54 EC 402 ELECTRONICS CIRCUITS II - SYLLABUS

1. FEEDBACK AMPLIFIRES

Types of feedback, Effect of feedback on noise, Distortion, Gain, Input and input impedance of the amplifier, Analysis of Voltage and Current feedback amplifiers.

2. OSCILLATORS

Negative feedback oscillator, Barhausen Criterion for oscillation in feedback oscillators, Mechanism for start of oscillation and stabilization of amplitude Analysis of RC oscillators using Cascade connection of Low Pass and High Pass filters, Wein phase shift and twin - T network, analysis of LC oscillators Colpitts, Hartley, Clapp, Franklin, Armstrong and Miller oscillators, Frequency rage of RC and LC oscillators, Quartz crystal construction, Electrical equivalent circuit of crystal oscillator circuits, Use of Logic Gates as linear amplifiers, Oscillator and clock generator circuits using Logic Gate circuits.

3. TUNED AMPLIFIERS

Coil losses, Unloaded and loaded Q of tank circuits, Analysis of single tuned amplifier, Double tuned, Stagger tuned amplifiers, Instability of tuned amplifiers, Stabilization technique, Narrow band neutralization using coil, Broad banding using Hazeltine neutralization, Class C tuned amplifiers and their applications, Efficiency of class C tuned amplifiers.

4. MULTIVIBRATOR CIRCUITS

Collector coupled and complimentary collector coupled stabled multivibrators, Emitter coupled astable multivibrators, Monostable and bistable multivibrators using similar and complimentary transistors, Trigger methods, Storage delay and calculation of switching times, Speed of capacitors, Schmitt trigger circuits.

5. BLOCK OSCILLATORS AND TIMEBASE GENERATORS

Monostable and astable blocking oscillators using Emitter-based timing, Frequency control using core saturation, Pushpull operation of astable blocking oscillators i.e. inverter, pulse transformers, RC and RL wave shaping circuits, Bootstrap and Miller saw tooth generators, current time base generators.

6. TUTORIALS

Text Books;

1. David A Bell 2002, “Solid state Pulse Circuits” - Prentice Hall of India
2. John D Ryder, 1999, “Electronic Fundamental and Applications – integrated and Discrete Systems” - Prentice Hall of India

Reference Books;

Millman J and Taub H 2001 “Pulse Digital and Switching wave form” - McGraw Hill

UNIT 1. FEEDBACK AMPLIFIERS

L – 1. Types of Feedback, Effect of Feedback on Gain, Noise and Distortion

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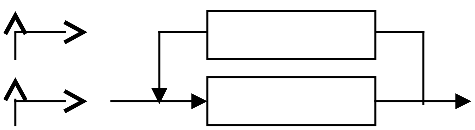
Types of feedback,

In the process of feedback, a part of output is sampled and fed back to the input of the amplifier; therefore, at input we have two signals;

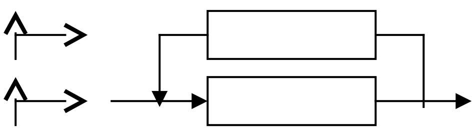
Input signal,

Part of the output that is fed back to the input

Both these signals may be in phase or out of phase.



Positive feedback is when input signal and part of the output signal are in phase.



Negative feedback is when signal and part of the output signal are out of phase.

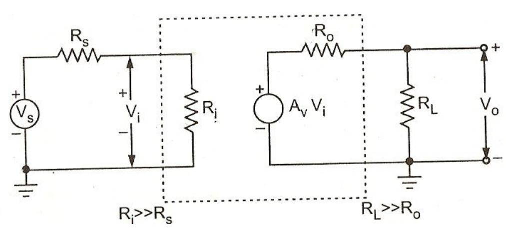
Use of positive feedback results in oscillations and therefore it is not used in amplifiers.

Classification of amplifiers

a. Voltage amplifier;

Is an amplifier with input resistance R_i large compared with the source resistance R_s *such that* $V_i \approx V_s$, and the external load resistance R_L large compared with the output resistance R_o *then* $Vo \approx A_v V_i \approx A_v V_s$

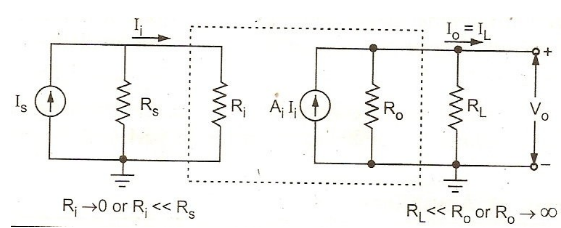
A voltage amplifier circuit provides a voltage output proportional to the voltage input and the proportionality factor is independent of the magnitude of the source and load resistance.



b. Current amplifier;

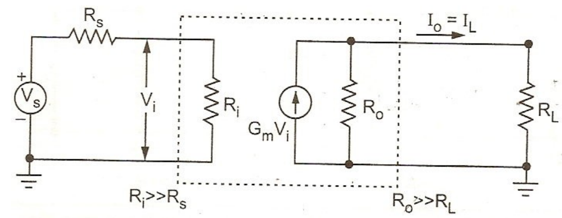
Is an amplifier with input resistance $R_i \to 0$, *such that* $I_i \approx I_s$ and the amplifier output resistance $R_o \to \infty$ *then* $I_L = A_i I_i \approx A_v V_s$

A current amplifier provides a current output proportional to the signal current and the proportionality factor is independent of the source and load resistance.



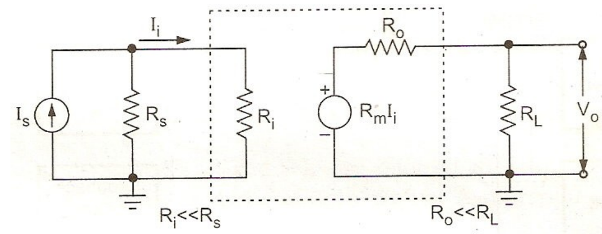
c. Transconductance amplifier;

An output current is proportional to the input signal voltage and the proportionality factor is independent of the magnitudes of the source and load resistance.



d. <u>Transresistance amplifier;</u>

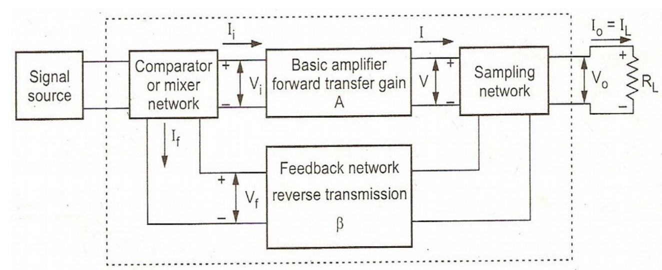
The output voltage is proportional to the input signal current and the proportionality factor is independent of the source and load resistance.



<u>Question:</u> Classify amplifiers in accordance to how input parameters are related to the output parameters.

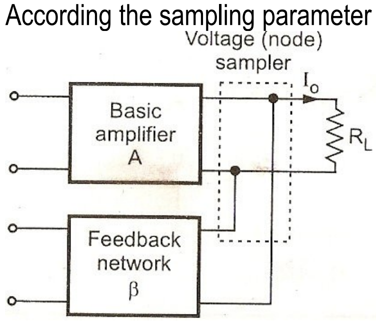
Feedback concepts;

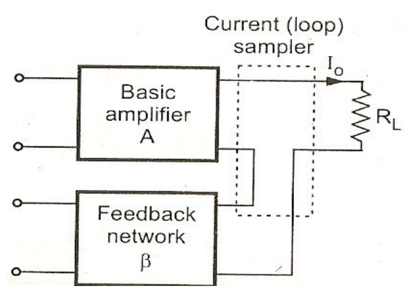
According to the type of an amplifier circuit, the output voltage or current is sampled by means of a suitable sampling network and applied to the input through a feedback network.



<u>Sampling Network:</u>

According the sampling parameter (voltage or current) there are two ways to sample the output;





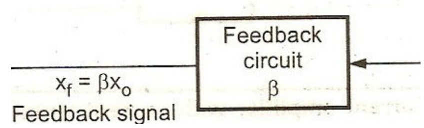
(a) Voltage or node sampling (b) Current or loop sampling

<u>Feedback network:</u>

A feedback network provides a reduced portion of the output as feedback signal to the input of the mixer network. The feedback signal X_f is given by $X_f = \beta X_o$

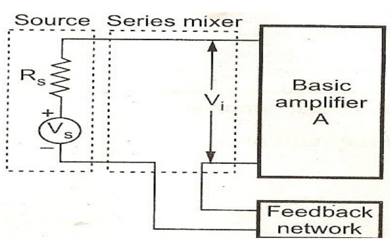
Where β - feedback factor (feedback ratio) $0 < \beta < 1$

X_o - Output signal

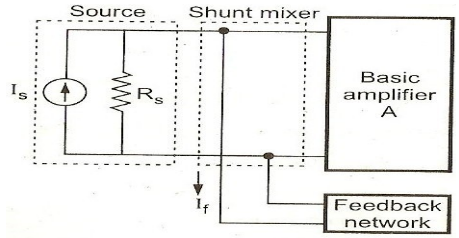


<u>Mixer network</u>

According the parameter (voltage or current) to be mixed, there are two ways to mix the input;



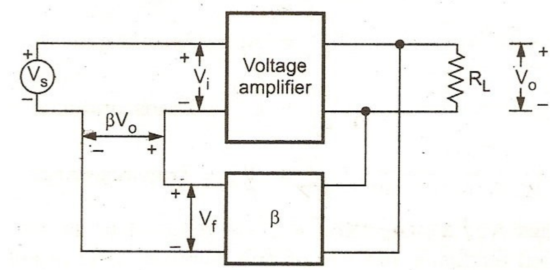
(a) Series mixing



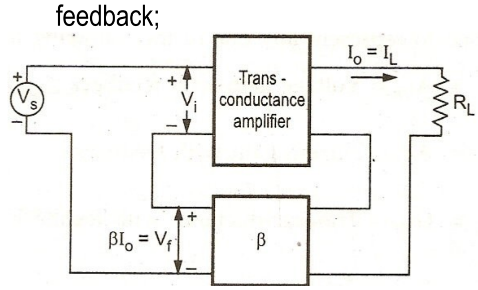
(b) Shunt mixing

Ways of introducing negative feedback in amplifiers

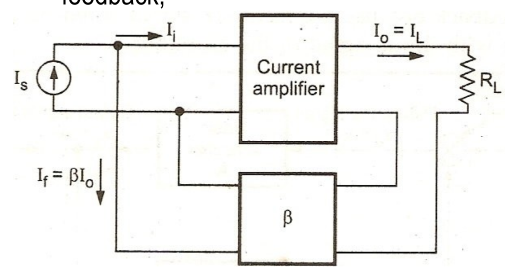
a. Voltage amplifier with voltage series feedback;



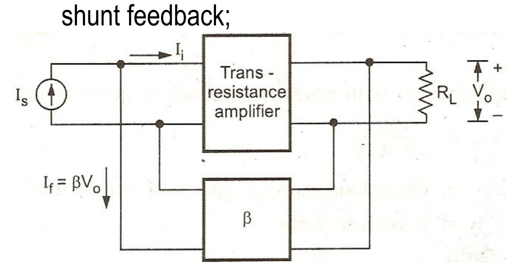
b. Transconductance amplifier with current series feedback;



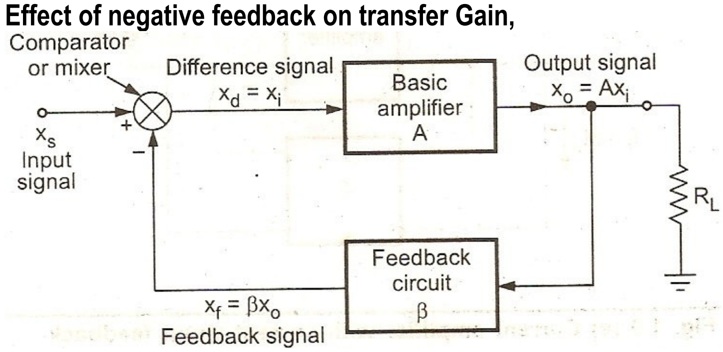
c. Current amplifier with current-shunt feedback;



d. Transresistance amplifier with voltage shunt feedback;



Effect of negative feedback on transfer Gain,



Let A represent transfer gain of amplifier without feedback where

$$A = \frac{X_o}{X_i}$$

A_f represent transfer gain of amplifier with feedback $A_f = \frac{X_o}{X_s}$

Where X_o – *Output voltage or current*;

X_i – *input voltage or input current*;

X_s – *source voltage or source current*;

As it is negative feedback $X_i = X_s + (-X_f)$; therefore $A_f = \frac{X_o}{X_s} = \frac{X_o}{X_i + X_f}$

Where X_f -feedback voltage or feedback current

Dividing by X_i to numerator and denominator; introducing $A = \frac{X_o}{X_i}$ and $\beta = \frac{X_f}{X_o}$ - feedback factor

$$\Rightarrow A_f = \frac{\frac{X_o}{X_i}}{\frac{X_i + X_f}{X_i}} = \frac{A}{1+\frac{X_f}{X_i}} = \frac{A}{1+\left(\frac{X_f}{X_o}\right)\left(\frac{X_o}{X_i}\right)} = \frac{A}{1+\beta A}$$

Hence $A_f = \frac{A}{(1+\beta A)}; and\ A = A_f(1+\beta A)$

Gain without feedback is always greater than gain with feedback and it decreases with increase in feedback factor.

Stability of Gain

The transfer gain of an amplifier depends on factors like temperature, operating point etc, that is, it is not constant. The introduction of negative feedback in amplifiers reduces the lack of stability.

Consider $A_f = \frac{A}{(1+\beta A)}$

Differentiate both sides with respect to A $\frac{\partial A_f}{\partial A} = \frac{(1+\beta A)1-\beta A}{(1+\beta A)^2} = \frac{1}{(1+\beta A)^2} \therefore \partial A_f = \frac{\partial A}{(1+\beta A)^2}$

Divide both sides by $A_f \Rightarrow \frac{\partial A_f}{A_f} = \frac{\partial A}{A_f(1+\beta A)^2}$ $\quad Since\ A_f = \frac{A}{(1+\beta A)}$

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

Where $\frac{\partial A_f}{A_f} - fractional\ change\ in\ amplification\ with\ feedback$; - sensitivity of transfer gain

$\frac{\partial A}{A} - fractional\ change\ in\ amplification\ without\ feedback$

Hence, the change in feedback with gain is less than the change in gain without feedback by a factor $1+\beta A$

If $\beta A >> 1$ then $A_f = \frac{A}{(1+\beta A)} = \frac{A}{\beta A} = \frac{1}{\beta}$ -the gain is dependent only on the feedback network.

<u>**Hence**</u>

- For voltage series feedback $A_{Vf} = \frac{1}{\beta}$
  Voltage gain is stabilized
- For current shunt feedback $A_{If} = \frac{1}{\beta}$
  Current gain is stabilized
- For current series feedback $G_{Mf} = \frac{1}{\beta}$
  Transconductance is stabilized
- For voltage shunt feedback $R_{Mf} = \frac{1}{\beta}$
  Transresistance is stabilized

L – 2 Frequency Response and Bandwidth

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From $A_f = \dfrac{A}{(1+\beta A)} \Rightarrow A_{fmid} = \dfrac{A_{mid}}{(1+\beta A_{mid})};\ A_{fLow} = \dfrac{A_{Low}}{(1+\beta A_{Low})};\ A_{fHigh} = \dfrac{A_{High}}{(1+\beta A_{High})}$

Analyzing the effect of negative feedback on the lower cut-off and upper cut-off frequencies of the amplifier;

- Lower cut-off frequency

The relation between gain at low frequency and gain at mid frequency is given as;

$$\frac{A_{low}}{A_{mid}} = \frac{1}{1-j\left(\frac{f_L}{f}\right)} \quad \therefore \quad A_{low} = \frac{A_{mid}}{1-j\left(\frac{f_L}{f}\right)}$$

Substituting the value A_{Low} we get

$$A_{flow} = \frac{\dfrac{A_{mid}}{1-j\left(\frac{f_L}{f}\right)}}{1+\beta\left(\dfrac{A_{mid}}{1-j\left(\frac{f_L}{f}\right)}\right)} = \frac{A_{mid}}{1-j\left(\frac{f_L}{f}\right)+A_{mid}\beta} = \frac{A_{mid}}{(1+A_{mid}\beta)-j\left(\frac{f_L}{f}\right)}$$

Dividing numerator and denominator by $(1+A_{mid}\beta)$

$$A_{flow} = \frac{\dfrac{A_{mid}}{(1+A_{mid}\beta)}}{1-j\left(\dfrac{\frac{f_L}{(1+A_{mid}\beta)}}{f}\right)} = \frac{A_{f\,mid}}{1-j\left(\dfrac{\frac{f_L}{(1+A_{mid}\beta)}}{f}\right)} \quad \because \quad A_{f\,mid} = \frac{A_{mid}}{(1+A_{mid}\beta)}$$

$\therefore \dfrac{A_{f\,Low}}{A_{f\,mid}} = \dfrac{1}{1-j\left(\frac{f_{Lf}}{f}\right)}$ Where the lower cut-off frequency with feedback $f_{Lf} = \dfrac{f_L}{(1+A_{mid}\beta)}$;

Hence, the lower cut-off frequency with feedback is less than the lower cut-off frequency without feedback by factor $1+A_{mid}\beta$. Therefore, by introducing negative feedback the low frequency response of the amplifier is improved.

Question:

Show that introducing negative feedback the low frequency response of the amplifier is improved. i.e. the lower cut-off frequency with feedback is less than the lower cut-off frequency without feedback.

- Upper cut-off frequency

The relation between gain at high frequency and gain at mid frequency is given as;

$$\frac{A_{high}}{A_{mid}} = \frac{1}{1 - j\left(\frac{f}{f_H}\right)} \quad \therefore \quad A_{high} = \frac{A_{mid}}{1 - j\left(\frac{f}{f_H}\right)}$$

Substituting the value A_{high} we get

$$A_{fhigh} = \frac{\frac{A_{mid}}{1 - j\left(\frac{f}{f_H}\right)}}{1 + \beta\left(\frac{A_{mid}}{1 - j\left(\frac{f}{f_H}\right)}\right)} = \frac{A_{mid}}{1 - j\left(\frac{f}{f_H}\right) + A_{mid}\beta} = \frac{A_{mid}}{(1 + A_{mid}\beta) - j\left(\frac{f}{f_H}\right)}$$

Dividing numerator and denominator by $(1 + A_{mid}\beta)$

$$A_{fhigh} = \frac{\frac{A_{mid}}{(1 + A_{mid}\beta)}}{1 - j\left(\frac{f}{(1 + A_{mid}\beta)f_H}\right)} = \frac{A_{f\,mid}}{1 - j\left(\frac{f}{(1 + A_{mid}\beta)f_H}\right)} \quad \because A_{f\,mid} = \frac{A_{mid}}{(1 + A_{mid}\beta)}$$

$\therefore \dfrac{A_{f\,high}}{A_{f\,mid}} = \dfrac{1}{1 - j\left(\dfrac{f}{f_{Hf}}\right)}$ Where $f_{Hf} = (1 + A_{mid}\beta)f_H$; higher cut-off frequency with feedback.

Hence, the higher cut-off frequency with feedback is greater than the higher cut-off frequency without feedback by factor $1 + A_{mid}\beta$. Therefore, by introducing negative feedback the high frequency response of the amplifier is improved.

<u>Questions:</u>

(i) *Show that introducing negative feedback the high frequency response of the amplifier is improved. i.e. the higher cut-off frequency with feedback is greater than the higher cut-off frequency* without feedback.

(ii) *An amplifier has mid-band voltage gain A_{mid} of 1000 with $f_L = 50Hz$ and $f_H = 50kHz$, if 5% feedback is applied then calculate (a) gain with feedback $A_{f\,mid}$, (b) upper cut-off frequency with feedback $f_{H\,f}$ and (c) lower cut-off frequency with feedback $f_{L\,f}$*

<u>Solution:</u>

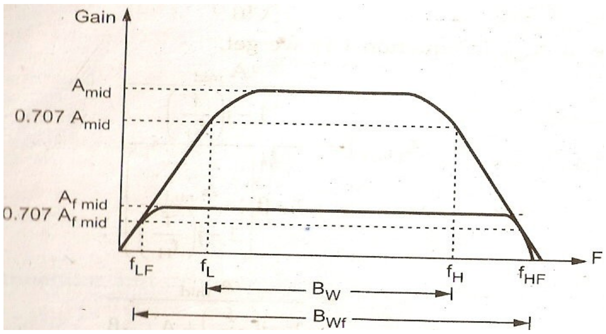
Given $\beta = \frac{5}{100} = 0.05,\ f_L = 50\ Hz,\ f_H = 50000\ Hz;\ A_{mid} = 1000$

(a) Gain with feedback $A_{f\,mid} = \dfrac{A_{mid}}{(1 + A_{mid}\beta)} = \dfrac{1000}{1 + 1000 \times 0.05} = 19.6$

(b) $f_{Lf} = \dfrac{f_L}{(1 + A_{mid}\beta)} = \dfrac{50}{1 + 1000 \times 0.05} = 0.98Hz$

(c) $f_{Hf} = f_H \times (1 + A_{mid}\beta) = 50000(1 + 1000 \times 0.05) = 2.55MHz$

- Bandwidth of the amplifier with feedback



Bandwidth of the amplifier =
= (Upper cut-off– Lower cut-off) frequency,
$\therefore BW = f_H - f_L$

Therefore Bandwidth of the amplifier with feedback is given as;

$$BW_f = f_{Hf} - f_{Lf} = (1 + A_{mid}\beta) f_H - \frac{f_L}{(1 + A_{mid}\beta)}$$ Since $(f_{Hf} - f_{Lf}) > (f_H - f_L)$

The bandwidth of an amplifier with feedback is greater than bandwidth of an amplifier without feedback.

Question
(iii) (d) Calculate the bandwidth with feedback BW_f in the above example,
Solution

$$(d)\quad BW_f = f_{Hf} - f_{Lf} = (2550000 - 0.98)\ Hz = 2.549902\ MHz$$

Effect of feedback on Distortion,

If the feedback network does not contain reactive elements, the overall gain is not a function of frequency. Under such conditions, frequency and phase distortion is substantially reduced.

If β is made up of reactive components, the reactance of these components will change with frequency, changing the β . As a result, gain will also change with frequency.
This factor is used in tuned amplifiers where the feedback network is designed in such a way that at tuned frequencies $\beta \to 0$ and at other frequencies $\beta \to \infty$, that is, an amplifier provides high gain for signals at tuned frequency and rejects all other frequencies.

Effect of feedback on noise and nonlinear distortion,

Signal feedback reduces the amount of noise signal and non-linear distortion. Both input noise and resulting non-linear distortion are reduced by the same factor $(1 + \beta A)$ as the gain.

L – 3. Input Impedance of the Amplifier,

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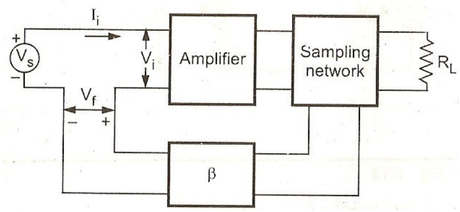
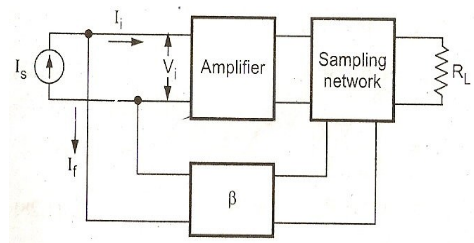


If the feedback signal is added to the input in series with the applied voltage, it increases the input resistance. Since the feedback voltage V_f opposes V_s, the input current I_i is less than it would be if V_f were absent.

Hence input resistance with feedback $R_{if} = \frac{V_s}{I_i}$ is greater than input resistance without feedback

If the feedback signal is added to the input in shunt with the applied voltage, it decreases the input resistance. Since $I_s = I_i + I_f$, the current I_s drawn from the signal source is increased over what it would be if there was no feedback current.

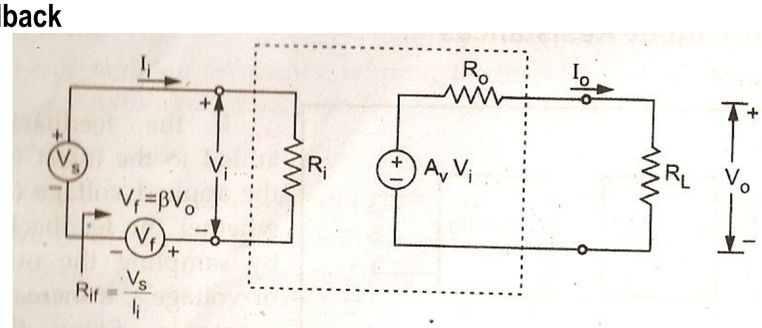


Hence input resistance with feedback $R_{if} = \frac{V_s}{I_s}$ is decreased.

Effect of negative feedback on input resistance in different feedback topologies;

- **Voltage series feedback**

In the voltage series feedback topology, the input and output circuits being replaced by Thevenin's equivalent circuit;



The input resistance with feedback $R_{if} = \frac{V_s}{I_i}$

Applying KVL to the input side; $V_s - I_i R_i - V_f = 0$

$V_s = I_i R_i + V_f = I_i R_i + \beta V_o$; The output voltage $V_o = \frac{A_v V_i R_L}{R_o + R_l} = A_V I_i R_i = A_V V_i$;

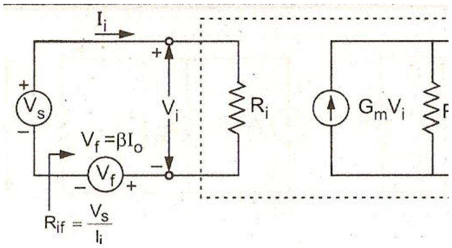
Where $A_V = \frac{V_o}{V_i} = \frac{A_v R_L}{R_o + R_L}$ - voltage gain without feedback taking the load into account;

Substituting $V_o \Rightarrow V_s = I_i R_i + \beta A_V I_i R_i$; hence $\frac{V_s}{I_i} = R_i + \beta A_v R_i$,

Input resistance with feedback $\therefore R_{if} = R_i(1 + \beta A_v)$

- **Current series feedback**

In the voltage series feedback topology, the amplifier input circuit is being replaced by Thevenin's equivalent circuit and output circuit being replaced by Norton's equivalent circuit;



The input resistance with feedback $R_{if} = \dfrac{V_s}{I_i}$

Applying KVL to the input side; $V_s - I_i R_i - V_f = 0$

$V_s = I_i R_i + V_f = I_i R_i + \beta I_o$

The output current $I_o = \dfrac{G_m V_i R_o}{R_o + R_L} = G_M V_i ;\, where\ G_M = \dfrac{G_m R_o}{R_o + R_L}$

Substituting I_o we get

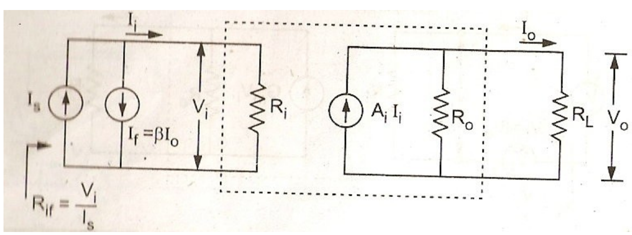
$V_s = I_i R_i + \beta A_v G_M V_i = I_i R_i + \beta G_M I_i R_i$; since $V_i = I_i R_i$, therefore $V_s = I_i R_i (1 + \beta G_M)$

Dividing both sides by I_i , we get $\dfrac{V_s}{I_i} = R_i (1 + \beta G_M)$,

Input resistance with feedback $\therefore R_{if} = R_i (1 + \beta G_M)$

- **Current shunt feedback**

In the current shunt feedback topology, the input and output circuits being replaced by Norton's equivalent circuit;



Applying KCL to the input node; $I_s = I_i + I_f = I_i + \beta I_o$

The output current $I_o = \dfrac{A_i I_i R_o}{R_o + R_L} = A_I I_i ;\, where\ A_I = \dfrac{A_I R_o}{R_o + R_L}$

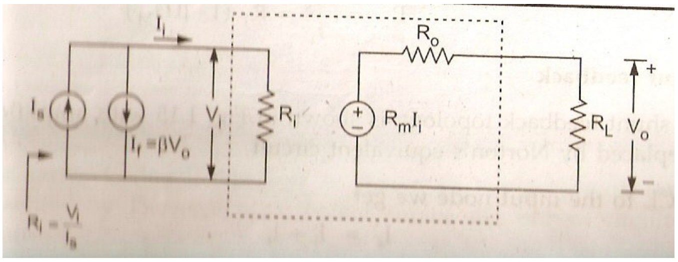
Substituting I_o we get $I_s = I_i + \beta A_I V_i = I_i (1 + \beta A_I)$;

The input resistance with feedback $R_{if} = \dfrac{V_i}{I_s} = \dfrac{V_i}{I_i (1 + \beta A_I)}$ since $R_i = \dfrac{V_i}{I_i}$

$$R_{if} = \frac{R_i}{(1 + \beta A_I)}$$

- **Voltage shunt feedback**

In the voltage shunt feedback topology, the input circuit is being replaced by Norton's equivalent circuit and output circuit is being replaced by Thevenin's equivalent;



Applying KCL to the input node; $I_s = I_i + I_f = I_i + \beta V_o$

The value of V_o is given as; $V_o = \frac{R_m I_i R_o}{R_o + R_L} = R_M I_i; \, where \, R_M = \frac{R_m R_o}{R_o + R_L}$

Where R_m - open circuit transresistance without feedback

R_M - Transresistance without feedback taking the load R_L into account.

Substituting the value of V_o we get $I_s = I_i + \beta R_M I_i = I_i(1 + \beta R_M)$

The input resistance with feedback $R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i(1+\beta R_M)}$ since $R_i = \frac{V_i}{I_i}$

$$R_{if} = \frac{R_i}{(1+\beta R_M)}$$

L – 4. Output Impedance of the Amplifier

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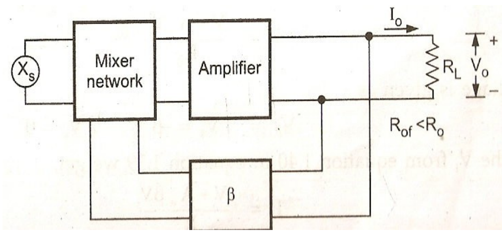
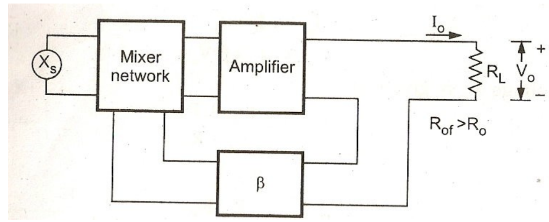
The negative feedback which samples the output voltage tends to decrease the output resistance.



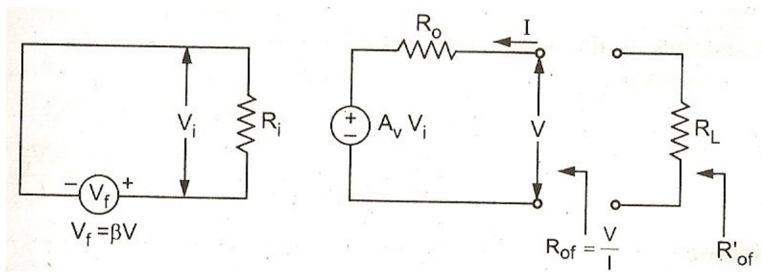
The negative feedback which samples the output voltage tends to increase the output resistance.



<u>Effect of negative feedback on output resistance in different feedback topologies;</u>

- **Voltage series feedback**

The output resistance is measured by shorting the input source $V_s = 0$ and looking into the output terminals with R_L disconnected.



Applying KVL to the output side $A_v V_i + IR_o - V = 0$,

Therefore $I = \frac{V - A_v V_i}{R_o}$

The input voltage is given by, $V_i = -V_f = -\beta V$ since $V_s = 0$

Substituting V_i $I = \frac{V + A_v \beta V}{R_o} = \frac{V(1+\beta A_v)}{R_o}$ Therefore $R_{of} = \frac{V}{I} = \frac{R_o}{(1+\beta A_v)}$;

Where A_v - open loop voltage with R_L disconnected

Let $R'_{of} = R_{of} \| R_L$ then

$$R'_{of} = \frac{R_{of} \times R_L}{R_{of} + R_L} = \frac{\frac{R_o}{(1+\beta A_v)} \times R_L}{\frac{R_o}{(1+\beta A_v)} + R_L} = \frac{R_o R_L}{R_o + R_L(1+\beta A_v)} = \frac{R_o R_L}{R_o + R_L + R_L \beta A_v}$$

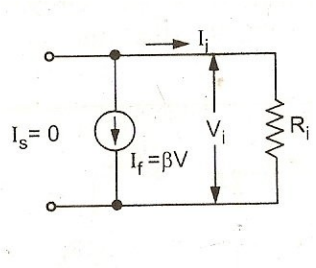
Dividing numerator and denominator by $R_o + R_L$

$$R'_{of} = \frac{\frac{R_o R_L}{R_o + R_L}}{1 + \frac{R_L \beta A_v}{R_o + R_L}} = \frac{R'_L}{1 + \beta A_V} \quad Therefore R'_o = \frac{R_o R_L}{R_o + R_L} \quad and \quad A_V = \frac{R_L A_v}{R_o + R_L}$$

Where A_V - Open loop gain taking into account R_L

- **Voltage shunt feedback**

The output resistance is measured by shorting the input source $V_s = 0$ and looking into the output terminals with R_L disconnected.



Applying KVL to the output side $R_m I_i + IR_o - V = 0$, Therefore $I = \frac{V - R_m I_i}{R_o}$

The input current is given by, $I_i = -I_f = -\beta V$

Substituting I_i $I = \frac{V + R_m \beta V}{R_o} = \frac{V(1 + R_m A_v)}{R_o}$

Therefore $R_{of} = \frac{V}{I} = \frac{R_o}{(1 + R_m \beta)}$; where R_m - open loop transresistance with R_L disconnected

Let $R'_{of} = R_{of} \| R_L$ then

$$R'_{of} = \frac{R_{of} \times R_L}{R_{of} + R_L} = \frac{\frac{R_o}{(1 + R_m \beta)} \times R_L}{\frac{R_o}{(1 + \beta R_m)} + R_L} = \frac{R_o R_L}{R_o + R_L(1 + \beta R_m)} = \frac{R_o R_L}{R_o + R_L + R_L \beta R_m}$$

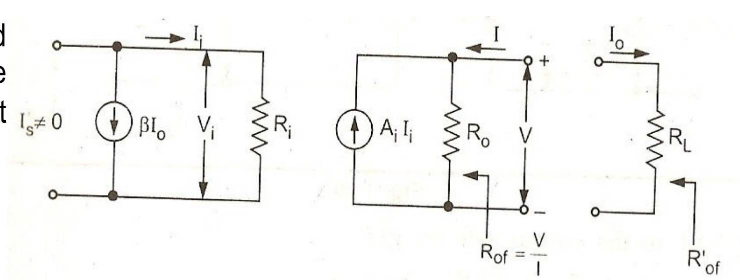
Dividing numerator and denominator by $R_o + R_L$

$$R'_{of} = \frac{\frac{R_o R_L}{R_o + R_L}}{1 + \frac{R_L \beta R_m}{R_o + R_L}} = \frac{R'_o}{1 + \beta R_M} \quad Therefore \;\; R'_o = \frac{R_o R_L}{R_o + R_L} \;\; and \;\; R_M = \frac{R_L R_m}{R_o + R_L}$$

Where R_M - open loop transresistance taking into account R_L .

- **Current shunt feedback**

The output resistance is measured by open circuiting the input source $I_s = 0$ and looking into the output terminals with R_L disconnected.



Applying KCL to the output node $I = \frac{V}{R_o} - A_i I_i$,

Input current $I_i = -I_f = -\beta I_o \quad \because I_s = 0$
$I_i = \beta I \quad \because I = -I_o$

Substituting I_i $I = \frac{V}{R_o} - A_i \beta I; \; Therefore \;\; I(1 + A_i \beta) = \frac{V}{R_o}$

$$\Rightarrow R_{of} = \frac{V}{I} = R_o(1 + A_i\beta)$$

Where A_i - open loop current gain with R_L disconnected

Let $R'_{of} = R_{of} \| R_L$ then $R'_{of} = \frac{R_{of} \times R_L}{R_{of} + R_L} = \frac{R_o(1+\beta A_i)R_L}{R_o(1+\beta A_i) + R_L} = \frac{R_o R_L(1+\beta A_i)}{R_o + R_L + \beta A_i R_o}$

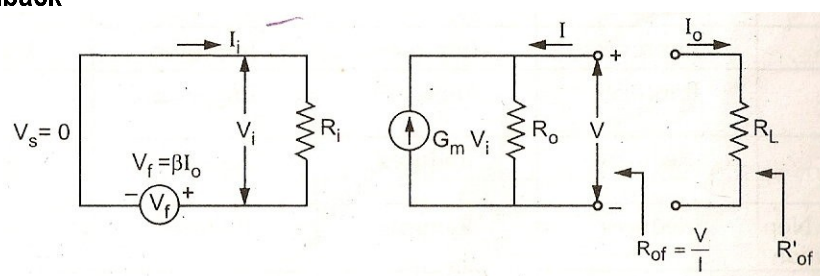
Dividing numerator and denominator by $R_o + R_L$

$$R'_{of} = \frac{\frac{R_o R_L(1+\beta A_i)}{R_o + R_L}}{1 + \frac{R_o \beta A_i}{R_o + R_L}} = \frac{R'_o(1+\beta A_i)}{(1+\beta A_i)} \quad Therefore \quad R'_o = \frac{R_o R_L}{R_o + R_L} \quad and \quad A_I = \frac{R_L A_i}{R_o + R_L}$$

Where A_I is the open loop current gain taking into account R_L

- **Current series feedback**

The output resistance is measured by shorting the input source $V_s = 0$ and looking into the output terminals with R_L disconnected.



Applying KCL to the output node $I = \frac{V}{R_o} - G_m I_i$,

Input voltage $V_i = -V_f = -\beta I_o$
$I_i = \beta I \qquad \because I_o = -I$

Substituting V_i we get $I = \frac{V}{R_o} - G_m \beta I;\ Therefore\ I(1 + G_m\beta) = \frac{V}{R_o}$

$\Rightarrow R_{of} = \frac{V}{I} = R_o(1 + G_m\beta)$ Where G_m - open loop transconductance with R_L disconnected.

Let $R'_{of} = R_{of} \| R_L$ then $R'_{of} = \frac{R_{of} \times R_L}{R_{of} + R_L} = \frac{R_o(1+\beta G_m)R_L}{R_o(1+\beta G_m) + R_L} = \frac{R_o R_L(1+\beta G_m)}{R_o + R_L + \beta G_m R_o}$

Dividing numerator and denominator by $R_o + R_L$

$$R'_{of} = \frac{\frac{R_o R_L(1+\beta G_m)}{R_o + R_L}}{1 + \frac{R_o \beta G_m}{R_o + R_L}} = \frac{R'_o(1+\beta G_m)}{(1+\beta G_M)} \quad Therefore \quad R'_o = \frac{R_o R_L}{R_o + R_L} \quad and \quad G_M = \frac{R_o G_m}{R_o + R_L}$$

Where G_M is the open loop current gain taking into account R_L

Summary of the effect of negative feedback on an amplifier

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	$R_{mf} = \frac{R_m}{(1+\beta R_{mf})}$ Decreases
Stability	Improves	Improves	Improves	Improves
Frequenc y response	Improves	Improves	Improves	Improves
Frequenc y distortion	Reduced	Reduced	Reduced	Reduced
Noise and Non-linear distortion	Reduced	Reduced	Reduced	Reduced
Input resistanc e	$\therefore R_{if} = R_i(1+\beta A_v)$ Increases	$\therefore R_{if} = R_i(1+\beta G_M)$ Increases	$R_{if} = \frac{R_i}{(1+\beta A_I)}$ Decreases	$R_{if} = \frac{R_i}{(1+\beta R_M)}$ Decreases
Output resistanc e	$R_{of} = \frac{V}{I} = \frac{R_o}{(1+\beta A_v)}$ Decreases	$R_{of} = R_o(1+G_M\beta)$ Increases	$R_{of} = R_o(1+A_i\beta)$ Increases	$R_{of} = \frac{R_o}{(1+R_m\beta)}$ Decreases

L – 5. Methodology of Feedback Amplifier Analysis

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Step 1: Identify type of feedback;

(a) To find the type of sampling network;

- By shorting the output i.e. $V_o = 0$; if feedback signal $X_f = 0$ - Voltage sampling,
- By opening the output loop, i.e. $I_o = 0$; if feedback signal $X_f = 0$ - current sampling,

(b) To find the type of mixing network;

- If the feedback signal is subtracted from the externally supplied signal as a voltage in the input loop – Series mixing.
- If the feedback signal is subtracted from the externally supplied signal as a current in the input loop – Shunt mixing.

Step 2: To find the input current;

(a) For Voltage sampling make $V_o = 0$; by shorting the output,

(b) For current sampling make $I_0 = 0$, by opening the output loop,

Step 3: To find the output circuit;

(a) For series mixing, make $I_i = 0$ by opening the input loop,

(b) For shunt mixing, make $V_i = 0$ by shorting the input,

Step 4 Replace each active device by its h-parameter model.

Step 5: find the open loop gain (gain without feedback), A of the amplifier,

Step 6: Indicate X_f and X_o on the circuit and evaluate $\beta = X_f \;\; X_o$

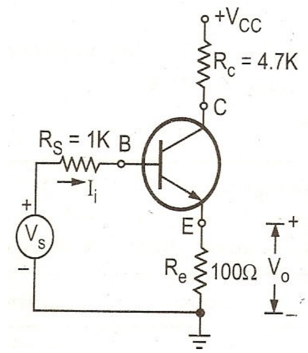
Step 7: From A and β, find $A_f,\ R_{if},\ R_{of}\ \ and\ \ R'_{of}$

Characteristic	Topology			
	Voltage series	Current series	Current shunt	Voltage shunt
Sampling signal X_o	Voltage	Voltage	Current	Current
Mixing signal	Voltage	Current	Current	Voltage
To find input loop, set	$V_o = 0$	$I_0 = 0$	$I_0 = 0$	$V_o = 0$
To find output loop, set	$I_i = 0$	$I_i = 0$	$V_i = 0$	$V_i = 0$
Single source	Thevenin	Thevenin	Norton	Norton
$\beta = \frac{X_f}{X_o}$	$\frac{V_f}{V_o}$	$\frac{V_f}{I_o}$	$\frac{I_f}{I_o}$	$\frac{I_f}{V_o}$
$A = \frac{X_o}{X_i}$	$A_V = \frac{V_o}{V_i}$	$G_M = \frac{I_o}{V_i}$	$A_I = \frac{I_o}{I_i}$	$R_M = \frac{V_o}{I_i}$
$D = 1 + \beta A$	$1 + \beta A_V$	$1 + \beta G_M$	$1 + \beta A_I$	$1 + \beta R_M$
A_f	$\frac{A_V}{D}$	$\frac{G_M}{D}$	$\frac{A_I}{D}$	$\frac{R_M}{D}$
R_{if}	$R_i D$	$R_i D$	$\frac{R_i}{D}$	$\frac{R_i}{D}$
R_{of}	$\frac{R_o}{(1+\beta A_v)}$	$R_o(1+\beta G_m)$	$R_o(1+\beta A_i)$	$\frac{R_o}{(1+\beta R_m)}$
$R'_{of} = R_{of} \Vert R_L$	$\frac{R_o}{(1+\beta A_v)}$	$\frac{R'_o(1+\beta G_m)}{(1+\beta G_M)}$	$\frac{R'_o(1+\beta I_I)}{(1+\beta I_I)}$	$\frac{R'_o}{(1+\beta R_M)}$

Analysis of Voltage Series Feedback Amplifiers

Analysis of the transistor emitter follower

In the transistor emitter follower circuit, the feedback voltage is the voltage across R_o and sampled signal is $V_o \; across \; R_e$



Analysis;

Step 1: Identify topology

By shorting output voltage $(V_o = 0)$, feedback signal becomes zero - voltage sampling

Feedback signal V_f is subtracted from externally applied signal V_s - series mixing

Hence the topology is a **voltage series feedback** amplifier

Step 2: Find input circuit;

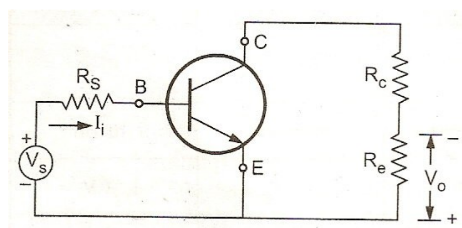
To find input circuit set $V_o = 0$,

V_s in series with R_s appears between $B \; and \; E$

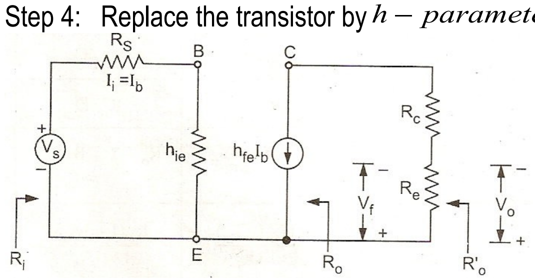
Step 3: Find Output circuit;

To find output circuit, set $I_i = I_b = 0$,

R_e appears only in the output loop only.



Step 4: Replace the transistor by $h - parameters$



Step 5: Find open loop voltage gain

$$A_V = \frac{V_o}{V_s} = \frac{h_{fe} I_b R_e}{V_s}$$

Applying KVL to input loop, we get

$$V_s = I_b\left(R_s + h_{ie}\right)$$

Substituting value of V_s $A_V = \dfrac{h_{fe} R_e}{R_s h_{ie}}$

Step 6: Indicate $V_o \; and \; V_f$ and calculate β , $\beta = \dfrac{V_f}{V_o} = 1$ (∵ both voltage present across R_e)

Step 7: Calculate $D; A_{Vf}; R_{if}; R_{of} \; and \; R'_{of}$

(i) $D = \left(1 + \beta A_V\right)$

(ii) $A_{Vf} = \dfrac{A_V}{1 + \beta A_V}$

(iii) $R_{if} = R_i D = \left(R_s + h_{ie}\right) D$

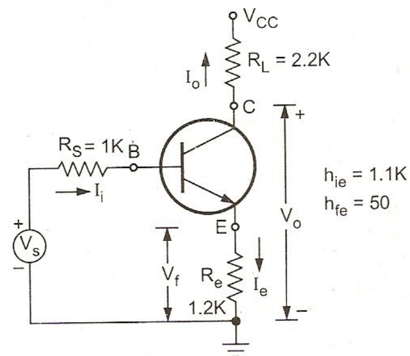
(iv) $R_{of} = R_o = \infty$

(v) $R'_{of} = \dfrac{R'_o}{D}; \quad where \; R'_o = R_e$

Analysis of Current Series Feedback Amplifiers

Common Emitter Configuration with Unbypassed R_e

Common emitter circuit with unbypassed R_e, resistor R_e is common to base – emitter input circuit as well as collector – emitter output circuit and input current I_b as well as output current I_c both flow through it.



The voltage drop across R_e; - $V_f = (I_b + I_c)R_e = I_e R_e = -I_o R_e$

This voltage drop shows that the output current I_o is being sampled and it is being converted to voltage by feedback network, at the input side V_f is being subtracted from V_s to produce V_i. Therefore the feedback applied is in series.

Analysis;

Step 1: Identify topology;

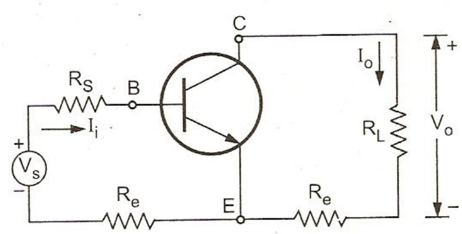
By opening output loop (output current, $I_o = 0$), feedback signal becomes zero; - Current sampling,

The feedback signal is subtracted from externally applied signal V_s; - Series mixing,

Hence the topology is a **current series feedback** amplifier.

Step 2: Find input circuit;

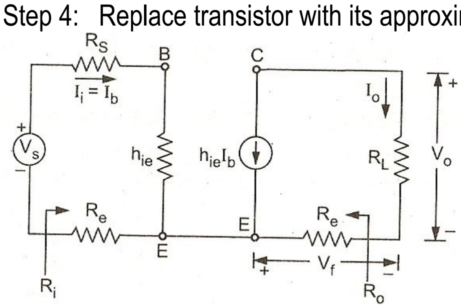
Set $I_o = 0$, then R_e appears at the input side.



Step 3: Find output circuit;

Set $I_i = 0$, then R_e appears in the input circuit

Step 4: Replace transistor with its approximate $h-parameter$ equivalent circuit



Step 5: Find open loop transfer gain;

$$G_M = \frac{I_o}{V_i} = \frac{-h_{fe}I_b}{V_s} = \frac{-h_{fe}I_b}{I_b(R_s + h_{ie} + R_e)} = \frac{-h_{fe}}{(R_s + h_{ie} + R_e)}$$

Step 6: Indicate $I_0\ and\ V_f$ and calculate β $\beta = \frac{V_f}{I_o} = \frac{I_e R_e}{I_o} = -R_e \because I_e = -I_o$

Step 7: Calculate $D; G_{Mf}; A_{Vf}; R_{if}; R_{of}\ and\ R'_{of}$

(i) $D = (1 + \beta G_M)$

(ii) $G_{Mf} = \dfrac{G_M}{D}$

(iii) $A_{Vf} = \dfrac{V_o}{V_s} = \dfrac{I_o R_L}{V_s} = G_{Mf} R_L \ \because G_{Mf} = \dfrac{I_o}{V_s}$

(iv) R_i - From the diagram $R_i = R_s + h_{ie} + R_e$; Hence $R_{if} = D \times R_i$

(v) $R_o = \infty; \ R_{of} = R_o \times D = \infty;$
$R'_{of} = R_{of} \| R_L = R_L \ \because R_{of} = \infty$

Preparatory Questions

PART - A

Que 1. Mention two types of feedback in electronic circuits,
Que 2. Mention the signals that are fed on the input of the amplifier with feedback,
Que 3. Identify the type of feedback used in;
- (i) Amplifiers,
- (ii) Oscillators,

Que 4. Define;
- (i) Voltage Amplifier,
- (ii) Transresistance Amplifier,

Que 5. Define;
- (i) Transconductance Amplifier,
- (ii) Current Amplifier,

Que 6. What is the function of;
- (i) Mixer network?
- (ii) Sampling network?

PART - B

Que 1. Classify amplifiers in accordance to how input parameters are related to the output parameters.
Que 2. Draw neat and well labeled diagrams to show (i) Voltage or node sampling (ii) Current or loop sampling
Que 3. Prove that gain without feedback is always greater than gain with feedback and it decreases with increase in feedback factor.
Que 4. Prove that the Stability of Gain with feedback is less than Stability of Gain without feedback by a factor $1 + \beta A$
Que 5. Draw the Characteristic of Gain Vs Frequency of an amplifier and indicate;
- (i) Lower and Upper cut-off frequencies without feedback
- (ii) Bandwidth of the amplifier without feedback
- (iii) Lower and Upper cut-off frequencies with feedback
- (iv) Bandwidth of the amplifier with feedback

Que 6. Account for the effect of negative feedback on;
- (i) Noise
- (ii) Distortion

PART - C

Que 1. Show that introducing negative feedback the high frequency response of the amplifier is improved. i.e. the higher cut-off frequency with feedback is greater than the higher cut-off frequency without feedback.

Que 2. An amplifier has mid-band voltage gain A_{mid} of 1000 with $f_L = 50Hz$ and $f_H = 50kHz$, if 5% feedback is applied then calculate (i) gain with feedback $A_{f\,mid}$, (ii) upper cut-off frequency with feedback f_{Hf} and (iii) lower cut-off frequency with feedback f_{Lf} (iv) Calculate the bandwidth with feedback BW_f.

Que 3. Show that introducing negative feedback the low frequency response of the amplifier is improved. i.e. the lower cut-off frequency with feedback is less than the lower cut-off frequency without feedback.

Que 4. (i) Identify the type of feedback topology used in the diagram Fig. 1

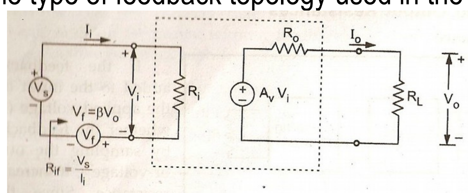


Fig. 1

(ii) Show that the input resistance with feedback $R_{if} = R_i(1+\beta A_v)$

Que 5. (i) Identify the type of feedback topology used in the diagram Fig. 2

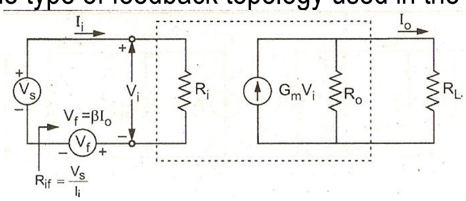


Fig. 2

(ii) Show that the input resistance with feedback $R_{if} = R_i(1+\beta G_M)$, $where\ G_M = \dfrac{G_m R_o}{R_o + R_L}$

Que 6. (i) Identify the type of feedback topology used in the diagram Fig. 3

(ii) Show that the output resistance with feedback $R_{of} = \dfrac{V}{I} = \dfrac{R_o}{(1+R_m\beta)}$,

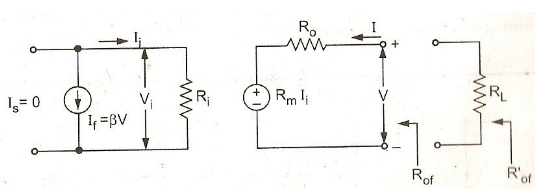


Fig.

Que 7. (i) Identify the type of feedback topology used in the diagram Fig.4 .

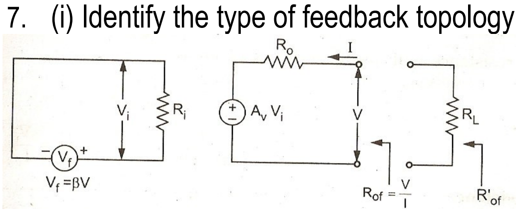


Fig. 4

(ii) Show that the output resistance with feedback $R_{of} = \dfrac{V}{I} = \dfrac{R_o}{(1+\beta A_v)}$

UNIT 2. OSCILLATORS

L – 6 Basic theory of Oscillators

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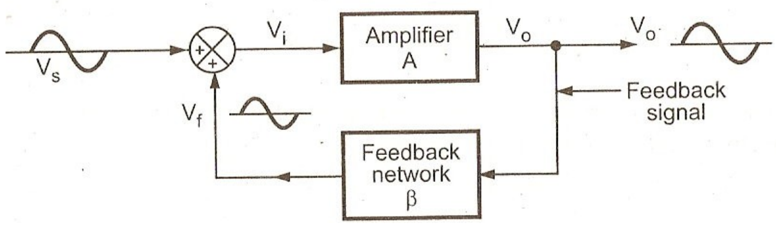
1. Introduction

An oscillator is a circuit which acts as a generator, generating the output signal which oscillates with constant amplitude and frequency.

An oscillator is an amplifier which uses a positive feedback without any external input signal.



Where $A = \frac{V_o}{V_i}$ - open loop gain of the amplifier,

$A_f = \frac{V_o}{V_s}$ - closed loop gain of the circuit or gain with feedback,

$V_i = V_s + V_f$ Since the feedback is positive and V_f is added to V_s to generate amplifier input

$V_f = \beta V_o$ The feedback voltage V_f depends on the feedback gain β . Hence substituting,

$V_i = V_s + \beta V_o \quad \therefore \quad V_s = V_i - \beta V_o$

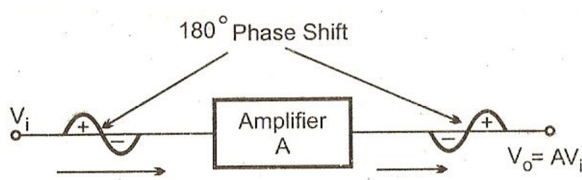
Substituting $A_f = \frac{V_o}{V_i - \beta V_o}$, dividing both the numerator and denominator by V_i

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

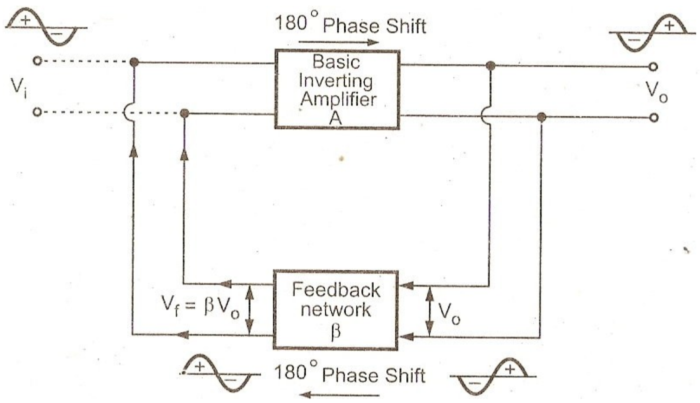
The gain with feedback increases as the amount of positive feedback increases. By increasing β , the circuit stops amplifying and starts oscillating.

2. Barkhausen Criterion for oscillation in feedback oscillators,

Consider an inverting amplifier which produces a phase shift of 180^0 between the input and output, with an open loop gain A .



The input V_i is derived from its output V_o using the feedback network, and then the voltage derived from output using the feedback network must be in phase with V_i . Thus the feedback network must introduce a phase shift of 180^0 while feeding back the voltage from output to input to ensure positive feedback.



Consider a fictitious voltage V_i applied at the input of the amplifier, hence $V_o = \beta V_i$. The feedback factor β decides the feedback to be to be given to the input $V_f = -\beta V_o$, the negative signal indicates 180^0 phase shift.

Hence $V_f = -\beta A V_i$

For the oscillator $V_f = V_i$, then $V_i = -A\beta V_i$ $Hence, -A\beta = 1$ - $(Barkhausen\ criterion)$

3. Classification of oscillators

The various ways in which oscillators are classified are;

(a) Based on the output wave form – sinusoidal or non-sinusoidal
(b) Based on circuit components - RC, LC or crystal
(c) Based on range of operating frequency – Low frequency (LF), High frequency (HF),
(d) Based on whether feedback is used or not – Feedback type, Non-feedback type.

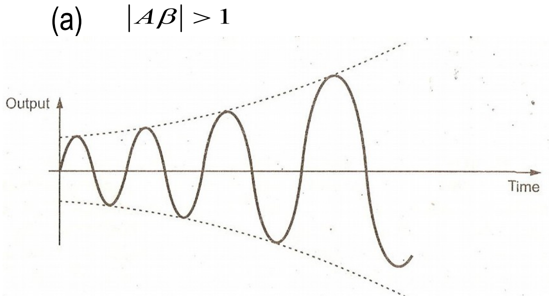
4. Barkhausen Criterion

- The total phase shift around a loop, as the signal proceeds from input through amplifier, feedback network back to amplifier again, completing a loop, is precisely 0^0 or 360^0 (or an integral multiple of 2π radians.
- The magnitude of the product of the open loop gain of the amplifier (A) and the feedback factor β is unity i.e. $|A\beta| = 1$

Oscillations will not be sustained if, at the oscillator frequency, the magnitude of the product of the transfer gain of the amplifier and the magnitude of the feedback factor of the feedback network are less than unity.

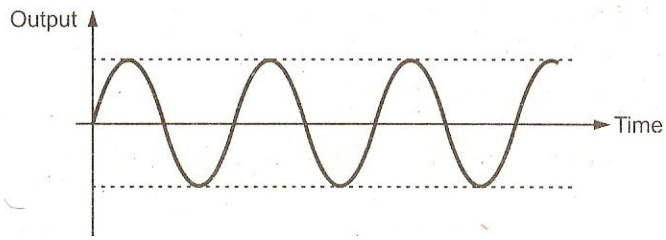
Consider the following situations

(a) $|A\beta| > 1$



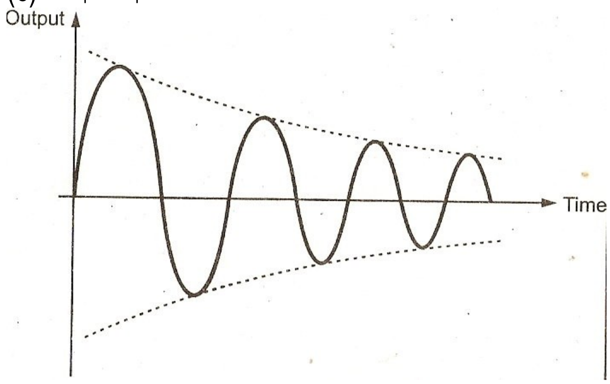
When the total phase shift around a loop is 0^0 or 360^0 and $|A\beta| > 1$, then the output oscillations are of **growing** type, the amplitude of oscillations goes on increasing.

(b) $|A\beta| = 1$



When the total phase shift around a loop is 0^0 or 360^0 ensuring positive feedback and $|A\beta| = 1$, then the oscillations are with constant frequency and amplitude – **sustained oscillations.**

(c) $|A\beta| < 1$



When total phase shift around a loop is 0^0 or 360^0 and $|A\beta| < 1$ then the oscillations are of **decaying** type, the amplitude of oscillations decreases exponentially and finally the oscillations ceases.

5. Mechanism for start of oscillation and stabilization of amplitude

In analysis the input voltage of an oscillator V_i is assumed as a fictitious input because the oscillator output supplies its own input. How the oscillator starts and where does the starting voltage come from?

Oscillators are self-starting and begin as soon as power is applied due to presence of electric noise. Every resistance has some free electrons, which under the influence of temperature, move randomly and can generate noise voltage across a resistance. Such noise voltages at a desired frequency f_o present across the resistance are amplified.

To amplify such small noise voltages and to start oscillations $|A\beta|$ *is kept greater than unity* at start, the amplified voltage appears at the output terminals and when the part of this output is sufficient to drive the input of the amplifier circuit, the circuit adjusts itself to get $|A\beta| = 1$ and with a phase shift of 360^0 sustained oscillations are obtained.

Example2.1

In a certain oscillator circuit, the gain of the amplifier $A = \dfrac{-16\times10^6}{j\omega}$ *and the feedback factor of the feedback network* $\beta = \dfrac{10^3}{\left[2\times10^3 + j\omega\right]^2}$.

a. *Verify Barkhausen Criterion for the sustained oscillations.*

b. *Find the frequency at which the circuit will oscillate.*

Solution

$$A\beta = -\frac{16\times10^6\times10^3}{j\omega\left[2\times10^3 + j\omega\right]^2} = -\frac{16\times10^9}{j\omega\left[4\times10^6 + 4\times10^3 j\omega + (j\omega)^2\right]} = -\frac{16\times10^9}{j\omega\left[4\times10^6 + 4j\omega\times10^3 - \omega^2\right]}$$

$$= -\frac{16\times10^9}{4\times10^6 j\omega + 4j^2\omega^2\times10^3 - j\omega^3} = -\frac{16\times10^9}{j\omega\left[4\times10^6 - \omega^2\right] - \left[\omega^2\times4\times10^3\right]};\ \text{since } j^2 = -1$$

After rationalizing the denominator $A\beta = \dfrac{16\times10^9\left[4\times10^3\omega^2 + j\omega\left(4\times10^6 - \omega^2\right)\right]}{16\times10^6\omega^4 + \omega^2\left(4\times10^6 - \omega^2\right)^2}$;

Now to have $\angle A\beta = 0^0$ *the imaginary part of* $A\beta$ *. i.e.* $\omega\left(4\times10^6 - \omega^2\right) = 0$

Therefore $\omega = 0$ *or* $\left(4\times10^6 - \omega^2\right) = 0$ *neglecting* $\omega = 0$

Hence $\omega^2 = 4\times10^6 \Rightarrow \omega = 2\times10^3 rad/\sec$*; at this frequency* $|A\beta|$ *is obtained;*

a. $|A\beta| = \dfrac{16\times10^9\left[4\times10^3\omega^2\right]}{16\times10^6\omega^4 + \omega^2\left(4\times10^6 - \omega^2\right)^2}$; *substituting* $\omega = 2\times10^3$

$|A\beta| = 1$

Therefore, at $\omega = 2\times10^3 rad/\sec$ *and* $\angle A\omega = 0^0$ *as imaginary part is zero while* $|A\beta| = 1$*. Thus Barkhausen Criterion is satisfied.*

b. *The frequency at which the circuit will oscillate is the value of at which* $|A\beta| = 1$ *and at the same time* $\angle A\beta = 0^0$ *is* $\omega = 2\times10^3 rad/\sec$

but $\omega = 2\pi f = 2\times10^3 \Rightarrow f = \dfrac{\omega}{2\pi} = \dfrac{2\times10^3}{2\pi} = 318.309 Hz$

L – 7: Analysis of RC Oscillators using Cascade connection of Low Pass and High Pass filters

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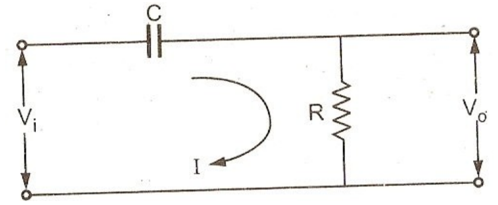
1. RC Phase shift oscillators



An RC phase shift oscillators consists of an amplifier and a feedback network made up of resistors and capacitors.

Let $V_i = V_m \sin \omega t$; The impedance of the circuit

$$Z = R - jX_c \quad where \quad X_c = \frac{1}{2\pi fC} = |Z|\angle - \phi$$

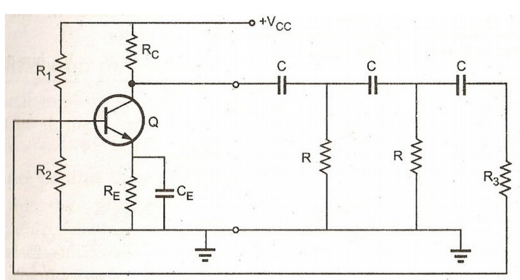
Where $\tan\phi = \frac{-X_c}{R} = \frac{1}{2\pi fRC}$;

$\phi - Phase\ angle$ of the circuit which depends on the values of R and C. Rand C are so selected that such a phase angle is 60^0

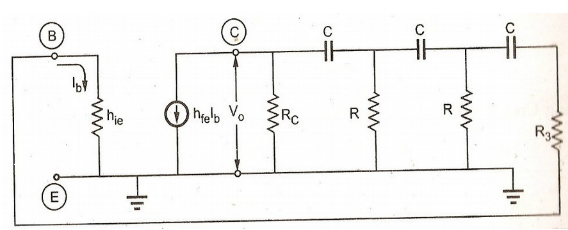
If the amplifier causes a phase shift of 180^0 then the feedback network should create a phase shift of 180^0, to satisfy *Barkhausen Criterion.* Hence three sections of RC are connected in cascade each introducing a phase shift of 60^0.

2. Transistorized RC phase shift oscillator

A transistor in common emitter is used as an active element of the amplifier stage. A phase shift network is the feedback network, i.e., the output of the feedback network is given as an input to the amplifier. All the resistance values and all the capacitor values are the same, so that for a particular frequency, each section of R and C produces a phase shift of 60^0.



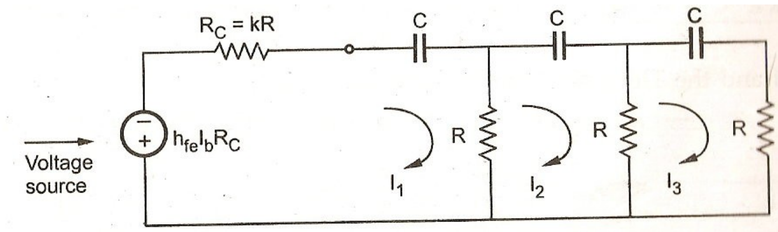
Transistorized RC phase shift oscillator



Equivalent oscillator circuit

Derivation for the Frequency of Oscillation

Replacing the transistor by its approximate $h - parameter$ model we get the equivalent oscillator circuit. Replacing $h_{ie} + R_e = R$; Replacing $h_{fe}I_b$ by its equivalent voltage source $h_{fe}I_b R_C$ and assuming the ratio of the resistance $\frac{R_c}{R} = k$, we get a modified equivalent circuit.



Modified equivalent circuit

Applying KVL for various loops

Loop 1:

$$-I_1R_C - \frac{1}{j\omega C}I_1 - I_1R + I_2R - h_{fe}I_bR_C = 0;\ \text{Replacing } R_C \text{ by } kR$$

$$I_1\left[(k+1)R + \frac{1}{j\omega C}\right] - I_2R = -h_{fe}I_bkR$$

Loop 2:

$$-\frac{1}{j\omega C}I_2 - I_2R - I_2R + I_1R + I_3R = 0;\quad Hence\quad -I_1R + I_2\left[2R + \frac{1}{j\omega C}\right] - I_3R = 0$$

Loop 3:

$$-I_3\frac{1}{j\omega C} - I_3R - I_3R + I_2R = 0;\quad Hence\quad -I_2R + I_3\left[2R + \frac{1}{j\omega C}\right] = 0$$

Using Cramer's rule to solve for I_3

$$D = \begin{vmatrix} (k+1)R + \frac{1}{j\omega C} & -R & 0 \\ -R & 2R + \frac{1}{j\omega C} & -R \\ 0 & -R & 2R + \frac{1}{j\omega C} \end{vmatrix} \quad and \quad D_3 = \begin{vmatrix} (k+1)R + \frac{1}{j\omega C} & \cdots \\ -R & 2R + \cdots \\ 0 & \end{vmatrix}$$

Solving for D we get $D = \dfrac{(j\omega CR)^3[3k+1] + (j\omega CR)^2[4k+6] + (j\omega CR) + 1}{(j\omega C)^3}$

Solving for D_3 we get $D_3 = -kR^3h_{fe}I_b$

Therefore $I_3 = \dfrac{D_3}{D} = \dfrac{-kR^3h_{fe}I_b(j\omega C)^3}{(j\omega CR)^3[3k+1] + (j\omega CR)^2[4k+6] + (j\omega CR) + 1}$

I_3 - Output current of the feedback circuit

I_b - Input current of the amplifier

$I_C = h_{fe}I_b$ - Input current of the feedback circuit

$$\beta = \frac{Output\ \ of\ \ feedback\ circuit}{Input\ of\ \ feedback\ circuit} = \frac{I_3}{h_{fe}I_b}$$

$$\therefore A\beta = \frac{I_3}{h_{fe}I_b} \times h_{fe} = \frac{I_3}{I_b} = \frac{-kR^3h_{fe}(j\omega C)^3}{(j\omega CR)^3[3k+1] + (j\omega CR)^2[4k+6] + (j\omega CR) + 1}$$

$$A\beta = \frac{-kh_{fe}(j^3)(\omega RC)^3}{(j^3)(\omega RC)^3[3k+1] + (j^2)(\omega RC)^2[4k+6] + (j\omega CR) + 1}$$

Substituting $j^2 = -1;\ j^3 = -j$

$$A\beta = \frac{jkh_{fe}\omega^3R^3C^3}{-j\omega^3R^3C^3[3k+1] - \omega^2R^2C^2[4k+6] + (j\omega CR) + 1}$$

Separating the real and imaginary part in the denominator, dividing numerator and denominator by $j\omega^3 R^3 C^3$ and replacing $-\frac{1}{j} = j$

$$A\beta = \frac{kh_{fe}}{j\left[\frac{1}{\omega^3 R^3 C^3} - \frac{4k}{\omega CR} - \frac{6}{\omega CR}\right] + \left[3k + 1 - \frac{5}{\omega^2 R^2 C^2} - \frac{k}{\omega^2 R^2 C^2}\right]} \text{***}$$

According to Barkhausen criterion, the angle of the term $\angle A\beta = 0$, hence the imaginary part of the denominator term must be zero.

Hence $j\left[\frac{1}{\omega^3 C^3 R^3} - \frac{4k}{\omega RC} - \frac{6}{\omega R}\right] = 0$; Substituting $\frac{1}{\omega RC} = \alpha$ for simplicity;

$\alpha^3 - 4k\alpha - 6\alpha = 0$ or $\alpha\left(\alpha^2 - 4k - 6\right) = 0$ neglecting zero value $\alpha^2 = 4k + 6$

$\alpha = \sqrt{4k+6} \Rightarrow \frac{1}{\omega RC} = \sqrt{4k+6}$

$Hence\ \omega = \frac{1}{RC\sqrt{4k+6}} \Rightarrow f = \frac{1}{2\pi RC\sqrt{4k+6}}$ The frequency at which $\angle A\beta = 0$

Replacing the value of $\omega = \frac{1}{RC\sqrt{4k+6}}$ on the equation ***

$$\mathrm{A}\beta = \frac{k\,h_{fe}}{-4k^2 - 23k - 29} \quad for\ \ |\mathrm{A}\beta| = 1 \quad h_{fe} = 4k + 23 + \frac{29}{k}$$

<u>Example</u>

For phase shift oscillator, the feedback network uses $R = 6\ k\Omega\ \ and\ \ C = 1500\ pF$. The transistor used has a collector resistance $R_C = 18k\Omega$. Calculate the frequency of oscillations and the minimum value h_{fe} of the transistor.

<u>Solution</u>

$$R = 6\ k\Omega\ ; C = 1500\ pF;\ R_C = 18k\Omega \Rightarrow k = \frac{R_C}{R} = \frac{18\times10^3}{6\times10^3} = 3$$

Hence, the frequency of oscillation

$$f = \frac{1}{2\pi RC\sqrt{4k+6}} = \frac{1}{2\pi\times6\times103\times1500\times10-12\sqrt{(4\times3)+6}} = 4.168kHz$$

$$\left(h_{fe}\right)_{\min} = 4k + 23 + \frac{29}{k} = 4\times3 + 23 + \frac{29}{3} = 44.67$$

L – 8 <u>Analysis of RC Oscillators using Cascade connection of Low Pass and High Pass filters (Cont..)</u>

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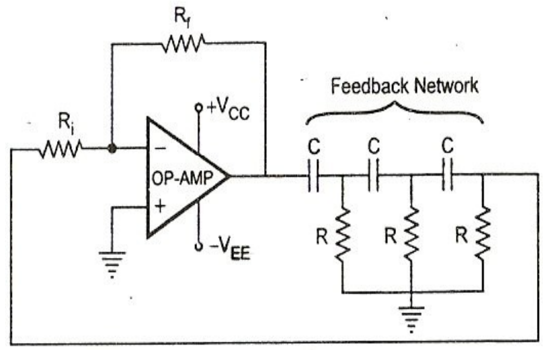
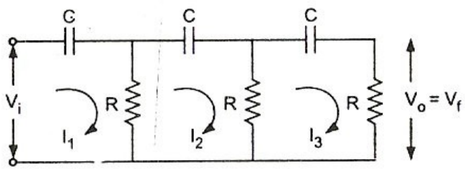
1. Phase shift oscillator using Op Amp



The Op Amp is used in inverting mode to provide 180^0 phase shift, the feedback circuit used is the same as in the transistorized phase shift oscillator. The output of the Op Amp is fed to three-section RC network which provides the needed 180^0 phase shift. The Op Amp gain (A) is adjusted with the help of resistances R_f and R_i such that the product of Op Amp gain A and the feedback network gain (β)is slightly greater than one to get the required oscillations. $|A\beta| \geq 1$

Transfer function of RC feedback network;

Apply KVL to various loops and then writing the equations in matrix form and using Cramer's rule to get I_3 ,



$$I_1\left(R+\frac{1}{j\omega C}\right)-I_2R=V_i$$

$$-I_1R+I_2\left(2R+\frac{1}{j\omega C}\right)-I_3R=0$$

$$0-I_2R+I_3\left(2R+\frac{1}{j\omega C}\right)=0$$

$$\begin{vmatrix} R+\frac{1}{j\omega C} & -R & 0 \\ -R & 2R+\frac{1}{j\omega C} & -R \\ 0 & -R & 2R+\frac{1}{j\omega C} \end{vmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} = \begin{bmatrix} V_1 \\ 0 \\ 0 \end{bmatrix}$$

Using Cramer's rule to obtain I_3

$$D = \begin{vmatrix} \frac{1+j\omega RC}{j\omega C} & -R & 0 \\ -R & \frac{1+2j\omega RC}{j\omega C} & -R \\ 0 & -R & \frac{1+2j\omega RC}{j\omega C} \end{vmatrix}$$

$$D=\frac{(1+j\omega RC)(1+j\omega RC)^2}{(j\omega)^3C^3}-\frac{R^2(1+j\omega RC)}{j\omega C}-\frac{R^2(1+j\omega RC)}{j\omega C}=\frac{1+5j\omega RC+6(j\omega)^2R^2C^2+(j\omega)^3R^3C^3}{(j\omega)^3C^3}$$

$$D_3 = \begin{vmatrix} \frac{1+j\omega RC}{j\omega C} & \\ -R & \frac{1+2\ldots}{\ldots} \\ 0 & -R \end{vmatrix}$$

$$I_3=\frac{D_3}{D}=\frac{V_iR^2(j\omega)^3C^3}{1+5j\omega RC+6(j\omega)^2R^2C^2+(j\omega)^3R^3C^3}$$

$$V_o = V_f = I_3 R = \frac{V_i R^3 (j\omega)^3 C^3}{1 + 5j\omega RC + 6(j\omega)^2 R^2 C^2 + (j\omega)^3 R^3 C^3}$$

$$\therefore \beta = \frac{V_o}{V_i} = \frac{R^3 (j\omega)^3 C^3}{1 + 5j\omega RC + 6(j\omega)^2 R^2 C^2 + (j\omega)^3 R^3 C^3} \quad replacing\ (j\omega)^2 = -\omega^2 and\ (j\omega)^3 = -j\omega^3$$

$$\beta = \frac{R^3 (j\omega)^3 C^3}{1 + 5j\omega RC - 6j\omega^2 R^2 C^2 - j\omega^3 R^3 C^3}$$

Dividing numerator and denominator by $-j\omega^3 R^3 C^3$

$$\beta = \frac{1}{1 - 5\left(\frac{1}{\omega RC}\right)^2 + j\left(\frac{1}{\omega RC}\right)\left(6 - \left(\frac{1}{\omega RC}\right)^2\right)}$$

To have a phase shift of 180^0, the imaginary part in the denominator must be zero

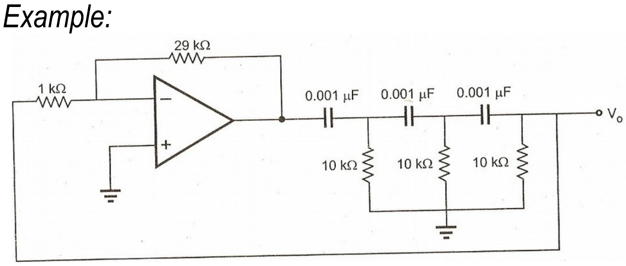
$$Therefore\ \left(\frac{1}{\omega RC}\right)\left(6 - \left(\frac{1}{\omega RC}\right)^2\right) = 0;\ Hence, either \left(\frac{1}{\omega RC}\right) = 0,\ \ or\ \left(\frac{1}{\omega RC}\right)^2 = 6$$

$$Then\ \frac{1}{\omega RC} = \sqrt{6} \Rightarrow f = \frac{1}{2\pi RC\sqrt{6}}\ \ The\ frequency\ with\ which\ circuit\ will\ oscillate$$

At this frequency $\beta = \frac{1}{1 - 5(\sqrt{6})^2} = -\frac{1}{29}$; the negative sign indicates a phase shift of 180^0

$|\beta| = \frac{1}{29}$. To have oscillations $|A||\beta| > 1;\ \ |A| > \frac{1}{|\beta|} > \frac{1}{\frac{1}{29}}\ \ Therefore\ \ |A| \geq 29$

Example:



Determine the frequency of RC phase shift oscillator.

Solution: Given

$R = 10\ k\Omega;\ C = 0.01\ \mu F$

The frequency of the RC phase shift oscillator is given by, $f = \frac{1}{2\pi\sqrt{6}RC}$

Hence

$$f = \frac{1}{2\pi\sqrt{6} \times 10 \times 10^3 \times 0.001 \times 10^6} = 6.467\ kHz$$

L – 9. Analysis of Op-Amp Wien Bridge oscillators,

Date: ……………..…
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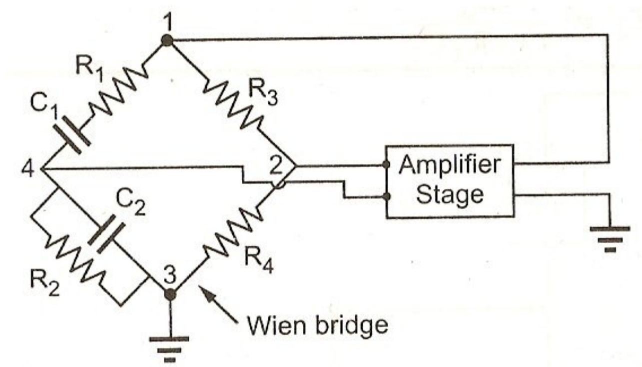
1. Wien Bridge Oscillator

The Wein Bridge Oscillator uses a non-inverting amplifier and hence does not provide any phase shift during amplifying stage. That means, no phase shift is necessary during feedback because the total phase shift around the loop is 0^0

Consider a basic Wein Bridge

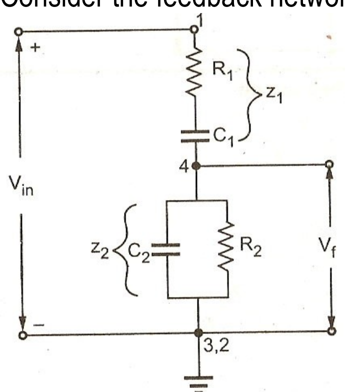


The output of the amplifier is applied between terminals 1 and 3 which is the input of the feedback network. The amplifier input is supplied from terminals 2 and 4 which is the output from the feedback network.

R_1, C_1 in series and R_2, C_2 in parallel are called – *frequency sensitive arms* because the component of these two arms decide the frequency of the oscillator.

V_{in} to the feedback network is 1 and 3 while output V_f of the feedback network is between 2 and 4. Such a feedback network is called lead-lag network since at very low frequencies it acts like a lead while at very high frequencies it acts like a lag.

Consider the feedback network of Wein Bridge;

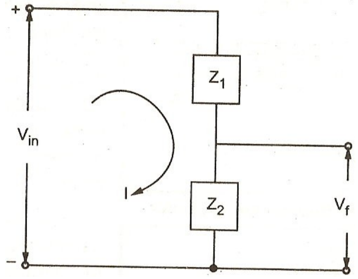


$$Z_1 = R_1 + \frac{1}{j\omega C_1} = \frac{1 + j\omega R_1 C_1}{j\omega C_1}$$

$$Z_2 = R_2 \left\| \frac{1}{j\omega C_1} \right. \Rightarrow \frac{1}{Z_2} = \frac{1}{R_2} + \frac{1}{\frac{1}{j\omega C}}$$

R_2

Using a simplified diagram



$$I = \frac{V_{in}}{Z_1 + Z_2}; V_f = IZ_2 = \frac{V_{in} Z_2}{Z_1 + Z_2}$$

$$Hence \beta = \frac{V_f}{V_{in}} = \frac{Z_2}{Z_1 + Z_2}$$

Substituting the values of Z_1 and Z_2

$$\beta = \frac{\frac{R_2}{1+j\omega R_2C_2}}{\frac{R_1}{1+j\omega R_1C_1}+\frac{R_2}{1+j\omega R_2C_2}} = \frac{\omega^2 C_1R_2\left(R_1C_1+R_2C_2+C_1R_2\right)+j\omega C_1R_2\left(1-\omega^2R_1R_2C_1C_2\right)}{\left(1-\omega^2R_1R_2C_1C_2\right)^2+\omega^2\left(R_1C_1+R_2C_2+C_1R_2\right)^2}$$

To have zero phase shift of the feedback network, its imaginary part must be equal to zero;

$$\therefore \omega C_1R_2\left(1-\omega^2R_1R_2C_1C_2\right)=0 \Rightarrow \omega^2R_1R_2C_1C_2=1 \; ; \; \omega^2=\frac{1}{R_1R_2C_1C_2} \; and \; f=\frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}$$

f - is the frequency of the oscillator and it shows that the components of the frequency sensitive arms are the deciding factors, for the frequency.

In practice $R_2 = R_1 = R \;\; and \;\; C_2 = C_1 = C \therefore f = \frac{1}{2\pi RC} \;\; and \;\; \beta = \frac{1}{3}$ the magnitude of the feedback network at resonating frequency of the oscillator. The positive sign of β indicates that the phase shift by the feedback is zero (0^o).

To satisfy Barkhausen criterion for sustained oscillations;

$\left|A\beta\right| \geq 1; \;\; \left|A\right| \geq \frac{1}{\beta} \geq \frac{1}{\frac{1}{3}}; \; hence \;\; \left|A\right| \geq 3$ required gain of amplifier stage without any phase shift;

2. Advantages of the Wien bridge oscillator;

- By varying the two capacitor values can be varied simultaneously by mounting them on a common shaft to get different frequency ranges.
- No phase shift is required in the feedback network because the total phase shift around the loop is 0^o and the gain of the amplifier is low.

Wien Bridge Oscillator using Op Amp

This is an oscillator in which the amplifier circuit is an Op Amp while the feedback circuit remains Wein Bridge circuit. The feedback is given to the non-inverting terminal of the Op Amp to ensure zero phase shift.

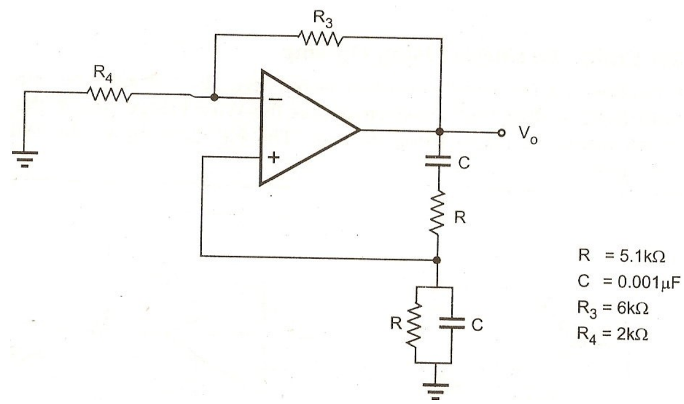
The resistance R and capacitor C are the components of *frequency sensitive arms* of the bridge. Resistance R_3 and R_4 form the part of the feedback path.

Gain of the Op amplifier $A = 1 + \frac{R_3}{R_4}$

According to the oscillating conditions $A \geq 3 \;\; \Rightarrow \; 1+\frac{R_3}{R_4} \geq 3 \;\; Hence \;\; \frac{R_3}{R_4} \geq 2$

Thus if the ratio of R_3 and R_4 is greater than or equal to two will provide sufficient loop gain for the circuit to oscillate at a frequency $f = \frac{1}{2\pi RC}$

Example 2.2



Determine whether the circuit shown will work as an oscillator or not. If yes, determine the frequency of the oscillator.

$Given\ R = 5.1k\Omega;\ R_3 = 6k\Omega;\ R_4 = 2k\Omega; C = 0.001\mu F$

Solution

The circuit is Wien Bridge oscillator using Op Amp. The gain of the loop $A = 1 + \frac{R_3}{R_4} = 1 + \frac{6}{2} = 4$

- *Since* $A > 3$, *the conditions necessary for oscillation are satisfied;*
- *Since the feedback is given to the non-inverting terminal the zero phase shift is ensured;*
- *The circuit will work as an oscillator.*

The frequency of oscillations $f = \frac{1}{2\pi RC} = \frac{1}{2\pi 5.1\times 10^3 \times 0.001\times 10^{-6}} = 31.2068\text{kHz}$

Comparison of RC phase shift and Wien Bridge Oscillator

RC phase shift Oscillator	Wien Bridge Oscillator
▪ Phase shift Oscillator used for low frequency range.	▪ Wien Bridge Oscillator used for low frequency range.
▪ Feedback network – Three RC section	▪ Feedback network – Lead-lag *Wien Bridge*
▪ Phase shift by feedback network – 180^0	▪ Phase shift by feedback network - 0^0
▪ Op Amp is used in an inverting mode	▪ Op Amp is used in a non-inverting mode
▪ Phase shift by Op Amp – 180^0	▪ Phase shift by Op Amp - 0^0
▪ Frequency of oscillation $f = \frac{1}{2\pi RC\sqrt{6}}$	▪ Frequency of oscillation $f = \frac{1}{2\pi RC}$
▪ Amplifier gain condition $\lvert A\rvert \geq 29$	▪ Amplifier gain condition $\lvert A\rvert \geq 3$
▪ Frequency variation is difficult	▪ Mounting the two capacitors on common shaft and varying their values, frequency can be varied

L – 10. Analysis of twin-T network,

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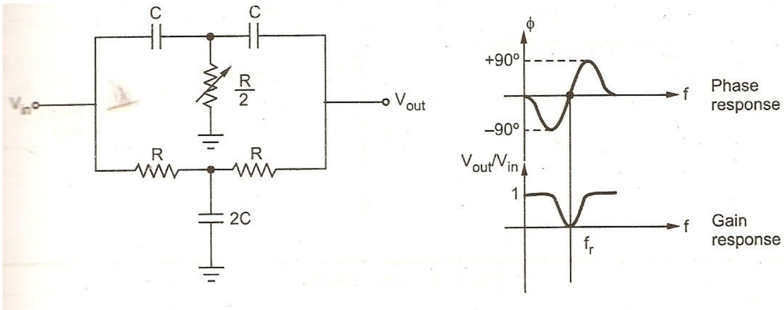
Twin - T Oscillator

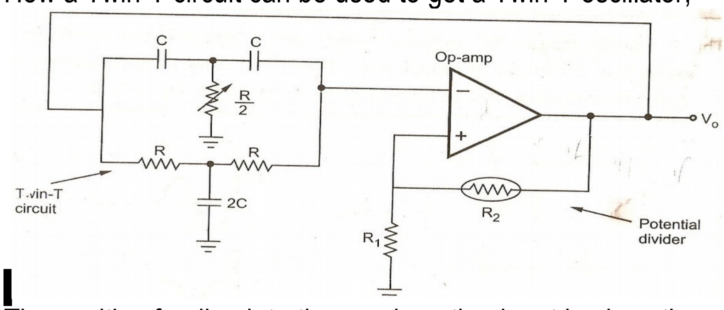
It is a RC Oscillator in which the Twin-T filter is a lead-lag circuit whose phase angle varies between $+90^0$ to -90^0 against the frequency. At $f = f_r$, its phase angle is 0^0 and it does not introduce any phase shift. At low and high frequencies its gain is 1, but at $f = f_r$ its gain is reduced to zero hence it notches out frequencies near f_r. The equation for its resonating frequency is $f_r = \dfrac{1}{2\pi RC}$.

A Twin-T filter is a combination of a *High Pass* filter and *Low Pass* filter that is why a Twin-T filter acts as a notch filter.



(a) Twin-T circuit (b) Frequency response

How a Twin-T circuit can be used to get a Twin-T oscillator;



The positive feedback to the non-inverting input is given through the potential divider (R_1 *and* R_2) whereby R_2 is a lamp. The negative feedback to the inverting input is given through twin-T filter. When the power is given to the circuit, lamp resistance R_2 is low thus positive feedback is maximum which helps to build oscillations. As the oscillations grows, the lamp resistance increases, decreasing the positive feedback. This controls the growing oscillations and makes the oscillations to sustain. The lamp acts as a stabilizer of output voltage level.

Oscillations occur only at the resonating frequency f_r because at frequencies other than f_r there is a negative feedback which is not conducive for oscillations. At f_r the negative feedback is negligible, hence positive feedback through R_1 *and* R_2 allows the circuit to oscillate.

Disadvantages of a twin-T oscillator

- It operates only at one frequency f_r
- R_2 of the potential divider must be larger enough $(10R_1 \ to \ 100R_1)$ to obtain oscillating frequency f_r

L – 11. Analysis of LC oscillators,

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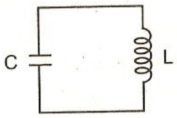
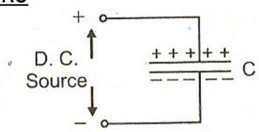
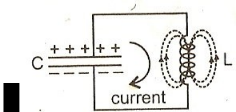
Analysis of LC Oscillators

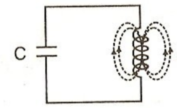
LC Oscillators are oscillators which use elements L *and* C to produce oscillations. The circuit using elements L *and* C is known as <u>tank</u> or <u>oscillatory</u> or <u>resonant</u> or <u>tuned</u> circuit. These oscillators are used for high frequency range (200 kHz to GHz)

How a $L\ C$ tank circuit works





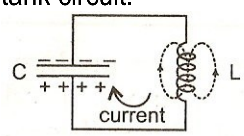


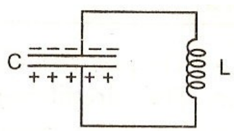


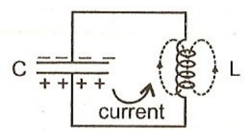
$L\ C$ tank circuit | Initial charging | Discharging | Magnetic fields in L

When the capacitor C gets charged, electrostatic energy gets stored in the capacitor. When the charged capacitor is connected across an inductor L , it discharges through L . Due to such current flow, the magnetic gets set up around the inductor L and the inductor starts storing the energy. When the capacitor is fully discharged, maximum current flows through the circuit, and at this instant all the electrostatic energy gets stored as magnetic energy in the inductor.

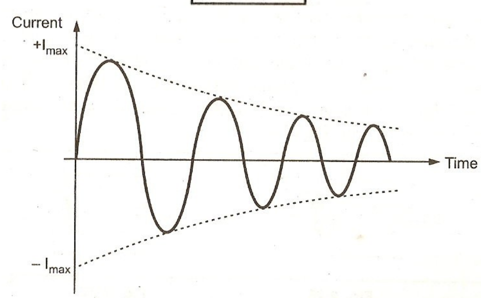
When the magnetic field around L starts collapsing, the capacitor is charged with opposite polarity (as per Lenz's law). After some time the capacitor gets fully charged with opposite polarities meaning that the entire magnetic energy has been converted back to electrostatic energy in capacitor, since the process in continuous an alternating (oscillatory) current is produced in the tank circuit.







The energy transfer from C to L and L to C losses occur due to which the amplitude of oscillating current decays exponentially. Damped oscillations stop after some time.



To make oscillations sustain, an amplifier can be used to compensate the energy which is lost. The frequency of oscillations generated by a tank circuit depends on $L\ and\ C$ values.

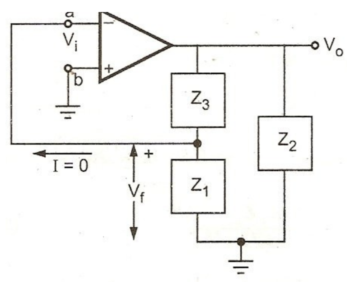
$$f = \frac{1}{2\pi\sqrt{LC}} Hz\ \ Where;\ L - henries;\ C - farads$$

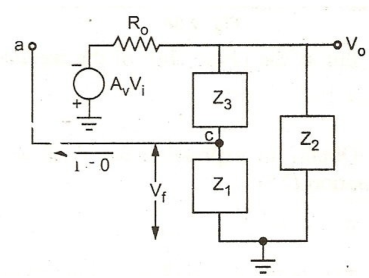
Depending upon the type of tank circuits, $L\ C$ oscillators are classified as

- *Colpitt's Oscillators;*
- *Hartley Oscillators;*
- *Clap Oscillators*;

Basic forms of LC circuit

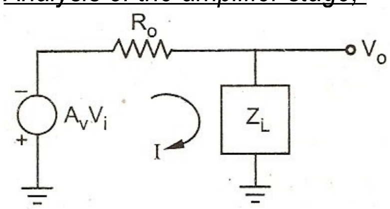
The circuit consists of an amplifier stage of gain A_V with its output to a feedback stage of impedances $Z_1, Z_2\ and\ Z_3$. The amplifier provides a phase shift of 180^0 while the feedback network provides an additional phase shift of 180^0





[redacted]Basic Circuit Equivalent Circuit

Analysis of the amplifier stage;



The input impedance of the amplifier is infinite; hence there is no current flowing towards the input terminal i.e. $I = 0$. Let R_0 be the output impedance of the amplifier stage.

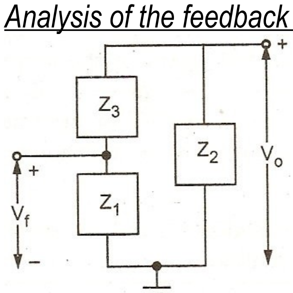
As $I = 0$, Z_1, Z_2 appear in series, Z_1, Z_2 and Z_3 form an equivalent load impedance Z_L.

Where $Z_L = \frac{Z_2(Z_1+Z_3)}{Z_1+Z_2+Z_3}$

$$\textit{Therefore}\ \mathrm{I} = \frac{-\mathrm{A_V}V_i}{R_0 + Z_L}\quad \textit{Since}\ V_0 = IZ_L \Rightarrow V_0 = \frac{-\mathrm{A_V}V_iZ_L}{R_0 + Z_L};\quad \textit{Hence}\ \frac{V_0}{V_i} = A = \frac{-\mathrm{A_V}Z_L}{R_0 + Z_L}$$

Where A, is the gain of the amplifier stage.

Analysis of the feedback stage;



$$V_f = V_0\left[\frac{Z_1}{Z_1 + Z_3}\right], \textit{Therefore}\ \frac{V_f}{V_0} = \beta = \left[\frac{Z_1}{Z_1 + Z_3}\right],$$

But as the phase shift of the feedback network is 180^0,

then $\beta = -\left[\frac{Z_1}{Z_1 + Z_3}\right]$.

To satisfy Barkhausen Criterion $-A\beta = 1$; $\textit{Therefore}\ -A\beta = \frac{-A_V Z_1 Z_L}{(R_0 + Z_L)(Z_1 + Z_3)}$

$$\text{Hence} -A\beta = \frac{-A_V Z_1\left[\frac{Z_2(Z_1 + Z_3)}{Z_1 + Z_2 + Z_3}\right]}{\left(R_0 + \left[\frac{Z_2(Z_1 + Z_3)}{Z_1 + Z_2 + Z_3}\right]\right)(Z_1 + Z_3)}$$

Dividing numerator and denominator by $\frac{Z_2(Z_1+Z_3)}{Z_1+Z_2+Z_3}$ and substituting

$$Z_1 = jX_1, Z_2 = jX_2 = and\ Z_3 = jX_3\ Where\ X = \omega L - \text{inductance} X = \frac{-1}{\omega C} - \text{capacitance}$$

$$-A\beta = \frac{-A_V X_1 X_2}{-X_2(X_1+X_3) + jR_0(X_1+X_2+X_3)}$$

To have a phase shift of 180^0, the imaginary part of the denominator must be equal to zero i.e. $X_1 + X_2 + X_3 = 0 \Rightarrow X_1 + X_3 = -X_2$

Hence $-A\beta = \frac{-A_V X_1 X_2}{-X_2(X_1+X_3)} or -A\beta = \frac{-A_V X_1}{-X_2} = A_V \frac{X_1}{X_2}$

According to Barkhausen Criterion $-A\beta$ must be positive and greater than unity. As A_V is positive $-A\beta$ will be positive only if $X_1\ and\ X_2$ will have the same sign. **This indicates that $X_1\ and\ X_2$ must be both inductive or both capacitive.**

Oscillator type	Reactance element in tank circuit		
	X_1	X_2	X_3
Hartley Oscillator	L	L	C
Colpitt's Oscillators	C	C	L

L – 12. Analysis of Hartley and Colpitt's Oscillators

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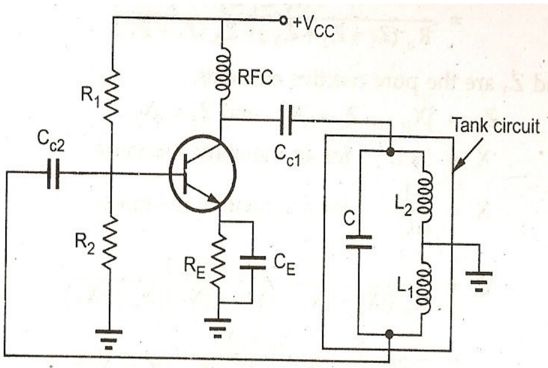
A. Analysis of Hartley Oscillator,

- **Transistorized Hartley Oscillator**

The amplifier stage uses a transistor in common emitter configuration.



Functions of amplifier stage components

R_1 *and* R_2 - biasing resistance;

RFC (Radio frequency choke)- isolation between a.c. and d.c. due to its high reactance at high frequencies and zero reactance for d.c.

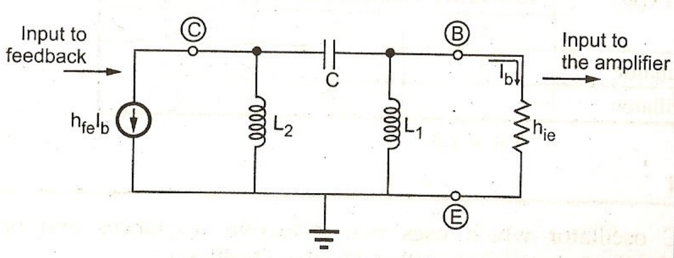
R_E -emitter circuit biasing resistor;

C_E -emitter bypass capacitor;

C_1 *and* C_2 -coupling capacitor

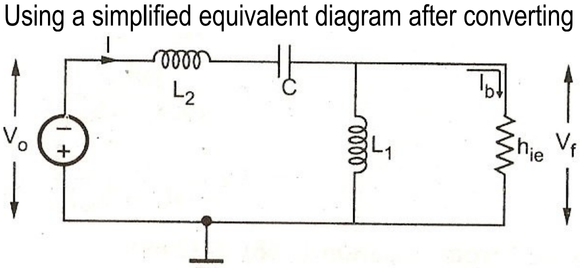
The common emitter amplifier provides a phase shift of 180^0. As the emitter is grounded, base and collector voltages are out of phase by 180^0. As the center of L_1 *and* L_2 is grounded, when the upper end becomes positive, the lower becomes negative and vice versa, so the LC feedback network gives an additional phase shift of 180^0 required to satisfy the necessary conditions for oscillation.

Derivation of Frequency of Oscillations



The output current is the collector current $I_C = h_{fe} I_b$ where I_b is the base current, h_{ie} the input impedance of the transistor. The output of the feedback is I_b, which is the input current of the transistor.

Using a simplified equivalent diagram after converting current source into voltage source;



$$V_0 = h_{fe} I_b X_{L2} = h_{fe} I_b j\omega L_2$$

L_1 *and* h_{fe} are in parallel, then the current I drawn from the supply

$$I = \frac{-V_0}{\left[X_{L2} + X_C\right] + \left[X_{L1} \| h_{ie}\right]}$$

NB – The –ve sign indicates current direction shown in opposite to the polarities of V_o

Substituting $X_{L2} + X_C = j\omega L_2 + \frac{1}{j\omega C}$ and $X_{L1} \| h_{ie} = \frac{j\omega L_1 h_{ie}}{j\omega L_1 + h_{ie}}$ in the current equation

$$I = \frac{-h_{fe} I_b j\omega L_2}{\left[j\omega L_2 + \frac{1}{j\omega C}\right] + \left[\frac{j\omega L_1 h_{ie}}{j\omega L_1 + h_{ie}}\right]} \qquad \text{(i)}$$

According to current division in parallel circuits $I_b = I \times \frac{X_{L1}}{X_{L1} + h_{ie}} = I \times \frac{j\omega L_1}{j\omega L_1 + h_{ie}}$ (ii)

Substituting the value of I and rationalizing the denominator we get;

$$I = \frac{\omega^4 h_{fe} L_1^2 L_2 C\left(1 - \omega^2 L_2 C\right) + j\omega^3 h_{fe} L_1 L_2 C\left[h_{ie} - \omega^2 C h_{ie}\left(L_1 + L_2\right)\right]}{\left[h_{ie} - \omega^2 C h_{ie}\left(L_1 + L_2\right)\right]^2 + \omega^2 L_1^2\left(1 - \omega^2 L_1^2 L_2 C\right)^2} \qquad \text{(iii)}$$

To satisfy the conditions, the imaginary part of RHS must be equal to zero;

$Therefore\ \ \omega^3 h_{fe} L_1 L_2 C\left[h_{ie} - \omega^2 C h_{ie}\left(L_1 + L_2\right)\right] = 0\ \ Hence\ \ \omega = \frac{1}{\sqrt{C\left(L_1 + L_2\right)}}$

$\Rightarrow f = \frac{1}{2\pi\sqrt{C\left(L_1 + L_2\right)}}$ - The frequency of oscillation. If $L_1 + L_2 = L_{eq}\ \ then\ f = \frac{1}{2\pi\sqrt{CL_{eq}}}$

Equating the value of $\omega = \frac{1}{\sqrt{C\left(L_1 + L_2\right)}}$ in eq, (iii) when the imaginary part is 0 the value of h_{fe} required to satisfy the oscillating conditions is $h_{fe} = \frac{L_1}{L_2}$. For mutual inductance of M $h_{fe} = \frac{L_1 + M}{L_2 + M}$

Example

In a transistorized Hartley oscillator two inductors are 2mH and 20μHwhile the frequency is to be varied from 950 kHz to 2050 kHz. Calculate the range over which the capacitor is to be varied.

Solution

The frequency is given by $f = \frac{1}{2\pi\sqrt{CL_{eq}}}\ where\ L_{eq} = L_1 + L_2$

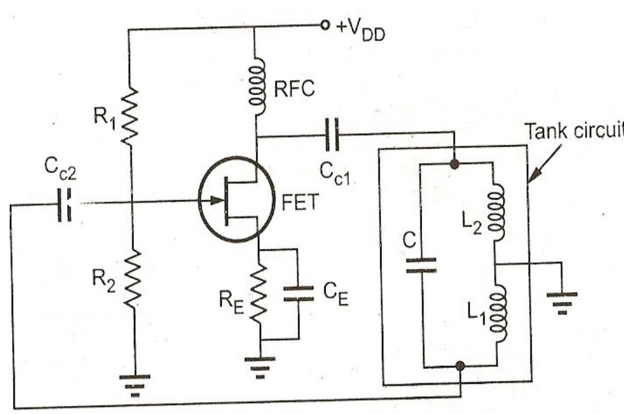
$$L_{eq} = 2\times10^{-3} + 20\times10^{-6} = 0.00202$$

$$When\ f = f_{max}\ ; 2050\times10^3 = \frac{1}{2\pi\sqrt{C\times0.00202}} \Rightarrow C = 2.98pF$$

$$When\ f = f_{min}\ ; 950\times10^3 = \frac{1}{2\pi\sqrt{C\times0.00202}} \Rightarrow C = 13.89pF$$

FET Hartley Oscillator

A FET is used as an active device in an amplifier stage. The resistors R_1, R_2 bias the FET along with R_3 to maintain Q point stable, coupling capacitors C_{C1}, C_{C2} with larger values compared to C are used.



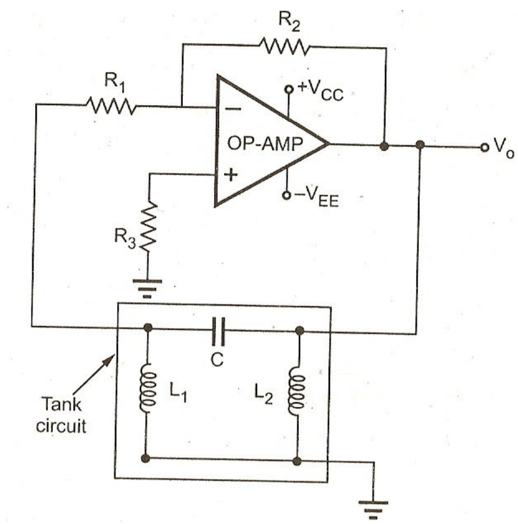
Refer; $X_1 + X_2 + X_3 = 0\ and\ X_1 = j\omega L_1, X_2 = j\omega L_2\ and\ X_3 = \frac{1}{j\omega LC}$

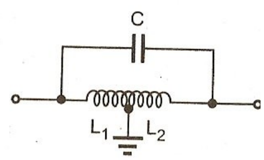
Solving for ω we get $f = \frac{1}{2\pi\sqrt{CL_{eq}}}$; $where\ \ L_{eq} = L_1 + L_2\ or\ L_1 + L_2 + 2M$

If $L_1 = L_2 = L$, then the frequency of oscillations becomes $f = \frac{1}{2\pi\sqrt{CL}}$

Hartley Oscillator using Op-amp

The amplifying stage uses an op-amp as an active device





The frequency of oscillations $f = \dfrac{1}{2\pi\sqrt{CL_{eq}}}$ $where L_{eq} = L_1 + L_2 \; or \; L_1 + L_2 + 2M$

For oscillations, the amplifier gain $A_V = \dfrac{L_1}{L_2}$

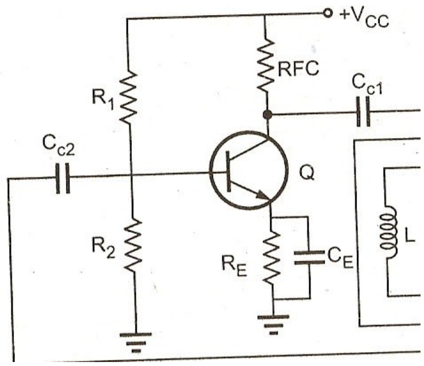
If mutual inductance exists between L_1 and L_2, then $A_V = \dfrac{L_1 + M}{L_2 + M}$

B. Analysis of Colpitt's Oscillator,

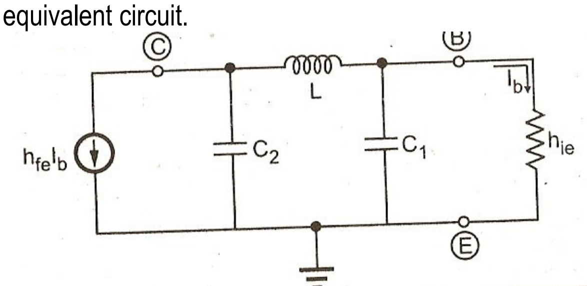
A Colpitts oscillator is a high frequency LC oscillator (between 1 MHz – 500 MHz) which uses two capacitive reactances and one inductive reactance in the tank circuit i.e. feedback network.

- **Transistorized Colpitts Oscillator;**

The amplifier stage uses a transistor in common emitter configuration as an active device which causes a phase shift of 180^0 and LC feedback network consisting of one inductor L and two capacitors $C_1 \; and \; C_2$, which adds further 180^0 phase shift to satisfy the oscillating conditions.

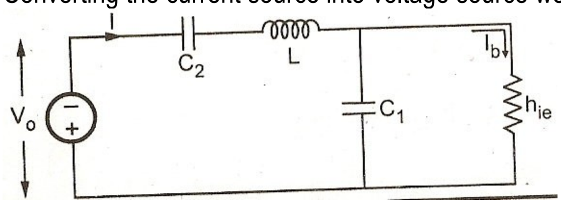


The output current $I_C = h_{fe} I_b$ acts as the input to the feedback network. While the base current I_b acts as the output current of the tank circuit, flowing through the input impedance h_{ie} of the amplifier, hence the equivalent circuit.



Derivation of frequency of oscillation

Converting the current source into voltage source we get a simplified equivalent circuit.



$V_o = h_{fe} I_b X_{C2} = h_{fe} I_b \dfrac{1}{j\omega C_2}$ .Total current drawn from supply $I = \dfrac{-V_o}{[X_{C2} + X_L] + [X_{C1} \| h_{ie}]}$

NB: (-ve sign: Current direction is assumed opposite to voltage V_o polarity)

$$X_{C2} + X_L = \frac{1}{j\omega C_2} + j\omega L \text{ and } X_{C1} \| h_{ie} = \frac{\frac{1}{j\omega C_1} \times h_{ie}}{\frac{1}{j\omega C_1} + h_{ie}} \Rightarrow \quad I = \frac{-h_{fe} I_b \frac{1}{j\omega C_2}}{\left[\frac{1}{j\omega C_2} + j\omega L\right] + \left[\frac{\frac{h_{ie}}{j\omega C_1}}{h_{ie} + j\omega C_1}\right]};$$

(i)

According to current division in parallel circuits $I_b = I \times \dfrac{X_{C1}}{X_{C1} + h_{ie}} = \dfrac{I \times \frac{1}{j\omega C_1}}{\frac{1}{j\omega C_1} + h_{ie}}$; $\qquad (ii)$

Substituting value of I from $eq^n\,(i)$ in $eq^n\,(ii)$ and replacing $(j\omega)^2 = -\omega 2;\ and\ (j\omega)^3 = -j\omega^3$

$$1 = \frac{-h_{fe}}{\left(1-\omega^2 LC_2\right) + j\omega h_{ie}\left[C_1 + C_2 - \omega^2 LC_1C_2\right]}; \qquad (iii)$$

To satisfy Barkhausen criterion, the imaginary part of the RHS denominator must be equal to zero.

i.e. $\omega h_{ie}\left[C_1 + C_2 - \omega^2 LC_1C_2\right] = 0 \Rightarrow C_1 + C_2 - \omega^2 LC_1C_2 = 0 \;\; hence\; \omega^2 = \dfrac{C_1 + C_2}{LC_1C_2}$

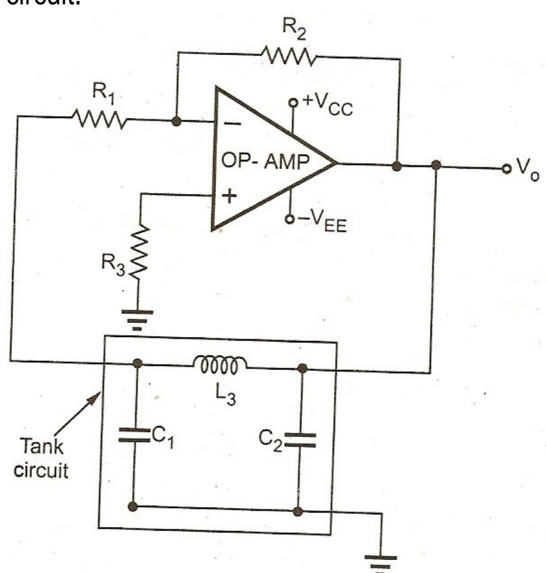
$$\omega = \frac{1}{\sqrt{LC_{eq}}} \Rightarrow f = \frac{1}{2\pi\sqrt{LC_{eq}}} \;\; where\;\; C_{eq} = \frac{C_1C_2}{C_1 + C_2}$$

$f = \dfrac{1}{2\pi\sqrt{LC_{eq}}}$ - Frequency of oscillation of the Colpitts oscillator.

Substituting the value of $f = \dfrac{1}{2\pi\sqrt{LC_{eq}}}$ in $eq^n\,(iii)$ and equating the magnitudes of both sides, the restriction of the value h_{fe} is obtained as, $h_{fe} = \dfrac{C_2}{C_1}$

- **Colpitts Oscillator using Op-amp**

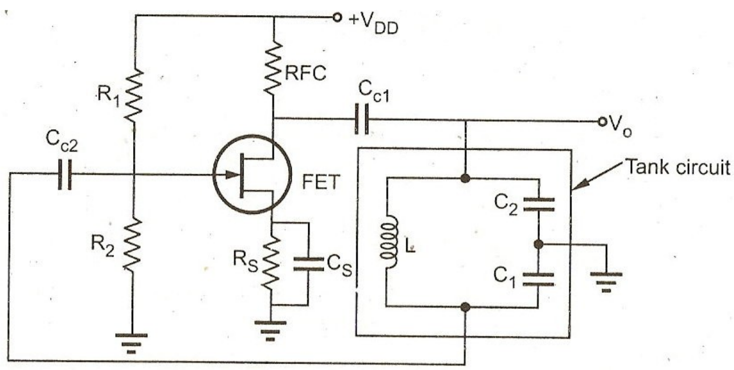
The Op-amp is used for an amplifier stage while the tank circuit remains as in a transistorized circuit.



The oscillator frequency $f = \dfrac{1}{2\pi\sqrt{LC_{eq}}}$; the condition of gain of the op-amp is now applicable.

- **Colpitts Oscillator using** FET

The FET is used as an active device in the amplifier stage, the tank circuit remain the same and the oscillating frequency also remain the same.



Example:

Find the frequency of oscillation of a transistorized Colpitts oscillator having $C_1 = 150pF,\ C_2 = 1.5nF\ and\ L = 50\mu H$

Solution

$$C_{eq} = \frac{C_1 C_2}{C_1 + C_2} = \frac{150\times10^{-12}\times1.5\times10^{-9}}{150\times10^{-12}+1.5\times10^{-9}} = 136.363\ pF$$

$$f = \frac{1}{2\pi\sqrt{LC_{eq}}} = \frac{1}{2\pi\sqrt{150\times10^{-12}\times136.363\times10^{-12}}} = 1.927\ MHz$$

L – 13. Analysis of Clapp and Armstrong Oscillators,

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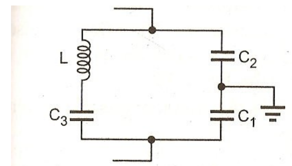
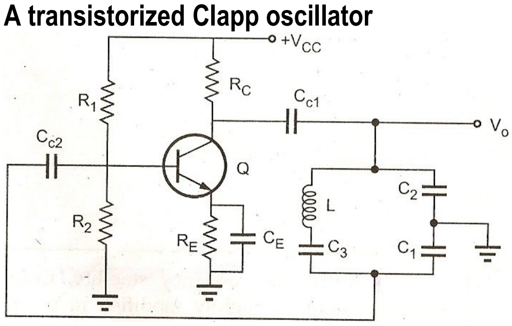
A. **Analysis of Clapp Oscillator;**

To achieve the frequency stability, the Colpitts oscillator circuit is slightly modified to get Clapp oscillator such that the basic tank circuit with two capacitive reactances and one inductive reactance remains the same, but one more capacitor C_3 is introduced in series with the inductance L.



The value of C_3 is much smaller than the values of C_1 and C_2, hence the equivalent capacitance becomes; $\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3} \approx \frac{1}{C_3}$. In practice the values C_1 and C_2 are neglected thus $C_3 = C_{eq}$ and the frequency of oscillation is given by $f = \frac{1}{2\pi\sqrt{LC_3}}$,

A transistorized Clapp oscillator



Derivation of frequency of oscillations

Advantages of the Clapp oscillator;

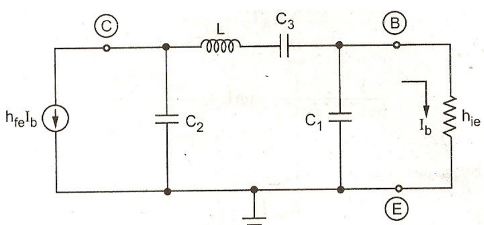
- Frequency stability;

There is no transistor parameter, like stray capacitance across C_3, hence the frequency is stable and accurate,

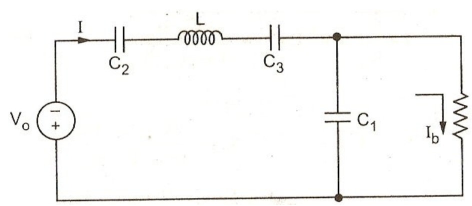
- The frequency can be varied in the desired range;

C_3 can be kept variable, and since the frequency depends on C_3, the frequency can be varied in the desired range

Derivation is similar to the Colpitts oscillator with C_3 in series with L in the equivalent circuit of the transistorized Clapp oscillator.



Equivalent circuit



Simplified equivalent circuit with current source converted to voltage source

$V_o = h_{fe} I_b X_{C2} = h_{fe} I_b \frac{1}{j\omega C_2}$ Total current drawn from supply $I = \frac{-V_o}{[X_{C2} + X_{C3} + X_L] + [X_{C1} \| h_{ie}]}$

NB: (-ve sign: Current direction is assumed opposite to voltage V_o polarity)

$X_{C2} + X_{C3} + X_L = \frac{1}{j\omega C_2} + \frac{1}{j\omega C_3} + j\omega L$ and $X_{C1} \| h_{ie} = \frac{\frac{1}{j\omega C_1} \times h_{ie}}{\frac{1}{j\omega C_1} + h_{ie}}$

Therefore $$I = \frac{-h_{fe} I_b \dfrac{1}{j\omega C_2}}{\left[\dfrac{1}{j\omega C_2} + \dfrac{1}{j\omega C_3} + j\omega L\right] + \left[\dfrac{\dfrac{h_{ie}}{j\omega C_1}}{h_{ie} + j\omega C_1}\right]}; \boldsymbol{(i)}$$

According to current division in parallel circuits $I_b = I \times \dfrac{X_{C1}}{X_{C1} + h_{ie}} = \dfrac{I \times \dfrac{1}{j\omega C_1}}{\dfrac{1}{j\omega C_1} + h_{ie}}$; $\boldsymbol{(ii)}$

Substituting value of I from $eq^n(i)$ in $eq^n(ii)$ and replacing $(j\omega)^2 = -\omega 2;\ and\ (j\omega)^3 = -j\omega^3$

$$1 = \frac{-h_{fe} C_3}{\left(C_2 C_3 - \omega^2 L C_2 C_3\right) + j\omega h_{ie}\left[\left\{C_1 C_2 + C_2 C_3 + C_3 C_1\right\} - \omega^2 L C_1 C_2 C_3\right]}; \quad \boldsymbol{(iii)}$$

To satisfy Barkhausen criterion, the imaginary part of the RHS denominator must be equal to zero.
i.e.

$$\omega h_{ie}\left[C_1 C_2 + C_2 C_3 + C_3 C_1 - \omega^2 L C_1 C_2 C_3\right] = 0 \Rightarrow C_1 C_2 + C_2 C_3 + C_3 C_1 - \omega^2 L C_1 C_2 C_3 = 0$$

$hence\ \omega^2 = \dfrac{C_1 C_2 + C_2 C_3 + C_3 C_1}{L C_1 C_2 C_3}$. Therefore $\omega^2 = \dfrac{\dfrac{1}{C_1} + \dfrac{1}{C_2} + \dfrac{1}{C_3}}{L} = \dfrac{\dfrac{1}{C_{eq}}}{L}$

$\omega^2 = \dfrac{1}{L C_{eq}} \Rightarrow \omega = \dfrac{1}{\sqrt{L C_{eq}}}\quad Hence\ f = \dfrac{1}{2\pi\sqrt{L C_{eq}}}$ But as

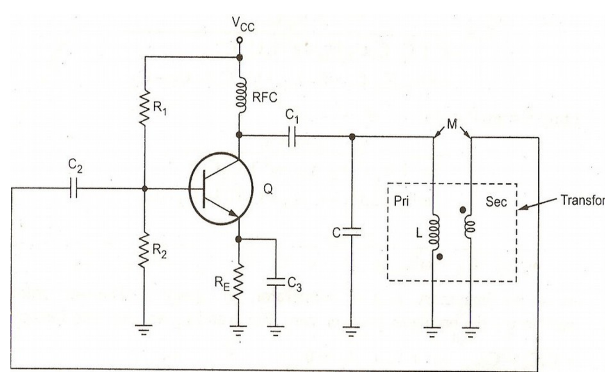
$C_3 << C_1;\ and\ C_3 << C_2 \Rightarrow C_{eq} = C_3;$

$Hence\ f = \dfrac{1}{2\pi\sqrt{L C_3}}$ The required frequency of oscillation of the Clapp oscillator

B. **Analysis of Armstrong Oscillator;**

In the Armstrong Oscillator a transformer is used, whose primary winding acts as L in the circuit while the voltage across secondary winding is used as a feedback.

C_1 and C_2 are coupling capacitors, the collector of transistor Q drives the LC resonating circuit with primary of transformer acting as L. The feedback signal is taken from the secondary winding which is small and given to the base of the transistor Q.



There is mutual inductance M between primary and secondary of the transformer which introduces a phase shift of 180^0, as indicated by the dots, in addition to the phase shift of 180^0 by transistor Q. Thus overall phase shift around the loop is 360^0 which satisfies the Barkhausen criterion.

The feedback fraction $\beta = \dfrac{M}{L}$ For oscillations to start $\beta A > 1$, hence $\dfrac{1}{\beta} > \dfrac{L}{M}$ or $A_{\min} = \dfrac{L}{M}$.

The frequency of sustained oscillations $f_r = \dfrac{1}{2\pi\sqrt{LC}}$

Disadvantages:

- The frequency of sustained oscillations is dependent on primary winding inductance L and C across it,
- Transformers are generally avoided due to their size and cost.

Frequency Stability of Oscillators

Meaning;

- Frequency stability of an oscillator is the measure of the ability of an oscillator to maintain the desired frequency as precisely as possible for as long a time as possible

Sources of instability (or factors that affect the frequency stability of an oscillator);

- In a transistorized Colpitts and Hartley oscillator, the base-collector junction is reverse biased, and there exists an internal capacitance which is dominant at high frequencies. This capacitive effect in transistor and stray capacitances affect the value of capacitance in the tank circuit and hence the oscillating frequency.
- Tank circuit component parameters are temperature sensitive. Changes in values of inductors and capacitors due to changes in temperature are the main cause due to which frequency does not remain stable.
- Active device parameters such as BJT, FET are temperature sensitive. As temperature changes the parameters get affected in-turn affect the oscillating frequency.
- Variation in power supply,
- Changes in atmospheric conditions, aging and unstable transistor parameters,
- Changes in the load connected, affect the effective resistance of the tank circuit,

Methods of improving the frequency stability

The frequency stability of an oscillator can be improved by the following modifications:

- Enclosing the circuit in a constant temperature chamber,
- Maintaining constant voltage by using Zener diodes,
- Couple the oscillator loosely to reduce the load effect.

L – 14. Quartz crystal construction and its electrical equivalent circuit,

Date: ……………..…

Hall: 21 Period: ……

Crystals are naturally occurring or synthetically manufactured compounds exhibiting **piezoelectric effects.**

Meaning of Piezoelectric Effect

- A phenomenon by which under the influence of mechanical pressure, a voltage gets generated across the opposite faces of the crystal and vice versa.

  i.e.

  - If a mechanical force is applied so as to force the crystal to vibrate, an $a.c.$ voltage gets generated across it, or conversely,
  - If a crystal is subjected to $a.c.$ voltage it vibrates.

Every crystal (depending on the cut) has its own resonating frequency, so under mechanical vibrations a crystal generates an electric signal of constant frequency. A crystal has greater stability in holding the constant frequency

Types of crystals exhibiting piezoelectric effect;

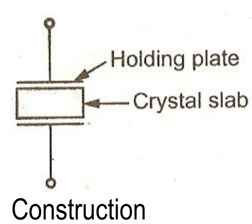
Types of crystal	Piezoelectric activities	Mechanical strength	Cost	Area of application
Rochelle Salts	Greatest	Weakest and can easily break	Intermediate	Microphones Headsets Loudspeakers
Quartz	Intermediate	Intermediate	Inexpensive	Transmitters Receivers RF oscillators RF Filters Watches
Tourmaline	Least	Strongest	Expensive	Rare in practice

- Rochelle Salts have the greatest piezoelectric activity, i.e. for a given $a.c.$ voltage they vibrate more than Quartz or Tourmaline
- Quartz is inexpensive and easily available in nature and hence very commonly used in crystal oscillators

A crystal oscillator is a tuned circuit oscillator using a piezoelectric crystal as its resonant tank circuit.

Constructional details

A quartz crystal is naturally hexagonal prism, but for its practical use it is cut to a rectangular slab. The slab is then mounted between two metal plates.



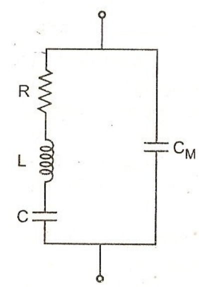
Construction

$A.C.$ Equivalent Circuit

- When the crystal is not vibrating, is equivalent to the capacitance due to the mechanical mounting of the crystal – mounting capacitance C_M

- When it is vibrating it experiences internal frictional losses – resistance R , while the mass of the crystal indicating its inertia being represented by an inductance L . In vibrating condition, it is having some stiffness, which can be represented by a capacitor C .

Let C_M - Mounting capacitance
The capacitance existing due to metal separation by dielectric-like crystal slab in shunt
R - Internal frictional losses due to vibrations,
L - Inertia due to the mass of the vibrating crystal,
C - Capacitance indicating stiffness in vibrating conditions.



Equivalent Circuit

$R\ L\ C$ forms a resonating circuit

$$f_r = \frac{1}{2\pi\sqrt{LC}}\sqrt{\frac{Q^2}{1+Q^2}}\ ;\ Where\quad Q = \frac{\omega L}{R} \text{ - quality factor of crystal}$$

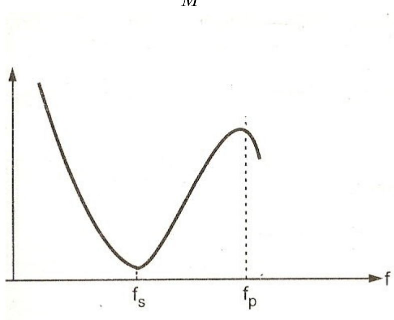
Since the quality factor of a crystal is very high $\sqrt{\frac{Q^2}{1+Q^2}} \approx 1$, hence $f_r = \frac{1}{2\pi\sqrt{LC}}$

The crystal frequency is inversely proportional to the thickness of the crystal $f\ \alpha\ \frac{1}{t}$.

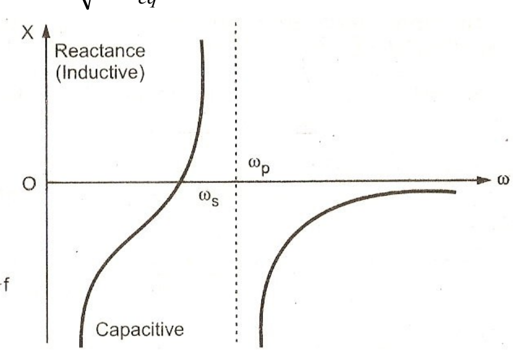
So to have very high frequencies, thickness of the crystal must be very small which makes the crystal mechanically weak and may get damaged under vibrations. Practical crystal oscillators are used up to $300\ kHz$

Series and Parallel Resonance

- Series resonance occurs when the reactances of the series $R\ L\ C$ leg are equal, i.e. $X_C = X_L$ hence $f_s = \frac{1}{2\pi\sqrt{LC}}$
- Parallel resonance (or antiresonance condition) occurs when the reactances of the series resonant leg equals the reactance of the mounting capacitance C_M . Under parallel resonance $C_{eq} = \frac{C_M C}{C_M + C}$ Hence $f_p = \frac{1}{2\pi\sqrt{LC_{eq}}}$



Crystal impedance Vs Frequency



Reactance Vs Frequency

Factors affecting frequency stability of the crystal are:

- Temperature stability (change in frequency per degree change in temperature). When greater stability is required a crystal is kept in an oven (constant temperature box)

- Aging of the crystal material – long term stability. (Aging rate $= 2 \times 10^{-8}$ per year - negligible)
- Frequency drift with time – short term stability $\approx 0.0001\%$ per day.

Crystal Oscillators

1. **Pierce Crystal Oscillator**

   Pierce Crystal Oscillator is a modification of Colpitts Oscillator circuit whereby a crystal has replaced the inductor in the tank circuit.

Note: A crystal behaves as an inductor for frequencies slightly higher than the series resonance frequency

Functions of amplifier stage components

R_1 *and* R_2 - biasing resistance;

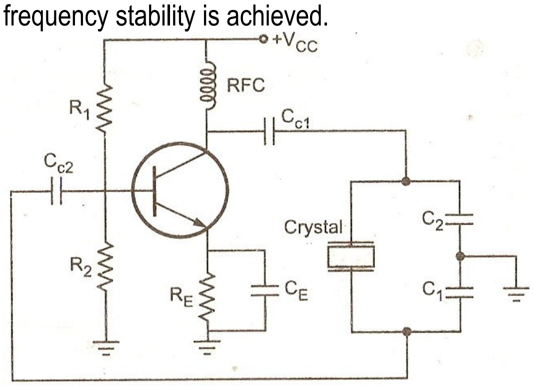
RFC (Radio frequency choke)- isolation between a.c. and d.c..

R_E -emitter circuit biasing resistor;

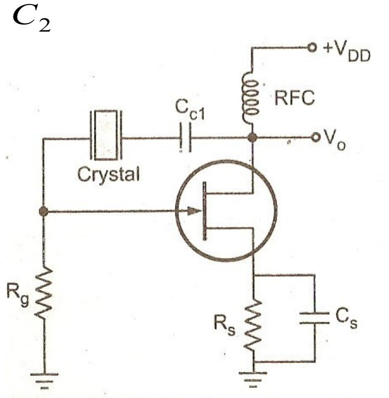
C_E -emitter bypass capacitor;

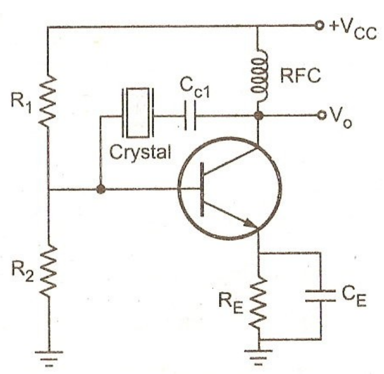
C_1 *and* C_2 -coupling capacitor

The resulting circuit frequency is set by the series resonating frequency of the crystal. Change in supply voltage, temperature and transistor parameters have no effect on the circuit operating condition hence good frequency stability is achieved.



Oscillator circuits can be modified by using the internal capacitance of transistor used instead of C_1 and C_2





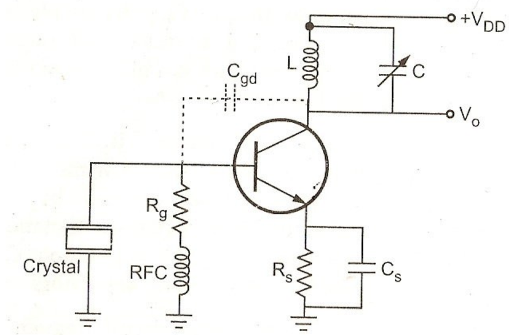
Pierce Crystal Oscillator using FET   Pierce Crystal Oscillator using transistor

2. **Miller Crystal Oscillator**

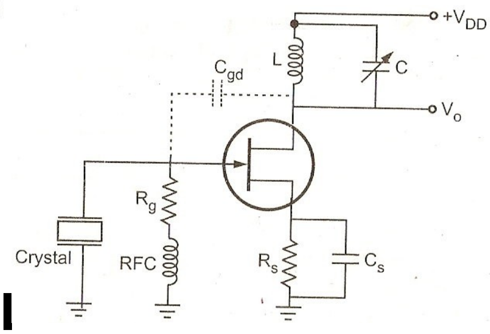
   Mille Crystal Oscillator is a modification of Hartley Oscillator circuit whereby one of the inductors in the tank circuit has been replaced by a crystal. A crystal behaves as an inductor for frequencies slightly higher than the series resonance frequency.

The tuned circuit of L and C is off-tuned to behave as an inductor L_1 . The crystal behaves as another inductance L_2 between base and ground. The internal capacitance of the transistor acts as a capacitor to fulfill the elements of the tank circuit. The crystal decides the operating frequency of the oscillator .

Transistorized Miller Crystal Oscillator



FET Miller Crystal Oscillator



Example

A crystal has $L = 2H;\ \ C = 0.01pF;\ and\ R = 2\ k\Omega$. Its mounting capacitance is $2\ pF$.

Calculate its (i) Series resonating frequency, (ii) Parallel resonating frequency

Solution

(i) $f_s = \dfrac{1}{2\pi\sqrt{LC}} = \dfrac{1}{2\pi\sqrt{2\times 0.01\times 10^{-12}}} = 1.125MHz$

(ii)

$$f_p = \frac{1}{2\pi\sqrt{LC_{eq}}} = \left| C_{eq} = \frac{(2\times 0.01)\times 10^{-12}}{(2+0.01)\times 10^{-12}} = 9.95\times 10^{-15} \right| = \frac{1}{2\pi\sqrt{2\times 9.95\times 10^{-15}}} = 1.128MHz$$

Frequency range of RC and LC oscillators

RC oscillators are widely used in audio frequency range

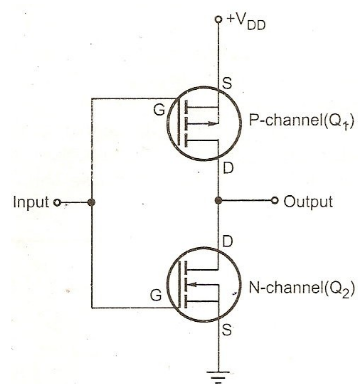
LC oscillators are commonly used higher radio frequency range

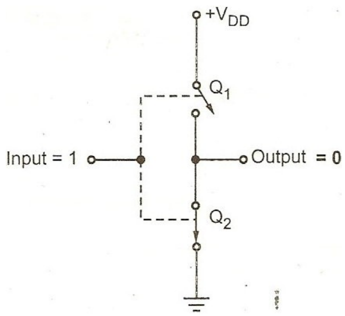
L – 15. Use of Logical Gates as Linear amplifiers,

Date: ……………..…
Hall: 21 Period: ……

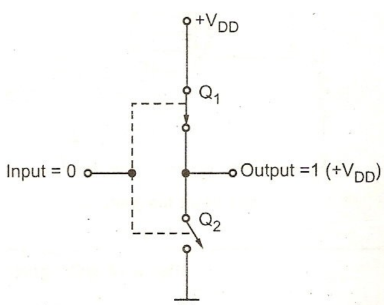
Introduction – The basic CMOS inverter

The basic CMOS inverter circuit consists of two MOSFETs ($PMOS$ transistors and $NMOS$ devices) in series in such a way that the $P-channel$ device has its source connected to $+V_{DD}$ and the $N-channel$ device has its source connected to the ground. The gates of the two devices are connected together to form common input and the drains are connected together to form common output.





When input is HIGH, the gate of Q_1 ($P-channel$) is at 0V relative to the source of Q_1, i.e $V_{GS1} = 0V$. Thus $Q_1 - OFF$ By then the gate of Q_2 ($N-channel$) is at $+V_{DD}$ relative to its source, i.e. $V_{GS2} = +V_{DD}$. Thus $Q_2 - ON$. This will produce $V_{OUT} \approx 0V$



When input is LOW, the gate of Q_1 ($P-channel$) is at negative potential relative to the source, while Q_2, has $V_{GS} = 0V$. Thus $Q_1 - ON$ and $Q_2 - OFF$. This will produce $V_{OUT} \approx +V_{DD}$

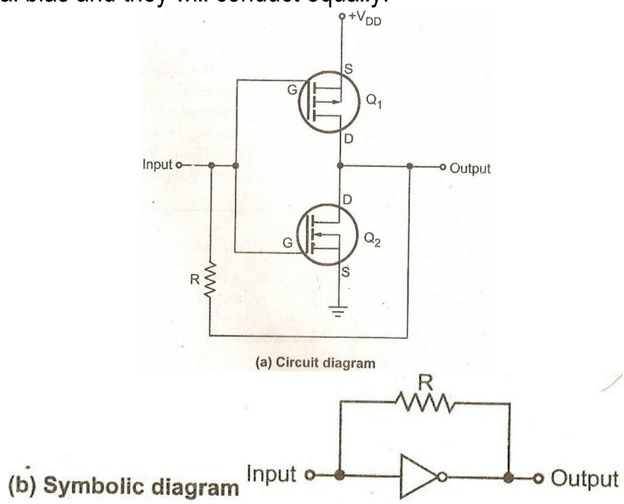
Truth Table of Inverter operating in ON - OFF

Input	Q_1	Q_2	Output
0	ON	OFF	1
1	OFF	ON	0

When the MOSFETs are operating in a linear region of their characteristics, the circuit becomes a **Linear Amplifier.**

A simple method of biasing an inverter circuit into its linear region

A resistor R connected between the output and input to provide negative feedback. This resistor sets the output d.c. voltage equal to $\frac{V_{DD}}{2}$ which is feedback to the common gate input. Due to this voltage both MOSFETs will have equal bias and they will conduct equally.



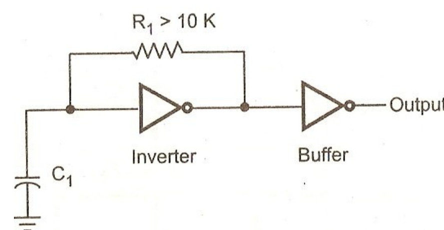
(a) Circuit diagram

(b) Symbolic diagram

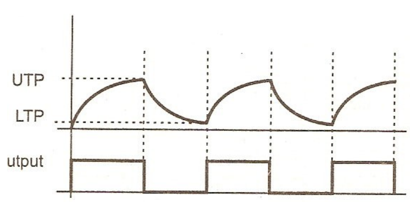
If a sinusoidal signal is applied at the input, then the output will vary linearly with respect to the input with a phase shift of 180^0 .

Oscillators and Clock generators circuits using Logic Gate Amplifiers

Using two Schmitt Inverters (inverters having upper and lower threshold levels) and RC components, the waveform shows that the input to the inverter 1 rises exponentially at a rate set by the components R_1C_1 . These components determine the ON and OFF time for each part cycle of the clock.



The frequency of the circuit is given by $f = \frac{0.53}{R_1C_1}$



Consider that initially input of $Inverter - 1$ is at $Logic - 0$ and hence the output at $Logic - 1$. As input is at $Logic - 0$, capacitor C_1 is not charged and it starts charging trough resistor R_1 . When capacitor C_1 charges above the upper threshold level (UTP) input is considered as a $Logic - 1$ and output switches to $Logic - 0$. This forces the capacitor C_1 to discharge through resistor R_1 . When capacitor C_1 discharges below the lower threshold level (LTP) input is considered as a $Logic - 0$ and output switches to $Logic - 1$.

Advantages
- Simple,
- Temperature stability

Disadvantages
- Frequency limited
- No duty cycle control

Questions for Revision

PART - A

Que 1. (i) Define an oscillator (ii) Give the classification of oscillators.
Que 2. State the Barkhausen Criterion for oscillation in feedback oscillators.
Que 3. Why is an RC phase shift needed in RC phase shift oscillators?
Que 4. Why is an LC oscillator preferred over RC oscillator at radio frequency?
Que 5. Draw a circuit diagram of a Wien Bridge oscillator using an op-amp.
Que 6. Brief about frequency ranges of RC and LC oscillators

PART - B

Que 1. What do you understand by piezoelectric effect?
Que 2. List (i) two advantages and (ii) one disadvantage of crystal oscillators.
Que 3. (i) What is a twin-T filter? (ii) is it a notch filter or not? (ii) explain how it is used to obtain twin-T oscillators.
Que 4. What are the advantages and disadvantages of a Wien Bridge oscillator?
Que 5. Given various LC oscillator types, show the reactances X_1 , X_2 and X_3

Oscillator type	Reactance elements in the tank circuit		
	X_1	X_2	X_3
Hartley Oscillator			
Colpitts Oscillator			

Que 6. Mention the expression for frequency of oscillations of (i) Colpitts oscillator and (ii) Hartley oscillator.
Que 7. Explain the working of a Pierce crystal oscillator.

PART - C

Que 1. (i) Determine whether the circuit in Fig.1 will work as an oscillator or not.
(ii) If yes determine the frequency of the oscillator

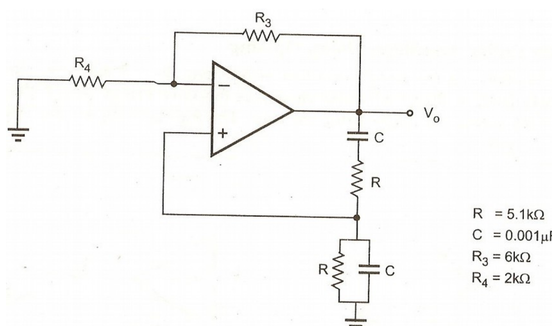


Fig1.

Que 2. Explain (i) where does the starting voltage of an oscillator come from, and (ii) how is the amplitude stabilized?
Que 3. The sensitive arm of the Wien Bridge oscillator uses $C_1 = C_2 = 0.001\mu F$ and $R_1 = 10\Omega$ while R_2 is kept variable. The frequency is to be varied from $10kHz - 50Hz$ by varying R_2 . Find the (i) minimum and (ii) maximum value of R_2
Que 4. Calculate the component values of the Wien Bridge suitable to be used in the oscillator to vary the frequency from $100Hz - 10kHz$ in the two ranges; given $R_1 = R_2 = R = 50\Omega$
Que 5. Identify and explain at least four factors that affect the stability of an oscillator.
Que 6. (i)Draw a circuit diagram of a Clapp oscillator and (ii) explain how frequency stability can be improved in Clapp oscillators.
Que 7. (i) Explain the concept of operation of LC tank circuits and (ii) show how sustained oscillations are achieved.

UNIT 3. AMPLIFIERS

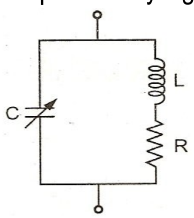
L – 16 Coil losses, Unloaded and loaded Q of tank circuits,

Date: ……………..…
Hall: 21 Period: ……

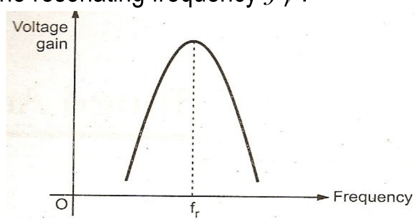
To amplify a selective range of frequencies, the resistive load, R_C is replaced by a tuned circuit, thus being capable of amplifying a signal over a narrow band of frequencies centered at the resonating frequency f_r. At resonance the tuned circuit is high impedance, while at all other frequencies it is a low impedance therefore only signals at the resonance frequency will be amplified.

Definition:

A tuned amplifier is an amplifier in which the resistive load R_C is replaced by a tuned circuit, and it amplifies only signals centered at the resonating frequency f_r.



Tuned Circuit



Frequency Response

At resonance, inductive and capacitive effects of the tuned circuit cancel each other i.e. $X_L = X_C$

$$Z \approx L \quad \Leftarrow \quad f_r \quad \Rightarrow \quad Z \approx C \qquad f$$

For frequencies above resonance, the circuit is capacitive,
For frequencies below resonance, the circuit is inductive,

Hence $f_r = \dfrac{1}{2\pi\sqrt{LC}}$ and $Z_r = \dfrac{L}{RC}$ as a result the tuned circuit is purely resistive at resonance and it can be used as a load for an amplifier.

Coil Losses

The tuned circuit consists of a coil which is not purely inductive; it consists of few losses which can be represented in form of leakage resistance in series with the inductor.

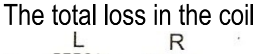
The total loss of the coil is represented;

- Copper loss: - This is equivalent to DC resistance of the coil at low frequency. Copper loss is inversely proportional to frequency; as frequency increases the copper loss decreases.

- Eddy current loss in iron and copper: - This is due to currents flowing within the copper or core cased by induction. The result of eddy current is a loss due to heating within the inductor s' copper or core. Eddy current losses are directly proportional to frequency.

- Hysteresis loss: - Hysteresis loss is proportional to the area enclosed by the hysteresis loop and the rate at which the loop is transversed. Hysteresis loss is independent of frequency.

The total loss in the coil or inductor is represented by inductance in series with leakage resistance of the coil



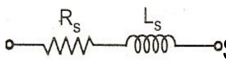
Inductor with leakage resistance

Quality factor Q

- Quality factor Q is a measure of how "pure" or "real" an inductor is. The higher the Q of an inductor, the fewer losses there are in the inductor. Therefore it's a ratio $Q = \dfrac{\text{Reactance}}{\text{Resistance}}$

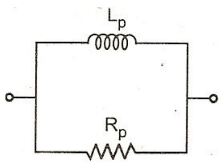
- Quality factor Q is a measure of efficiency with which the inductor can store energy.

The dissipation factor D which is referred to as the total loss within a component is $D = \frac{1}{Q}$



Series circuit

Inductive impedance $\frac{\omega L_S}{R_s}$



Parallel circuit

Inductive admittance $\frac{R_P}{\omega L_P}$

Quality Factor Equations

Quality factor equation $Q = \frac{1}{D} = \frac{\omega L_S}{R_s} = \frac{R_P}{\omega L_P}$

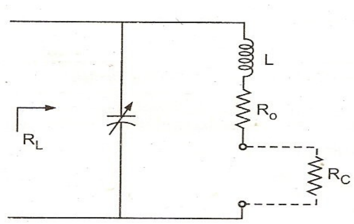
Unloaded and Loaded Q

- Unloaded $Q\ (Q_u)$: - The ratio of stored energy to dissipated energy in a resonator;

  $Q_u = \frac{Stored\ Energy}{Dissipated\ Energy} = \frac{X}{R_S}$; Where X – Reactance, R_S – Series Resistance

- Loaded $Q\ (Q_L)$ of the resonator is determined by how tight the resonator is coupled to its terminations.

Consider the circuit



$R\ and\ C$ - represent the tank circuit

R_o - represents the internal circuit losses of the inductor

R_C - represents the coupled-in load

$\therefore Q_u = \frac{\omega_o L}{R_o}$ - unloaded Q and $Q_L = \frac{\omega_o L}{R_c}$ loaded Q

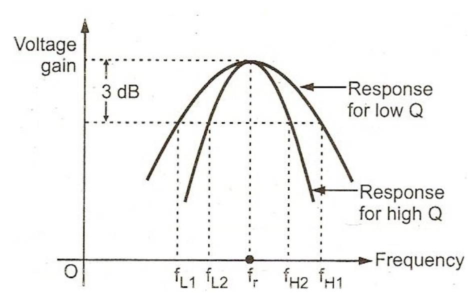
Hence the efficiency of the circuit $\eta = \frac{I^2 R_c}{I^2 (R_c + R_o)} = \frac{Q_u}{Q_u + Q_L} \times 100\%$

For high overall power efficiency;

Coupled-in load R_c must exceed the internal circuit losses of the inductor R_o i.e. $R_c > R_o$

If the Quality factor Q is large the bandwidth is small and the circuit is highly selective. For small values of Q, the bandwidth is high but the selectivity of the circuit is low.

Variation of 3dB bandwidth with variation in Q



Thus in tuned amplifier the Quality factor Q is kept as high as possible to get better selectivity. Such tuned amplifiers are used in communication receivers, broadcasting receivers where it is necessary to amplify only a selected band of frequencies.

Classification of Tuned Amplifiers

In order to get a large overall gain, cascaded multistage tuned amplifiers are used. Multistage tuned amplifiers can be categorized as:

- Single tuned amplifier;
- Double tuned amplifier,
- Stagger tuned amplifier.

According to the coupling used to cascade the stages of the multistage amplifier;

- Capacitive coupled,
- Inductive coupled,
- Transformer coupled

L – 17. Tuned Amplifiers,

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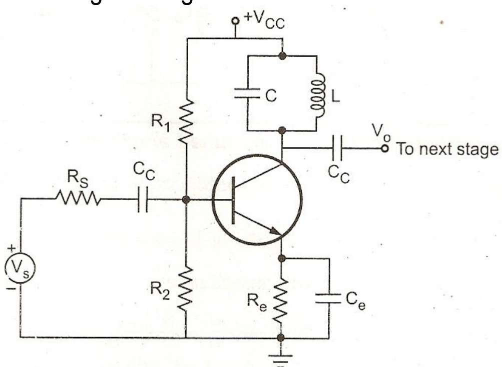
Analysis of Single Tuned Amplifier,

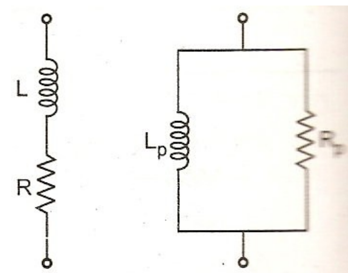
A single tuned amplifier uses one parallel tuned circuit as a load. A single tuned amplifier can be multistage, each stage having one tuned circuit and all tuned circuits tuned to the same frequency.



The diagram shows a single tuned amplifier in CE-configuration, a capacitive coupled transistor amplifier.

- $L \; and \; C$ - form a tuned circuit acting as a collector load and resonating at f_r
- R_1, R_2, R_E, C_E - provide self-biasing for the circuit

Representing the series RL circuit by its equivalent parallel circuit and establishing the conditions for equivalency by equating the admittances of the two circuits;



- Admittance of the series combination of RL: $Y = \frac{1}{R + j\omega L}$

Multiplying the numerator and denominator by $R - j\omega L$

$$Y = \frac{R - j\omega L}{R_2 + (\omega L)^2} = \frac{R - j\omega L}{R_2 + (\omega L)^2} = \frac{R}{R_2 + (\omega L)^2} - \frac{j\omega L}{R_2 + (\omega L)^2} = \frac{R}{R_2 + (\omega L)^2} - \frac{j\omega^2 L}{\omega\left(R_2 + (\omega L)^2\right)}$$

$$Y = \frac{1}{R_p} + \frac{1}{j\omega L_p} \; where \; R_p = \frac{R^2 + \omega^2 L^2}{R}; \; and \; L_p = \frac{R^2 + \omega^2 L^2}{\omega^2 L}$$

Centre (Resonating) Frequency f_r ; - $f_r = \frac{1}{2\pi\sqrt{L_p C_{eq}}}$;

Where $L_p = \frac{R^2 + \omega^2 L^2}{\omega^2 L}$ and $C_{eq} = C_o + C$ being the summation of the transistor output capacitance C_o and the tuned circuit capacitance C .

Quality Factor Q : - $Q = \frac{\omega_r L}{R}$; where ω_r - centre frequency

The quality factor Q of a coil is usually large so that $\omega L >> R$ in the frequency range of operation;

$$\therefore R_p = \frac{R^2 + \omega^2 L^2}{R} = \frac{R^2}{R} + \frac{\omega^2 L^2}{R} = R + \frac{\omega^2 L^2}{R}; \; as \; \frac{\omega^2 L^2}{R} >> 1 \;\Rightarrow\; R_p = \frac{\omega^2 L^2}{R}$$

$$L_p = \frac{R^2 + \omega^2 L^2}{\omega^2 L} = \frac{R^2}{\omega^2 L} + \frac{\omega^2 L^2}{\omega^2 L} = \frac{R^2}{\omega^2 L} + L; \quad as \; \omega L >> R; \quad \Rightarrow L_p \cong L$$

Therefore the Quality Factor at centre frequency $Q_r = \frac{R_p}{\omega_r L_p} = \frac{\omega_r^{\,2} L^2}{R} \times \frac{1}{L} = \frac{\omega_r^{\,2} L}{R}$

Effect of cascading single tuned amplifiers on bandwidth;

In order to obtain a high overall gain, several identical stages of tuned amplifiers can be used in cascade. The overall gain becomes the product of the voltage gains of the individual stages.

If $n-stages$ of single tuned direct coupled amplifiers are connected in cascade, then the bandwidth of $n-stages$ of identical amplifiers;

$$BW_n = BW_1\sqrt{2^{1/n}-1};$$ where BW_1 - bandwidth of a single stage.

Therefore, the bandwidth decreases with increasing number of stages.

L – 18. Tuned Amplifiers (Cont),

Date: ……………..…
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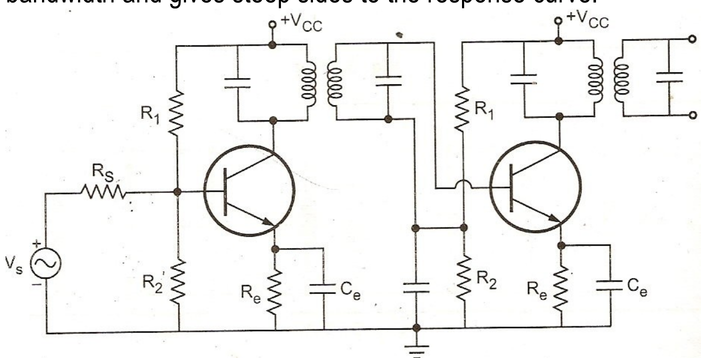
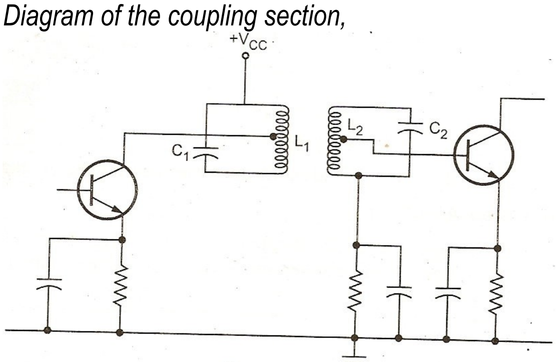
Analysis of Double Tuned Amplifiers

A double tuned amplifier is used in intermediate-frequency amplifier. It operates at a fixed frequency, so the tuning adjustments are set when the transformer is being manufactured. The circuit can provide a wide bandwidth and gives steep sides to the response curve.



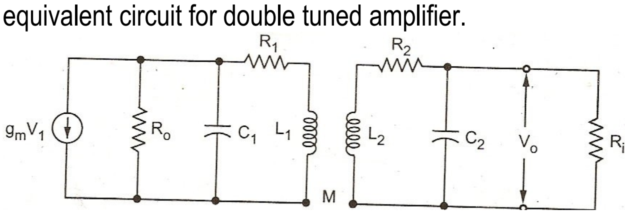
The figure shows a double tuned RF amplifier in CE configuration. The voltage developed across one tuned circuit is coupled inductively to another tuned circuit. Both circuits are tuned to the same frequency.

Diagram of the coupling section,



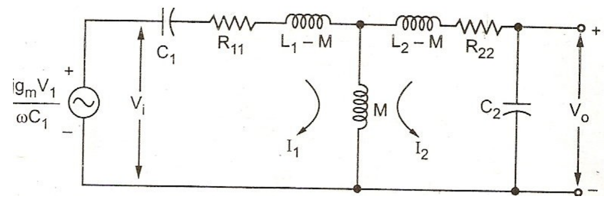
A transistor which is a current source with its output resistance R_o is replaced by a voltage source. C_1 *and* L_1 are tank circuit components of the primary side, while C_2 *and* L_2 represents tank circuit components of the secondary side.

In the *equivalent circuit* the series and parallel resistances are combined into series elements to form an equivalent circuit for double tuned amplifier.



In a *simplified equivalent circuit* diagram the current generator in parallel with C_1 is replaced with a voltage generator in series with C_1 source

$$\therefore R_{11} = \frac{\omega_0^2 L_1^2}{r_1} + R_1 \text{ and } R_{22} = \frac{\omega_0^2 L_2^2}{r_2} + R_2$$

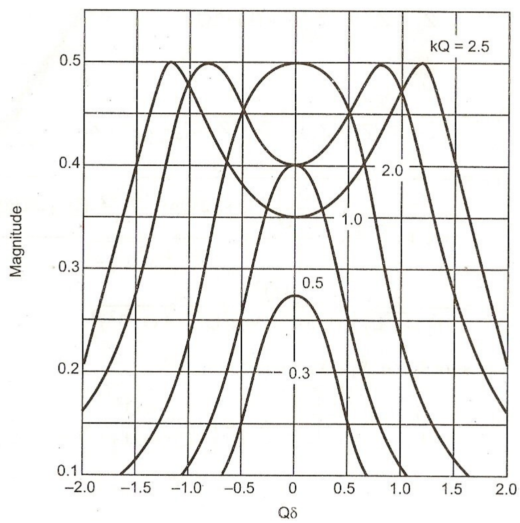


Advantages of tuned amplifiers

- They amplify defined frequencies;
- Signal to noise ration at output is good;
- They are well suited for radio transmitters and receivers
- The band of frequency over which amplification is required can be varied;

Disadvantages;

- Since they inductors and capacitors as tuning elements, the circuit is bulky and costly;
- If the band of frequency is increased, design becomes complex;
- They are suitable to amplify audio frequencies.



L – 19. Stagger Tuned Amplifiers,

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A double tuned amplifier gives a greater bandwidth with steeper sides and flat top, but alignment of a double tuned amplifier is difficult. To overcome this problem two single tuned cascaded amplifiers having certain bandwidth are taken and their resonant frequencies adjusted so that they are separated by an equal interval to the bandwidth of each stage. Since their resonant frequencies are displaced or staggered, they are known as stagger tuned amplifiers.

Advantages of stagger tuned amplifiers;-

- They have better flat, wideband characteristics as compared to single tunes amplifiers.

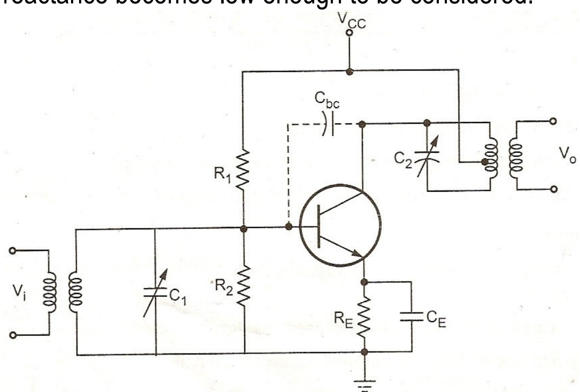
L – 20. Instability of tuned amplifiers and Stabilization techniques,

Date: ……………..…
Hall: 21 Period: ……

Instability of Tuned Amplifiers

In tuned RF amplifiers, transistors are used at frequencies nearer to their unity gain bandwidths to amplify a narrow band of frequencies centered about a radio frequency.

At this frequency, interjunction capacitance between base and collector, C_{bc} of the transistor is high, i.e. its reactance becomes low enough to be considered.



Considering an amplifier in C-E configuration, capacitance C_{bc} come across input and output circuits of an amplifier. As the reactance of C_{bc} at RF is low enough it provides the feedback path from collector to base. If some feedback signal manages to reach the input from the output in a positive manner with proper phase shift, then, there is a possibility of the circuit to be converted to an unstable one, generating its own oscillations hence stopping working as an amplifier.

Such a circuit will oscillate if enough energy is fed back from the collector to the base in the correct phase to overcome circuit losses. Since the conditions for best gain and selectivity are those which also promote oscillations, in order to prevent oscillations in tuned RF amplifiers it is necessary to reduce the stage gain to a level that ensures circuit stability.

Methods of lowering the quality factor Q of tuned circuits:

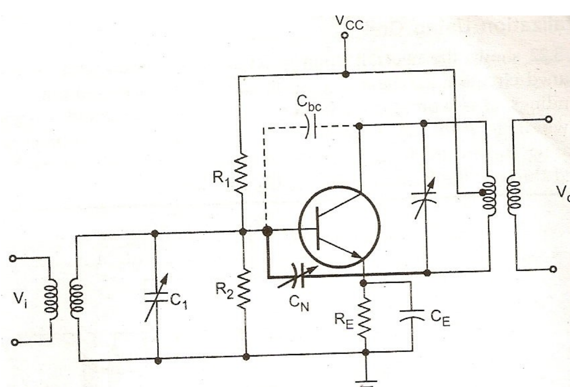
- Stagger tuning,
- Loose coupling between the stages,
- Inserting a "loser" element into the circuit,

These methods of reduced gain, detuning and Q reduction has disadvantageous effects on selectivity.

Stabilization techniques that does not need to loose the circuit performance in order to achieve stability are;-

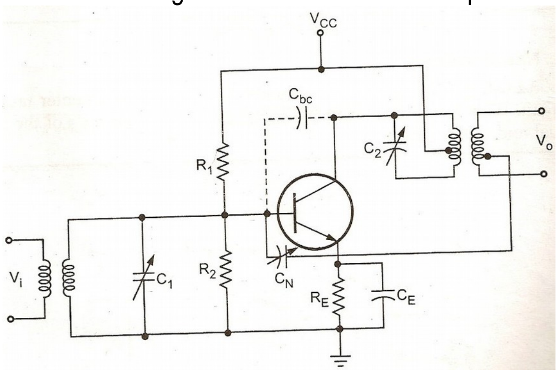
- **Hazeltine neutralization,**

Hazeltine neutralization is achieved by feeding back a portion of the output signal to the input in such a way that it has the same amplitude as the unwanted feedback but the opposite phase. This is done by connecting a small value of variable capacitance C_N from the bottom of the coil, point B, to the base. The neutralization capacitor can be adjusted correctly to completely nullify the signal fed through the C_{bc}



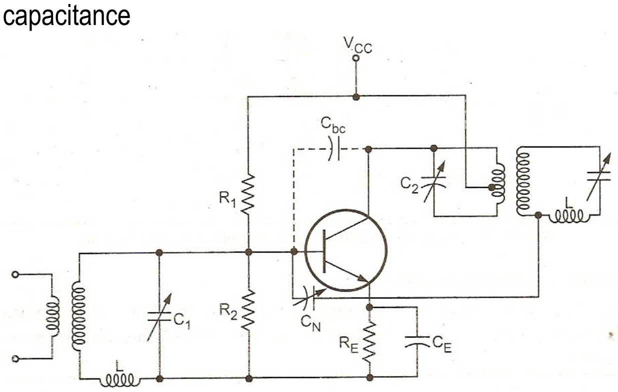
- **Neutrodyne neutralization**

In Neutrodyne neutralization technique, the neutralization capacitor is connected from the lower end of base coil of the next stage to the base of the transistor. The circuit functions as the Hazeltine neutralization circuit with an advantage that the neutralization capacitor does not have the supply voltage across it.



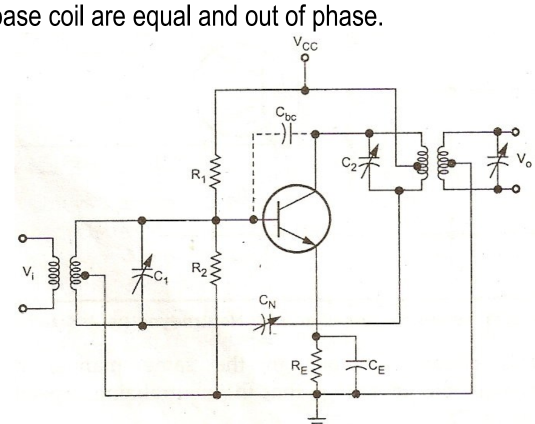
- **Neutralization using coil**

Neutralization of RF amplifier using coil L part of the tuned circuit at the next stage is oriented for maximum coupling to the other winding. It wound on a separate form and is mounted at right angles to the coupled windings. If the windings are properly polarized, the voltage across L due to the circulating current in the base circuit will have proper phase to cancel the signal coupled through the base to collector, C_{bc} capacitance



- **Rice neutralization,**

Rice neutralization uses a center tapped coil in the base circuit. With this arrangement the signal voltages at the ends of the tuned base coil are equal and out of phase.



L – 21. Class “C” Tuned Amplifiers, their Efficiency and Applications:

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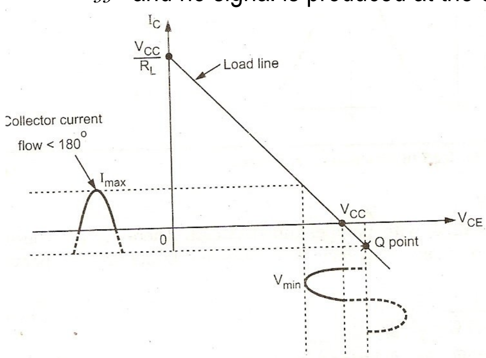
Class "C" Tuned Amplifier

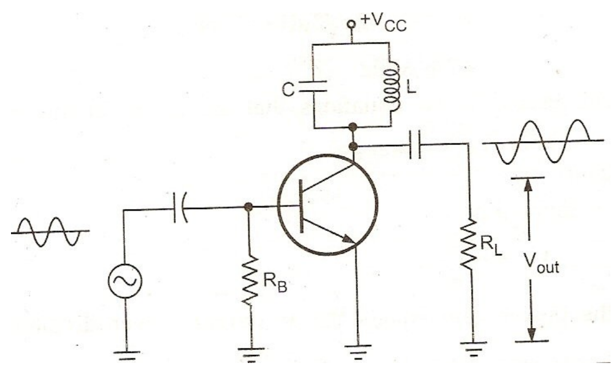
A Class "C" Tuned Amplifier is an amplifier which the Q point and the input signal are selected such that the output signal is obtained for less than a half cycle, for a full input cycle.

Due to such a selection of the Q point, the transistor remains active, for less than a half cycle, hence only that part is reproduced at the output. For the remaining cycle of the input cycle, the transistor remains $cut-off$ and no signal is produced at the output.



Current and voltage waveforms for Class "C" amplifier operation shows that it is apparent that the total angle through which current flows is less than 180^0. This angle is called the conduction angle, θ_c

A parallel resonant circuit acts as load impedance. As collector current flows for less than half a cycle, the collector current consists of a series of pulses with the harmonics of the input signal. A parallel tuned circuit acting as load impedance is tune to the input frequency; therefore it filters the harmonic frequencies and produces a sine wave output voltage consisting of fundamental component of the input signal. The output voltage is maximum at the resonant frequency of the parallel tuned circuit. $f_r = \frac{1}{2\pi\sqrt{LC}}$

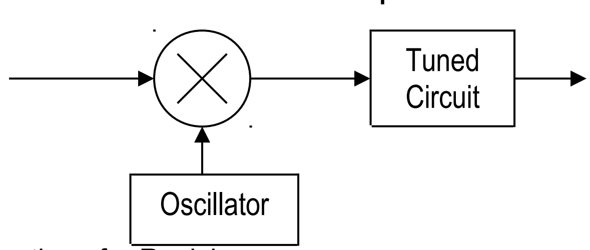


Efficiency of Class "C" Tuned Amplifiers:

The efficiency of a Class "C" Tuned Amplifier is given by $\eta = \frac{P_{out}}{P_{dc}} \times 100\% = \frac{P_{out}}{V_{cc} \times I_{dc}} \times 100\%$

When the conduction angle $\theta_c = 180^0$ the efficiency $\eta = 75\%$. The efficiency of class "C" amplifiers increases when the conduction angle decreases and at very small conduction angle, maximum efficiency $(\eta \cong 100\%)$ is approached. Bandwidth $BW = f_2 - f_1 = \frac{f_r}{Q}$; where Q - quality factor of the circuit.

Applications of Class "C" Tuned Amplifiers:



Class "C" Tuned Amplifiers are used in mixer (frequency converter) circuits. Frequency conversion is the process of translating a modulated signal to a higher or lower frequency while still retaining the original transmitted information.

Questions for Revision

PART - A

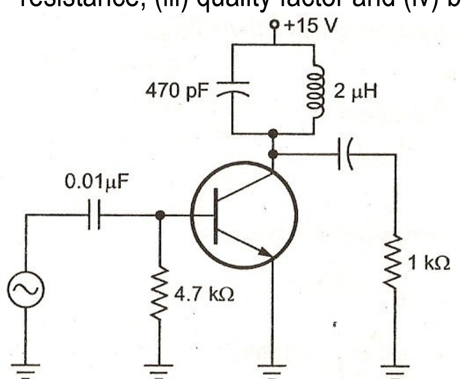
Que 1. Define a synchronously tuned amplifier.
Que 2. What do you understand by unloaded Q and loaded Q?
Que 3. Discus the advantages of tuned amplifiers.
Que 4. Discus the disadvantages of tuned amplifiers.
Que 5. Compare the frequency response of single tuned, double tuned and stagger tuned amplifiers.
Que 6. Define Q

PART - B

Que 1. Explain the effect of double tuned amplifier on bandwidth.
Que 2. What are the advantages of double tuned amplifier compared to a single tuned amplifier?
Que 3. A Class-C tuned amplifier has inductance of $3\mu H$ and capacitance of $470pF$ in the tank circuit. Calculate the resonant frequency.
Que 4. Write short notes on coil losses.
Que 5. Mention the need for stagger tuned amplifier.
Que 6. What do you understand by instability of tuned amplifiers?
Que 7. What do you understand by neutralization?

PART - C

Que 1. Explain the Hazeltine method of neutralization in tuned amplifiers.
Que 2. Explain the Neutrodyne method of neutralization in tuned amplifiers.
Que 3. Explain the method of neutralization in tuned amplifiers using coil.
Que 4. For the circuit in Fig., assuming $Q_L = 100$ calculate the (i) resonant frequency, (ii) ac collector resistance, (iii) quality factor and (iv) bandwidth



Que 5. (i) Draw the circuit diagram of a single tuned amplifier and (ii) explain the function of each component indicated in the circuit. (iii) If the bandwidth of a single tuned amplifier is $20kHz$, calculate the bandwidth if such three stages are cascaded.
Que 6. Write notes on Class-C amplifier highlighting the following; (i) meaning of Class-C amplifier, (ii) Current and voltage waveforms, (iii) Application of Class-C tuned amplifiers and (iv) Mixer circuit using Class-C tuned amplifier.

UNIT 4. MULTIVIBRATOR CIRCUITS

L – 22. Introduction to multivibrator circuits

Date: ……………..…
Hall: 21 Period: ……

Multivibrators are electronic circuits that are used to generate non-sinusoidal waveforms, such as, square, rectangular, triangular, ramp or sawtooth. A multivibrator is a two-stage resistance coupled amplifier with the output of each stage coupled regeneratively to the other, i.e., the output of the first stage is fed to the input of the second stage while the output of the second stage is feedback to the input of the first stage.

Types of multivibrators:

i. Bistable multivibrator (flip-flop);
   - Has two stable states. The multivibrator can exist indefinitely in either of the two states; hence it requires an external trigger pulse to change from one stable state to the other.

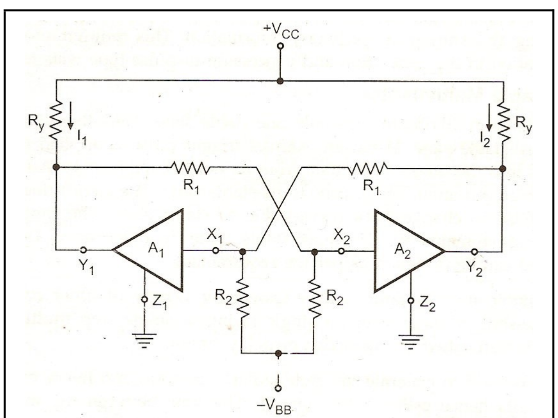
ii. Monostable multivibrator (single-shot);
   - Has only one stable state and another unstable (quasi-stable) state; hence when an external trigger pulse is applied to the circuit, the circuit goes into the quasi-stable state from its normal stable state in which it remains for an interval of time determined by the circuit components, then automatically returns to its stable state.

iii. Astable multivibrator (free-running);
   - Has both the states as quasi-stable states, i.e., none of the state is state; hence it automatically makes successive transitions from one quasi-stable to another without any external trigger pulse. The rate of transition from one state quasi-stable state to the other is determined by the circuit components.

Bistable multivibrator

A_1 and A_2 are active devices which are npn transistors. X , Y and Z are the three terminals of the transistor. The output of each amplifier stage is directly coupled to the input of the other stage.

Assume that the two devices are in equilibrium and are in the active region, the quiescent currents I_1 and I_2 must be the same.



Assume that the current I_1 increases under this condition of equilibrium. Due to this, voltage at Y_1 will decrease, this will further decrease the voltage at X_2. This change in voltage at X_2 will be amplified and inverted by A_2 and the output voltage at Y_2 will increase. Thus input voltage at X_1 will increase. This will decrease the voltage at Y_1 further, by further increase in current I_1. The cycle will repeat, I_1 will keep on increasing while I_2 will keep on decreasing and the circuit will be driven away from equilibrium.

Thus in a stable state of Bistable multivibrator, one of the transistor is OFF the other one is ON, i.e., one is below CUT-OFF and the other is in SATURATION. Both transistors are never OFF or ON simultaneously.

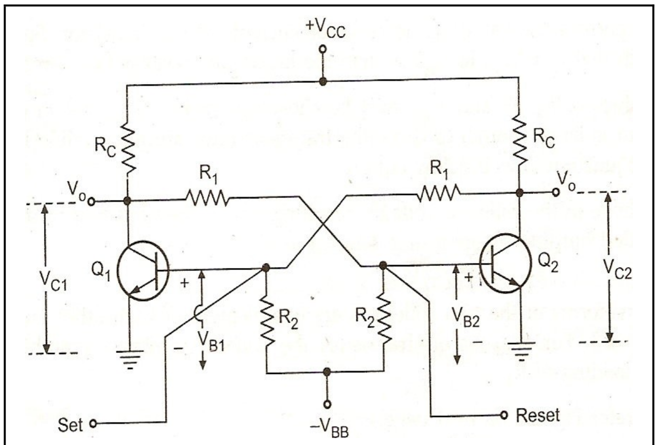
Types of transistor bistable multivibrator circuit,

i. Fixed Bias transistor circuit,
ii. Self bias transistor circuit,

L – 23. Fixed Bias $n-p-n$ Transistor Bistable Multivibrator

Date: ……………..…
Hall: 21 Period: ……

The circuit uses two $n-p-n$ transistors Q_1 and Q_2 whereby the collector of Q_2 is coupled to the base of Q_1 through resistor R_1 while the collector of Q_1 is connected to the base of Q_2 through an identical resistor R_1. Since the characteristics of both transistors can never be identical, hence giving supply one of the transistors starts conducting ahead of the other.



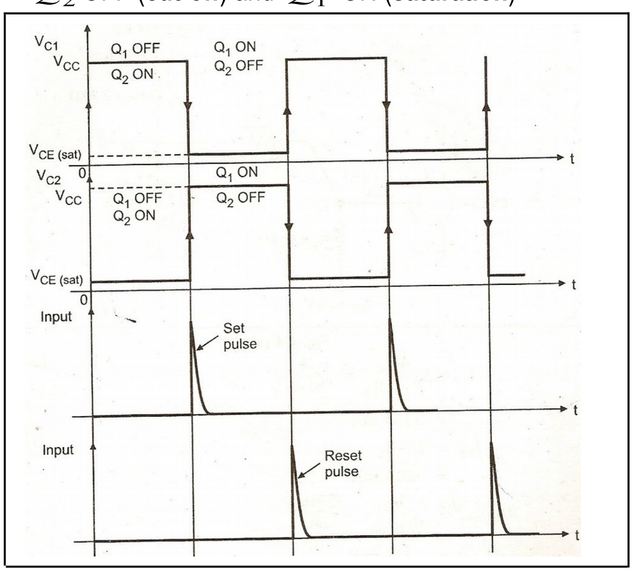
Assume Q_2 starts conducting ahead of Q_1 and hence the current drawn by Q_2 is more than the current drawn by Q_1. Due to regenerative feedback, current drawn by Q_2 keeps on increasing and that drawn by Q_1 keeps on decreasing. This cumulative process drives Q_2 to **saturation** and Q_1 to **cut-off**. This is a stable state of the multivibrator, and the circuit remains in this stable state till an external trigger pulse is applied at the **set** or **reset** terminal.

If a positive going pulse is applied at the **set** or **reset** terminal, it will drive Q_1 to **saturation** and Q_2 to **cut-off** – the second stable state of the multivibrator.

The two states of a bistable multivibrator;

Q_1 OFF (**cut-off**) and Q_2 ON (**saturation**)

Q_2 OFF (**cut-off**) and Q_1 ON (**saturation**)



Under saturation condition;

Collector current I_C is maximum,

$$I_C = \frac{V_{CC} - V_{CE(sat)}}{R_C} \approx \frac{V_{CC}}{R_C},$$

Therefore R_C is chosen such that I_C must be less than the permissible current.

Analysis of a collector-coupled fixed biased multivibrator circuit using an example:

Example:

The fixed bias multivibrator uses the following parameters:

$V_{CC} = +12V;\ V_{BB} = -8V;\quad R_1 = 10k\Omega;\ R_2 = 50k\Omega;\ R_C = 2.2k\Omega$

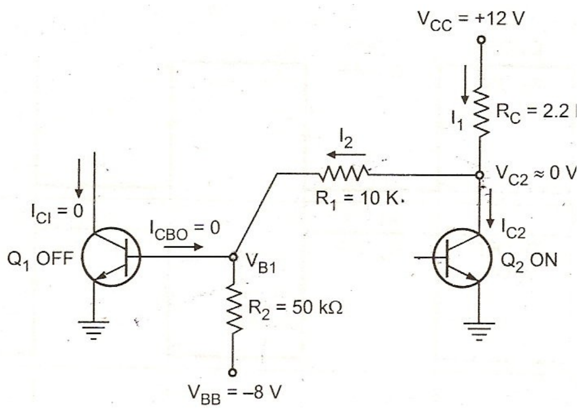
The transistors are silicon transistors with a minimum value of $h_{fe} = 30$. Calculate the stable state currents I_{C1} ; I_{C2}, I_{B1}, I_{B2} and stable state voltages; $V_{C1}, V_{C2}, V_{B1}, V_{B2}$ when;

(i) *All junction voltages are neglected;*
(ii) *Assuming $V_{CE(sat)} = 0.2V$ and $V_{BE(sat)} = 0.7V$*

<u>Solution</u>

Case (i) All junction voltages of ON transistor are neglected, i.e. $V_{CE2} = 0V$ *and* $V_{BE2} = 0V$;

As the emitter is grounded $V_{C2} = 0V$ *and* $V_{B2} = 0V$ hence the equivalent circuit in part from base of Q_1 to the collector of Q_2 is as below;



$$V_{B1} = -V_{BB}\left(\frac{R_1}{R_1 + R_2}\right) = -8\left(\frac{8}{8+50}\right) = -1.33V$$

$$I_1 = \frac{V_{CC}}{R_C} = \frac{12}{2.2\times10^3} = 5.45mA$$

$$I_2 = \frac{V_{BB}}{R_1 + R_2} = \frac{8}{10+50} = 0.133mA$$

$$I_{C2} = I_1 - I_2 = 5.45mA - 0.133mA = 5.316mA$$

The equivalent circuit showing the collector of Q_1 to base of Q_2 is shown below;

$$I_3 = \frac{V_{CC}}{R_C + R_1} = \frac{12}{(2.2+10)\times10^3} = 0.9836mA$$

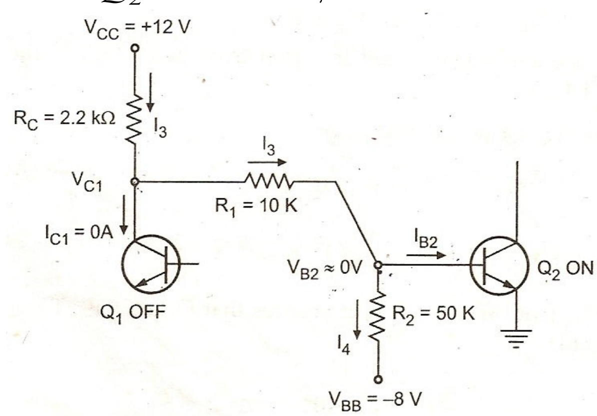
The current through R_C and R_1 as $I_{C1} = 0$

$$I_4 = \frac{V_{B2} - V_{BB}}{R_2} = \frac{0-(-8)}{50} = 0.16mA$$

$$I_{B2} = I_3 - I_4 = 0.8236mA$$

$I_{B2} > (I_{B2})_{\min}$ - Transistors Q_2 is saturated

$$V_{C1} = V_{CC} - I_3R_C = 12 - 0.98396\times2.2 = 9$$



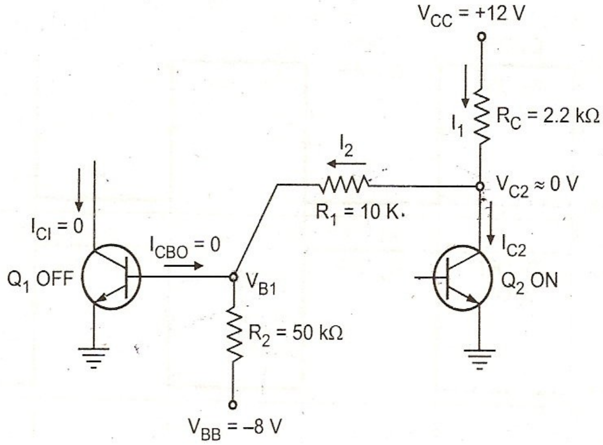
Hence the stable state currents are; $I_{C1} = 0A;$ $I_{C2} = 5.316mA;$ $I_{B1} = 0A;$ $I_{B2} = 0.8236mA$

And stable state voltages are; $V_{C1} = 9.836V;$ $V_{C2} = 0;$ $V_{B1} = -1.33V;$ and $V_{B2} = 0$

Case (ii) $V_{CE(sat)} = 0.2V$ *and* $V_{BE(sat)} = 0.7V$

For the transistor Q_2 as emitter is grounded $V_{C2} = 0.2V$ *and* $V_{B2} = 0.7V$ hence the equivalent circuit in part from base of Q_1 to the collector of Q_2 is as below;

V_{B1} will be due to V_{C2} and V_{BB} , hence using superposition principle, i.e., considering the effect of each independently,



$$V_{B1} = -V_{BB}\left(\frac{R_1}{R_1+R_2}\right)\Bigg|_{V_{C2}=0} + V_{C2}\left(\frac{R_1}{R_1+R_2}\right)\Bigg|_{V_{BB}=0}$$

$$V_{B1} = -8\left(\frac{10}{10+50}\right)+(0.2)\left(\frac{50}{50+10}\right) = -1.16V$$

As $V_{B1} < V_{BE(cut-off)}$ hence Q_1 is OFF

$$I_1 = \frac{V_{CC} = V_{C2}}{R_C} = \frac{12-0.2}{2.2\times 10^3} = 5.36mA$$

$$I_2 = \frac{V_{C2}+V_{BB}}{R_1+R_2} = \frac{0.2+8}{10+50} = 0.136mA$$

$$I_{C2} = I_1 - I_2 = 5.36mA - 0.136mA = 5.223mA$$

To calculate I_{B2} apply the equivalent circuit showing the collector of Q_1 to base of Q_2 as shown below:

$$I_3 = \frac{V_{CC}-V_{B2}}{R_C+R_1} = \frac{12-0.7}{(2.2+10)\times 10^3} = 0.926m$$

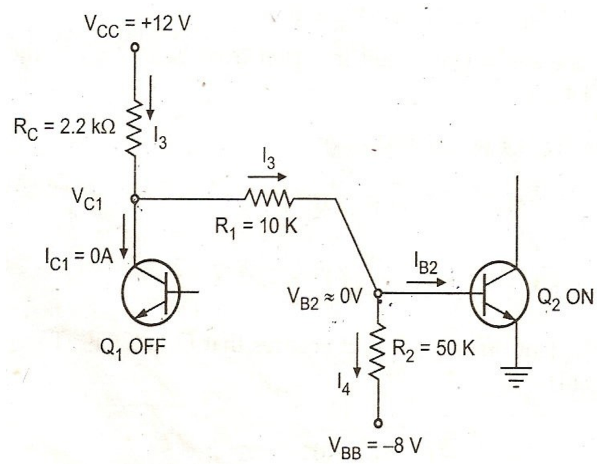
The current through R_C and R_1 as $I_{C1} = 0$

$$I_4 = \frac{V_{B2}-V_{BB}}{R_2} = \frac{0.7-(-8)}{50} = 0.174mA$$

$$I_{B2} = I_3 - I_4 = (0.926-0.174)mA = 0.752m$$

$I_{B2} > (I_{B2})_{\min}$ - Transistors Q_2 is saturated

$$V_{C1} = V_{CC} - I_3 R_C = 12 - 0.926\times 2.2 = 9.7$$



Hence the stable state currents are; $I_{C1} = 0A;$, $I_{C2} = 5.223mA$, $I_{B1} = 0mA$ and $I_{B2} = 0.752mA$

And the stable state voltages are; $V_{C1} = 9.7628V$; $V_{C2} = 0.2V$; $V_{B1} = -1.16V$ and $V_{B2} = 0.7V$

Loading effect:

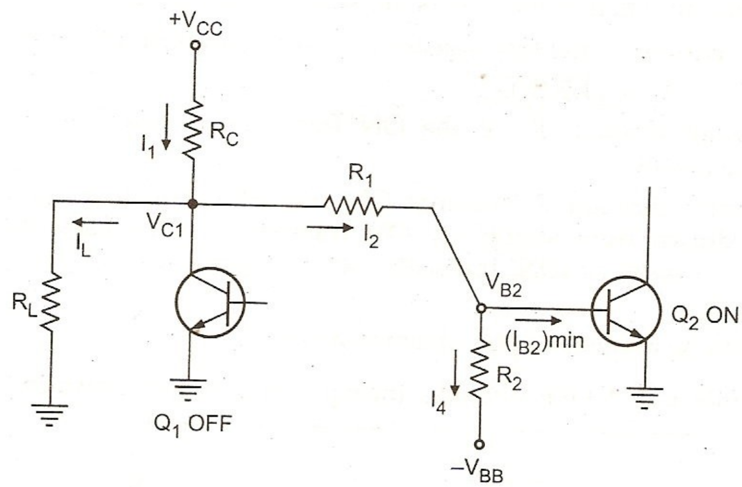
The bistable multivibrator is used to drive other circuits, therefore the loading effect must be considered during designing to ensure that one transistor remains in **saturation** while the other in **cut-off.**

Calculation of heaviest load (load resistance connected to transistor Q_1);

$I_4 = \dfrac{V_{B2} - V_{BB}}{R_2}$. Now $I_2 = I_4 + (I_{B2})_{\min}$ and $I_2 = \dfrac{V_{C1} - V_{B2}}{R_1}$

$\Rightarrow V_{C1}$ can be determined from $V_{C1} = V_{CC} - I_1 R_C$. Hence current I_1 is calculated from $I_1 = I_L + I_2$

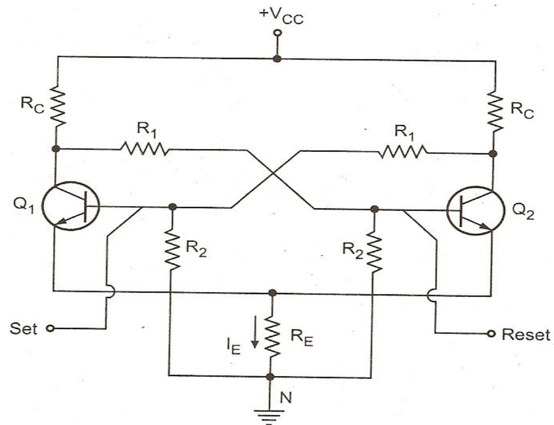
Thus the corresponding load $(R_L)_{\min}$ can be obtained from $R_L = \dfrac{V_{C1}}{I_L}$



L – 24. Collector-Coupled Self-Biased Transistor Bistable Multivibrators:

Date: …………..…
Hall: 21 Period: ……

A self-biased transistor bistable multivibrator is obtained when a negative power supply in a fixed bias multivibrator is eliminated by using a common emitter resistance R_E. The resistance provides the biasing to keep one transistor ON and the other OFF in a stable state. The resistance R_E is connected between common emitter terminals.



The circuit uses two $n-p-n$ transistors Q_1 and Q_2. For $p-n-p$ the polarity V_{CC} must be reversed. The required biasing voltage is provided by the $I_E R_E$ drop. The waveforms at the two collectors are complimentary of each other as in fixed bias multivibrator.

Analysis of a collector-coupled self-biased multivibrator circuit using an example:

<u>Example:</u>

The self-biased multivibrator uses the following parameters: $V_{CC} = +12V; \quad R_1 = 30k\Omega;$ $R_2 = 10k\Omega; \quad R_C = 4k\Omega; and\ R_E = 500\Omega$; *The transistors are* $n-p-n$ *silicon transistors.*

Calculate:

(i) Stable state currents $I_{C1}\ ; I_{C2}, I_{B1}, I_{B2}$

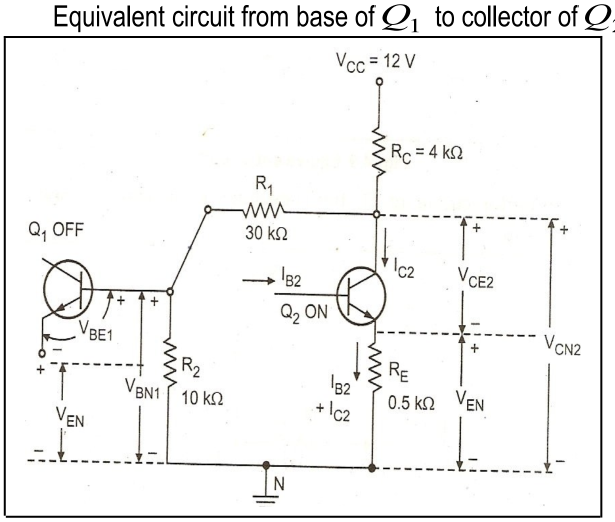
(ii) Stable state voltages; $V_{C1}, V_{C2}, V_{B1}, V_{B2}$ *and*

(iii) Minimum value of h_{fe} *which will keep the* ON *transistor in saturation.*

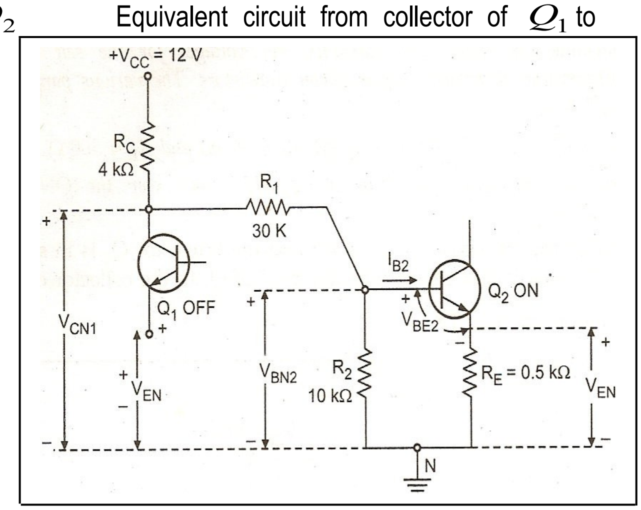
<u>Solution</u>

Assume that transistor Q_1 is cut-off and transistor Q_2 is in saturation and draw the equivalent circuit from base of Q_1 to collector of Q_2, and another equivalent circuit from collector of Q_1 to base of Q_2

Equivalent circuit from base of Q_1 to collector of Q_2



Equivalent circuit from collector of Q_1 to



To calculate the various voltages it is necessary to calculate the current I_{C1}, I_{B2} as Q_2 is ON. The currents I_{C1} = $I_{B1} = 0$ as Q_1 is OFF.

Obtaining Thevenin's equivalent circuit across collector and ground while another across base and ground of the same transistor Q_2 assuming it is loaded.

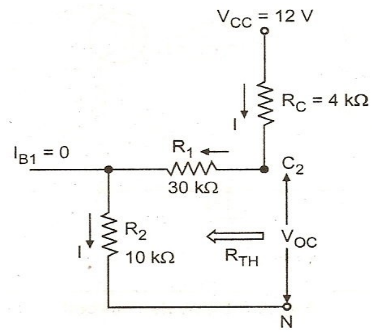
To replace the collector circuit of transistor Q_2 by Thevenin's equivalent, consider Q_2 as open. Refer Fig.

$$V_{OC} = I(R_1 + R_2) = \frac{V_{CC}}{(R_1 + R_2 + R_C)}(R_1 + R_2)$$

$$V_{OC} = \frac{12}{(30+10+4)} \times 40 = 10.9V$$

$R_{TH} = (R_1 + R_2) \| R_C$ with $V_{CC} - N$ short

$$R_{TH} = \frac{40 \times 4}{40+4} = 3.363k\Omega$$



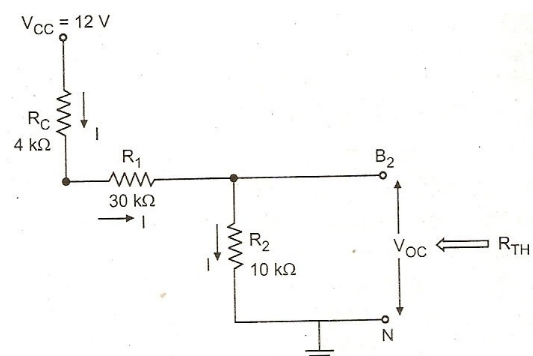
To replace the base circuit of transistor Q_2 by Thevenin's equivalent, consider Q_2 open. Refer Fig. below;

$$V_{OC} = IR_2 = \frac{V_{CC}}{(R_1 + R_2 + R_C)} \times R_2$$

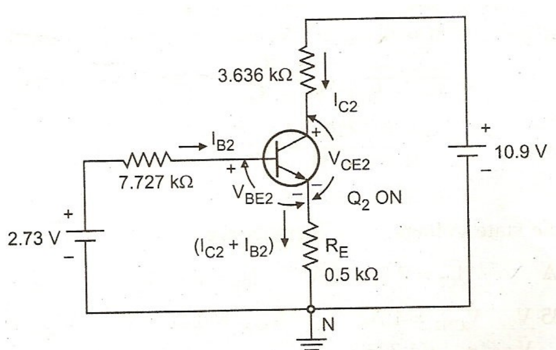
$$V_{OC} = \frac{12}{(30+10+4)} \times 10 = 2.73V$$

$$R_{TH} = R_2 \| (R_1 + R_C)$$

$$R_{TH} = \frac{10 \times 34}{10+34} = 7.727k\Omega$$



Thus the equivalent circuit for transistor Q_2 ON, of the calculated above is given in the figure below;



For silicon transistors,

$V_{BE(Sat)} = 0.8V$ *and* $V_{CE(Sat)} = 0.4V$

Applying KVL to base-emitter loop

$$-7.727I_{B2} - V_{BE2} - (I_{C2} + I_{B2}) \times 0.5 + 10.9 = 0$$

With $V_{BE2} = 0.8V$;

$$I_{B2} + 0.06075I_{C2} = 0.2345$$

Applying KVL to collector-emitter loop

$$-3.636I_{C2} - V_{CE2} - (I_{C2} + I_{B2}) \times 0.5 + 10.9 = 0$$

With $V_{CE2} = 0.4V$; $4.14I_{C2} + 0.5I_{B2} = 10.5$

Solving the two equations simultaneously we get:

$$I_{C2} = 2.526mA;\ and\ I_{B2} = 0.0847mA;\ (h_{fe})_{min} = \frac{I_{C2}}{I_{B2}} = \frac{2.526}{0.0847} = 29.815$$

The various voltages can be obtained by referring equivalent circuits from base of Q_1 to collector of Q_2 and from collector of Q_1 to base of Q_2 .

$$V_{EN} = (I_{B2} + I_{C2}) R_E$$

$$= 1.305 \text{ V}$$

$$V_{CN2} = V_{CE2} + V_{EN}$$

$$= 0.4 + 1.305 = 1.705 \text{ V}$$

$$V_{BN2} = V_{BE2} + V_{EN} = 0.8 + 1.305$$

$$= 2.105 \text{ V}$$

$$V_{BN1} = V_{CN2} \times \frac{R_2}{R_1 + R_2} = 1.705 \times \left(\frac{10}{40}\right) = 0.4262 \text{ V}$$

$$V_{BE1} = V_{BN1} - V_{EN} = 0.4262 - 1.305 = -0.8788 \text{ V}$$

As $V_{BE1} < V_{BE}(\text{sat})$ which is about 0.8 V, the transistor Q_1 is indeed OFF.

$$V_{CN1} = \frac{V_{CC} R_1}{(R_C + R_1)} + \frac{V_{BN2} R_C}{R_C + R_1} \quad \text{using Superposition principle}$$

$$= \frac{12 \times 30}{34} + \frac{2.105 \times 4}{34}$$

$$= 10.8358 \text{ V}$$

Thus the stable state voltages and currents are :

$I_{C1} = 0$ mA $\quad I_{C2} = 2.526$ mA $\quad I_{B1} = 0$ mA $\quad I_{B2} = 0.0847$ mA

$V_{CN1} = 10.835$ V $\quad V_{CN2} = 1.705$ V $\quad V_{BN1} = 0.4262$ V $\quad V_{BN2} = 2.105$ V

and $\quad V_{EN} = +1.305$ V

The voltage V_{EN} provides the required self bias.

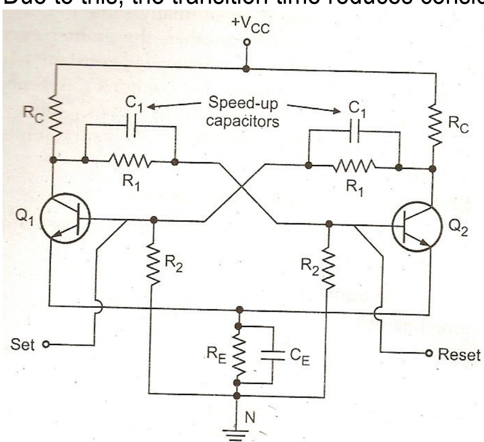
L – 25. Speed-up Capacitors / Commutating Capacitors:

Date: ……………..…
Hall: 21 Period: ……

The bistable multivibrator remains in the stable state till a **trigger pulse** is applied to **set** or **reset** terminals. In some applications it is necessary that the transition from one state to another to occur instantaneously, when abruptly changing pulse is applied to the circuit.
$\Rightarrow$ This means: **Transition time** of the circuit from one state to another should be as small as possible.

Transition time: the time interval during which conduction transfers from one transistor to another.

To improve the switching characteristics, **small capacitors** are used in **shunt** with **coupling resistors** R_1 . Due to this, the transition time reduces considerably without affecting the stable states.



Speed-up capacitors allow fast rise and fast fall thus avoiding distortion in output wave forms. Speed-up capacitors are a.k.a. **Commutating Capacitors** a.k.a. **Transpose Capacitors**, since they help the multivibrator in making instantaneous transition between the states.

The smallest allowable interval between triggers is called **Resolving Time** of the binary.
$\Rightarrow$ Resolving Time should be sufficient so that all transients die out completely and hence the flip-flop can be triggered reliably. Resolving Time decides the maximum frequency of triggering.

$$f_{\max} = \frac{1}{\text{Resolving Time}} = \frac{1}{2C_1\left(R_1 \| R_2\right)} = \frac{R_1 + R_2}{2C_1 R_1 R_2}$$

Application of Bistable Multivibrators:

- Used for performance of many digital operations, e.g. counting and storing of digital information.
- Used as memory elements in registers, counters etc.
- Used in processing of pulse type waveforms,
- Used to generate symmetrical square waves, i.e. trigger pulses of equal intervals, corresponding to the frequency required.
- Can be used as frequency dividers.

L – 26. Triggering of Bistable Multivibrators:

Date: ……………..
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To achieve a transition from one state to another, a **trigger signal** (a **pulse**, positive-going or negative-going **step voltage**) is required. Such signals are used to produce **symmetrical** or **unsymmetric**al triggering.

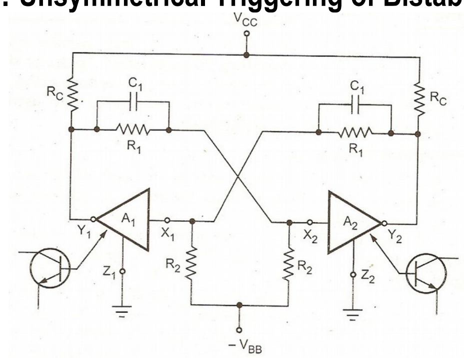
Unsymmetrical Triggering: - Two trigger inputs are used;

i. To **set** the circuit in one particular state (e.g. Q_1 OFF and Q_2 ON)
ii. To **reset** the circuit to the opposite state (e.g. Q_2 OFF and Q_1 ON)

Unsymmetrical triggering is designed with two separate triggering sources to turn OFF the transistor which is ON. The other name for such triggering is set-reset triggering.

Symmetrical Triggering: - One trigger input to the input of any one transistor is used;
Each successive triggering signal induces a transition. e.g. trigger input is applied to Q_1 which is ON then it makes Q_1 OFF and Q_2 ON. In this type of triggering, if one triggering input makes one particular transistor ON to OFF, then the next triggering signal makes the same transistor OFF to ON.

1. Unsymmetrical Triggering of Bistable Multivibrators

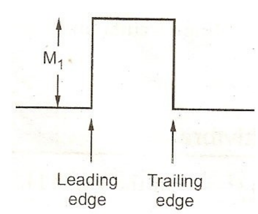


Consider A_1 and A_2 are $n-p-n$ transistors. Let A_1 be OFF, so to turn it ON a positive step is applied to the base of A_1 through C_1.

Difficulties in triggering a multivibrator using a positive step;

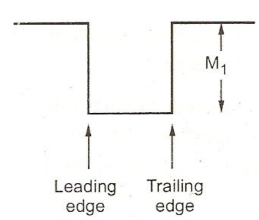
(i) To have the transition of A_1 from OFF to ON a positive pulse is required which must exceed the voltage by which the transistor is below cut off.

(ii) If a positive step is applied to OFF transistor and if amplitude M_1 is more than the voltage by which OFF transistor is below cut-off, then transition occurs and OFF transistor becomes ON. But in the positive pulse, there is a negative-going step at trailing edge and as circuit is more sensitive to negative step the same transistor which was made ON at leading edge will become again OFF at trailing edge.



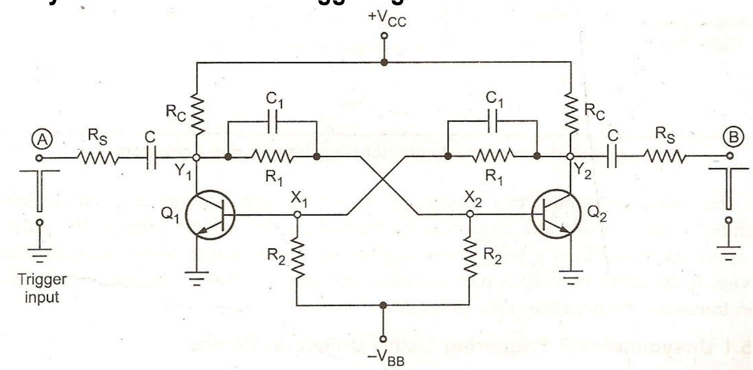
Triggering a multivibrator using a positive step;

If a negative pulse is applied to make ON transistor OFF then its magnitude M_1 at leading edge can be adjusted small to which transition results, as the same magnitude M_1 which is positive-going at trailing edge is smaller than the voltage by which OFF transistor is below cut-off and reverse transition can not occur at trailing edge.



Due to this advantage the negative pulse is used for unsymmetrical triggering to achieve transition by making ON transistor OFF.

Unsymmetrical Collector triggering

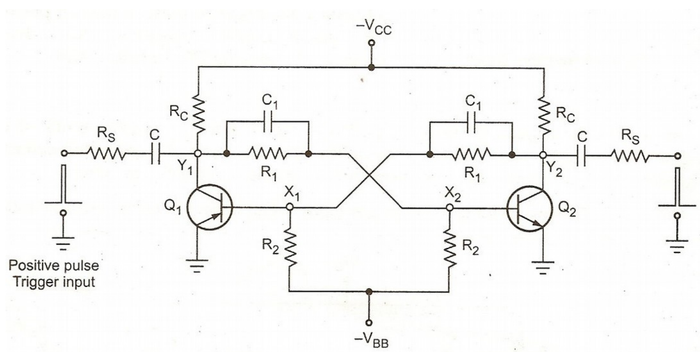


The trigger signal is applied at collector i.e. output of one of the stages of the binary instead of applying at one of the base, i.e. input.

Analysis of Unsymmetrical collector triggering:

Let Q_2 be ON and Q_1 be OFF. If a negative pulse is applied to point A (output of Q_1), due to transmission through **speed-up capacitor** C_1 , the signal will immediately appear at base input X_2 of Q_2 . Since R_S increases the sensitivity of $n-p-n$ transistor to a negative pulse, Q_2 quickly turns OFF and Q_1 turns ON. To have next transition, a second triggering signal is required at Y_2 which quickly appears at the base of Q_1 which turns Q_1 OFF and makes Q_2 ON.

If $p-n-p$ transistors are used the supply voltage polarities must be opposite, then a positive is required to turn ON the OFF transistor.



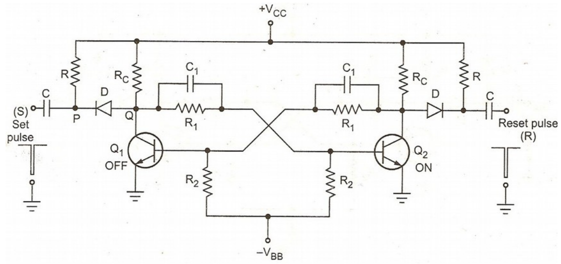
Conclusion

Unsymmetrical triggering of a bistable multivibrator on the leading edge is best achieved by applying a pulse to the output of the non-conducting device.

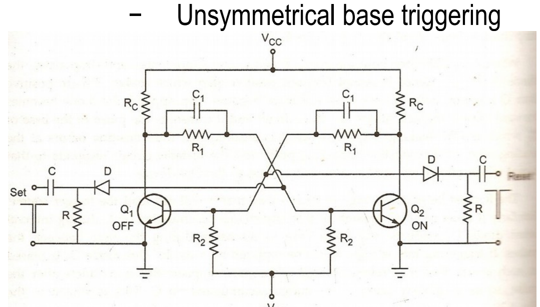
- For a $n-p-n$ a negative pulse is required,
- For a $p-n-p$ a positive pulse is required,

Other unsymmetrical triggering methods;

- Unsymmetrical triggering using unilateral device
    - Unsymmetrical triggering using diodes



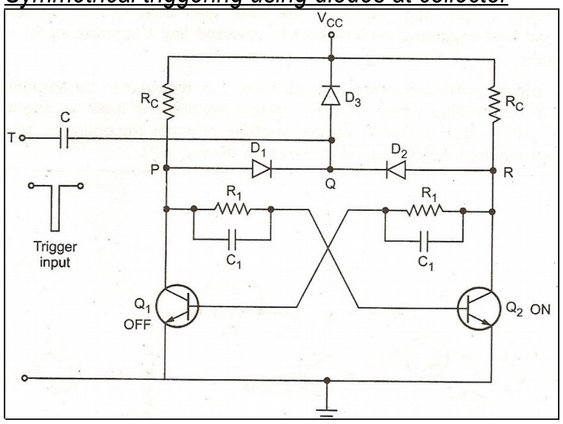
    - Unsymmetrical base triggering



2. Symmetrical Triggering of Bistable Multivibrators

The trigger pulse is applied to only one transistor either at its base or its collector. For each successive trigger signal one of the two transistors changes its state from ON to OFF. Symmetrical triggering using diodes at the collector of one of the transistors is shown in figure below.

Symmetrical triggering using diodes at collector

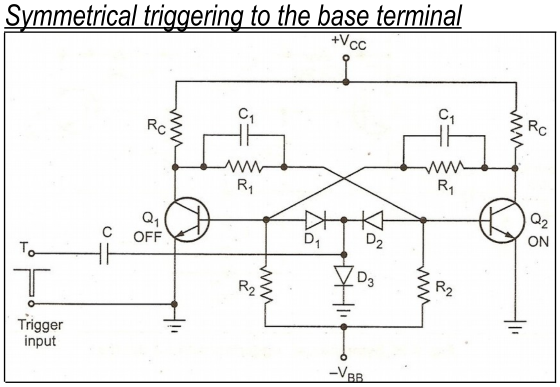


Assume Q_2 is ON and Q_1 is OFF. The voltage drop across R_C of Q_2 is large and due to D_2 is reverse biased. Since Q_1 is OFF, there is no voltage drop across R_C of Q_1, hence points P and Q are equipotential, i.e. no voltage drop across D_1.

When a negative going pulse is applied at input T, point Q goes negative due to which D_1 becomes forward biased and acts as a short circuit allowing the pulse to reach point P which is collector of Q_1. Then the negative pulse is passed to the base of Q_2 through R_1 and C_1. This turns OFF Q_2 and turns ON Q_1. Thus transition occurs. D_1 and D_2 are steering diodes.

NB (i) The circuit will not respond to a positive pulse;
(ii) If $p-n-p$ transistors are used, then diode directions must be reversed and a positive pulse must be used as trigger.

Symmetrical triggering to the base terminal



In the circuit, D_3 is connected to the ground instead of connecting to $+V_{CC}$. When a negative pulse is applied, it passes to the base of Q_2 through forward biased D_2 and Q_2 becomes OFF and Q_1 turns ON due to regenerative action.

When next negative pulse is applied, D_1 becomes forward biased and the pulse gets applied to Q_1 which is then ON, turning it OFF.

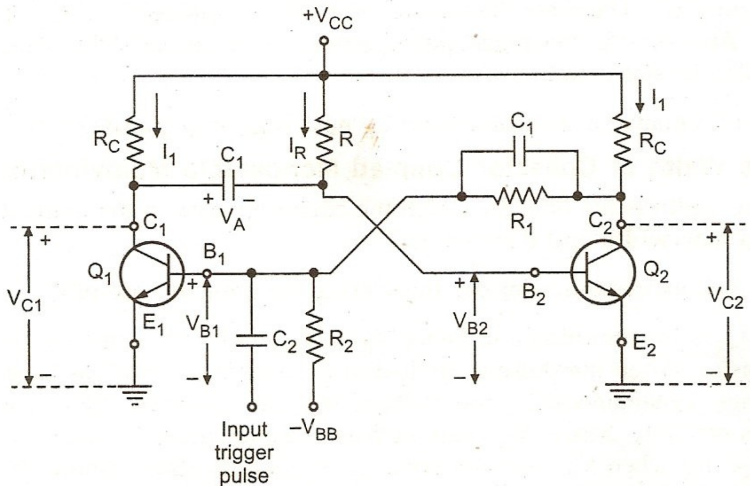
L – 27. Monostable Multivibrators:

Date: ……………..…
Hall: 21 Period: ……

A Monostable multivibrator has one stable state, when an external trigger is applied the circuit changes to quasi-stable state and ten automatically after some time interval T , the circuit returns back to its stable state. The time T is dependent on the circuit components.

Collector-Coupled Monostable Multivibrator

Q_1 and Q_2 are identical $n-p-n$ transistors with collector (output) of Q_2 coupled to the base (input) of Q_1 through resistance R_1 which is shunted by a small capacitor C_1 (speed-up capacitor). The collector of Q_2 is coupled to the base of Q_2 through capacitor C for capacitive coupling. Resistance R at input of Q_2 is returned to the supply voltage $+V_{CC}$.



The values of R_2 and $-V_{BB}$ are chosen such that transistor Q_1 is OFF by reverse biasing it. Q_2 is saturated i.e. ON, by forward biasing it with $+V_{CC}$ and resistance R .

Thus Q_2 is **ON** and Q_1 is **OFF** is the **normal stable state**.

If an unsymmetrical positive pulse of sufficient magnitude and duration is applied to the base of Q_1 through capacitor C_2 , the transistor Q_1 starts conducting. Due to this, voltage at its collector V_{C1} (which is coupled to base of Q_2 through C)decreases.

But voltage across capacitor cannot change instantaneously. Hence decrease in V_{C1} directly causes a decrease in base voltage of Q_2 i.e. V_{B2} . The voltage drop is I_1R_C which decreases the forward bias of Q_2 and hence collector current I_2 decreases. Thus the collector voltage of Q_2 increases which is applied to the base of Q_1 through R_1 . This further increases the base potential of Q_1 and Q_1 is quickly driven into saturation and at the same time Q_2 gets driven into cut-off. This is a quasi-stable state of the circuit.

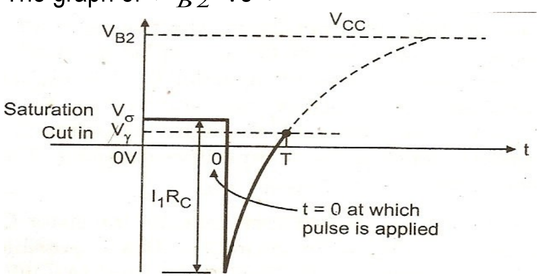
The circuit remains in this quasi-stable state for time T when capacitor C is charging through the path $V_{CC}R$ and ON transistor Q_1 . As it starts charging towards $+V_{CC}$, the base of Q_2 experiences rise in voltage. When this voltage becomes more than cut-in voltage V_γ of Q_2 ,then Q_2 starts conducting. And due to regenerative action Q_1 is turned OFF, and the circuit returns back to its stable state.

Pulse Width of Collector Coupled Monostable Multivibrator

The pulse width (or gate width) is the time for which the circuit remains in the quasi-stable state.

Assume Q_2 is ON (saturation), hence $V_{B2} = V_{BE2(Sat)} = V_\gamma$. When a trigger pulse is applied at $t = 0$ then at $t = 0^+$, as capacitor voltage cannot change instantaneously, the voltage V_{B2} decreases by $I_1 R_C$. The capacitor charges exponentially hence V_{B2} also increases exponentially and at $t \to \infty$ $V_{B2} \to V_{CC}$. But when $V_{B2} = V_\gamma$, Q_2 starts conducting and the circuit returns back to its stable state.

The graph of V_{B2} Vs t



$$T = \tau \ln(2) + \tau \ln \left\{ \frac{V_{CC} - \left(\frac{V_{CE(Sat)} + V_{BE(Sat)}}{2} \right)}{V_{CC} - V_\gamma} \right\}$$

Where: V_γ - Cut-in voltage

τ Time constant for the charging path of capacitor C $\tau = RC$

The gate width is dependent on:

- Transistor characteristics,
- Temperature (as $V_{CE(sat)}$, $V_{BE(Sat)}$ and V_γ depend on temperature),
- Resistance values
- Supply voltages

Waveforms of Monostable Multivibrator;

Assume the trigger pulse is applied at $t = 0$ and the reverse transition from quasi-stable to stable state occurs at $t = T$

- **Stable state:**

  Assume Q_2 is **ON** and Q_1 is **OFF** is the **normal stable state**.

  For Q_2 $V_{B2} = V_{BE2(Sat)} = V_\gamma$; and $V_{C2} = V_{CE(sat)}$

  For Q_1 $V_{C1} = V_{CC}$ since Q_1 is OFF and current is zero

$$V_{B1} = -V_{BB} \left(\frac{R_1}{R_1 + R_2} \right) \Bigg|_{V_{C2}=0} + V_{C2} \left(\frac{R_2}{R_1 + R_2} \right) \Bigg|_{V_{BB}=0} \approx V_F$$

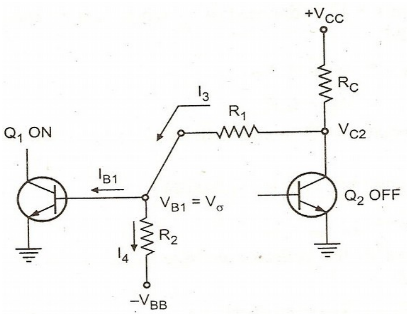
  To have Q_1 OFF, $|V_F| \le 0$ for silicon while $|V_F| \le 0.1$ for germanium transistor.

- **Quasi-stable state**

  When pulse is applied at $t = 0$, Q_2 becomes **OFF** and Q_1 becomes **ON.**

  The voltage at V_{C1} and V_{B2} drops instantaneously by the amount $I_1 R_C$ where I_1 is current drawn by Q_1 when it starts conducting. Then Q_1 gets driven into saturation, hence,

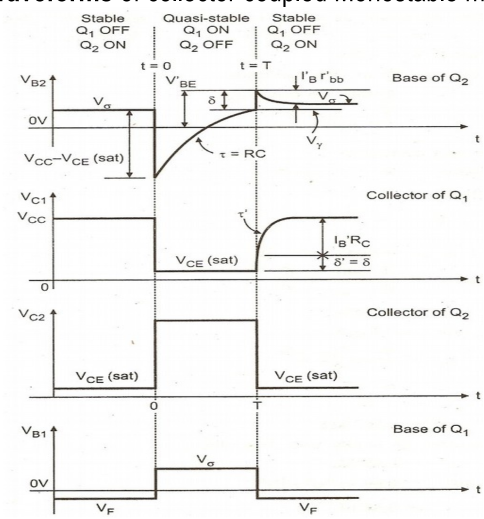
  $V_{B1} = V_\sigma$; $V_{C1} = V_{CE(sat)}$ and $I_1 R_C = V_{CC} - V_{CE(sat)}$



To calculate V_{C2}, consider the equivalent circuit and apply the superposition principle;

$$V_{C2} = V_{CC}\left(\frac{R_1}{R_1 + R_2}\right)\Bigg|_{V_B=0} + V_\sigma\left(\frac{R_C}{R_1 + R_C}\right)\Bigg|_{V_{CC}=0}$$

Waveforms of collector coupled Monostable multivibrator

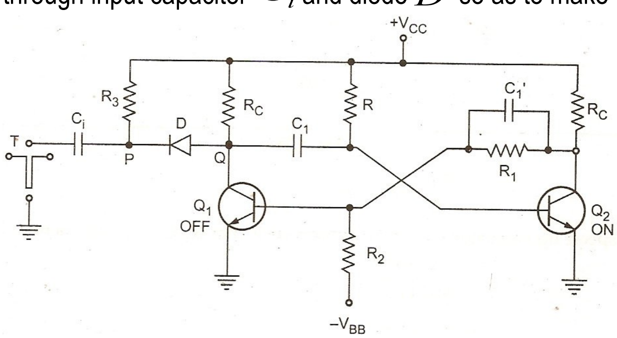


Applications of Monostable multivibrator;

(i) Used to produce rectangular waveforms and hence can be used as gating circuit,
(ii) Used to produce time delays as gate width is adjustable,
(iii) Used to generate uniform width pulses from variable width input pulse train.

Triggering of Monostable multivibrator

Since the sensitivity of $n-p-n$ transistors is more to the negative pulse, a negative pulse is applied through input capacitor C_i and diode D so as to make the ON transistor OFF.



Assume the normal stable state: - Q_2 is **ON** and Q_1 is **OFF**

Diode D is at zero bias as Q_1 is OFF, P and Q are equipotential. When a negative pulse is applied diode D conducts and acts as a short circuit, hence the negative pulse passes to the base of Q_2 which is ON. This decreases Q_2 current and due to regenerative action Q_1 gets driven to **saturation** and Q_2 becomes OFF. This is the **quasi-stable state**.

L – 28. Astable Multivibrators:

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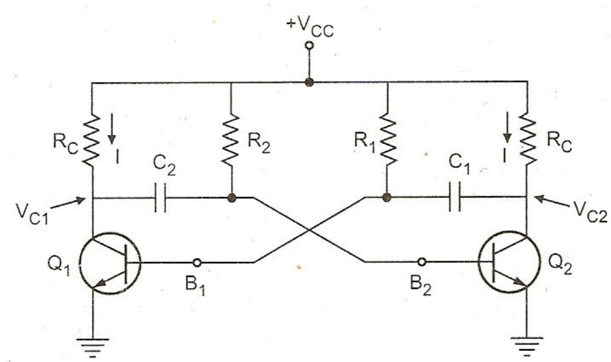
Hall: 21 Period: ……

Collector Coupled Astable Multivibrator

Astable multivibrator has two states, both are quasi-stable, which means, the multivibrator cannot remain in any of the two states indefinitely but it keeps on alternating the states.

Construction:

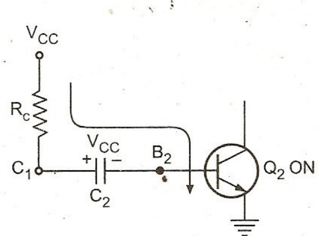
Q_2 and Q_1 are identical transistors; The collector of Q_2 is coupled to base of Q_1 through capacitor C_1 while collector of Q_1 is coupled to base of Q_2 through capacitor C_1. The capacitive coupling is used between stages, due to which neither can remain permanently cut-off. The circuit has two quasi-stable states and it makes periodic switching between these states without any external trigger signal.

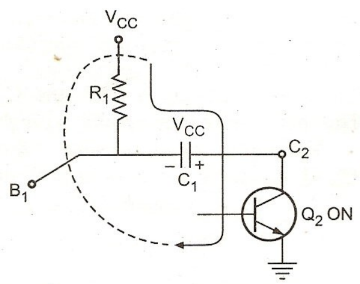


Principle of work:

At start assume Q_2 is **ON** and Q_1 is **OFF**. Capacitor C_2 starts charging towards V_{CC} through path $R_C - C_2 - Q_2(ON)$ till finally the voltage across C_2 becomes equal to V_{CC} with proper polarity.

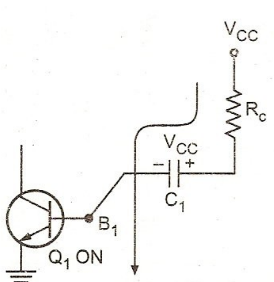
At the same time capacitor C_1 which is charged to V_{CC} in the earlier state, starts discharging through path $Q_2 - V_{CC} - R_1 - C_1$

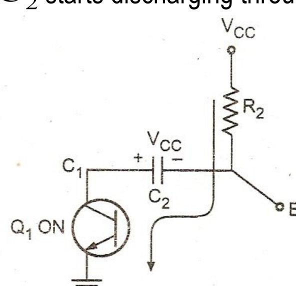




The base of Q_1 is at $-V_{CC}$ at the beginning, but as C_1 starts discharging, it becomes less and less negative, i.e. becomes more positive and finally becomes equal to V_γ the cut-in voltage of Q_1. When $V_{B1} \geq V_\gamma$ transistor Q_1 starts conducting, so Q_1 becomes **ON** and Q_2 becomes **OFF.**

As Q_1 becomes **ON** and Q_2 becomes **OFF**, capacitor C_1 starts charging again through $R_C - C_1 - Q_1(ON)$ while C_2 starts discharging through path $V_{CC} - R_2 - C_2 - Q_1(ON)$





As C_2 discharges, V_{B2} the potential at B_2, becomes less negative i.e. it increases towards positive. When $V_{B2} \geq V_\gamma$ of Q_2 the Q_2 starts conducting, and thus, Q_2 becomes **ON** and Q_1 becomes **OFF.** The changes in the two states is automatic and without any external triggering signal.

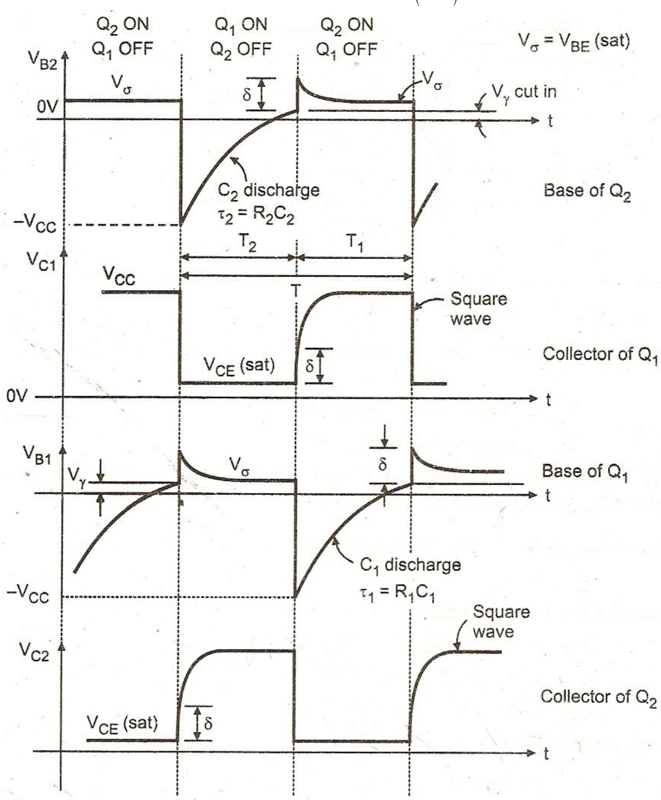
Waveforms of Astable Multivibrator at collectors of Q_1 and Q_2

When Q_1 is **OFF** and Q_2 is **ON,** C_1 discharges and V_{B1} increases. This increases exponentially with time constant R_1C_1 . When Q_1 is in saturation;

$$V_{B1} = V_{BE(sat)}$$
$$V_{C1} = V_{CE(sat)}$$
$$V_{C2} = V_{CC}$$

When Q_1 is **ON** and Q_2 is **OFF,** C_2 discharges and V_{B2} increases. This increases exponentially with time constant R_2C_2 . When Q_1 is in saturation;

$$V_{B2} = V_{BE(sat)}$$
$$V_{C1} = V_{CC}$$
$$V_{C2} = V_{CE(sat)}$$



Expression for time period

$$T_1 = 0.69R_1C_1$$
$$T_2 = 0.69R_2C_2$$
since $T = T_1 + T_2 \Rightarrow T = 0.69(R_2C_2 + R_1C_1)$

If $R_1 = R_2 = R \; and \; C_1 = C_2 = C \quad \Rightarrow T = 0.69(2RC) = 1.38RC$

Applications of Astable Multivibrator:

- Used as a square wave generator,
- Used as a voltage to frequency converter,
- Used as a clock for binary logic signals,
- Used in digital voltmeters and switched mode power supplies,
- As an oscillator to generate wide range of audio and radio frequencies.

L – 29. Astable Multivibrators (Conti….)

Date: ……………..

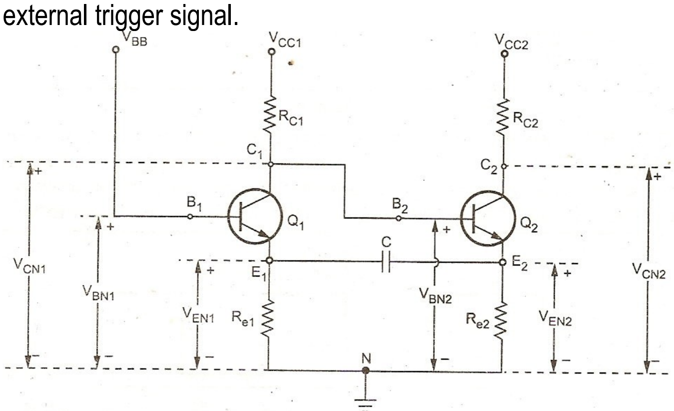
Hall: 21 Period: ……

Emitter Coupled Astable Multivibrator

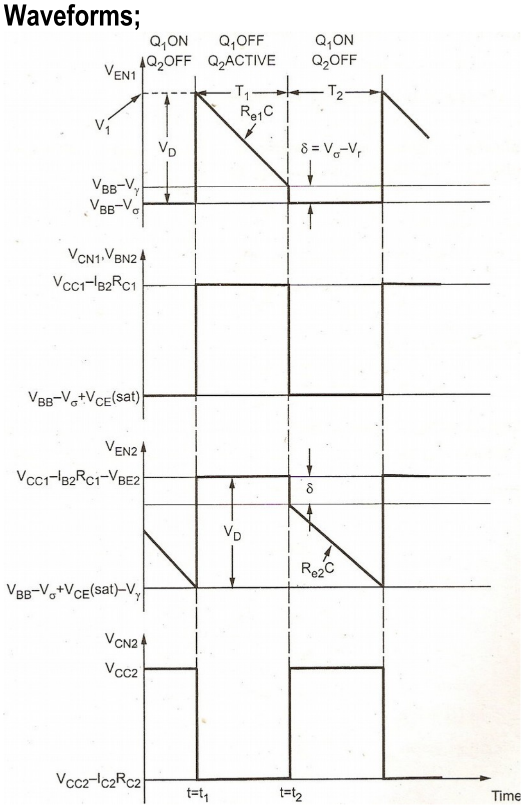
Construction:

The collector of Q_1 is connected to the base of Q_2, the capacitive coupling is used to connect emitters of Q_1 and Q_2. Additional resistances R_{e1} and R_{e2} are used in the emitter leg of Q_1 and Q_2 respectively.

The circuit has two quasi-stable states and it makes periodic switching between these states without any external trigger signal.



Waveforms;



Expression for the Time Period

Assuming the supply voltages are large compared to junction voltages;

$$T = T_1 + T_2 \Rightarrow T = \left(R_{e1} + R_{e2}\right) C \ln\left(\frac{V_{CC1}}{V_{BB}}\right) \quad \text{and}$$

$$f = \frac{1}{T} = \frac{1}{\left(R_{e1} + R_{e2}\right) C \ln\left(\frac{V_{CC1}}{V_{BB}}\right)}$$

Te frequency is not dependent on transistor parameters, and if V_{CC1} and V_{BB} are selected such that they are proportional to each other the frequency can be made insensitive to supply voltages.

Advantages of Emitter Coupled Astable Multivibrator

- No external triggering signal is required. It is self starting,
- The output can be used to drive heavy loads without affecting the circuit performance,
- The distortion in the output waveform due to transients is absent,
- Easy to synchronize since the input terminal which is the base of Q_1 is isolated;
- A single capacitor C controls the frequency; hence it is easy to change frequency by varying C. In collector coupled, it is necessary to change both C_1 and C_2

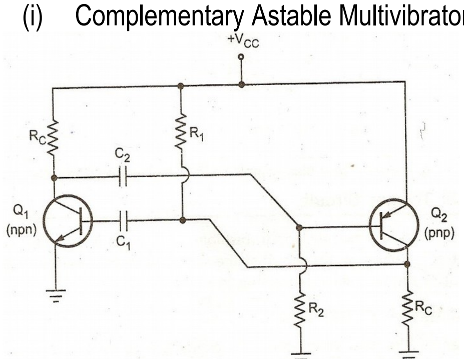
Disadvantages of Emitter Coupled Astable Multivibrator

- The number of components are more compared to the collector coupled circuit,
- The design is complicated because the quiescent points of Q_1 and Q_2 are required to be designed such that Q_1 switches between **cut-off** and **saturation** while Q_2 switches between **cut-off** and **active region**.
- Emitter resistances R_{e1} and R_{e2} cannot differ much, hence with single capacitor C, widely different T_1 and T_2 cannot be obtained.
- As design is complicated and more components are required, it is more expensive.

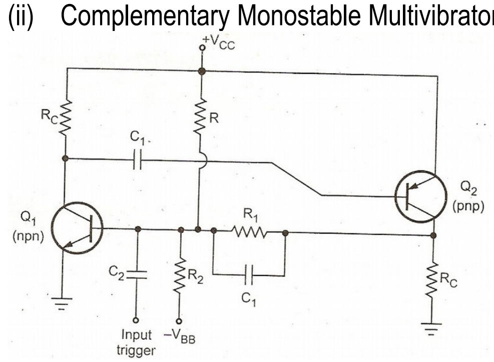
Complementary Multivibrators

A complementary multivibrator is obtained by replacing one of the transistors by its complimentary so that one transistor is $n-p-n$ while the other is $p-n-p$.

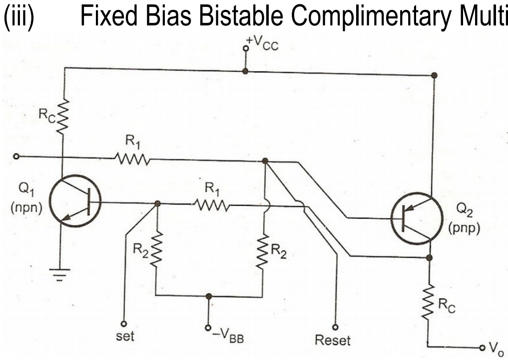
(i) Complementary Astable Multivibrator



(ii) Complementary Monostable Multivibrator



(iii) Fixed Bias Bistable Complimentary Multivibrator



L – 30. Schmitt Trigger Circuit

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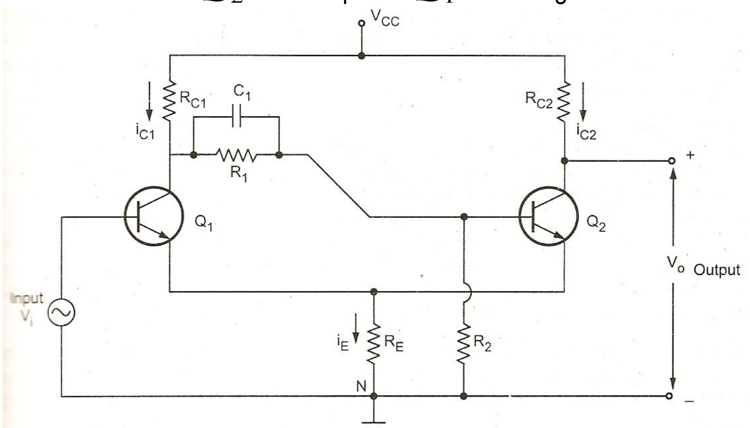
Schmitt trigger looks like a basic bistable configuration but it differs by the fact that the coupling from collector of Q_2 to the input of Q_1 is missing in this circuit.



The emitter of the two transistors are connected to each other and grounded through resistance R_E. The feedback is obtained through the resistance R_E. There exist two stable states of the output of the circuit. The resistance R_{C1} is kept smaller than R_{C2} so that regeneration cannot take place.

Operation of the circuit:

Let a sinusoidal input voltage V_i be applied to the circuit. Assume Q_2 is conducting and saturated. And as $V_i = 0$ at start, Q_1 is cut-off, hence current $i_{C1} = 0$ while $i_{C2} = \max$.

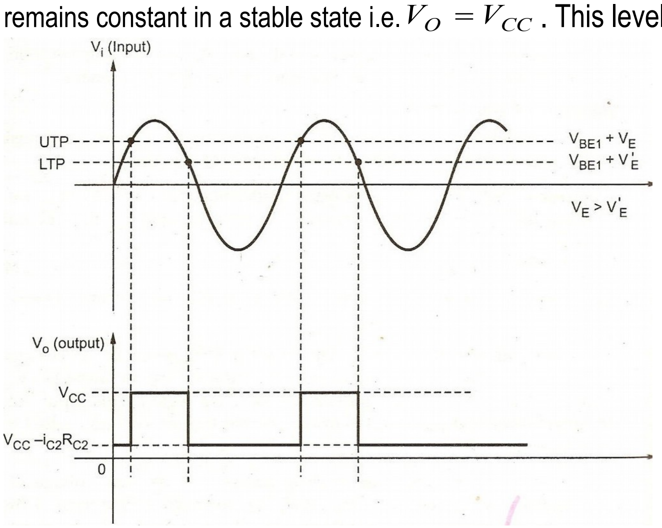
Neglecting base current of Q_2; $i_{C2} = i_E$ flowing through R_E. The drop across R_E is $i_{C2}R_E$

Therefore; $V_O = V_{CC} - i_{C2}R_C$

This drop across R_E raises the emitter voltage of Q_1 and reverse biases transistor Q_1.

Then V_i increases and to make Q_1 ON, it must increase to the level equal to cut-in voltage V_{BE1} of Q_1, plus the amount by which emitter voltage is raised V_E. When $V_i = V_{BE1} + V_E$ transistor Q_1 gets driven to active region. This input voltage level is the **upper threshold point** (**UTP**) of the Schmitt trigger.

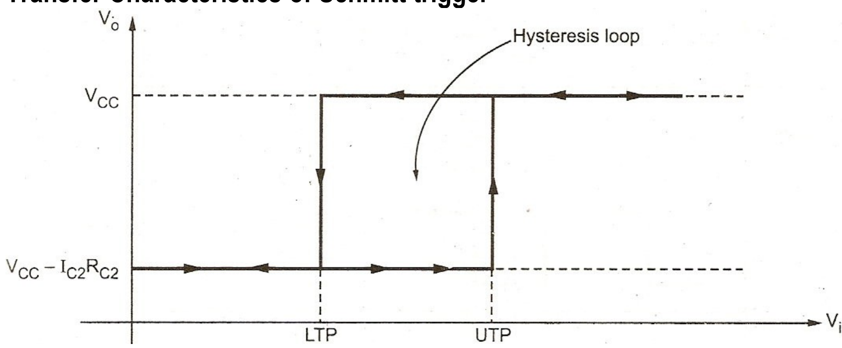
As Q_1 is **ON**, i_{C1} starts flowing. Due to drop across R_{C1} base voltage of Q_2 reduces and as the process is regenerative, Q_1 is driven into **saturation** and simultaneously Q_2 is **cu-off**. The output remains constant in a stable state i.e. $V_O = V_{CC}$. This level will not change automatically.



By this time, the collector current of Q_1 is i_{C1}, neglecting base current $i_E = i_{C1}$ which causes a drop of $V_E^{'}$ across R_E. When V_i starts decreasing, the base voltage of Q_1 starts decreasing, and when it reaches the level of $V_{BE1} + V_E^{'}$, then Q_1 stops conducting. This makes Q_2 **ON**, and instantaneously Q_2 goes into **saturation** and the output level reduces to $V_{CC} - i_{C2}R_{C2}$.

The level of V_i at which Q_1 becomes OFF and Q_2 **ON** is the **lower threshold point** (**LTP**)

Transfer Characteristics of Schmitt trigger

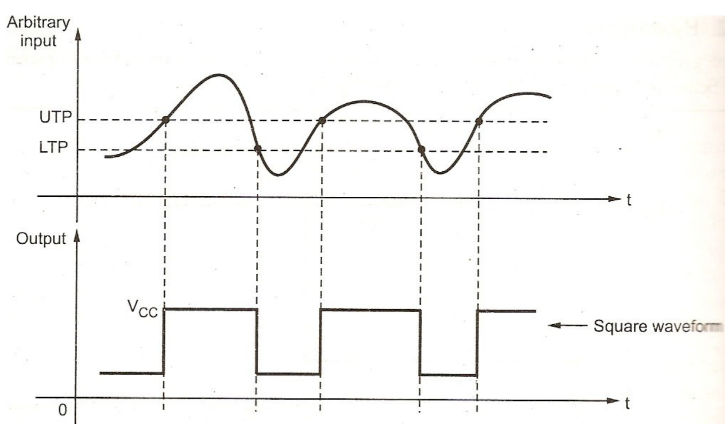


The transfer characteristics of Schmitt trigger is called **hysteresis**, i.e. once the output changes its state, it remain there indefinitely until the input voltage crosses any of the threshold levels. The difference between UTP and LTP is the width of hysteresis, i.e. **UTP – LTP = hysteresis width**.

Applications

(i) Amplitude comparator
To identify the moment at which any arbitrary waveform attains a particular reference level.

(ii) Squaring circuit
Any arbitrary (random) input can be converted to a square wave by applying the arbitrary waveform to its input. The output square wave has amplitude which is independent of the amplitude of the input wave form.



(iii) Flip-flop
When Schmitt trigger is triggered its two stable states by two alternate positive and negative pulses, it becomes a flip-flop.

Revision questions

PART - A

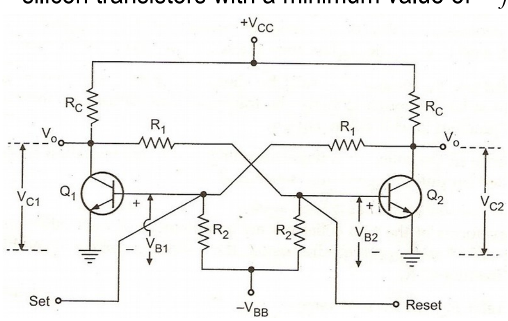
Que 1. What are the applications of Bistable multivibrators, (at least two)
Que 2. What are the applications of Astable multivibrators, (at least two)
Que 3. What are the applications of Monostable multivibrators, (at least two)
Que 4. What is the role played by commutating capacitors in bistable multivibrators?
Que 5. Draw the typical waveforms at base and collector of a Collector Coupled Astable Multivibrator.
Que 6. Draw the waveforms of a Collector Coupled Monostable Multivibrator.

PART - B

Que 1. Explain why in bistable multivibrators both transistors cannot remain in active condition.
Que 2. Using circuit diagrams differentiate fixed biased from self biased bistable multivibrators.
Que 3. Differentiate unsymmetrical triggering from symmetrical triggering.
Que 4. Brief about UTP and LTP of Schmitt trigger.
Que 5. Draw a circuit diagram of a Complementary Astable Multivibrator.
Que 6. Draw a circuit diagram of a Complementary Bistable Multivibrator.
Que 7. Draw a circuit diagram of a Complementary Monostable Multivibrator.

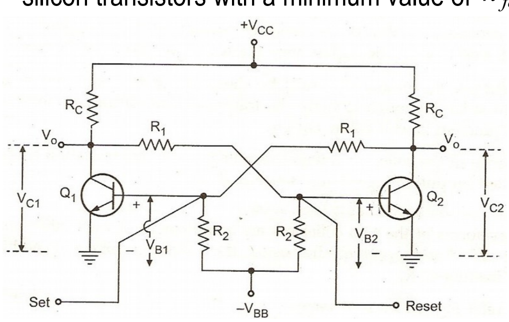
PART - C

Que 1. The fixed bias bistable multivibrator uses the following parameters; $V_{CC} = +12\,V,\ V_{BB} = -8\,V,\ R_1 = 10\,k\Omega,\ R_2 = 50\,k\Omega,\ R_C = 2.2\,k\Omega$ the transistors are silicon transistors with a minimum value of h_{fe} as 30.



Neglecting all the junction voltages, Calculate;
(i) Stable state currents $I_{C1}\ ; I_{C2}, I_{B1}, I_{B2}$
(ii) Stable state voltages; $V_{C1}, V_{C2}, V_{B1}, V_{B2}$

Que 2. The fixed bias bistable multivibrator uses the following parameters; $V_{CC} = +12\,V,\ V_{BB} = -8\,V,\ R_1 = 10\,k\Omega,\ R_2 = 50\,k\Omega,\ R_C = 2.2\,k\Omega$ the transistors are silicon transistors with a minimum value of h_{fe} as 30.



Assuming $V_{CE(Sat)} = 0.2V$; and $V_{BS(Sat)} = 0.7V$, calculate;
(i) Stable state currents $I_{C1}\ ; I_{C2}, I_{B1}, I_{B2}$
(ii) Stable state voltages; $V_{C1}, V_{C2}, V_{B1}, V_{B2}$

Que 3. Using a labeled neat circuit diagram of a bistable multivibrator, explain unsymmetrical triggering through collector, using negative pulse.
Que 4. What are the (i) advantages and (ii) disadvantages of Emitter Coupled Astable Multivibrator?
Que 5. Using a labeled neat circuit diagram of Schmitt Trigger Circuit, explain (i) operation (ii) input and output waveforms and (iii) applications of the circuit,

Que 6. Determine the (i) period and (ii) frequency of oscillation for an astable multivibrator with component values $R_1 = 1\,k\Omega,\ R_2 = 5\,k\Omega,\ C_1 = 0.1\,\mu\text{F}, C_2 = 0.03\,\mu\text{F}$

ASSIGNMENT

Que 1. Determine the values of capacitors to be used in an astable multivibrator to provide a train of pulse $2\mu\sec$ wide at a repetition of $75kHz$ with $R_1 = R_2 = 10k\Omega$.

Que 2. If $R_1 = 10k\Omega$ and $R_2 = 5k\Omega$ and $C_1 = C_2 = 0.1\mu F$, find the frequency and duty cycle of astable output.

Que 3. Circuit parameter of a fixed bias multivibrator are: $V_{CC} = V_{BB} = 5V; R_C = 1k\Omega, R_1 = 5k\Omega, R_2 = 25k\Omega$. The $n-p-n$ silicon transistors used have $(h_{fe})_{\min} = 20$. Assume all junction voltages to be zero.
Calculate stable state currents and voltages. Verify that one transistor is saturated and the other in cut-off.

Que 4. the fixed bias binary uses transistors with $(h_{fe})_{\min} = 20$. The circuit parameters are: $V_{CC} = 12V,\ V_{BB} = -3V,\ R_C = 1k\Omega,\ R_1 = 5k\Omega,\ R_2 = 10k\Omega,\ V_{CE(sat)} = 0.3V,\ V_{BE(sat)} = 0.7V$
Find the steady state voltages and currents.

Que 5. A fixed bias binary uses a d.c. supply of $\pm 12V$. The parameters of the circuit are: $R_C = 2k\Omega$, $R_1 = 10k\Omega$, and $R_2 = 47k\Omega$. The circuit uses $n-p-n$ silicon transistors with $V_{CE(sat)} = 0.1V$, and $V_{BE(sat)} = 0.7V$. Assuming $(h_{fe})_{\min} = 30$.

a. Draw the circuit diagram and obtain stable state currents and voltages assuming Q_2 **ON** and Q_1 **OFF**

b. Verify the device states.

UNIT 5. BLOCKING OSCILLATORS AND TIMEBASE GENERATORS

L – 31. Introduction and Pulse Transformers

Date: ………..…
Hall: 21 Period: ……

Definition;
A blocking oscillator is a special type of wave generator which is used to produce a single narrow pulse or a train of pulses.

Important elements of the blocking oscillator;

- An active device like transistor,
- A pulse transformer which is used to couple the output of the transistor back to the input.

The nature of such a feedback through pulse transformer is controlled by relative winding polarities of a pulse transformer. The properly controlled winding polarities produce a regenerative feedback.

Types of blocking oscillators;

- Monostable blocking oscillator,
  If the circuit is used to produce a single pulse, then the circuit is called Monostable operation of the blocking oscillator
- Astable blocking oscillator,
  If the circuit is used to produce pulse train then the circuit is called astable operation of the blocking oscillator.

Uses of blocking oscillators;
A transformer coupled configuration of a blocking oscillator is used in many practical applications which are concerned with the timing of some other circuits like;

- Frequency dividers,
- Counter circuits,
- Circuits for switching other circuits ON and OFF at specific times.

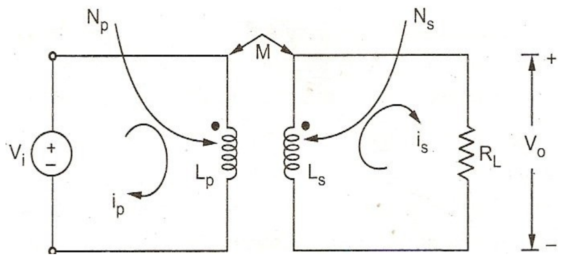
Pulse Transformer
A Pulse Transformer is a transformer which couples a source of pulses of electrical energy to the load keeping the shape of other properties of the pulse unchanged. The voltage level of the pulse can be raised or lowered b designing proper turns ratio of the pulse transformer.

Characteristics of Pulse transformers

- Iron cored
- Small in size
- Minimum leakage inductance of the pulse transformer windings.
- Very low interwinding capacitance.
- The core of pulse transformer have high permeability,
- High magnetizing inductance

Since pulse transformers are designed to handle fast waveforms like train of pulses, they can be approximated to be ideal transformers from analysis point of view.

Ideal pulse transformer model



Let

L_p – *primary* inductance
L_s – secondary inductance
M – *mutual* inductance
V_i – *source*
V_o – *output response*
R_L – *load* resistance
N_p – *primary turns*
N_s – secondary *turns*
i_p – *primary current*
i_s – secondary *current*

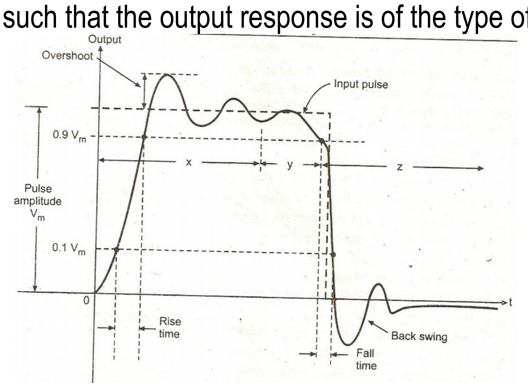
$K = \frac{M}{\sqrt{L_p L_s}}$ Coefficient of coupling between primary and secondary. For an ideal transformer $K = 1$

Let $n = \frac{N_s}{N_p}$ transformer ratio; Then for an ideal transformer, various ratios can be obtained as

$\frac{V_o}{V_i} = \frac{i_p}{i_s} = \frac{N_s}{N_p} = n = \sqrt{\frac{L_s}{L_p}}$ - The voltage and current ratios are inversely proportional to each other

Pulse response characteristics,

Due to various inductances and capacitance, the pulse transformer output is distorted. The transformers are designed such that the output response is of the type of damped oscillations.



The total pulse characteristics can be divided into three sections;

(i) Leading edge response;
At start there is an overshoot and then the pulse settles down.
- Rise time – Time required for the pulse to rise from 10% to 90% of its amplitude. The rise time sets the limit on the maximum pulse repetition rate, which can be handled by the pulse transformer.
- Overshoot – The amount by which the output exceeds its amplitude, during first attempt.

(ii) Trailing edge response - The portion of response from backswing till it settles down.

(iii) Flat top response – the portion of the response between the trailing edge and leading edge

The pulse distortion can be avoided by designing the pulse transformer with infinite magnetizing inductance L and $zero$ $\sigma\sigma$ and C $alues$ by using high permeability material for the core. Hence primary turns are kept very small.

Applications of pulse transformer;

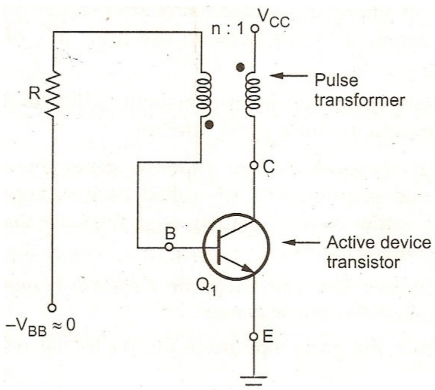
(i) Pulse generating circuits like blocking oscillators as coupling elements,
(ii) To change the amplitude and impedance level of a pulse,
(iii) For fast signal transmission required in transmission lines,
(iv) To invert the polarity of the pulse,
(v) To provide equal positive and negative pulses simultaneously using center-tapped winding
(vi) To provide d.c isolation between source and load,
(vii) To differentiate a pulse,
(viii) For coupling the stages of pulse amplifiers,
(ix) Used in digital signal transmission

L – 32. Monostable blocking oscillators using base timing

Date: ……….…..
Hall: 21 Period: ……

The blocking oscillator is closely related to the two-transistor astable circuit, except that it uses only one amplifying device, the other being replaced by a pulse transformer, which provides strong feedback at all frequencies.
As a Monostable, it was useful in 1950's for producing what were then, **short pulses**, in the micro-second range. It was faster than the Abraham-Bloch Monostable.
In the transistor era, it fell from grace because it could not be miniaturized, since it requires a transformer.

Construction:



- One winding of the pulse transformer is in the collector circuit while the other in the base circuit.
- Number of turns of transformer winding in base circuit is n times the number of turns of transformer winding in the collector circuit.
- Pulse transformer core is made up of iron or ferrites (not indicated to avoid diagram complexity)
- A Monostable blocking oscillator using base timing is a.k.a. **triggered transistor blocking oscillator**.

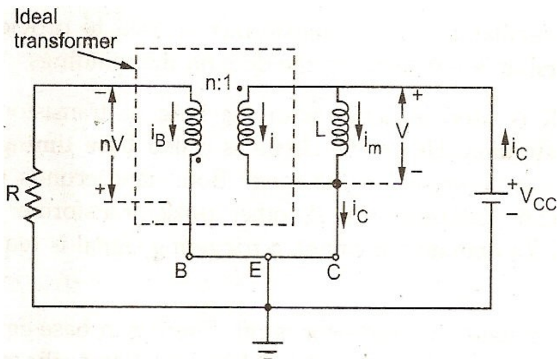
- A pulse transformer is used to produce polarity inversions indicated by dots on the windings.
- Resistance R is used in series with base of transistor controls the timing i.e. pulse duration, hence the circuit is a.k.a. base timing blocking oscillator.
- The pulse width depends on R, pulse transformer parameters and other circuit parameters.
- To operate the circuit, a trigger signal is required to the collector momentarily.

A pulse transformer accepts a pulse at one winding and tries to produce a similar at the other winding. It can produce inverted pulse of that applied to one winding, if winding polarities are properly designed.

Operation:
In the quiescent state, the transistor is OFF. Thus if a small voltage pulse is applied to the collector, then as the turns ratio of the pulse transformer $n:1$, it can produce a large signal which is enough to start the operation of the circuit.

V_{BB} is applied at the base to keep $B-E$ junction reverse biased to avoid false triggering that might occur due to noise voltage at increased temperature.
- Let a triggering signal be applied momentarily to the collector of Q_1 such that the collector voltage level reduces suddenly, then the voltage applied at the primary (collector) produces inverted signal at the secondary (base). Thus base potential increases as collector potential decreases.
- If $V_B > V_{cut-off}$ the transistor starts conducting drawing current from V_{CC}.
- This increases collector current which leads to increase in drop across transformer winding in collector, which further lowers collector potential and increases base potential.
- If a.c. loop gain $A_{ac} > 1$, regenerative action takes place and Q_1 gets driven into saturation from OFF state. The action happens very quickly.

Mathematical analysis of the circuit



Using the equivalent circuit and neglecting;

- Leakage inductance σ
- Saturation voltages $V_{BE(sat)}$, $V_{CE(sat)}$

For an ideal transformer $\frac{V_2}{V_1} = n = \frac{i}{i_B}$

Where

V_1 -primary voltage (across collector winding)

V_2 -secondary voltage (across base winding)

From current ratio $i = n.i_B \quad \therefore i - n.i_B = 0$

For a transistor in saturation $V_1 = V = V_{CC}$, $\Rightarrow V = V_{CC}$

According to transformer ratio $V_2 = n.V_1 = nV = nV_{CC}$

Applying KVL to the base circuit;

$$i_B R = nV = nV_{CC} \quad \Rightarrow i = n \times \frac{nV_{CC}}{R} = \frac{n^2 V_{CC}}{R}$$

The voltage V across collector winding = voltage across magnetizing inductance L

$$V = L\frac{\partial i_m}{\partial t} \quad \Rightarrow \quad \partial i_m = \frac{V}{L}\partial t \quad \therefore \int \partial i_m = \frac{V}{L}\int_0^t \partial t, \quad hence \ i_m = \frac{V}{L}t = \frac{V_{CC}}{L}t$$

From the equivalent circuit $i_c = i + i_m = \frac{n^2 V_{CC}}{R} + \frac{V_{CC}}{L}t$

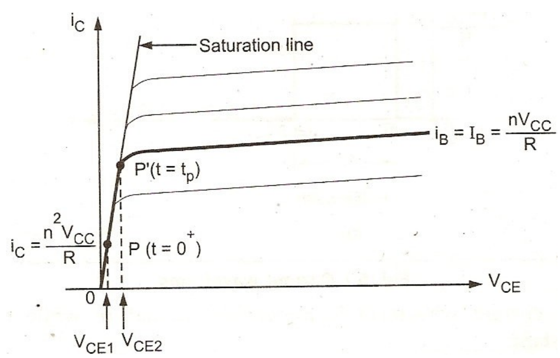
Just after the transistor enters into saturation $(t = 0^+)$;

The currents i_B, i_c are given;

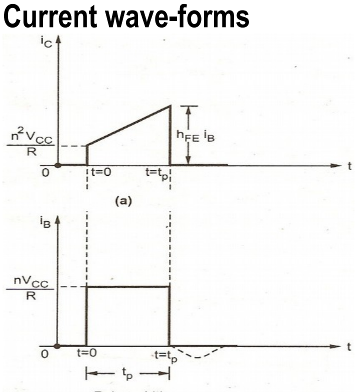
$$i_B = \frac{nV_{CC}}{R}, \ and \ i_c = \frac{n^2 V_{CC}}{R}$$

For $t > 0$, $i_B = \frac{nV_{CC}}{R}$: where P – operation point

At $t = t_p$, the transistor enters into cut-off and base current is reduced to zero



Current wave-forms



Pulse width

At point P' the pulse terminates, hence $i_c = h_{fe}.i_b$.

From $i_c = i + i_m \quad \Rightarrow \quad \frac{n^2 V_{CC}}{R} + \frac{V_{CC}}{L}t_p = h_{fe}\frac{nV_{CC}}{R}$

$$\therefore \ \frac{V_{CC}}{L}t_p = h_{fe}\frac{nV_{CC}}{R} - \frac{n^2 V_{CC}}{R} = \frac{nV_{CC}}{R}\left(h_{fe} - n\right)$$

$$\therefore \ t_p = \frac{nL}{R}\left(h_{fe} - n\right) \approx \frac{nL}{R}h_{fe}$$ since $h_{fe} >> n$

Conclusions:

From the equation of the pulse width $t_p = \frac{nL}{R} h_{fe}$ the following disadvantages can be noted;

- The pulse width is a linear function of h_{fe} which is temperature dependent,
- The value of h_{fe} changes from transistor to transistor and hence the pulse width gets affected due to transistor replacement.

The Monostable blocking oscillators using base timing circuit is not used if a stable pulse width is required

L – 33. Monostable blocking oscillators using emitter timing

Date: ……………..…

This is a circuit that makes pulse width insensitive to h_{fe}. It uses a resistance in the emitter circuit which controls the pulse width. The circuit is a.k.a. **triggered transistor blocking oscillator** with emitter timing.

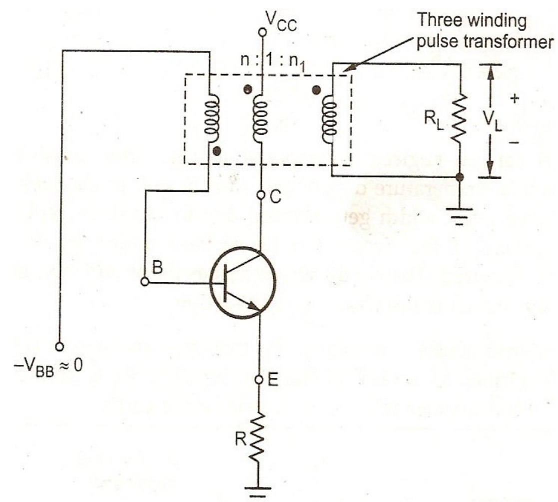
Construction:

The pulse transformer used is a **three winding transformer**;

- Primary winding (collector winding)
- Base circuit winding ($n-times$ as many turns as collector winding)
- A load resistance R_L is connected across third winding which has $n_1 - times$ as many turns as the collector winding

The base and collector windings must produce polarity inversion as indicated by dots wile relative direction of third winding can be arbitrary.

The resistance R_L acts as load and also helps in improving damping.



Mathematical analysis:

Using the equivalent circuit, assuming an ideal transformer and neglecting; Leakage inductance σ, Capacitance, Winding resistance and Saturation voltages $V_{BE(sat)}, V_{CE(sat)}$.

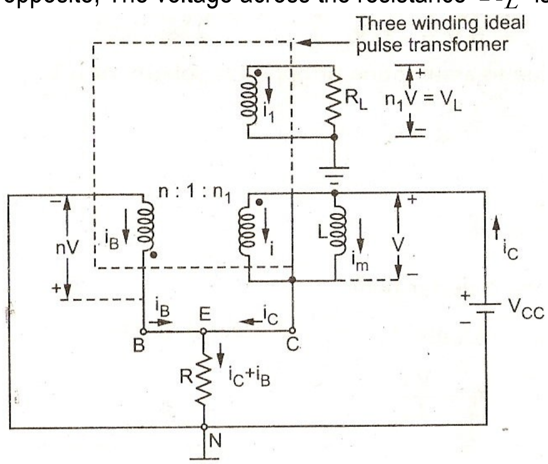
For analysis simplicity, assume;

Number of turns of primary of transformer (collector winding) $= 1$

Number of turns of secondary of transformer (base winding) $= n$

Number of turns of tertiary winding of transformer (circuit used to connect R_L) $= n_1$

If V is the voltage across the primary collector winding, when transistor is in saturation, the corresponding voltage across secondary winding in base circuit is nV. Due to inversion, polarities of V and nV are opposite; The voltage across the resistance R_L is n_1V and the polarity is the same as that of V.



Applying KVL to the outermost loop (including base and collector loops together);

$$-V_{CC} + nV + V = 0 \quad \therefore \quad V = \frac{V_{CC}}{n+1}$$

Applying KVL to the base circuit

$$+nV-(i_B+\mathrm{i}_C)R=0 \quad \therefore \quad nV=(i_B+\mathrm{i}_C)R \;\Rightarrow\; (\, i_B+\mathrm{i}_C)=\mathrm{i}_E=\frac{nV}{R}=\frac{nV_{CC}}{(n+1)R} \quad \text{...(i)}$$

The current i_E is the emitter current coming from the active device. From the equation, $\mathrm{i}_E = const$.

According to the principle of sum of the ampere turns in an ideal transformer which is always zero;

$\therefore$ Ampere turns of primary $=i\times 1=i$

Ampere turns of secondary $=i_B\times n=n.i_B$

Ampere turns of tertiary $=i_1\times n_1=i_1 n_1$ ……(ii)

The signs of primary and tertiary are the same while sign of secondary ampere turns is opposite to primary and tertiary ampere turns

$$\therefore \quad i-n.i_B+i_1 n_1=0 \quad \text{……(iii)}$$

From the load circuit loop;

$V_L=n_1\,V=-i_1R_L$ - Negative sign is due assumption of opposite polarity of $n_1\,V$; $\therefore\; i_1=-\dfrac{n_1\,V}{R_L}$ ..(iv)

Substituting (iii) in (i) $\quad i-n.i_B+\left(-\dfrac{n_1V}{R_L}\right)n_1=0 \;\Rightarrow\; i-n.i_B-\dfrac{n_1^2V}{R_L}=0$ ……. (v)

Applying KVL at the collector node;

$i_c=i+i_m$ and $i_m=\dfrac{V}{L}t$ *(as derived in L - 32)* hence $i=i_c-\dfrac{V}{L}t$ ….(vi)

Substituting (iv) in (v) $\quad i_c-\dfrac{V}{L}t-n.i_B-\dfrac{n_1^2V}{R_L}=0$ …..(vii)

Solving (i) and (vii) simultaneously, individual expressions for i_c and i_B are obtained

$$\begin{aligned} i_B+\mathrm{i}_C&=\frac{nV_{CC}}{(n+1)R}\\ i_c-\frac{V}{L}t-n.i_B-\frac{n_1^2V}{R_L}&=0\end{aligned} \quad \Rightarrow i_B+\frac{V}{L}t+n.i_B+\frac{n_1^2V}{R_L}=\frac{nV_{CC}}{(n+1)R}$$

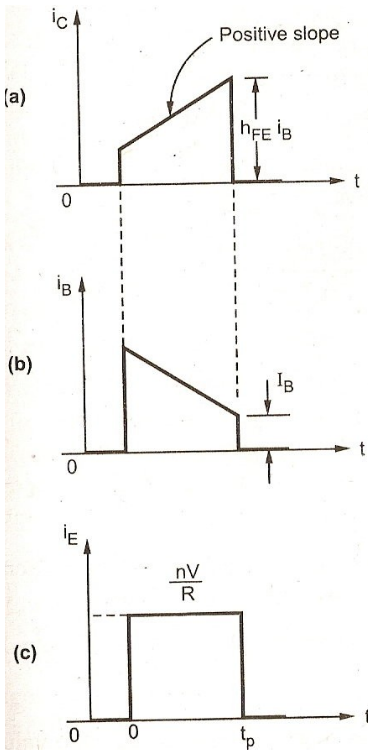
Hence $(n+1)i_B=\dfrac{nV_{CC}}{(n+1)R}-\dfrac{V}{L}t-\dfrac{n_1^2V}{R_L}$ ….(viii)

Substituting $V=\dfrac{V_{CC}}{n+1}$ in $(n+1)i_B=\dfrac{nV_{CC}}{(n+1)R}-\dfrac{V}{L}t-\dfrac{n_1^2V}{R_L}$

$$(n+1)i_B=\frac{nV_{CC}}{(n+1)R}-\frac{V_{CC}}{(n+1)L}t-\frac{n_1^2V_{CC}}{(n+1)R_L} \quad \Rightarrow i_B=\frac{V_{CC}}{(n+1)^2}\left[\frac{n}{R}-\frac{t}{L}-\frac{n_1^2}{R_L}\right] \text{...(ix)}$$

Substituting i_B in $i_c-\dfrac{V}{L}t-n.i_B-\dfrac{n_1^2V}{R_L}=0$; we obtain $i_C=\dfrac{V_{CC}}{(n+1)^2}\left[\dfrac{n^2}{R}-\dfrac{t}{L}-\dfrac{n_1^2}{R_L}\right]$ ..(x)

Due to presence of t in both expressions for i_B and i_C; the i_B and i_C waveforms are trapezoidal. The slope of i_B is negative while the slope of i_C is positive. The current $\mathrm{i}_E = const$ during the pulses.



Expression of the pulse width;

At $(t = 0^+)$ the current i_C is small, as time passes, the current i_C increases, at the same time current i_B starts decreasing. At point P' when $i_B = I_B$ the pulse ends.

This time corresponds to $t = t_p$ and $i_c = h_{fe}.i_b$

Substituting $i_B = \frac{V_{CC}}{(n+1)^2}\left[\frac{n}{R} - \frac{t}{L} - \frac{n_1^2}{R_L}\right]$ and

$i_C = \frac{V_{CC}}{(n+1)^2}\left[\frac{n^2}{R} + \frac{t}{L} + \frac{n_1^2}{R_L}\right]$ in $i_c = h_{fe}.i_b$ with $t = t_p$

we get $\frac{V_{CC}}{(n+1)^2}\left[\frac{n^2}{R} + \frac{t_p}{L} + \frac{n_1^2}{R_L}\right] = \frac{V_{CC}}{(n+1)^2}\left[\frac{n}{R} - \frac{t_p}{L} - \frac{n_1^2}{R_L}\right]h_{fe}$

Therefore the required **pulse width** $t_p = \frac{nL}{R}\frac{(h_{fe} - n)}{(h_{fe} + 1)} - \frac{n_1^2 L}{R_L}$

Usually n << 1 and $h_{fe} >> n$ hence $\frac{(h_{fe} - n)}{(h_{fe} + 1)} \approx 1$

Therefore the **pulse width** $t_p = \frac{nL}{R} - \frac{n_1^2 L}{R_L}$: this shows that the pulse width t_p is not dependent on the parameter h_{fe} and depends on the passive parameters $n,\ L,\ R, and\ R_L$. Thus Monostable blocking oscillator with emitter timing gives a stable pulse width.

Limiting value of R_L

From the equation $t_p = \frac{nL}{R} - \frac{n_1^2 L}{R_L}$ if $\frac{n_1^2 L}{R_L}$ is greater than $\frac{nL}{R}$, the value of t_p becomes negative. Since negative time is not possible in practice, the value of t_p must be positive. For positive t_p and regenerative actions to take place in the circuit, $\frac{n_1^2 L}{R_L} < \frac{nL}{R}\frac{(h_{fe} - n)}{(h_{fe} + 1)}$

Therefore $R_L > \frac{n_1^2 R}{n}\frac{(h_{fe} + 1)}{(h_{fe} - 1)}$

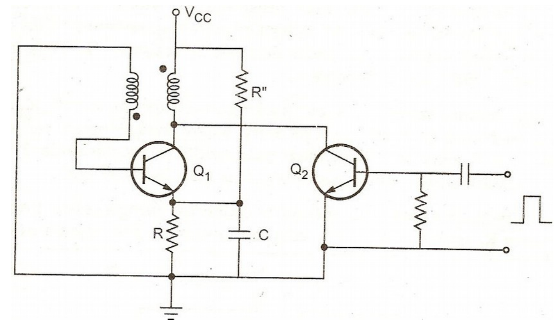
L – 34. Triggering Circuits for Monostable Blocking Oscillator Frequency Control using Core Saturation,

Date: ……………..

Hall: 21 Period: ……

Triggering Circuits

(i) Triggering Circuit using Transistor.



For triggering, it is necessary to lower the collector voltage of Q_1 momentarily by the triggering circuit. If momentarily a positive-going pulse is applied to the base of Q_2, will drive Q_2 into saturation, hence providing a short across the collector of Q_1 which momentarily brings the collector voltage of Q_1 to zero.

Due to phase inversion, such a voltage gets induced in the secondary which makes the base of Q_1 positive, hence it starts conducting.

Thereafter, the working of Monostable Blocking Oscillator starts. When the pulse is removed from Q_2, Q_2 becomes off (open circuit) hence the triggering source causes no interaction with the blocking oscillator.

N.B.

(i) The applied trigger pulse must have enough steep leading edge to ensure that the induced transformer voltage will bring Q_1 out of cut-off region.

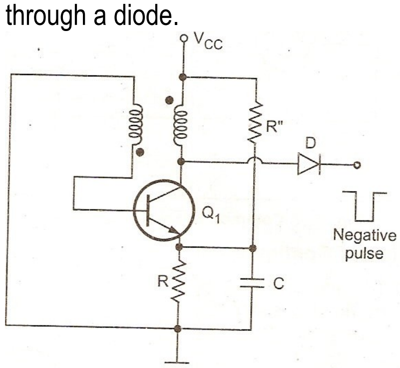
(ii) The resistance R'' is kept very large compared to R to avoid false triggering due to noise voltages and need of V_{BB} is avoided.

(iii) A small capacitor C connected across R improves rise time of the pulse which is very important in blocking oscillators.

(iv) The trigger pulse should be wider than the oscillator pulse, because if the trailing edge of the trigger pulse has short fall time, then it may terminate the blocking oscillator pulse prematurely.

(ii) Triggering Circuit using Diode.

A positive-going pulse is applied at the base of Q_1 or a negative-going pulse to the collector of Q_1 through a diode.



The pulse is applied momentarily to makes diode forward biased, which lowers the collector voltage of Q_1 momentarily which in-turn brings Q_1 out of cut-off state. When circuit works, during pulse formation, the diode D is reversed biased, this avoids the blocking oscillator to react with the triggering source or influence the triggering source or oscillator.

N.B.

- The trigger pulse should be wider than the oscillator pulse, because if the trailing edge of the trigger pulse has short fall time, then it may terminate the blocking oscillator pulse prematurely.

Frequency Control

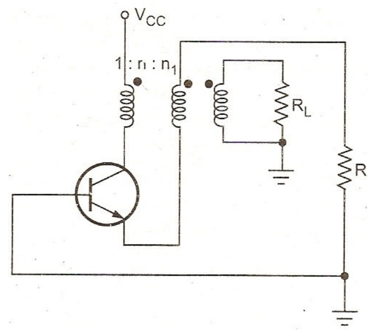
Apart from the common emitter timing method of pulse control, other circuits to obtain and control the pulse are:

- Use of Common Base Configuration,
- Use of Common Collector Configuration,
- Use of Core Saturation method,
- Use of Shorted Delay Line,

a. Frequency Control using **Common Base Configuration**,

The primary winding of pulse transformer is placed in the collector while the other winding is placed in the emitter. The timing resistor R is placed in emitter to avoid dependence of t_p on h_{fe} of transistor.

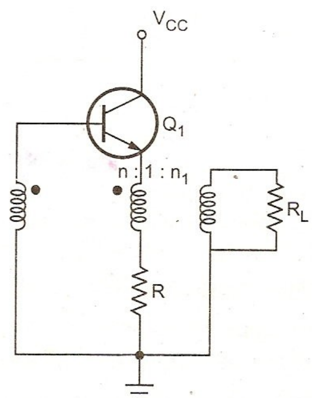
N.B. There is no phase inversion between the voltages associated with the two windings.



b. Frequency Control using **Common Collector Configuration**,

The primary winding is placed in the emitter of the transistor Q_1 while the secondary winding is placed in the base of the transistor Q_1. The timing resistor R is placed in the emitter to avoid dependence of t_p on h_{fe} of the transistor. The load is connected using tertiary winding of the pulse transformer using n_1 turns.

N.B. There is no phase inversion between the voltages associated with the two windings as in common base configuration.



c. Frequency Control using **Core Saturation**,

The core having characteristics such that it saturates when the flux density B in the core reaches a maximum value B_m. Inductance L is inversely proportional to the current $L \propto \left({}^{1}\!/_{i} \right)$.

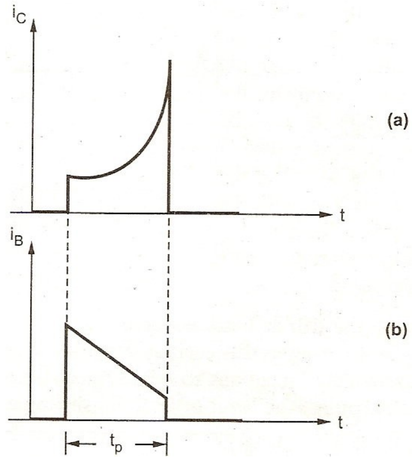
i.e. when current increases, the inductance L decreases, and as current increases, the magnetic field strength H increases which increases flux density B. Hence $L \Rightarrow 0$ as $B \Rightarrow B_m$

(NB: in earlier derivation of t_p magnetizing Inductance L was assume constant for simplicity)

As derived earlier;

$$i_C = \frac{V_{CC}}{(n+1)^2}\left[\frac{n^2}{R} + \frac{t}{L} + \frac{n_1^2}{R_L}\right] \text{ and } i_B = \frac{V_{CC}}{(n+1)^2}\left[\frac{n}{R} - \frac{t}{L} - \frac{n_1^2}{R_L}\right];$$

As $L \Rightarrow 0$ i_C increases to a value $h_{fe} i_B$ when the pulse ends, while i_B decreases to zero. Hence as the core saturates at $t = t_p$, the pulse ends. Waveforms of i_C and i_B due to core saturation are no longer trapezoidal.



Derivation for the pulse width t_p

Assuming that the pulse ends when the core is completely saturated;

Let; N = number of turns in collector winding

ϕ = magnetic flux in the core

A = cross-sectional area of core;

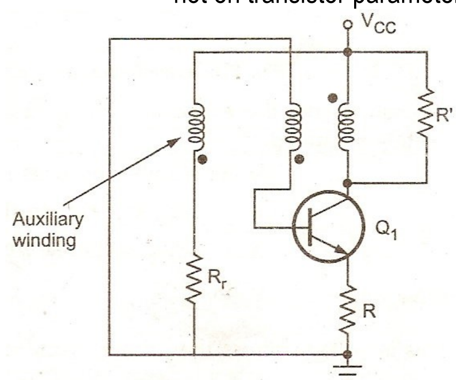
$V = N\frac{\partial \phi}{\partial t} = \frac{V_{CC}}{n+1}$ voltage across primary winding

Now $\phi = BA$

Therefore, $\frac{V_{CC}}{n+1} = NA\frac{\partial B}{\partial t} \Rightarrow \partial t = \frac{(n+1)NA}{V_{CC}}\partial B$

Integrating between limits $t = 0$ **to** $t = t_p$, $\Rightarrow \int \partial t = \frac{(n+1)NA}{V_{CC}}\int_0^{t_p} \partial B$

$t_p = \frac{(n+1)NA}{V_{CC}}B_m$ - Pulse duration depends on the supply voltage and characteristics of the core and not on transistor parameters. This is the advantage of the method.



Applications of blocking oscillator

- To obtain abrupt pulse from slowly varying input triggering voltage, Monostable blocking oscillator is used.
- An Astable blocking oscillator is used as a main device to supply triggers for synchronization of a system having pulse type waveforms.
- To produce large peak power pulses, both Monostable and Astable oscillators can be used.
- The output winding can be isolated from ground whenever required by using tertiary winding of pulse transformer.
- Output pulses of either polarity can be obtained due to the use of tertiary winding of pulse transformer.
- Blocking oscillators are used as frequency divider circuits or as counter circuit.
- Blocking oscillators can be used as low impedance switch used to discharge a capacity very quickly.
- The output of oscillator can be used to produce gating waveforms.

L – 35 RC and RL wave shaping circuits,

Date: ……………..…
Hall: 21 Period: ……

Introduction

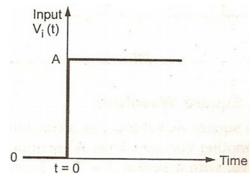
The reproduction of the input waveform at the output is possible only in case of sinusoidal signals and not in case of other signals. The process by which the shape of a sinusoidal signal is changed by passing the signal through a network consisting of linear elements is called linear wave shaping.

Examples of commonly occurring non-sinusoidal waveforms;

(i) Step waveform

Has a characteristic that its value is zero for all $t < 0$ and its value A is constant for all $t > 0$, and the switch over from the zero value to A units takes instantaneously at $t = 0$.
Mathematical expression;

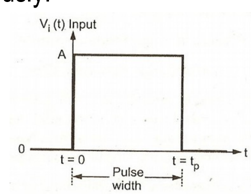
$$V_i(t) = 0 \quad t < 0$$
$$V_i(t) = A \quad t \geq 0$$



(ii) Pulse waveform

The waveform has the value zero for all $t < 0$ and for all $t > t_p$, while between $t = 0$ and $t = t_p$ its value is A units. The transition from zero to level A takes place at $t = 0$ instantaneously, while transition from A to 0 takes place at $t = t_p$ instantaneously.
Mathematical expression;

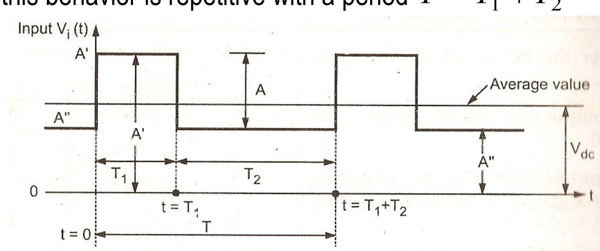
$$V_i(t) = 0 \quad for \quad t < 0 \quad and \quad t > t_p$$
$$V_i(t) = A \quad for \quad 0 \leq t \leq t_p$$



(iii) Square waveform

Has a constant level A' maintained for a time period T_1 while it has another constant level A'' maintained for a time period T_2 and this behavior is repetitive with a period $T = T_1 + T_2$
Mathematical expression;

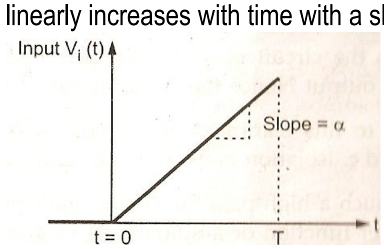
$$V_i(t) = A' \quad for$$
$$V_i(t) = A'' \quad for$$



(iv) Ramp waveform

A waveform which is zero for all $t < 0$ and then linearly increases with time with a slope α for all $t > 0$
Mathematical expression;

$$V_i(t) = 0 \quad for \quad t < 0$$
$$V_i(t) = \alpha\, t \quad for \quad t \geq 0$$



Effect of RC and RL linear networks on different non-sinusoidal inputs

a. RC **circuit**

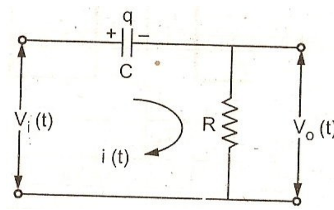
RC circuits can be categorized as

(i) High pass RC circuits
(ii) Low pass RC circuits

High pass RC circuits

Construction;

The output is taken across the resistance R. The input is $V_i(t)$, while the output is $V_o(t)$. "q" is the charge on the capacitor C. The capacitive reactance offered by the capacitor depends on the frequency $X_C = \frac{1}{2\pi fC}$ where f frequency of input waveform.



Functioning of the circuit;

As X_C is inversely proportion to f, the reactance decreases as frequency increases. At very high frequency, the capacitor almost acts as a short circuit and all the input appears at the output. At zero frequency (i.e. for d.c) the reactance becomes infinite and hence offers open circuit.
Thus the circuit obstructs the low frequencies while it allows high frequencies to reach the output. Hence the circuit is a High Pass RC circuit.

Uses of the circuit;

Due to this feature the circuit is commonly used as the coupling circuit to provide d.c. isolation between input and output.

Cut-off frequency;

The magnitude of the ratio of the output to input (a.k.a. transfer function, a.k.a. amplification, a.k.a. gain)

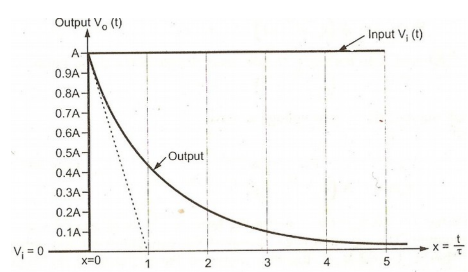
$$|A| = \frac{1}{\sqrt{1+\left(\frac{f_1}{f}\right)^2}}; \; where \; f_1 = \frac{1}{2\pi RC}$$, f = input frequency

The frequency f_1 where the magnitude of the capacitive reactance X_C is equal to the resistance R i.e., the gain $|A| = \frac{1}{\sqrt{1+1^2}} = \frac{1}{\sqrt{2}} = 0.707$ is called the lower 3-dB frequency or cut-off frequency or corner frequency

Effect on Step input voltage

When the excitation is applied to the circuit, the current starts flowing instantaneously, as the capacitor charges exponentially, the current decays exponentially and the output voltage also decays exponentially. When the capacitor charges equal to the input voltage, current stops and in steady state the output voltage is zero.

$$V_o(t) = V_i(t)e^{-t/RC} = Ae^{-t/RC}$$



Effect on Pulse input voltage

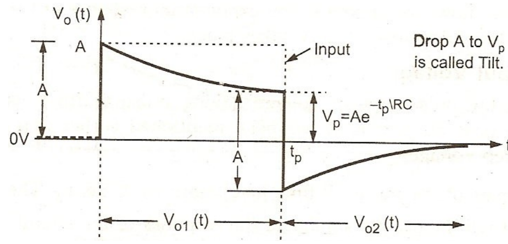
When a pulse type voltage having pulse width t_p is applied to the input of a High pass RC circuits, considering the pulse as a sum of two step voltages;

The response of the circuit to the first part of the input;

$$V_{o1}(t) = Ae^{-t_p/RC} = V_P \quad at \;\; t = t_p$$

The response of the second part of the input for $t > t_p$

$$V_{o2}(t) = A\left(e^{-t_p/RC} - 1\right)e^{-(t-t_p)/RC}$$



Effect on Square wave input voltage

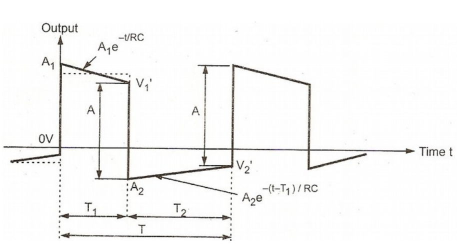
The square wave response high pass RC circuit can be obtained from the equation

- The voltage level of the positive part of the output square wave is A_1, hence after time T_1

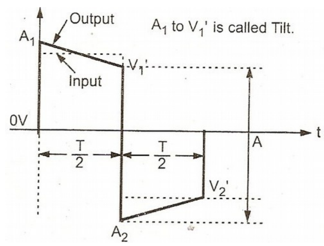
  $V_1^{'} = A_1 e^{-T_1/RC}$

- Then there will be an exponential decay of the output towards zero for the time period T_2,

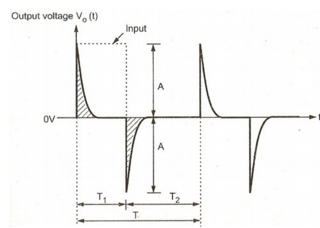
  $V_2^{'} = A_2 e^{-T_2/RC}$ The behaviour repeats with the time period T



General square wave response



Response to symmetrical square wave

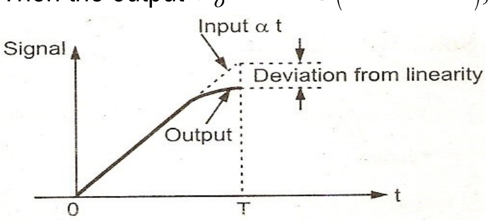


RC circuit with small τ

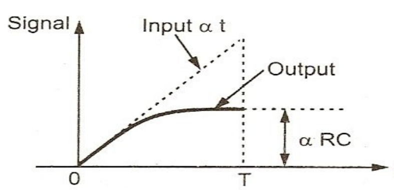
Effect on Ramp input voltage

Let the input ramp $V_i(t) = \alpha\, t \;\; for \;\; t \geq 0$,

Then the output $V_o = \alpha\, RC\left(1 - e^{-t/RC}\right); \;\; where \;\; V_o = 0 \;\; at \;\; t = 0$



Response for $\dfrac{RC}{T} >> 1$



Response for $\dfrac{RC}{T} << 1$

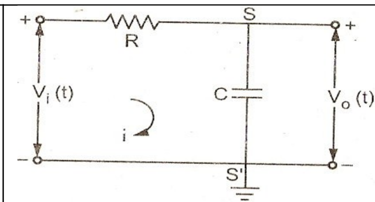
High Pass RC circuit as a Differentiator

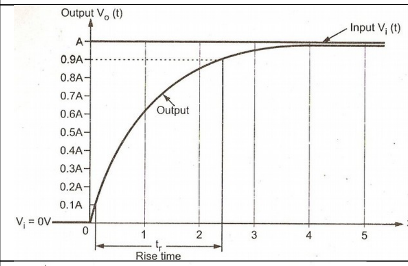
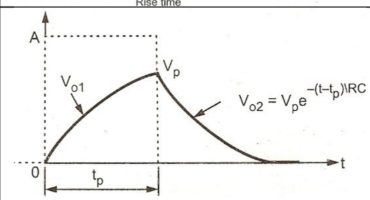
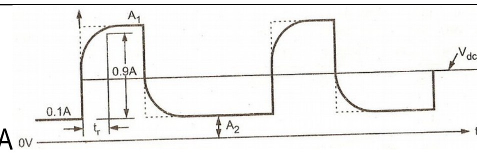
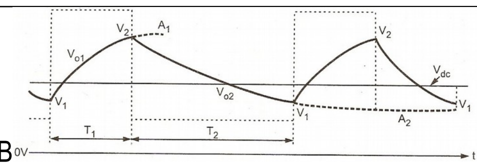
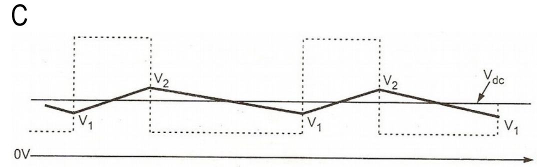
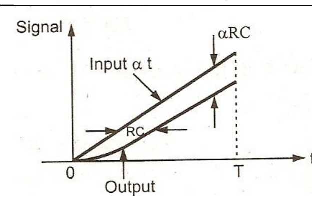
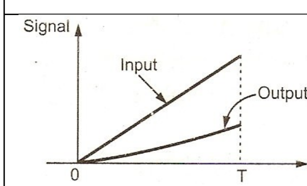
For a High Pass RC circuit, if the time constant is very small as compared to the time required by the input signal to make an appreciable change, then the circuit will act as a differentiator. Under this case the drop across R is negligible compared to drop across C. thus entire input voltage can be considered to be appearing at C.

Then the current $i = C\dfrac{\partial V_C}{\partial t} = C\dfrac{\partial V_i}{\partial t}$, The output which is the drop across R is; $V_o = iR = RC\dfrac{\partial V_i}{\partial t}$

The equation shows that the output is proportional to the derivative of the input; hence the circuit is a **differentiator.**

<u>Low pass</u> RC circuits

Input voltage $V_i(t)$, output voltage $V_0(t)$ is taken across capacitor C. The capacitive reactance X_C depends on frequency. At high frequency the capacitor acts as a virtual short and hence output falls to zero. Thus high frequencies get attenuated and low frequencies pass readily	

Magnitude of transfer function (gain)	$\lvert A\rvert = \dfrac{1}{\sqrt{1+\left(\dfrac{f}{f_2}\right)^2}}$; $where\ \ f_2 = \dfrac{1}{2\pi RC}$
Effect on Step input voltage of magnitude A ▪ Output voltage $V_o(t) = A\left(1-e^{-t/RC}\right)$ ▪ Rise time: time required by the output response to rise from $10\% - 90\%$ of its final steady value.	
Effect on Pulse input voltage of amplitude A $V_{o1}(t) = A\left(1-e^{-t/RC}\right) \quad for\ \ 0 < t < t_p$ $V_{o1}(t) = A\left(1-e^{-t_p/RC}\right) = V_P \quad at\ \ t = t_p$ $V_{o2}(t) = V_p e^{-(t-t_p)/RC} \quad for\ t > t_p$	
Effect on Square wave input voltage of amplitude A $V_{o1} = A_1 + (V_1 - A_1)e^{-t/RC}; \quad V_{o2} = A_2 + (V_2 - A_2)e^{-(t-}$	 A
B – Output when time constant RC is comparable with T	 B
C – Output when time constant RC is largest	C 
Effect on Ramp input voltage of amplitude A $V_o = \alpha\,(t - RC) + \alpha RCe^{-t/RC}$	 $RC << T$
	 $RC >> T$

Low pass RC circuit as an integrator;

For the Low pass RC circuit, if the time constant is very large as compared to the time required by the input signal to make appreciable change, the circuit acts as an **integrator**;

Under this case, the drop across C is negligible compared to the drop across R.

Thus the entire input $V_i(t)$ can be assumed to be appearing across R. Then $V_R = V_i = iR \therefore i = \frac{V_i}{R}$

Hence the output voltage i.e. voltage across the capacitor C $V_o = V_C = \frac{1}{C}\int i\partial t = \frac{1}{RC}\int V_i(t)\partial t$

Thus the output is proportional to the integration of the input; hence the circuit is an **integrator**.

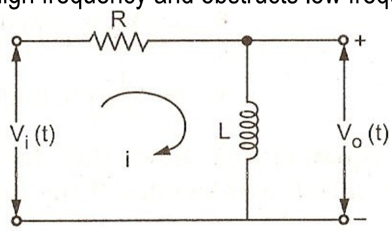
b. RL **circuit**

If the capacitor C and resistor R of the RC circuit are replaced by resistor R' and inductor L, respectively, and if the time constant $\frac{L}{R'}$ is equal to the time constant RC, then all the results derived earlier are applicable without any change.

Practically RL are rarely used if a large time constant $\frac{L}{R'}$ is required because, to get a large time constant L should be large. This is possible only with iron core inductor which is very large, heavy and expensive compared to capacitor for a similar application.

High pass RL circuits

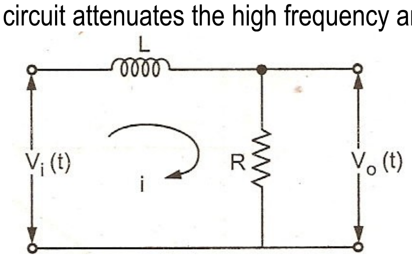
The output $V_0(t)$ is taken across inductor L. The inductive reactance offered by the inductor L directly proportional to the input frequency; $X_L = 2\pi fL$. At $f = 0$,i.e. d.c. input $X_L = 0$, the inductor behaves as a short, hence in this circuit, low frequency component cannot reach to the output. The circuit passes high frequency and obstructs low frequency.



Low pass RL circuits

The inductive reactance offered by the inductor L directly proportional to the input frequency; $X_L = 2\pi fL$.

At $f = 0$, i.e. d.c. input $X_L = 0$, the inductor behaves as a short circuit, hence the maximum current flows through the circuit and the output is maximum. At high frequencies, the reactance is very high and the inductor acts as an open circuit and current falls to zero. The output falls to zero for high frequency. The circuit attenuates the high frequency and passé low frequency.



L – 36 Linear Time-Base or Sawtooth Generators (Sweep) Circuits

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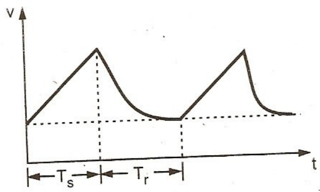
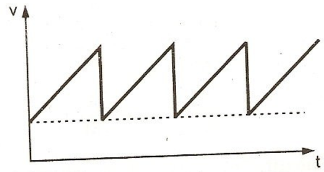
Definitions:

- Linear Time-Base or Sawtooth Generators (Sweep) Circuits
  Circuits which provide output wave form with linear variation of voltage or current with respect to time.



(a) Typical timebase or sweep voltage



(b) A sawtooth voltage waveform

The voltage starting from some initial value increases linearly with time to a maximum value, after which it returns again to its initial value.

- Restoration time T_r (a.k.a. return time, a.k.a. flyback time)
  The time required for the return to the initial value

- Sweep time T_s
  The period during which voltage increases linearly,

In such circuit, if restoration time is extremely short in comparison with sweep time ($T_r << T_s$) the waveform is called sawtooth voltage waveform.

- Sawtooth generator or ramp generator;
  The circuit which generates sawtooth voltage waveform,

There are seven basic sweep circuits by which sweep linearly can be achieved:

- Exponential charging;
  A capacitor is charged through a resistor to a voltage which is small in comparison with the supply voltage.
- Constant-current charging;
  A capacitor is charged with a constant current source.
- Miller circuit;
  Integrator is used to convert a step wave form into ramp waveform.
- Phantastron circuit;
  It is a modified Miller circuit which requires only pulse input to get the ramp wave form.
- Bootstrap circuit;
  A constant current is obtained by maintaining nearly constant voltage across a fixed resistor in series with a capacitor.
- Compensation networks;
  Compensation circuits are added to improve the linearity of bootstrap and Miller time-base generators.
- Inductor circuit;
  Linear capacitor charging is achieved by introducing RLC series circuit.

L – 37 Miller Integrator voltage sweep generator,

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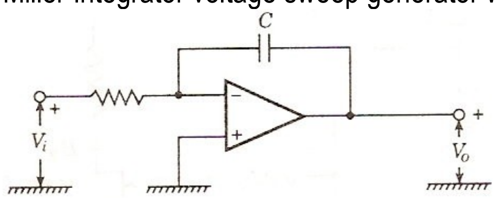
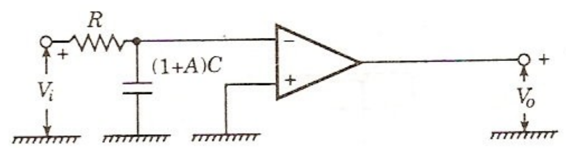
The integrator is used to convert a step waveform into ramp waveform. This generator is used to produce a precise and linear ramp voltage. By using an active device with appropriate feedback, the effective time constant and power supply voltage get enhanced. The feedback may be negative voltage feedback or negative current feedback.

Miller Integrator voltage sweep generator with negative voltage feedback using Op-amp;



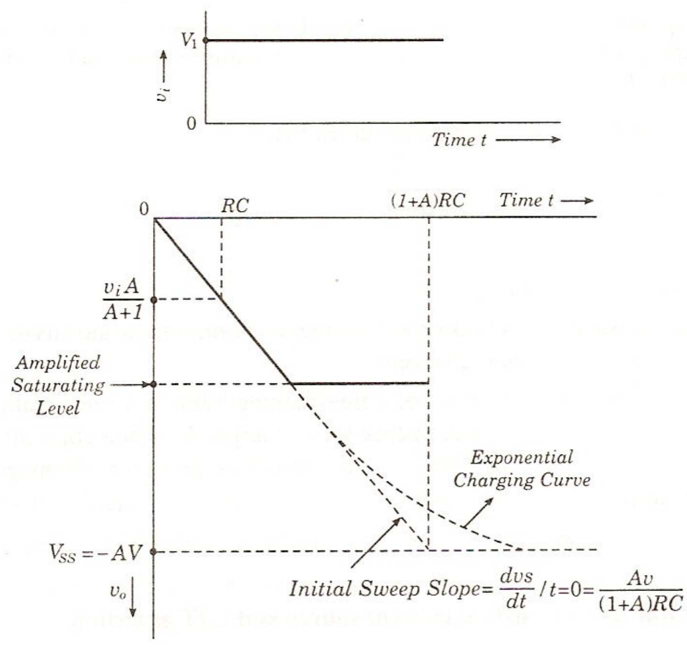
Miller Integrator voltage sweep generator using Op-amp



Equivalent circuit showing timing capacitor having been multiplied by (1+A)

The transfer function $\frac{V_o(s)}{V_i(s)} = -\frac{A}{1+(A+1)RCs} = -\frac{1}{\frac{1}{A}+\left(1+\frac{1}{A}\right)RCs}$

If $A \to \infty$ then $\frac{V_o(s)}{V_i(s)} = -\frac{1}{RCs}$, thus for large values of gain, (say 1000 or above) the circuit approaches an ideal integrator.



Step response of Miller sweep generator.

L – 38 Bootstrap saw tooth generators

Date: ……………..

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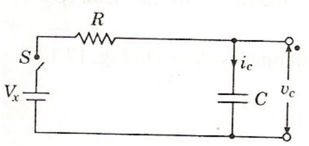
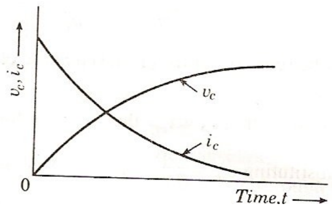
In Miller voltage sweep generator, a negative feedback is used for linearization purpose. In bootstrap method, linearization is achieved by using positive feedback.

Consider diagram (a); Let the switch S be closed at time $t = 0$, then $v_C\big|_{t=0} = 0$ and $i_C\big|_{t=0} = \frac{V_x}{R}$

For the time interval $(0 < t < \infty)$ we have $v_C = V_x\left[1 - e^{-1/RC}\right]$ and $i_c = \left[i_c\big|_{t=0}\right]e^{-1/RC}$

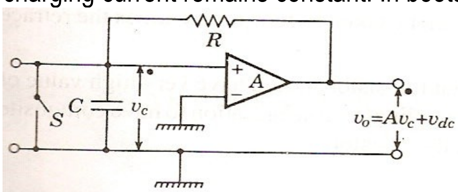




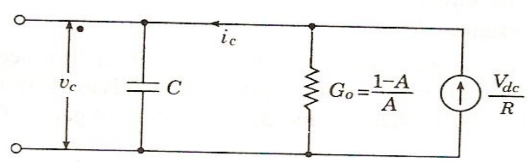
In diagram (b)

Plots of the waveforms of voltage v_C and current i_C shows that the current decays exponentially while the voltage rises exponentially towards a steady state value V_x

One of the methods of sweep linearization is to charge the capacitor C through a constant current source, or another method, as v_C rises, the supply voltage V_x can be made to rise suitably such that the charging current remains constant. In bootstrap method this is achieved by positive feedback.



(*a*) General bootstrap circuit employing positive feedback for sweep linearization.



(*b*) Equivalent circuit of general bootstrap sweep circuit

Consider diagram (a) which shows the circuit arrangement of bootstrap linear voltage sweep generator.

The feedback resistor R provides positive feedback. The capacitor C is a timing capacitor. The voltage across C at any time t after the switch is opened is denoted by v_C. This voltage is amplified A times by the amplifier and the output is given by;

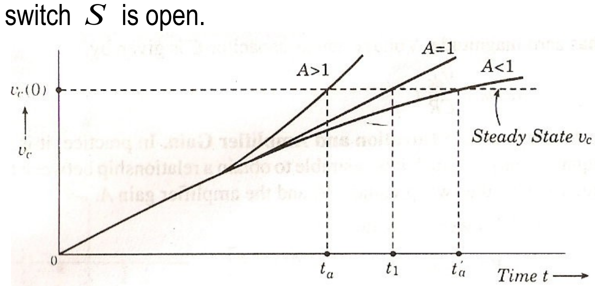
$v_o = Av_c + v_{dc}$ where A – amplifier gain; v_{dc} - the $d.c.$ level of the amplifier at the output terminals.

The dc level v_{dc} is a necessary requirement due to special nature of the circuit. In the absence of such a $d.c.$ level, capacitor C will not begin to charge when switch S is open.

The charging current

$$i_C = \frac{v_{dc} + Av_c - v_c}{R} = \frac{v_{dc}}{R} - \frac{(1-A)}{R}v_c \text{ ....(i)}$$

Diagram (b) gives the equivalent circuit diagram for equation (i) which shows that to keep the charging current i_C constant, the amplifier gain must be positive and have unity magnitude.



(*c*) Capacitor voltage for different values of amplifier gain.

L – 39 Current time base generators.

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A current time-base generator provides a current waveform which increases linearly with time. Such current time-base generators are used in television deflection systems, where magnetic deflection is employed for greater deflection sensitivity.

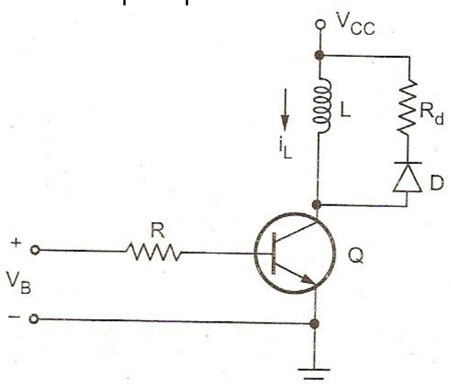
Definition;

Current time-base generators are circuits which produce current that linearly increase with time.

Inductors are used in practical current time-base generators. Such inductors are in forms of coil or sets of coil known as yoke.

Simple current time-base generator (simple current sweep circuit)

The basic principle is based on the characteristic behaviour of an inductor current.



Construction and principle of operation;

If at $t = 0$, voltage V is applied to a coil of inductance L , with initial zero current, then

$$V = L\frac{\partial i_L}{\partial t}; \ \ therefore \ \ \partial i_L = \frac{V}{L}\partial t \quad \Rightarrow i_L = \frac{V}{L}t$$

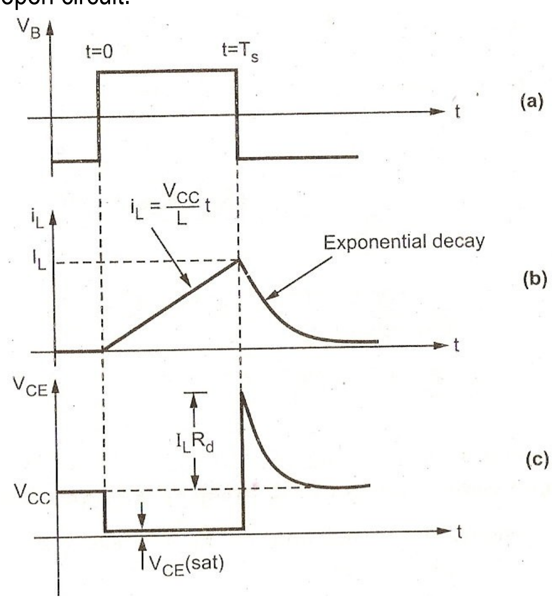
This shows that the inductor current varies linearly with time.

A diode D with R_d in series is connected across L .

Transistor Q is used as a switch.

A gating waveform V_B applied at the base of Transistor Q at $t = 0$ makes the base positive and hence Q gets ON , when gating waveform attains lower level, Q gets OFF ,

When Q is ON , i_L increases linearly with time from $t = 0$ till $t = T_s$ when the sweep ends as the gating waveform makes Q gets OFF . As i_L increases with time D is reverse biased and acts as an open circuit.



At $t = T_s$, $Q \ is \ cut-off$ and D becomes forward biased, then the inductor current i_L flow through D and R_d till it decays to zero.

$\tau = \frac{L}{R_d}$ - decay time constant.

The equation of the inductor current during the sweep; $i_L = \frac{V_{cc}}{L}t$

The equation of the exponential decay of inductor current; $i_L = I_L e^{-\frac{R_d}{L}(t-T_s)}$

At the instant when Q becomes OFF a spike of amplitude $I_L R_d$ appears across L which decays with the same time constant τ as that of i_L .

Effects of other resistances;

Under practical conditions the yoke has a finite internal resistance R_L and the collector saturation resistance R_{cs} cannot be neglected.

Let; R_L - Internal resistance of inductor yoke

R_{cs} - Collector saturation resistance

The inductor current equation due to these two resistors R_L and R_{cs}

$$i_L = \frac{V_{cc}}{R_L + R_{cs}}\left[1 - e^{-\left(\frac{R_L + R_{cs}}{L}\right)t}\right] \quad \text{..........(i)}$$

To show that the current departs from a linear increase in time i.e., produces a **slope error**, let us determine the slope difference ratio e_s .

$$e_s = \frac{\left.\frac{\partial i_L}{\partial t}\right|_{t=0} - \left.\frac{\partial i_L}{\partial t}\right|_{t=T_s}}{\left.\frac{\partial i_L}{\partial t}\right|_{t=0}}$$ - if the current rises to a maximum value I_L

Differentiating equation (i)

$$\frac{\partial i_L}{\partial t} = \frac{\partial}{\partial t}\left[\frac{V_{cc}}{R_L + R_{cs}}\left[1 - e^{-\left(\frac{R_L + R_{cs}}{L}\right)t}\right]\right] = \frac{\partial}{\partial t}\left[\left(\frac{V_{cc}}{R_L + R_{cs}}\right) - \left(\frac{V_{cc}}{R_L + R_{cs}}\right)e^{-\left(\frac{R_L + R_{cs}}{L}\right)t}\right]$$

Recall $\partial e^x = e^x \partial x \Rightarrow \frac{\partial i_L}{\partial t} = -\left(\frac{V_{cc}}{R_L + R_{cs}}\right)e^{-\left(\frac{R_L + R_{cs}}{L}\right)t}\left(-\frac{R_L + R_{cs}}{L}\right) = \frac{V_{cc}}{L}e^{-\left(\frac{R_L + R_{cs}}{L}\right)t}$

$$Therefore \quad \left.\frac{\partial i_L}{\partial t}\right|_{t=0} = \frac{V_{cc}}{L}; \quad \left.\frac{\partial i_L}{\partial t}\right|_{t=T_s} = \frac{V_{cc}}{L}e^{-\left(\frac{R_L + R_{cs}}{L}\right)T_s}$$

Hence $e_s = \dfrac{\frac{V_{cc}}{L} - \frac{V_{cc}}{L}e^{-\left(\frac{R_L + R_{cs}}{L}\right)T_s}}{\frac{V_{cc}}{L}} = 1 - e^{-\left(\frac{R_L + R_{cs}}{L}\right)T_s}$;

Let $e^{-\left(\frac{R_L + R_{cs}}{L}\right)T_s} = e^{-x}$

Expanding exponential term in terms of power series $e^{-x} = 1 - x + \frac{x^2}{2} - \frac{x^3}{3} + \ldots$

Neglecting the lower terms $e^{-x} \approx 1 - x$, therefore $e_s = 1 - \left(1 - \frac{R_L + R_s}{L}T_s\right) = \frac{R_L + R_s}{L}T_s$

Refer the graph $I_L = \frac{V_{cc}}{L}T_s \Rightarrow T_s = \frac{I_L \times L}{V_{cc}}$

Therefore $e_s = \frac{R_L + R_s}{L} \times \frac{I_L \times L}{V_{cc}} = I_L\left(\frac{R_L + R_s}{V_{cc}}\right)$

L – 40 Linearization using constant current circuit

Date: ……………..…

Hall: 21 Period: ……

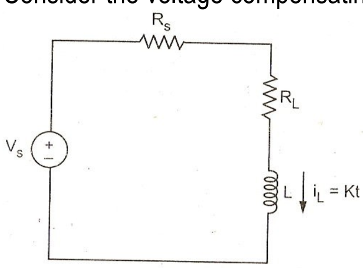
In the current time-base generator circuits as the yoke current increases, the drop across its internal resistance increases. This reduces the overall voltage across the inductor.

The voltage across the inductor is given by; $V = L\frac{\partial i_L}{\partial t}$, so as the voltage across the yoke decreases, the rate of change of current also decreases $V \downarrow \Rightarrow \frac{\partial i_L}{\partial t} \downarrow$.

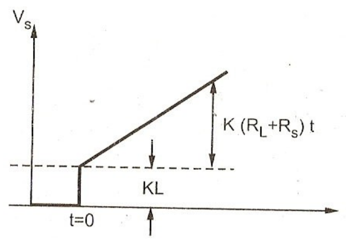
This produces a non-linearity in the linear wave form of current with time.

To compensate for decrease in voltage across the inductor, a linearly increasing voltage is applied to an inductor externally, and hence linear increase of current with time is maintained.

Consider the voltage compensating circuit;



(a)



The external source V_s *with* R_s in series, Nature of voltage V_s

Total circuit resistance $R_s + R_L$

It is required to obtain inductor current perfectly linearly varying with time. $i_L = kt$

From this requirement, the nature of applied voltage V_s can be obtained as;

$V_s = L\frac{\partial i_L}{\partial t} + i_L(R_s + R_L)$, substituting $i_L \Rightarrow V_s = L\frac{\partial}{\partial t}(kt) + kt(R_s + R_L)$

Hence $V_s = kL + kt(R_s + R_L)$ i.e. at $t = 0$; $V_s = kL$

The equation shows that at $t = 0$; $V_s = kL$ and then increases linearly with time. At the end of the sweep current reduces to zero exponentially with $\tau = \frac{L}{R_s + R_L}$, but usually $R_s >> R_L \Rightarrow \tau = \frac{L}{R_s}$

Note;

- When R_s is small $\Rightarrow$ τ is large;
  The current will decay very slowly and a long period will elapse before the next sweep. But due to slowly varying current, the peak voltage developed across the current source will be small.
- When R_s is large $\Rightarrow$ τ is small;
  The current reduces rapidly to zero after the end of the sweep, then a large peak voltage will appear across the source, which is undesirable.

A compromise is to be made while designing R_s

Revision Questions

PART - A

Que 1. Mention at least four applications of pulse transformer.
Que 2. Mention at least four applications of blocking oscillators.
Que 3. How the linearity of current sweep generators can be improved?
Que 4. What is the function of a pulse transformer in blocking oscillators?
Que 5. (i) Draw the equivalent circuit of a pulse transformer. (ii) Explain the various elements.
Que 6. What is the difference between High Pass RL and Low Pass RL circuits?

PART - B

Que 1. How can a Low Pass RC circuit be used as an integrator?

Que 2. How can a High Pass RC circuit be used as an integrator?

Que 3. Derive the expression for the *sweep-speed error* in current time base generators.

Que 4. Determine the upper $3-dB\ frequency$ for low pass RC circuit, if a pulse of $0.4\mu\sec$ is required to pass without distortion. Find the value of resistance if the capacitor is $0.001\mu F$.

Que 5. Explain the core saturation method of controlling the pulse width in Monostable blocking oscillators.

Que 6. Derive an expression for the pulse width of Emitter Timing Monostable Blocking oscillator.

PART - C

Que 1. Explain with a suitable circuit diagram, the performance of Monostable blocking oscillator (with base timing).

Que 2. Discus about linearization using constant current circuit.

Que 3. Draw the circuit diagram of a Monostable transistor blocking oscillator with emitter timing. Explain its operation with equivalent circuit during the pulse formation.

Que 4. Using a neat and well labeled circuit diagram, explain the triggering circuit for Monostable Blocking Oscillator using a **transistor**

Que 5. Using a neat and well labeled circuit diagram, explain the triggering circuit for Monostable Blocking Oscillator using a **diode**

Que 6. Derive and draw the response of low pass RC circuit to the following waveforms:
(i) Step (ii) square (iii) Ramp

APPENDIX

A. **Continuous Assessment Tests**

CAT – I

ST JOSEPH COLLEGE of ENGINEERING and TECHNOLOGY

IV SEM – BE DEGREE

54 EC 402 ELECTRONIC CIRCUITS II

Time: 1 Hr 30 mins. CAT – ONE Total Marks: 50

Part A: (2 x 7 = 14 marks)

Que 1. Define; (i) Voltage Amplifier, (1%) (ii) Transresistance Amplifier, (1%)

<u>ANS (i)</u> *A voltage amplifier is a circuit that provides a voltage output proportional to the voltage input and the proportionality factor is independent of the magnitude of the source and load resistance.* $Vo \approx A_v V_i \approx A_v V_s$

<u>ANS</u> *(ii) Transresistance amplifier is a circuit with an output voltage proportional to the input signal current and the proportionality factor is independent of the source and load resistance.*

Que 2. Classify amplifiers in accordance to how input parameters are related to the output parameters. (2%)

<u>ANS</u> (i) Voltage amplifier;

A voltage amplifier circuit provides a voltage output proportional to the voltage input and the proportionality factor is independent of the magnitude of the source and load resistance.

(ii) Current amplifier;

A current amplifier provides a current output proportional to the signal current and the proportionality factor is independent of the source and load resistance.

(iii)Transconductance amplifier;

An output current is proportional to the input signal voltage and the proportionality factor is independent of the magnitudes of the source and load resistance.

(iv) Transresistance amplifier;

The output voltage is proportional to the input signal current and the proportionality factor is independent of the source and load resistance.

Que 3. State Barkhausen criterion for sustained oscillations. (2%)

<u>ANS</u> *Barkhausen Criterion*

- *The total phase shift around a loop, as the signal proceeds from input through amplifier, feedback network back to amplifier again, completing a loop, is precisely* 0^0 *or* 360^0 *(or an integral multiple of* 2π *radians.*
- *The magnitude of the product of the open loop gain of the amplifier (A) and the feedback factor* β *is unity i.e.* $|A\beta| = 1$

OR

Oscillations will not be sustained if, at the oscillator frequency, the magnitude of the product of the transfer gain of the amplifier and the magnitude of the feedback factor of the feedback network are less than unity.

Que 4. List two Advantages of the Wien bridge oscillator (2%)

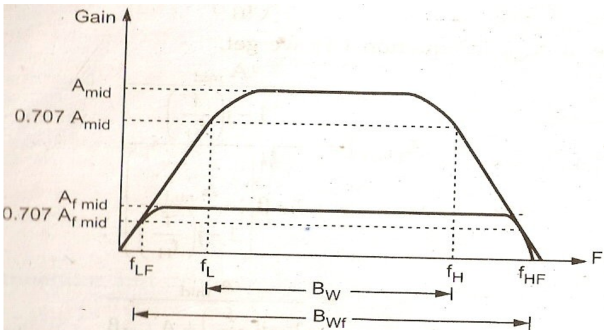
<u>ANS</u> *Advantages of the Wien bridge oscillator;*

- *By varying the two capacitor values can be varied simultaneously by mounting them on a common shaft to get different frequency ranges.*
- *No phase shift is required in the feedback network because the total phase shift around the loop is* 0^0*,*
- *The gain of the amplifier is low.*

Que 5. Draw the Characteristic of Gain Vs Frequency of an amplifier and indicate;

i. Lower and Upper cut-off frequencies without feedback (0.5%)
ii. Bandwidth of amplifier without feedback (0.5%)
iii. Lower and Upper cut-off frequencies with feedback (0.5%)
iv. Bandwidth of amplifier with feedback (0.5%)

<u>ANS</u>



Que 6. In not less than four points, compare/ contrast the RC phase shift from Wien Bridge Oscillator. (2%)

<u>*ANS*</u>

Comparison of RC phase shift and Wien Bridge Oscillator

RC phase shift Oscillator	Wien Bridge Oscillator
▪ Phase shift Oscillator used for low frequency range.	▪ Phase shift Oscillator used for low frequency range.
▪ Feedback network – Three RC section	▪ Feedback network – Lead-lag *Wien Bridge*
▪ Phase shift by feedback network – 180^0	▪ Phase shift by feedback network - 0^0
▪ Op Amp is used in an inverting mode	▪ Op Amp is used in a non-inverting mode
▪ Phase shift by Op Amp – 180^0	▪ Phase shift by Op Amp - 0^0
▪ Frequency of oscillation $f = \dfrac{1}{2\pi RC\sqrt{6}}$	▪ Frequency of oscillation $f = \dfrac{1}{2\pi RC}$
▪ Amplifier gain condition $\lvert A\rvert \geq 29$	▪ Amplifier gain condition $\lvert A\rvert \geq 3$
▪ Frequency variation is difficult	▪ Mounting the two capacitors on common shaft and varying their values, frequency can be varied

Que 7. (i) Is the Op Amp in a Wien Bridge Oscillator used in inverting or non-inverting mode? (1%)
(ii) Justify your answer. (1%)

<u>ANS(i)</u> Non-inverting mode

(ii) The feedback is given to the non-inverting terminal of the Op Amp to ensure zero phase shift.

<u>Part B:</u> (4 x 3 = 12 marks)

Que 8. (A) Show that introducing negative feedback the low frequency response of the amplifier is improved. i.e. the lower cut-off frequency with feedback is less than the lower cut-off frequency without feedback. (4%)

<u>*ANS*</u>

▸ <u>*Lower cut-off frequency*</u>

The relation between gain at low frequency and gain at mid frequency is given as;

$$\frac{A_{low}}{A_{mid}} = \frac{1}{1 - j\left(\dfrac{f_L}{f}\right)} \quad \therefore \quad A_{low} = \frac{A_{mid}}{1 - j\left(\dfrac{f_L}{f}\right)}$$

Substituting the value A_{Low} we get

$$A_{flow} = \frac{\dfrac{A_{mid}}{1-j\left(\dfrac{f_L}{f}\right)}}{1+\beta\left(\dfrac{A_{mid}}{1-j\left(\dfrac{f_L}{f}\right)}\right)} = \frac{A_{mid}}{1-j\left(\dfrac{f_L}{f}\right)+A_{mid}\beta} = \frac{A_{mid}}{(1+A_{mid}\beta)-j\left(\dfrac{f_L}{f}\right)}$$

Dividing numerator and denominator by $(1+A_{mid}\beta)$

$$A_{flow} = \frac{\dfrac{A_{mid}}{(1+A_{mid}\beta)}}{1-j\left(\dfrac{\dfrac{f_L}{(1+A_{mid}\beta)}}{f}\right)} = \frac{A_{f\,mid}}{1-j\left(\dfrac{\dfrac{f_L}{(1+A_{mid}\beta)}}{f}\right)} \quad \because A_{f\,mid} = \frac{A_{mid}}{(1+A_{mid}\beta)}$$

$\therefore \dfrac{A_{f\,Low}}{A_{f\,mid}} = \dfrac{1}{1-j\left(\dfrac{f_{Lf}}{f}\right)}$ *Where the lower cut-off frequency with feedback* $f_{Lf} = \dfrac{f_L}{(1+A_{mid}\beta)}$;

Hence, the lower cut-off frequency with feedback is less than the lower cut-off frequency without feedback by factor $1+A_{mid}\beta$. Therefore, by introducing negative feedback the low frequency response of the amplifier is improved.

<u>OR</u>

(B) Show that introducing negative feedback the high frequency response of the amplifier is improved. i.e. the higher cut-off frequency with feedback is greater than the higher cut-off frequency without feedback. (4%)

<u>*ANS*</u>

- *Upper cut-off frequency*

The relation between gain at high frequency and gain at mid frequency is given as;

$$\frac{A_{high}}{A_{mid}} = \frac{1}{1-j\left(\dfrac{f}{f_H}\right)} \quad \therefore \quad A_{high} = \frac{A_{mid}}{1-j\left(\dfrac{f}{f_H}\right)}$$

Substituting the value A_{high} we get

$$A_{fhigh} = \frac{\dfrac{A_{mid}}{1-j\left(\dfrac{f}{f_H}\right)}}{1+\beta\left(\dfrac{A_{mid}}{1-j\left(\dfrac{f}{f_H}\right)}\right)} = \frac{A_{mid}}{1-j\left(\dfrac{f}{f_H}\right)+A_{mid}\beta} = \frac{A_{mid}}{(1+A_{mid}\beta)-j\left(\dfrac{f}{f_H}\right)}$$

Dividing numerator and denominator by $(1+A_{mid}\beta)$

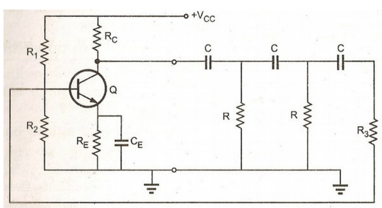
$$A_{fhigh} = \frac{\frac{A_{mid}}{(1+A_{mid}\beta)}}{1-j\left(\frac{f}{(1+A_{mid}\beta)f_H}\right)} = \frac{A_{f\,mid}}{1-j\left(\frac{f}{(1+A_{mid}\beta)f_H}\right)} \quad \because A_{f\,mid} = \frac{A_{mid}}{(1+A_{mid}\beta)}$$

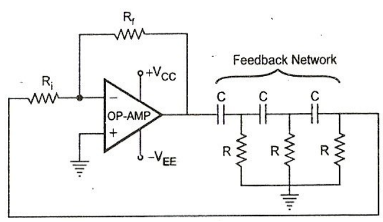
$$\therefore \frac{A_{f\,high}}{A_{f\,mid}} = \frac{1}{1-j\left(\frac{f}{f_{Hf}}\right)}$$ *Where* $f_{Hf} = (1+A_{mid}\beta)f_H$; *higher cut-off frequency with feedback.*

Hence, the higher cut-off frequency with feedback is greater than the higher cut-off frequency without feedback by factor $1+A_{mid}\beta$ *. Therefore, by introducing negative feedback the high frequency response of the amplifier is improved.*

Que 9. (A)(i) Draw a diagram of an RC oscillator using cascade connections of low pass or high pass filters. (2%)
(ii) Write the equation for the frequency of oscillation of the oscillator in sec. (i) (2%)

<u>ANS</u>





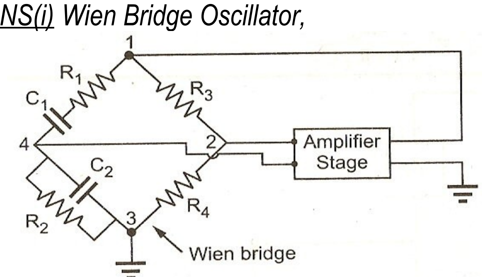
The frequency of transistorized RC phase shift oscillator $f = \frac{1}{2\pi RC\sqrt{4k+6}}$ where $k = \frac{R_C}{R}$

The frequency of the RC phase shift oscillator using an op-amp $f = \frac{1}{2\pi\sqrt{6}RC}$

<u>OR</u>

(B) (i) Draw the diagram of a Wien Bridge oscillator. (2%)
(ii) Write the equation for the frequency of oscillation of the oscillator in sec. (i) (2%)

<u>ANS(i)</u> Wien Bridge Oscillator,



<u>ANS(ii)</u> $f = \frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}$

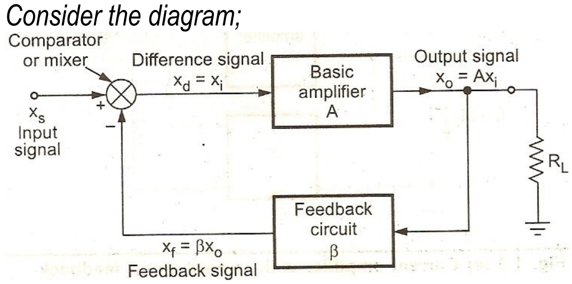
if $R_2 = R_1 = R$ *and* $C_2 = C_1 = C$

then $f = \frac{1}{2\pi RC}$

Que 10. (A) Prove that gain without feedback is always greater than gain with feedback and it decreases with increase in feedback factor β . (4%)

<u>ANS</u>

Consider the diagram;



Let A *represent transfer gain of amplifier without feedback where* $A = \frac{X_o}{X_i}$

A_f *represent transfer gain of amplifier with feedback* $A_f = \frac{X_o}{X_s}$

Where X_o $- Output\ voltage\ or\ current;$

X_i $- input\ voltage\ or\ input\ current;$

X_s $- source\ voltage\ or\ source\ current;$

As it is negative feedback $X_i = X_s + \left(-X_f\right)$*; therefore* $A_f = \frac{X_o}{X_s} = \frac{X_o}{X_i + X_f}$

Where X_f *-feedback voltage or feedback current*

Dividing by X_i *to numerator and denominator; introducing* $A = \frac{X_o}{X_i}$ *and* $\beta = \frac{X_f}{X_o}$ *- feedback factor*

$$\Rightarrow A_f = \frac{\frac{X_o}{X_i}}{\frac{X_i + X_f}{X_i}} = \frac{A}{1 + \frac{X_f}{X_i}} = \frac{A}{1 + \left(\frac{X_f}{X_o}\right)\left(\frac{X_o}{X_i}\right)} = \frac{A}{1 + \beta A}$$

Hence $A_f = \frac{A}{(1 + \beta A)}; and\ A = A_f(1 + \beta A)$

Gain without feedback is always greater than gain with feedback and it decreases with increase in feedback factor.

OR

(B) Prove that stability of gain with feedback is less than stability of gain without feedback by $1 + \beta A$ (4%)

ANS

The introduction of negative feedback in amplifiers reduces the lack of stability.

Consider $A_f = \frac{A}{(1 + \beta A)}$

Differentiate both sides with respect to A $\frac{\partial A_f}{\partial A} = \frac{(1 + \beta A)1 - \beta A}{(1 + \beta A)^2} = \frac{1}{(1 + \beta A)^2}$ $\therefore \partial A_f = \frac{\partial A}{(1 + \beta A)^2}$

Divide both sides by A_f $\Rightarrow \frac{\partial A_f}{A_f} = \frac{\partial A}{A_f(1 + \beta A)^2}$ $\quad Since\ A_f = \frac{A}{(1 + \beta A)}$

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

Where $\frac{\partial A_f}{A_f} - fractional\ change\ in\ amplification\ with\ feedback;$ *- sensitivity of transfer gain*

$\frac{\partial A}{A} - fractional\ change\ in\ amplification\ without\ feedback$

Hence, the change in feedback with gain is less than the change in gain without feedback by a factor $1 + \beta A$

Part C: **(2 x 12 = 24 marks)**

Que 11. (A) (i) With justifications, identify the type of feedback topology used in Fig. 1(6%)

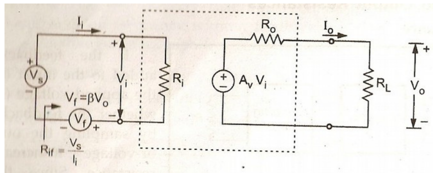


Fig. 1

(ii) Show that the input resistance with feedback $R_{if} = R_i(1+\beta A_v)$ (6%)

<u>*ANS*</u>

<u>OR</u>

(B) (i) With justifications, identify the type of feedback topology used in Fig.2 . (6%)

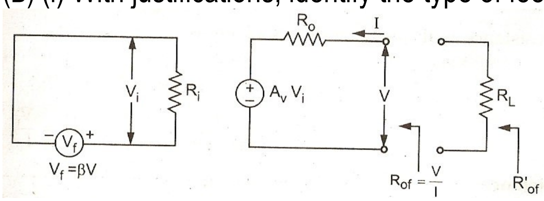


Fig. 2

(ii) Show that the output resistance with feedback $R_{of} = \frac{V}{I} = \frac{R_o}{(1+\beta A_v)}$ (6%)

<u>*ANS*</u>

Que 12. (A) (i) Determine whether the circuit shown will work as an oscillator or not; (6%)
(ii) If yes, determine the frequency of the oscillator. (6%)

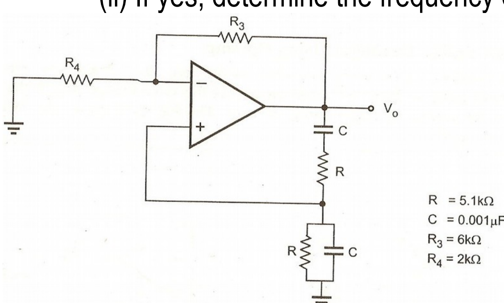


Fig 3

<u>*ANS(i)*</u>

The circuit is Wien Bridge oscillator using Op Amp. The gain of the loop $A = 1 + \frac{R_3}{R_4} = 1 + \frac{6}{2} = 4$

- *Since* $A > 3$ *, the conditions necessary for oscillation are satisfied;*
- *Since the feedback is given to the non-inverting terminal the zero phase shift is ensured;*
- *The circuit will work as an oscillator.*

<u>*ANS(ii)*</u>

The frequency of oscillations $f = \frac{1}{2\pi RC} = \frac{1}{2\pi 5.1\times 10^3 \times 0.001\times 10^{-6}} = 31.2068\text{kHz}$

CAT – II

St JOSEPH COLLEGE OF ENGINEERING AND TECNOLOGY

DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

54 EC 402 - ELECTRONIC CIRCUITS II

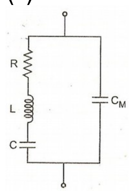
CAT – II **Marking Scheme** **Jun 2009**

TIME – 90 Min

PART A (7 x 2 = 14 Marks)

Que 1. (i) Define the frequency stability of an oscillator and (0.5%)

(ii) Given an equivalent circuit of a crystal,



prove that $f_r = \dfrac{1}{2\pi\sqrt{LC_{eq}}}$; where $C_{eq} = \dfrac{C_M C}{C_M + C}$ (1%)

(iii) Mention, with brief explanations, at least tree factors that affect the frequency stability of an oscillator. (0.5%)

ANS

(i) Frequency stability of an oscillator is the measure of the ability of an oscillator to maintain the desired frequency as precisely as possible for as long a time as possible

(ii) $X_M = X_L - X_C \Rightarrow \dfrac{1}{2\pi f_r C_M} = 2\pi f_r L - \dfrac{1}{2\pi f_r C} \Rightarrow 2\pi f_r L = \dfrac{1}{2\pi f_r}\left[\dfrac{1}{C_M} + \dfrac{1}{C}\right]$

Hence: $f_r = \dfrac{1}{2\pi\sqrt{LC_{eq}}}$ *where* $C_{eq} = \dfrac{C_M C}{C_M + C}$

(iii) Factors affecting frequency stability of the crystal are:

a. Temperature stability (change in frequency per degree change in temperature). When greater stability is required a crystal is kept in an oven (constant temperature box)

b. Aging of the crystal material – long term stability. (Aging rate $= 2 \times 10^{-8}$ *per year - negligible)*

c. Frequency drift with time – short term stability $\approx 0.0001\%$ *per day.*

Que 2. (i) What is the meaning of piezoelectric effect? (0.5%)

(ii) List down any two (2) crystals that exhibit piezoelectric effect (1%)

(iii) Identify the element in a Miller Crystal Oscillator that decides its operating frequency (0.5%)

ANS

(i) Meaning of Piezoelectric Effect

Piezoelectric Effect is a phenomenon by which under the influence of mechanical pressure, a voltage gets generated across the opposite faces of the crystal and vice versa. - i.e.

- *If a mechanical force is applied so as to force the crystal to vibrate, an* $a.c.$ *voltage gets generated across it, or conversely,*
- *If a crystal is subjected to* $a.c.$ *voltage it vibrates.*

(ii) Types of crystals exhibiting piezoelectric effect

a. Rochelle Salts *b. Quartz* *c. Tourmaline*

(iii) The crystal decides the operating frequency of the oscillator

Que 3. (i) What is a tuned amplifier? (1mark)

(ii) Write down the equation for the resonating frequency of a tuned parallel LC circuit (1mark)

ANS

(i) *A tuned amplifier is an amplifier in which the resistive load R_C is replaced by a tuned circuit, and it amplifies only signals centred at the resonating frequency f_r .*

(ii) $f_r = \dfrac{1}{2\pi\sqrt{L_p C_{eq}}}$ *Where* $L_p = \dfrac{R^2 + \omega^2 L^2}{\omega^2 L}$ *and* $C_{eq} = C_o + C$ *being the summation of the transistor output capacitance* C_o *and the tuned circuit capacitance* C .

Que 4. With brief clarifications, define the Q-factor. (2 marks)

ANS

Quality factor Q

- *Quality factor* Q *is a measure of how "pure" or "real" an inductor is. The higher the* Q *of an inductor, the fewer losses there are in the inductor. Therefore it's a ratio* $Q = \dfrac{\text{Reactance}}{\text{Resistance}}$
- *Quality factor* Q *is a measure of efficiency with which the inductor can store energy.*

Que 5. What is the effect of cascading single tuned amplifiers on the bandwidth? (2 marks)

ANS

The bandwidth decreases with increasing number of stages.

If $n - stages$ *of single tuned direct coupled amplifiers are connected in cascade, then the bandwidth of* $n - stages$ *of identical amplifiers;* $BW_n = BW_1\sqrt{2^{1/n} - 1}$; *where* BW_1 *- bandwidth of a single stage.*

Que 6. A tuned circuit consists of a coil which is not ideally inductive hence it has losses. With brief explanations, mention any two (2) losses which are usually represented in form of leakage resistance in series with the inductor. (2 marks)

ANS

Inductor with leakage resistance

The losses of the coil represented in form of leakage resistance in series with the inductor are;

- *Copper loss: - This is equivalent to DC resistance of the coil at low frequency. Copper loss is inversely proportional to frequency; as frequency increases the copper loss decreases.*
- *Eddy current loss in iron and copper: - This is due to currents flowing within the copper or core cased by induction. The result of eddy current is a loss due to heating within the inductor s' copper or core. Eddy current losses are directly proportional to frequency.*
- *Hysteresis loss: - Hysteresis loss is proportional to the area enclosed by the hysteresis loop and the rate at which the loop is transversed. Hysteresis loss is independent of frequency.*

Que 7. List any two (2) advantages and two (2) disadvantages of tuned amplifiers. (2 marks)

ANS

Advantages of tuned amplifiers

- *They amplify defined frequencies;*
- *Signal to noise ration at output is good;*
- *They are well suited for radio transmitters and receivers*
- *The band of frequency over which amplification is required can be varied;*

Disadvantages;

- *Since they inductors and capacitors as tuning elements, the circuit is bulky and costly;*
- *If the band of frequency is increased, design becomes complex;*
- *They are suitable to amplify audio frequencies.*

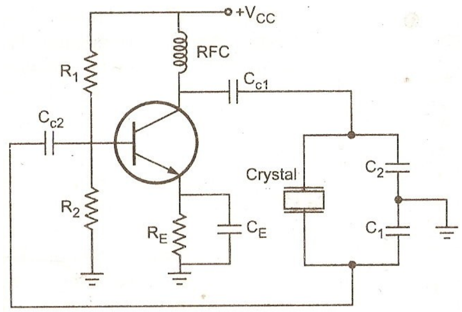
PART B (3 x 4 = 12 Marks)

Que 8. **(a)** (i) Draw a labeled circuit diagram of a PIERCE CRYSTAL OSCILLATOR. (2 marks)

(ii) Calculate series resonance frequency f_S - (1 mark)and parallel resonance frequency f_P - (1mark) of a crystal oscillator with $L = 3H,\ C = 0.01pF,$ and its mounting capacitance $C_M = 3pF$;

ANS

(i) *Pierce Crystal Oscillator is a modification of Colpitts Oscillator circuit whereby a crystal has replaced the inductor in the tank circuit.*



(ii)

$$f_S = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{3\times 0.01\times 10^{-12}}} = \frac{1}{2\pi\sqrt{3\times 10^{-14}}} = \frac{1}{2\times 3.14\times 1.732\times 10^{-7}} = \frac{10^7}{18.8394} = 0.531\text{MHz}$$

$$f_p = \frac{1}{2\pi\sqrt{LC_{eq}}} = \left|C_{eq} = \frac{(3\times 0.01)\times 10^{-24}}{(3+0.01)\times 10^{-12}} = 0.0903\times 10^{-12}\right| = \frac{1}{2\pi\sqrt{3\times 9.03\times 10^{-14}}} = \frac{10^7}{27.09} = 0.369MHz$$

OR

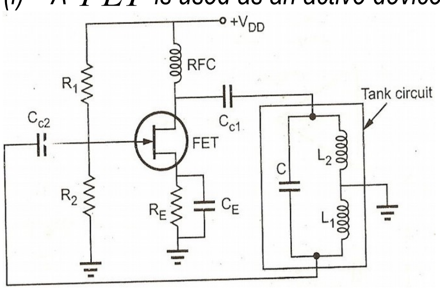
(b) (i) Draw a labeled circuit diagram of a HARTLEY OSCILLATOR using FET. (2 marks)

(ii) In a Hartley Oscillator, $L_1 = 15mH;\ C = 50pF;$ Calculate

- L_2 for a frequency of $168kHz$. The mutual inductance between L_1 and L_2 is $5\mu H$ (1 mark)
- The required gain of the transistor to be used for the oscillations, (1 mark)

ANS

(i) A FET is used as an active device in an amplifier stage.



(ii)

$$f = \frac{1}{2\pi\sqrt{CL_{eq}}};\ where\ L_{eq} = L_1 + L_2 + 2M \Rightarrow 168\times 10^3 = \frac{1}{2\times 3.14\sqrt{50\times 10^{-12}L_{eq}}} \Rightarrow L_{eq} = 17.95\text{mH}$$

$$h_{fe} = \frac{L_1 + M}{L_2 + M} = \frac{15\times 10^{-3} + 5\times 10^{-6}}{2.945\times 10^{-3} + 5\times 10^{-6}} = 5.08$$

Que 9. (a) (i) Explain the cause of instability of tuned RF amplifiers using the diagram in Fig.2. (2%)

(ii) Explain using the diagram of a tuned RF stage (Fig.2) how Hazeltine neutralization can be executed

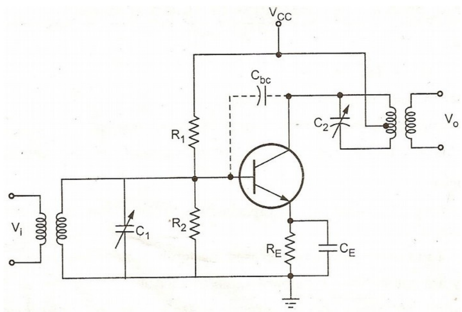


Fig 2. (2%)

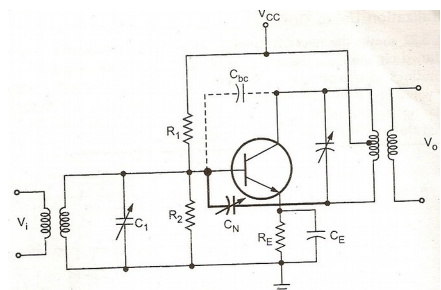
ANS

(i) The cause of instability of tuned RF amplifiers;

The capacitance C_{bc} come across input and output circuits of an amplifier. As the reactance of C_{bc} at RF is low enough it provides the feedback path from collector to base. If some feedback signal manages to reach the input from the output in a positive manner with proper phase shift, then, there is a possibility of the circuit to be converted to an unstable one, generating its own oscillations hence stopping working as an amplifier. Ref. Fig. 2 above.

(ii) Hazeltine neutralization

Hazeltine neutralization is achieved by feeding back a portion of the output signal to the input in such a way that it has the same amplitude as the unwanted feedback but the opposite phase. This is done by connecting a small value of variable capacitance C_N from the bottom of the coil, point B, to the base. The neutralization capacitor can be adjusted correctly to completely nullify the signal fed through the C_{bc}



OR

(b) Differentiate Q_U from Q_L and show how they affect the practical efficiency of a circuit (4 marks)

ANS

- *Unloaded Q (Q_u) : - The ratio of stored energy to dissipated energy in a resonator;*
- *Loaded Q (Q_L) of the resonator is determined by how tight the resonator is coupled to its terminations.*

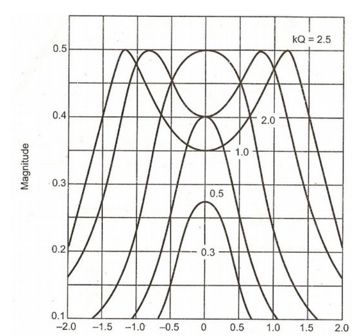
Que 10. **(a)** Draw universal response curves on the same graph for the double tuned amplifier with;

(i) Insufficient coupling i.e. $k < \frac{1}{Q}$ (1 mark)

(ii) Over-coupled i.e. $k > \frac{1}{Q}$ (1 mark)

(iii) Critically coupled i.e. $k = k_c = \frac{1}{Q}$ (2 marks)

ANS



OR

(b) (i) Mention 2 advantages of the double tuned compared to a single tuned amplifier. (2 marks)

(ii) What are the effects of cascading double tuned amplifiers on bandwidth? (1 mark)

(iii) The bandwidth for double tuned amplifier is 20 kHz. Calculate the bandwidth if such three stages are cascaded (1 mark)

ANS

(i) Advantages of tuned amplifiers

- *They amplify defined frequencies;*
- *Signal to noise ration at output is good;*
- *They are well suited for radio transmitters and receivers*
- *The band of frequency over which amplification is required can be varied;*

o *Effects of cascading double tuned amplifiers on bandwidth*

- *The overall bandwidth of the system is narrowed,*
- *The steepness of the sides of the response is increased.*

o $BW_n = BW_1 \times \left(2^{1/n} - 1\right)^{1/4} = 20 \times 10^3 \left(2^{1/3} - 1\right)^{1/4} = 14.28 kHZ$

PART C (2 x 12 = 24 Marks)

Que 11. (a) Applying the basic forms of LC amplifier circuits;

(i) Analyse the amplifier stage (3 marks)

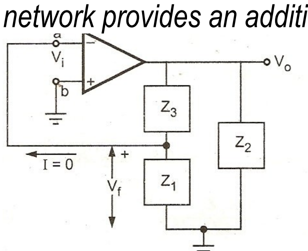
(ii) Analyse of the feedback stage; (3 marks)

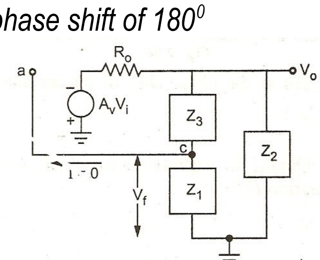
(iii) According to Barkhausen Criterion show that $-A\beta$ will be positive only if X_1 and X_2 will have the same sign. (3 marks)

(iv) Draw any scientific conclusion with reference to X_1 and X_2. (3 marks)

ANS

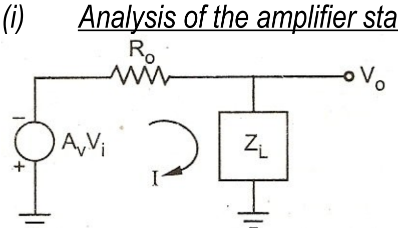
The circuit consists of an amplifier stage of gain A_V with its output to a feedback stage of impedances Z_1, Z_2 and Z_3. The amplifier provides a phase shift of 180^0 while the feedback network provides an additional phase shift of 180^0





[redacted]*Basic Circuit* *Equivalent Circui*

(i) <u>*Analysis of the amplifier stage;*</u>



The input impedance of the amplifier is infinite; hence there is no current flowing towards the input terminal i.e. $I = 0$. Let R_0 be the output impedance of the amplifier stage.

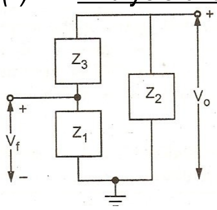
As $I = 0$, Z_1, Z_2 appear in series, Z_1, Z_2 and Z_3 form an equivalent load impedance Z_L.

Where $Z_L = \frac{Z_2(Z_1+Z_3)}{Z_1+Z_2+Z_3}$

$$Therefore\ I = \frac{-A_V V_i}{R_0 + Z_L}\ Since\ V_0 = IZ_L \Rightarrow V_0 = \frac{-A_V V_i Z_L}{R_0 + Z_L};\ Hence\ \frac{V_0}{V_i} = A = \frac{-A_V Z_L}{R_0 + Z_L}$$

Where A, is the gain of the amplifier stage.

(ii) *Analysis of the feedback stage;*



$$V_f = V_0\left[\frac{Z_1}{Z_1 + Z_3}\right], Therefore \frac{V_f}{V_0} = \beta = \left[\frac{Z_1}{Z_1 + Z_3}\right],$$

Phase shift of feedback network is 180^0, *then* $\beta = -\left[\frac{Z_1}{Z_1 + Z_3}\right]$.

To satisfy Barkhausen Criterion $-A\beta = 1;\ Therefore\ -A\beta = \frac{-A_V Z_1 Z_L}{(R_0 + Z_L)(Z_1 + Z_3)}$

Hence $$-A\beta = \frac{-A_V Z_1\left[\frac{Z_2(Z_1 + Z_3)}{Z_1 + Z_2 + Z_3}\right]}{\left(R_0 + \left[\frac{Z_2(Z_1 + Z_3)}{Z_1 + Z_2 + Z_3}\right]\right)(Z_1 + Z_3)}$$

Dividing numerator and denominator by $\frac{Z_2(Z_1 + Z_3)}{Z_1 + Z_2 + Z_3}$ *and substituting*

$$Z_1 = jX_1, Z_2 = jX_2 = and\ Z_3 = jX_3\ Where\ X = \omega L - \text{inductance} X = \frac{-1}{\omega C} - \text{capacitance}$$

$$-A\beta = \frac{-A_V X_1 X_2}{-X_2(X_1 + X_3) + jR_0(X_1 + X_2 + X_3)}$$

(iii) According to Barkhausen Criterion show that $-A\beta$ *will be positive only if* $X_1\ and\ X_2$ *will have the same sign.*

To have a phase shift of 180^0, *the imaginary part of the denominator must be equal to zero i.e.* $X_1 + X_2 + X_3 = 0 \Rightarrow X_1 + X_3 = -X_2$

Hence $-A\beta = \frac{-A_V X_1 X_2}{-X_2(X_1 + X_3)} or -A\beta = \frac{-A_V X_1}{-X_2} = A_V \frac{X_1}{X_2}$

According to Barkhausen Criterion $-A\beta$ *must be positive and greater than unity. As* A_V *is positive* $-A\beta$ *will be positive only if* $X_1\ and\ X_2$ *will have the same sign.*

(iv) Scientific conclusion with reference to $X_1\ and\ X_2$.

- *This indicates that* $X_1\ and\ X_2$ *must be both inductive or both capacitive.*

OR

(b) Explain the Twin-T oscillator highlighting the following;

(i) Construction of the Twin-T filter, its resonating frequency and frequency responses; (3 marks)

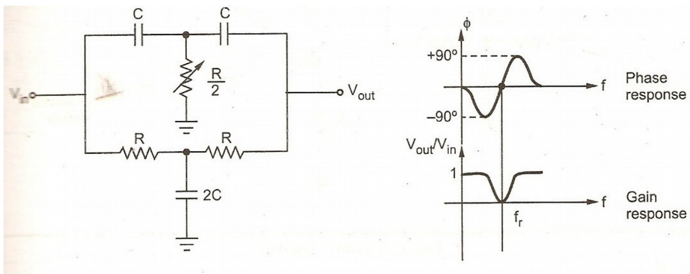
(ii) How a Twin-T circuit is used to obtain a Twin-T oscillator; (3 marks)

(iii) How do oscillations grow and become sustained oscillations (3 marks)

(iv) Any two disadvantages of the oscillator, (3 marks)

ANS

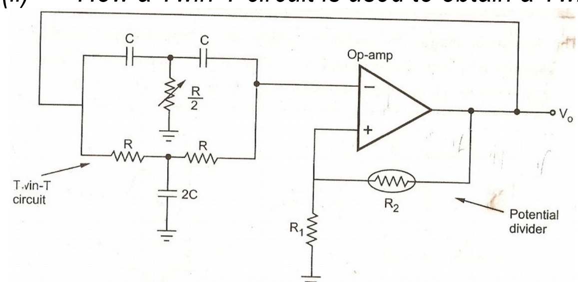
(i) *Construction of the Twin-T filter, its resonating frequency and frequency responses*



(a) Twin-T circuit (b) Frequency response

A Twin-T filter is a combination of a High Pass filter and Low Pass filter that is why a Twin-T filter acts as a notch filter.

(ii) How a Twin-T circuit is used to obtain a Twin-T oscillator



The positive feedback to the non-inverting input is given through the potential divider (R_1 and R_2) whereby R_2 is a lamp.

(iii) How do oscillations grow and become sustained oscillations

The negative feedback to the inverting input is given through twin-T filter. When the power is given to the circuit, lamp resistance R_2 is low thus positive feedback is maximum which helps to build oscillations. As the oscillations grows, the lamp resistance increases, decreasing the positive feedback. This controls the growing oscillations and makes them to sustain. The lamp acts as a stabilizer of output voltage level.

(iv) Two disadvantages of the oscillator

- *It operates only at one frequency f_r*
- *R_2 of potential divider must be larger enough $(10R_1 \text{ to } 100R_1)$ to obtain f_r*

Que 12. **(a)** For the circuit in Fig. 2, Assuming $Q_L = 100$ calculate;

(i) Resonant frequency, (3 marks)
(ii) AC collector resistance, (3 marks)
(iii) Quality factor (3 marks)
(iv) Bandwidth (3 marks)

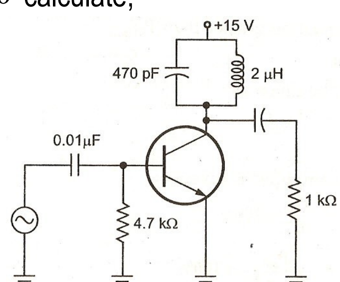


Fig. 2

ANS

(i) Resonant frequency, $f_r = \dfrac{1}{2\pi\sqrt{LC}} = \dfrac{1}{2\pi\sqrt{2\mu H \times 470pF}} = 5.19MHz$

(ii) Collector resistance

$$r_c = R_p \| R_L = Q_L 2\pi f_r L \| R_L = \left(100 \times 2\pi \times 5.19 \times 10^6 \times 2\mu H\right) \| 1k\Omega = 867\Omega$$

(iii) Quality factor $Q = \dfrac{r_c}{2\pi f_r L} = \dfrac{867}{2\pi \times 5.19 \times 10^6 \times 2 \times 10^{-6}} = 13.29$

(iv) Bandwidth $BW = \dfrac{f_r}{Q} = \dfrac{5.19 \times 10^6}{13.29} = 390.5kHz$

OR

(B) The fixed bias bistable multivibrator (Fig 3) uses the following parameters; $V_{CC} = +12\,V,\ V_{BB} = -8\,V,\ R_1 = 10\,k\Omega,\ R_2 = 50\,k\Omega,\ R_C = 2.2\,k\Omega$ the transistors are silicon transistors with a minimum value of $h_{fe}\ as\ 30$.

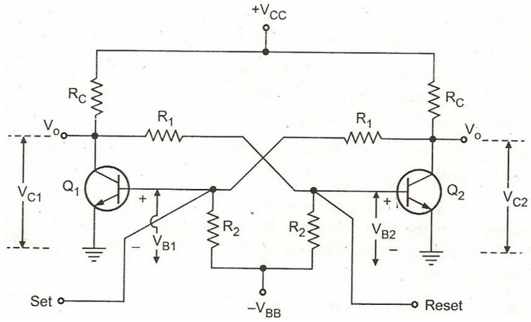


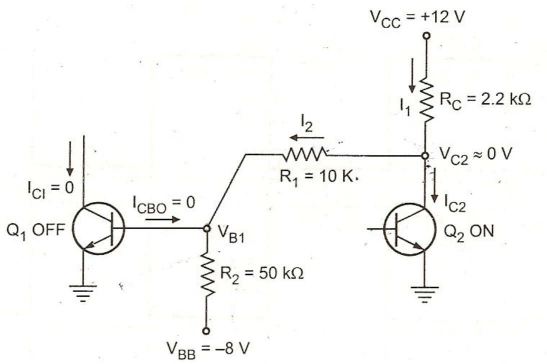
Fig 3

Neglecting all the junction voltages, Calculate;
(i) Stable state currents I_{C1} ; I_{C2}, I_{B1}, I_{B2} (6 marks)
(ii) Stable state voltages; $V_{C1}, V_{C2}, V_{B1}, V_{B2}$ (6 marks)

ANS

When all junction voltages of ON transistor are neglected, $V_{CE2} = 0V$ *and* $V_{BE2} = 0V$ *;*

As the emitter is grounded $V_{C2} = 0V$ *and* $V_{B2} = 0V$ *hence the equivalent circuit in part from base of* Q_1 *to the collector of* Q_2 *is as below;*



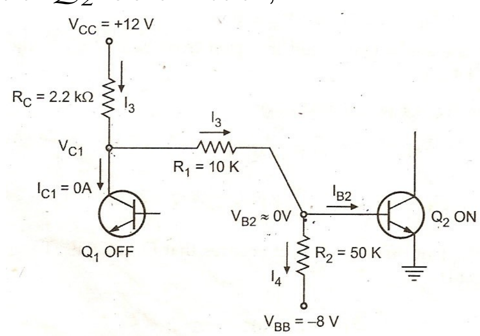
$$V_{B1} = -V_{BB}\left(\frac{R_1}{R_1 + R_2}\right) = -8\left(\frac{8}{8+50}\right) = -1.33V$$

$$I_1 = \frac{V_{CC}}{R_C} = \frac{12}{2.2\times 10^3} = 5.45mA$$

$$I_2 = \frac{V_{BB}}{R_1 + R_2} = \frac{8}{10+50} = 0.133mA$$

$$I_{C2} = I_1 - I_2 = 5.45mA - 0.133mA = 5.316mA$$

The equivalent circuit showing the collector of Q_1 *to base of* Q_2 *is shown below;*



$$I_3 = \frac{V_{CC}}{R_C + R_1} = \frac{12}{(2.2+10)\times 10^3} = 0.9836mA$$

The current through R_C *and* $R_1\ as\ I_{C1} = 0$

$$I_4 = \frac{V_{B2} - V_{BB}}{R_2} = \frac{0-(-8)}{50} = 0.16mA$$

$$I_{B2} = I_3 - I_4 = 0.8236mA$$

$I_{B2} > (I_{B2})_{\min}$ *- Transistors* Q_2 *is saturated*

$$V_{C1} = V_{CC} - I_3 R_C = 12 - 0.98396\times 2.2 = 9.$$

Hence the stable state currents are; $I_{C1} = 0A;\quad I_{C2} = 5.316mA;\quad I_{B1} = 0A;\quad I_{B2} = 0.8236mA$

And stable state voltages are; $V_{C1} = 9.836V;\ V_{C2} = 0;\ V_{B1} = -1.33V;$ *and* $V_{B2} = 0$

CAT – III

St JOSEPH COLLEGE OF ENGINEERING AND TECNOLOGY

DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

4th SEM B.E. DEGREE

54 EC 402 - ELECTRONIC CIRCUITS II

CAT – III **(Marking Scheme)** **Nov 2008**

TIME – 90 Min

PART A (7 x 2 = 14 Marks)

Que 1. (I) Explain why commutating capacitors are used in multivibrators and (1 mark)

- ✓ *The transition from one state to another should occur instantaneously when abrupt changing pulse is applied to the circuit, which means the transition time should be as small as possible. Commutating / speedup capacitors allow fast rise and fast fall thus avoiding distortion in output waveforms.*

(ii) Define rise time of a pulse (1mark)

- ✓ *Rise time a pulse is the time required by the output response to rise from 10% to 90% of its final steady state value.*

Que 2. (i) What is a complementary multivibrator? (1 mark)

- ✓ *Is a multivibrator circuit in one transistor is replaced by its complimentary and turning half the circuit upside down. So one transistor is n-p-n while the other is p-n-p.*

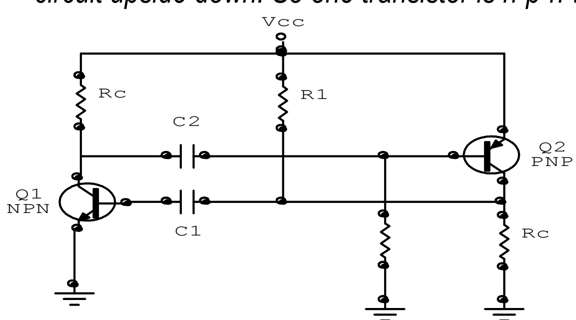


Fig 1

(ii) The circuit diagram (Fig 1) is a complementary multivibrator. Identify its category. (1mark)

- ✓ *Complementary astable multivibrator.*

Que 3. (i) What do you understand by symmetrical triggering? (1mark)

- ✓ *Symmetrical triggering uses only one triggering input to the input of any one transistor. The state of the circuit is changed each time a trigger pulse is applied. Thus each successive triggering signal induces a transition, regardless of the state in which the bistable circuit is.*

(ii) How the linearity of current sweep generators can be improved? (1mark)

- ✓ *Linearization using voltage source*

Que 4. (i) Define a Multivibrator (1mark)

- ✓ *A two stage amplifier operating in two modest used to generate non-sinusoidal waveforms.*

(ii) Why in bistable multivibrator, both transistors cannot remain in active condition? (1mark)

- ✓ *Because there is no state of equilibrium, when collector current in one transistor is increasing driving that transistor into saturation, the collector current in the other transistor is decreasing causing cut-off.*

Que 5. Mention the application of blocking oscillator. (2 marks)

- ✓ *To obtain abrupt pulses from slowly varying input triggering voltage,*
- ✓ *To provide triggers for system synchronization.*
- ✓ *To provide large peak power pulses,*
- ✓ *As a low impedance switch used to discharge a capacitor very quickly.*

✓ *To produce gating waveforms with very low mark space ratio.*

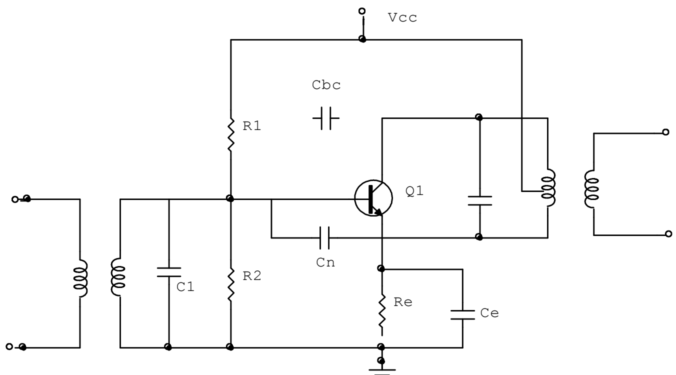
Que 6. What do you understand by symmetrical triggering? (2 marks)

Que 7. Define a current time base generator. (2 marks)

✓ *A circuit which produces currents that are linearly increasing with time.*

PART B (3 x 4 = 12 Marks)

Que 8. **(a)** Given a circuit diagram of a tuned RF amplifier with Hazeltine neutralization explain how the circuit functions. (4 marks)



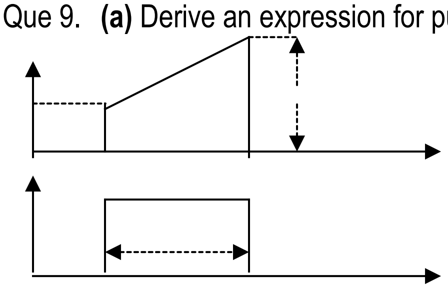
✓ *A small value of variable capacitance C_N is connected from the bottom of coil, point B to the base. Therefore, the internal capacitance C_{bc} feeds a signal from the top end of the coil, point A, to the transistor base and the C_N feeds a signal of equal magnitude but opposite polarity from the bottom of coil, point B, to the base. The neutralizing capacitor C_N can be adjusted to nullify the signal fed through C_{bc}.*

OR

Que.8 **(b)** A Class-C tuned amplifier has inductance of $3\ \mu H$ and capacitance of $470\ pF$ in the tank circuit. Calculate the resonant frequency. (4 marks)

✓ $f_r = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{3\mu H \times 470 pF}} = 4.238\ MHz$

Que 9. **(a)** Derive an expression for pulse width of emitter timing Monostable blocking oscillator.(4 marks)



At point P the pulse terminates, hence $i_c = h_{FE} i_B$

From $i_c = i + i_m \rightarrow \frac{n^2 V_{cc}}{R} + \frac{V_{cc} t_p}{L} = h_{FE} \frac{n V_{cc}}{R}$

$\therefore t_p = \frac{nL}{R}\left(h_{FE} - n\right) \approx \frac{nL}{R} h_{FE}$

OR

Que. 9 **(b)** When designing a linearization circuit using a constant current source, a compromise should be made on choosing Rs of the voltage compensating circuit. Explain. (4 marks)

✓ *From* $\tau = \frac{L}{R_S}$;

- ✓ *When R_S is small, τ is large, the current will decay very slowly and a long period will elapse before the next sweep. But due to slowly varying current, the peak voltage developed across the current source will be small.*
- ✓ *When R_S is large τ is small, the current reduces to zero after the end of the sweep, then a large peak voltage will appear across the current source, which is undesirable.*

Que 10. **(a)** Derive the equation for a *slope error* in a current base generator circuit when practical conditions are considered, i.e. the yoke having finite internal resistance R_L and the collector saturation resistance R_{CS} not neglected. (4 marks)

✓

$$e_s = \frac{\left.\frac{diL}{dt}\right|_{t=0} - \left.\frac{diL}{dt}\right|_{t=T_s}}{\left.\frac{diL}{dt}\right|_{t=0}} = \frac{\frac{V_{cc}}{L} - \frac{V_{cc}}{L}e^{-\frac{(R_L+R_{CS})}{L}T_s}}{\frac{V_{cc}}{L}} = 1 - e^{-\frac{(R_L+R_{CS})}{L}T_s}; hence\ e_s = I_L\left(\frac{R_L+R_{CS}}{L}\right)T_S$$

OR

Que 10 **(b)** With an aid of a simple current time base generator circuit (1.5 marks), illustrate that inductor current varies linearly with time (1 mark), showing the corresponding waveforms. (1.5 marks)

PART C (2 x 12 = 24 Marks)

Que 11. **(a)** (i) Draw a circuit diagram of a fixed bias bistable multivibrator, (4 marks)

(ii) The fixed bias bistable multivibrator uses the following parameters:

$V_{CC} = +12\ V;\ V_{BB} = -8\ V;\ R_1 = 10\ k\Omega k\Omega;_2 = 50\ k\Omega k$
$R_C = 2.2\ k\Omega;\ V_{CE}(sat) = 0.2; and\ V_{BE}(sat) = 0.7V$

(i) Calculate the stable currents $(I_{C1}; I_{C2}; I_{B1}; and\ I_{B2})$ (4 marks)

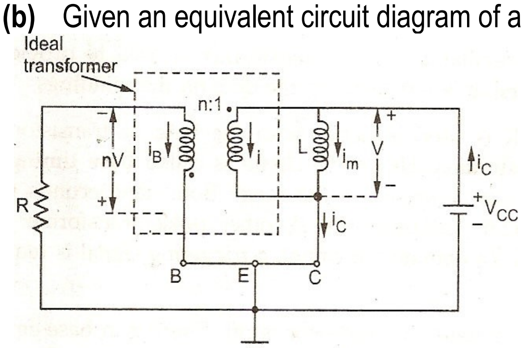
✓ $I_{C1} = 0\ mA; \quad I_{C2} = 5.223\ mA; \quad I_{B1} = 0\ mA; \quad and \quad I_{B2} = 0.752\ mA$

(ii) Calculate the stable voltages $(V_{B1}; V_{B2}; V_{C1}; and\ V_{C2}.)$ (4 marks)

✓ $V_{B1} = -1.16\,V; \quad V_{B2} = 0.7\,V; \quad V_{C1} = 9.9628\,V; \quad and \quad V_{C2} = 0.2\,V.$

OR

Que 11 **(b)** Given an equivalent circuit diagram of a Monostable blocking oscillator using base timings,



i. Analyze the circuit, (4 marks)

ii. Derive the expression for the pulse width (4 marks)

ANS

✓ $i_c = i + i_m \rightarrow \frac{n^2 V_{cc}}{R} + \frac{V_{cc} t_p}{L} = h_{FE} \frac{n V_{cc}}{R}$

$\therefore t_p = \frac{nL}{R}(h_{FE} - n) \approx \frac{nL}{R} h_{FE}$

iii. Mention any two (2) disadvantages of the circuit. (4 marks)

- ✓ The pulse width is a linear function of h_{fe} which is temperature dependent,
- ✓ The value of h_{fe} changes from transistor to transistor and hence the pulse width gets affected due to transistor replacement.

Que 12. **(a)** Determine the values of capacitors to be used in an astable multivibrator to provide a train of pulse 2 μsec wide at a repetition rate of 75 kHz with $R_1 = R_2 = 10$ kΩ (12 marks)

$$T_1 = 2\,\mu\text{sec}; f = 75\,kHz \therefore T = \frac{1}{75 \times 10^3} = 13.33\,\mu\text{sec};\ T = T_1 + T_2 \therefore T_2 = 13.33 - 2 = 11.33\mu\text{sec}$$

$$T_1 = 0.69 R_1 C_1 \quad hence\, 2 \times 10^{-6} = 0.69 \times 10 C_1 \rightarrow C_1 = 289.85\,pF;$$

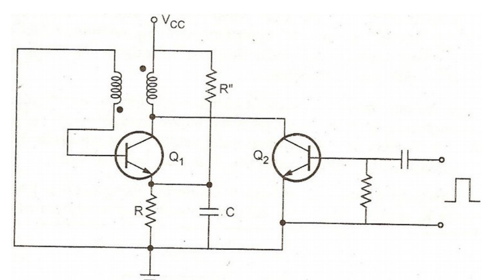
$$But\, T_2 = 0.69 R_2 C_2 \rightarrow C_2 = \frac{11.33}{0.69 \times 10 \times 10^3} = 1.642\,pF$$

OR

Que 12. **(b)** With a circuit diagram, explain the triggering for Monostable blocking oscillator (12 marks)

ANS;

Triggering Circuit using Transistor.



For triggering, it is necessary to lower the collector voltage of Q_1 momentarily by the triggering circuit. If momentarily a positive-going pulse is applied to the base of Q_2, will drive Q_2 into saturation, hence providing a short across the collector of Q_1 which momentarily brings the collector voltage of Q_1 to zero.

Due to phase inversion, such a voltage gets induced in the secondary which makes the base of Q_1 positive, hence it starts conducting.

Thereafter, the working of Monostable Blocking Oscillator starts. When the pulse is removed from Q_2, Q_2 becomes off (open circuit) hence the triggering source causes no interaction with the blocking oscillator.

B. **Model Examinations**

MODEL EXAMINATION **Nov 2008**

TIME – 3 Hrs

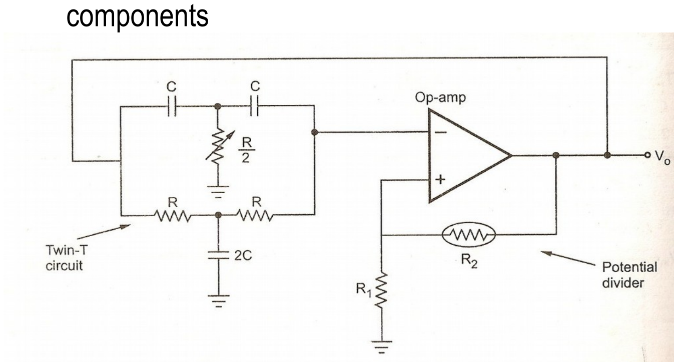
PART A (10 x 2 = 20 Marks)

1. Give topology for various types of feedback amplifiers.
2. Define (i) feedback factor β , (ii) Negative feedback.
3. State the Barkhausen Criterion for feedback oscillators?
4. Classify feedback oscillators.
5. Mention any two (2) applications of Class-C tuned amplifiers.
6. What do you understand by tuned amplifier?
7. Why quality factor is kept as high as possible in tuned circuit?
8. Explain why commutating capacitors are used in multivibrators.
9. Define blocking oscillator.
10. Why inductors are used in current time base generators?

PART B (5 x 4 = 20 Marks)

11. **(a)** What are the advantages of negative feedback over positive feedback?
 OR
 (b) Given the open loop gain of an amplifier A_v and the feedback factor β , show for a voltage amplifier how gain with feedback A_{vf} is related to A_v and A_v .

12. **(a)** List and explain the advantages and disadvantages of Wien Bridge Oscillator.
 OR
 (b) (i) Identify the circuit in Fig.1, (ii) explain its used and (iii) mention the functions of its components



Fig, 1

13. **(a)** (i) Draw and (ii) explain the basic mixer circuit.
 OR
 (b) The bandwidth of a double-tuned amplifier is 20 kHz. Calculate the bandwidth if such three stages are cascaded.
14. **(a)** Mention any four (4) applications of Blocking Oscillators.

OR

(b) Why in bistable multivibrator, both transistors cannot remain in active condition? Explain.

15. **(a)** (i) Draw the circuit for simple current time base generator.
(ii) How linearization is achieved in current time base generators? Explain
OR
(b) Mention any four (4) applications of pulse transformers.

PART C (5 x 12 = 60 Marks)

16. **(a)** Using a diagram of a typical feedback connection around an amplifier, explain
(i) The feedback concept, (3 marks)
(ii) Ways of sampling and functions of the sampling network, (3 marks)
(iii) The feedback network, (3 marks)
(iv) Ways of mixing and functions of the mixer network. (3 marks)

OR

(b) An amplifier with open loop gain of $A = 2000 \pm 150$ is available, it is necessary to have the amplifier whose voltage gain varies by not more than $\pm 0.2\%$. Calculate β and A_f .

17. **(a)** The frequency sensitive arm of the Wien bridge oscillator uses $C_1 = C_2 = 0.001\mu F$ and $R_1 = 10k\Omega,$ while R_2 is kept variable. The frequency is to be varied from $10kHz\ to\ 50\ kHz$ by varying R_2 . Find the minimum and maximum values of R_2 .
OR
(b) In not less than 5 points, give similarities and differences between RC phase shift oscillator and Wien bridge oscillator

18. **(a)** (i) Identify the circuit given in Fig. 2. (2 marks)
(ii) Explain the purpose of the circuit (2 marks)
(iii) Mention the circuit configuration and (2 marks)
(iv) The working of each individual component.(6 marks)

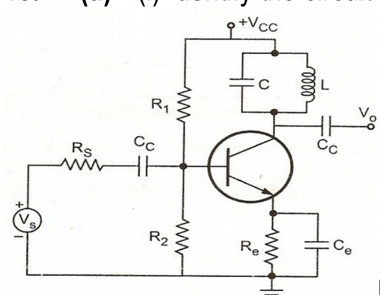


Fig. 2

OR

(b) (i) Identify the circuit given in Fig. 3 (2 marks)
(ii) Explain the purpose of the circuit (2 marks)
(iii) Mention the configuration (2 marks)
(iv) Explain the working of each individual component. (6 marks)

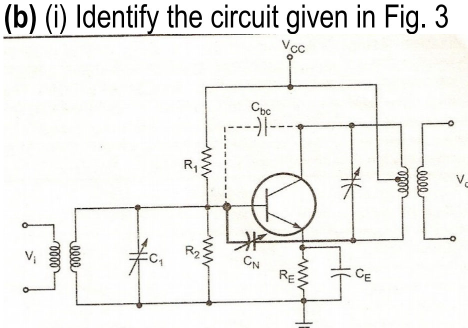


Fig. 3

19. **(a)** For the multivibrator circuit in Fig 4

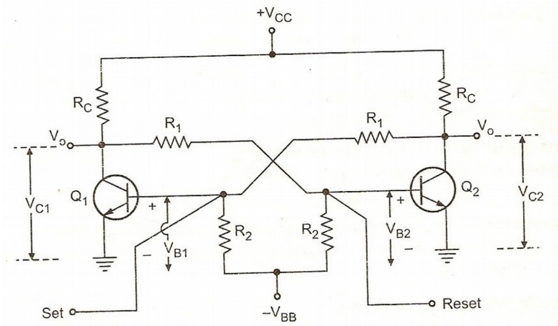


Fig.4.

(i) Identify type of the multivibrator and draw its output waveforms, (4 marks)
Using the following parameters
$V_{cc} = +12\ V;\ V_{BB} = -8\ V,\ R_1 = 10\ k\Omega k\Omega,_2 = 50\ k\Omega k\Omega,_C = 2.2\ k\Omega$ and
neglecting the junction voltages, calculate;
(ii) Stable voltages ($V_{C1},\ V_{C2},\ V_{B1},\ V_{B2}$) (4 marks)
(iii) Stable currents ($I_{C1},\ I_{C2},\ I_{B1},\ I_{B2}$) (4 marks)

OR

(b) Determine the values of capacitors in Fig.5 to be used in an astable multivibrator to provide a

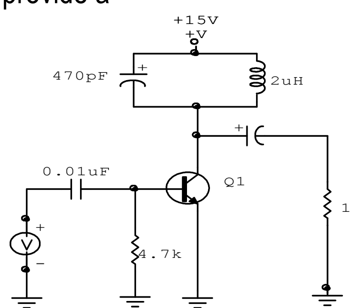


Fig.5

train of pulse 2 μsec wide at a repetition rate of 75 kHz with $R_1 = R_2 = 10\ k\Omega$ (12 marks)

20. (a) With a
(i) Clearly labeled diagram for triggering circuit for Monostable Blocking Oscillator ((6 marks)
(ii) Explain the triggering circuit for Monostable Blocking Oscillator. (6 marks)

OR

(b) Establish the relation between, (indicating clearly the assumption made):
(i) Tilt and time constant in high pass circuit, (6 marks)
(ii) Rise time and time constant (6 marks)

End Semester Examinations

St JOSEPH COLLEGE OF ENGINEERING and TECHNOLOGY

END SEMESTER EXAMINATION KEY ANSWERS - Nov 2008

54 EC 402 ELECTRONIC CIRCUITS II

PART A (10 x 2 = 20 Marks)

1. What is meant by Voltage amplifier?

ANS: A voltage amplifier is an amplifier circuit which provides a voltage output proportional to the voltage input, and the proportionality factor does not depend on the magnitudes of the source and load resistances. (2%)

OR

ANS: A voltage amplifier is an amplifier circuit with an input resistance R_i which is large compared to the source resistance R_s, then $V_i \approx V_s$; and the external load resistance R_L is large compared with the output resistance R_o, then $V_o \approx V_V V_i \approx V_V V_s$ (2%)

2. Define negative and positive feedback.

ANS: (i) Negative feedback – the input signal and part of the output which is fed back to the input are out of phase. (1%)

(ii) Positive feedback - the input signal and part of the output which is fed back to the input are in phase. (1%)

3. Identify, where does the starting voltage of an oscillator come from?

ANS: Every resistance has some free electrons. Under the influence of normal room temperature, these free electrons move randomly in various directions. Such a movement of the free electrons generates a voltage called noise voltage, across the resistance. Such noise voltages are amplified and to start oscillations, $|A\beta|$ is kept greater than unity at start. Such amplified voltage appears at the output terminals. When the part of this output is sufficient to drive the input of the amplifier circuit, the circuit adjusts to get $|A\beta| = 1$ and with phase shift of 360^0 we get sustained oscillations. (2%)

4. Explain the need of RC phase shift in RC phase shift oscillators?

ANS: The amplifier causes a phase shift of 180^0 then the feedback network should create a phase shift of 180^0 to satisfy the Barkhausen criterion. In an RC circuit the output voltage leads the input voltage by phase angle ϕ ; $\left(\tan\phi = -\frac{1}{2\pi fRC}\right)$. The values of R and C in the feedback network are selected so that the phase angle $\phi = 60^0$. Hence in a phase shift oscillator, three sections of RC circuit are connected in cascade, each introducing a phase shift of 60^0, thus introducing a total phase 180^0 due to the feedback network.(2%)

5. What is meant by unloaded Q of tank circuit?

ANS: Unloaded Q is the ratio of stored energy to dissipated energy in a reactor or resonator. (2%)

OR Unloaded Q of an inductor or capacitor is $\frac{X}{R_s}$ where X -reactance R_s -series resistance. (2%)

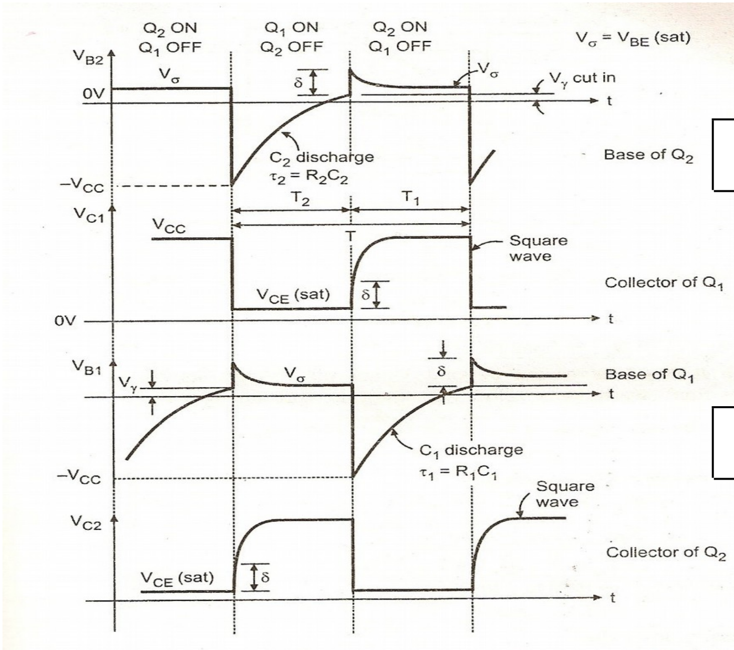
6. Mention the applications of Class C tuned amplifier.

ANS: (i) In a mixer circuit of a radio receiver as a down-converter, to convert incoming signal to a lower frequency where it is easier to obtain the high gain and selectivity required. (1%)

(ii) In a mixer circuit as up-converter to translate signal to higher frequency. (1%)

7. Draw the typical waveforms of base and collector of a collector coupled astable multivibrator.

ANS:



0.5 %

0.5%

0.5 %

0.5%

Fig. 4.36 Waveforms of collector coupled astable multivibrator

8. Mention the applications of Schmitt trigger circuit.

ANS: (i) Amplitude comparator: - it identifies the moment at which any arbitrary waveform attains a particular reference level. (1%)

(ii) Squaring circuit: - any arbitrary input can be converted to a square wave having amplitude which is independent of the amplitude of the wave form. (1%)

9. State the effect of saturation voltages on pulse width.

ANS:

From the equation of pulse width $t_p = \frac{nL(h_{FE} - n)}{R(h_{FE} + n)} - \frac{n_1^2 L}{R_L}$ it is assumed that the transistor saturation voltages are small compared to V_{CC} and are neglected. Hence the pulse width is independent of the supply voltage

When not neglected $t_p = \frac{nL(h_{FE} - n)}{R(h_{FE} + n)} \frac{V_{CC} - V_{CE(sat)} - \left[\frac{V_{BE(sat)}}{n}\right]}{V_{CC} - V_{CE(sat)} + V_{BE(sat)}} - \frac{n_1^2 L}{R_L}$ Hence the pulse width is depends on the supply voltage

10. Mention <u>any two applications</u> of blocking oscillator.

ANS:

(i) To obtain abrupt pulses from slowly varying input triggering voltage, Monostable blocking oscillator is used.

(ii) An astable blocking oscillator is used to as a main device to supply triggers for synchronization of a system having pulse type waveforms,

(iii) To produce large peak power pulses,

(iv) The output winding can be isolated from ground whenever required by using tertiary winding of pulse transformer,

(v) Output pulses of either polarity can be obtained due to the use of tertiary winding of pulse transformer,

(vi) Blocking oscillators are used as frequency dividers or as a counter circuit,

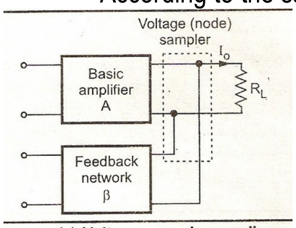
(vii) The blocking oscillator can be used as low impedance switch used to discharge a capacitor very quickly,
(viii) The output of a blocking oscillator can be used to produce gating waveform with very low mark space ratio.

PART B (5 x 4 = 20 Marks)

11. (a) Explain the sampling network

ANS: A sampling network is a circuit that samples portion of the output and feeds it to the feedback network. (2%)

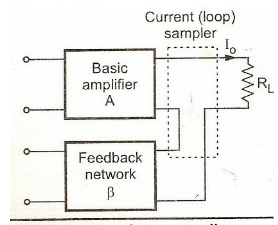
According to the sampling parameter;



(a) Voltage or node sampling

- A network that samples the output voltage by connecting the feedback network in shunt across the output is referred to as voltage or node sampling.

- A network that samples the output current by connecting the feedback network in series with the output is referred to as current or loop sampling. (0.5%)



(b) Current or loop sampling

(0.5%)

OR

(b) (i) Define desensitivity D. (ii) For large value of D what is Af ? (iii) What is the significance of this result?

ANS: (i) Desensitivity D is the reciprocal of sensitivity;
Sensitivity is defined as the fractional change in amplification with feedback divided by the fractional change in amplification without feedback; (1%)

Given $\frac{dA_f}{A_f}$ - Fractional change in amplification with feedback

$\frac{dA}{A}$ - fractional change in amplification without feedback

Hence, $D = \frac{\frac{dA_f}{A_f}}{\frac{dA}{A}}$ from $\frac{dA_f}{A_f} = \frac{dA}{A}\left(\frac{1}{1+\beta A}\right) \Rightarrow D = (1+\beta A)$ (1%)

(ii) For large value of D , $\beta A >> 1 \Rightarrow A_f = \frac{A}{1+\beta A} = \frac{A}{\beta A} = \frac{1}{\beta}$ (1%)

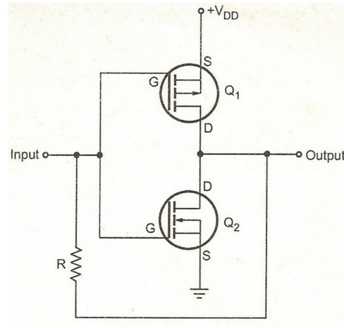
(iii) The significance of this result is that the stability of the amplifier increases with increase in desensitivity. (1%)

12 (a) Explain with suitable example, how logic gates are used as linear amplifiers,

ANS:

When MOSFETs operate in a linear region of their characteristics by giving appropriate biasing, the circuit works as a linear amplifier.

A feedback resistor is connected between input and output to provide negative feedback. The resistor sets the output dc voltage $\approx \frac{V_{DD}}{2}$ which is feedback to the common gate input. Due to this gate voltage $\left(\frac{V_{DD}}{2}\right)$, both MOSFETs will have equal bias and they will conductor equally. If a sinusoidal signal is applied at the input, biasing is varied accordingly and output will change linearly with respect to the input voltage, hence **logic gates** can be used as linear amplifiers. (2%)



(a) Circuit diagram

Diagram (2%)

OR

(b) Explain how amplitude is maintained in oscillators with negative feedback.

ANS: The noise voltages are amplified, and to start oscillations $|A\beta|$ is kept greater than unity at start. Such amplified voltage appears at the output terminals. When the part of this output is sufficient to drive the input of the amplifier circuit, the circuit adjusts to get $|A\beta| = 1$ and with phase shift of 360^0 we get sustained oscillations. (4%)

13 (a) Explain the Hazeltine method of neutralization;

ANS:

Instability in tuned amplifiers is due to inter junction capacitance between base and collector C_{bc} of the transistor. Hazeltine neutralization is a circuit in which a small value of neutralizing capacitor of variable capacitance C_N is connected from the bottom of the coil, point B, to the base. Therefore the internal capacitance C_{bc}, feeds a signal from the top end of the coil, point A, to the transistor base and C_N feeds a signal of equal magnitude but opposite polarity from the bottom of the coil, point B, to the base. The neutralizing capacitor C_N, can be adjusted to nullify the signal fed through C_{bc} (2%)

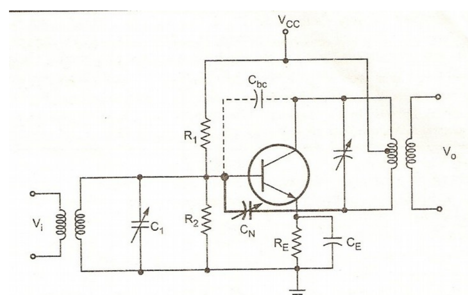


Fig. 3.20 Tuned RF amplifier with Hazeltine neutralization (2%)

OR

(b) Draw and explain the basic mixer circuit.

ANS:

The mixer accepts two inputs: the signal f_s which is translated to another frequency and sine wave f_o from the oscillator. The mixer performs mathematical multiplication of its two input signals and produce output signals f_s, f_o, $f_o + f_s$ *and* $f_o - f_s$. A tuned circuit or filter is used at the output of the mixer to select the desired frequency. (1%)

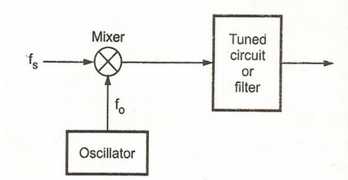


Fig. 3.30 Block schematic of mixer circuit

(1%)

The transistor is biased to operate as a class C amplifier so that the collector current does not vary linearly with variations in the base current. This results in analog multiplication which produces the sum and difference frequencies whereby the tuned circuit selects the sum or difference frequency. (1 mark)

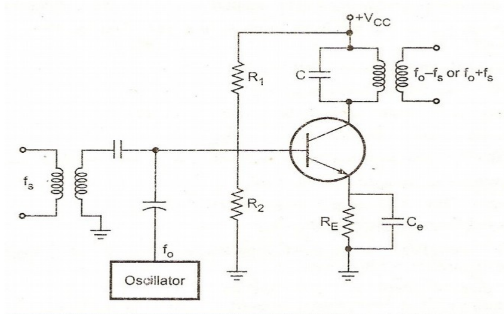


Fig. 3.31 Mixer circuit using class-C-turned amplifier (1%)

14 (a) Brief about UTP and LTP of Schmitt Trigger,

ANS: UTP- Upper Threshold Point of Schmitt trigger is the input turn-on voltage level which makes Q_1 get driven to active region, i.e., ON. At this point the output voltage instantaneously increases to V_{cc} (1%)

LTP – Lower Threshold Point of Schmitt trigger is the input turn-off voltage level which makes Q_1 become OFF and the output instantaneously switches over from V_{cc} to a lower level. (1%)

Schmitt trigger Waveforms (2%)

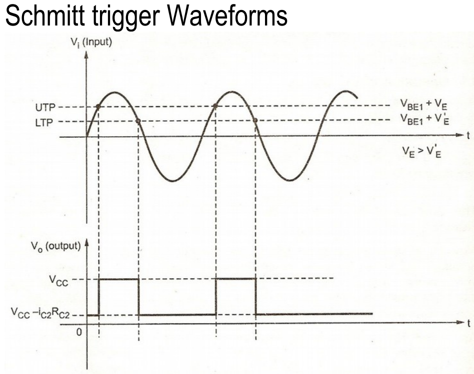


Fig. 4.48 Schmitt trigger waveforms

OR

(b) Draw the circuit diagram of complementary transistor Monostable multivibrator and explain its operation.

ANS: Explanation of the circuit construction: (1.5%)

One transistor is n-p-n while the other is p-n-p. The two corrector resistances are equal R_C. The output of Q_2, i.e. collector of Q_2 is coupled to the base of Q_1 through R_1 which is shunted by a speed-up capacitor C_1 to reduce the transition time.

The collector of Q_1 is coupled to the base of Q_2 through capacitor C_1 to make capacitive coupling. The resistance R at the input of Q_2 is returned to the supply voltage V_{cc}. The values of R_2 and $-V_{BB}$ are chosen such that the transistor Q_1 is OFF by reverse biasing it and transistor Q_2 is ON (in saturation) by forward biasing it with V_{cc} and R.

Explanation on the circuit operation: (1.5%)

When a positive trigger of sufficient magnitude and duration is applied to the base of Q_1, Q_1 starts conducting. Due to this, voltage at its collector decreases. This is coupled to the base of Q_2 through C_1. Decrease in V_{C1} directly cause a decrease in the base voltage V_{B2} which decreases the forward bias of Q_2 and hence collector current I_2 decreases. Thus the collector voltage of Q_2 increases which is applied to the base of Q_1, and makes Q_1 to become driven to saturation while Q_2 gets driven to cut-off. This is quasi-state of the circuit in which it remains for a finite time T.

Circuit diagram (1%)

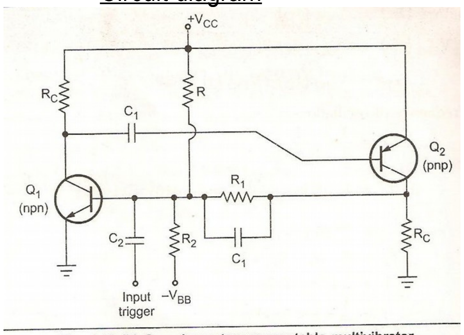


Fig. 4.46 (b) Complementary monostable multivibrator

15 (a) Derive the expression for the time period of an astable blocking oscillator,

ANS:

The voltage V across collector is the same as voltage across L

$$V = L\frac{di_m}{dt} \;\therefore\; di_m = \frac{V}{L}dt \;\Rightarrow \int di_m = \frac{V}{L}\int_0^t dt; \quad Hence\ i_m = \frac{V}{L}t = \frac{V_{CC}}{L}t$$

For equivalent circuit

$$\mathrm{i_C} = i + i_m = \frac{n^2 V_{CC}}{R} + \frac{V_{CC}}{L}t \;\therefore\; at\ t = 0;\; i_B = \frac{nV_{CC}}{R},\ and\ i_C = \frac{n^2 V_{CC}}{R}$$

$$t = t_p;\;\; i_m = \frac{V_{CC}t_p}{(n+1)L} = I_0,\ and\ t_p = \frac{nL}{R}$$

The diode network D in the circuit is replaced by it model with a battery of V_γ in series with $R_f \approx 0$. Due battery of V_γ, the collector voltage V_C increases by V_γ above V_{CC}. Neglecting R_f, the voltage across L is $-V_\gamma$; $\quad L\frac{di_m}{dt} = -V_\gamma \quad \therefore di_m = \frac{-V_\gamma}{L}dt$

Integrating and using constant of integration as initial current I_0, $i_m = \frac{-V_\gamma t}{L} + I_o$

The negative sign indicates that current i_m decreases linearly and becomes zero at $t = t_f$

$$\therefore 0 = \frac{-V_\gamma t_f}{L} + I_0;\;\; but\ I_0 = \frac{n}{n+1}\frac{V_{CC}}{R};\; Hence\ t_f = \frac{n}{n+1}\frac{L}{R}\frac{V_{CC}}{V_\gamma}$$

As i_m reduces to zero, the diode becomes an open circuit and forms a tank circuit that produces an oscillatory response with $f_r = \frac{1}{2\pi\sqrt{LC}} \quad \therefore T' = \frac{1}{f_r} = 2\pi\sqrt{LC}$

And $t_a = \frac{1}{4}T'$ =quarter of a cycle; $t_a = \frac{\pi}{2}\sqrt{LC'}$ Hence the period of oscillation T $T = t_p + t_f + t_a$

OR

(b) Draw and explain the triggering circuit used for Monostable blocking oscillator.

ANS: The triggering circuit consists of transistor Q_2. A positive going pulse is momentarily applied to the base of Q_2 which drives Q_2 into saturation making the drop across Q_2 to be zero and provide a short across the collector of Q_1. This momentarily brings the collector voltage of Q_1 to zero. Due to phase inversion, such a short gets induced in the secondary which makes the base of

Q_1 positive and brings it out of cut-off state and thereafter normal working of Monostable blocking oscillator starts. (2%)

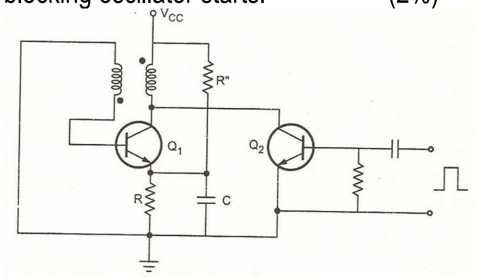


Fig. 5.15 Triggering circuit using transistor (2%)

PART C (5 x 12 = 60 Marks)

16 (a) With an aid of a block diagram and explain (i) voltage amplifier with voltage series feedback and (ii) transconductance amplifier with current series feedback

ANS: (i) Voltage amplifier with voltage series feedback

The sampling parameter is the voltage which is done in parallel with the output, and the feedback parameter is the voltage which is fed in series with the signal voltage; (3%)

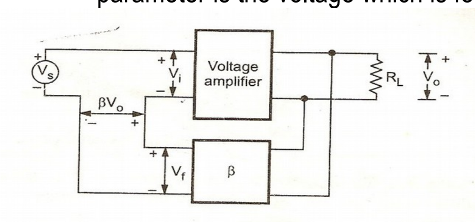


Fig. 1.9 (a) Voltage amplifier with voltage series feedback (3%)

(ii) Transconductance amplifier with current series feedback

The sampling parameter is the current which is done in series with the output, and the feedback voltage is fed in series with the signal voltage;

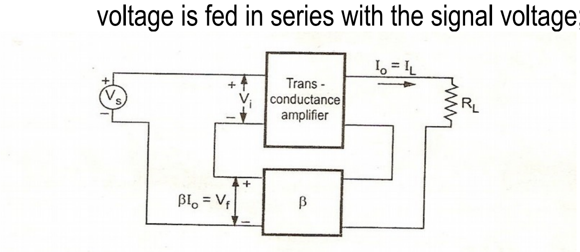


Fig. 1.9 (b) Transconductance amplifier with current series feedback (3%)

OR

(b) With an aid of a block diagram and explain (i) current amplifier with current-shunt feedback and (ii) transresistance amplifier with voltage-shunt feedback

ANS: (i) Current amplifier with current-shunt feedback;

The sampling parameter is the current which is done in parallel with the output, and the feedback parameter is the current which is fed in parallel with the signal current; (3 marks)

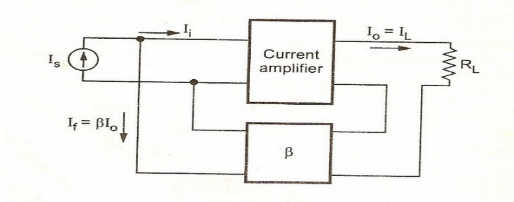


Fig. 1.9 (c) Current amplifier with current-shunt feedback (3%)

(ii) Transresistance amplifier with voltage-shunt feedback

The sampling parameter is the voltage which is done in parallel with the output, and the feedback parameter is the current which is fed in parallel with the signal current; (3%)

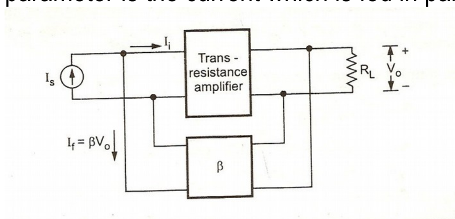


Fig. 1.9 (d) Transresistance amplifier with voltage shunt feedback (3%)

17 (a) (i) Draw the circuit diagram of a pierce crystal oscillator and (ii) explain its operation.

ANS: (i) Circuit diagram of a pierce crystal oscillator (6%)

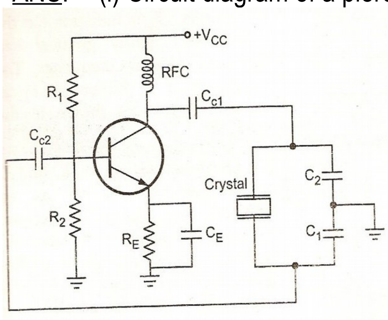


Fig. 2.68 Pierce crystal oscillator

(ii) Operation of a pierce crystal oscillator

The inductor in the Colpitts oscillator is replaced by the crystal which behaves as an inductor for frequency slightly higher than the series resonance frequency f_s. The resistance R1 R2, RE provide dc bias while the capacitor CE is emitter bypass capacitor. RFC provides isolation between ac and dc operation. Cc1 and Cc2 are coupling capacitors. The resulting circuit frequency is set by the series resonant frequency of the crystal. Changes in the supply voltages, temperature, and transistor parameters have no effect on the circuit operating conditions hence good frequency stability is obtained. (6%)

OR

(b) In a transistorized Hartley oscillator the two inductances are $2mH$ and $20\mu H$ while the frequency is to be changed from $950kHz$ to $2050kHz$. Calculate the range over the capacitor is to be varied.

ANS: The frequency is given by

$$f = \frac{1}{2\pi\sqrt{C(L_{eq})}};\ L_{eq} = L_1 + L_2 = 2\times10^{-3} + 20\times10^{-6} = 0.00202kHz$$ (4%)

$$for\ f_{\max} = 2050\times10^3;\ \ hence\ \ 2050\times10^3; = \frac{1}{2\pi\sqrt{C(0.0020)}} \Rightarrow C = 2.98pF$$ (4%)

$$for\ f_{\min} = 950\times10^3;\ \ hence\ \ 950\times10^3; = \frac{1}{2\pi\sqrt{C(0.0020)}} \Rightarrow C = 13.89pF$$

(4%)

18 (a) Explain the concept of staggering tuned amplifier with the help of frequency response,

ANS:

The concept of staggering tuned amplifier; (6%)

The double tuned amplifier give greater 3dB bandwidth having steeper side and flat top, but alignment of double tuned amplifier is difficult. To overcome this problem two single tuned cascaded amplifiers are so adjusted that the resonant frequencies are staggered (displaced by an amount equal to the bandwidth of each stage $2\times\left(\frac{\Delta_1}{2}\right)$.

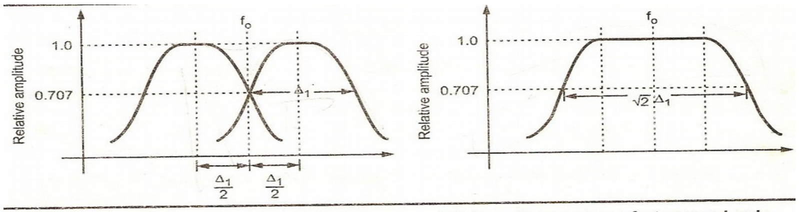


Fig. 3.17 (a) Response of individual stages (b) Overall response of staggered pair

The advantage of staggered tuned amplifier is a better flat, wideband characteristics (6%)

OR

(b) Explain the <u>stabilization techniques</u> used in tuned amplifiers.

<u>ANS</u>:

Occurrence of instability (3%)

Instability in tuned amplifiers is due to inter junction capacitance between base and collector C_{bc} of the transistor. As reactance of C_{bc} at RF is low enough to provide the feedback path from the collector to the base, which if it reaches the input in a positive manner with a proper phase shift, then there is the possibility of the circuit to become unstable and generate its own oscillations thus stopping working as an amplifier. Diagram (3%)

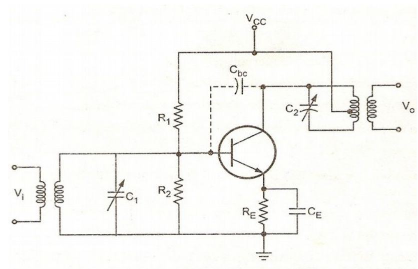


Fig. 3.19 Tuned RF stage

<u>Stabilization techniques</u> **<u>(Any one of the listed options)</u>**

(i) <u>Hazeltine neutralization technique</u>;

Is a circuit in which a small value of neutralizing capacitor of variable capacitance C_N is connected from the bottom of the coil, point B, to the base. Therefore the internal capacitance C_{bc} , feeds a signal from the top end of the coil, point A, to the transistor base and C_N feeds a signal of equal magnitude but opposite polarity from the bottom of the coil, point B, to the base. The neutralizing capacitor C_N , can be adjusted to nullify the signal fed through C_{bc} (3%)

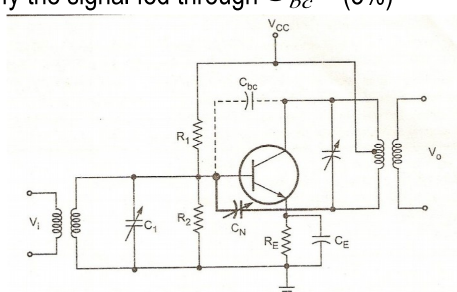


Fig. 3.20 Tuned RF amplifier with Hazeltine neutralization

Diagram (3%)

(ii) <u>Rice neutralization technique</u>; (3%)

The Rice circuit of neutralization uses a centre tapped coil in the base circuit. With this arrangement the signal voltages at the end of the tuned base coil are equal and out of phase.

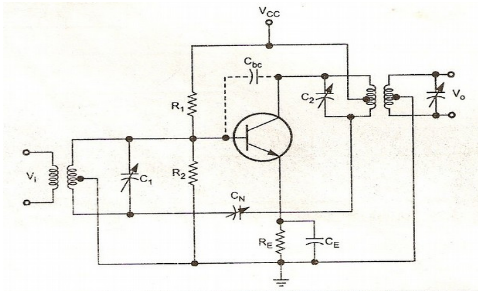


Fig. 3.23 Tuned RF amplifier using Rice neutralization

Diagram (3%)

(iii) Neutralization using coil (3%)

The L part of the tuned circuit at the base of the next stage is oriented for minimum coupling to the other windings, wound on a separate form and is mounted at right angles to the coupled windings. If the windings are properly polarized, the voltage across L due to the circulating current in the base circuit will have the proper phase to cancel the signal coupled through the base to collector, C_{bc} capacitance

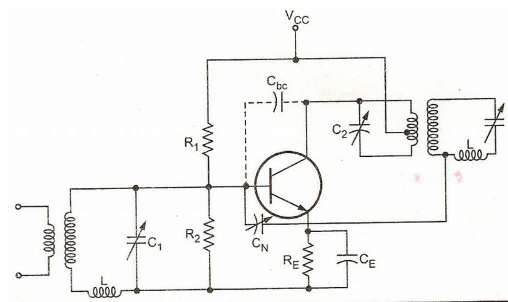


Fig. 3.22 Tuned RF amplifier using coil

Diagram (3%)

19 (a) If class C tuned amplifier has $R_L = 6k\Omega$ and required tank circuit Q = 80. Calculate values of (i) L, (ii) C of the tank circuit. Assume V_{CC}-20 V, resonant frequency = 5 MHz and worst case power dissipation = 20 mW.

ANS:

From $P_{D\max} = \frac{(V_{pp\max})^2}{40r_c} = \frac{(2V_{CC})^2}{40r_c} \therefore\Rightarrow r_c = \frac{(2V_{CC})^2}{40P_{D\max}} = \frac{(2\times 20)^2}{40\times 20\times 10^{-3}} = 2k\Omega$ (3%)

Since $r_c = R_P \| R_L \therefore\Rightarrow \frac{1}{R_P} = \frac{1}{r_c} - \frac{1}{R_L} = \frac{1}{2\times 10^3} - \frac{1}{6\times 10^3} = 3.33\times 10^{-4} \because R_P = 3k\Omega$ (3%)

But $R_P = Q_L \times \omega_r \times L = Q_L \times 2\pi f_r \times L \therefore\Rightarrow L = \frac{R_P}{Q_L \times 2\pi \times f_r} = \frac{3000}{80\times 2\pi \times 5\times 10^6} = 1.19\mu H$

(3%)

We know that

$f_r = \frac{1}{2\pi\sqrt{LC}}$ $hence \Rightarrow C = \frac{1}{(2\pi)^2 L f_r^2} = \frac{1}{(2\pi)^2 \times 1.19\times 10-6\times (5\times 10^6)^2} = 851pF$ (3%)

OR

(b) For the circuit if Fig 1, calculate (i) resonant frequency, (ii) collector resistance (iii) quality factor and (iv) Bandwidth. Assume Q_L = 120

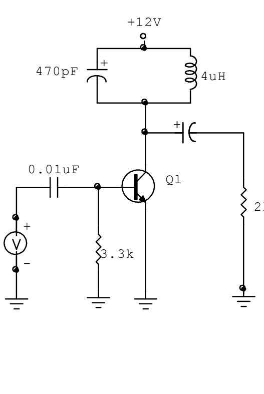


Fig 1

ANS:

(i) Resonant frequency f_r (3%)

$$f_r = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{4\times10^{-6}\times470\times10^{-12}}} = \frac{1}{2\pi\sqrt{18.8\times10^{-16}}} = \frac{10^8}{2\pi\times4.335} = 0.0367\times10^8 = 3.67\ MHz$$

Since $r_c = R_P \| R_L$

$$R_P = Q_L\times\omega_r\times L = Q_L\times2\pi f_r\times L = 120\times2\pi\times3.67\times10^6\times4\times10^{-6} = 11.062\times10^3 = 11.062\text{k}\Omega$$

(ii) ac collector resistance r_c (3%)

$$r_c = R_P \| R_L \ \therefore\Rightarrow \frac{1}{r_c} = \frac{1}{R_P} + \frac{1}{R_L} = \frac{1}{11.062\times10^3} + \frac{1}{2\times10^3} = (0.0903+0.5)\times10^{-3} = 0.5908\times10^{-3}$$

$$\because\ r_c = 1.6926\ k\Omega = 1692.6\ \Omega$$

(iii) Quality factor Q for the overall circuit (3%)

$$Q = \frac{r_c}{2\pi f_r L} = \frac{1692.6}{2\pi\times3.67\times10^6\times4\times10^{-6}} = \frac{1692.6}{92.19} = 18.36$$

(iv) Bandwidth Bw (3%)

$$Bw = \frac{f_r}{Q} = \frac{3670\times10^3}{18.36} = 199.89\times10^3 = 199.89\ kHz$$

20 (a) Explain with suitable circuit diagrams the performance of Monostable blocking oscillator (with base timing).

ANS:

Construction (2%)

One winding of the pulse transformer is in the collector circuit while the other is in the base circuit with n-time turns as in the collector circuit.

A pulse transformer is used to provide polarity inversion. A resistance R in series with the base of transformer controls the timing, i.e. pulse duration. To operate the circuit, a triggering signal is required to the collector, momentarily.

Operation (2%)

$-V_{BB}$ is applied to the base keeping the base-emitter junction reverse biased to avoid false triggering at increased temperature.

When a triggering signal is applied momentarily to the collector of Q1 such that the collector voltage reduces suddenly, an inverted signal at the base appears. The base potential increases as the collector potential decreases, and when this voltage is more than the cut in voltage the transistor starts conducting drawing current from the supply. This draws more current resulting further decrease in collector potential.

Diagrams @ (2%)

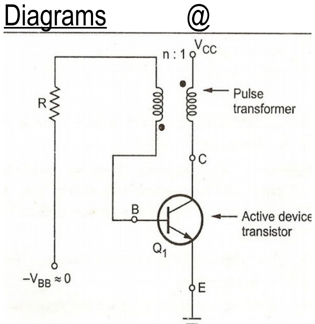


Fig. 5.4 Monostable blocking oscillator

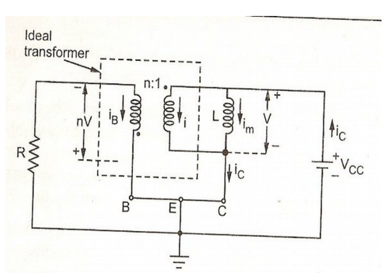


Fig. 5.5 Equivalent circuit

For the pulse width equation consider the equivalent circuit

Circuit analysis (2%)

$$\frac{V_2}{V_1} = n = \frac{i}{i_B} \Rightarrow i = n \times i_B \ \ and\ \ V_2 = n \times V_1 = n \times V_{CC}$$

Applying KVL $i_B = \frac{nV}{R} = \frac{nV_{CC}}{R};\ Hence\ i = n \times \frac{nV_{CC}}{R} = \frac{n^2 V_{CC}}{R}$

The voltage V across collector is the same as voltage across L

$$V = L\frac{di_m}{dt}\ \therefore\ di_m = \frac{V}{L}dt \ \Rightarrow \int di_m = \frac{V}{L}\int_0^t dt;\quad Hence\ i_m = \frac{V}{L}t = \frac{V_{CC}}{L}t$$

For the equivalent circuit

$$i_C = i + i_m = \frac{n^2 V_{CC}}{R} + \frac{V_{CC}}{L}t$$

Therefore, at

$$t = 0;\ \ i_B = \frac{nV_{CC}}{R},\ and\ i_C = \frac{n^2 V_{CC}}{R}$$

$$t = t_p;\ \ i_m = \frac{V_{CC}t_p}{(n+1)L} = I_0,\ and\ t_p = \frac{nL}{R}$$

Current waveforms (2%)

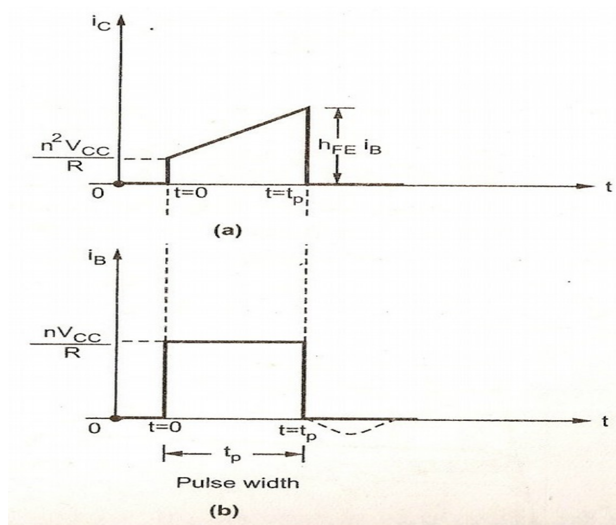


Fig. 5.7 Current waveforms

OR

(b) With appropriate diagrams, explain how Sawtooth wave forms are generated using UJT.

ANS:

Construction (3%)

R_1 and R_2 are biasing resistance which are selected such that they are lower than the interbase resistances R_{B1} and R_{B2}. The resistance R_T and capacitance C_T decide the oscillating rate. The value of R_T is so selected that the operating point of the UJT remains in the negative resistance region.

Circuit and characteristics (3%)

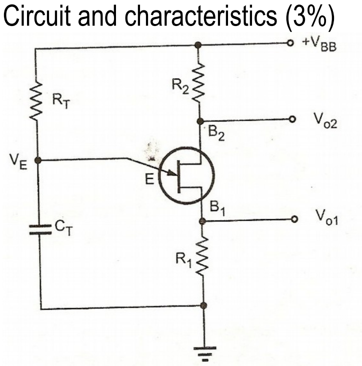


Fig. 5.87 UJT sawtooth generator

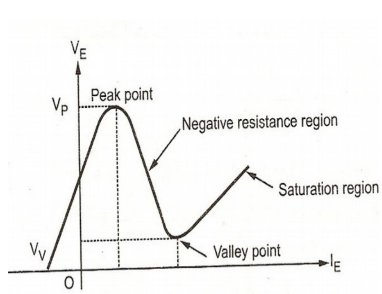


Fig. 5.88 UJT characteristics

Operation (3%)

Capacitor C_T gets charged through R_T towards supply voltage V_{BB}.

- When $V_C < V_P$ the emitter appear as an open circuit

$V_P = \eta V_{BB} + V_D$ where V_D - cutin voltage of diode, η - stand off ratio of UJT.

- When $V_C \geq V_P$, the UJT **FIRES**. The capacitor starts discharging though $R_1 + R_{B1}$

Due to design of R_1 , the discharge is fast and produces a pulse across R_1 .

- When $V_C = V_E = V_V$ the UJT turns **OFF**. The capacitor starts charging again.

Waveforms (3%)

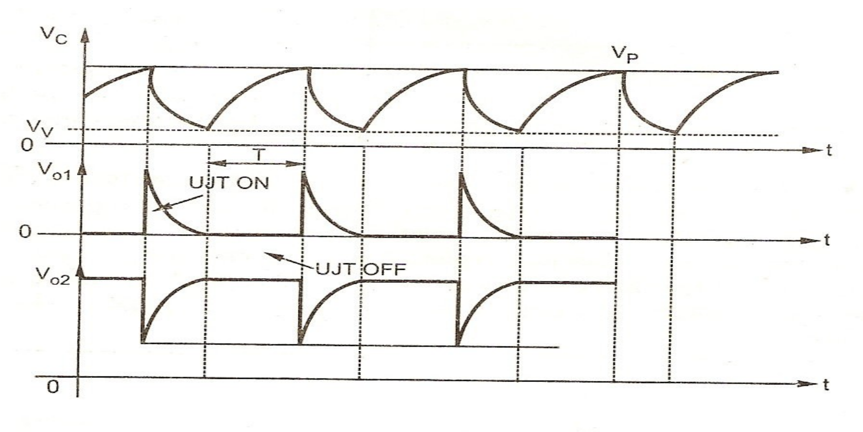


Fig. 5.89 Waveforms of UJT sawtooth oscillator

The discharge time of the pulse is controlled by $C_T \ R_1$ while the charging time is controlled by $C_T \ R_T$

C. TUTORIALS

Text Books;

1. Electronic Circuits – II Semester – IV (ECE)
   Second Revised Edition (Technical Publications Pune)
   AP Godse & UA Bakshi
2. David A Bell 2002, “Solid state Pulse Circuits” - Prentice Hall of India
3. John D Ryder, 1999, “Electronic Fundamental and Applications – integrated and Discrete Systems” - Prentice Hall of India

Reference Books;

Millman J and Taub H 2001 “Pulse Digital and Switching wave form” - McGraw Hill