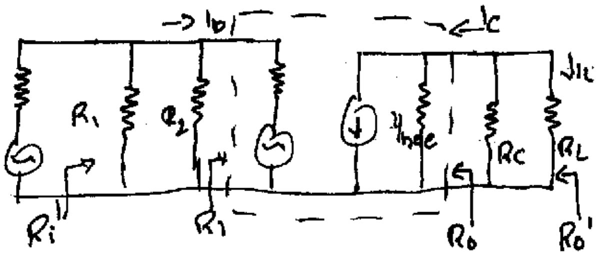
$R_o = \frac{1}{Y_o} = \infty$

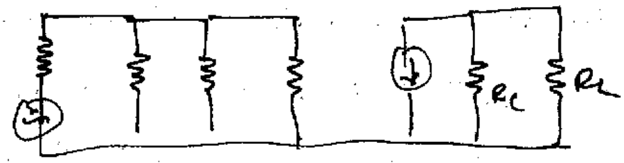
$R_o' = R_o \| R_L = R_L$

Go for Example in 10

Consider single stage CE amp with $R_s = 1K$, $R_1 = 50K$, $R_2 = 2K$, $R_C = 2K$, $R_L = 2K$

$h_{fe} = 50$, $h_{ie} = 1.1K$, $h_{re} = 2.5\times10^{-4}$, $h_{oe} = 25\,\mu A/V$. Find A_I, R_i, A_V, R_o, ~~A_{IS} A_{VS}~~ A_{IS}, A_{VS}





$$A_I = \frac{-h_{fe}}{1+h_{oe}R_L'}$$

$$R_L' = R_C \| R_L = 1K$$

$$= -\frac{50}{1+(25\times10^{-6}\times10^{3})} = \frac{-50}{1.025} = -48.78$$

$$R_i = h_{ie} + h_{re}A_I R_L'$$

$$= 1.1K + (2.5\times10^{-4})(-48.78)\times1K.$$

$$R_i = 1.1K - 12.195 = 1087.805\,\Omega$$

$$A_V = \frac{A_I R_L'}{R_i} = \frac{(-48.78)\times1000}{1087.805} = -44.84$$

$$A_{IS} = \frac{I_L}{I_s} =$$

$$Y_o = h_{oe} - \frac{h_{re}h_{fe}}{h_{ie}+R_s}$$

$$= 25\times10^{-6} - \frac{2.5\times10^{-4}\times50}{1100+1000}$$

$$= 2.5\times10^{-6} - 0.059\times10^{-4}$$

$$= +3.4\times10^{-6}$$

$$R_o = 294\,K\Omega$$

$$R_o' = R_o \| R_L'$$

$$= 1K\Omega$$

$$A_{IS} = \frac{I_L}{I_C}\times\frac{I_C}{I_b}\times\frac{I_b}{I_s} = -15.9$$

$$A_{VS} = \frac{V_C}{V_b}\times\frac{V_b}{V_s}$$

$$A_{VS} = A_V\times\frac{R_i'}{R_s+R_i'} = -18.71$$

$$A_I = -h_{fe} = -50$$

$$R_i = h_{ie} = 1.1K.$$

$$R_i' = h_{ie}\|R_1\|R_2 = 700\,\Omega$$

$$A_V = \frac{A_I R_L'}{R_i} = \frac{-50\times(2K\|2K)}{1.1K} = -45.45$$

$$R_o = \frac{1}{Y_o} = \infty$$

$$R_o' = R_o\|R_L' = 1K$$

$$A_{VS} = \frac{V_o}{V_s} = \frac{V_o}{V_b}\times\frac{V_b}{V_s}$$

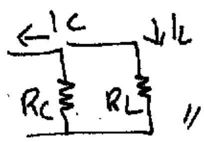
$$A_V = \frac{V_o}{V_b} =$$

$$\frac{V_b}{V_s} = \frac{R_i'}{R_i'+R_s}$$

$$A_{VS} = \frac{A_V R_i'}{R_i'+R_s} = -18.71$$

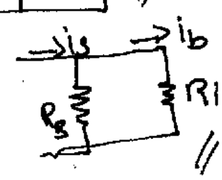
$$A_{IS} = \frac{I_L}{I_s} = \frac{I_L}{I_C}\times\frac{I_C}{I_b}\times\frac{I_b}{I_s}$$

$$\frac{I_L}{I_C} = -\frac{R_C}{R_C+R_L} = -0.5$$



$$\frac{I_C}{I_b} = h_{fe} = 50.$$

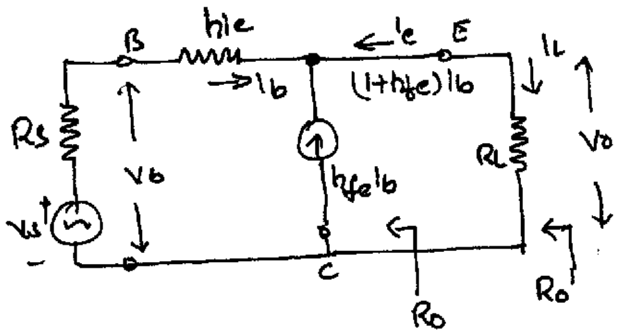
$$\frac{I_b}{I_s} = \frac{R_B}{R_B+R_i} = 0.636$$



$$\frac{I_L}{I_s} = -0.5\times50\times0.636$$

$$= -15.9$$

Simplified CC model



$h_{fe}I_b$ direction is exactly opposite that of CE model. because the current $h_{fe}I_b$ always points towards emitter

Current gain: Apply KCL

$$A_i = \frac{I_o}{I_b} = \frac{-I_e}{I_b} = 1+h_{fe}$$

$$I_e + I_c + I_b = 0$$

$$I_e = -h_{fe}I_b - I_b$$

$$I_e = -I_b(1+h_{fe})$$

$$\frac{-I_e}{I_b} = (1+h_{fe})$$

Ri: $R_i = \frac{V_b}{I_b}$

$$V_b - h_{ie}I_b - I_oR_L = 0$$

$$V_b = I_bh_{ie} + I_oR_L$$

$$\frac{V_b}{I_b} = h_{ie} + \frac{I_o}{I_b} \cdot R_L$$

$$R_i = h_{ie} + (1+h_{fe})R_L$$

$V_1 = h_i \cancel{I_1} + h_r V_2$

I_oR_L

Go for Example in (10)

Av: $A_v = \frac{V_o}{V_i} = \frac{A_iR_L}{R_i}$

$$A_v = \frac{(1+h_{fe})R_L}{h_{ie}+(1+h_{fe})R_L}$$ always < 1

$\because (1+h_{fe})R_L \gg h_{ie}$.

Ro: $R_o = \frac{V_o}{I_e}\Big|_{V_s=0}$.

KVL to outside ~~o/p loop~~

$$V_s - I_bR_s - I_bh_{ie} - V_o = 0$$

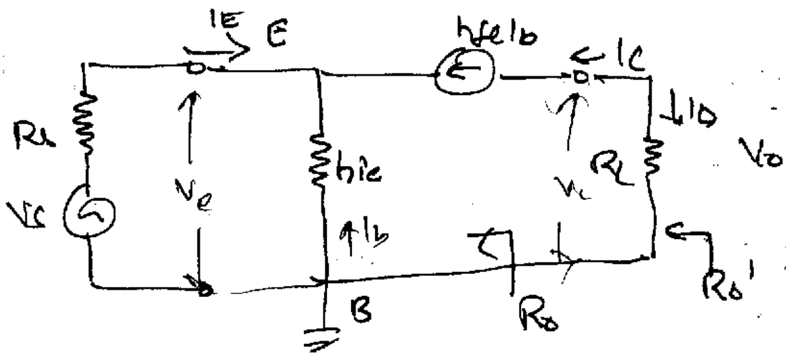
$$V_o = -I_bR_s - I_bh_{ie}$$

$$i_e = -(1+h_{fe})I_b$$

$$R_o = \frac{V_o}{I_e} = \frac{R_s+h_{ie}}{1+h_{fe}}$$

$$R_o' = R_o \| R_L$$

Common Base



$I_e = I_c + I_b$

$h_{fe} I_b + I_b$

$I_e = (h_{fe}+1) I_b$

current gain:

$$A_I = \frac{I_0}{I_e} = \frac{-I_c}{I_e} = \frac{-h_{fe} I_b}{-(1+h_{fe}) I_b} = \frac{h_{fe}}{1+h_{fe}} \quad \text{always} < 1$$

Input impedance:

$$R_i = \frac{V_e}{I_e} = \frac{\cancel{-h_{fe} I_b}}{\cancel{-(1+h_{fe})} I_b} = \frac{\cancel{h_{fe}}}{1\cancel{+h_{fe}}}$$

$V_e = -h_{ie} I_b \qquad I_e = -(1+h_{fe}) I_b$

$$R_i = \frac{h_{ie}}{1+h_{fe}}$$

Voltage gain.

$$A_V = \frac{V_o}{V_e} = \frac{I_0 R_L}{I_e R_i} = \frac{A_i R_L}{R_i} = \frac{h_{fe} R_L}{h_{ie}}$$

Output resistance:

$$R_o = \frac{V_o}{I_c}\Big|_{V_s=0}$$

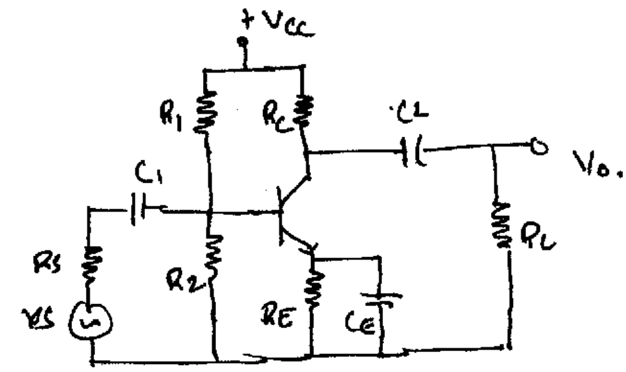
Since $I_e = 0$ $\qquad I_b = 0, \quad I_c = 0$

$R_o = \infty$

$R_o' = R_o \| R_L = R_L$

Go for example in (1)

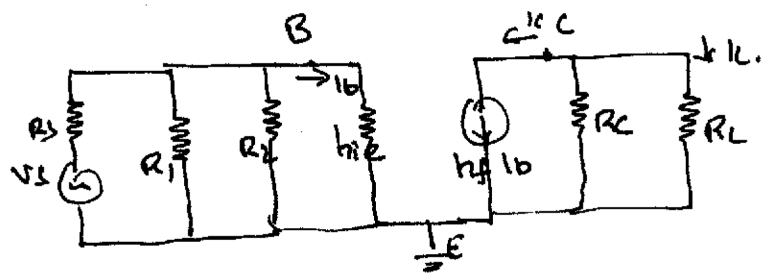
Consider a single stage CE amplifier with $R_s = 1K\Omega$, $R_1 = 50K\Omega$, $R_2 = 2K\Omega$, $R_C = 2K\Omega$, $R_L = 2K$, h_{fe} 50, h_{ie} 1100, h_{oe} $25\mu A/V$, h_{re} 2.5×10^{-4}



Find $A_i, R_i, A_v, A_i, A_{vs}, R_o$.

Since $h_{oe} R_L' = 25\times10^{-6} \times (2000 \| 2000)$
$= 0.025$ which is less than 0.1, then we use approximate analysis.

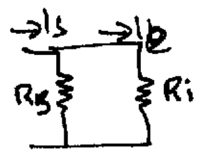
The simplified hybrid model is as shown below.



$A_i = -h_{fe} = -50$

$R_i = h_{ie} + (\times) = 1.1K$; $R_i' = R_1 \| R_2 \| R_i = 700\Omega$

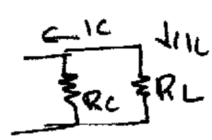
$A_v = \dfrac{A_i R_L'}{R_i} = \dfrac{-50\,(2K\|2K)}{1.1K} = \cancel{700\Omega}\; -45.45$



$\dfrac{I_b}{I_s} = \dfrac{R_B}{R_B + R_i} = 0.636$

$R_o = \dfrac{1}{Y_o} = \dfrac{1}{0} = \infty$

$R_o' = R_o \| R_L' = 1K.$



$A_{vs} = \dfrac{V_o}{V_s} = \dfrac{V_o}{V_b} \times \dfrac{V_b}{V_s}$

$\dfrac{I_L}{I_C} = -\dfrac{R_C}{R_C + R_L} = -0.5$

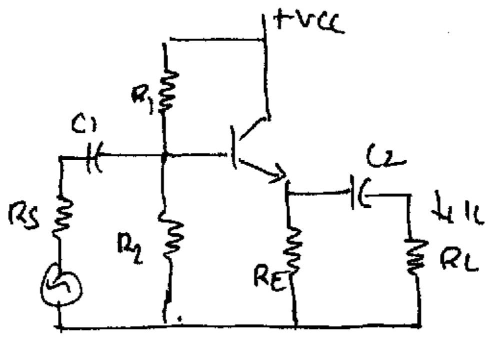
$A_{vs} = \dfrac{A_v \cdot R_i'}{R_i' + R_s} = \dfrac{-45.45 \times 700}{700 + 1K} = -18.71$

$A_{is} = \dfrac{I_L}{I_s} = \dfrac{I_L}{I_C} \times \dfrac{I_C}{I_b} \times \dfrac{I_b}{I_s}$

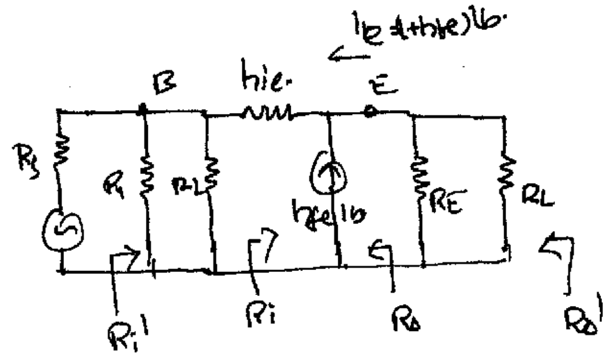
$\therefore A_{is} = -0.5 \times 50 \times 0.636$
$= -15.9$

$\dfrac{I_C}{I_b} = +50$

A common collector circuit as shown in fig! has $R_1 = 27K\Omega$, $R_2 = 27K\Omega$, $R_E = 5.6K\Omega$, $R_L = 47K\Omega$, $R_s = 600\Omega$, $h_{ie} = 1K\Omega$, $h_{fe} = 85$ $h_{oe} = 2\mu A/V$. & calculate. A_i R_i A_v R_o A_{vs} A_i



$\Leftrightarrow$



$A_i = 1 + h_{fe} = 1 + 85 = 86$

$R_i = h_{ie} + (1+h_{fe}) R_L'$.

$R_L' = R_E \| R_L$.

$R_i = 431.33\,\Omega$

$R_i' = R_1 \| R_2 \| R_i = 13.09K$.

$$A_v = \frac{A_i R_L'}{R_i} = \frac{(1+h_{fe}) R_L'}{h_{ie} + (1+h_{fe}) R_L'} = 0.997.$$

$$R_o = \frac{R_s' + h_{ie}}{1+h_{fe}}$$

$R_s' = R_s \| R_1 \| R_2$

$= 18.3\Omega$.

$R_o' = R_o \| R_E \| R_L = 18.23\Omega$.

$$A_{vs} = \frac{V_o}{V_s} = \frac{V_o}{V_b} \times \frac{V_b}{V_s}$$

$$\frac{V_b}{V_s} = \frac{R_i'}{R_i' + R_s}$$

$$\therefore \frac{A_v R_i'}{R_i' + R_s} = 0.953$$

$$A_{is} = \frac{I_o}{I_s} = \frac{I_o}{I_e} \times \frac{I_e}{I_b} \times \frac{I_b}{I_s}$$

$$\frac{I_o}{I_e} = \frac{-R_e}{R_e + R_L} = -0.106$$

$$\frac{I_b}{I_s} = \frac{R_B}{R_B + R_i} = 0.03$$

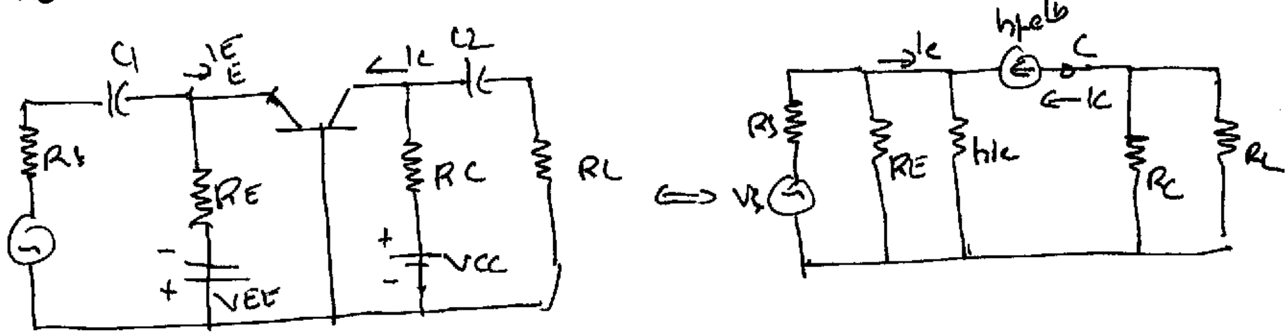
$\therefore$

$$\frac{I_o}{I_s} = (-0.106) \times (-86) \times (0.03)$$

$= 0.273$

A common base amplifier as shown has, $R_s = 600\Omega$, $R_C = 5.6K\Omega$, $R_E = 5.6K\Omega$, $R_L = 39K\Omega$. $h_{ie} = 1K$, h_{fe} 85, h_{oe} $2\mu A/V$. Calculate

R_i R_o A_v A_{vs}



$$A_I = \frac{h_{fe}}{1+h_{fe}} = 0.988$$

$$R_i = \frac{h_{ie}}{1+h_{fe}} = 11.627$$

$$R_i' = R_i \| R_E = 11.6\Omega \qquad R_L' = 4.89K\Omega$$

$$A_v = \frac{h_{fe} R_L'}{h_{ie}} = 416\Omega$$

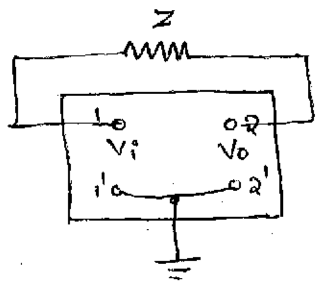
O/p Res. $R_o = \infty$

$$R_o' = R_o \| R_L' = 4.89K.$$

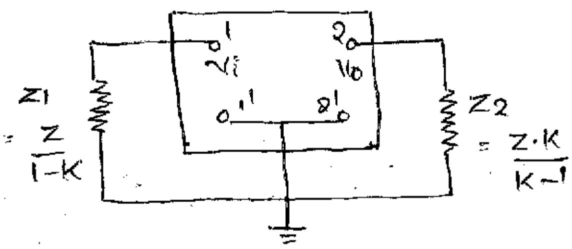
$$A_{vs} = \frac{A_v R_i'}{R_i' + R_s} = 7.89.$$

$$A_{Is}: \frac{I_o}{I_s}.$$

Millers Theorem:



(a)



(b)

Miller's theorem is used for converting any circuit having configuration of fig (a) to another configuration shown in (b).

If Z is the impedance connected between two nodes node1 and node2, it can be replaced by two seperate impedances Z_1 and Z_2; where Z_1 is connected b/w node1 and ground & Z_2 is connected b/w node 2 and ground.

The V_i and V_o are voltages at node1 and node 2 respectively. The values of Z_1 & Z_2 can be derived from ratio of V_o and V_i denoted as K.

The values of Z_1 & Z_2 are given as

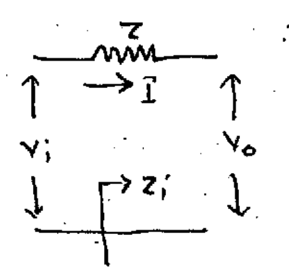
$$Z_1 = \frac{Z}{1-K} \quad \text{and} \quad Z_2 = \frac{Z \cdot K}{K-1} \qquad \left\{ \because K = \frac{V_o}{V_i} \right\}$$

Proof: Millers theorem states that, the effectiv of resistance Z on input circuit is a ratio of input voltage V_i to current I which follows from input to output.

$$Z_i = \frac{V_i}{I} \quad \text{where} \quad I = \frac{V_i - V_o}{Z}$$

$$I = \frac{V_i\left[1 - \frac{V_o}{V_i}\right]}{Z} \Rightarrow \frac{V_i}{I} = \frac{Z}{1-A_V} = \frac{Z}{1-K}.$$

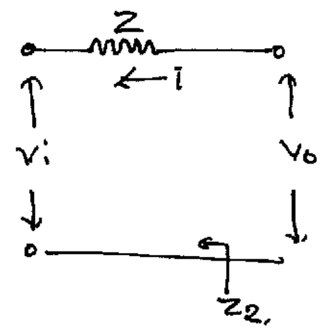
$$\because \frac{V_o}{V_i} = A_V = K$$



(a)

Millers theorem states that effect of resistance z on o/p circuit is the ratio of output voltage to the current I which flows from output to input

$z_2 = \frac{V_o}{I}$



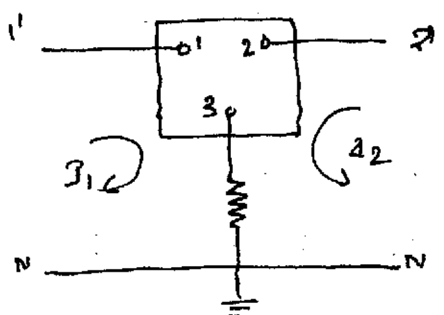
where

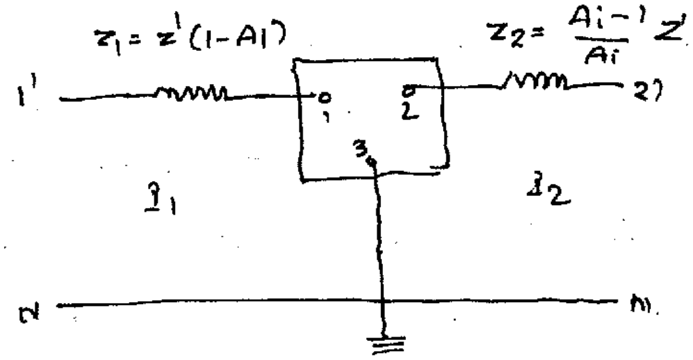
$$I = \frac{V_o - V_i}{z} = \frac{V_o\left[1 - \frac{V_i}{V_o}\right]}{z} = \frac{V_o\left[1 - \frac{1}{A_v}\right]}{z}$$

$$I = \frac{V_o\left[\frac{A_v - 1}{A_v}\right]}{z}$$

$$z_2 = \frac{V_o}{I} = \frac{Z A_v}{A_v - 1} = \frac{ZK}{K-1} \quad \because K = A_v = \frac{V_o}{V_i}$$

Dual of miller's Theorem:





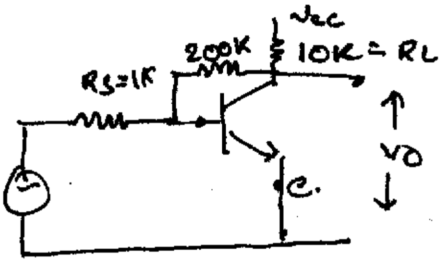
Consider network shown in A. Here z' is impedance b/w node 3 and ground N. According to dual of miller theorem z' can be split into z_1 and z_2 such that z_1 is placed in mesh 1 and z_2 is placed to mesh2 as shown in fig b.

$$A_i = \frac{-I_2}{I_1}$$

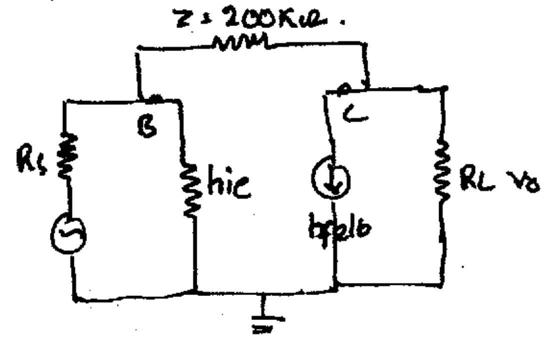
⇒ Dual of millers theorem, where node3 is grounded, an impedance z_1 is placed in mesh 1 & z_2 is added in mesh 2.. It is verified that the voltage $I_1 z_1$ equals the drop $(I_1 + I_2)z'$ across z' if $z_1 = z'(1+A_i)$

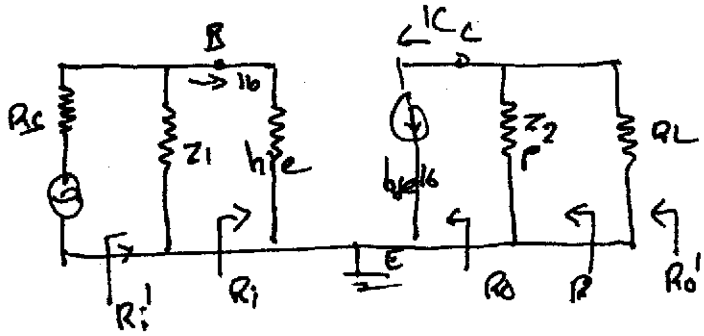
$$z_2 = \frac{z'(A_i - 1)}{A_i}$$

Ex: Calculate R_i A_v A_i A_{vs}, R_o.
h_{ie} 1.1K h_{fe} 50 $h_{oe} = h_{re} = 0$



h parameter equivalent is shown below





$$Z_2 = \frac{ZK}{K-1} \approx Z = 200K\Omega.$$

Since $h_{oe}, h_{re} = 0$ we use approximate analysis

$A_i = -h_{fe} = -50$

$R_i = h_{ie} = 1100\Omega.$

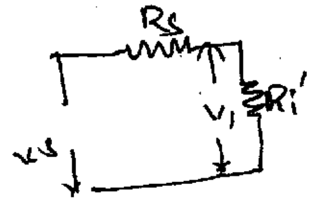
$$A_V = \frac{A_i R_L'}{R_i} \qquad R_L' = R_L \| Z_2 = 9.52K.$$

$$A_v = \frac{-50 \times 9.52 \times 10^3}{1100} = -432.72$$

$$R_i' = R_i \| Z_1 \Rightarrow Z_1 = \frac{Z(}{1-K} = \frac{200}{1-(-432.72)} = 461.12\,\Omega$$

$R_i' = 461.12 \| 1100 = 324.9\Omega.$

overall voltage gain. $\frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s}$

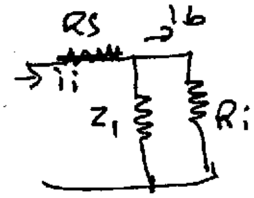


$$A_{vs} = \frac{A_v \cdot R_i'}{R_i' + R_s}$$

$$= A_v \times 0.245 = -432.72 \times 0.245 = -106.$$

overall current gain.

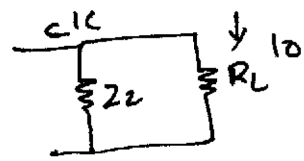
$$A_i = \frac{I_o}{I_i} = \frac{I_o}{I_c} \times \frac{I_c}{I_b} \times \frac{I_b}{I_i}$$



$$I_b = \frac{Z_1 I_i}{R_i + Z_1} = 0.295$$

∴ $A_i = -0.952 \times 50 \times 0.295 =$

$-14.$



$$\frac{I_o}{I_c} = \frac{-Z_2}{Z_2 + R_L} = -0.952$$

Ig

$R_o = \infty$

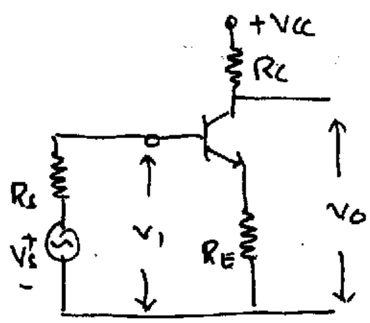
$R_o' = \infty \| R_L' = 9.52 k\Omega$ //

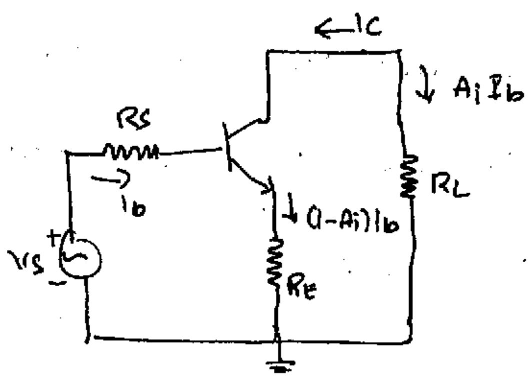
CE Amplifier with an emitter resistance:

Whenever the gain provided by single stage amplifier is not sufficient, it is necessary to cascade the number of stages of amplifier. In such situations, it becomes important to stabilize the voltage amplification of each stage, because instability of first stage is amplified in the next (2nd) stage and it is further amplified in next. This is not desired.

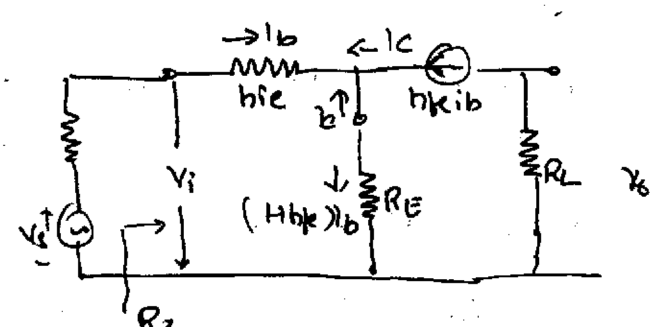
The simple way & effective way to obtain voltage gain stabilization is to add an emitter resistance R_E to CE stage as shown.

The presence of emitter resistance has number of better effects on amplifier performance. These can be analyzed with help of h-parameter equivalent circuit.




AC equivalent.

Approximate Analysis:



current gain =

$$A_i = \frac{-I_c}{I_b} = \frac{-h_{fe} i_b}{I_b} = -h_{fe}.$$

$$R_i = \frac{V_i}{I_b} = h_{ie} + (1+h_{fe}) R_E$$

$$A_V = \frac{A_i R_L}{R_i} = \frac{-h_{fe} R_L}{h_{ie} + (1+h_{fe}) R_E}$$

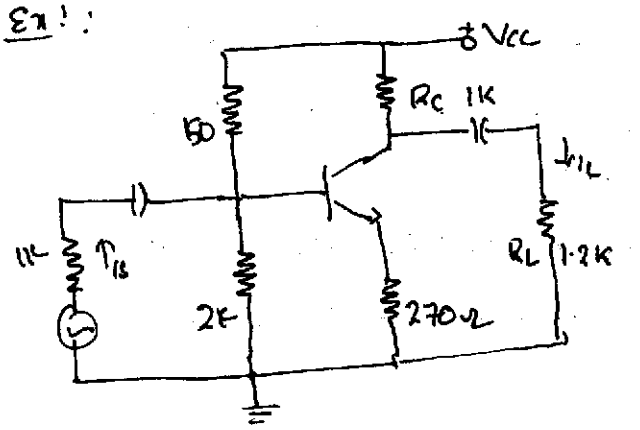
$R_o = \left.\frac{V_o}{I_o}\right|_{V_s=0}$ = it is the resistance of amplifier without considering source & load. ($V_s = 0$ & $R_L = \infty$)

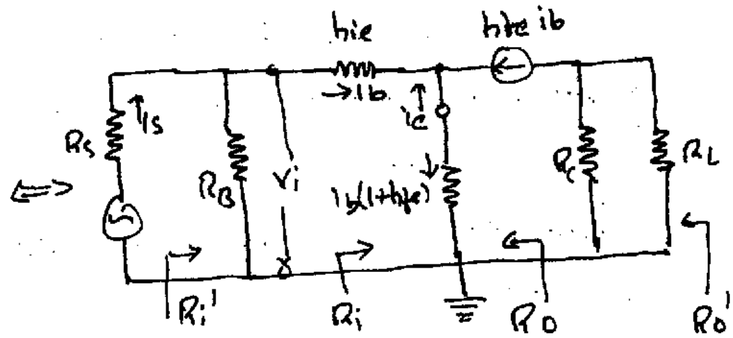
$R_o = \frac{V_o}{I_o}\Big|_{V_s=0}$ when $V_s=0$, current through loop $I_b=0$, hence I_C & I_o both are zero.

$\therefore \quad R_o = \infty$

$R_o' = R_o \| R_L = \infty \| R_L = R_L$.

Ex! :





Since $h_{oe} R_L < 0.1$ Go for approximate analysis.

Current gain $A_i = -h_{fe} = -50$

Input resistance $R_i = \frac{V_b}{I_b} = h_{ie} + (1+h_{fe})R_E = 14.87\ K\Omega$

545.45

Voltage gain $A_v = \frac{A_i R_L}{R_i} = -1.834$

Overall input resistance $R_i' = R_B \| R_i$ $\quad \because R_B = R_1 \| R_2$

$= 1.7K$.

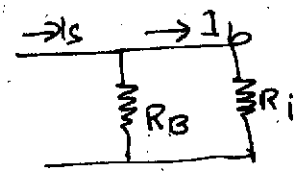
O/L

Overall output resistance $R_o' = R_o \| R_C \| R_L = 545.45\Omega$.

Overall voltage gain $A_{vs} = \frac{A_v R_i'}{R_i' + R_s} = -1.15$

Overall current gain

$A_{is} = \frac{I_L}{I_s} = \frac{I_L}{I_C} \times \frac{I_C}{I_b} \times \frac{I_b}{I_s}$

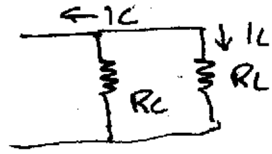


$I_b = \frac{I_S R_B}{R_B + R_i}$

$\therefore A_{I_s} = \frac{I_L}{I_S} = -2.6$.

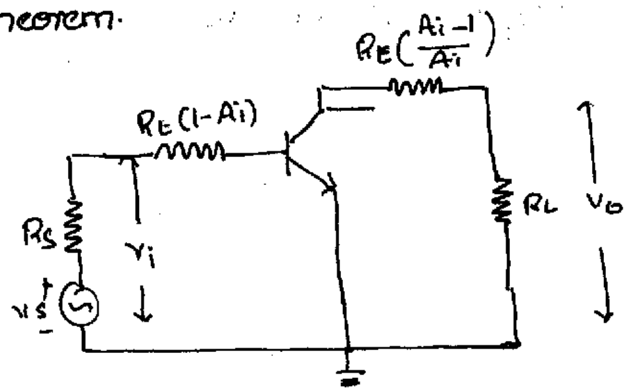
$I_L = \frac{-I_C R_C}{R_L + R_C}$

$\frac{I_L}{I_C} = \frac{-R_C}{R_L + R_C}$

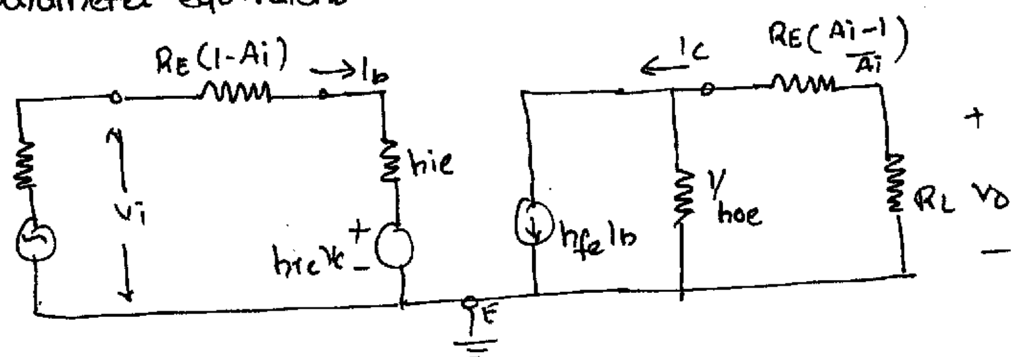


Exact Analysis:

To make the analysis of CE amplifier with R_E simple, we have to use dual of miller's theorem.



the h-parameter equivalent circuit



current gain $A_i = \dfrac{-h_{fe}}{1+h_{oe}R_L'} = \dfrac{-h_{fe}}{1+h_{oe}\left[R_L + R_E\left(\frac{A_i-1}{A_i}\right)\right]}$

$\Rightarrow \quad A_i + A_i h_{oe}R_L + A_i h_{oe}R_E - h_{oe}R_E = -h_{fe}$

$A_i\left[1+h_{oe}(R_L+R_E)\right] = h_{oe}R_E - h_{fe} \Rightarrow$

$$A_i = \frac{h_{oe}R_E - h_{fe}}{1+h_{oe}(R_L+R_E)}$$

Input Resistance R_i

$$R_i = \frac{V_i}{I_b} = h_{ie} + h_{re}A_iR_L'$$

From circuit we can see that $R_E(1-A_i)$ is in series with hie. & $R_L' = R_L + \left[\frac{A_i-1}{A_i}\right]R_E$

$$\therefore \quad R_i = \frac{V_i}{I_b} = (1-A_i)R_E + h_{ie} + h_{re}A_iR_L' \; //$$

Voltage gain $A_V = \frac{A_i R_L}{R_i}$

output resistance $R_O = \frac{V_o}{I_o} = \frac{1}{h_o}$?

$R_C = 5.7K$

$R_S = 470$

$R_E = 2.2K\Omega$

$R_L = 1.2K\Omega$

$R_1 = 100\Omega$

R_L 200KΩ

$$Y_o = h_{oe} - \frac{h_{fe} h_{re}}{R_s + h_{ie} + (1 - A_i) R_E}$$

$$R_O = \frac{1}{Y_o} =$$

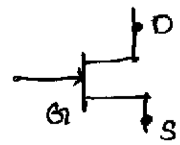
JFET Small signal model

JFET amplifiers provide excellent voltage gain with added advantage of high input impedance.

many concepts that relate to amplifiers using BJT apply equally to FET amplifiers. There are three Basic configurations.

- Common Source
- common drain and
- Common gate.

The only difference is that BJT controls large o/p current by means of small base current. whereas FET controls o/p current by means of small input (gate) voltage.



- We know that R_{DS} of FET is controlled by V_{gs}. The change in drain current due to change in V_{gs} can be determined using transconductance factor g_m.

$$\Delta I_d = g_m \Delta V_{gs}$$

In BJT, relation b/w i/p & o/p is given by amplification factor β.
In JFET " " " " " " transconductance factor g_m

Another important parameter of JFET is drain resistance

$$r_d = \left.\frac{\Delta V_{DS}}{\Delta I_D}\right|_{V_{GS} = \text{Constant}}$$

JFET low frequency ac equivalent circuit

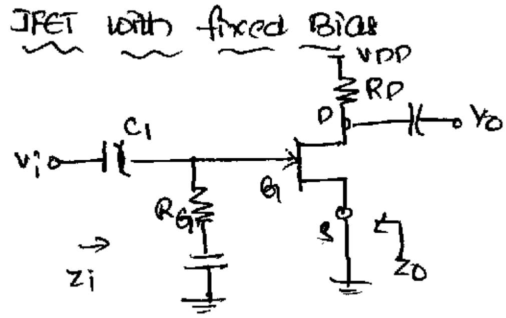
Fig: n channel JFET.

- The relation b/w of I_d by V_{gs} is included as current source $g_m V_{gs}$ connected from drain to source.
- Input impedance is represented by open circuit since $I_g = 0$
- o/p impedance is represented by r_d from drain to source.

Common source Amplifier:

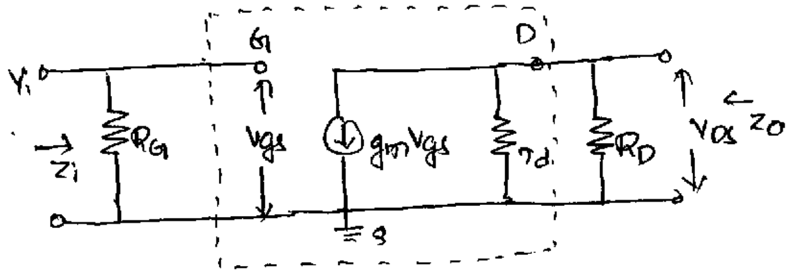
in CS amplifier input is applied b/w gate & source & o/p is taken from drain & source

JFET with fixed Bias



low frequency equivalent model for CS amplifier circuit is obtained by replacing

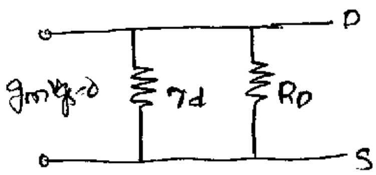
- All capacitors & dc supply voltages with short circuit
- JFET with its low frequency equivalent circuit.



Input impedance Zi: From the circuit $\boxed{Z_i = R_G}$

Output impedance Zo: The o/p impedance measured looking from o/p side with input voltage (V_i) equal to 0;

As $V_i = 0$, $V_{gs} = 0$ & hence $g_m V_{gs} = 0$



$\boxed{Z_o = R_D \parallel r_d}$

If $r_d >> R_D$ then $\boxed{Z_o \simeq R_D}$

Voltage gain: Av:

$$A_V = \frac{V_{ds}}{V_{gs}} = \frac{V_o}{V_i}$$

From Fig. we can write $V_o = -(g_m V_{gs})(r_d \parallel R_D)$

As we know $V_i = V_{gs}$ then $V_o = -g_m V_i (r_d \parallel R_D)$

$$A_V = -g_m (r_d \parallel R_D)$$

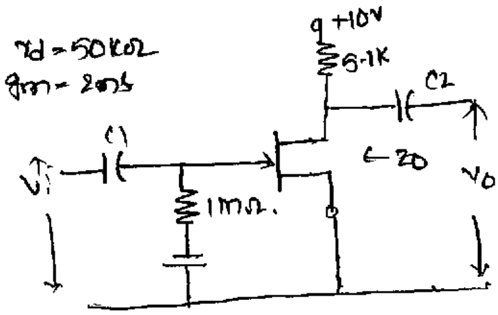
If $r_d >> R_D$ $\boxed{A_V = -g_m \cdot R_D}$

-ve sign for Av indicates there is phase shift of 180° b/w i/p & o/p voltage

1) calculate Z_i, Z_o, A_V for given circuit

$r_d = 50\,k\Omega$

$g_m = 2\,mS$



$Z_i = R_G = 1\,M\Omega$.

$Z_o = r_d \| R_D = 4628\,\Omega$.

$A_V = -g_m (r_d \| R_D)$

$= -9.256$.

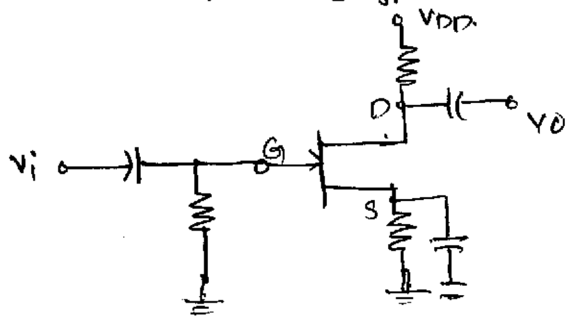
$g_{mo} = \frac{2 I_{DSS}}{|V_P|}$

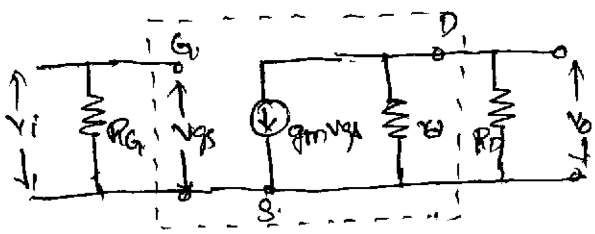
$g_m = g_{mo}\left(1 - \frac{V_{GSQ}}{V_P}\right)$

$\therefore r_d = \frac{1}{Y_{os}}$

2) calculate g_m, r_d, Z_i, Z_o & A_V. The value of $Y_{os} = 20\,\mu S$, $V_{GSQ} = -2V$, $I_{DSS} = 8\,mA$, $V_P = -8V$. for above circuit.

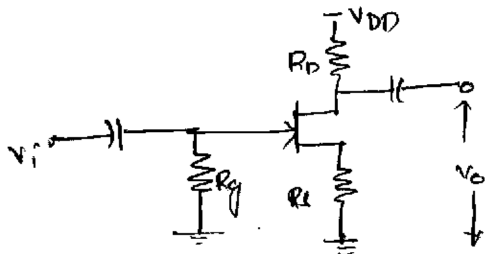
JFET with self Bias (Bypassed R_S)



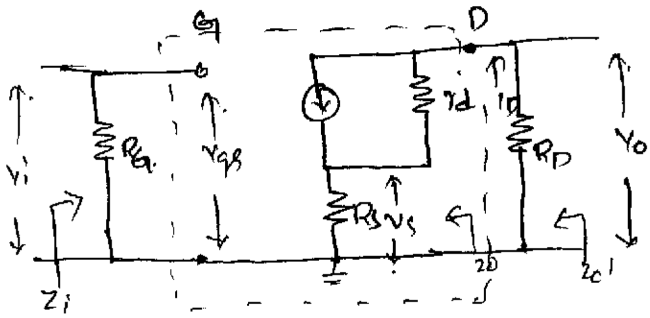


$\boxed{Z_i = R_G}$ $\boxed{Z_o = r_d \| R_D}$ $\boxed{Z_o \simeq R_D}$ $\boxed{A_V = -g_m (r_d \| R_D)}$ $\boxed{A_V = -g_m R_D}$

JFET with self Bias (unbypassed R_S)



⇒



$\boxed{Z_i = R_G}$ $Z_o' = Z_o \| R_D$ $Z_o = \left.\frac{V_o}{I_d}\right|_{V_i = 0}$

Applying KVL on o/p ckt $V_o = (I_d - g_m V_{gs}) r_d + I_d R_s$.

$V_{gs} = V_{in} - I_d R_s$. Since $V_{in} = 0$ $V_{gs} = -I_d R_s$.

$$V_o = [I_d - g_m(-I_d R_s)] r_d + I_d R_s = I_d (r_d + g_m R_s r_d + R_s).$$

$$Z_o = \frac{V_o}{I_d} = r_d + g_m R_s r_d + R_s$$ $\boxed{\mu = g_m r_d}$

$\boxed{Z_o = r_d + R_s(\mu + 1)}$ $\boxed{Z_o' = [r_d + R_s(\mu+1)] \| R_D}$.

Voltage gain A_V:

$$A_V = \frac{V_o}{V_i} \qquad V_o = -I_d R_D \quad -①$$

Applying KVL to o/p ckt

$$(I_d - g_m V_{gs}) r_d + I_d R_s + I_d R_D = 0 \quad -②$$

$$V_{gs} = V_{in} - I_d R_s \qquad I_d(r_d + R_s + R_D + g_m R_s r_d) = g_m V_i r_d$$

$$I_d = \frac{g_m V_i r_d}{r_d + R_s + R_D + g_m R_s r_d} \quad -③$$

sub ③ in ①

$$V_o = \frac{-g_m V_i r_d R_D}{r_d + R_s + R_D + g_m R_s r_d}$$

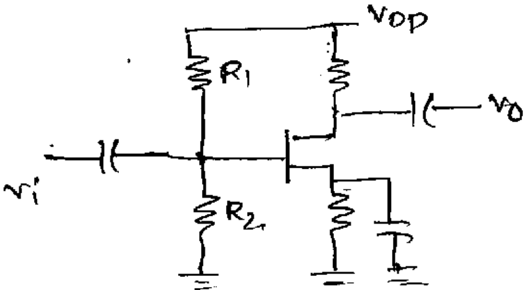
$$A_V = \frac{V_o}{V_i} = \frac{-g_m R_D r_d}{r_d + R_s + R_D + g_m R_s r_d}$$

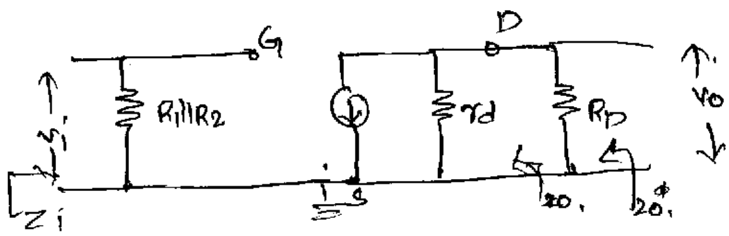
$$\boxed{A_V = \frac{-g_m R_D}{1 + g_m R_s + \frac{R_s + R_D}{r_d}}}$$

If $r_d \gg R_s + R_D$

$$A_V = \frac{V_o}{V_i} = \frac{-g_m R_D}{1 + g_m R_s} \quad //$$

JFET with voltage divider Bias: (Bypassed R_s)

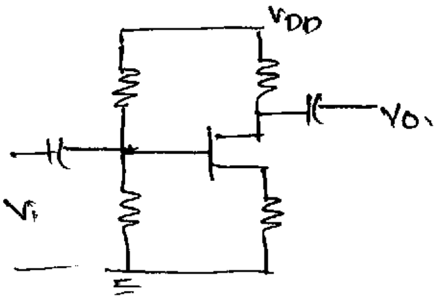


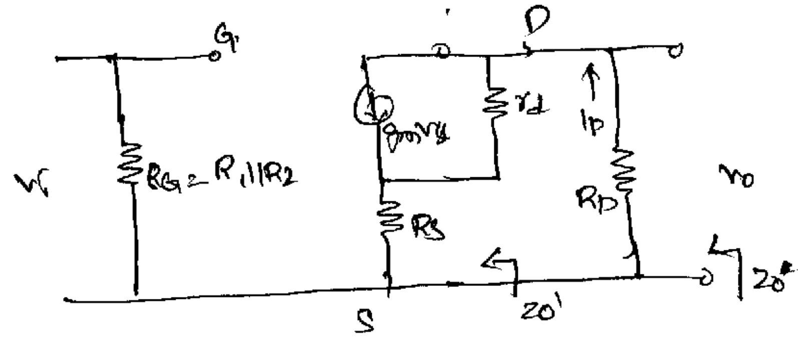


$\boxed{Z_i = R_G = R_1 \| R_2}$ $\boxed{Z_o = r_d \| R_D}$ $\boxed{Z_o \simeq R_D}$ $A_V = -g_m (r_d \| R_D)$

$\boxed{A_V = -g_m R_D}$

JFET Voltage divider Bias with Unbypassed R_s.





$\boxed{Z_i = R_G = R_1 || R_2}$

$Z_o' = r_d + g_m R_s r_d + R_s$

$Z_o' = r_d + R_s(\mu+1)$

$Z_o = Z_o' || R_D$

$\boxed{Z_o = [r_d + R_s(\mu+1)] || R_D}$

$$\boxed{A_V = \frac{-g_m R_D}{1 + g_m R_s + \frac{R_s + R_D}{r_d}}}$$ //

Ex: Find A_V, Z_i, Z_o, Z_o' $g_m = 2mA/V$, $r_d = 10K$, $R_D = 50K$, $R_L = 50K$, $R_s = 1K$, $R_G = 1M\Omega$

$\boxed{VD, URs}$ $A_V = -8.928$, $Z_i = 1M\Omega$ $Z_o = 19.135K$ $Z_o' = 18.838K$.

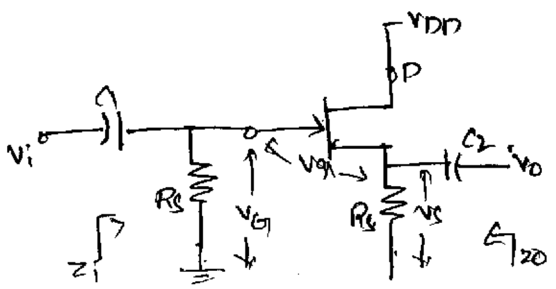
Ex Calculate g_m, r_d, Z_i, Z_o, A_V for CS Amp. operating point defined by $V_{GSQ} = -2.5V$, $V_P = -6V$ & $I_{DQ} = 2.5mA$ with $I_{DSS} = 8mA$. $R_S = 1K$ $R_G = 1M\Omega$ $R_D = 2.2K\Omega$, $Y_{os} = 20\mu S$

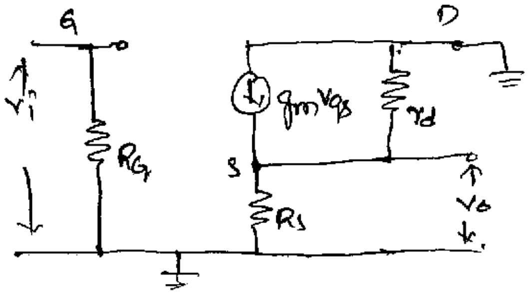
$\boxed{CS URs}$

$g_{mo} = 2.67mS$ $\boxed{g_m = 1.58mS}$, $r_d = \frac{1}{Y_{os}}$, $Z_i = 1M\Omega$ $Z_o = 2163.4\Omega$, $A_V = -1.315$

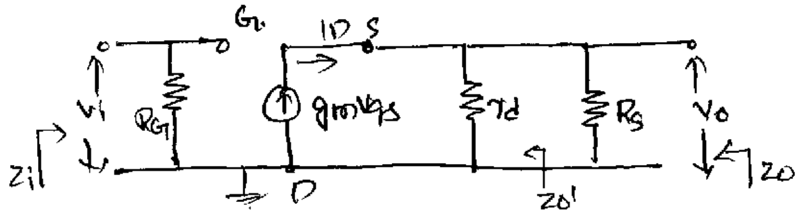
	JFET	MOSFET
r_d	0.1 – 1M	1 – 50K
g_m	0.1 – 10 mA/V	0.1 – 20 mA/V
C_{ds}	0.1 – 1 PF	0.1 – 1 PF
C_{gs}, C_{gd}	1 – 10 pF	1 – 10 pF
r_{gs}	$>10^8\Omega$	$>10^{10}$
r_{gd}	$>10^8\Omega$	$>10^{14}$

Common Drain (CD) Amplifier:





$\boxed{Z_i = R_G}$



$Z_o = Z_o' \| R_s$; $\quad Z_o' = \left.\frac{V_o}{I_d}\right|_{V_i=0}$

Applying KVL to outer loop $\quad V_i + V_{gs} - V_o = 0$

As $V_i = 0 \quad V_{gs} = V_o$

$g_m V_{gs} = I_d$

$g_m V_o = I_d$

$Z_o' = \frac{V_o}{I_d} = \frac{1}{g_m} \Rightarrow Z_o = \frac{1}{g_m} \| R_s$

$A_V = \frac{V_o}{V_i}$

$V_o = -I_d (r_d \| R_s)$

$I_d = g_m V_{gs}$

$V_o = -g_m V_{gs} (r_d \| R_s)$

$V_i = -V_{gs} + V_o \Rightarrow V_i = -V_{gs} + (-g_m V_{gs} (r_d \| R_s))$

$A_V = \frac{V_o}{V_i} = \frac{g_m (r_d \| R_s)}{1 + g_m (r_d \| R_s)}$ $\quad$ If $r_d \gg R_s$ $\quad A_V = \frac{g_m R_s}{1 + g_m R_s}$

A_V is always less than 1.

$R_s = 3.3K \quad R_G = 1M \quad r_d = 25K \quad g_m = 2.5\,ms$ Calculate Z_i, Z_o & A_V.

UNIT-2 FORMULAE

EXACT ANALYSIS

1) Current Gain: $A_i = \dfrac{-h_{fe}}{1 + h_{oe} R_L}$

2) Input impedance $= R_i = h_i + h_r A_i R_L = h_i - \dfrac{h_r h_f R_L}{1 + h_o R_L}$

$= h_i - \dfrac{h_r h_f}{1/R_L + h_o} = h_i - \dfrac{h_r h_f}{Y_L + h_o}$

3) Overall current gain:

$A_{is} = \dfrac{A_i R_s}{R_i + R_s}$

4) Voltage gain: $A_v = \dfrac{A_i R_L}{Z_i}$

5) Overall voltage gain $A_{vs} = \dfrac{V_2}{V_s} = \dfrac{V_2}{V_1} \times \dfrac{V_1}{V_s}$

$A_v \times \dfrac{V_1}{V_s} = A_v \cdot \dfrac{R_i}{R_i + R_s} = \dfrac{A_i R_L}{R_i + R_s}$ //

6) Output Admittance:

$Y_o = h_o - \dfrac{h_f h_r}{h_i + h_s}$

7) power gain. $A_p = A_v \cdot A_i = \dfrac{A_i^2 R_L}{Z_i}$

8) $A_{vs} = A_{is} \cdot \dfrac{R_L}{R_s}$

	CE	CB	CB
h_{ie}	1100Ω	$h_{ic} = h_{ie}$	$h_{ib} = \dfrac{h_{ie}}{1+h_{fe}}$
h_{re}	2.5×10^{-4}	$h_{rc} = 1$	$h_{rb} = \dfrac{h_{ie} h_{oe}}{1+h_{fe}} - h_{re}$
h_{fe}	50Ω	$h_{fc} = -(h_{fe}+1)$	$h_{fb} = \dfrac{-h_{fe}}{1+h_{fe}}$
h_{oe}	$25\mu A/V$	$h_{oc} = h_{oe}$	$h_{ob} = \dfrac{h_{oe}}{1+h_{fe}}$

Simplified model formulae.

CE:

$A_i = -h_{fe}$

$R_i = h_{ie}$

$A_v = \frac{A_i R_L}{h_{ie}}$

$Y_o = 0 \quad R_o = \infty$

$R_o' = R_o \| R_L = R_L$

CC

$A_i = \frac{-I_e}{I_b} = 1 + h_{fe}$

$R_i = (1+h_{fe}) R_L + h_{ie}$

$A_v = \frac{A_i R_L}{R_i} = \frac{(1+h_{fe}) R_L}{h_{ie} + (1+h_{fe}) R_L}$

$R_o = \frac{V_o}{I_e} = \frac{R_s + h_i}{1+h_{fe}}$

$R_o' = R_o \| R_L$

CB

$A_i = \frac{-I_c}{I_e} = \frac{h_{fe}}{1+h_{fe}}$

$R_i = \frac{h_{ie}}{1+h_{fe}}$

$A_v = \frac{h_{fe} R_L}{h_{ie}}$

$R_o = \infty$

$R_o' = R_o \| R_L = R_L$

h_{fe} h_{ie} h_{re}

CE

$h_{oe} \rightarrow$

h_{fe}

Miller's theorem

$Z_1 = Z(1-K) \qquad Z_2 = \frac{ZK}{K-1}$

Dual $Z_1 = R_E(1-A_i) \qquad Z_2 = \frac{(A_i - 1) R_E}{A_i}$

CE with R_E!

UNIT - IV

multi stage amplifiers.

introduction:

we need an amplifier which can amplify a signal from a weak source such as a microphone to a level which is suitable for the operation of another transducer such as loud speaker. If sufficient amount of amplification is not achieved with the single stage amplifier, we go for multi stage amplifier.

When the amplification of single stage amplifier is not sufficient. (or) when the input and output impedance is not of correct magnitude, for a particular application two or more amplification stages are connected called as multi stage amplifier.

There are two types of multi stage amplifiers

→ cascade

→ cascode.

In cascade, the stages are connected such that the output of first stage is connected to input of second stage.

cascode amplifiers consist of a common emitter amp. stage in series with common base amp. stage

Analysis of cascaded amplifier:

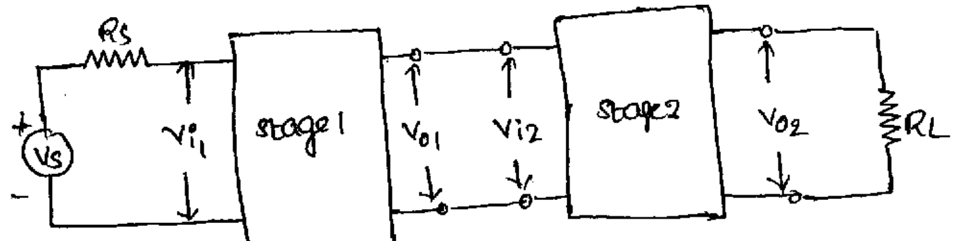


Fig: Block diagram of two stage cascade amplifier.

As shown in fig: V_{i1} input of first stage

V_{o1} is output of first stage

V_{i2} is input of second stage

V_{o2} is output of second stage.

The overall voltage gain of two stage amplifier is given by

$$A_V = \frac{V_{o2}}{V_{i1}} = \frac{V_{o2}}{V_{i2}} \times \frac{V_{i2}}{V_{i1}}$$

we know that $V_{o1} = V_{i2}$.

$$\therefore \quad A_V = \frac{V_{o2}}{V_{i2}} \times \frac{V_{o1}}{V_{i1}}$$

$$\boxed{A_V = A_{V2} \times A_{V1}} \quad \text{--- (1)}$$

So that we can say that voltage gain of multistage amplifier is the product of voltage gains of individual stages.

For N stages of cascaded amplifiers, the voltage gain is given by

$$A_V = A_{V1} \times A_{V2} \times A_{V3} \cdots \times A_{Vn} \quad \text{--- (2)}$$

To analyse multistage amplifier circuit, we make use of general expressions of A_I, Z_i, A_V & Y_o which were derived in UNIT-1.

NOTE: In cascaded amplifiers, collector resistance of one stage is shunted by input impedance of next stage. Hence it is advantageous to start analysis with last stage.

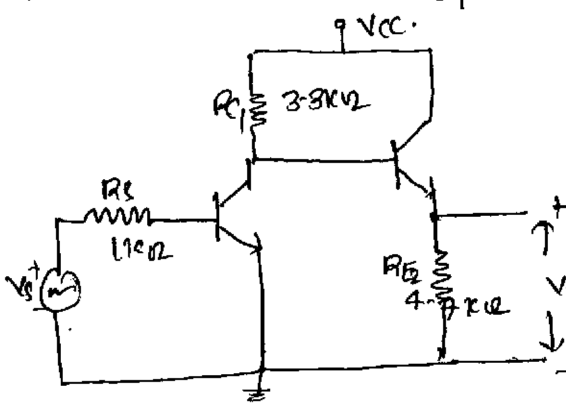
Pb 1. Consider a two stage amplifier circuit in CE CE configuration as shown in fig; the transistor parameters at the corresponding quiescent point are

$h_{ie} = 2K$

$h_{fe} = 50$

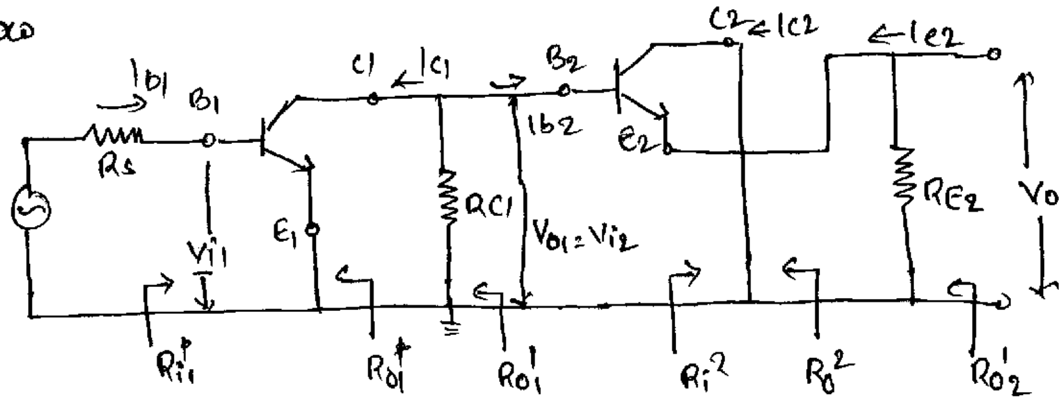
$h_{re} = 0$

$h_{oe} = 0$.

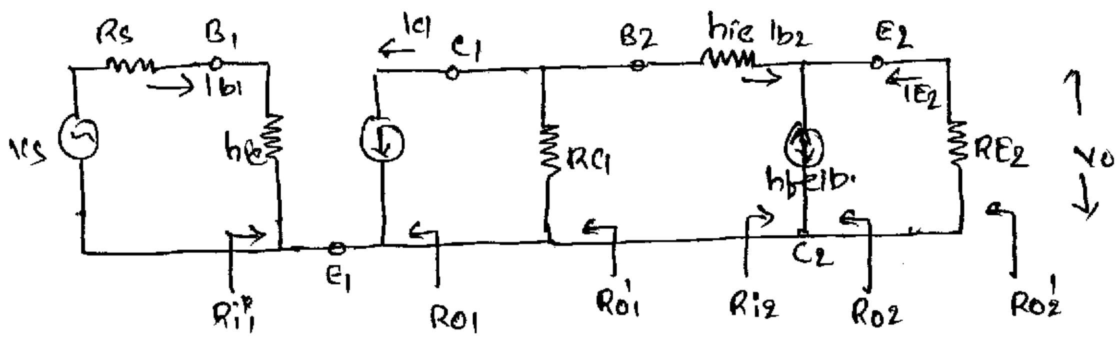


Calculate input impedance, output impedance, overall voltage and current gains

The AC equivalent circuit for two stage amplifier is shown below



h parameter equivalent model is given as below.



Analysis of second stage (CC amplifier)

current gain $A_{i2} = 1+h_{fe} = 1+50 = 51$

Input resistance $R_{i2} = h_{ie} + (1+h_{fe})R_{E2} = 2K + 51 \times 4.7K = 241.7K\Omega$

Voltage gain $A_{V2} = \frac{A_I R_{L2}}{R_{i2}} = \frac{A_{i2} R_{E2}}{R_{i2}} = \frac{51 \times 4.7K}{241.7K} = 0.991.$

Analysis of first stage (CE amplifier)

current gain $A_{i1} = -h_{fe} = -50$

Input resistance $R_{i1} = h_{ie} = 2K$

Voltage gain $A_{V1} = \frac{A_{i1} R_{L1}}{R_{i1}} = \frac{A_{i1} \times R_{C1} \| R_{i2}}{R_{i1}}$

* The net load resistance R_{L1} of first stage is parallel combination of R_{C1} and R_{i2}.

$\therefore \quad R_{L1} = R_{C1} \| R_{i2} = 3.3K \| 241.7K = 3.25K.$

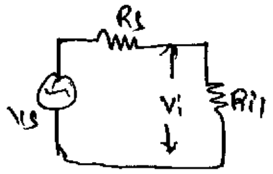
$A_{V1} = -81.25$

Total voltage gain: $A_V = A_{V1} \times A_{V2}$

$$= -81.25 \times 0.991 = -80.51$$

overall voltage gain A_{VS}

$$A_{VS} = \frac{V_o}{V_s} = \underbrace{\frac{V_o}{V_i}}_{A_V} \times \frac{V_i}{V_s}$$



$$V_i = \frac{V_s R_{i1}}{R_{i1}+R_s}$$

$$\frac{V_i}{V_s} = \frac{R_{i1}}{R_{i1}+R_s}$$

$$\therefore A_{VS} = \frac{A_V \times R_{i1}}{R_{i1}+R_s} = -80.51 \times \frac{2K}{1K+2K} =$$

$2K = 40.25$

$$= -53.67$$

output impedance R_o

$R_{o1} = \infty$

$R_{o1}' = R_{o1} \| R_{C1} = \infty \| 3.3K = 3.3K$

$$R_{o2} = \frac{R_s + h_{ie}}{1+h_{fe}}$$

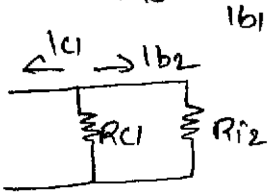
$$= \frac{R_{o1}' + h_{ie}}{1+h_{fe}} = \frac{3.3K + 2K}{1+50}$$

$R_{o2} = 103.9\Omega$

$R_o = R_{o2} \| R_{E2} = 103.9 \| 4.7K = 101.65\Omega$

overall current gain A_{is}

$$A_{is} = \frac{I_o}{I_{b1}} = \frac{I_o}{I_{b2}} = \frac{-I_{e2}}{I_{b2}} \times \frac{I_{b2}}{I_{c1}} \times \frac{I_{c1}}{I_{b1}} = (A_{i2}) \frac{I_{b2}}{I_{c1}} (-A_{i1})$$



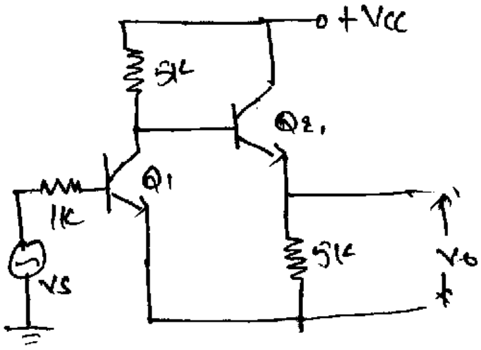
$$\frac{I_{b2}}{I_{c1}} = \frac{-R_{C1}}{R_{C1}+R_{i2}} = -0.01346$$

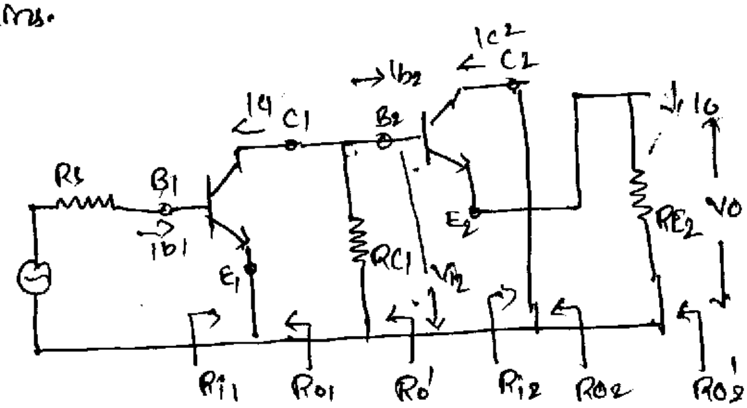
$$\therefore A_i = -51 \times -0.01346 \times -(-50) = -34.323$$

Problem 2.2 Consider a two stage amplifier circuit in CE & CC config. as shown in fig. The transistor parameters at quiscent point are.

$h_{ie} = 2k\Omega$, $h_{fe} = 50$, $h_{re} = 6\times10^{-4}$, $h_{oe} = 25\,\mu A/V$, $h_{ic} = 2k\Omega$, $h_{fc} = -51$, $h_{rc} = 1$

$h_{oc} = 25\mu A/V$. Find. input and output impedance & individual and overall voltage and current gains.





h-parameter equivalent model is shown below.

2nd stage $\quad h_{oc} \times R_{E2} = 25\times10^{-6} \times 5k = 125\times10^{-3} = 0.125 > 0.1$

go for exact analysis.

1st stage $\quad h_{oe} \times (R_{c1} \| R_{i2}) \quad 25\times10^{-6} \times (4.89k) = \;>0.1$ exact analysis.

CC Amplifier Exact Analysis:

$$A_{i_2} = \frac{-h_{fc}}{1+h_{oc}R_{L2}} = \frac{51}{1+25\times10^{-6}\times5k} = 45.3$$

$$R_{i_2} = h_{ic} + h_{rc}A_{i_2}R_{L2} = 2k + (1\times45.3\times5k) = 228.5\,k\Omega.$$

$$A_V = \frac{V_o}{V_{i2}} = \frac{A_{i2}R_{L2}}{R_{i2}} = \frac{45.3\times5k}{228.5k} = 0.991$$

CE amplifier first stage:

The net load resistance R_{L1} of first stage is the $\|el$ combination of R_{C1} & R_{i2}

$$\therefore R_{L1} = R_{C1} \| R_{i2} = 4.9k\Omega.$$

$$\text{current gain } A_{i_1} = \frac{-I_{c1}}{I_{b1}} = \frac{-h_{fe}}{1+h_{oe}R_{L1}} = \frac{-50}{1+25\times10^{-6}(4.9k)} = -44.5$$

current gain A_{I1} =

input impedance $R_{i1} = h_{ie} + h_{re} A_{I1} R_{L1}$

$= 2K + (6 \times 10^{-4} \times 44.5 \times 4.9k)$

$= 1.87 \, K\Omega$.

The voltage gain for 1st stage is

$$A_{V1} = \frac{A_{I1} R_{L1}}{R_{i1}} = \frac{-44.5 \times 4.9k}{1.87k} = -116.5$$

The output impedance.

$$Y_{01} = h_{oe} - \frac{h_{fe} h_{re}}{h_{ie} + R_s} = 2.5 \times 10^{-6} - \frac{50 \times 6 \times 10^{-4}}{2k + 1k}$$

$$= 15 \times 10^{-6} \, \mho$$

$R_{01} = 66.7 \, k\Omega$.

effective o/p impedance of first stage, taking R_{C1} into account is $R_{01} \| R_{C1}$

$$R_{01}' = R_{01} \| R_{C1} = \frac{R_{01} R_{C1}}{R_{01} + R_{C1}} = 4.65 \, K\Omega$$

output resistance of last stage

source resistance R_{s2}' for Q_2 is $R_{01} \| R_{C1} \Rightarrow R_{s2}' = R_{01}' = 4.65 \, K\Omega$

$$Y_{02} = h_{oc} - \frac{h_{fc} h_{rc}}{h_{ic} + R_{s2}'} = 7.70 \times 10^{-3} \, A/V$$

$R_{02} = 130 \, \Omega$.

$R_{02}' = R_{02} \| R_{e2} = 127 \, \Omega$

Total voltage gain:

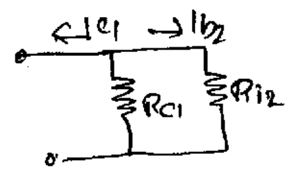
$A_V = A_{V1} \times A_{V2} = 0.991 \times (-116.5) = -115$.

overall voltage gain.

$$A_{Vs} = A_V \cdot \frac{R_{i1}}{R_{i1} + R_s} = \frac{-115 \times 1.87k}{1.87k + 1k} = -75.3$$

$$A_I = \frac{-I_{e2}}{I_{b2}} \times \frac{I_{b2}}{I_{c1}} \times \frac{I_{c1}}{I_{b1}} = A_{I2} \, \frac{I_{b2}}{I_{c1}} \times (-A_{I1})$$

A_I =



$$\frac{I_{b2}}{I_{C1}} = \frac{-R_{C1}}{R_{C1} + R_{i2}}$$

=

$$\therefore A_{IS} = 45.3 \times 44.5 \times \frac{-5k}{(228.5 + 5)k} =$$

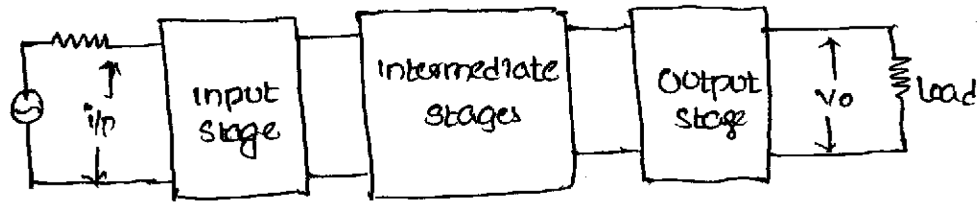
$= -43.2$

$$A_{Is} = A_I \times \frac{R_s}{R_s + R_{i1}}$$

CHOICE OF TRANSISTOR CONFIGURATION IN A CASCADE:

It is important to note that the previous calculations of input and output impedances and voltage and current gains are applicable for any connection of cascaded stages. They could be CC, CB or combination of all three possible connections.

The multi stage amplifiers in cascade can be categorized into three types (ways).



The input and output stages are selected on the basis of impedance considerations.

The intermediate stages are generally selected based on voltage & current gain considerations.

→ For getting maximum voltage gain in cascade, one cannot use CC configuration in intermediate stages because voltage gain of such stages is less than unity. Hence it is not possible to increase overall voltage amplification by cascading common collector stages.

→ Grounded base (common base) stages are often cascaded because the voltage gain of such stage is approximately same as that of output stage alone.

This is because the voltage gain is given by $A_V = \frac{A_i R_L}{R_i}$

The effective load resistance R_L is parallel combination of actual collector resistance R_C and input resistance R_i of following stage ie; $R_L = R_C \| R_i$

This parallel combination is certainly less than R_i and hence for identical stage (CB-CB), the effective load resistance R_L is less than R_i ($R_L < R_i$)

The maximum current gain is h_{fb}, which is always less than unity (but ≤ 1) $A_i \leq 1$

Hence voltage gain of any stage except last or o/p stage is less than unity (≤ 1)

Therefore, the intermediate stages can be considered with CE configuration stage because current gain h_{fe} & voltage gain of CE stage are very much greater than unity.

We can now state that, in cascade, the intermediate transistor should be connected in a common emitter configuration.

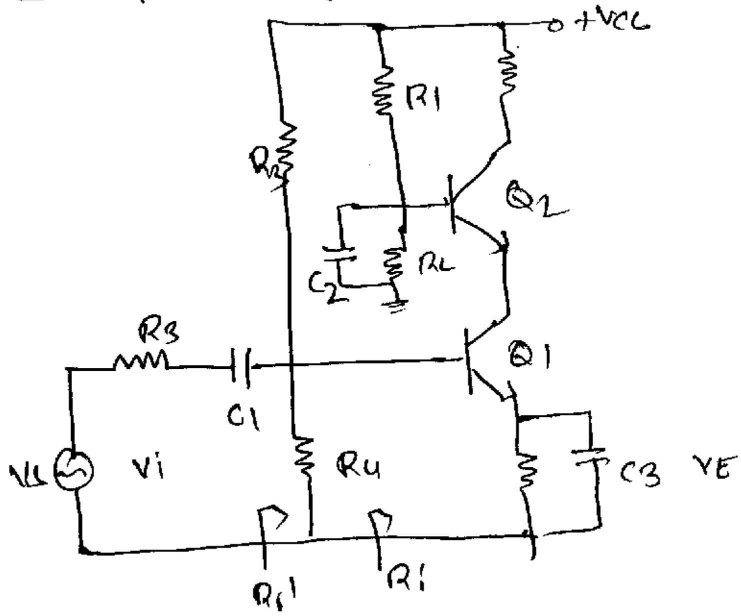
(2) The choice of input stage is decided by criteria other than voltage gain. For ex; the amplitude or freq response of transducer (source) \/ce may depend on impedance into which it operates. In many cases the common collector or common base stage is used at input because of impedance.

Noise is another important consideration which may determine selection of particular configuration of input stage

(3) The output stage is selected also on the basis of impedance considerations. Since a CC stage has very low input resistance it is often used for last stage if it is required to drive a low impedance load (capacitive loads).

CASCODE Amplifier.

The Cascode amplifier consists of CE amplifier stage with in series with CB amplifier stage. as shown in fig.

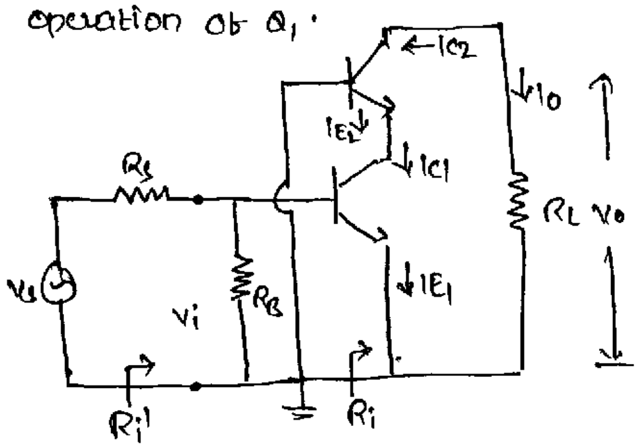


It is one approach to solve low impedance problem of CB circuit. Here Q1 & its associated components operate as CE amplifier stage, while the circuits of Q2 acts as CB amp O/p stage.

The cascode amplifier gives high i/p impedance of CE amp. as well as good voltage gain & high freq performance of CB circuit.

⑥

For DC bias, conditions of circuit, it is seen that the emitter current for Q_1 is set by V_{E1} & R_{E1}. collector current I_{C1} is approximately equal to I_{E1}, and I_{E2} is same as I_{C1}. Therefore I_{C2} is " " " I_{E1}. This current remains constant regardless of level of V_{B2} as long as V_{E1} remains large enough for current operation of Q_1.



The fig shows ac equivalent circuit for cascade amplifier. It is drawn by shorting the DC supply and capacitors.

$\rightarrow R_B = R_B \| R_u$.

The simplified h-parameter circuit for cascade amplifier is shown below.

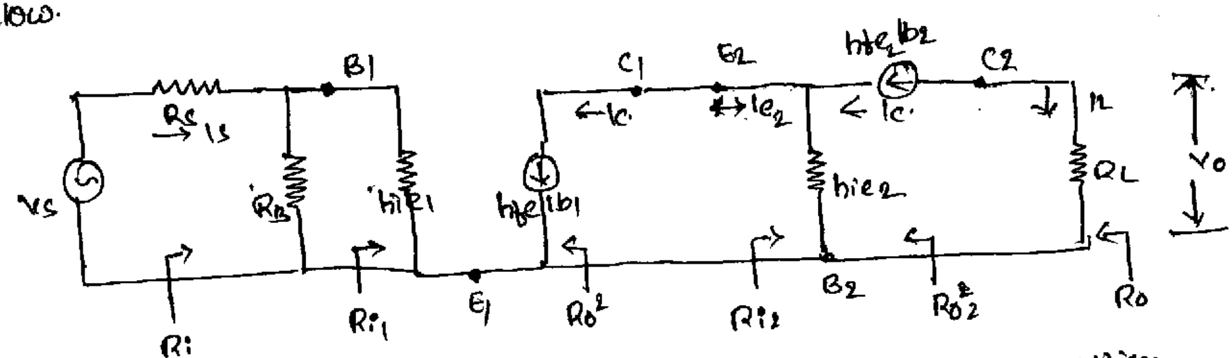


fig. Simplified hybrid model for cascode amplifier

Let us consider parameters $R_s = 1K$, $R_B = 200K$, $R_u = 10K$ $R_L = 3K$ $h_{ie} 1.1K$ h_{fe} 50.

Analysis of second stage (CB)

current gain $A_{i2} = \frac{h_{fe}}{1+h_{fe}} = 0.98$

input resistance $R_{i2} = \frac{h_{ie}}{1+h_{fe}} = 21.56\Omega$

voltage gain $A_V = \frac{A_{i2} R_{L2}}{R_{i2}} = \frac{0.98 \times 3000}{21.56} = 136.36$

Analysis of first stage (CE)

current gain $A_{i_1} = -h_{fe} = -50$

input resistance $R_{i_1} = h_{ie} = 1.1k\Omega$

Voltage gain $A_{V_1} = \frac{A_{i_1} R_{L_1}}{R_i} = \frac{-50 \times 21.56}{1.1K} = -0.98$

$\because (R_{L_1} = R_{i_2})$

Total voltage gain $A_V = A_{V_1} \times A_{V_2} = -0.98 \times 136.36 = -133.63$

Total input resistance $R_i = R_{i_1} || R_B = 986.1\Omega$

overall voltage gain A_{VS}

$$A_{VS} = \frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s} = A_V \cdot \frac{R_i}{R_i + R_s} = \frac{-133.68 \times 986.1}{986.1 + 1000}$$

$$A_{VS} = -66.35$$

Overall current gain A_{IS}

$$A_{IS} = \frac{I_o}{I_s} = \left(\frac{I_o}{I_{C_2}}\right) \times \frac{I_{C_2}}{I_{e_2}} \times \frac{I_{e_2}}{I_{C_1}} \times \frac{I_{C_1}}{I_{b_1}} \times \frac{I_{b_1}}{I_s}$$

$$\frac{I_o}{I_{C_2}} = -1 \qquad \frac{I_{C_2}}{I_{e_2}} = -A_{i_2} \qquad \frac{I_{e_2}}{I_{C_1}} = -1, \quad \frac{I_{C_1}}{I_{b_1}} = -A_i$$

$$\frac{I_{b_1}}{I_s} = \frac{R_B}{R_B + R_{i_1}}$$

$$A_{IS} = -1 \times -A_{i_2} \times -1 \times -A_{i_1} \times \frac{R_B}{R_B + R_{i_1}}$$

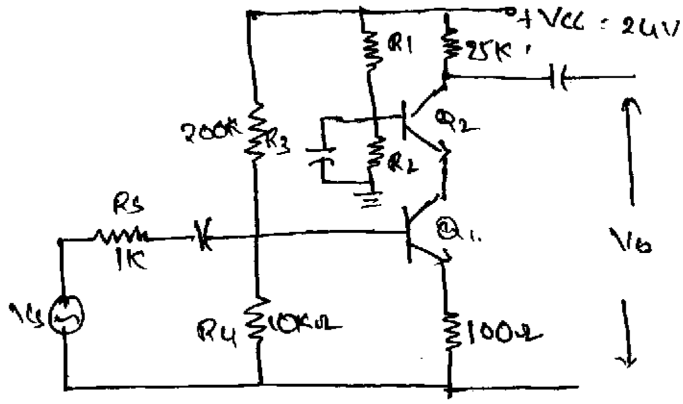
$$= -43.9$$

output Resistance R_o =

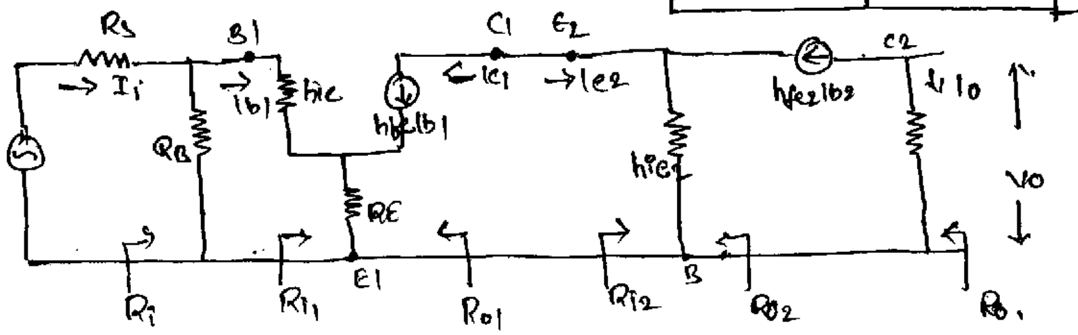
$R_{o_1} = \infty \qquad R_{o_2} = \infty$

$R'_{o_2} = R_{o_2} || R_L = \infty || 3k = 3k\Omega$

Calculate $A_I = \frac{I_0}{I_1}$, A_V, A_{VS}, R_i & R_o for the cascode circuit shown. Assume transistor parameters are identical with $h_{fe}=100$, $h_{ie}=2K$, $h_{oe}=h_{re}=0$.



The simplified h-parameter equivalent circuit for CE-CB cascode amplifier is shown in fig.



Analysis of second stage (CB)

$$A_{I2} = \frac{h_{fe}}{1+h_{fe}} = \frac{100}{1+100} = 0.99$$

$$R_{i2} = \frac{h_{ie}}{1+h_{fe}} = \frac{2K}{101} = 19.8\Omega$$

$$A_{V2} = \frac{A_{I2} R_{L2}}{R_{i2}} = \frac{0.99 \times 25K}{19.8} = 1250$$

Analysis of first stage (CE):

$$A_{I1} = -h_{fe} = -100$$

$$R_{i1} = h_{ie} + (1+h_{fe})R_E = 2K + 101 \times 100 = 12.1 K\Omega$$

$$A_{V1} = \frac{A_{I1} R_{L1}}{R_{i1}} = \frac{-100 \times 19.8}{12.1K}$$

$(\because R_{L1} = R_{i2})$ $= -0.1636$

Total gain $A_V = A_{V1} \times A_{V2}$

$= -204.5$

Total input resistance

$R_i = R_B \| R_{i1}$ $\quad R_B = 9.523K$

$= 5.328 K\Omega$.

Overall voltage gain.

$$A_{VS} = \frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s}$$

$$A_V \times \frac{R_i}{R_i + R_s} = -204.5 \times \frac{5.328K}{6.328K}$$

$A_{VS} = -172.18$

Overall current gain A_{IS}

$$A_{IS} = \frac{I_0}{I_S} = \frac{I_0}{I_{C2}} \times \frac{I_{C2}}{I_{E2}} \times \frac{I_{E2}}{I_{C1}} \times \frac{I_{C1}}{I_{b1}} \times \frac{I_{b1}}{I_S}$$

$$\frac{I_{b1}}{I_S} = \frac{R_B}{R_B + R_{i1}}$$

$$\therefore A_{IS} = -1 \times -A_{I2} \times -1 \times -A_{I1} \times \frac{R_B}{R_B + R_i}$$

$= -43.6$

Output resistance R_o

$R_{o1} = \infty$ $\quad R_{o2} = \infty$

$R_o = R_{o2} \| R_L = 25K\Omega$

Darlington pair with emitter follower:

We know that the common collector or emitter follower has high input impedance. Typically it is 200KΩ to 300KΩ. A single stage emitter follower circuit can give input impedance upto 500KΩ. However the input impedance considering biasing resistor is significantly less because $R_i' = R_1 \| R_2 \| R_i$

The input impedance can be increased using two techniques

→ Using Direct Coupling (Darlington Connection)

→ Using Bootstrap technique

Darlington transistors:

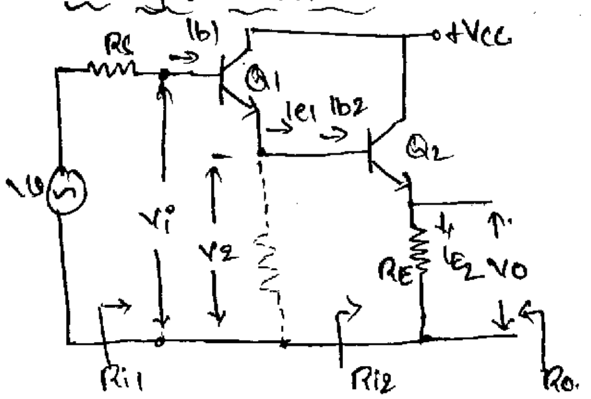
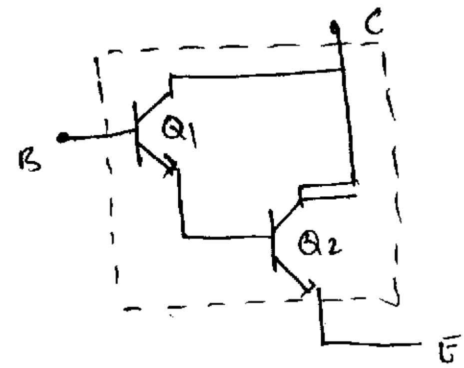


Fig: Darlington Emitter follower circuit

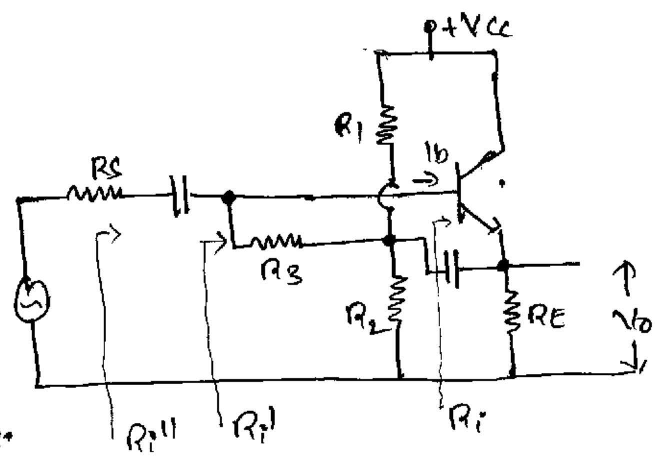


Darlington transistor.

Bootstrap emitter follower

In bootstrap EF, two additional components are used, resistance R_3 and capacitor C_2. T

The capacitor C_2 is connected b/w emitter and junction of R_1, R_2, R_3.



– For DC signal, C_2 acts as open circuit & therefore R_1, R_2, R_3 provides necessary biasing to keep transistor in active region.

– For AC signal, capacitor acts as short circuit. Its value is chosen such that it provides very low reactance. SC at lowest operating frequency. Hence for AC, the other end of R_3 is effectively connected b/w input node & O/P node.

For such connection, effective resistance is given by miller theorem.

The theorem states that impedance connected b/w two nodes can be resolved into two components, one from each node. The two components are:

$$\frac{Z}{1-K} \quad \text{and} \quad \frac{ZK}{K-1}$$

In our case R_3 is impedance b/w output voltage and input voltage

Fig: shows the bootstrap circuit with two resolved components of R_3 Using miller theorem.

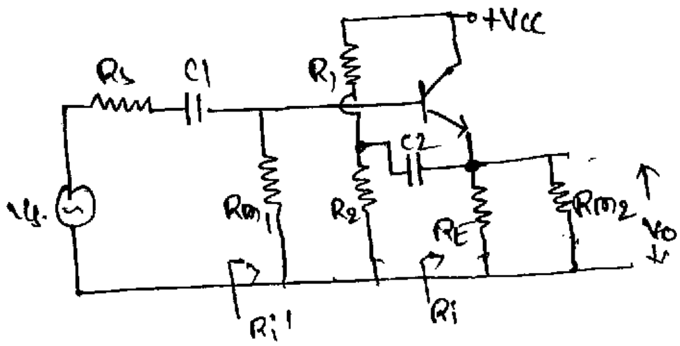


Fig: resolved boot strap Emitter follower circuit

where $R_{m_1} = \dfrac{R_3}{1-A_V}$

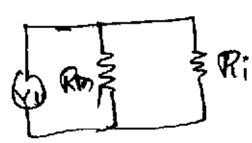
$$R_{m_2} = \frac{R_3(A_V)}{(A_V-1)}$$

Since, for EF, A_V approaches unity, then R_{m_2} becomes extremely large.

For Example: $A_V = 0.99$ & $R_3 = 200K$, then $R_{m_2} = 20M\Omega$.

The effective resistance R_i' for circuit can be given as

$R_i' = R_i \| R_{m_1}$ where $R_i = h_{ie} + (1+h_{fe})R_e$

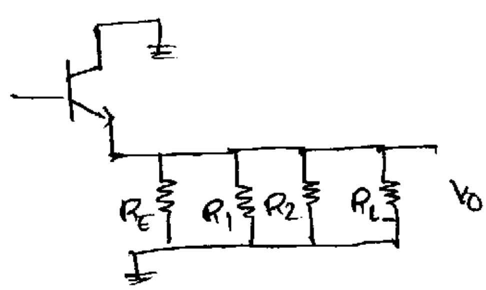


The above effect when $A_V = 1$, is called "bootstrapping." The name arises from the fact that R_3 changes in voltage, the other end of R_3 moves through same potential difference it is as if R_3 is pulling itself by its "bootstrapping."

The effective load on EF can be given as.

$R_{eff} = R_E \| R_1 \| R_2 \| R_{m_2}$

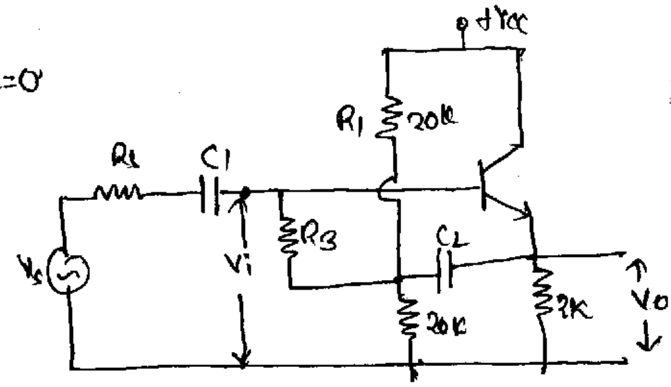
Because of capacitor C_2, biasing resistors R_1 & R_2 come on output side shunting effective load resistance



The R_{m_2} is very large, so it can be neglected

$$\therefore \quad R_{Leff} = R_E \| R_1 \| R_2.$$

For the circuit shown in fig. calculate R_{Leff}, R_i and R_i'.



$h_{fe}=100$, $h_{re}=0$, $h_{ie}=2K$, $h_{oe}=0$

Sol:

Here $R_{Leff} = R_1 || R_2 || R_E$

$= 1.67 K\Omega$

$R_i = h_{ie} + (1+h_{fe}) R_{Leff}$.

$R_i = 2K + (1+100)\, 1.67 \times 10^3 = 170.67 K\Omega$

where $A_V = 1 - \frac{h_{ie}}{R_i} = 1 - \frac{2K}{170.67K} = 0.988$

$R_i' = R_i || \frac{R_3}{1-A_V} = 141.66 K\Omega$

Darlington pair with Bootstrapped emitter follower

The fig. shows the Darlington pair with Bootstrap emitter follower circuit.

The AC equivalent circuit for Darlington amplifier is shown below.

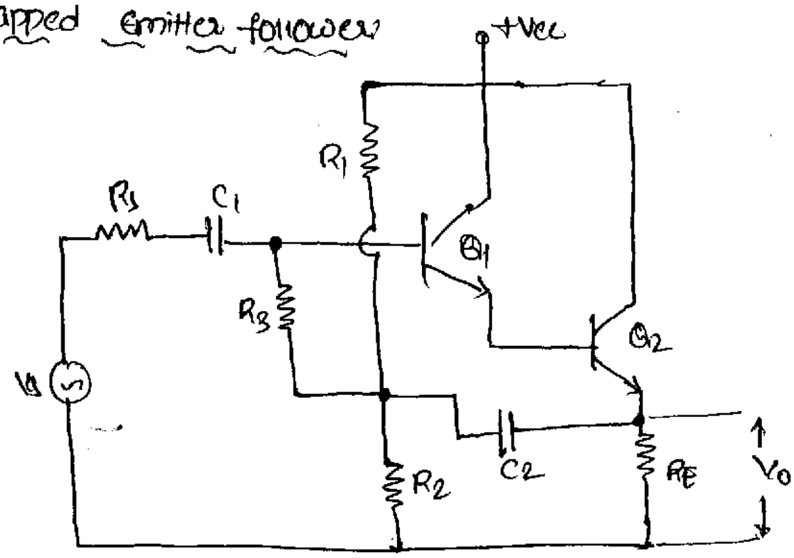


Fig: 2.12 Bootstrap Darlington ckt

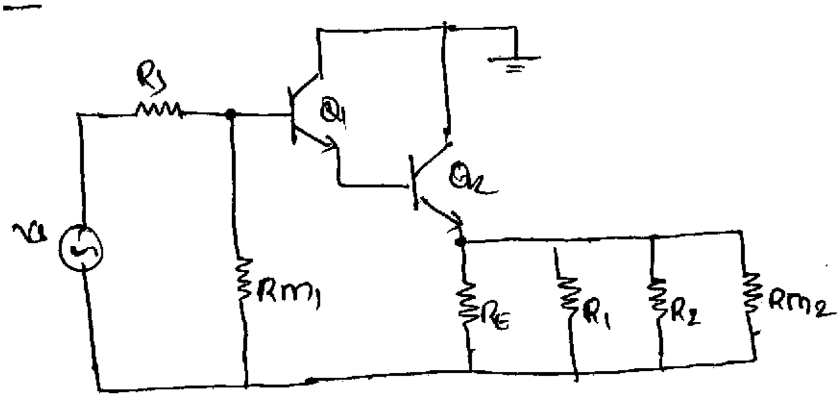


Fig: AC equivalent of 2.12.

let us consider $R_s = 10k\Omega$, $R_1 = 100k\Omega$, $R_2 = 10k\Omega$, $R_3 = 50k\Omega$, $R_E = 1k\Omega$
$h_{ie} = 1k\Omega$, $h_{fe} = 100$, $h_{re} = 2.5\times10^{-4}$ $h_{oe} = 25\mu A/V$.

Note that R_3 is connected b/w i/p & o/p, & has been splitted into R_{m1} & R_{m2} using millers theorem.

Analysis of second stage:

The load resistance R_{L2} for second stage is given by

$$R_{L2} = R_E || R_1 || R_2 || R_{m2}.$$

By miller theorem $R_{m2} = \frac{R_3 A_v}{A_v - 1}$ as $A_v >> 1$, in CC amplifier R_{m2} is very high & neglected.

Hence $R_{L2} = R_E || R_1 || R_2 = 900.9\Omega$.

$h_{oe} R_{L2} = 25\times10^{-6} \times 900.9 = 0.0225 < 0.1$, we go for approximate analysis.

current gain: $A_{I2} = 1+h_{fe} = 101$

input resistance $R_{i2} = h_{ie} + (1+h_{fe}) R_{L2} = 91.99\,k\Omega$

voltage gain $A_{V2} = 1 - \frac{h_{ie}}{R_{i2}} = 0.989.$

Analysis of ~~second~~ first stage:

$R_{L1} = R_{i2} = 91.99\,k\Omega.$

$h_{oe} R_{L1} = 2.299 > 0.1$, go for exact analysis

current gain $A_{I1} = \frac{1+h_{fe}}{1+h_{oe}R_{L1}} = \frac{1+100}{1+25\times10^{-6}(91.99k)} = 30.6.$

input resistance $R_{i1} = h_{ie} + A_{I1}R_{L1} = 1K + (30.6 \times 91.99K)$

voltage gain $A_{V1} = 1 - \frac{h_{ie}}{R_{i1}}$.

overall voltage gain $A_V = A_{V1} \times A_{V2}$.

Overall input resistance (R_i)

$R_i = R_{i1} \| R_{m1}$ $\quad R_{m1} = \frac{R_3}{1-A_V}$

or =

Overall voltage gain A_{VS}

$$A_{VS} = \frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s} = \frac{A_V R_i}{R_i + R_s}$$

Output resistance R_o.

$R_{o1} = \frac{R_{s1} + h_{ie}}{1+h_{fe}}$ $\quad R_{s1} = R_s \| R_{m1}$

$R_{o2} = \frac{R_{s2} + h_{ie}}{1+h_{fe}} = \frac{R_{o1} + h_{ie}}{1+h_{fe}}.$

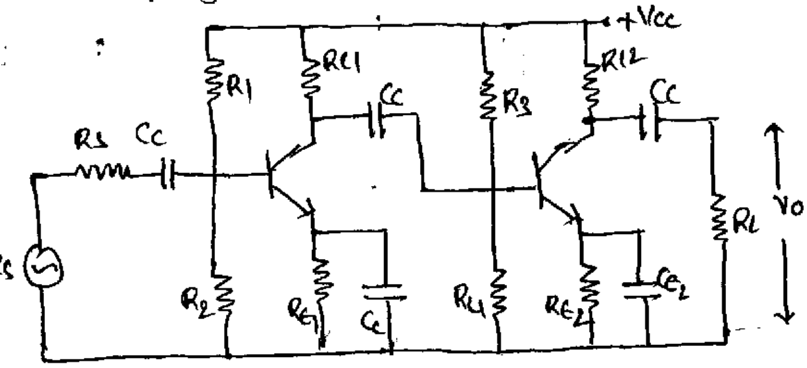
$R_o = R_{o2} \| R_{L2} = ?$

Different coupling schemes used in amplifiers:

In multistage amplifier, the output of preceeding stage is to be coupled to input of p succeding stage. For this interstage coupling, different types of coupling elements are used
Technique

* RC coupling
* Transformer coupling
* Direct coupling.

RC COUPLING:

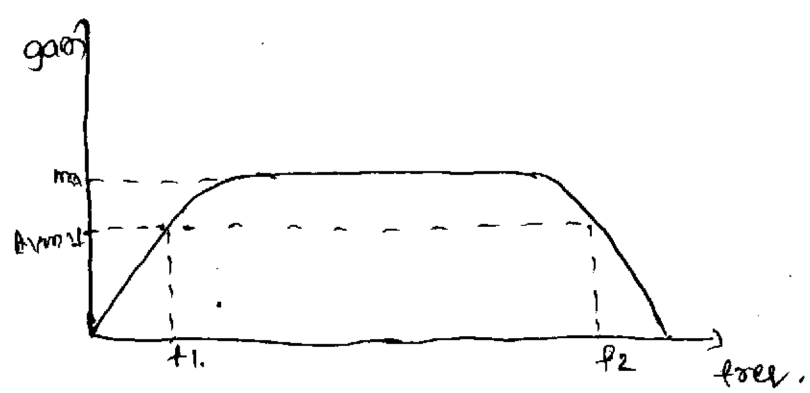


The Fig 2.14 shows the RC coupled amplifier, in which the output signal of first stage is connected to the input of next stage through coupling capacitors & resistive load at output of first stage.

The coupling does not affect Q point of next stage since the coupling capacitor Cc blocks dc voltage of first stage from reaching the base of second stage.

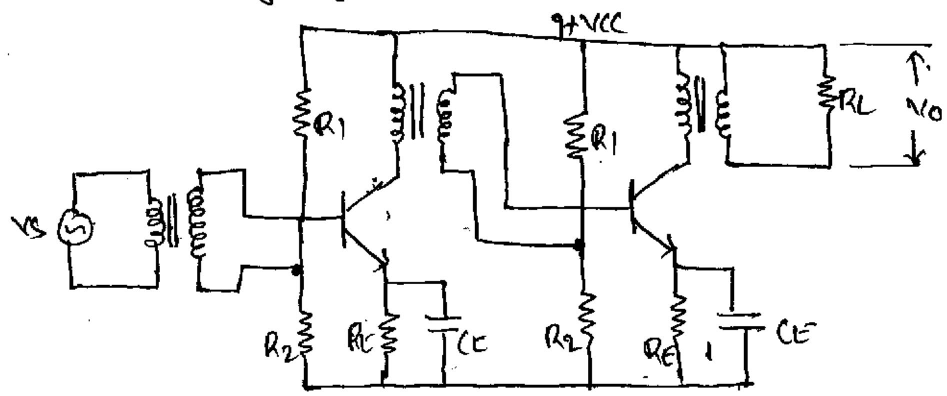
The RC network is broad band in nature. Therefore, it gives a wide band frequency response without peak at any frequency & hence used to cover a complete audio frequency band. However the frequency response drops off at very low frequencies due to coupling capacitors & also at high frequencies due to shunt capacitance such as stray capacitance.

The frequency response of RC coupled amplifier is shown below.



Transformer coupling:

The following figure shows the transformer coupled amplifier.



The output signal of first stage is coupled to input of next stage through an impedance matching transformer.

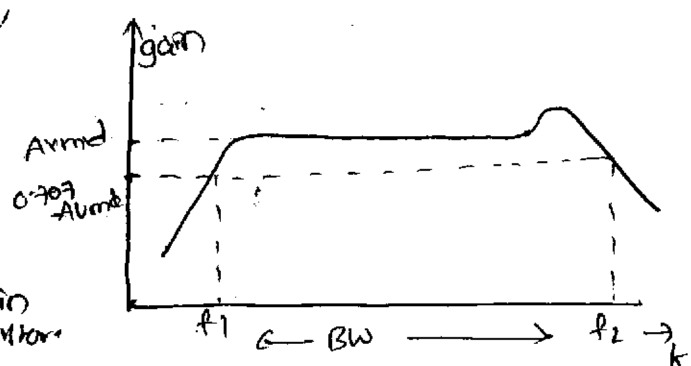
This type of coupling is used to match impedance b/w output and input cascaded stage. Usually it is used to match larger o/p resistance of AF power amplifiers to a low impedance load like loud speaker. We know that transformer blocks dc, providing dc isolation b/w two stages. Therefore transformer coupling does not effect the Quiscent point of next stage.

Frequency response of transformer coupled amplifier is poor compared with RC coupled. Its leakage inductance & interwinding capacitance does not allow amplifier to amplify signals of different frequencies equally. Interwinding capacitance of transformer coupled may give rise resonance at certain freq, which makes amplifier to give very high gain at that frequency. By putting shunting capacitors across each winding of transformer, we can get resonance at any desired RF frequency. Such amps are called "Tuned voltage ampl"

These provide high gain at desired freq i.e., they amplify selected frequencies. For this reason, TC Amps are used in radio and TV receivers to amplify RF signals.

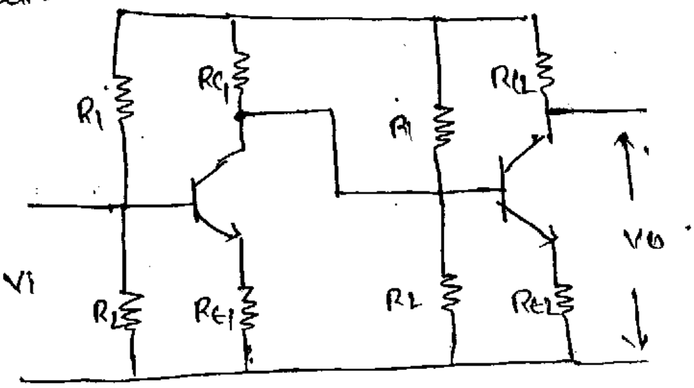
The dc resistance of T winding is very low, almost all dc voltages applied by Vcc is available at collector.

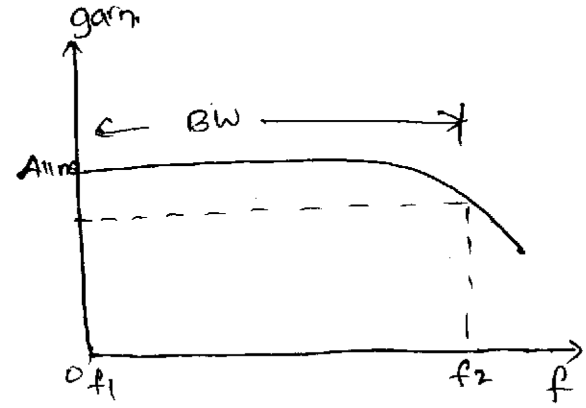
Due to absence of collector resistance, it also eliminates unnecessary power loss in resistors.



Direct coupling:

The following figure shows Direct coupled amplifier using transistor.





The o/p signal of first stage is directly connected to input of next stage. This direct coupling allows Quiscent DC collector current of first stage to pass through base of next stage affecting biasing condition.

Due to the absence of RC components, its low frequency response is good but at higher frequencies shunting capacitance such as stray capacitances reduce gain of amplifier.

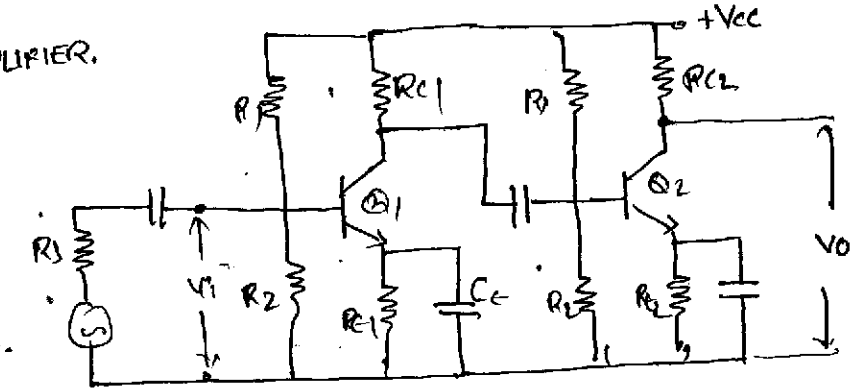
The transistor parameters such as V_{BE} & β change with temperature causing collector current & voltage to change. Because of direct coupling these changes appear at base of next stage & hence in o/p. Such unwanted change in o/p is called current drift & it is serious problem in direct coupled amplifiers.

COMPARISON.

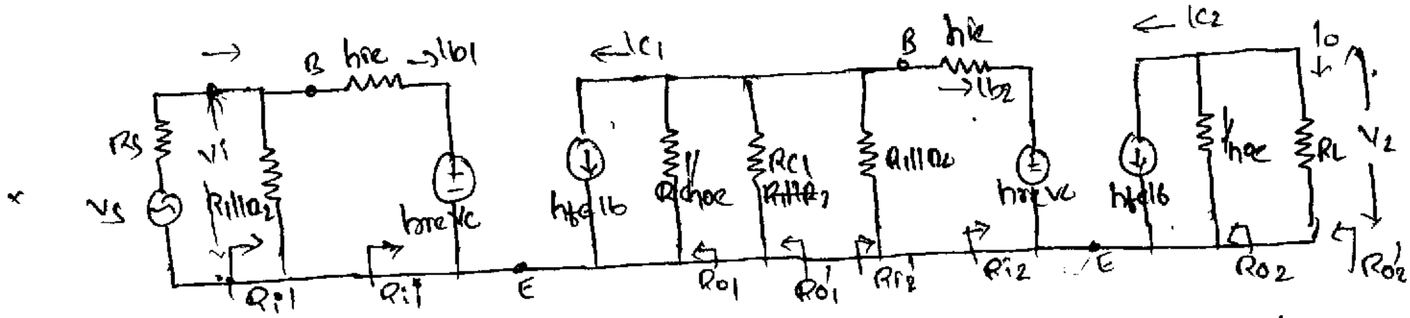
PARAMETER	RC coupled	Transformer	Direct coupled
coupling components	Resistor & capacitor	impedance matching transformer	—
Frequency response	flat at middle frequencies	uniform, high at resonant frequency	flat at middle frequencies
weight	light	Bulky & heavy	—
Application	used in audio, small signal amplifiers	used in amplifiers where impedance matching is an important criteria	used in amplification of slow varying parameters & where DC amplification is required.

ANALYSIS OF RC COUPLED AMPLIFIER.

The figure shows CE-CE cascaded amplifier with RC coupling and their biasing arrangements.



The h-parameter equivalent circuit for CE-CE cascade amplifier is shown in below figure.



Let us calculate R_i, A_I, A_V, R_i', A_{VS} & A_{IS} for $R_s = 1K$, $R_{C1} = 15K$, $R_{E1} = 100\Omega$, $R_{C2} = 4K$, $R_{E2} = 300\Omega$, $R_1 = 200\Omega$ & $R_2 = 20K\Omega$ for first stage, R_{E2} $47K\Omega$, $R_2 = 4.7K$ for second stage. Assume $h_{ie} = 1.2K\Omega$, $h_{fe} = 50$, $h_{re} = 2.5\times10^{-4}$, $h_{oe} = 25\times10^{-6}$ A/V.

Analysis of 2nd stage.

$h_{oe}R_L = h_{oe}R_{C2} = 0.1$

so for approximate

$A_i = -h_{fe} = -50$

$R_{i2} = h_{ie} = 1200\Omega$

$A_{V2} = \dfrac{A_{i2}R_{L2}}{R_{i2}} = \dfrac{-50\times4K}{1.2K} = -166.67$

Analysis of 1st stage

$R_L' = R_{C1}\|R_1\|R_2\|R_{i2} = 881.8\Omega$

$h_{oe}R_L' = 25\times10^{-6}\times881.8 = 0.022 < 1$

so for approximate

$A_i = -h_{fe} = -50$

$R_{i1} = h_{ie} = 1200\Omega$

$A_{V1} = \dfrac{A_{i1}R_L'}{R_{i1}} = \dfrac{-50\times881.8}{1.2K} = -36.74$

Total gain $A_V = A_{V1}\times A_{V2}$

$= -(166.67)(-36.74) = 6123.45$

$A_{VS} = \dfrac{A_V R_i'}{R_i' + R_s}$

$R_i' = R_1\|R_2\|R_{i1} = 1.13K\Omega$

$A_{VS} = \dfrac{6123.45\times1.13K}{1.13K+1K} = 3248.6$

Output res R_o

$R_{o1} = \infty$ $R_{o1}' = R_{o1}\|R_{C1} = R_{C1} = 15K$

$R_{o2} = \infty$ $R_{o2}' = R_{o2}\|R_{C1} = R_{C2} = 4K\Omega$.

$A_{IS}: \dfrac{I_o}{I_s} = \underbrace{\dfrac{I_o}{I_{c2}}}_{-1}\times\underbrace{\dfrac{I_{c2}}{I_{b2}}}_{-A_{i2}}\times\underbrace{\dfrac{I_{b2}}{I_{c1}}}_{-1}\times\underbrace{\dfrac{I_{c1}}{I_{b1}}}_{-A_{i1}}\times\dfrac{I_{b1}}{I_s}$

$\dfrac{I_{b1}}{I_s} = \dfrac{R_B}{R_B\|R_{i1}}$ $\dfrac{I_{b2}}{I_{c1}} = -$

Boot strap problems.

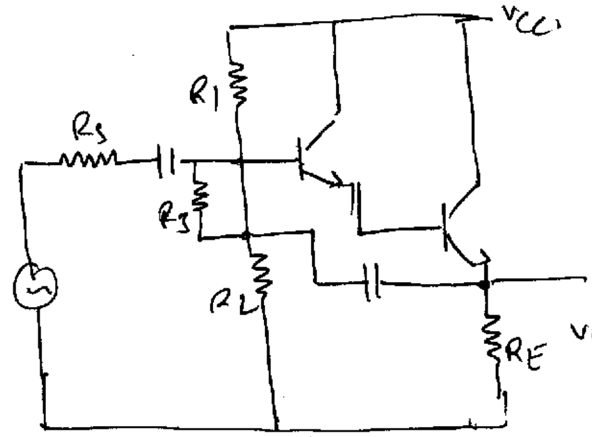
$R_s = 10K$ $200K$.

$R_1 = 470K$

$R_2 = 10K$

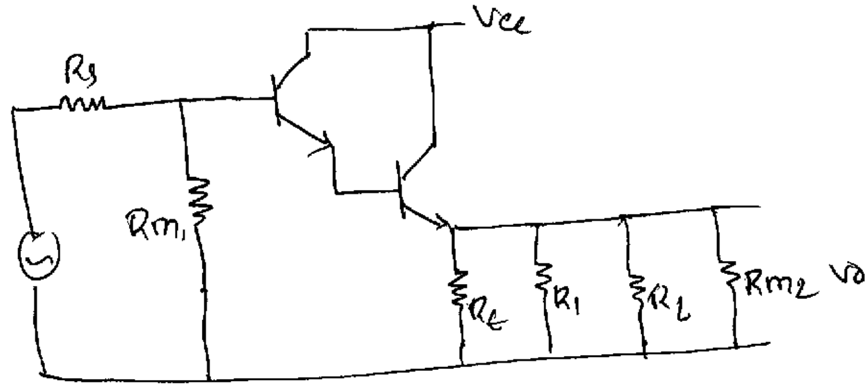
$R_3 = 68K$.

$R_E = 1.2K$.



Find. A_i, A_{vs} R_i, R_o.

h parameter are typical.



R_o.

R_L for 2nd stage is given by

$= R_E \| R_1 \| R_2 \| R_{m2}$.

$R_{L2} = R_E \| R_1 \| R_2$ R_{m2} is neglected

$h_{oe} R_{L2} = 0.0267 < 0.1$

Approximate

$A_{i2} = 1 + h_{fe} = 1 + 50 = 51$

$R_{i2} = h_{ie} + (1+h_{fe}) R_{L2} = 556.2 K\Omega$.

$A_{v2} = 1 - \dfrac{h_{ie}}{R_{i2}} = 0.98$

Analysis of 1st stage

$h_{oe} R_{L1} = h_{oe} R_{i2} = 1.39 > 0.1$

Exact analysis.

$A_{i1} = \dfrac{1+h_{fe}}{1 + h_{oe} R_{L1}} =$

$R_{i1} = h_{ie} + A_{i1} R_{L1}$

$A_{v1} = 1 - \dfrac{h_{ie}}{R_{i1}}$

$A_v = A_{v1} \times A_{v2}$

$R_i = R_{i1} \| R_{m1}$ $R_{m1} = \dfrac{R_E}{1 - A_v}$.

$R_i = 868 K\Omega$.

$A_{vs} =$

$\dfrac{V_o}{V_s} = \dfrac{V_o}{V_i} \times \dfrac{V_i}{V_s}$

$\dfrac{V_o}{V_i} \times \dfrac{R_i}{R_i + R_s} = 0.781$

$R_o =$

$R_{o1} = \dfrac{R_{s1} + h_{ie}}{1 + h_{fe}}$

$R_{s1} = R_{m1} \| R_s = 206 K\Omega$

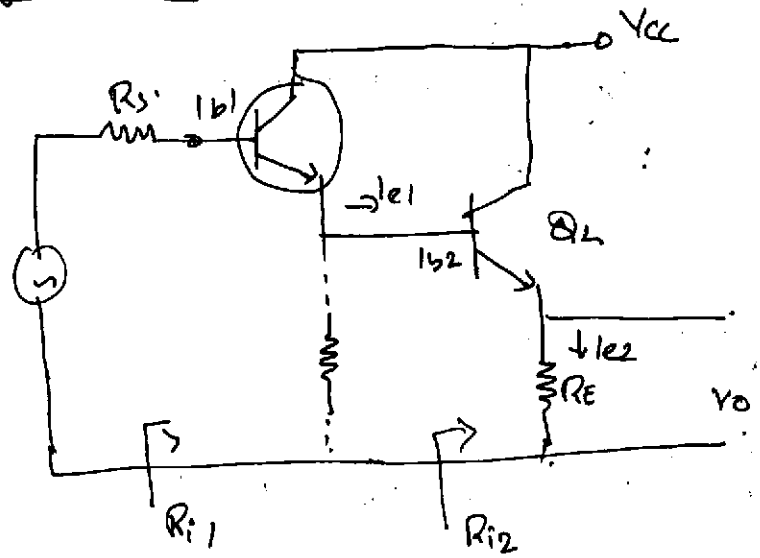
$R_{o1} = 4.06 K\Omega$

$R_{o2} = \dfrac{R_{s2} + h_{ie}}{1+h_{fe}} = \dfrac{R_{o1} + h_{ie}}{1+h_{fe}}$

$= 160\Omega$

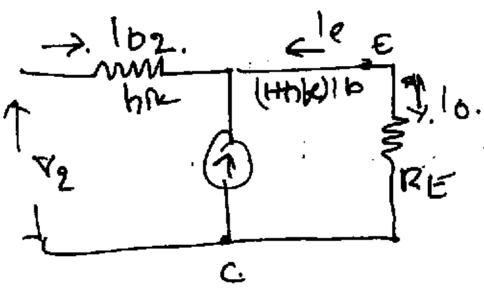
$R_o = R_{o2} \| R_{L2} = 12.8\Omega$.

Darlington Analysis.



Assume $h_{oe} R_L < 0.1$

Analysis of second stage.



$$A_{i2} = \frac{I_o}{I_b} = \frac{-I_e}{I_b}$$

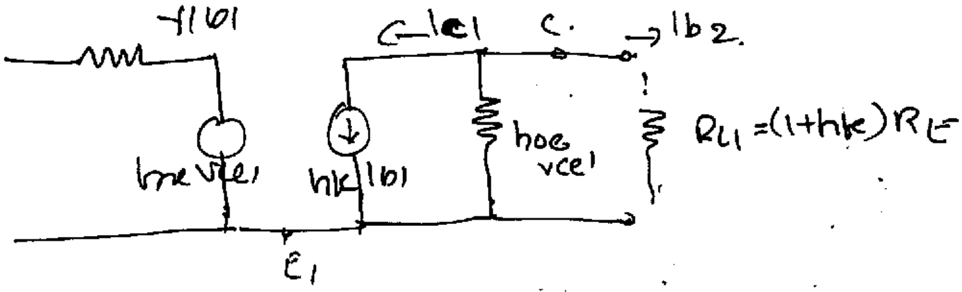
$$= \frac{I_b + h_{fe} I_b}{I_b}$$

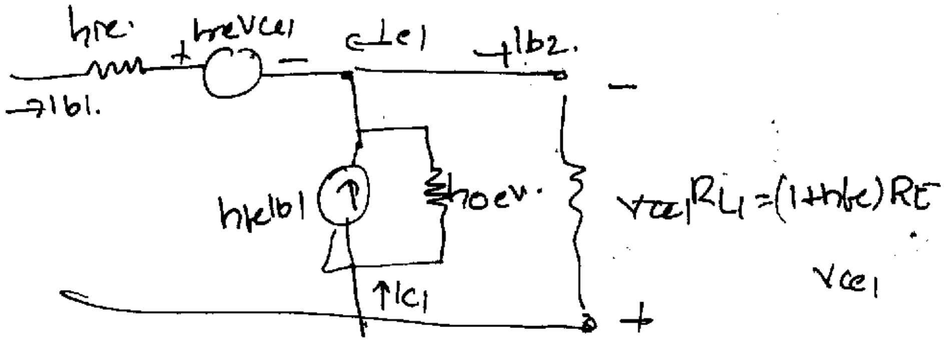
$$A_{i2} = (1 + h_{fe})$$

$$R_{i2} = \frac{V_2}{I_{b2}} = (1 + h_{fe}) R_E$$

Analysis of 1st stage

the R_L of 1st stage is input resistance of 2nd stage. ie R_{i2}. As R_{i2} is high, it does not meet requirement of $h_{oe} R_{i2} < 0.1$, hence we have to go for exact analysis for 1st stage.





$$A_{i1} = \frac{I_{b2}}{I_{b1}} = \frac{I_{e1}}{I_{b1}}$$

$$I_{e1} = -(I_{b1} + I_{c1})$$

$$I_{c1} = h_{fe} I_{b1} + h_{oe} V_{ce1} = h_{fe} I_{b1} + h_{oe}(-I_{b2} R_{L1})$$

$$= h_{fe} I_{b1} + h_{oe} I_{e1} R_{L1}$$

$$I_{e1} = -(I_{b1} + h_{fe} I_{b1} + h_{oe} I_{e1} R_{L1}) = -I_{b1} - h_{fe} I_{b1} - h_{oe} I_{e1} R_{L1}$$

$$I_{e1} + h_{oe} I_{e1} R_{L1} = -I_{b1}(1+h_{fe})$$

$$\frac{-I_{e1}}{I_{b1}} = \frac{1+h_{fe}}{1+h_{oe} R_{L1}}$$

WKT. $R_{L1} = (1+h_{fe}) R_E$

$$A_{i1} = \frac{1+h_{fe}}{1+h_{oe}(1+h_{fe})R_E} = \frac{1+h_{fe}}{1+h_{oe}(h_{fe})R_E} \quad \because h_{fe} >> 1$$

R_{i1} $\frac{V_o}{I_{b1}}$

Apply KVL to o/p loop

$$V_i - I_{b1}h_{ie} - h_{re}V_{ce} + V_{ce1} = 0$$

$$V_i = I_{b1}h_{ie} + h_{re}V_{ce1} - V_{ce1}.$$

$h_{re}V_{ce}$ is negligible.

$$V_i = I_{b1}h_{ie} - (-I_{b2}R_{L1})$$

$$V_i = I_{b1}h_{ie} + I_{b2}R_{L1}$$

$$R_{i1} = \frac{V_i}{I_{b1}} = h_{ie} + \frac{I_{b2}}{I_{b1}}R_{L1} = h_{ie} + A_{i1}R_{L1}$$

$$\therefore R_{i1} = h_{ie} + A_{i1}(1+h_{fe})R_E$$

Sub A_{i1}

$$R_{i1} = h_{ie} + \frac{(1+h_{fe})^2 R_E}{1+h_{oe}h_{fe}R_E}$$

$$R_{i1} \cong \frac{(1+h_{fe})^2 R_E}{1+h_{oe}h_{fe}R_E} \quad \because h_{ie} <<$$

Current gain A_i

$$A_i = A_{i1} \times A_{i2}$$

$$A_i = \frac{(1+h_{fe})^2}{1+h_{oe}(1+h_{fe})R_E}$$

	single	Darlington.
R_i	168.3K	$R_i = 1.69M\Omega$ for
A_i	51	$A_i = 500$

$R_E = 3.3K$.
h_{ie} 1100 $h_{re} = 2.5\times10^{-4}$.

$$A_v = \quad A_v = \frac{A_i R_L}{R_i}$$

Subtracting 1 on both sides

$$1 - A_V = 1 - \frac{A_i R_L}{R_i} = \frac{R_i - A_i R_L}{R_i}$$

$$= \frac{h_{ic} + h_{rc}A_i R_L - A_i R_L}{R_i}$$

$$= \frac{h_{ie}}{R_i} \quad \text{since } h_{rc} = h_{ie} \ \& \ h_{rc} = 1 - h_{re} \simeq 1$$

$$A_V = 1 - \frac{h_{ie}}{R_i}$$

W.K.T. $A_V = A_{V1} \times A_{V2}$

$$= \left(1 - \frac{h_{ie}}{R_{i1}}\right)\left(1 - \frac{h_{ie}}{R_{i2}}\right).$$

$$A_V = 1 - \frac{h_{ie}}{R_{i2}} - \frac{h_{ie}}{R_{i1}} + \frac{h_{ie}^2}{R_{i1}R_{i2}}$$

$$A_V \cong 1 - \frac{h_{ie}}{R_{i2}} \qquad \because R_{i1} >> R_{i2}$$

$\underline{R_{o2}}$: $\quad R_o = \frac{1}{Y_o}$

$$Y_{o1} = h_{oc} - \frac{h_{fe} h_{rc}}{h_{ic} + R_s}$$

$$= h_{oc} - \frac{-(1+h_{fe})}{h_{ie} + R_s} \qquad \because h_{oc} = h_{oe}, \; h_{fc} = -(1+h_{fe}), \; h_{ic} = h_{ie}$$

$$Y_{o1} = h_{oe} + \frac{(1+h_{fe})}{h_{ie} + R_s}$$

$$Y_{o1} = \frac{1+h_{fe}}{h_{ie} + R_s} \qquad \because h_{oe} << \frac{1+h_{fe}}{h_{ie} + R_s}$$

$$R_{o1} = \frac{h_{iei} + R_s}{1 + h_{fe}}.$$

R_{o1} becomes source resistance of 2nd stage.

$$R_{s2} = R_{o1}$$

$$R_{o2} = \frac{R_{s2} + h_{ie2}}{1 + h_{fe}}$$

$$= \frac{\left(\frac{h_{ie1} + R_s}{1 + h_{fe}}\right) + h_{ie2}}{1 + h_{fe}}$$

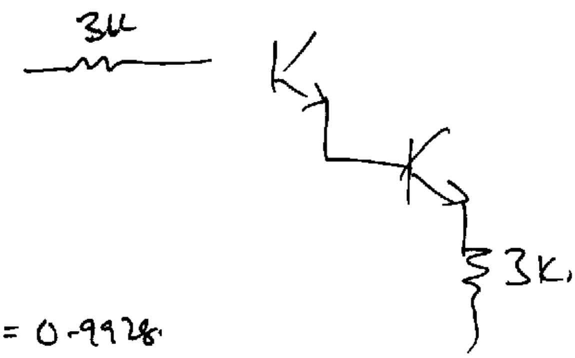
$$R_{o2} = \frac{h_{ie1} + R_s}{(1 + h_{fe})^2} + \frac{h_{ie2}}{1 + h_{fe}}$$

current in T_2 is $1+h_{fe}$ times current in T_1

$$h_{ie1} \simeq (1 + h_{fe})\, h_{ie2}.$$

$$R_{o2} = \frac{R_s}{(1 + h_{fe})^2} + \frac{2h_{ie2}}{1 + h_{fe}}$$

Ex:



2nd

$A_{i2} = 51$

$R_{i2} = 154.1K.$

$A_{v2} = 1 - \frac{h_{ie}}{R_{i2}} = 0.9928.$

1st.

$R_E = R_{i2} = 154.1K.$

$$A_{i1} = \frac{1 + h_{fe}}{1 + h_{oe}(1 + h_{fe})R_{i2})}$$

$R_{i1} =$

$A_{v1} = 0.999$

$A_v = 0.9918$

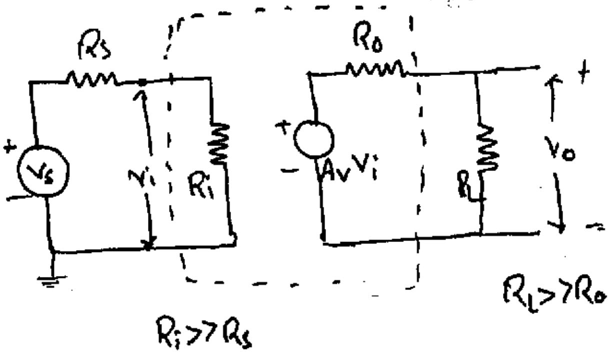
$R_{o1} = 80.39.$

$R_{o2} = 23.145$

$R_o = R_{o2} \| R_L$

$= 22.97\,\Omega.$

Voltage amplifier:



Thevinins equivalent of voltage amp.

If $R_i >> R_s$ then $V_i \simeq V_s$.

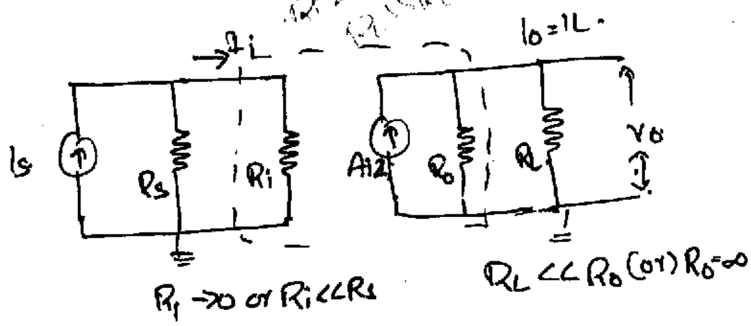
If $R_L >> R_o$ then $V_o = A_vV_i = A_vV_s$

Such amp ckt provides a voltage o/p proportional to input voltage & the proportionality factor does not depend on magnitude of R_L & R_s. Hence it is called voltage amplifier.

ideal $R_i \to \infty$

practical $R_i >> R_s$ $R_L >> R_o$.

current amplifier:



If $R_i \to 0$ then $I_i = I_s$

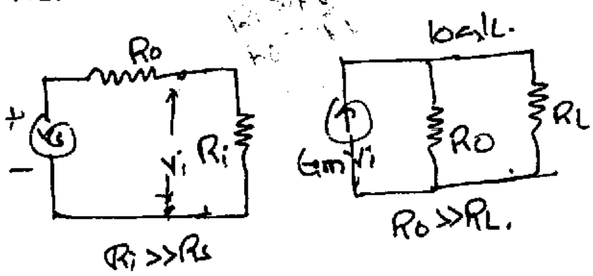
If $R_o \to \infty$ then $I_L = A_i I_i$

Such amp. provides current o/p proportional to signal current. factor independent of R_s & R_L.

ideal $R_i = 0$ $R_o = \infty$

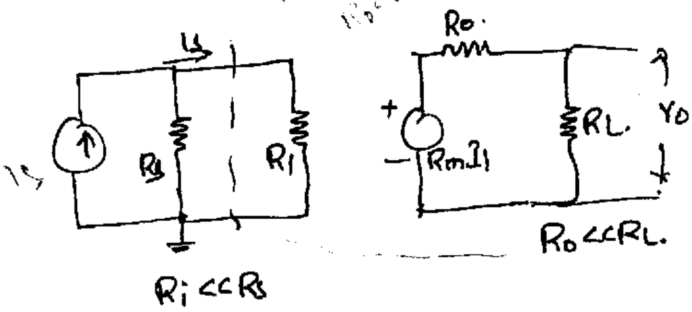
prac $R_i << R_s$ $R_o >> R_L$

Transconductance



$V_i = V_s$, $I_L = G_mV_i$

Trans resistance



$I_i = I_s$ $V_o = R_mI_s$

Concept of feedback:

We have seen four basic amplifier types & their ideal - characteristics. In each of these circuits we can sample the o/p voltage or current by means of a suitable sampling network and apply this signal to input through feedback two port network. as shown in fig. At input, feedback signal is combined with input signal through a mixer network & is fed to amplifier.

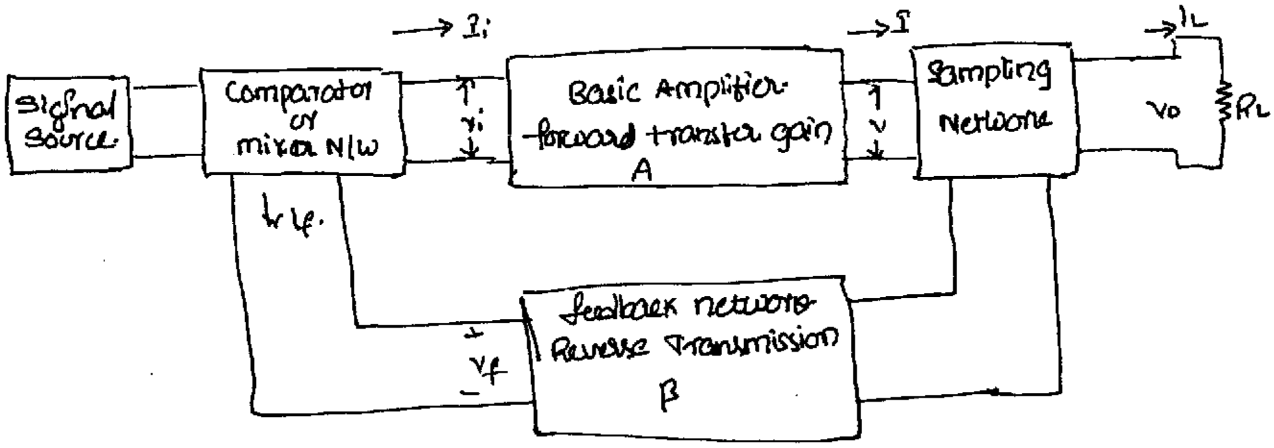


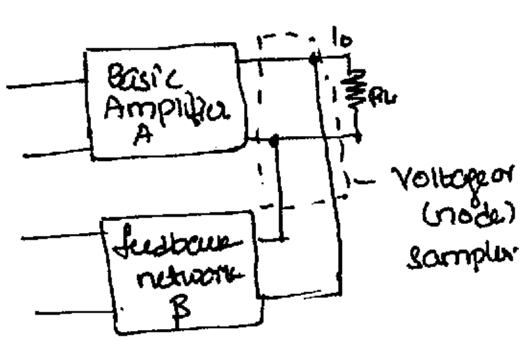
Fig: Typical feedback connection around basic amplifier

a). Sampling N/W b) Feedback N/W c) mixer N/W

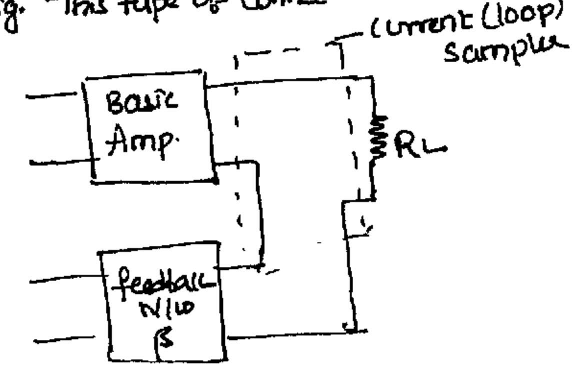
Sampling N/W:

There are two ways to sample the output, voltage or current, according to the sampling parameter.

The o/p voltage is sampled by connecting feedback network in shunt across o/p as shown in fig. This type of Connec. is called Voltage & node sampling.



Voltage or node sampling



Current or loop sampling.

The o/p current is sampled by connecting FB N/W in series with o/p as shown. This type of connection is referred to as current or loop sampling.

Feedback Network:

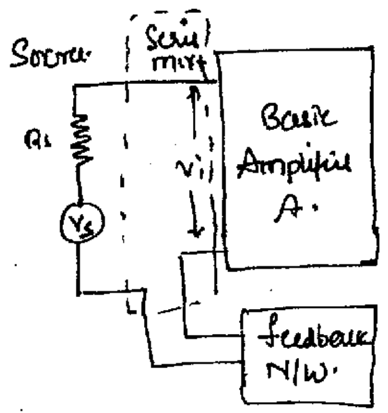
It may consist of resistors, capacitors & inductors. Most often it is simply a resistive configuration. It provides reduced portion of output as feedback-signal to input mixer network. It is given as.

$$\boxed{V_f = \beta V_o}$$

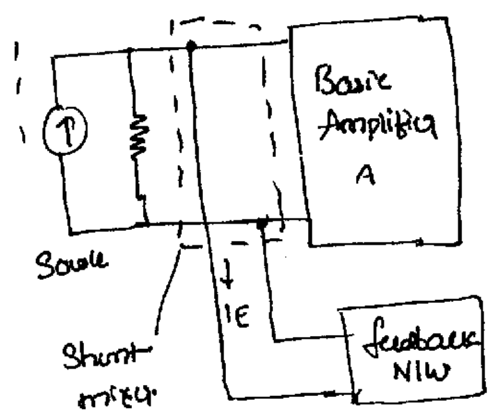
Where β is a feedback factor (or) feedback ratio. It always lies between 0 and 1. β is totally different from β symbol used to represent current gain in CE amp., which is greater than 1.

Mixer Network:

Like sampling, there are two ways of mixing feedback signal with input signal. They are: Series input connection
shunt input connection



Series mixer N/W.

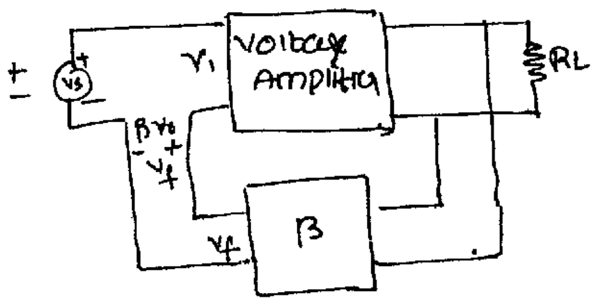


Shunt mixer N/W

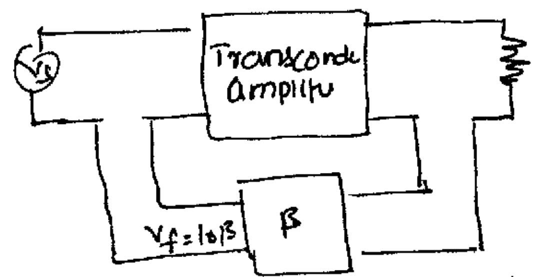
The basic amplifier shown in fig(1) may be a voltage, current transconductance or trans-resistance amplifier. these can be connected in feedback configuration as shown below.

There are four basic ways of connecting feedback. they are.

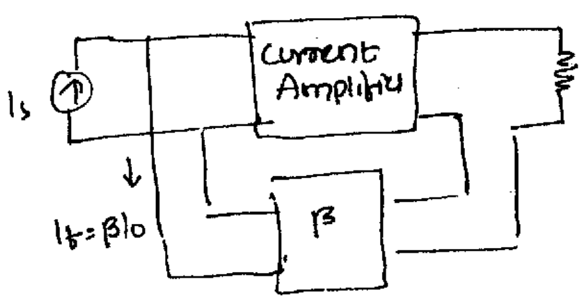
* Voltage series FB
  current series FB
  Voltage shunt FB
  current shunt FB



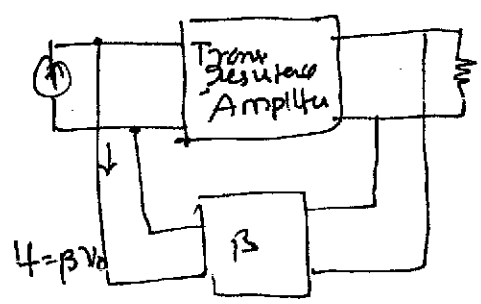
Voltage Amp with voltage series FB.



Transconductance Amp with current series FB.



Current amp. with current shunt FB



Transresistance Amp with voltage shunt FB.

Transfer Ratio or gain:

The ratio of o/p signal to input signal of basic amplifier is represented by symbol A. The suffix of A given next, represents different transfer ratios.

$\frac{V}{V_i} = A_V$ = Voltage gain

$\frac{I}{I_i} = A_i$ = Current gain

$\frac{I}{V_i} = G_m$ = Transconductance

$\frac{V}{I_i} = R_m$ = Trans Resistance.

These four quantities A_V, A_I, G_m, R_m are referred to as transfer gain of basic amplifier without feedback. & use of only symbol A represent any one of these quantities.

The transfer gain with feedback is represented by A_f. It is defined as ratio of output signal to input signal of amp config. A_f is used to represent any one of following four ratios.

$\frac{V_o}{V_s} = A_{Vf}$ = Voltage gain with feedback; $\frac{I_o}{I_s} = A_{if}$ = Current gain with feedback

$\frac{I_o}{V_s} = G_{mf}$ = Transconductance with FB; $\frac{V_o}{I_s} = R_{mf}$ = Transresistance with FB.

There are two types of feed back amplifiers.

(1) +ve feedback amplifiers
(2) -ve feedback amplifiers

Positive Feedback: If the feedback signal V_f is inphase with i/p signal V_s then the net effect of feed back increase i.e.) $V_i = V_s + V_f$. Hence i/p voltage applied to basic amplifier increases thereby increasing V_o. This type of feedback is known as +ve feedback (or) regenerative feed back.

$\because V_i = V_s + V_f$.
$V_s = V_i - V_f$.

$$A_f = \frac{V_o}{V_s} = \frac{V_o}{V_i - V_f}$$

$$A_f = \frac{1}{\frac{V_i}{V_o} - \frac{V_f}{V_o}} = \frac{1}{\frac{1}{A} - \beta} = \frac{A}{1-\beta}$$

(2) **-ve feedback Amp:** If the feed back signal V_f is out of phase with input signal, then $V_i = V_s - V_f$. So, the i/p voltage applied to basic amplifier is decreased, correspondingly, the o/p is also decreased. Hence voltage gain reduces. This type of feedback is called -ve feedback. It is also known as degenerative feed back (or) return difference feedback.

$$A_f = \frac{V_o}{V_s} = \frac{V_o}{V_i + V_f} = \frac{A}{1 + A\beta}.$$

Stability of gain:

The transfer gain of amp is not constant as it depends on factors such as operating point, temperature etc. the lack of stability in ampr can be reduced by introducing negative feedback.

We know that $A_f = \dfrac{A}{1+A\beta}$

Differentiate w.r.to A on bot sides

$$\frac{dA_f}{dA} = \frac{(1+\beta A)1 - \beta A}{(1+\beta A)^2} = \frac{1}{(1+A\beta)^2}.$$

$$dA_f = \frac{dA}{(1+A\beta)^2}.$$

Dividing both sides by A_f we get.

$$\frac{dA_f}{A_f} = \frac{dA}{(1+A\beta)^2} \times \frac{1}{A_f} = \frac{dA}{(1+A\beta)^2} \times \frac{1+\beta A}{A}.$$

(2) —— $\dfrac{dA_f}{A_f} = \dfrac{dA}{A}\,\dfrac{1}{1+A\beta}$ where $\dfrac{dA_f}{A_f}$ is fractional change in amplification with feedback.

$\dfrac{dA}{A}$ fractional change in amp without feed back.

$\partial A_f/A_f$ is < $\dfrac{\partial A}{A}$ by factor $(1+\beta A)$. The f.c.A w F divided by F.C A without F is called Sensitivity of transfer gain. Its reciprocal is Desensitivity D.

∴ The stability of amplifier increases with increase in Desensitivity.

If $\beta A >> 1$ then $A_f = \frac{1}{\beta}$ i.e, gain dependent only on FB N/W.

since A represents either Av Gm Ai or Rm &
Af " " Avf Gmf Aif or Rmf , the equation signifies that.

For Voltage series FB $A_{vf} = \frac{1}{\beta}$ ⇒ Voltage gain is stabilized.
Current series FB $G_{mf} = \frac{1}{\beta}$ ⇒ Transconductance G is "
Voltage shunt FB $R_{mf} = \frac{1}{\beta}$ ⇒ Trans resistance G is "
Current shunt FB $A_{if} = \frac{1}{\beta}$ ⇒ Current Gain " "

Frequency response & Bandwidth:

We know that $A_f = \frac{A}{1+\beta A}$

Using this equation we can write

$$A_{fmid} = \frac{A_{mid}}{1+\beta A_{mid}} \quad —(1) \quad ; \quad A_{flow} = \frac{A_{low}}{1+\beta A_{low}} \quad —(2) \quad \& \quad A_{f\,high} = \frac{A_{high}}{1+\beta A_{high}} \quad —(3)$$

Lower cutoff frequency:

We know the relation b/w gain at low frequency & gain at mid freq. is given as.

$$\frac{A_{low}}{A_{mid}} = \frac{1}{1-j(f_L/f)} \quad \therefore \quad A_{low} = \frac{A_{mid}}{1-j\,f_L/f} \quad —(4)$$

Substituting eqn (4) in eqn (2)

$$A_{flow} = \frac{\frac{A_{mid}}{1-jf_L/f}}{1+\beta\frac{A_{mid}}{1-jf_L/f}} = \frac{A_{mid}}{1-jf_L/f+\beta A_{mid}} = \frac{A_{mid}}{(1+\beta A_{mid})-j\,f_L/f}.$$

Dividing by $(1+A_{mid}\beta)$, both Numerator & denominator.

$$A_{flow} = \frac{\frac{A_{mid}}{1+A_{mid}\beta}}{1-j\frac{\frac{f_L}{1+A_{mid}\beta}}{f}} = \frac{A_{fmid}}{1-j\frac{\frac{f_L}{1+A_{mid}\beta}}{f}} = \frac{A_{fmid}}{1-j\frac{f_{Lf}}{f}}.$$

$$\therefore \quad \frac{A_{flow}}{A_{fmid}} = \frac{1}{1-j\frac{f_{Lf}}{f}} \quad \text{where } f_{Lf} = \text{lower cut off freq with FB} = \frac{f_L}{1+A_{mid}\beta}.$$

f_{Lf} is $< f_L$ by a factor $1+A_{mid}\beta$

Upper cutoff frequency:

we know that relation b/w gain at high freq & gain at mid freq. is given as

$$\frac{A_{high}}{A_{mid}} = \frac{1}{1-j\,f/f_H} \Rightarrow A_{high} = \frac{A_{mid}}{1-j\,f/f_H} \quad -⑤$$

Sub. ⑤ in ③

$$A_f\,high = \frac{A_{mid}}{1+A_{mid}\beta - j\,f/f_H}$$

Dividing both N & D with $(1+\beta A_{mid})$

$$A_f\,high = \frac{\frac{A_{mid}}{1+\beta A_{mid}}}{1-j\frac{f}{(1+\beta A_{mid})f_H}} = \frac{A_{fmid}}{1-j\,f/f_{HF}}$$

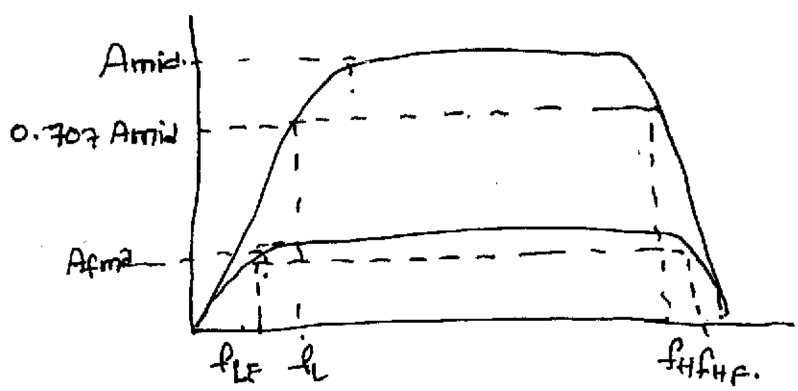
where upper cutoff frequency with feedback is $f_{Hf} = (1+A_{mid}\beta)f_H$ —⑥

from ⑥ we can say that upper cutoff frequency with feedback is greater than upper cutoff frequency without feedback by a factor $(1+\beta A_{mid})$ ∴ introducing -ve feedback., high freq response of amp is improved.

Bandwidth = Upper cutoff Freq − lower cutoff freq.

$$BW_f = f_{Hf} - f_{Lf} = (1+A\beta_{mid})f_H - \frac{f_L}{(1+A\beta_{mid})}$$

It is very clear that $(f_{Hf} - f_{Lf}) > (f_H - f_L)$. & hence BW of amp with feedback is greater than BW of amp without feedback.



Frequency distortion:

If the feedback network does not contain reactive elements, the overall gain is not a function of frequency. Under such conditions frequency & phase distortion is substantially reduced.

If β is made up of reactive components, reactance of components will change with frequency, changing the β. As a result gain will also change with frequency.

This fact is used in tuned amplifiers. In TA, feedback network is designed such that at tuned frequency $\beta \to 0$ & at other frequencys $\beta \to \infty$. As a result, amplifier provides high gain for signal at tuned frequency & relatively reject all frequencies.

Noise & Non linear Distortion:

Signal feedback reduces amount of noise signal & non linear distortion. The factor $(1+A\beta)$ reduces both input noise & resulting non linear distortion for considerable improvement. Thus, noise & " " " also reduced by same factor as the gain.

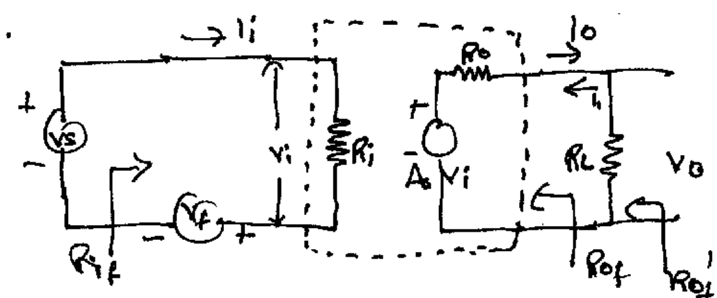
INPUT RESISTANCE:

We discuss the effect of topology of feedback amplifier upon input resistance.

If the feedback signal is returned to input in "series" with applied voltage, it increases input resistance.

Since feedback voltage V_f opposes V_s, the input current I_i is less than it would be if V_f were absent. Hence the input resistance with feedback.

$$R_{if} = \frac{V_s}{I_i}$$ is greater than input resistance with out feedback, as shown.

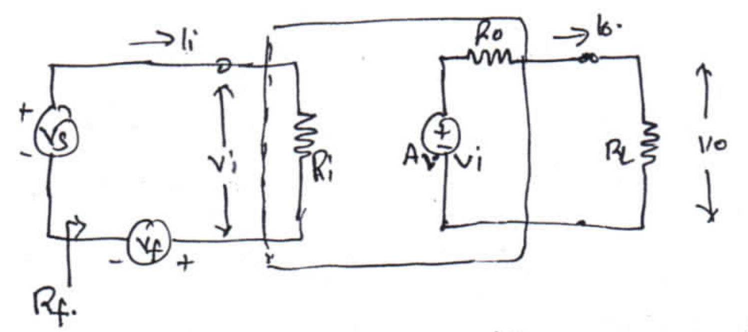


If the feedback signal is added to input in shunt with applied voltage (regardless of sampled V or C), it decreases the input resistance.

Since $I_s = I_i + I_f$ Hence i/p resistance with feedback is decreased

Now we'll see effect of -ve feedback on different topologies.

(i) <u>voltage series feedback</u> i/p & o/p thevenin



$$R_{if} = \frac{V_s}{I_i}$$

Apply KVL to input. $V_s - I_i R_i - V_f = 0$

$$V_s = I_i R_i + V_f = I_i R_i + \beta V_o \quad \text{—(1)}$$

The o/p voltage V_o is given as. $V_o = \frac{A_v V_i R_L}{R_o + R_L}$ —(2)

$$\Rightarrow \quad AV = \frac{V_o}{V_i} = \frac{A_v R_L}{R_o + R_L} \quad \text{—(2)}$$

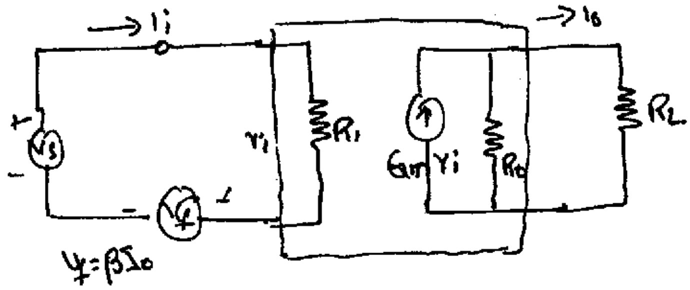
Note: A_v represents open circuit voltage gain without feedback. & AV is voltage gain without feedback taking R_L into account

(2) in (1) $V_s = I_i R_i + \beta AV I_i R_i$

$$\frac{V_s}{I_i} = R_i(1 + \beta AV).$$

Current Series feedback: i/p Thevenin o/p Norton

$R_{if} = \frac{V_s}{I_i}$



Apply KVL to I/p

$V_s - I_i R_i - V_f = 0$

$V_s = I_i R_i + \beta I_o$

The o/p current I_o is

$I_o = \frac{G_m V_i R_o}{R_o + R_L} = G_M V_i$

where $G_M = \frac{G_m R_o}{R_o + R_L}$.

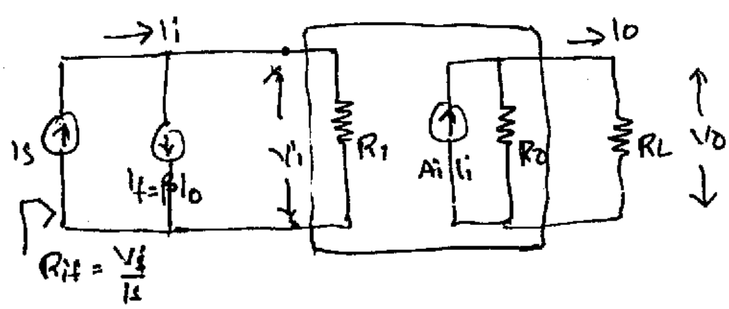
$V_s = I_i R_i + \beta G_M V_i$

$V_s = I_i R_i + \beta G_M I_i R_i$

$R_{if} = \frac{V_s}{I_i} = R_i (1 + \beta G_M)$

Current shunt feedback:

i/p & o/p replaced by Norton's equivalent.



Applying KCL at input node. $I_s = I_i + I_f$.

$= I_i + \beta I_o$

I_o is given by $I_o = \frac{A_i I_i R_o}{R_o + R_L} = A_I I_i$ where $A_I = \frac{A_i R_o}{R_o + R_L}$.

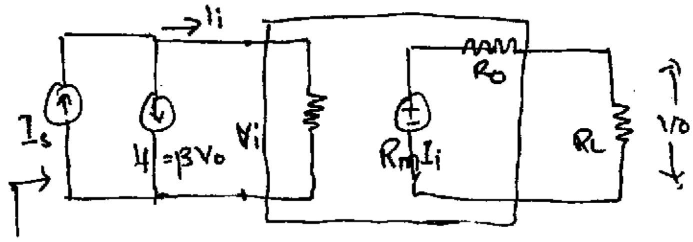
Note A_i represents open circuit current gain without feedback &
A_I " " " " " " taking R_L in count.

$I_s = I_i + \beta A_I I_i = I_i (1 + \beta A_I)$

$\therefore R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i (1 + \beta A_I)} = \frac{R_i}{1 + \beta A_I}$ //

Voltage shunt feedback:

The VSF topology is shown with amplifier input represented by Norton equivalent & O/p represented by thevinin equivalent.



Apply KCL at input

$I_s = I_i + I_f$

$= I_i + \beta V_o$

$R_i = \frac{V_i}{I_s}$

$V_o = \frac{R_m I_i R_L}{R_o + R_L}$

$V_o = R_M I_i$

where $R_M = \frac{R_m R_L}{R_o + R_L}$

$I_s = I_i + \beta R_M I_i$

$I_s = I_i (1 + \beta R_M)$

$R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i (1 + \beta R_M)} = \frac{R_i}{1 + \beta R_M}$ //.

Output resistance:

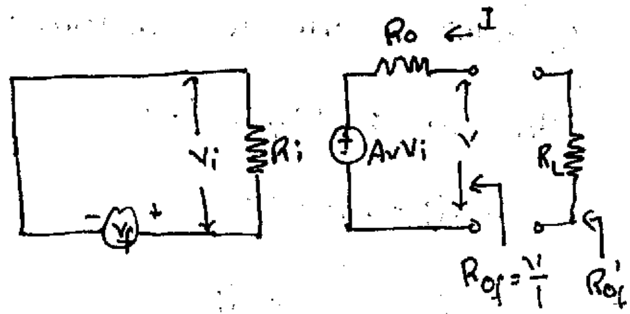
The negative feedback which samples the output voltage, tends to decrease the O/p resistance.

The negative feedback which samples the output current tends to increase the O/p resistance.

Voltage series feedback:

In this topology, O/p resistance can be measured by shorting input source $V_s=0$ & looking into O/p terminals with R_L disconnected as shown in fig:

Apply KVL to output loop, we get



$$A_vV_i + IR_o - V = 0$$

$$I = \frac{V - A_vV_i}{R_o} \quad \text{--- (1)}$$

The input voltage $V_i = -V_f = -\beta V \quad \because V_s=0 \quad \text{--- (2)}$

(2) in (1)

$$I = \frac{V - A_v(-\beta V)}{R_o} = \frac{V(1+A\beta)}{R_o}$$

$$\therefore \quad R_{of} = \frac{V}{I} = \frac{R_o}{1+\beta A_v} \quad \text{--- (4)}$$

Note: A_v is open loop voltage gain without taking R_L in account.

$$R_{of}' = R_{of} \parallel R_L$$

$$= \frac{\frac{R_o}{1+\beta A_v} \times R_L}{\frac{R_o}{1+\beta A_v} + R_L} = \frac{R_oR_L}{R_o + R_L + \beta A_vR_L}$$

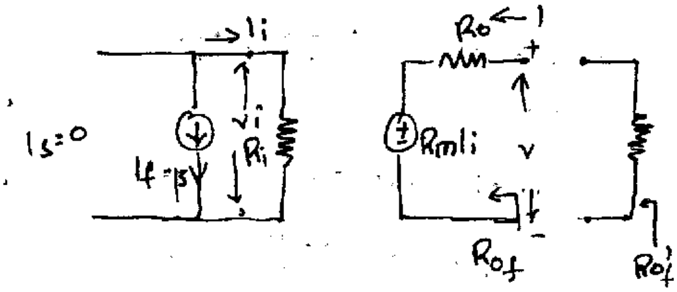
Dividing Numerator & Denominator by (R_o+R_L)

$$R_{of}' = \frac{\frac{R_oR_L}{R_o+R_L}}{1+\beta \frac{A_vR_L}{R_o+R_L}} = \frac{R_L'}{1+\beta A_V}$$

$$\because R_o' = \frac{R_oR_L}{R_o+R_L}, \quad A_V = \frac{A_vR_L}{R_o+R_L}$$

Voltage shunt feedback:

In this topology, the o/p resistance can be measured by ~~shorting i/p~~ opening input source $I_s = 0$ with & looking into o/p terminall with R_L disconnected as shown in fig:



Apply KVL to output side.

$$R_m I_i + I R_o - V = 0$$

$$I = \frac{V - R_m I_i}{R_o} \quad \text{--- (1)}$$

The input current $I_i = -I_f = -\beta V$ --- (2)

(2) in (1)

$$I = \frac{V - R_m(-\beta V)}{R_o} = \frac{V(1+\beta R_m)}{R_o} \quad \text{--- (3)}$$

$$\cancel{R_{of}} = \cancel{\frac{V}{I}} = \frac{1+\beta R_m}{R_o}$$

$$R_{of} = \frac{V}{I} = \frac{R_o}{1+\beta R_m}$$

Note: R_m is openloop transresistance without taking R_L in account

$$R_{of}' = R_{of} \parallel R_L$$

$$= \frac{\frac{R_o}{1+\beta R_m} * R_L}{\frac{R_o}{1+\beta R_m} + R_L} = \frac{R_o R_L}{R_o + R_L + \beta R_m R_L}$$

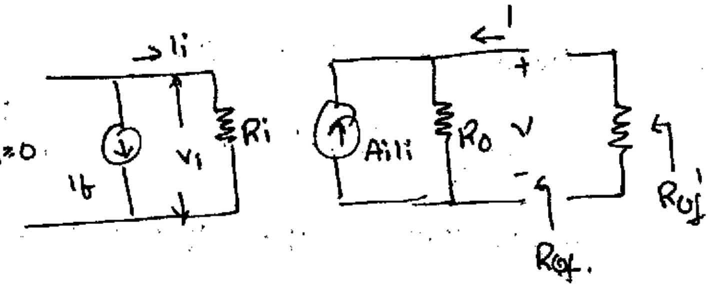
Dividing both Numerator & denominator with $(R_o + R_L)$

$$R_{of} = \frac{\frac{R_o R_L}{R_o + R_L}}{1 + \beta \frac{R_m R_L}{R_o + R_L}} = \frac{R_o'}{1+\beta R_M}$$

$\because R_o' = \frac{R_o R_L}{R_o + R_L}$ $\quad R_M = \frac{R_m R_L}{R_o + R_L}$.

Current shunt feedback:

In this topology, the o/p resistance can be measured by open circuiting input source $I_s=0$ & looking into o/p terminals with R_L disconnected as shown in fig.



Applying KCL to output, we get.

$$\frac{V}{R_o} - A_i I_i = I \quad \text{—(1)}$$

The input current is given as

$$I_i = -I_f = -\beta I_o$$

$$I_i = +\beta I \qquad \because I = -I_o \quad \text{—(2)}$$

(2) in (1)

$$\frac{V}{R_o} = I + A_i \beta I \Rightarrow \frac{V}{I} = R_o(1+\beta A_i) \quad \text{—(3)}$$

Note A_i: open loop current gain without R_L in account

$$R_{of}' = R_{of} \| R_L$$

$$= \frac{R_o(1+A_i\beta)R_L}{R_o(1+A_i\beta)+R_L} = \frac{R_o R_L(1+\beta A_i)}{R_o+R_L+\beta A_i R_o}$$

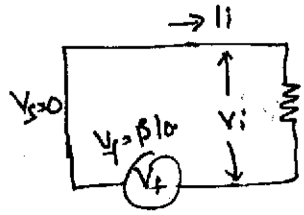
Dividing both numerator & denominator by (R_o+R_L)

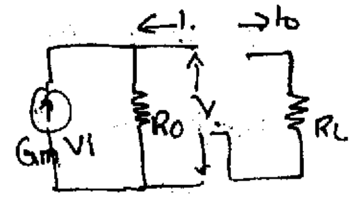
$$R_{of}' = \frac{\frac{R_o R_L}{R_o+R_L}(1+\beta A_i)}{1+\beta \frac{A_i R_o}{R_o+R_L}} = \frac{R_o'(1+\beta A_i)}{1+\beta A_I} \ //$$

$$\therefore R_o' = \frac{R_o R_L}{R_o+R_L}, \quad A_I = \frac{A_i R_o}{R_o+R_L} \ //$$

Current series feedback:

In this topology, the o/p resistance is measured by shorting input source $V_s=0$ & looking into output terminals with R_L disconnected as shown in fig.





Applying KCL to o/p node:

$$I = \frac{V}{R_o} - G_m V_i \quad \text{---} (1)$$

input voltage $V_i = -V_f = -\beta I_o = \beta I$ --- (2)

$$\frac{V}{R_o} = I + G_m(\beta I)$$

$$\frac{V}{R_o} = I(1+\beta G_m) \Rightarrow R_{of} = \frac{V}{I} = R_o(1+\beta G_m) \quad \text{---} (3)$$

Note: G_m is trans conductance of open loop ckt without R_L in account

$$R'_{of} = R_{of} \parallel R_L$$

$$= \frac{R_o(1+\beta G_m)R_L}{R_o(1+\beta G_m)+R_L} = \frac{R_o R_L(1+\beta G_m)}{R_o+R_L+\beta G_m R_L}$$

Dividing both numerator & Denominator by (R_o+R_L)

$$R'_{of} = \frac{\frac{R_o R_L}{R_o+R_L}(1+\beta G_m)}{\cancel{R_o}+ 1+\beta\frac{G_m R_L}{R_o+R_L}} = \frac{R'_o(1+\beta G_M)}{1+\beta G_M}$$

$$\because R'_o = \frac{R_o R_L}{R_o+R_L} \qquad G_M = \frac{G_m R_L}{R_o+R_L}$$

Effects of Negative feedback on Amplifier Characteristics:

Parameter	voltage series	current series	current shunt	voltage shunt
Gain with feedback.	$A_{vf} = \frac{A_v}{1+\beta A_v}$ decreases	$G_{mf} = \frac{G_m}{1+\beta G_m}$ decreases	$A_{if} = \frac{A_i}{1+\beta A_i}$ decreases	$R_{mf} = \frac{R_m}{1+\beta R_m}$ decreases
Stability	improves	improves	improves	improves
Frequency Response	improves	improves	improves	improves
Frequency Distortion	Reduces	Reduces	Reduces	Reduces
Noise & Non linear distortion	Reduces	Reduces	Reduces	Reduces
input Resistance	$R_{if} = R_i(1+\beta A_v)$ increases	$R_{if} = R_i(1+\beta G_m)$ increases	$R_{if} = \frac{R_i}{1+\beta A_i}$ decreases	$R_{if} = \frac{R_i}{1+\beta R_m}$ decreases
output resistance.	$R_{of} = \frac{R_o}{1+\beta A_v}$ decreases	$R_{of} = R_o(1+\beta G_m)$ increases	$R_{of} = R_o(1+\beta A_i)$ increases	$R_{of} = \frac{R_o}{1+\beta R_m}$ decreases

ADVANTAGES & DISADVANTAGES OF NEGATIVE FEEDBACK.

The negative feedback of amplifier circuit reduces gain of amplifier. This is the main disadvantage using -ve FB in amplifier.

inspite of this, -ve FB is used in almost every amp, due to its no. of advantages.

* Negative feedback stabilizes the gain of amplifier
* -ve FB increases BW of amplifier, so that it provides constant gain for large freq range of i/p signal.
* -ve FB reduces the distortion in amplifier output.
* Series -ve FB increases input resistance of amp.
* Shunt voltage -ve FB decreases output resistance of amp
* -ve FB also stabilizes operating point.

method of analysis of a feedback amplifier:

To calculate the important characteristics of feedback system (A_f, R_{if} & R_{of}), it is desirable to seperate feedback amplifier into two blocks, basic amplifier & feedback network β.

The basic amplifier configuration without feedback but taking loading effect of β network into account is to be obtained.

The complete analysis of feedback (system) amplifier is obtained by carrying out the following steps.

1) Identify the topology.

a) To find the type of topology sampling network.

(i) By shorting o/p i.e., $V_o=0$, if feedback signal X_f becomes zero, then it is voltage sampling.

(ii) By opening the o/p loop i.e., $I_o=0$, if feedback signal X_f becomes zero, then it is current sampling.

b) To find the type of mixing network.

(i) If feedback signal is given back to externally applied signal as voltage in i/p loop, then it is series mixing.

(ii) If feedback signal is given back to externally applied signal as current, in i/p loop, then it is shunt mixing.

Thus type of feedback amplifier is determined.

(2) Draw basic amplifier without feedback.

a) To find input circuit

(i) set $V_o=0$ for voltage sampling (short circuit o/p)

(ii) set $I_o=0$ for current sampling. (open circuit o/p)

b) To find output circuit

(i) set $V_i=0$ for shunt mixing (short circuit i/p)

(ii) set $I_i=0$ for series mixing (open circuit i/p)

step② ensures that the feedback is reduced to zero without altering the loading of basic amplifier.

(3) use thevinin's source if x_f is voltage & Norton's source if x_f is current.

(4) Replace each active element by proper model (Hybrid π for High freq & h-parameter for low frequency)

(5) Indicate x_f & x_o on circuit obtained, by carrying steps ② ③ & ④. Evaluate $\beta = \frac{x_f}{x_o}$

(6) Evaluate A by applying KVL & KCL to equivalent circuit obtained.

(7) From A & β, Find D, A_f, R_{if}, R_{of} & R_{of}'

The following table summarizes above procedure.

Characteristics	Voltage series	current series	current shunt	voltage shunt.
feedback signal x_f	voltage	voltage	current	current.
sampled signal x_o	voltage	current	current	voltage.
To find i/p loop, set	$V_o = 0$	$I_o = 0$	$I_o = 0$	$V_o = 0$
To find o/p loop set,	$I_i = 0$	$I_i = 0$	$V_i = 0$	$V_i = 0$
signal source	Thevinin	Thevinin	Norton	Norton
$\beta = \frac{X_f}{X_o}$	$\frac{V_f}{V_o}$	$\frac{V_f}{I_o}$	$\frac{I_f}{I_o}$	$\frac{I_f}{V_o}$
$A = X_o/X_i$	$A_{V_1} = \frac{V_o}{V_i}$	$G_{m1} = \frac{I_o}{V_i}$	$A_{i1} = \frac{I_o}{I_i}$	$R_{m1} = \frac{V_o}{I_i}$
$D = 1+A\beta$	$1+\beta A_{V1}$	$1+\beta G_{m1}$	$1+\beta A_{i1}$	$1+\beta R_{m1}$
A_f	$\frac{A_V}{D}$	$\frac{G_m}{D}$	$\frac{A_i}{D}$	$\frac{R_m}{D}$
R_{if}	$R_i D$	$R_i D$	R_i/D	R_i/D
R_{of}	$\frac{R_o}{1+\beta A_V}$	$R_o(1+\beta G_m)$	$R_o(1+\beta A_i)$	$\frac{R_o}{1+\beta R_m}$
$R_{of}' = R_{of} \parallel R_L$	$\frac{R_o'}{D}$	$\frac{R_o'(1+\beta G_m)}{D}$	$\frac{R_o'(1+\beta A_i)}{D}$	$\frac{R_o'}{D}$

* This procedure gives basic amplifier circuit without feedback but taking loading of β, R_L & R_s into account.

Voltage series feedback:

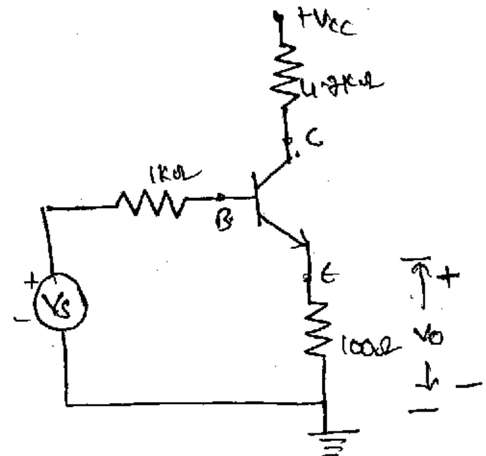


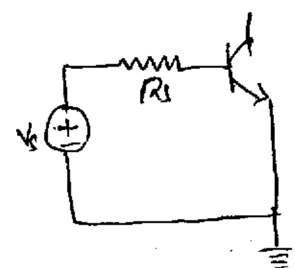
Fig. shows emitter follower circuit. Here feedback voltage is voltage across R_o & sampled signal is V_o across R_e.

Step1: Identify topology.

By shorting output voltage ($V_o = 0$), feedback signal becomes zero & hence it is voltage sampling. In circuit, feedback signal V_f is subtracted from externally applied signal V_s & hence it is series mixing.

Thus, given amplifier is voltage series feedback amplifier.

Step2: Find i/p & o/p circuits



To find i/p circuit, set $V_o = 0$ & hence V_s is in series with R_s appears b/w base & emitter.

To find o/p circuit set $I_i = I_b = 0$ & hence R_e appears only in o/p loop. With these we obtain the circuit as shown below.

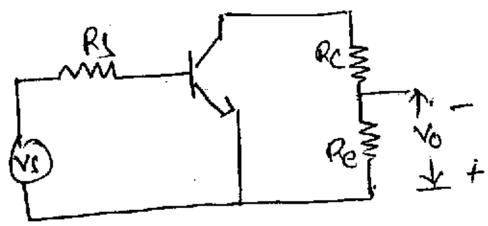


Fig: Amp without FB

Step 3&4: Replace transistor by its h-parameter model.

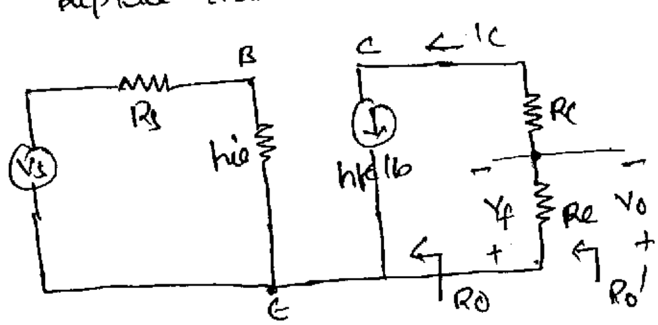


Fig: Transistor replaced by its appropriate model at low freq.

Here $V_o = V_f$.

Then $\beta = \frac{V_f}{V_o} = 1$ //

The voltage gain A_V is given by:

$$A_V = \frac{V_o}{V_i} = \frac{V_o}{V_s}$$

$$V_o = h_{fe} I_b R_e$$

$$V_s = I_b (R_s + h_{ie})$$

$$\therefore A_V = \frac{V_o}{V_s} = \frac{h_{fe} I_b R_e}{I_b (R_s + h_{ie})} = \frac{h_{fe} R_e}{R_s + h_{ie}} \quad 3.38$$

$$D = 1 + \beta A_V = 1 + \left(\frac{h_{fe} R_e}{R_s + h_{ie}}\right)\beta = \quad 3.38$$

$$A_{Vf} = \frac{A_{V_1}}{D} = \frac{h_{fe} R_e}{R_s + h_{ie} + h_{fe} \beta R_e} \quad 3.38$$

For $R_s + h_{ie} \ll h_{fe} R_E$ $A_{Vf} \cong 1$, For CC $A_V = 1$

The input resistance without feedback is $R_i = R_s + h_{ie} = 2.1\,k\Omega$.

The input resistance with feedback is

$$R_{if} = R_i D = (R_s + h_{ie}) \frac{R_s + h_{ie} + \beta h_{fe} R_e}{(R_s + h_{ie})} = 7.098\,k\Omega$$

The output resistance without feedback $R_o = \infty$

The output resistance with feedback is

$$R_{of} = \frac{R_o}{1 + \beta A_V} = \infty$$

The effective output resistance with feedback is

$$R_{of}' = R_o \| R_e = \infty \| R_e = R_e.$$

$$R_{of}' = \frac{R_e (R_s + h_{ie})}{R_s + h_{ie} + \beta h_{fe} R_e} = 29.58\,\Omega \;//$$

CURRENT SERIES

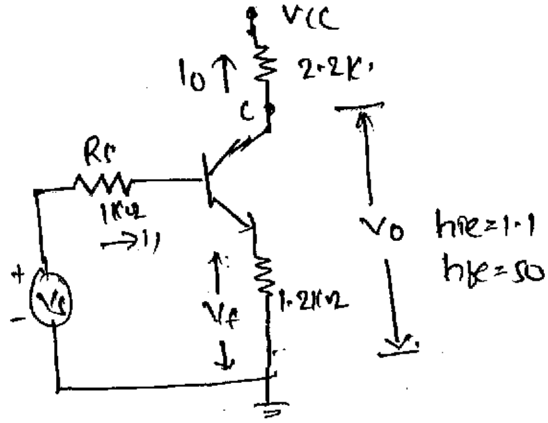


Fig shows common emitter with unbypassed Re.
In this config, Re is common to Base & emitter i/p circuit as well as collector to emitter o/p circuit. i.e., the i/p current I_b as well as o/p current I_c both flow through it.

The voltage drop across Re is $V_f = (I_b + I_c) R_e$

$= I_c R_e = -I_o R_e$.

This voltage shows that o/p current I_o being sampled & is converted to voltage by feedback network. At input side voltage V_f is subtracted from V_s to produce V_i. Therefore the feedback applied in series.

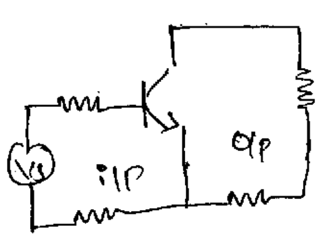
Step 1: By opening output loop ($I_o = 0$), feedback signal becomes zero. & hence it is current sampling.

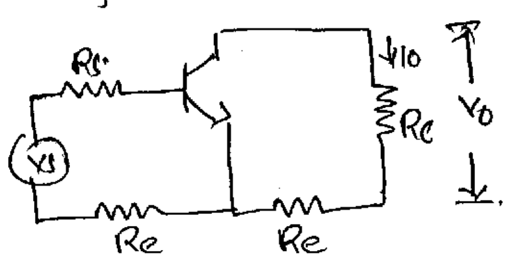
From fig, the feedback signal V_f is subtracted from externally applied signal V_s & hence it is series mixing.

Thus the circuit is current series feedback amplifier.

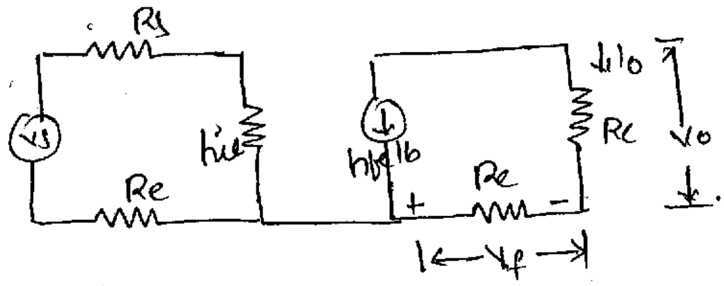
Step ②: To find input circuit, set $I_o = 0$, then Re appears at input side.

To find output circuit, set $I_i = 0$, then Re appears across output circuit. The resulting circuit is shown below





Step 4: Replace transistor with its approximate h-parameter model as shown.



The openloop gain. $G_M = \frac{I_o}{V_i} = \frac{-h_{fe} I_b}{V_b(R_s + h_{ie} + R_e)} = \frac{-h_{fe}}{R_s + h_{ie} + R_e}$.

$= -0.015$

With I_o & V_f, we get $\beta = \frac{V_f}{I_o} = \frac{I_e R_e}{I_o}$ $(\because I_e = -I_o)$

$= -R_e = -1200$

Desensitivity

$D = 1 + \beta G_M = 1 + (-1200)(-0.015) = 19.18$

$G_{Mf} = \frac{G_M}{D} = \frac{-0.015}{19.18} = -0.782 \times 10^{-3}$

$A_{vf} = \frac{V_o}{V_s} = \frac{I_o R_L}{V_s} = G_{Mf} R_L = -1.72$.

From fig:

$R_i = R_s + h_{ie} + R_e = 3.3 K\Omega$

$R_{if} = R_i D = 63.294K$.

The output resistance $R_o = \infty$.

$R_{of} = R_o(1 + \beta G_m) \Rightarrow R_{of} = R_o D = \infty$ $\qquad R_o' = R_o \| R_L$.

$R_{of}' = R_{of} \| R_L = 2.2K\Omega$.

current shunt feedback:

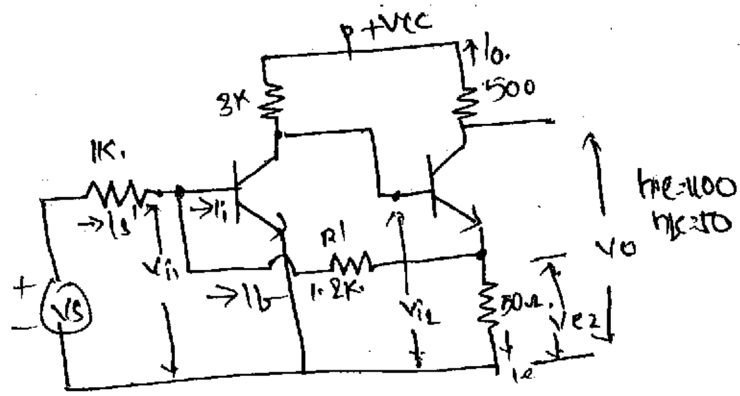


Fig: Shows the two transistors in cascode connection with feedback from second emitter to first base through resistor R'.

Here the feedback network is formed by R' and Re2 drives current ie. Since Ie = −Io, the feedback network gives current as feedback.

At input side $i_i = I_S - I_f$. ie If is subtracted from Is to get Ii, so it is shunt mixing. Therefore the config is current shunt feedback.

Step1: Identify topology

By shorting o/p voltage ($V_o=0$), feedback signal does not become zero & hence it is not voltage sampling.

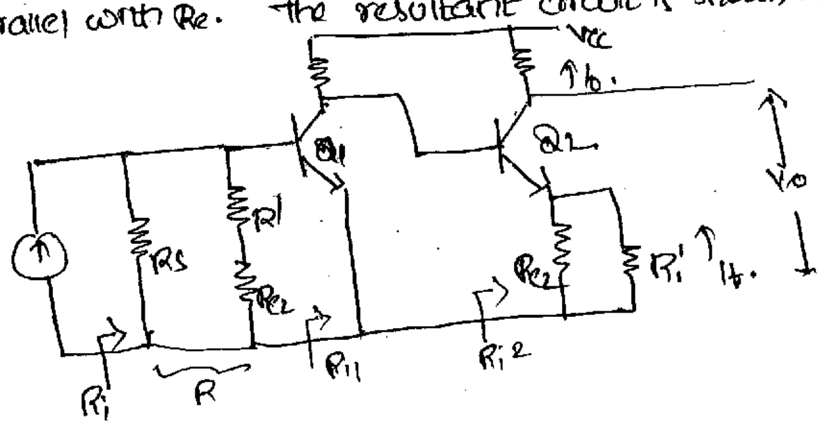
By opening o/p loop ($I_o=0$), feedback becomes zero & hence it is current sampling.

The feedback signal appears in shunt with input, hence, the topology is current shunt feedback amplifier.

Step 2: Find i/p & o/p circuit

The input circuit of amplifier without feedback is obtained by opening the o/p loop at emitter Q2 ($I_o=0$). This places R' in series with Re from base to emitter of Q1.

The o/p circuit is found by shorting i/p node ies $V_i=0$. this places R' in parallel with Re. The resultant circuit is shown below:



The open circuit Transfer Gain is given by.

$$A_I = \frac{-I_{C2}}{I_s} = \frac{-I_{C2}}{I_{b2}} \times \frac{I_{b2}}{I_{C1}} \times \frac{I_{C1}}{I_{b1}} \times \frac{I_{b1}}{I_s}$$

we know that $\frac{-I_{C2}}{I_{b2}} = A_{i2} = -h_{fe} = -50$

$$\frac{-I_{C1}}{I_{b1}} = A_{i1} = -h_{fe} = -50 \Rightarrow \frac{I_{C1}}{I_{b1}} = 50$$

From fig: $\frac{I_{b2}}{I_{C1}} = \frac{-R_{C1}}{R_{C1}+R_{i2}} = -0.457$ //

$R_{i2} = h_{ie} + (1+h_{fe})(R_{e2} \| R_C) \longrightarrow$ Since 2nd stage is having R_{e2}, R_C as common for both i/p & o/p. we need to calculate R_{i2} as $h_{ie} + A_i R_L$.

$A_i = \frac{-I_{e2}}{I_{b2}} = (1+h_{fe})$

$R_C' = R_{e2} \| R_C)$

$$\frac{I_{b2}}{I_{C1}} = \frac{-3K}{3K+3.55K} = -0.457 //$$

$\frac{I_{b1}}{I_s} = \frac{R}{R+h_{ie}}$ where $R = R_s \| (R' + R_e) = \frac{1.2K \times 1.25K}{1.2K + 1.25K} = 0.612K$ //

.55

$$\frac{I_{b1}}{I_s} = \frac{0.612K}{0.612K + 1.1K} = 0.358 //$$

$\therefore$ $A_I = (-50)(-0.457)(50)(0.358) = 406$ //

Calculate β

$\beta = \frac{I_f}{I_o}$ $\quad I_f = \frac{-I_e R_{e2}}{R_{e2}+R'} = \frac{-I_{C2} R_{e2}}{R_{e2}+R'}$ $(\because I_e \cong I_C)$

$$I_f = \frac{I_o R_{e2}}{R_{e2}+R'} \quad (\because I_o = -I_{C2})$$

$$\therefore \beta = \frac{I_f}{I_o} = \frac{R_{e2}}{R_{e2}+R'} = 0.04 //.$$

calculate D, R_i, R_{if}, R_{of}, A_{if}, A_{vf}, R_o.

$$D = 1 + \beta A_i = 1 + (0.04)(406) = 17.2 //$$

$$A_{if} = \frac{A_I}{D} = \frac{406}{17.2} = 23.6 //$$

$$A_{vf} = \frac{V_o}{V_s} = \frac{-I_{C2} R_{C2}}{I_s R_s} = A_{if} \frac{R_{C2}}{R_s} = 9.83 //$$

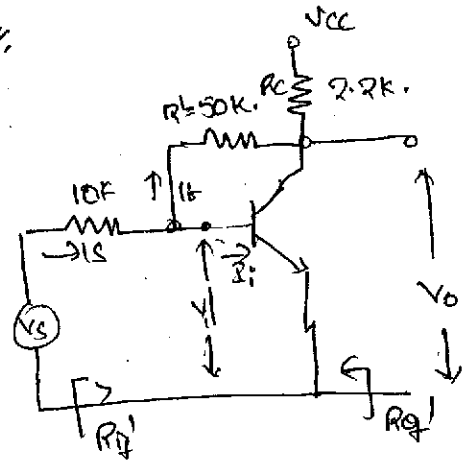
$R_i = R \| h_{ie} = \dfrac{0.612K \times 1.1K}{1.612K} = 0.394K.$

$R_{if} = \dfrac{R_i}{D} = \dfrac{0.394K}{17.2} = 23\Omega.$

$R_o = \infty$; $R_{of} = R_o D = \infty$

$R_o' = R_o \| R_{C2} = R_{C2} = 500\Omega.$

$R_{of}' = R_o' \dfrac{(1+\beta A_i)}{D} = R_o' = R_{C2} = 500\Omega$ //.



Voltage shunt feedback.

Fig shows CE amplifier with resistor R' from o/p to input.

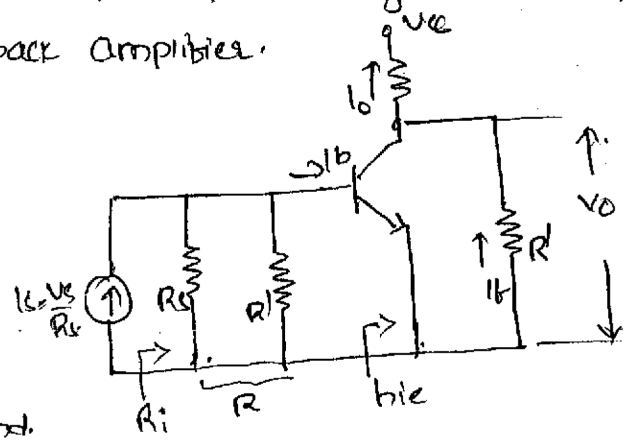
Step 1 Identify topology.

The feedback current $I_f = \dfrac{V_i - V_o}{R'}$ but $V_o > \beta V_i$

$$I_f = \frac{-V_o}{R'}$$

By shorting output voltage ($V_o=0$), feedback reduces to zero, hence it is voltage sampling. As $I_i = I_s - I_f$ mixing is shunt type & topology is voltage shunt feedback amplifier.

Step 2. Find i/p & o/p circuits.



To find i/p circuit, set $V_o = 0$, this places R' b/w base & ground.

To find o/p circuit, set $V_i = 0$, this places R' b/w collector & ground.

The feedback signal is current I_f in resistor R' which is in output circuit as shown in fig.

we seen that

$$I_f = \frac{V_i - V_o}{R'} = \frac{-V_o}{R'}$$

$$\frac{I_f}{V_o} = \beta = \frac{-1}{R'}$$

Find open circuit transresistance.

$$R_m = \frac{V_o}{I_s} = \frac{I_o R_c'}{I_s} = \frac{-I_c R_c'}{I_s}$$

$$R_c' = R_c \| R' = 2.2K \| 50K = 2.1K.$$

$$\& \quad \frac{-I_c}{I_s} = \frac{-I_c}{I_b}\,\frac{I_b}{I_s} = A_i \frac{I_b}{I_s}$$

$$\frac{I_b}{I_s} = \frac{R}{R + h_{ie}} \Rightarrow R = R_s \| R' = 8.33K.$$

$$\frac{I_b}{I_s} = \frac{8.33K}{8.33K + 1.1K} = 0.883$$

$$A_i = -h_{fe} = -50$$

$$\therefore \quad R_m = \frac{-I_c R_c'}{I_s} = (-50)(0.883) \times 2.1K =$$
$$= -92.715K.$$

Calculate β

$$\beta = \frac{-1}{R'} = \frac{-1}{50K} = -2 \times 10^{-5}$$

Calculate D

$$D = 1 + \beta R_m = 2.854.$$

$$R_{mf} = \frac{R_m}{D} = -32.48K$$

$$A_{vf} = \frac{V_o}{V_s} = \frac{V_o}{I_s R_s} = \frac{R_{mf}}{R_s} = -3.248$$

$$R_i = R \| h_{ie} = \frac{R h_{ie}}{R + h_{ie}} = 0.971K.$$

$$R_{if} = \frac{R_i}{D} = 340.22\Omega.$$

$R_o = \infty$

$R_{of} = \infty$

$$R_o' = R_o \| R_c' = R_c' = 2.1K.$$

$$R_{of}' = \frac{R_o'}{D} = 735.8\Omega$$

UNIT V OSCILLATORS

The device which works on the principle of positive feedback is called Oscillator.

An Oscillator is a circuit which basically acts as a generator. It generates o/p signal with constant amplitude and constant desired frequency.

An oscillator does not require any input signal. In short, an oscillator is an amplifier, which uses a positive feedback and without any external input signal, generates an o/p waveform at desired frequency.

concept of positive feedback:

The feedback is said to be positive whenever the part of output that is fed back to the amplifier as its input is in-phase with the original input signal applied to the amplifier.

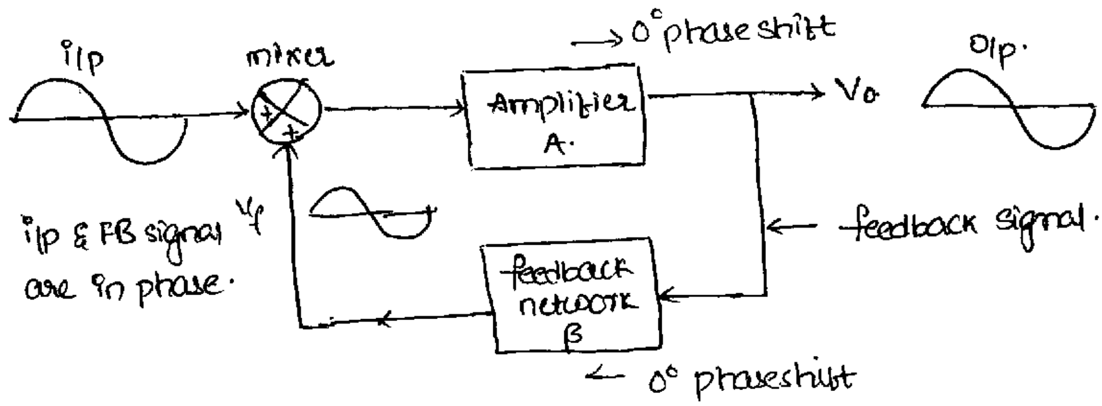


Fig: concept of positive feedback

As the phase of feedback signal is same as that of the input applied, the feedback is called positive feedback.

consider a non inverting amplifier with voltage gain A. Vs is applied to ckt. As amplifier o/p Vo is in phase with input Vs. some part of o/p is feed back to i/p with help of feedback network. Feed back voltage V_f is inphase with Vs.

Amplifier gain $A = \frac{V_o}{V_i}$ = open loop gain of amplifier.

$A_f = \frac{V_o}{V_s}$ = closed loop gain (or) Gain with feedback.

$V_i = V_s + V_f$ —①

V_f is depending on feedback element gain β.

So. $\beta = \frac{V_f}{V_o}$

$V_f = \beta V_o$ —②

$V_i = V_s + \beta V_o$

$V_s = V_i - \beta V_o$ —③

$$A_f = \frac{V_o}{V_s} = \frac{V_o}{V_i - \beta V_o}$$

$$= \frac{V_o/V_i}{1 - \beta \frac{V_o}{V_i}}$$

$$\boxed{A_f = \frac{A}{1 - A\beta}} \quad —④$$

Consider various values of β & corresponding values of A_f for constant amplifier gain of A = 20

A	β	A_f
20	0.005	22.22
20	0.04	100
20	0.045	200
20	0.05	∞

The above result shows that the gain with feedback increase as the amount of positive feedback increases.

In limiting case, the gain becomes infinite. It means that the circuit can provide output without external input ($V_s = 0$) just by feeding the part of output as its own input.

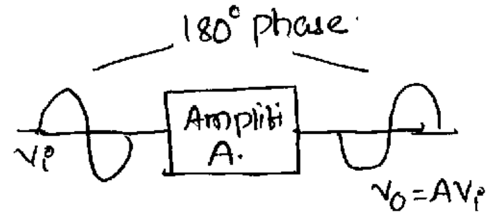
Similarly, output cannot be infinite but gets driven into oscillations. In otherwords, the circuit stops amplifying & starts oscillating.

" Thus without an input, the output will continue to oscillate whose frequency depends upon feedback network or the amplifier or the both. Such a circuit is called an oscillator".

Difference b/w positive and negative feedback.

positive feedback	Negative feedback.
1) The feedback signal is in phase with applied input signal	1) The feedback signal is out of phase with applied input signal
2) It increases gain of amplifier	2) It reduces gain of amplifier
3) It is regenerative or Direct FB	3) It is degenerative (or) Inverse FB.
4) It makes amplifier unstable	4) It makes amplifier stable
5) It reduces Bandwidth	5) It increases Bandwidth
6) It is used in Oscillators	6) It is used in small signal amps.

Barkhausen Criterion



Consider a basic inverting amplifier with an open loop gain A. The feedback network attenuation factor β is less than unity.

As basic amplifier is inverting, it produces phase shift of 180° between input & output as shown in fig.

Now the input V_i applied to amplifier is to be derived from its output V_o & feedback network.

But the feedback must be positive ie., the o/p voltage using feedback must be in phase with V_i. Thus feedback network introduces a phase shift of 180° while feeding back voltage from output to input. This ensures positive feedback.

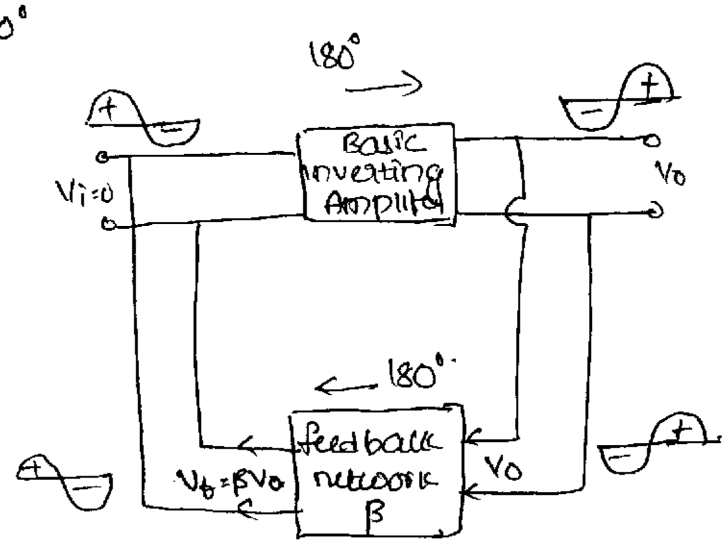


Fig: Block diagram of Oscillator ckt

consider the voltage V_i, applied at input of amplifier

$$V_o = AV_i \quad —①$$

The feedback factor β decides the feedback to be given to input

$$V_f = -\beta V_o \quad —②$$ –ve sign indicates 180° phase shift.

$$\therefore \quad V_f = -\beta A V_i \quad —③$$

For oscillator, we want that feedback should drive amplifier & hence V_f must act as V_i.

Now if V_f has to be equal to V_i, then. It will be satisfied only when.

$$\boxed{-A\beta = 1} \quad —④$$

$-A\beta = 1$ condition is called "Barkhausen Criterion".

$$\therefore \quad -A\beta = 1 + j0$$

$$\boxed{|A\beta| = 1} \quad —⑤$$

The phase of V_f must be same as V_i i.e., feedback network should introduce 180° phase shift in addition to 180° phase shift introduced by a inverting amplifier. so total phase shift around loop is 360°.

Barkhausen criterion states that:

Total phase shift around loop, as signal proceeds from input through amplifier, FB n/w back to input again, completing a loop, is precisely 0° or 360°. or an integral multiple of 2π.

The magnitude of product of open loop gain of amplifier A & feedback factor β is unity i.e., $|A\beta| = 1$

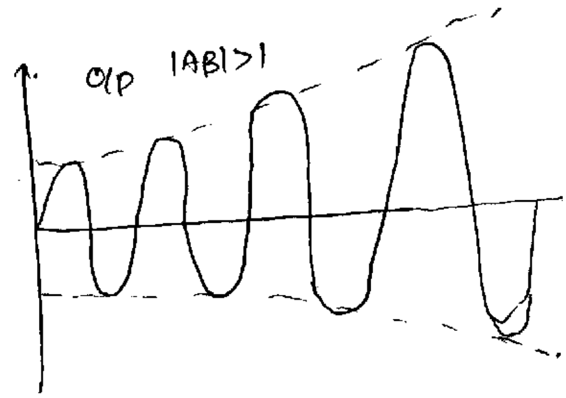
If the above two conditions are satisfied, then the circuit works as an oscillator producing sustained oscillations of constant frequency & amplitude.

Stability of oscillator:

The oscillations produced at the o/p of circuit with constant frequency and amplitude mainly depends on parameter $A\beta$. Based on the value of $A\beta$ product, the oscillations are produced in oscillator circuit at desired frequency and amplitude.

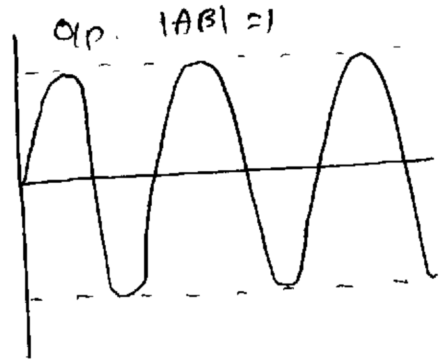
(1) $|A\beta| > 1$

When total phase shift around loop is 0° or 360° and $|A\beta| > 1$, then the output oscillates but the oscillations are of growing type i.e., the amplitude goes on increasing as shown in fig.



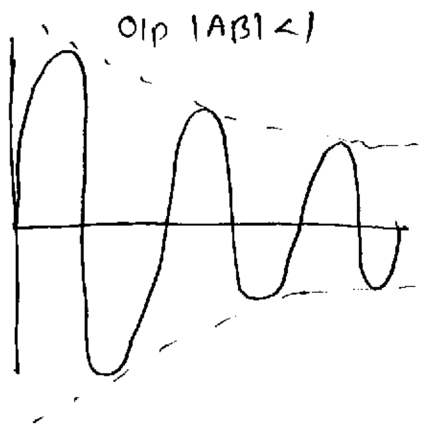
(2) If $|A\beta| = 1$

When the total phase shift around loop is 0° or 360° and $|A\beta|=1$, then the oscillations are with constant frequency & constant amplitude called sustained oscillations as shown in fig.



(3) $|A\beta| < 1$

When the total phase shift around loop is 0° or 360° & $|A\beta|<1$; then the oscillations are of decaying type which have the amplitude decreasing exponentially. as shown in fig



Classification of Oscillators

The oscillators are classified

i based on nature of output waveform

ii based on parameters used

iii based on the range of frequency.

(i) Based on the O/p wave, the Oscillators are classified as sinusoidal and non sinusoidal oscillators.

Sinusoidal oscillators generate purely sinusoidal wave at O/p whereas non sinusoidal " generate O/p as triangular, square, saw. etc.

(ii) Based on circuit components

RC oscillator — contains Resistor & Capacitor as components
LC oscillator — contains inductor & capacitor as components
Crystal oscillator — which uses crystal.

(iii) Based on range of frequency:

Low frequency oscillators (or) Audio freq. oscillators — used to generate oscillations at audio frequency range (20Hz – 20KHz)

High frequency oscillators or Radio frequency oscillators — used to generate oscillations from 200–300KHz upto few GHz

RC Oscillators are used at low frequency range.
LC oscillators are used at high frequency range.

At low freq, the value of inductor required is large. The large inductor is larger in size & occupies a large space. It increases size and cost of circuit. Hence LC oscillators are not used at low freq.

Based on whether feedback is used or not:

The Oscillators in which feedback is used, which satisfies the required condition are called feedback type Osc.

The Oscillators which doesnot use any feedback to generate oscillations are called as non feedback type Oscillators.

Ex: UJT Relaxation Oscillator

Oscillators

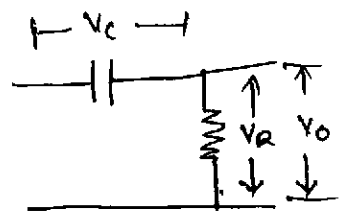
- Sinusoidal
  - low freq.
    - RC-phaseshift
    - wein bridge
  - High freq
    - LC oscillators
      - Hartley
      - colpitt
      - clapp
      - crystal.
- non-sinusoidal
  - UJT relaxation
  - multivibrators
    - Astable
    - monostable
    - Bistable.
  - Time Based Generators

Fig: classification of oscillators.

RC PHASESHIFT OSCILLATOR:

RC phase shift oscillator basically consists of an amplifier and a feedback network consisting of resistors and capacitors arranged in ladder fashion. Hence such an oscillator is also called ladder type RC phase shift oscillator.

The following figure shows basic RC circuit



a) RC-circuit.

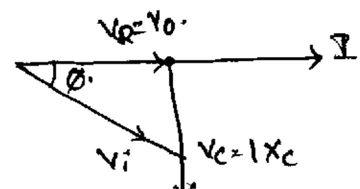
The capacitor C & resistor R are in series

The reactance $X_C = \frac{1}{j\omega C}\ \Omega = \frac{1}{j2\pi f C}$.

The total impedance of circuit is

$$Z = R - jX_C = R - j\frac{1}{2\pi f C}\ \Omega.$$

$$= |Z| \angle -\phi\ \Omega.$$



b) phasor diagram

The rms value of input voltage be V_i volts.

then the current is given by.

$$I = \frac{V_i \angle 0^\circ}{Z \angle -\phi}$$

$$I = \frac{V_i}{Z} \angle +\phi \ A. \quad \text{where} \quad |Z| = \sqrt{R^2 + X_C^2}$$

$$\theta = \tan^{-1}\left(\frac{X_C}{R}\right).$$

From expression of current I, it can be seen that current 'I' leads input voltage by angle ϕ."

The output voltage V_o is the drop across resistance R is given by

$$V_o = V_R = IR.$$

The voltage across capacitor $V_C = IX_C$.

The drop V_R is in phase with current I while V_C lags current I by 90° i.e., I leads V_C by 90° as shown in fig.

By using proper values of R and C, the angle of ϕ is adjusted to get 60°.

In RC phase shift oscillator feedback network must introduce a phase shift of 180° to obtain total phase shift around loop as 360°. Thus to produce phase shift of 180°, such three RC networks must be connected in phase cascade as shown in fig:

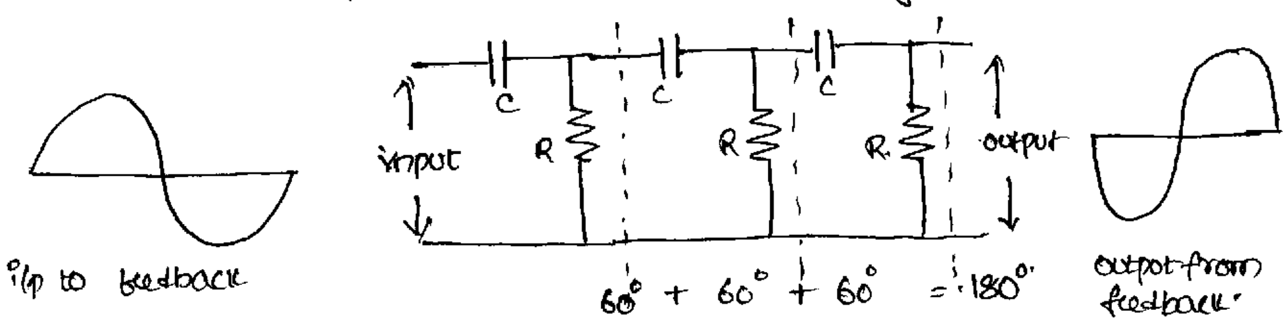


Fig: feedback network in RC phase shift Osc.

The network is also called ladder network. All the resistance and capacitance values are same so that for particular frequency, each section of R & C network produces a phase shift of 60°.

Transistorized RC phase shift oscillator

Figure shows a practical transistorised RC-phase shift Oscillator which uses a single stage CE amplifier & a phase shifting network consisting of three identical RC sections.

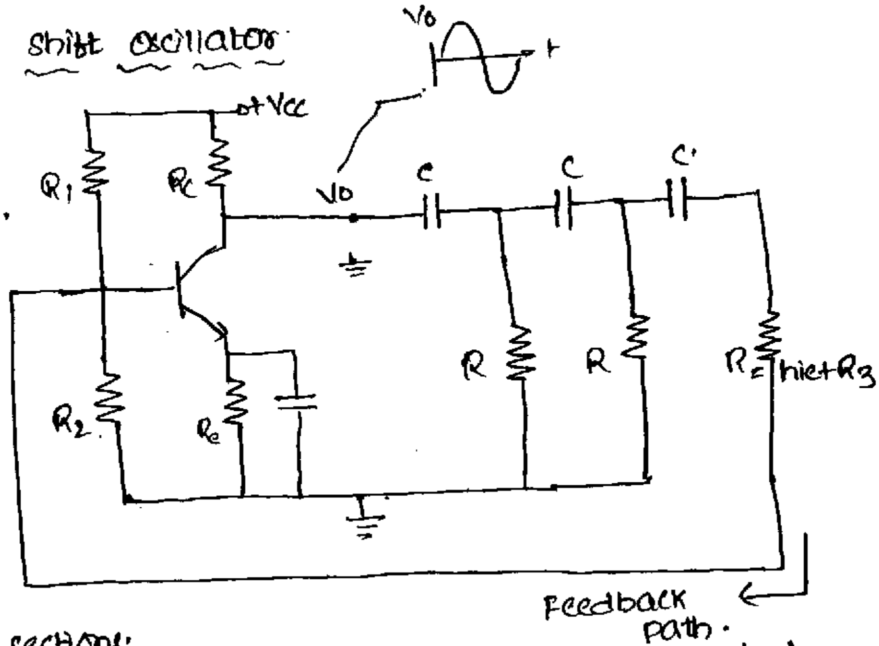


Fig: Transistorised RC phase shift oscillator.

The output of feedback network gets loaded due to low input impedance of transistor. Hence an emitter follower input stage before the CE amplifier stage can be used to avoid the problem of (low i/p impe) But if only single stage is to be used then voltage shunt feedback is used.

Neglecting the values of R_1 and R_2 as these are sufficiently large, we get.

h_{ie} = input impedance of amplifier stage.

Now the resistance R_3 and h_{ie} are in series and the value of R_3 is so selected such that the resultant of two resistances is R. which is required value of resistance in last section of RC network.

$$R = R_3 + h_{ie}. \quad \text{--- (1)}$$

This ensures that, all the three sections of phase shifting n/ws are identical.

NOTE: If the resistance R_1 & R_2 are not neglected then the input impedance of amplifier stage becomes

$$R_i' = R_1 \| R_2 \| h_{ie}. \quad \text{--- (2)}$$

In such a case, the value of R_3 must be selected such that.

$$R = R_i' + R_3. \quad \text{--- (3)}$$

Derivation of frequency of oscillations:

Replacing the transistor by its approximate h-parameter model, we get the equivalent oscillator circuit as shown in fig:

Now, we can replace $h_{ie}+R_S$ as R.. Similarly replace current source $h_{fe}I_b$ by its equivalent voltage source.

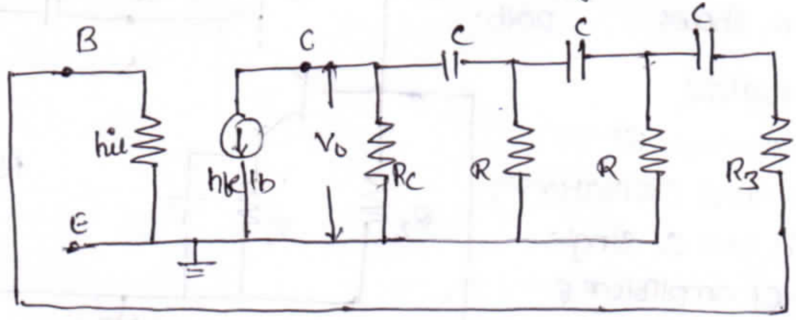
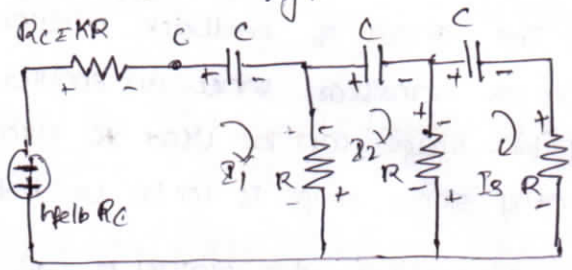


Fig: Equivalent circuit using h-parameter model.

Assume the ratio of R_C & R as $K = \frac{R_C}{R}$. — (4)

The modified equivalent circuit is shown in fig:

Applying KVL for various loops in the modified equivalent circuit, we get. →



For loop 1

$$-I_1 R_C - I_1 X_C - (I_1 - I_2) R - h_{fe} I_b R_C = 0$$

$$-I_1 R_C - I_1 \frac{1}{j\omega C} - (I_1 - I_2) R - h_{fe} I_b R_C = 0$$

Replacing R_C by KR and '$j\omega$' by 's' we get

$$I_1 \left((K+1) R + \frac{1}{sC} \right) - I_2 R = -h_{fe} I_b K R \quad \text{— (5)}$$

For loop 2.

$$-I_2 \frac{1}{j\omega C} - (I_2 - I_1) R - (I_2 - I_3) R = 0$$

$$-I_1 R + I_2 \left[2R + \frac{1}{sC} \right] - I_3 R = 0 \quad \text{— (6)}$$

For loop 3

$$-I_3 \frac{1}{j\omega C} - I_3 R - (I_3 - I_2) R = 0$$

$$-I_2 R + I_3 \left(2R + \frac{1}{sC} \right) = 0 \quad \text{— (7)}$$

Representing equations ⑤, ⑥ & ⑦ in matrix form we get:

$$\begin{bmatrix} (K+1)R+\frac{1}{SC} & -R & 0 \\ -R & 2R+\frac{1}{SC} & -R \\ 0 & -R & 2R+\frac{1}{SC} \end{bmatrix}\begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} = \begin{bmatrix} -h_{fe} I_b KR \\ 0 \\ 0 \end{bmatrix}$$

Using Cramer's rule to solve $I_3 = \frac{D_3}{D}$.

$$D = \begin{vmatrix} (K+1)R+\frac{1}{SC} & -R & 0 \\ -R & 2R+\frac{1}{SC} & -R \\ 0 & -R & 2R+\frac{1}{SC} \end{vmatrix} = \left[(K+1)R+\frac{1}{SC}\right]\left[\left(2R+\frac{1}{SC}\right)^2-R^2\right]-(-R)\left[-R\left(2R+\frac{1}{SC}\right)\right]$$

$$= \left[(K+1)R+\frac{1}{SC}\right]\left(2R+\frac{1}{SC}\right)^2 - R^2\left[(K+1)R+\frac{1}{SC}\right] - R^2\left(2R+\frac{1}{SC}\right)$$

$$= \frac{SRC(K+1)+1}{SC}\,\frac{(2SRC+1)^2}{S^2C^2} - R^2\frac{[SCR(K+1)+1]}{SC} - R^2\frac{[2SCR+1]}{SC}$$

$$D = \underbrace{\frac{[SRC(K+1)+1][4S^2R^2C^2+4SRC+1]}{S^3C^3}}_{\text{1st term}} - \underbrace{R^2\frac{[SCR(K+1)+2SCR+2]}{SC}}_{\text{2nd term}}$$

1st term = 4

$$\text{2nd term} = \frac{-(SCR^3K+SCR^3+2SCR^3+2R^2)}{SC} = \frac{-[2R^2+3SCR^3+SCR^3K]}{SC}$$

$$\therefore\quad D = \frac{S^3C^3R^3(3K+1)+S^2C^2R^2(4K+6)+SRC(5+K)+1}{S^3C^3} \quad \text{—— ⑧}$$

$$\text{Now } D_3 = \begin{vmatrix} (K+1)R+\frac{1}{SC} & -R & -h_{fe}I_bKR \\ -R & 2R+\frac{1}{SC} & 0 \\ 0 & -R & 0 \end{vmatrix} = -R^2[h_{fe}I_bKR] \quad \text{—— ⑨}$$

$$I_3 = \frac{D_3}{D} = \frac{-KR^3h_{fe}I_b S^3C^3}{S^3C^3R^3(3K+1)+S^2C^2R^2(4K+6)+SRC(5+K)+1} \quad \text{—— ⑩}$$

Where I_3 = output current of feedback circuit.

I_b = input current of amplifier.

$I_0 = -I_c = -h_{fe}I_b$ = input current of feedback circuit.

$$\beta = \frac{I_B}{I_C}, \quad A = \frac{-I_C}{I_b} \Rightarrow A\beta = \frac{I_B}{I_C} \times \frac{-I_C}{I_b} = \frac{-I_B}{I_b} \quad — (11)$$

$$\therefore A\beta = \frac{I_B}{I_b} = \frac{KR^3s^3C^3 h_{fe}}{s^3R^3C^3(3K+1) + s^2C^2R^2(4K+6) + sRC(5+K) + 1} \quad — (12)$$

substituting $s = j\omega$, $s^2 = j^2\omega^2 = -\omega^2$, $s^3 = j^3\omega^3 = -j\omega^3$ in eqn (12).

$$A\beta = \frac{-j\omega^3 KR^3C^3 h_{fe}}{-j\omega^3R^3C^3(3K+1) - \omega^2C^2R^2(4K+6) + j\omega RC(5+K) + 1}$$

seperating real & imaginary parts in denominator, we get

$$A\beta = \frac{-j\omega^3 KR^3C^3 h_{fe}}{[1-(4K+6)\omega^2R^2C^2] - j[(3K+1)\omega^3R^3C^3 - (5+K)\omega RC]}$$

multiplying numerator & denominator by $\frac{-1}{j\omega^3R^3C^3}$, we get

$$= \frac{Kh_{fe}}{-\frac{[1-(4K+6)\omega^2R^2C^2]}{j\omega^3R^3C^3} + \frac{j[(3K+1)\omega^3R^3C^3 - (5+K)\omega RC]}{j\omega^3R^3C^3}}$$

$$A\beta = \frac{Kh_{fe}}{j\left[\frac{1}{\omega^3R^3C^3} - \frac{4K}{\omega RC} - \frac{6}{\omega RC}\right] + \left[3K+1 - \frac{5}{\omega^2R^2C^2} - \frac{K}{\omega^2R^2C^2}\right]}$$

Replacing $\frac{1}{\omega RC} = \alpha$, we get

$$A\beta = \frac{Kh_{fe}}{[3K+1 - 5\alpha^2 - K\alpha^2] + j[\alpha^3 - 4K\alpha - 6\alpha]} \quad — (13)$$

As per Barkhausen Criterion, $\angle A\beta = 0^\circ$, so the angle of term in eqn. (13) is 0° hence the imaginary part of term must be zero.

$$\alpha^3 - 4K\alpha - 6\alpha = 0 \Rightarrow \alpha[\alpha^2 - 4K - 6] = 0$$

$$\alpha^2 = 4K+6 \Rightarrow \alpha = \sqrt{4K+6} \quad — (14)$$

$$\frac{1}{\omega RC} = \sqrt{4K+6} \Rightarrow \omega = \frac{1}{RC\sqrt{4K+6}}$$

$$\boxed{\therefore f = \frac{1}{2\pi RC\sqrt{4K+6}}} \quad — (15)$$

At the same frequency $|A\beta|=1$ we get.

Substituting $\alpha = \sqrt{4K+6}$ in eqn ⑭ then.

$$A\beta = \frac{Kh_{fe}}{(3K+1)-(5+K)(4K+6)} = \frac{Kh_{fe}}{3K+1-20K-30-4K^2-6K}$$

$$\therefore A\beta = \frac{Kh_{fe}}{-4K^2-23K-29}$$

$$|A\beta|=1 \Rightarrow \left|\frac{Kh_{fe}}{-4K^2-23K-29}\right| = 1$$

$$Kh_{fe} = 4K^2+23K+29$$

$$\boxed{h_{fe} = 4K+23+\frac{29}{K}} \quad \text{—— ⑯}$$

This must be the value of h_{fe} for oscillations.

minimum value of h_{fe} for oscillation is given by.

$$\frac{dh_{fe}}{dK}=0 \Rightarrow \frac{d}{dK}\left[4K+23+\frac{29}{K}\right]=0$$

$$4-\frac{29}{K^2}=0 \Rightarrow K^2=\frac{29}{4}$$

$\boxed{K = 2.6925}$ for minimum h_{fe} —⑰

Substituting ⑰ in ⑯ we get. $\boxed{h_{fe\,min} = 44.54}$ —⑱.

Thus for the circuit to oscillate we must select the transistor whose $h_{fe\,min}$ should be greater than 44.54.

Advantages

1. The circuit provides sinusoidal output waveform.
2. Simple to design & can produce output over audio frequency
3. It is a fixed frequency oscillator

Disadvantages

By changing values of R&C, the freq of oscillator can be changed. But values of R&C of all three sections must be changed simultaneously to satisfy oscillating conditions. But this is practically impossible. Hence phase shift oscillator is considered as fixed freq. oscillator.

The frequency stability is poor due to changes in values of various components due to effect of temperature, aging etc.

WEINBRIDGE OSCILLATOR:

Generally in an oscillator, amplifier stage introduces 180° phase shift and feedback network introduces additional 180° phase shift to obtain a phase shift of 360° (2π-radians) around a loop.

But weinbridge oscillator uses a non inverting amplifier and hence does not require any phase shift during amplifier stage.

As total phase shift required is 0° (or $2n\pi$ radians), in wein bridge type, no phase shift is necessary through feedback.

A basic weinbridge used in this oscillator and an amplifier stage is as shown in fig.

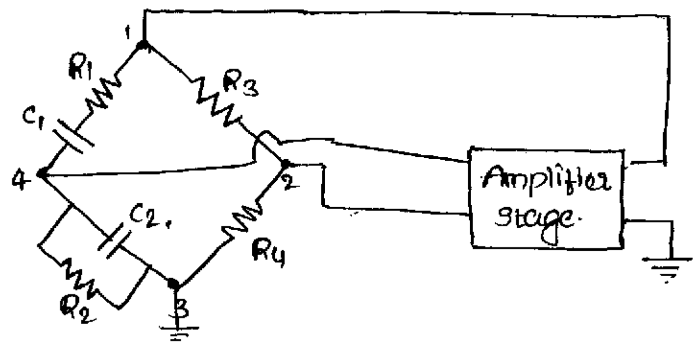


Fig: Basic circuit of weinbridge oscillator.

The output of amplifier is applied between the terminals '1' and '3' which is the input to the feedback network. while the amplifier input is supplied from diagonal terminals 2 and 4, which is the output from feedback network. thus amplifier supplied its own input through the weinbridge as a feedback network.

The two arms of the bridge R_1, C_1 in series and R_2, C_2 are in parallel are called "frequency sensitive arms". This is because the components of these two arms decide the frequency of oscillator.

Here Vin be the input voltage applied to the feedback network. between the terminals "1 and 3".

V_f be the output of feedback network. between the terminals "2 and 4" as shown in fig.

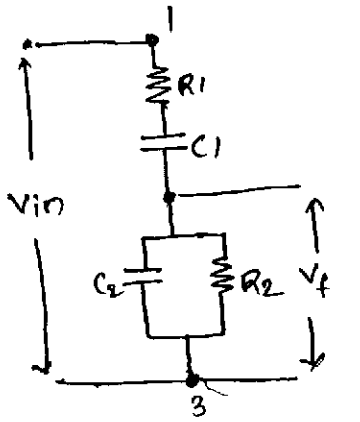
$$Z_1 = R_1 + \frac{1}{j\omega C_1}$$

$$Z_1 = \frac{1 + j\omega R_1 C_1}{j\omega C_1} \quad \text{—(1)}$$

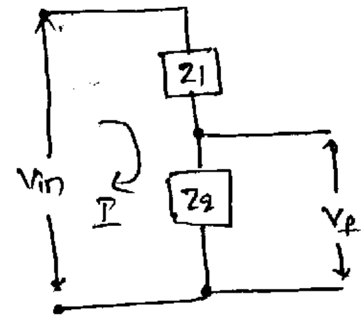
$$Z_2 = R_2 \parallel \frac{1}{j\omega C_2}$$

$$Z_2 = \frac{R_2 \times \frac{1}{j\omega C_2}}{R_2 + \frac{1}{j\omega C_2}}$$

$$Z_2 = \frac{R_2}{1 + j\omega C_2 R_2} \quad \text{—(2)}$$



FB N/W of WB osc.



simplified circuit

Replacing $s = j\omega$; $Z_1 = \frac{1 + sR_1C_1}{sC_1}$, $Z_2 = \frac{R_2}{1 + sR_2C_2}$

The current I is given by.

$$I = \frac{V_{in}}{Z_1 + Z_2}$$

The feedback voltage $V_f = IZ_2$.

$$V_f = \frac{V_{in} Z_2}{Z_1 + Z_2} \quad \text{—(3)}$$

The feedback factor β is given by $\beta = \frac{V_f}{V_{in}} = \frac{Z_2}{Z_1 + Z_2}$ —(4)

$$\beta = \frac{\frac{R_2}{1 + sR_2C_2}}{\frac{1 + sR_1C_1}{sC_1} + \frac{R_2}{1 + sR_2C_2}} = \frac{R_2 s C_1}{(1 + sR_1C_1)(1 + sR_2C_2) + R_2 s C_1}$$

$$\beta = \frac{sC_1R_2}{1 + s(R_1C_1 + R_2C_2 + R_2C_1) + s^2 R_1R_2C_1C_2}.$$

Replacing $s = j\omega$, $s^2 = -\omega^2$. then $\beta = \frac{j\omega C_1 R_2}{j\omega(R_1C_1 + R_2C_2 + R_2C_1) + (1 - \omega^2 R_1R_2C_1C_2)}$

$$\beta = \frac{\cancel{j}\omega R_2 C_1}{\cancel{j}\left[\omega(R_1C_1 + R_2C_2 + C_1R_2) + \frac{1}{j}(1 - \omega^2 R_1R_2C_1C_2)\right]} \qquad \because \frac{1}{j} = -j$$

$$\beta = \frac{\omega C_1 R_2}{\omega[R_1C_1 + R_2C_2 + R_2C_1] - j(1 - \omega^2 R_1R_2C_1C_2)} \quad \text{—(5)}$$

To have zero phase shift for feedback network, its imaginary part must be zero.

$$1-\omega^2 R_1R_2C_1C_2=0$$

$$\omega^2=\frac{1}{R_1R_2C_1C_2} \Rightarrow \boxed{\omega=\frac{1}{\sqrt{R_1R_2C_1C_2}}} \quad —(6)$$

$$\boxed{f=\frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}} \quad —(7)$$

This is the frequency of oscillator which shows that the components of frequency sensitive arms decide the frequency.

In practice $R_1=R_2=R$ and $C_1=C_2=C$ then.

$$f=\frac{1}{2\pi\sqrt{R^2C^2}} \Rightarrow \boxed{f=\frac{1}{2\pi RC}} \quad —(8)$$

then $\beta=\dfrac{C_1R_2}{R_1C_1+R_2C_2+R_2C_1}$

At $R_1=R_2=R$, $C_1=C_2=C$ then $\beta=\dfrac{\cancel{RC}}{3\cancel{RC}} \Rightarrow \boxed{\beta=\frac{1}{3}}$ —(9)

The positive sign of β indicates that the phase shift of feedback is $0°$.

To satisfy Barkhausen criterion for sustained oscillations:

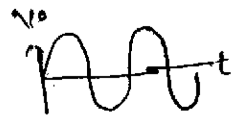
$$|A\beta|\geq 1 \Rightarrow |A|\geq\frac{1}{\beta}\geq\frac{1}{(1/3)} \Rightarrow \boxed{|A|\geq 3}$$

This is required gain of amplifier stage without any phase shift.

$$\boxed{f=\frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}} \quad \boxed{\beta=\frac{C_1R_2}{R_1C_1+R_2C_2+R_2C_1}} \quad \boxed{A\geq\frac{R_1C_1+R_2C_2+R_2C_1}{C_1R_2}} \quad —(10)$$

The advantage of Wein bridge oscillator is that by varying the two capacitor values simultaneously by making them on the common shaft different frequency ranges can be provided.

If the bridge is balanced, then the oscillations will not take place as $|A\beta|\geq 1$ will not be satisfied.



Transistorized wein bridge oscillator

in this circuit, two stage CE amplifier is used. each stage contributes 180° phase shift, hence the total phase shift due to amplifier stage becomes 360° or 0° which is necessary as per the oscillator conditions.

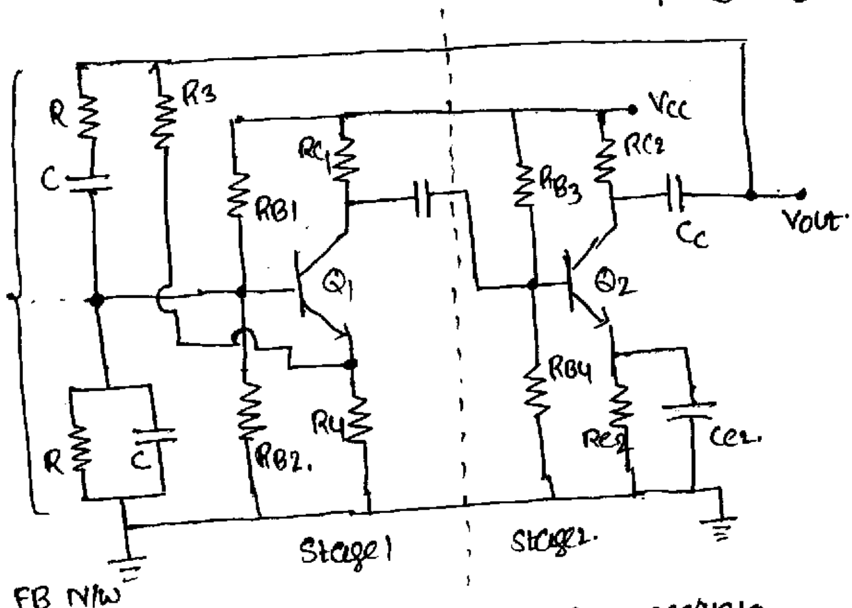


Fig: Transistorized wein bridge oscillator

The bridge consist of R & C in series, R & C in parallel, R_3 & R_4. The feedback is applied from collector of Q_2 through capacitor C_c to the bridge circuit. The resistor R_4 serves as dual purpose of emitter resistance of transistor Q_1 and also the element of wein bridge.

The two stage amplifier provides a gain ($A \gg 3$) and it is necessary to reduce it. To reduce gain, the negative feedback is used without by passing the resistance R_4. The negative feedback can accomplish the gain stability & can control the output magnitude.

The negative feedback also reduces the distortion and therefore output is pure sinusoidal in nature. The amplitude stability can be improved using a non linear resistor R_4. Due to this, the loop gain depends on the amplitude of oscillations.

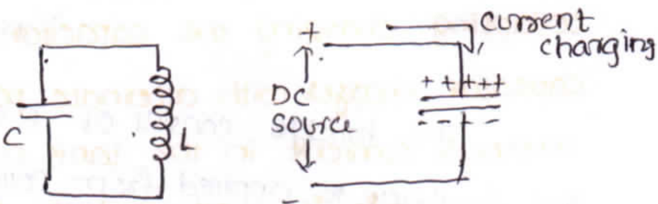
increase in amplitude of oscillations increases the current through non linear resistance, which results into an increase in value of non linear resistance R_4. when this value increases, a greater amount of negative feedback is applied. this reduces the loop gain A which in turn reduces and controls the signal amplitude.

LC OSCILLATORS:

The Oscillators which use the elements L and C to produce the Oscillations are called LC Oscillators. The circuit using elements L and C is called "tank circuit" or "Oscillatory circuit" which is an important part of LC Oscillators. This circuit is also referred as resonating circuit or tuned circuit. These oscillators are used for high frequency range from 200KHz upto few GHz. Due to high frequency range, these oscillators are often used for sources of RF energy.

Operation of LC Tank circuit:

The LC tank circuit consists of elements L and C connected in parallel as shown in fig.

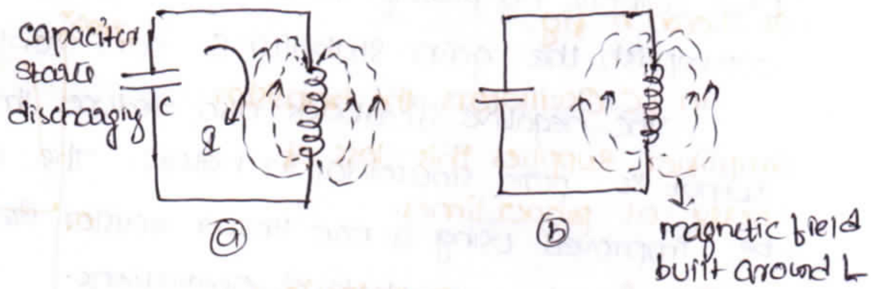


Tank circuit. initial charging

capacitor is initially charged with polarities as shown in fig.

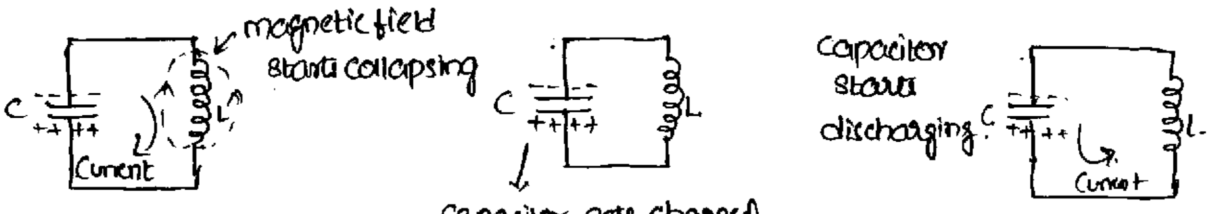
When the capacitor gets charged, the energy gets stored in capacitor is called electrostatic energy. When such a charged capacitor is connected across inductor 'L' in a tank circuit, the capacitor starts discharging through 'L' as shown in fig.

The arrow mark indicates direction of flow of current. Due to such current flow, magnetic field is induced around the inductor.



Thus inductor starts storing the energy. When the capacitor is fully discharged, maximum current flows through circuit. At this instant all the electrostatic energy get stored as magnetic energy in the inductor as shown in fig (b)

Now the magnetic field around L starts collapsing. As per lenz law, this starts discharging the capacitor with polarity as shown in fig. After some time, capacitor gets fully charged with opposite polarities as compared to its initial polarities as shown.



(a) (b) (c)

Now the capacitor again starts discharging through inductor 'L'. But the direction of current through circuit is now opposite to the direction of current earlier in the circuit as shown in fig c.

Again electrostatic energy is converted to magnetic energy. When the capacitor is fully discharged, the magnetic field starts collapsing charging the capacitor again in opposite direction. Thus the capacitor charges with alternate polarities & discharges producing alternate current in the tank circuit.

This is nothing but oscillatory current. But every time when energy is transferred from C to L and L to C, the losses occur due to which amplitude of oscillating current keeps on decreasing every time when energy transfer takes place. Hence we get exponentially decaying oscillations called damped oscillations as shown in fig.

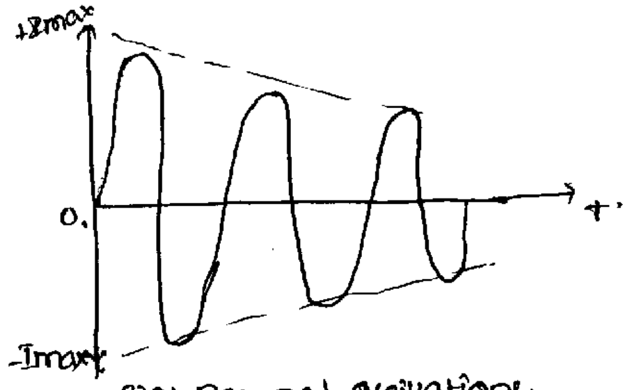


Fig: Damped oscillations.

In LC oscillators, the transistor amplifier supplies this loss of energy at proper times.

The frequency of oscillations generated by LC Tank circuit depends on the values of L & C and is given by.

$$f = \frac{1}{2\pi\sqrt{LC}} \ Hz. \quad \text{—— (1)}$$

where L is in henries & C is in farads

Basic form of LC circuit:

LC tuned circuit forms feedback network with an Op-amp, FET or BJT as an active device in amplifier. The following fig. shows the basic form of LC oscillators circuit with gain of the amplifier as Av.

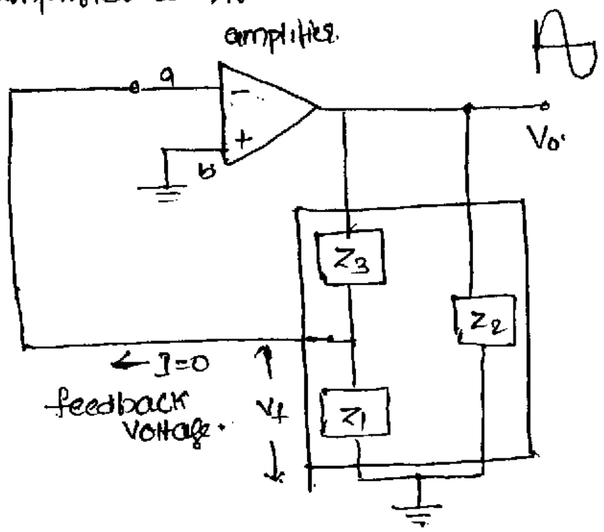


Fig: Basic circuit

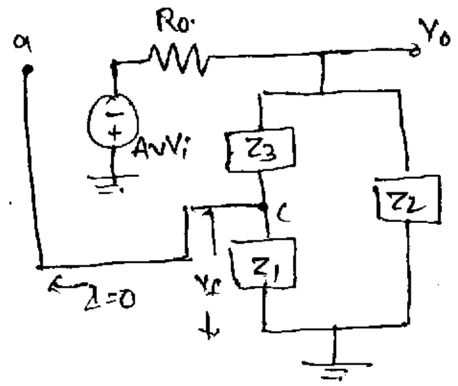


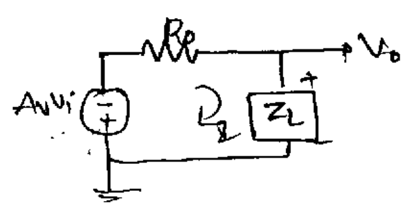
Fig: Equivalent circuit

The amplifier network consisting of impedances Z_1, Z_2 and Z_3. As Assume an active device with infinite impedance such as FET or Opamp. The basic circuit can be replaced by its linear equivalent circuit as shown in above fig.

Amplifier provides a phase shift of 180°, while the feedback network provides an additional phase shift of 180° to satisfy the Barkhausen Condition.

As the input impedance of amplifier is infinite, there is no current flowing towards the input terminals. Let R_o be the o/p impedance of amplifier stage.

As I=0, Z_1, Z_3 appears in series & the combination is in parallel with Z_2. The equivalent Z_L i.e., load resistance. so the circuit can be reduced as shown in fig.



$$I = \frac{-A_V V_i}{R_o + Z_L} \quad \text{--- (2)}$$

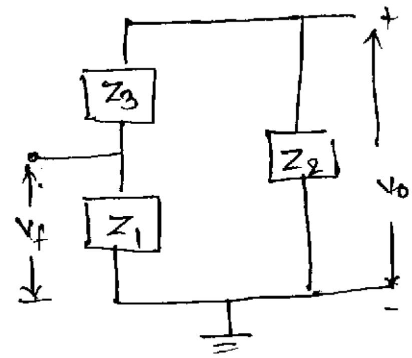
$$V_o = I Z_L .$$

$$V_o = \frac{-A_V V_i Z_L}{R_o + Z_L} \Rightarrow \boxed{\frac{V_o}{V_i} = A = \frac{-A_V Z_L}{R_o + Z_L}} \quad \text{--- (3)}$$

A is the gain of amplifier stage. The negative sign indicates the amplifier stage introduces 180° phase shift.

Analysis of Feedback:

For calculation of feedback factor (β), consider only the feedback circuit as shown in fig.



From the voltage division in parallel circuit we have.

$$V_f = V_o \left[\frac{Z_1}{Z_1 + Z_3} \right] \quad \text{--- (4)}$$

$$\beta = \frac{V_f}{V_o} = \frac{Z_1}{Z_1 + Z_3} \quad \text{--- (2)}$$

But as the phase shift of feedback is 180°. $\beta = \frac{-Z_1}{Z_1 + Z_3}$ --- (3)

Obtain an expression for $-A\beta$ as basic barkhausen criterion.

$$-A\beta = 1$$

$$-A\beta = \frac{-A_V Z_1 Z_L}{(R_o + Z_L)(Z_1 + Z_3)} \quad \text{--- (3)}$$

$$Z_L = (Z_1 + Z_3) \| Z_2 = \frac{(Z_1 + Z_3) Z_2}{Z_1 + Z_2 + Z_3} \quad \text{--- (4)}$$

$$-A\beta = \frac{-A_V Z_1 \left(\frac{(Z_1 + Z_3) Z_2}{Z_1 + Z_2 + Z_3} \right)}{\left(R_o + \frac{(Z_1 + Z_3) Z_2}{Z_1 + Z_2 + Z_3} \right) (Z_1 + Z_3)} = \frac{-A_V Z_1 Z_2 \frac{(Z_1 + \cancel{Z_3})}{Z_1 + \cancel{Z_2} + Z_3}}{\left[\frac{R_o(Z_1 + Z_2 + Z_3) + Z_2}{Z_1 + Z_3} \right] \frac{(Z_1 + Z_3)}{Z_1 + \cancel{Z_2} + Z_3}}$$

$$\boxed{-A\beta = \frac{-A_V Z_1 Z_2}{R_o(Z_1 + Z_2 + Z_3) + Z_2(Z_1 + Z_3)}} \quad \text{--- (5)}$$

As Z_1, Z_2 & Z_3 are pure reactances.

$Z_1 = jX_1$, $Z_2 = jX_2$ & $Z_3 = jX_3$, $X = \omega L$ for inductive reactance

$X = \frac{-1}{\omega C}$ for capacitive reactance

$$-A\beta = \frac{-A_v \, jX_1 \, jX_2}{-X_2(X_1+X_3) + jR_o(X_1+X_2+X_3)} \quad \text{--- (6)}$$

To have 180° phase shift, the imaginary part must be zero

$$\therefore \; X_1 + X_2 + X_3 = 0 \quad \text{--- (7)}$$

Substituting in equation (6)

$$-A\beta = \frac{-A_v X_1 X_2}{X_2(X_1+X_3)}$$

But from (7) $\quad -X_2 = X_1 + X_3$

$$-A\beta = \frac{-A_v X_1}{-X_2} = A_v \frac{X_1}{X_2} \quad \text{--- (8)}$$

According to Barkhausen criterion, $-A\beta$ must be positive & must be greater than or equal to unity. As A_v is positive, the $-A\beta$ will be positive only when X_1 and X_2 will have same sign.

This indicates that X_1 and X_2 must be of same type of reactances. either both inductive or capacitive.

Types of LC Oscillators :

Based on the design of X_1, X_2 and X_3 (Z_1, Z_2 & Z_3) in the feedback network, we have

	X_1	X_2	X_3
HARTLEY Oscillator	L	L	C
COLPITTS Oscillator	C	C	L.

56, 72, 74, 58, 67, 71

Hartley Oscillator:

A LC Oscillator which uses two inductive reactances & one capacitive reactance in its feedback network is Hartley oscillator.

Transistorized Hartley Oscillator:

The amplifier stage uses an active device, a transistor in CE configuration as shown in fig.

The resistances R_1 & R_2 are the biasing resistances.

The RFC is the "Radio frequency choke".

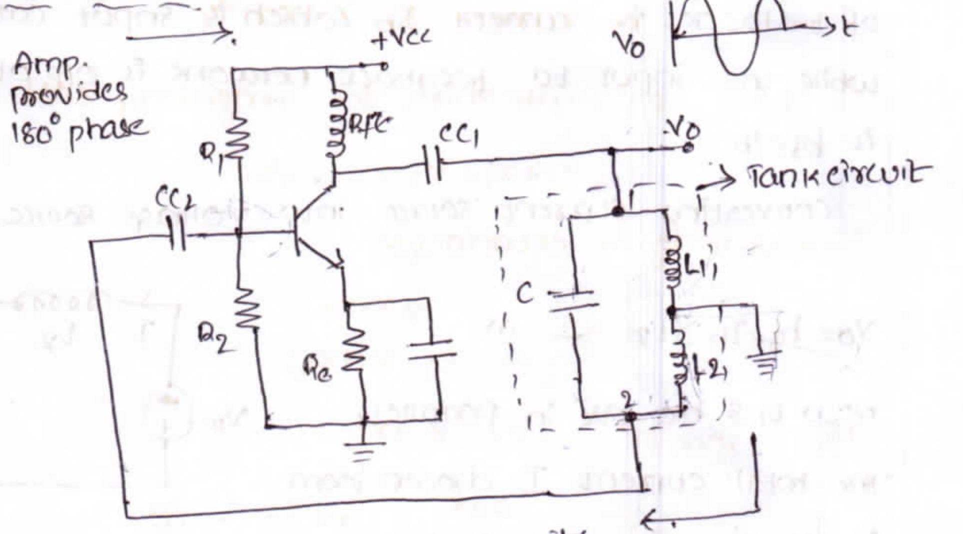


Fig: Transistorized Hartley Oscillator.

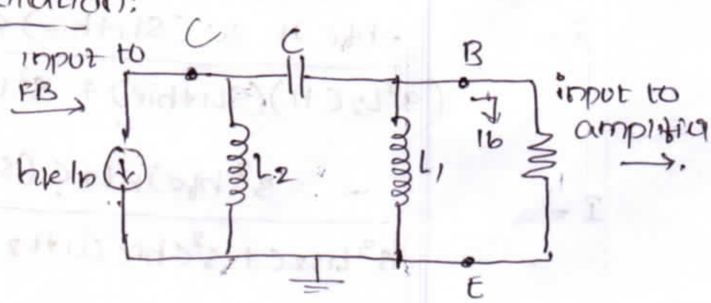
Its reactance value is very high. hence it can be treated as open circuit. while for DC conditions, the reactance is zero, hence causes no problem for dc capacitances.

Hence due to RFC, the isolation between a.c & dc operation is achieved. Th

The CE amplifier provides phase shift of 180°. As the centre of L_1 & L_2 is grounded, the upper end becomes positive & the lower end becomes negative & viceversa. So the LC feedback network gives an additional phase shift of 180°, necessary to satisfy oscillation condition.

Derivation of frequency of oscillation:

The o/p current which is collector current is $h_{fe} I_b$. where I_b is base current.

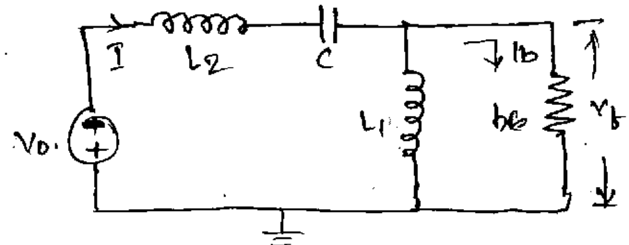


Equivalent circuit.

Assuming coupling capacitor C is b/w base and collector.
L_1 is b/w base and emitter, L_2 is b/w base and collector & emitter.
the equivalent circuit is shown above.

As hie is the input impedance of transistor. The output of feedback is current I_b which is input current of transistor, while the input to feedback network is o/p of transistor which is $h_{fe} I_b$.

Converting current source into voltage source, we get.



Simplified equivalent circuit.

$V_o = h_{fe} I_b X_{L2}$ —①

Now L_1 & hie are in parallel,
The total current I drawn from supply is

$$I = \frac{-V_o}{(X_{L2}+X_C)+(X_{L1}\|h_{ie})} \quad —②$$

Note - -ve sign indicate that current direction shown in opposit to polarities of V_o.

$$X_{L2}+X_C = j\omega L_2 + \frac{1}{j\omega C} \; ; \quad X_{L1}\|h_{ie} = \frac{j\omega L_1 h_{ie}}{j\omega L_1 + h_{ie}}$$

$$\therefore \quad I = \frac{-h_{fe} I_b \, j\omega L_2}{\left[j\omega L_2 + \frac{1}{j\omega C}\right] + \left[\frac{j\omega L_1 h_{ie}}{j\omega L_1 + h_{ie}}\right]} \quad —③$$

Replacing $j\omega$ by s; we get.

$$I = \frac{-h_{fe} I_b SL_2}{\left(SL_2 + \frac{1}{SC}\right) + \frac{SL_1 h_{ie}}{SL_1 + h_{ie}}} = \frac{-h_{fe} I_b SL_2}{\frac{S^2 L_2 C + 1}{SC} + \frac{SL_1 h_{ie}}{SL_1 + h_{ie}}}$$

$$I = \frac{-h_{fe} I_b \, SC\,(SL_1 + h_{ie})\, SL_2}{(S^2 L_2 C + 1)(SL_1 + h_{ie}) + S^2 L_1 C h_{ie}}$$

$$I = \frac{-S^2 h_{fe} I_b L_2 C\,(SL_1 + h_{ie})}{S^3 L_1 L_2 C + S^2 C h_{ie}(L_1 + L_2) + SL_1 + h_{ie}} \quad —④$$

According to current Divider in parallel circuit.

$$I_b = I \times \frac{X_{L1}}{X_{L1}+h_{ie}} = I \times \frac{j\omega L_1}{j\omega L_1 + h_{ie}} = I \times \frac{SL_1}{SL_1+h_{ie}} \quad \text{—— (5)}$$

Substituting value of I from eqn (4) in eqn (5)

$$I_b = \frac{-s^2 h_{fe} I_b L_2 C (SL_1+h_{ie})}{s^3 L_2 L_1 C + s^2 C h_{ie}(L_1+L_2)+SL_1+h_{ie}} \times \frac{SL_1}{SL_1+h_{ie}}$$

$$\Rightarrow \frac{-s^3 h_{fe} L_1 L_2 C}{s^3 L_1 L_2 C + s^2 C h_{ie}(L_1+L_2) + SL_1 + h_{ie}} = 1 \quad \text{—— (6)}$$

replacing s by $j\omega$, $\therefore s^2 = -\omega^2$ $s^3 = -j\omega^3$ we get.

$$\frac{j\omega^3 h_{fe} L_1 L_2 C}{-j\omega^3 L_1 L_2 C - \omega^2 C h_{ie}(L_1+L_2) + j\omega L_1 + h_{ie}} = 1$$

$$\frac{j\omega^3 h_{fe} L_1 L_2 C}{j\left[(\omega L_1 - \omega^3 L_1 L_2 C) + [jh_{ie} - j\omega^2 C h_{ie}(L_1+L_2)]\right]} = 1$$

$$\Rightarrow \frac{\omega^3 h_{fe} L_1 L_2 C}{(\omega L_1 - \omega^3 L_1 L_2 C) + j h_{ie}[1-\omega^2 C(L_1+L_2)]} = 1$$

To satisfy the condition, the imaginary part should be zero.

$$h_{ie}(1-\omega^2 C(L_1+L_2)) = 0$$

$$\omega^2 = \frac{1}{C(L_1+L_2)} \Rightarrow \omega = \frac{1}{\sqrt{(L_1+L_2)C}}$$

$$\therefore \boxed{f = \frac{1}{2\pi\sqrt{C(L_1+L_2)}}} \quad \text{—— (7)}$$

This is the frequency of Oscillations. At this frequency, the value of h_{fe} can be obtained as.

$$1 = \frac{\omega^3 h_{fe} L_1 L_2 C}{\omega(L_1 - \omega^2 L_1 L_2 C)} = \frac{h_{fe} L_1 L_2 C \omega^2}{L_1 - \omega^2 L_1 L_2 C}$$

$$1 = \frac{\omega^2 h_{fe} \cancel{K_1} L_2 C}{\cancel{K_1}(1-\omega^2 L_2 C)} = \frac{\omega^2 h_{fe} L_2 C}{1-\omega^2 L_2 C}.$$

$$1 = \frac{\frac{1}{C(L_1+L_2)} h_{fe} \cdot L_2 C}{1 - \frac{1}{C(L_1+L_2)} L_2 C} \qquad \because \omega^2 = \frac{1}{C(L_1+L_2)}$$

$$1 = \frac{h_{fe} \frac{L_2}{L_1+L_2}}{1 - \frac{L_2}{L_1+L_2}} = h_{fe} \frac{L_2}{L_1}$$

$$\therefore \boxed{h_{fe} = \frac{L_1}{L_2}} \quad \text{—(8)}$$

This is the value of h_{fe} required to satisfy condition of oscillation.

For a mutual inductance of M.

$$\boxed{h_{fe} = \frac{L_1+M}{L_2+M}} \quad \text{—(9)}$$

L_1+L_2 is equivalent inductance of two inductances L_1 & L_2 connected in series denoted as

$$L_{eq} = L_1+L_2.$$

$$\boxed{\therefore \quad f = \frac{1}{2\pi\sqrt{C L_{eq}}}} \quad \text{—(10)}$$

So, if the capacitor C is kept variable, frequency can be varied over a large range as per the requirement.

In practice L_1 and L_2 may be wound on a single core so that there exists mutual inductance between them denoted as M.

In such a case, mutual inductance is considered while determining equivalent inductance L_{eq}. Hence $\boxed{L_{eq} = L_1+L_2+2M}$ —(11)

If L_1 & L_2 are assisting each other then sign of 2M is positive. While L_1 and L_2 are in series opposition then sign of 2M is negative.

COLPITTS OSCILLATOR:

An LC Oscillator which uses two capacitive reactances and one inductive reactance in the feedback network. i.e.) tank circuit is called Colpitts oscillator.

Transistorised colpitts oscillator:

The amplifier stage uses an active device as a transistor in CE configuration.

The practical circuit is shown in fig:

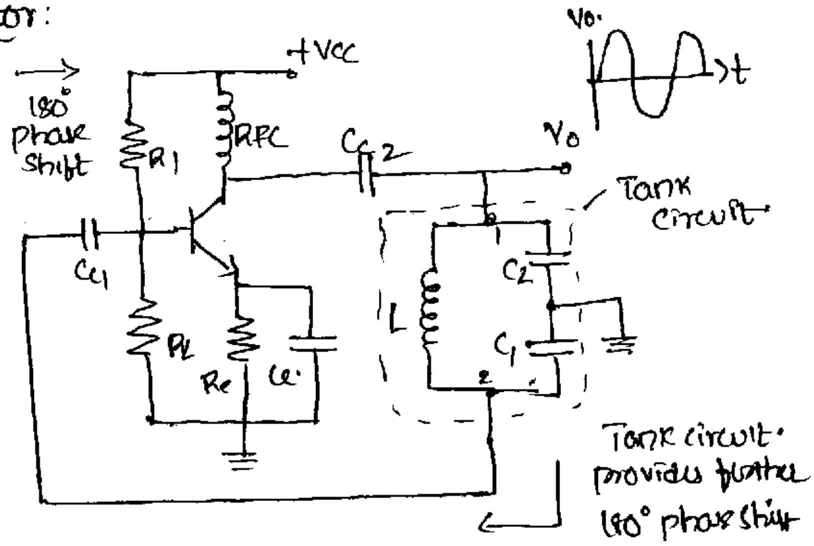
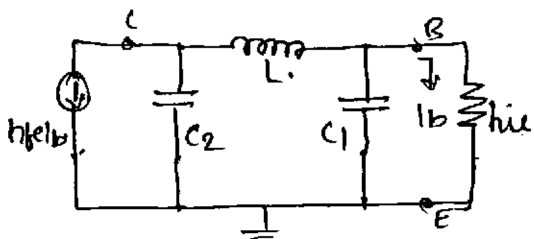


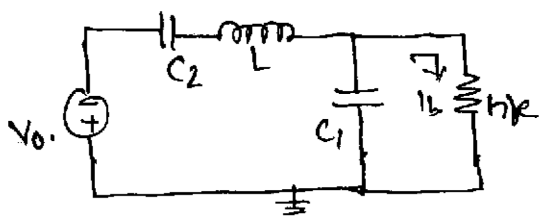
Fig: Transistorized colpitts oscillator.

Derivation of frequency of oscillation:

The output current I_c which is $h_{fe}I_b$ acts as input to the feedback which the base current acts as output current of tank circuit flowing through the input impedance of the amplifier hie. The equivalent circuit of tank circuit is shown in fig:



(a) equivalent circuit



(b) simplified circuit

converting current source into voltage source, we get circuit as shown in (b).

$$V_o = h_{fe} I_b X_{C_2} = h_{fe} I_b \frac{1}{j\omega C_2} \quad \text{--- (1)}$$

The total current drawn from circuit supply is

$$I = \frac{-V_o}{(X_{C_2}+X_L) + (X_{C_1} \| h_{ie})} \quad \text{--- (2)}$$

$X_{C2} + X_L = \frac{1}{j\omega C_2} + j\omega L$ and $X_{C1} \| h_{ie} = \frac{\frac{1}{j\omega C_1} \cdot h_{ie}}{\frac{1}{j\omega C_1} + h_{ie}}$

substituting in eqn ② we get.

$$I = \frac{-h_{fe} I_b \left(\frac{1}{j\omega C_2}\right)}{\left[\frac{1}{j\omega C_2} + L\right] + \left[\frac{\frac{h_{ie}}{j\omega C_1}}{h_{ie} + \frac{1}{j\omega C_1}}\right]} \quad \text{—③}$$

Replacing $j\omega$ by 's', we have $I = \dfrac{-h_{fe} I_b \frac{1}{sC_2}}{\frac{1}{sC_2} + L + \frac{\frac{h_{ie}}{sC_1}}{h_{ie} + \frac{1}{sC_1}}}$.

$$I = \frac{-h_{fe} I_b \frac{1}{sC_2}}{\frac{1+LsC_2}{sC_2} + \frac{h_{ie}}{1+h_{ie}sC_1}}$$

$$= \frac{-h_{fe} I_b \frac{1}{\cancel{sC_2}} \times (1+h_{ie}sC_1)(\cancel{sC_2})}{(1+h_{ie}sC_1)(1+LsC_2) + h_{ie}sC_2}$$

$$I = \frac{-h_{fe} I_b (1+sC_1 h_{ie})}{s^3 LC_1C_2 h_{ie} + s^2 LC_2 + s h_{ie}(C_1+C_2)+1} \quad \text{—④}$$

According to current divider rule in parallel circuit.

$$I_b = I \frac{X_{C1}}{X_{C1}+h_{ie}} = \frac{I \frac{1}{j\omega C_1}}{\frac{1}{j\omega C_1} + h_{ie}} = \frac{I}{1+h_{ie} j\omega C_1} = \frac{I}{1+sC_1 h_{ie}} \quad \text{—⑤.}$$

Substitute I_b in eqn ④, we get.

$$I = \frac{-h_{fe} \cancel{I_b} (1+\cancel{sC_1 h_{ie}}) \frac{\cancel{I}}{(1+\cancel{sC_1 h_{ie}})}}{1+h_{ie}s(C_1+C_2) + s^2 LC_2 + s^3 LC_1C_2 h_{ie}}$$

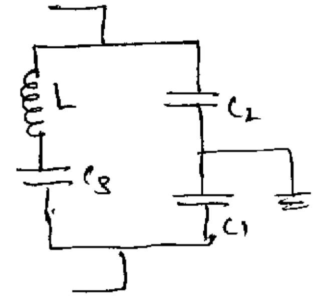
$$1 = \frac{-h_{fe}}{s^3 LC_1C_2 h_{ie} + s^2 LC_2 + s h_{ie}(C_1+C_2)+1}.$$

Replacing $s = j\omega$, $s^2 = -\omega^2$, $s^3 = -j\omega^3$.

$$1 = \frac{-h_{fe}}{-j\omega^3 LC_1C_2 h_{ie} - \omega^2 LC_2 + j\omega h_{ie}(C_1+C_2)+1}$$

CLAPP OSCILLATOR:

To achieve frequency stability, colpitts circuit is slightly modified, in practice called clapp oscillator. A one more capacitance C_3 is introduced in series with the inductance as shown in fig.



The value of C_3 is much smaller than the value of C_1 & C_2.

The equivalent capacitance becomes $\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$ —(1)

The oscillator frequency is given by

$$\boxed{f = \frac{1}{2\pi\sqrt{LC_{eq}}}} \quad \text{—(2)}$$

As $\frac{1}{C_1}$ & $\frac{1}{C_2}$ are very small, they can be neglected.

Now $\frac{1}{C_{eq}} = \frac{1}{C_3} \Rightarrow C_{eq} = C_3$.

$$\therefore \boxed{f = \frac{1}{2\pi\sqrt{LC_3}}} \quad \text{—(3)}$$

Here across C_3, there is no transistor parameter & hence the frequency of clapp oscillator is stable & accurate.

The transistor & stray capacitance have no effect on C_3. hence good frequency stability is achieved in clapp oscillator.

Hence practically clapp oscillator is preferred over colpitts oscillator.

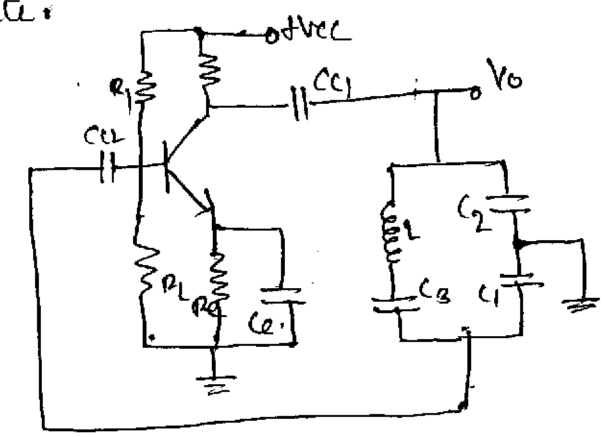


Fig Transistorized clapp Oscillator.

The transistorized clapp oscillator is shown in fig.

Another advantage of C_3 is that it can be kept variable. As frequency is dependent on C_3, the frequency can be varied in desired range.

$$1 = \frac{-h_{fe}}{(1-\omega^2 LC_2) + j\omega h_{ie}(C_1+C_2-\omega^2 LC_1C_2)} \quad \text{--- (6)}$$

To satisfy barkhausen criterion, the imaginary term in eqn should be zero.

$$\omega h_{ie}(C_1+C_2-\omega^2 LC_1C_2) = 0$$

$$\omega^2 LC_1C_2 = C_1+C_2$$

$$\omega^2 = \frac{C_1+C_2}{LC_1C_2}$$

$$\omega^2 = \frac{1}{L\frac{C_1C_2}{C_1+C_2}} = \frac{1}{LC_{eq}}$$

$$\boxed{\omega = \frac{1}{\sqrt{LC_{eq}}}} \quad \text{--- (7)}$$

$$\boxed{f = \frac{1}{2\pi\sqrt{LC_{eq}}}} \quad \text{--- (8)}$$

This is the frequency of oscillation in colpitts oscillator.

Substituting this frequency in equation (6) we get.

$$1 = \frac{-h_{fe}}{1-\omega^2 LC_2} = \frac{-h_{fe}}{1-\frac{1}{\cancel{L}\frac{C_1C_2}{C_1+C_2}} \times \cancel{L}C_2} \cdot \frac{-h_{fe}}{1-\frac{1}{L\frac{C_1C_2}{C_1+C_2}} \times LC_2}$$

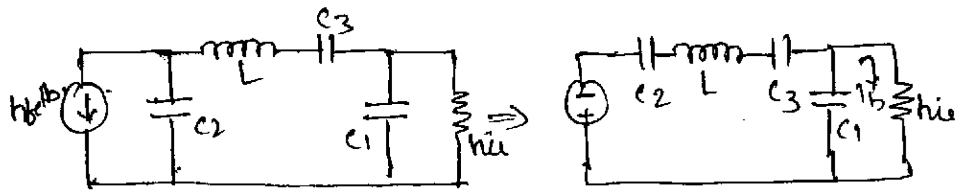
$$1 = \frac{-h_{fe}}{1-\frac{C_1+C_2}{C_1}} \Rightarrow 1 = \frac{-h_{fe}}{\frac{\cancel{C_1C_2}-C_2^2+\cancel{C_1C_2}}{C_1C_2}}$$

$$1 = \frac{+h_{fe}C_1C_{\cancel{2}}}{+C_2^{\cancel{2}}} \Rightarrow h_{fe}\frac{C_1}{C_2} = 1$$

$$\boxed{h_{fe} = \frac{C_2}{C_1}} \quad \text{--- (9)}$$

Derivation of Frequency of Oscillation.

The equivalent circuit of Clapp Oscillator feedback network is shown below



$V_0 = h_{fe} I_b X_{C_2}$.

$$V_0 = h_{fe} I_b \frac{1}{j\omega C_2} \quad \text{—} ①$$

$$X_{C_2} + X_{C_3} + X_{L} = \frac{1}{j\omega C_2} + \frac{1}{j\omega C_3} + j\omega L.$$

$$I = \frac{-V_0}{(X_{C_2}+X_{C_3}+X_L)+(X_{C_1} \| h_{ie})} \quad \text{—} ②$$

$$X_{C_1} \| h_{ie} = \frac{\frac{h_{ie}}{j\omega C_1}}{\frac{1}{j\omega C_1} + h_{ie}}.$$

$$\therefore I = \frac{-h_{fe} I_b \frac{1}{j\omega C_2}}{\frac{1}{j\omega C_2} + \frac{1}{j\omega C_3} + j\omega L + \frac{h_{ie}}{1 + j\omega C_1 h_{ie}}}$$

Replacing $j\omega$ by s, we get

$$I = \frac{-h_{fe} I_b \frac{1}{sC_2}}{\frac{1}{sC_2} + \frac{1}{sC_3} + sL + \frac{h_{ie}}{1 + sC_1 h_{ie}}}.$$

$$I = \frac{-h_{fe} I_b}{sC_2\left[\frac{1}{sC_2} + \frac{1}{sC_3} + sL + \frac{h_{ie}}{(1+sC_1 h_{ie})}\right]}$$

This gives

$$I = \frac{-h_{fe} I_b C_3 (1 + sC_1 h_{ie})}{s^3 L C_1 C_2 C_3 h_{ie} + s^2 L C_2 C_3 + s h_{ie}[C_2C_3 + C_1C_2 + C_1C_3] + C_2 + C_3} \quad \text{—} ③$$

The base current $I_b = I \frac{X_{C_1}}{X_{C_1} + h_{ie}} = \frac{I}{1 + h_{ie} s C_1}$. —④

④ in ③ we get

$$1 = \frac{-h_{fe} C_3}{C_2 + C_3 - \omega^2 L C_2 C_3 + j\omega h_{ie}[(C_1C_2 + C_2C_3 + C_1C_3) - \omega^2 L C_1 C_2 C_3]} \quad \text{—} ⑤$$

The imaginary part of eq⑤ should be zero to fulfill Barkhausen Criterion

$$j\omega h_{ie}[(C_1C_2 + C_2C_3 + C_3C_1) - \omega^2 L C_1 C_2 C_3] = 0$$

$$\omega^2 = \frac{C_1C_2 + C_2C_3 + C_3C_1}{L C_1 C_2 C_3} = \frac{1}{L C_{eq}}$$

$$\omega = \frac{1}{\sqrt{L C_{eq}}} \Rightarrow \boxed{f = \frac{1}{2\pi\sqrt{L C_{eq}}}} \quad \text{—} ⑥$$

where $\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$

$C_3 << C_1 \& C_2 \Rightarrow C_{eq} = C_3$

$$\boxed{f = \frac{1}{2\pi\sqrt{LC_3}}} \quad \text{—} (6)$$

This is the required frequency for clapp Oscillator.

Frequency stability of oscillator

For an oscillator, the frequency of oscillation must remain constant. The analysis of dependance of oscillating frequency on various factors like stray capacitance temperature etc is called 'Frequency stability analysis'

"The measure of ability of an oscillator to maintain the desired frequency as precisely as possible for a long time is called Frequency stability of Oscillator."

In a transistorized colpitts (or) Hartley Oscillator, the base collector junction is reverse biased and there exist an internal capacitance which is dominant at high frequencies. This capacitance affects the value of capacitances in the tank circuit and hence the oscillator frequency.

Similarly the transistor parameters are temperature sensitive. As temperature changes, the oscillating frequency also changes and no longer remains stable. Thus practically, the circuit cannot provide stable frequency.

Factors Affecting Frequency stability

The factors which affect the frequency stability of oscillator are

i) Due to changes in temperature, the values of components of tank circuit get affected. So, the changes in values of inductors and capacitors due to changes in temperature is the main cause for change in frequency.

2) Due to changes in temperature, the parameters of active devices like BJT, FET get affected which in turn affect the frequency.

3) Variation in power supply

4) Changes in atmospheric conditions, aging & unstable transistor parameters.

5) Changes in the load connected, affect effective resistance of tank circuit.

6) Capacitive effect in transistor, & stray capacitances affect capacitance of T.C.

7) The variation of frequency with temperature is given by the factor denoted by 'S'.

$$S_{\omega,T} = \frac{\Delta\omega/\omega_r}{\Delta T/T_r} \text{ parts per million per } ^\circ C. \quad \text{—(1)}$$

where ω_r - desired frequency, $\Delta\omega$ = change in frequency
T_r - operating temperature, ΔT - change in temperature.

The frequency stability is defined as

$$S_\omega = \frac{d\theta}{d\omega} \quad \text{—(2)}$$ where $d\theta$ = phase shift introduced for small change in desired freq. fr.

larger the value of $\frac{d\theta}{d\omega}$, more stable is the oscillator.

The frequency stability can be improved by following modifications.

- enclosing the circuit in constant temperature chamber
- maintaining constant voltages by using zener diodes
- The load effect is reduced by coupling the oscillator to the load loosely or with the help of circuit having high input impedance and low output impedance

CRYSTAL OSCILLATOR

The crystals are either normally occuring or synthetically manufactured; exhibiting the "piezoelectric effect". The piezo electric effect means, under the influence of mechanical pressure, the voltages gets generated across opposite faces of crystal. If the mechanical pressure is applied in such a way that to force the crystal to vibrate, the a.c voltage generated across it. Conversely if the crystal is subjected to a.c voltage, it vibrates causing mechanical distortion in crystal shape. every crystal has its own resonating frequency depending on its cut.

Under the influence of mechanical stress, the crystal generates an electrical signal of constant frequency. The crystal has a greater stability in holding constant frequency.

A crystal oscillator is basically a tuned circuit oscillator using piezoelectrical crystal as its resonating tank circuit. The crystal oscillators are preferred when greater frequency stability is required. Hence crystals are used in watches, communication Txrs & Rxrs etc.

The main substances exhibiting piezoelectric effect are Quartz, Rochelle salt and tourmaline.

Rochelle salts have greater piezo electric stability. For a given a.c. voltage, they vibrate more. When it is vibrating, there are some internal frictional losses which are denoted by a resistance R. While the mass of crystal, which is indication of its inertia is represented by an inductance 'L'. In vibrating conditions, it is having some stiffness, represented by a capacitor C. The mounting capacitance is a shunt capacitance and hence the overall equivalent circuit of crystal can be as shown in fig:

RLC forms resonating circuit, the expression for resonating frequency f_r is

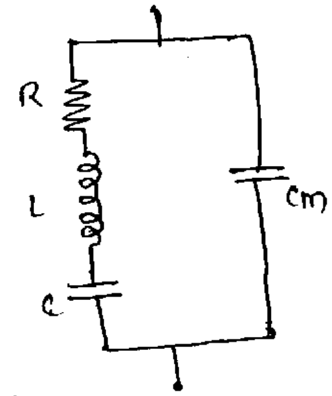


Fig: equivalent ckt of Crystal

$$f_r = \frac{1}{2\pi\sqrt{LC}}\sqrt{\frac{Q^2}{1+Q^2}} \quad \text{---} (1)$$

where Q is quality factor of crystal.

$$Q = \frac{\omega L}{R} \quad \text{---} (2)$$

The Q factor of crystal is very high, typically 20000 to 10^6. Hence $\sqrt{\frac{Q^2}{1+Q^2}}$ factor approaches to unity & we get the resonating frequency as $f = \frac{1}{2\pi\sqrt{LC}}$ — (3)

The crystal frequency is infact inversely proportional to the thickness of crystal

$f \propto \frac{1}{t}$ t is thickness.

So, we have very high frequencies, thickness of crystal should be very small. But it makes crystal mechanically weak & hence it may get damaged under the vibrations. Hence, practically crystal oscillator are used upto 200 to 300 kHz only.

The crystal has two resonating frequencies: series & parallel.

SERIES & PARALLEL RESONANCES

The series resonance occurs when reactance of series RLC are equal i.e. $X_L = X_C$. This is nothing but the impedance offered by this branch under resonant condition. is minimum which is resistance R. The series resonant frequency is given by

$$\boxed{f_s = \frac{1}{2\pi\sqrt{LC}} = f_r} \quad \text{---} (4)$$

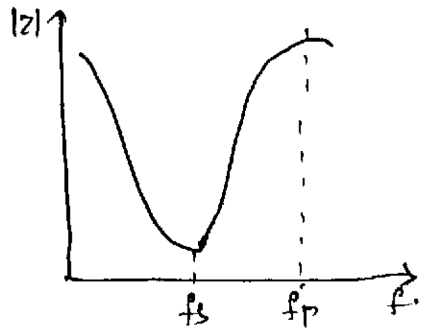
The parallel resonance occurs when reactances of series resonant leg is equal to the reactance of mounting capacitance Cm. Under this condition, the impedance offered by crystal to external circuit is maximum. It is also known as "anti resistance".

The parallel resonant frequency is given by

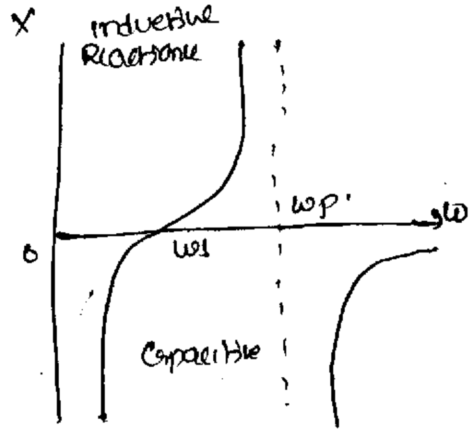
$$f_p = \frac{1}{2\pi\sqrt{L C_{eq}}} \quad \text{——} (5) \qquad C_{eq} = \frac{C C_m}{C + C_m} \quad \text{——} (6)$$

When the crystal capacitance C is much smaller than Cm, then the behaviour of crystal impedance versus frequency is shown in fig.

Generally value of f_s & f_p are close to each other & practically it can have only one resonating frequency for the crystal.



Crystal impedance Vs frequency



reactance Vs frequency.

The higher value of Q is main advantage of crystal. Due to high Q of resonant circuit, it provides very good frequency stability. If we neglect resistance R, the impedance of crystal is a reactance jX which depends on frequency as

$$jX = \frac{-j}{\omega C_m} \frac{\omega^2 - \omega_s^2}{\omega^2 - \omega_p^2}$$

ω_s - Series Resonant frequency
ω_p - parallel Resonant frequency.

The sketch of reactance against frequency was shown in fig (6).

The oscillating frequency lies between ω_s & ω_p.

Crystal stability:

The frequency of crystal tends to change slightly with time due to temperature, again etc.

Temperature stability:

It is defined as change in frequency per degree change in temperature (Hz or MHz/°C).

For one degree change (1°C) in temperature, frequency changes by 10 to 12 Hz in 1MHz. This is negligiably small & it is considered to be constant for all practical purposes. But to keep freq. constant, the crystal is kept in a box where temperature is maintained at constant.

Long term stability:

It is basically defined due to aging of crystal. Aging rates are 2×10^{-8} per year for quartz crystal which is negligable.

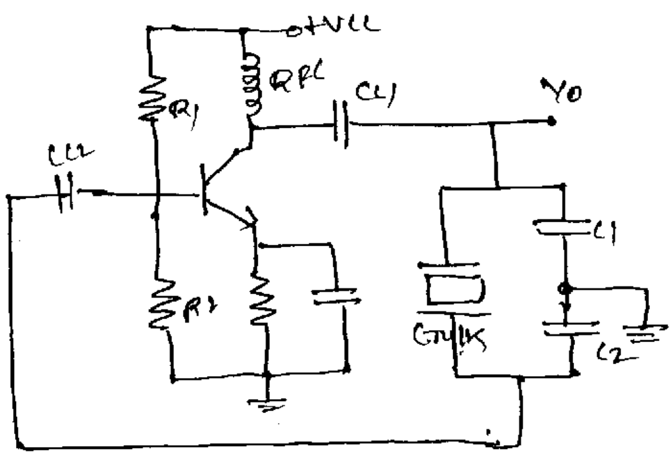
Short term stability:

In a quartz crystal, the frequency drift (change with time is typically less than one part in 10^6 i.e., 0.00001% per year which is very small value.

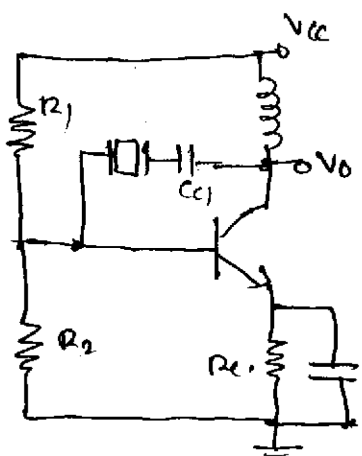
Overall, crystal has good frequency stability. Hence it is used in computers, counters, basic timing devices in electronic wrist watches etc.

PIERCE CRYSTAL OSCILLATOR

The Colpitts oscillator can be modified by using crystal to behave as inductor. The circuit is called pierce crystal oscillator.



Pierce crystal oscillator



modified pierce crystal osc

The crystal behaves as an inductor for a frequency slightly higher than series resonant frequency fs. The two capacitances C_1 & C_2 required in tank circuit along with an inductor is used.

The basic working principle of pierce cryst osc is same as that of colpitts osc as shown in fig.

The resulting circuit freq. is set by series resonant freq. of crystal. Change in supply voltages, temperature, transistor parameters have no effect on circuit operating conditions & hence good freq stability is obtained.

The oscillator can be modified using internal capacitance of transistor used instead of C_1 & C_2 as shown in fig(b)

Miller crystal oscillator.

Amplitude limiting (stability)

Need: The oscillator o/p amplitude, if not limited, attains extreme levels of saturation. This can cause distortion in o/p waveform, due to either clipping of some part of waveform. Or it drives the amplifier into saturation. The circuits used in oscillators for this purpose are called amplitude limiting circuits