DEPARTMENT : ELECTRICAL AND ELECTRONICS ENGINEERING

SUBJECT : EC1209 – ELECTRON DEVICES AND CIRCUITS

YEAR : II

SEMESTER : III

ANNA UNIVERSITY TIRUCHIRAPPALLI
Tiruchirappalli – 620 024

Regulations 2008

Curriculum

B.E. ELECTRICAL AND ELECTRONICS ENGINEERING

SEMESTER III

S. No.	Subject Code	Subject	L	T	P	C
Theory						
1	MA1201	Transforms and Partial Differential Equations	3	1	0	4
2	EI1202	Measurements and Instrumentation	3	0	0	3
3	EE1201	Electromagnetic Theory	3	1	0	4
4	HS1201	Environmental Science and Engineering	3	0	0	3
5	EC1209	Electron Devices and Circuits	3	0	0	3
6	CS1201	Data Structures	3	0	0	3
Practical						
7	EC1210	Electron Devices and Circuits Laboratory	0	0	3	2
8	CS1203	Data Structures Laboratory	0	0	3	2
9	EI1203	Measurements and Instrumentation Laboratory	0	0	3	2
					Total	**26**

EC1209 – ELECTRON DEVICES AND CIRCUITS

UNIT I SEMICONDUCTOR DIODE AND BJT

PN Junction – Current components in a PN diode – Junction capacitance – Junction diode switching time – Zener diode – Varactor diode – Tunnel diode – Schottky diode – Transistor Structure – Basic Transistor operation – Transistor characteristics and parameters – Transistor as a switch and amplifier – Transistor bias circuit – Voltage divider bias circuits – Base bias circuits – Emitter bias circuits – Collector feedback bias circuits – DC load line – AC load line – Bias stabilization – Thermal runaway and thermal stability.

UNIT II FET, UJT and SCR

JFET characteristics and parameters – JFET biasing – Self bias – Voltage divider bias – Q point – Stability over temperature – MOSFET – D-MOSFET and E-MOSFET – MOSFET characteristics and parameters – MOSFET biasing – Zero bias – Voltage divider bias – Drain feedback bias – Characteristics and applications of UJT, SCR, DIAC, TRIAC.

UNIT III AMPLIFIERS

CE, CC and CB amplifiers – Small-signal low frequency transistor amplifier circuits – h-parameter representation of a transistor – Analysis of single stage transistor amplifier circuits – Voltage gain – Current gain – Input impedance and output impedance – Frequency response – RC coupled amplifier – Classification of Power amplifiers – Class A, B, AB and C Power amplifiers – Push-Pull and Complementary-Symmetry amplifiers – Design of power output, efficiency and cross-over distortion.

UNIT IV FEEDBACK AMPLIFIERS AND OSCILLATORS

Advantages of negative feedback – Voltage/current, series/shunt feedback – Positive feedback – Conditions for oscillation – Phase shift – Wein Bridge – Hartley – Colpitts and Crystal oscillators.

UNIT V PULSE CIRCUITS AND POWER SUPPLY

RC wave shaping circuits – Diode clampers and clippers – Multivibrators – Schmitt triggers – UJT saw-tooth oscillators – Single and poly-phase rectifiers and analysis of filter circuits – Design of zener and transistor series voltage regulators – Switched mode power supplies.

Total: 45

TEXT BOOKS

1. Robert T. Paynter, "Introductory Electronic Devices and Circuits", 7th Edition**,** Pearson Education, 2006.

2. Millman and Halkias, "Electronic Devices and Circuits", Tata McGraw Hill, 2007.

REFERENCES

1. Mottershead, A., "Electronic Devices and Circuits an Introduction", Prentice Hall of India, 2003.

2. Boylsted and Nashelsky, "Electronic Devices and Circuit Theory", Prentice Hall of India, 6th Edition, 1999.

3. Bell, D.A., "Electronic Devices and Circuits", Oxford University Press, 4th Edition, 1999.

UNIT I

SEMICONDUCTOR DIODE AND BJT

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UNIT I

SEMICONDUCTOR DIODE AND BJT

Semiconductor diodes

A modern semiconductor diode is made of a crystal of semiconductor like silicon that has impurities added to it to create a region on one side that contains negative charge carriers (electrons), called n-type semiconductor, and a region on the other side that contains positive charge carriers (holes), called p-type semiconductor. The diode's terminals are attached to each of these regions. The boundary within the crystal between these two regions, called a PN junction, is where the action of the diode takes place. The crystal conducts conventional current in a direction from the p-type side (called the anode) to the n-type side (called the cathode), but not in the opposite direction.

Another type of semiconductor diode, the Schottky diode, is formed from the contact between a metal and a semiconductor rather than by a p-n junction.

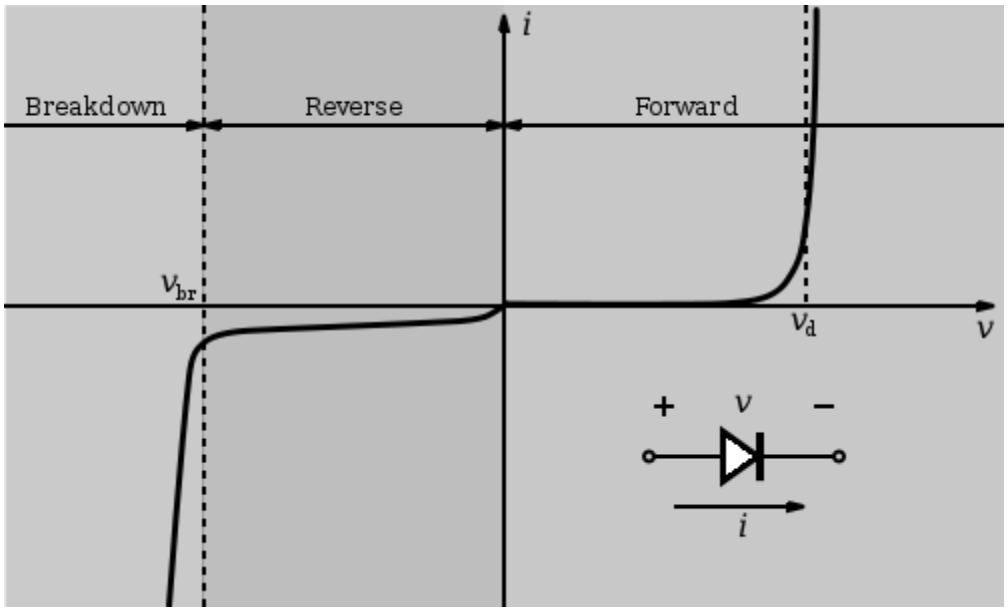
Current–voltage characteristic

A semiconductor diode's behavior in a circuit is given by its current–voltage characteristic, or I–V graph (see graph below). The shape of the curve is determined by the transport of charge carriers through the so-called depletion layer or depletion region that exists at the p-n junction between differing semiconductors. When a p-n junction is first created, conduction band (mobile) electrons from the N-doped region diffuse into the P-doped region where there is a large population of holes (vacant places for electrons) with which the electrons "recombine". When a mobile electron recombines with a hole, both hole and electron vanish, leaving behind an immobile positively charged donor (dopant) on the N-side and negatively charged acceptor (dopant) on the P-side. The region around the p-n junction becomes depleted of charge carriers and thus behaves as an insulator.

However, the width of the depletion region (called the depletion width) cannot grow without limit. For each electron-hole pair that recombines, a positively charged dopant ion is left behind in the N-doped region, and a negatively charged dopant ion is left behind in the P-doped region. As recombination proceeds more ions are created, an increasing electric field develops through the depletion zone which acts to slow and then finally stop recombination. At this point, there is a "built-in" potential across the depletion zone.

If an external voltage is placed across the diode with the same polarity as the built-in potential, the depletion zone continues to act as an insulator, preventing any significant electric current flow (unless electron/hole pairs are actively being created in the junction by, for instance, light. see photodiode). This is the reverse *bias* phenomenon. However, if the polarity of the external voltage opposes the built-in potential, recombination can once again proceed, resulting in substantial electric current through the p-n junction (i.e. substantial numbers of electrons and holes recombine at the junction). For silicon diodes, the built-in potential is approximately 0.7 V (0.3 V for Germanium and 0.2 V for Schottky). Thus, if an external current is passed through the diode, about 0.7 V will be developed across the diode such that the P-doped region is positive with respect to the N-doped region and the diode is said to be "turned on" as it has a forward *bias*.

A diode's ***'I–V characteristic'*** can be approximated by four regions of operation.



I–V characteristics of a P-N junction diode

At very large reverse bias, beyond the peak inverse voltage or PIV, a process called reverse breakdown occurs which causes a large increase in current (i.e. a large number of electrons and holes are created at, and move away from the pn junction) that usually damages the device permanently. The avalanche diode is deliberately designed for use in the avalanche region. In the zener diode, the concept of PIV is not applicable. A zener diode contains a heavily doped p-n junction allowing electrons to tunnel from the valence band of the p-type material to the conduction band of the n-type material, such that the reverse voltage is "clamped" to a known value (called the *zener voltage*), and avalanche does not occur. Both devices, however, do have a limit to the maximum current and power in the clamped reverse voltage region. Also, following the end of forward conduction in any diode, there is reverse current for a short time. The device does not attain its full blocking capability until the reverse current ceases.

The second region, at reverse biases more positive than the PIV, has only a very small reverse saturation current. In the reverse bias region for a normal P-N rectifier diode, the current through the device is very low (in the μA range). However, this is temperature dependent, and at sufficiently high temperatures, a substantial amount of reverse current can be observed (mA or more).

The third region is forward but small bias, where only a small forward current is conducted.

As the potential difference is increased above an arbitrarily defined "cut-in voltage" or "on-voltage" or "diode forward voltage drop (V_d)", the diode current becomes appreciable (the level of current considered "appreciable" and the value of cut-in voltage depends on the application), and the diode

presents a very low resistance. The current–voltage curve is exponential. In a normal silicon diode at rated currents, the arbitrary "cut-in" voltage is defined as 0.6 to 0.7 volts. The value is different for other diode types — Schottky diodes can be rated as low as 0.2 V, Germanium diodes 0.25-0.3 V, and red or blue light-emitting diodes (LEDs) can have values of 1.4 V and 4.0 V respectively.

At higher currents the forward voltage drop of the diode increases. A drop of 1 V to 1.5 V is typical at full rated current for power diodes.

Shockley diode equation

The *Shockley* ideal diode equation or *the diode law* (named after transistor co-inventor William Bradford Shockley, not to be confused with tetrode inventor Walter H. Schottky) gives the I–V characteristic of an ideal diode in either forward or reverse bias (or no bias). The equation is:

$$I = I_S \left(e^{V_D/(nV_T)} - 1\right),$$

where

I is the diode current,

I_S is the reverse bias saturation current (or scale current),

V_D is the voltage across the diode,

V_T is the thermal voltage, and

n is the *ideality factor*, also known as the *quality factor* or sometimes *emission coefficient*. The ideality factor n varies from 1 to 2

depending on the fabrication process and semiconductor material and in many cases is assumed to be approximately equal to 1 (thus the notation *n* is omitted).

The thermal voltage V_T is approximately 25.85 mV at 300 K, a temperature close to "room temperature" commonly used in device simulation software. At any temperature it is a known constant defined by:

$$V_T = \frac{kT}{q},$$

where *k* is the Boltzmann constant, *T* is the absolute temperature of the p-n junction, and *q* is the magnitude of charge on an electron (the elementary charge).

The *Shockley ideal diode equation* or the *diode law* is derived with the assumption that the only processes giving rise to the current in the diode are drift (due to electrical field), diffusion, and thermal recombination-generation. It also assumes that the recombination-generation (R-G) current in the depletion region is insignificant. This means that the Shockley equation doesn't account for the processes involved in reverse breakdown and photon-assisted R-G. Additionally, it doesn't describe the "leveling off" of the I–V curve at high forward bias due to internal resistance.

Under *reverse bias* voltages (see Figure 5) the exponential in the diode equation is negligible, and the current is a constant (negative) reverse current value of $-I_S$. The reverse *breakdown region* is not modeled by the Shockley diode equation.

For even rather small *forward bias* voltages (see Figure 5) the exponential is very large because the thermal voltage is very small, so the subtracted '1' in

the diode equation is negligible and the forward diode current is often approximated as

$$I = I_S e^{V_D/(nV_T)}$$

The use of the diode equation in circuit problems is illustrated in the article on diode modeling.

Several types of junction diodes

There are several types of junction diodes, which either emphasize a different physical aspect of a diode often by geometric scaling, doping level, choosing the right electrodes, are just an application of a diode in a special circuit, or are really different devices like the Gunn and laser diode and the MOSFET:

Normal (p-n) diodes, which operate as described above, are usually made of doped silicon or, more rarely, germanium. Before the development of modern silicon power rectifier diodes, cuprous oxide and later selenium was used; its low efficiency gave it a much higher forward voltage drop (typically 1.4–1.7 V per "cell", with multiple cells stacked to increase the peak inverse voltage rating in high voltage rectifiers), and required a large heat sink (often an extension of the diode's metal substrate), much larger than a silicon diode of the same current ratings would require. The vast majority of all diodes are the p-n diodes found inCMOS integrated circuits, which include two diodes per pin and many other internal diodes.

Avalanche diodes

Diodes that conduct in the reverse direction when the reverse bias voltage exceeds the breakdown voltage. These are electrically very similar to Zener diodes, and are often mistakenly called Zener diodes, but break down by a

different mechanism, the *avalanche effect*. This occurs when the reverse electric field across the p-n junction causes a wave of ionization, reminiscent of an avalanche, leading to a large current. Avalanche diodes are designed to break down at a well-defined reverse voltage without being destroyed. The difference between the avalanche diode (which has a reverse breakdown above about 6.2 V) and the Zener is that the channel length of the former exceeds the "mean free path" of the electrons, so there are collisions between them on the way out. The only practical difference is that the two types have temperature coefficients of opposite polarities.

Constant current diodes

These are actually a JFET with the gate shorted to the source, and function like a two-terminal current-limiter analog to the Zener diode, which is limiting voltage. They allow a current through them to rise to a certain value, and then level off at a specific value. Also called *CLDs*, *constant-current diodes*, *diode-connected transistors*, or *current-regulating diodes*.

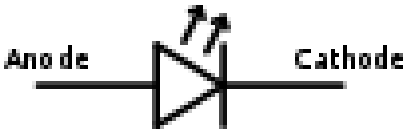
Esaki or tunnel diodes

These have a region of operation showing negative resistance caused by quantum tunneling, thus allowing amplification of signals and very simple bistable circuits. These diodes are also the type most resistant to nuclear radiation.

Gunn diodes

These are similar to tunnel diodes in that they are made of materials such as GaAs or InP that exhibit a region of negative differential resistance. With appropriate biasing, dipole domains form and travel across the diode, allowing high frequency microwave oscillators to be built.

Light-emitting diodes (LEDs)



In a diode formed from a direct band-gap semiconductor, such as gallium arsenide, carriers that cross the junction emit photons when they recombine with the majority carrier on the other side. Depending on the material, wavelengths (or colors)[11] from the infrared to the near ultraviolet may be produced.[12] The forward potential of these diodes depends on the wavelength of the emitted photons: 1.2 V corresponds to red, 2.4 V to violet. The first LEDs were red and yellow, and higher-frequency diodes have been developed over time. All LEDs produce incoherent, narrow-spectrum light; "white" LEDs are actually combinations of three LEDs of a different color, or a blue LED with a yellow scintillator coating. LEDs can also be used as low-efficiency photodiodes in signal applications. An LED may be paired with a photodiode or phototransistor in the same package, to form an opto-isolator.

Laser diodes

When an LED-like structure is contained in a resonant cavity formed by polishing the parallel end faces, a laser can be formed. Laser diodes are commonly used in optical storage devices and for high speed optical communication.

Thermal diodes

This term is used both for conventional PN diodes used to monitor temperature due to their varying forward voltage with temperature, and for Peltier heat pumps for thermoelectric heating and cooling.. Peltier heat

pumps may be made from semiconductor, though they do not have any rectifying junctions, they use the differing behaviour of charge carriers in N and P type semiconductor to move heat.

Photodiodes

All semiconductors are subject to optical charge carrier generation. This is typically an undesired effect, so most semiconductors are packaged in light blocking material. Photodiodes are intended to sense light(photodetector), so they are packaged in materials that allow light to pass, and are usually PIN (the kind of diode most sensitive to light).[13] A photodiode can be used in solar cells, in photometry, or in optical communications. Multiple photodiodes may be packaged in a single device, either as a linear array or as a two-dimensional array. These arrays should not be confused with charge-coupled devices.

Point-contact diodes

These work the same as the junction semiconductor diodes described above, but their construction is simpler. A block of n-type semiconductor is built, and a conducting sharp-point contact made with some group-3 metal is placed in contact with the semiconductor. Some metal migrates into the semiconductor to make a small region of p-type semiconductor near the contact. The long-popular 1N34 germanium version is still used in radio receivers as a detector and occasionally in specialized analog electronics.

PIN diodes

A PIN diode has a central un-doped, or *intrinsic*, layer, forming a p-type/intrinsic/n-type structure.[14] They are used as radio frequency switches and attenuators. They are also used as large volume ionizing radiation detectors and as photodetectors. PIN diodes are also used in power electronics, as their central layer can withstand high voltages. Furthermore, the PIN structure can be found in many power semiconductor devices, such as IGBTs, power MOSFETs, and thyristors.

Schottky diodes

Schottky diodes are constructed from a metal to semiconductor contact. They have a lower forward voltage drop than p-n junction diodes. Their forward voltage drop at forward currents of about 1 mA is in the range 0.15 V to 0.45 V, which makes them useful in voltage clamping applications and prevention of transistor saturation. They can also be used as low loss rectifiers although their reverse leakage current is generally higher than that of other diodes. Schottky diodes are majority carrier devices and so do not suffer from minority carrier storage problems that slow down many other diodes — so they have a faster "reverse recovery" than p-n junction diodes. They also tend to have much lower junction capacitance than p-n diodes which provides for high switching speeds and their use in high-speed circuitry and RF devices such as switched-mode power

varactor diodes

Varactors are operated reverse-biased so no current flows, but since the thickness of the depletion zone varies with the applied bias voltage, the capacitance of the diode can be made to vary. Generally, the depletion region thickness is proportional to the square root of the applied voltage; andcapacitance is inversely proportional to the depletion region thickness. Thus, the capacitance is inversely proportional to the square root of applied voltage.

All diodes exhibit this phenomenon to some degree, but specially made varactor diodes exploit the effect to boost the capacitance and variability range achieved - most diode fabrication attempts to achieve the opposite.

In the figure we can see an example of a crossection of a varactor with the depletion layer formed of a p-n-junction. But the depletion layer can also be made of a MOS-diode or a Schottky diode. This is very important in CMOS and MMIC technology.

Zener diodes

Diodes that can be made to conduct backwards. This effect, called Zener breakdown, occurs at a precisely defined voltage, allowing the diode to be used as a precision voltage reference. In practical voltage reference circuits Zener and switching diodes are connected in series and opposite directions to balance the temperature coefficient to near zero. Some devices labeled as high-voltage Zener diodes are actually avalanche diodes (see above). Two (equivalent) Zeners in series and in reverse order, in the same package, constitute a transient absorber (or Transorb, a registered trademark). The Zener diode is named for Dr. Clarence Melvin Zener of Carnegie Mellon University, inventor of the device.

Other uses for semiconductor diodes include sensing temperature, and computing analog logarithms (see Operational amplifier applications#Logarithmic).

Zener diode is a type of diode that permits current not only in the forward direction like a normal diode, but also in the reverse direction if the voltage is larger than the breakdown voltage known as "Zener knee voltage" or "Zener voltage". The device was named after Clarence Zener, who discovered this electrical property.

A conventional solid-state diode will not allow significant current if it is reverse below its reverse breakdown voltage. When the reverse bias breakdown voltage is exceeded, a conventional diode is subject to high current due to avalanche breakdown. Unless this current is limited by circuitry, the diode will be permanently damaged. In case of large forward bias (current in the direction of the arrow), the diode exhibits a voltage drop due to its junction built-in voltage and internal resistance. The amount of the voltage drop depends on the semiconductor material and the doping concentrations.

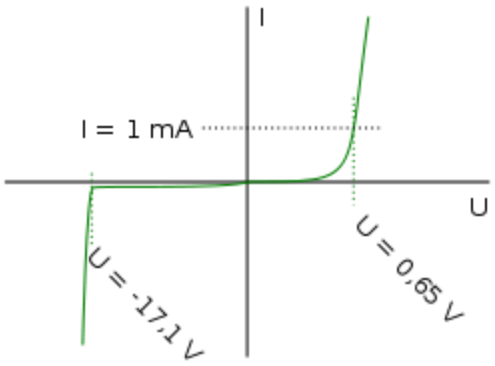
A Zener diode exhibits almost the same properties, except the device is specially designed so as to have a greatly reduced breakdown voltage, the so-called Zener voltage. By contrast with the conventional device, a reverse-biased Zener diode will exhibit a controlled breakdown and allow the current to keep the voltage across the Zener diode at the Zener voltage. For example, a diode with a Zener breakdown voltage of 3.2 V will exhibit a voltage drop of 3.2 V even if reverse bias voltage applied across it is more than its Zener voltage. The Zener diode is therefore ideal for applications such as the generation of a reference voltage (e.g. for an amplifier stage), or as a voltage stabilizer for low-current applications.

The Zener diode's operation depends on the heavy doping of its p-n junction allowing electrons to tunnel from the valence band of the p-type

material to the conduction band of the n-type material. In the atomic scale, this tunneling corresponds to the transport of valence band electrons into the empty conduction band states; as a result of the reduced barrier between these bands and high electric fields that are induced due to the relatively high levels of dopings on both sides.[1] The breakdown voltage can be controlled quite accurately in the doping process. While tolerances within 0.05% are available, the most widely used tolerances are 5% and 10%. Breakdown voltage for commonly available zener diodes can vary widely from 1.2 volts to 200 volts.

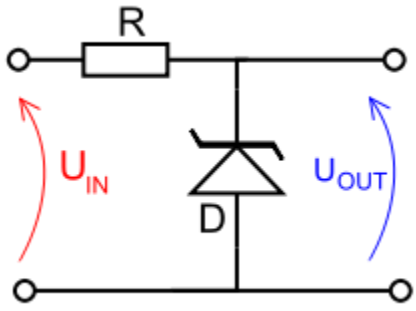
Another mechanism that produces a similar effect is the avalanche effect as in the avalanche diode. The two types of diode are in fact constructed the same way and both effects are present in diodes of this type. In silicon diodes up to about 5.6 volts, the Zener effect is the predominant effect and shows a marked negative temperature coefficient. Above 5.6 volts, the avalanche effect becomes predominant and exhibits a positive temperature coefficient[1]. In a 5.6 V diode, the two effects occur together and their temperature coefficients neatly cancel each other out, thus the 5.6 V diode is the component of choice in temperature-critical applications. Modern manufacturing techniques have produced devices with voltages lower than 5.6 V with negligible temperature coefficients, but as higher voltage devices are encountered, the temperature coefficient rises dramatically. A 75 V diode has 10 times the coefficient of a 12 V diode.

All such diodes, regardless of breakdown voltage, are usually marketed under the umbrella term of "Zener diode".



Current-voltage characteristic of a Zener diode with a breakdown voltage of 17 volt. Notice the change of voltage scale between the forward biased (positive) direction and the reverse biased (negative) direction.

Zener diodes are widely used as voltage references and as shunt regulators to regulate the voltage across small circuits. When connected in parallel with a variable voltage source so that it is reverse biased, a Zener diode conducts when the voltage reaches the diode's reverse breakdown voltage. From that point on, the relatively low impedance of the diode keeps the voltage across the diode at that value.



In this circuit, a typical voltage reference or regulator, an input voltage, U_{IN}, is regulated down to a stable output voltage U_{OUT}. The intrinsic voltage drop of diode D is stable over a wide current range and holds U_{OUT} relatively constant even though the input voltage may fluctuate over a fairly wide

range. Because of the low impedance of the diode when operated like this, Resistor R is used to limit current through the circuit.

In the case of this simple reference, the current flowing in the diode is determined using Ohms law and the known voltage drop across the resistor R. $I_{Diode} = (U_{IN} - U_{OUT}) / R_{\Omega}$

The value of *R* must satisfy two conditions:

1. *R* must be small enough that the current through D keeps D in reverse breakdown. The value of this current is given in the data sheet for D. For example, the common BZX79C5V6[2] device, a 5.6 V 0.5 W Zener diode, has a recommended reverse current of 5 mA. If insufficient current exists through D, then U_{OUT} will be unregulated, and less than the nominal breakdown voltage (this differs to voltage regulator tubes where the output voltage will be higher than nominal and could rise as high as U_{IN}). When calculating *R*, allowance must be made for any current through the external load, not shown in this diagram, connected across U_{OUT}.
2. *R* must be large enough that the current through D does not destroy the device. If the current through D is I_D, its breakdown voltage V_B and its maximum power dissipation P_{MAX}, then $I_D V_B < P_{MAX}$.

A load may be placed across the diode in this reference circuit, and as long as the zener stays in reverse breakdown, the diode will provide a stable voltage source to the load.

Shunt regulators are simple, but the requirements that the ballast resistor be

small enough to avoid excessive voltage drop during worst-case operation (low input voltage concurrent with high load current) tends to leave a lot of current flowing in the diode much of the time, making for a fairly wasteful regulator with high quiescent power dissipation, only suitable for smaller loads.

Zener diodes in this configuration are often used as stable references for more advanced voltage regulator circuits.

These devices are also encountered, typically in series with a base-emitter junction, in transistor stages where selective choice of a device centered around the avalanche/Zener point can be used to introduce compensating temperature co-efficient balancing of the transistor PN junction. An example of this kind of use would be a DC error amplifier used in a regulated power supply circuit feedback loop system.

Zener diodes are also used in surge protectors to limit transient voltage spikes.

Another notable application of the zener diode is the use of noise caused by its avalanche breakdown in a random number generator that never repeats.

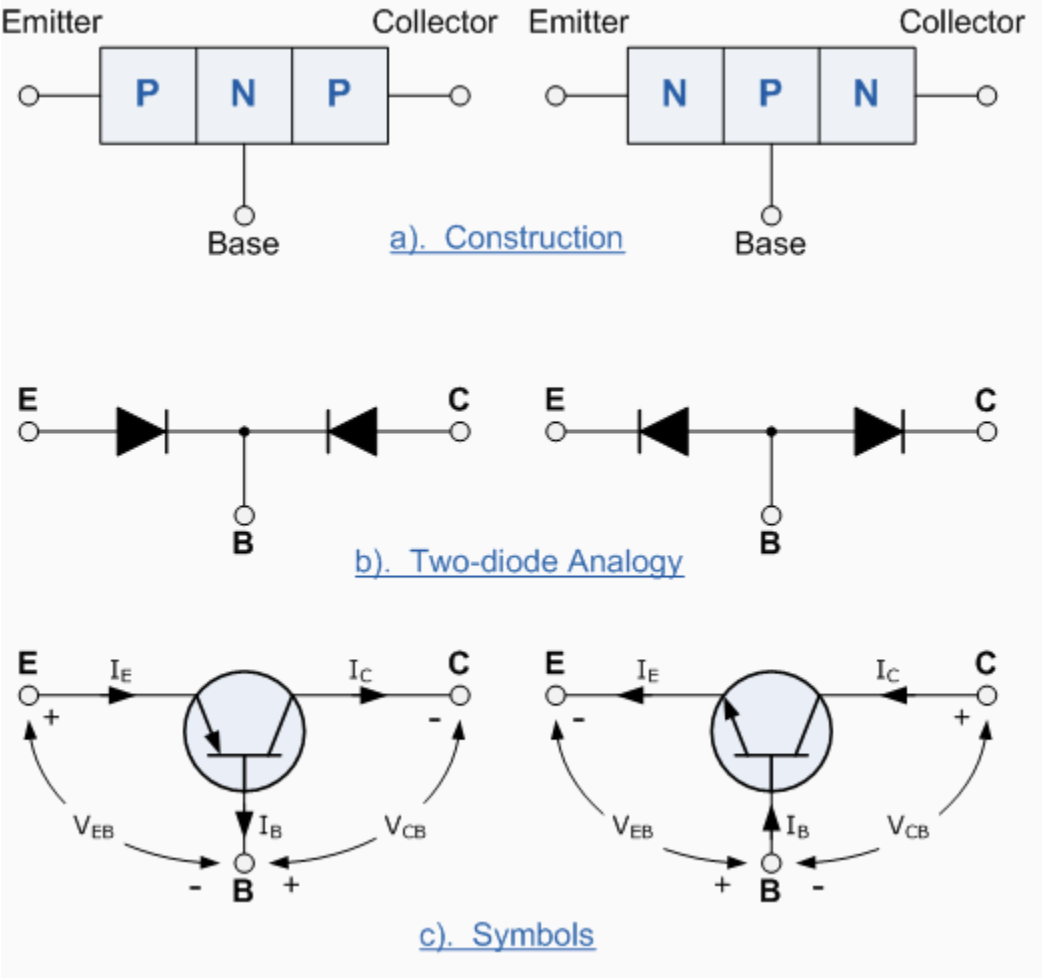
Bipolar Transistor Basics

In the *Diode* tutorials we saw that simple diodes are made up from two pieces of semiconductor material, either Silicon or Germanium to form a simple PN-junction and we also learnt about their properties and characteristics. If we now join together two individual diodes end to end giving two PN-junctions connected together in series, we now have a three layer, two junction, three terminal device forming the basis of a **Bipolar Junction Transistor**, or **BJT** for short. This type of transistor is generally

known as a **Bipolar Transistor**, because its basic construction consists of two PN-junctions with each terminal or connection being given a name to identify it and these are known as the Emitter, Base and Collector respectively.

The word Transistor is an acronym, and is a combination of the words Transfer Varistor used to describe their mode of operation way back in their early days of development. There are two basic types of bipolar transistor construction, NPN and PNP, which basically describes the physical arrangement of the P-type and N-type semiconductor materials from which they are made. Bipolar Transistors are "CURRENT" Amplifying or current regulating devices that control the amount of current flowing through them in proportion to the amount of biasing current applied to their base terminal. The principle of operation of the two transistor types NPN and PNP, is exactly the same the only difference being in the biasing (base current) and the polarity of the power supply for each type.

Bipolar Transistor Construction



The construction and circuit symbols for both the NPN and PNP bipolar transistor are shown above with the arrow in the circuit symbol always showing the direction of conventional current flow between the base terminal and its emitter terminal, with the direction of the arrow pointing from the positive P-type region to the negative N-type region, exactly the same as for the standard diode symbol.

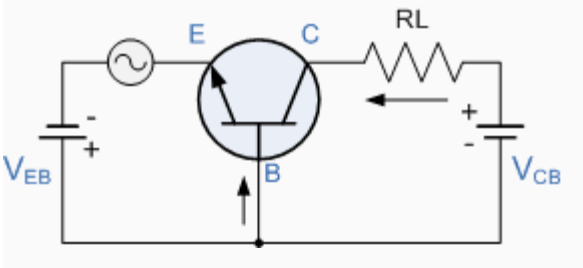
There are basically three possible ways to connect a **Bipolar Transistor** within an electronic circuit with each method of connection responding differently to its input signal as the static characteristics of the transistor vary with each circuit arrangement.

- 1. Common Base Configuration - has Voltage Gain but no Current Gain.
- 2. Common Emitter Configuration - has both Current and Voltage Gain.
- 3. Common Collector Configuration - has Current Gain but no Voltage Gain.

The Common Base Configuration.

As its name suggests, in the **Common Base** or Grounded Base configuration, the BASE connection is common to both the input signal AND the output signal with the input signal being applied between the base and the emitter terminals. The corresponding output signal is taken from between the base and the collector terminals as shown with the base terminal grounded or connected to a fixed reference voltage point. The input current flowing into the emitter is quite large as its the sum of both the base current and collector current respectively therefore, the collector current output is less than the emitter current input resulting in a Current Gain for this type of circuit of less than "1", or in other words it "Attenuates" the signal.

The Common Base Amplifier Circuit



This type of amplifier configuration is a non-inverting voltage amplifier circuit, in that the signal voltages Vin and Vout are In-Phase. This type of arrangement is not very common due to its unusually high voltage gain characteristics. Its Output characteristics represent that of a forward biased diode while the Input characteristics represent that of an illuminated photo-diode. Also this type of configuration has a high ratio of Output to Input resistance or more importantly "Load" resistance (RL) to "Input" resistance (Rin) giving it a value of "Resistance Gain". Then the Voltage Gain for a common base can therefore be given as:

Common Base Voltage Gain

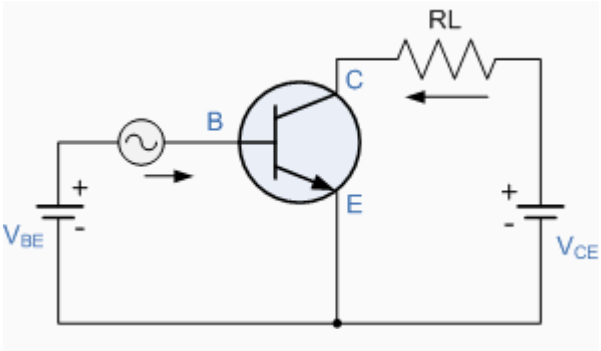
$$A_V = \frac{I_c \times R_L}{I_e \times R_{IN}} = \alpha \times \frac{R_L}{R_{IN}}$$

The Common Base circuit is generally only used in single stage amplifier circuits such as microphone pre-amplifier or RF radio amplifiers due to its very good high frequency response.

The Common Emitter Configuration.

In the **Common Emitter** or Grounded Emitter configuration, the input signal is applied between the base, while the output is taken from between the collector and the emitter as shown. This type of configuration is the most commonly used circuit for transistor based amplifiers and which represents the "normal" method of connection. The common emitter amplifier configuration produces the highest current and power gain of all the three bipolar transistor configurations. This is mainly because the input impedance is LOW as it is connected to a forward-biased junction, while the output impedance is HIGH as it is taken from a reverse-biased junction.

The Common Emitter Amplifier Circuit



In this type of configuration, the current flowing out of the transistor must be equal to the currents flowing into the transistor as the emitter current is given as Ie = Ic + Ib. Also, as the load resistance (RL) is connected in series with the collector, the Current gain of the Common Emitter Transistor Amplifier is quite large as it is the ratio of Ic/Ib and is given the symbol of Beta, (β). Since the relationship between these three currents is determined by the transistor itself, any small change in the base current will result in a large change in the collector current. Then, small changes in base current will thus control the current in the Emitter/Collector circuit.

By combining the expressions for both Alpha, α and Beta, β the mathematical relationship between these parameters and therefore the current gain of the amplifier can be given as:

$$I_E = I_C + I_B$$

$$\alpha = \frac{I_C}{I_E} \quad \text{and} \quad \beta = \frac{I_C}{I_B}$$

$$\alpha = \frac{\beta}{\beta + 1} \qquad \beta = \frac{\alpha}{1 - \alpha}$$

Where: "Ic" is the current flowing into the collector terminal, "Ib" is the current flowing into the base terminal and "Ie" is the current flowing out of the emitter terminal.

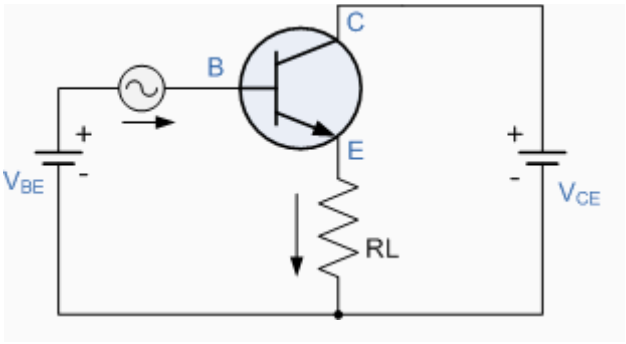
Then to summarise, this type of bipolar transistor configuration has a greater input impedance, Current and Power gain than that of the common Base configuration but its Voltage gain is much lower. The common emitter is an inverting amplifier circuit resulting in the output signal being 180^o out of phase with the input voltage signal.

The Common Collector Configuration.

In the **Common Collector** or Grounded Collector configuration, the collector is now common and the input signal is connected to the Base, while the output is taken from the Emitter load as shown. This type of configuration is commonly known as a **Voltage Follower** or **Emitter Follower** circuit. The Emitter follower configuration is very useful for

impedance matching applications because of the very high input impedance, in the region of hundreds of thousands of Ohms, and it has relatively low output impedance.

The Common Collector Amplifier Circuit



The Common Emitter configuration has a current gain equal to the β value of the transistor itself. In the common collector configuration the load resistance is situated in series with the emitter so its current is equal to that of the emitter current. As the emitter current is the combination of the collector AND base currents combined, the load resistance in this type of amplifier configuration also has both the collector current and the input current of the base flowing through it. Then the current gain of the circuit is given as:

$$I_E = I_C + I_B$$

$$A_i = \frac{I_E}{I_B} = \frac{I_C + I_B}{I_B}$$

$$A_i = \frac{I_C}{I_B} + 1$$

$$\therefore A_i = \beta + 1$$

This type of bipolar transistor configuration is a non-inverting amplifier circuit in that the signal voltages of Vin and Vout are "In-Phase". It has a voltage gain that is always less than "1" (unity). The load resistance of the common collector amplifier configuration receives both the base and collector currents giving a large current gain (as with the Common Emitter configuration) therefore, providing good current amplification with very little voltage gain.

Bipolar Transistor Summary.

The behaviour of the bipolar transistor in each one of the above circuit configurations is very different and produces different circuit characteristics with regards to Input impedance, Output impedance and Gain and this is summarised in the table below.

Transistor Characteristics

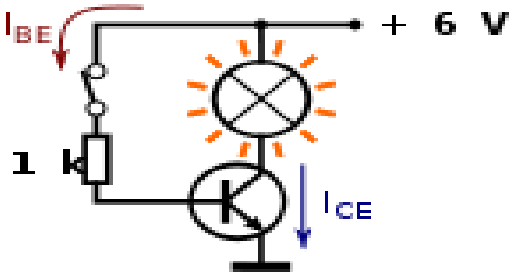
The static characteristics for **Bipolar Transistor** amplifiers can be divided into the following main groups.

Input Characteristics:-	Common Base -	$I_E \div V_{EB}$
	Common Emitter -	$I_B \div V_{BE}$
Output Characteristics:-	Common Base -	$I_C \div V_C$
	Common Emitter -	$I_C \div V_C$
Transfer Characteristics:-	Common Base -	$I_E \div I_C$
	Common Emitter -	$I_B \div I_C$

with the characteristics of the different transistor configurations given in the following table:

Characteristic	Common Base	Common Emitter	Common Collector
Input impedance	Low	Medium	High
Output impedance	Very High	High	Low
Phase Angle	0^o	180^o	0^o
Voltage Gain	High	Medium	Low
Current Gain	Low	Medium	High
Power Gain	Low	Very High	

Transistor as a switch



BJT used as an electronic switch, in grounded-emitter configuration.

Transistors are commonly used as electronic switches, for both high power applications including switched-mode power supplies and low power applications such as logic gates.

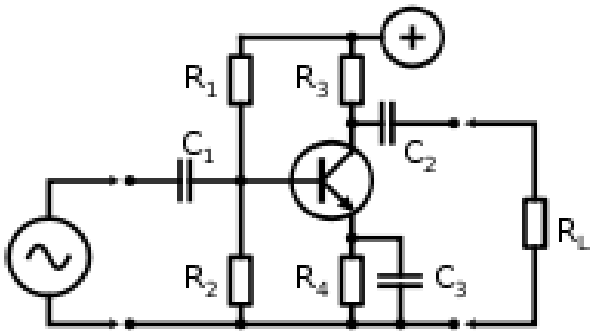
In a grounded-emitter transistor circuit, such as the light-switch circuit shown, as the base voltage rises the base and collector current rise exponentially, and the collector voltage drops because of the collector load resistor. The relevant equations:

$V_{RC} = I_{CE} \times R_C$, the voltage across the load (the lamp with resistance R_C)

$V_{RC} + V_{CE} = V_{CC}$, the supply voltage shown as 6V

If V_{CE} could fall to 0 (perfect closed switch) then Ic could go no higher than V_{CC} / R_C, even with higher base voltage and current. The transistor is then said to be saturated. Hence, values of input voltage can be chosen such that the output is either completely off,[13] or completely on. The transistor is acting as a switch, and this type of operation is common in digital circuits where only "on" and "off" values are relevant.

Transistor as an amplifier



Amplifier circuit, standard common-emitter configuration.

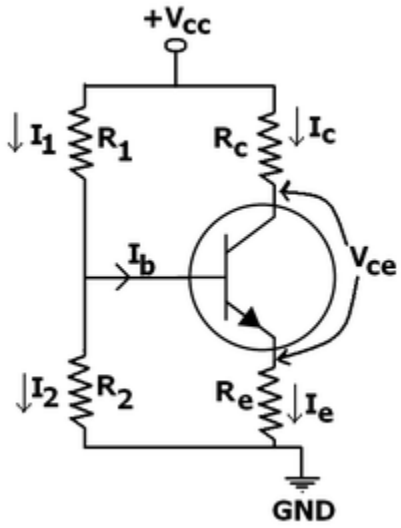
The common-emitter amplifier is designed so that a small change in voltage in (V_{in}) changes the small current through the base of the transistor and the transistor's current amplification combined with the properties of the circuit mean that small swings in V_{in} produce large changes in V_{out}.

Various configurations of single transistor amplifier are possible, with some providing current gain, some voltage gain, and some both.

From mobile phones to televisions, vast numbers of products include amplifiers for sound reproduction, radio transmission, and signal processing. The first discrete transistor audio amplifiers barely supplied a few hundred milliwatts, but power and audio fidelity gradually increased as better transistors became available and amplifier architecture evolved.

Modern transistor audio amplifiers of up to a few hundred watts are common and relatively inexpensive

Voltage divider bias



Voltage divider bias

The voltage divider is formed using external resistors R_1 and R_2. The voltage across R_2 forward biases the emitter junction. By proper selection of resistors R_1 and R_2, the operating point of the transistor can be made independent of β. In this circuit, the voltage divider holds the base voltage fixed independent of base current provided the divider current is large compared to the base current. However, even with a fixed base voltage, collector current varies with temperature (for example) so an emitter resistor is added to stabilize the Q-point, similar to the above circuits with emitter resistor.

In this circuit the base voltage is given by:

$$V_B = \text{voltage across } R_2 = V_{cc}\frac{R_2}{(R_1+R_2)} - I_B\frac{R_1R_2}{(R_1+R_2)}$$

$$\approx V_{cc}\frac{R_2}{(R_1+R_2)} \text{provided } I_B << I_2 = V_B/R_2.$$

Also $V_B = V_{be} + I_E R_E$

For the given circuit,

$$I_B = \frac{\frac{V_{CC}}{1+R_1/R_2} - V_{be}}{(\beta+1)R_E + R_1 \parallel R_2}.$$

Merits:

- Unlike above circuits, only one dc supply is necessary.
- Operating point is almost independent of β variation.
- Operating point stabilized against shift in temperature.

Demerits:

- In this circuit, to keep I_C independent of β the following condition must be met:

$$I_C = \beta I_B = \beta \frac{\frac{V_{CC}}{1+R_1/R_2} - V_{be}}{(\beta+1)R_E + R_1 \parallel R_2} \approx \frac{\frac{V_{CC}}{1+R_1/R_2} - V_{be}}{R_E},$$

which is approximately the case if

$$(\beta+1)R_E >> R_1 \parallel R_2$$

where $R_1 \parallel R_2$ denotes the equivalent resistance of R_1 and R_2 connected in parallel.

- As β-value is fixed for a given transistor, this relation can be satisfied either by keeping R_E fairly large, or making $R_1 \| R_2$ very low.

- If R_E is of large value, high V_{CC} is necessary. This increases cost as well as precautions necessary while handling.
- If $R_1 \parallel R_2$ is low, either R_1 is low, or R_2 is low, or both are low. A low R_1 raises V_B closer to V_C, reducing the available swing in collector voltage, and limiting how large R_C can be made without driving the transistor out of active mode. A low R_2 lowers V_{be}, reducing the allowed collector current. Lowering both resistor values draws more current from the power supply and lowers the input resistance of the amplifier as seen from the base.

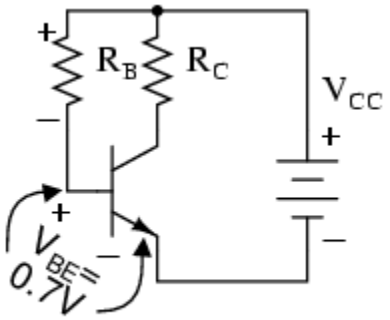
- AC as well as DC feedback is caused by R_E, which reduces the AC voltage gain of the amplifier. A method to avoid AC feedback while retaining DC feedback is discussed below.

Base Bias

The simplest biasing applies a *base-bias* resistor between the base and a base battery V_{BB}. It is convenient to use the existing V_{CC} supply instead of a new bias supply. An example of an audio amplifier stage using base-biasing is "Crystal radio with one transistor . . . " crystal radio, Ch 9 . Note the resistor from the base to the battery terminal. A similar circuit is shown in Figure below.

Write a KVL (Krichhoff's voltage law) equation about the loop containing the battery, R_B, and the V_{BE} diode drop on the transistor in Figure below. Note that we use V_{BB} for the base supply, even though it is actually V_{CC}. If β

is large we can make the approximation that $I_C = I_E$. For silicon transistors $V_{BE} \cong 0.7V$.



$$V_{BB} - I_B R_B - V_{BE} = 0$$

$$V_{BB} - V_{BE} = I_B R_B \qquad \text{(KVL)}$$

$$I_B = \frac{V_{BB} - V_{BE}}{R_B}$$

$$I_E = (\beta+1) I_B \approx \beta I_B$$

$$\mathbf{I_E = \frac{V_{BB} - V_{BE}}{R_B/\beta}} \qquad \text{(IE base-bias)}$$

Base-bias

Silicon small signal transistors typically have a β in the range of 100-300. Assuming that we have a β=100 transistor, what value of base-bias resistor is required to yield an emitter current of 1mA?

Solving the IE base-bias equation for R_B and substituting β, V_{BB}, V_{BE}, and I_E yields 930kΩ. The closest standard value is 910kΩ.

$$\beta = 100 \qquad V_{BB} = 10V \qquad I_C \approx I_E = 1ma$$

$$R_B = \frac{V_{BB} - V_{BE}}{I_E/\beta} = \frac{10 - 0.7}{1mA/100} = 930k$$

What is the the emitter current with a 910kΩ resistor? What is the emitter current if we randomly get a β=300 transistor?

$\beta = 100 \quad V_{BB} = 10V \quad R_B = 910k \quad V_{BE} = 0.7V$

$$I_E = \frac{V_{BB} - V_{BE}}{R_B/\beta} = \frac{10 - 0.7}{910k/100} = 1.02mA$$

$\beta = 300$

$$I_E = \frac{10 - 0.7}{910k/300} = 3.07mA$$

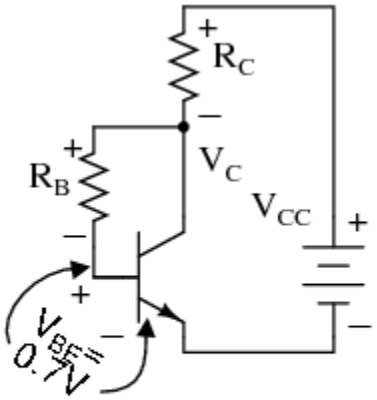
The emitter current is little changed in using the standard value 910kΩ resistor. However, with a change in β from 100 to 300, the emitter current has tripled. This is not acceptable in a power amplifier if we expect the collector voltage to swing from near V_{CC} to near ground. However, for low level signals from micro-volts to a about a volt, the bias point can be centered for a β of square root of (100·300)=173. The bias point will still drift by a considerable amount . However, low level signals will not be clipped.

Base-bias by its self is not suitable for high emitter currents, as used in power amplifiers. The base-biased emitter current is not temperature stable. *Thermal run away* is the result of high emitter current causing a temperature increase which causes an increase in emitter current, which further increases temperature.

Collector-feedback bias

Variations in bias due to temperature and beta may be reduced by moving the V_{BB} end of the base-bias resistor to the collector as in Figure below. If the emitter current were to increase, the voltage drop across R_C increases, decreasing V_C, decreasing I_B fed back to the base. This, in turn, decreases the emitter current, correcting the original increase.

Write a KVL equation about the loop containing the battery, R_C , R_B , and the V_{BE} drop. Substitute $I_C \cong I_E$ and $I_B \cong I_E/\beta$. Solving for I_E yields the IE CFB-bias equation. Solving for I_B yields the IB CFB-bias equation.



$$I_C = \beta I_B \qquad I_C \approx I_E \qquad I_E \approx \beta I_B$$

$$V_{CC} - I_C R_C - I_B R_B - V_{BE} = 0 \qquad \text{(KVL)}$$

$$V_{CC} - I_E R_C - (I_E/\beta) R_B - V_{BE} = 0$$

$$V_{CC} - V_{BE} = I_E R_C + (I_E/\beta) R_B$$

$$V_{CC} - V_{BE} = I_E((R_B/\beta) + R_C)$$

$$\mathbf{I_E = \frac{V_{CC} - V_{BE}}{R_B/\beta + R_C}} \qquad \text{(IE CFB-bias)}$$

$$\mathbf{R_B = \beta\left[\frac{V_{CC} - V_{BE}}{I_E} - R_C\right]} \qquad \text{(RB CFB-bias)}$$

Collector-feedback bias.

Find the required collector feedback bias resistor for an emitter current of 1 mA, a 4.7K collector load resistor, and a transistor with β=100 . Find the collector voltage V_C. It should be approximately midway between V_{CC} and ground.

$$\beta = 100 \qquad V_{CC} = 10V \qquad I_C \approx I_E = 1ma \qquad R_C = 4.7k$$

$$R_B = \beta\left[\frac{V_{CC} - V_{BE}}{I_E} - R_C\right] = 100\left[\frac{10 - 0.7}{1mA} - 4.7k\right] = 460k$$

$$V_C = V_{CC} - I_C R_C = 10 - (1mA)\cdot(4.7k) = 5.3V$$

The closest standard value to the 460k collector feedback bias resistor is 470k. Find the emitter current I_E with the 470 K resistor. Recalculate the emitter current for a transistor with β=100 and β=300.

$\beta = 100 \quad V_{CC} = 10V \quad R_C = 4.7k \quad R_B = 470k$

$$I_E = \frac{V_{CC} - V_{BE}}{R_B/\beta + R_C} = \frac{10 - 0.7}{470k/100 + 4.7k} = 0.989mA$$

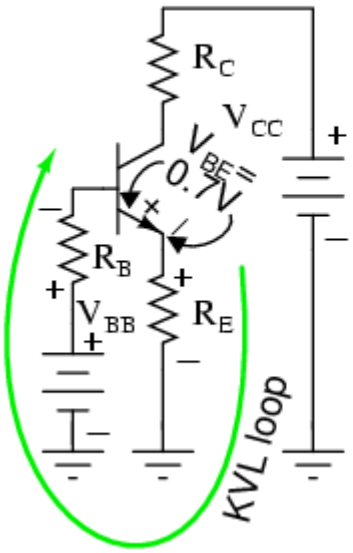
$\beta = 300$

$$I_E = \frac{V_{CC} - V_{BE}}{R_B/\beta + R_C} = \frac{10 - 0.7}{470k/300 + 4.7k} = 1.48mA$$

We see that as beta changes from 100 to 300, the emitter current increases from 0.989mA to 1.48mA. This is an improvement over the previous base-bias circuit which had an increase from 1.02mA to 3.07mA. Collector feedback bias is twice as stable as base-bias with respect to beta variation.

Emitter-bias

Inserting a resistor R_E in the emitter circuit as in Figure below causes *degeneration*, also known as negative feedback. This opposes a change in emitter current I_E due to temperature changes, resistor tolerances, beta variation, or power supply tolerance. Typical tolerances are as follows: resistor— 5%, beta— 100-300, power supply— 5%. Why might the emitter resistor stabilize a change in current? The polarity of the voltage drop across R_E is due to the collector battery V_{CC}. The end of the resistor closest to the (-) battery terminal is (-), the end closest to the (+) terminal it (+). Note that the (-) end of R_E is connected via V_{BB} battery and R_B to the base. Any increase in current flow through R_E will increase the magnitude of negative voltage applied to the base circuit, decreasing the base current, decreasing the emitter current. This decreasing emitter current partially compensates the original increase.



$$V_{BB} - I_B R_B - V_{BE} - I_E R_E = 0$$

$$I_E = (b+1) I_B \approx \beta I_B$$

$$V_{BB} - (I_E/\beta) R_B - V_{BE} - I_E R_E = 0$$

$$V_{BB} - V_{BE} = I_E ((R_B/\beta) + R_E)$$

$$\mathbf{I_E = \frac{V_{BB} - V_{BE}}{R_B/\beta + R_E}} \qquad \text{(IE emitter-bias)}$$

$$R_B/\beta + R_E = \frac{V_{BB} - V_{BE}}{I_E}$$

$$\mathbf{R_B = \beta \left[\frac{V_{BB} - V_{BE}}{I_E} - R_E \right]} \qquad \text{(RB emitter-bias)}$$

Emitter-bias

Note that base-bias battery V_{BB} is used instead of V_{CC} to bias the base in Figure above. Later we will show that the emitter-bias is more effective with a lower base bias battery. Meanwhile, we write the KVL equation for the loop through the base-emitter circuit, paying attention to the polarities on the components. We substitute $I_B \cong I_E/\beta$ and solve for emitter current I_E. This equation can be solved for R_B , equation: RB emitter-bias, Figure above.

Before applying the equations: RB emitter-bias and IE emitter-bias, Figure above, we need to choose values for R_C and R_E . R_C is related to the collector supply V_{CC} and the desired collector current I_C which we assume is approximately the emitter current I_E. Normally the bias point for V_C is set to half of V_{CC}. Though, it could be set higher to compensate for the voltage drop across the emitter resistor R_E. The collector current is whatever we require or choose. It could range from micro-Amps to Amps depending on

the application and transistor rating. We choose I_C = 1mA, typical of a small-signal transistor circuit. We calculate a value for R_C and choose a close standard value. An emitter resistor which is 10-50% of the collector load resistor usually works well.

$$V_C = V_{CC}/2 = 10/2 = 5V$$
$$R_C = Vc/I_C = 5/1mA = 5k \text{ (4.7k standard value)}$$
$$R_E = 0.10R_C = 0.10(4.7K) = 470\Omega$$

Collector-to-base biased bipolar amplifier

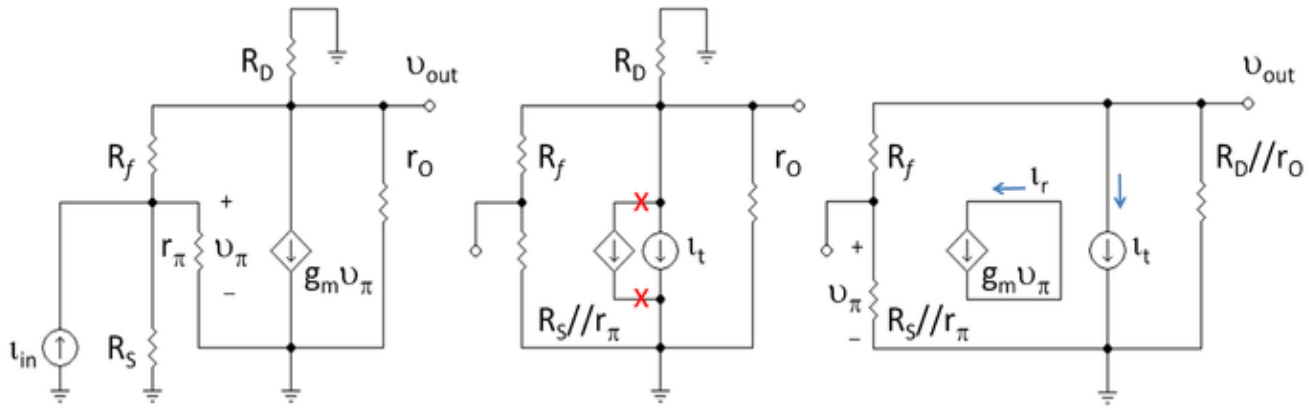


Figure 2: Left - small-signal circuit corresponding to Figure 1; center - inserting independent source and marking leads to be cut; right - cutting the dependent source free and short-circuiting broken leads

Figure 1 (top right) shows a bipolar amplifier with feedback bias resistor R_f driven by a Norton signal source. Figure 2 (left panel) shows the corresponding small-signal circuit obtained by replacing the transistor with its hybrid-pi model. The objective is to find the return ratio of the dependent current source in this amplifier.[9] To reach the objective, the steps outlined

above are followed. Figure 2 (center panel) shows the application of these steps up to Step 4, with the dependent source moved to the left of the inserted source of value i_t, and the leads targeted for cutting marked with an *x*. Figure 2 (right panel) shows the circuit set up for calculation of the return ratio *T*, which is

$$T = -\frac{i_r}{i_t} \ .$$

The return current is

$$i_r = g_m v_\pi \ .$$

The feedback current in R_f is found by current division to be:

$$i_f = \frac{R_D // r_O}{R_D // r_O + R_F + r_\pi // R_S} \ i_t \ .$$

The base-emitter voltage v_π is then, from Ohm's law:

$$v_\pi = -i_f \ (r_\pi // R_S) \ .$$

Consequently,

$$T = g_m (r_\pi // R_S) \ \frac{R_D // r_O}{R_D // r_O + R_F + r_\pi // R_S} \ .$$

Application in asymptotic gain model

The overall transresistance gain of this amplifier can be shown to be:

$$G = \frac{v_{out}}{i_{in}} = \frac{(1 - g_m R_F) R_1 R_2}{R_F + R_1 + R_2 + g_m R_1 R_2} \ ,$$

with $R_1 = R_S \,||\, r_\pi$ and $R_2 = R_D \,||\, r_O$.

This expression can be rewritten in the form used by the asymptotic gain model, which expresses the overall gain of a feedback amplifier in terms of several independent factors that are often more easily derived separately than the overall gain itself, and that often provide insight into the circuit. This form is:

$$G = \ G_\infty \frac{T}{1+T} + G_0 \frac{1}{1+T} \ ,$$

where the so-called **asymptotic gain** G_∞ is the gain at infinite g_m, namely:

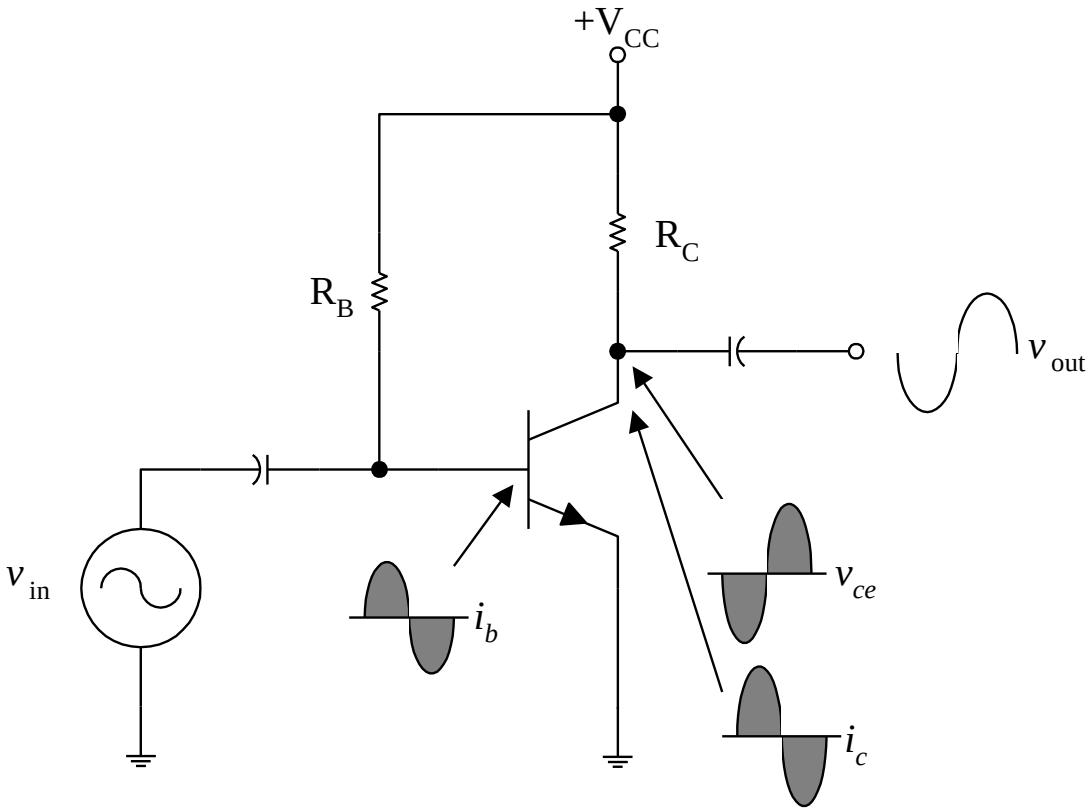
$$G_\infty = -R_F \ ,$$

and the so-called **feed forward** or **direct feedthrough** G_0 is the gain for zero g_m, namely:

$$G_0 = \frac{R_1 R_2}{R_F + R_1 + R_2} \ .$$

For additional applications of this method, see asymptotic gain model.

1. DC Biasing Circuits
2. The **ac** operation of an amplifier depends on the initial **dc** values of *IB*, *IC*, and *VCE*.
3. By varying *IB* around an initial dc value, *IC* and *VCE* are made to vary around their initial dc values.
4. **DC** biasing is a static operation since it deals with setting a **fixed (steady)** level of current (through the device) with a desired fixed voltage drop across the device.



Purpose of the DC biasing circuit

1. To turn the device "ON"
2. To place it in operation in the region of its characteristic where the device operates most linearly, i.e. to set up the initial dc values of *IB*, *IC*, and *VCE*

Voltage-Divider Bias

- The voltage – divider (or potentiometer) bias circuit is by far the most commonly used.

- RB1, RB2

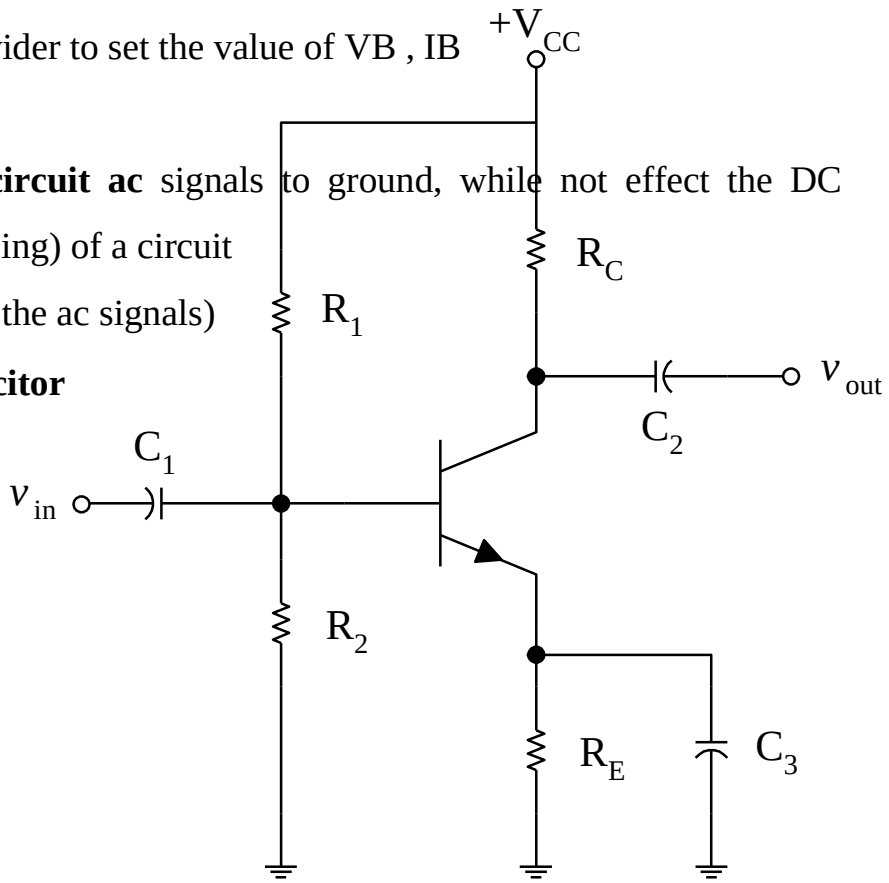
  ⇒ voltage-divider to set the value of VB , IB

- C3

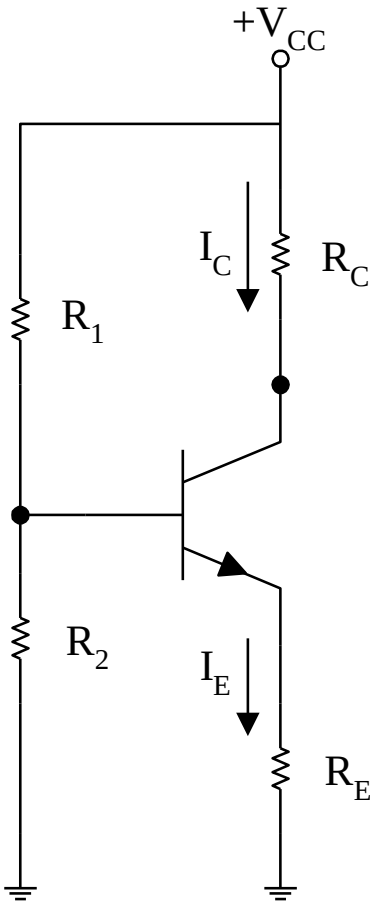
  ⇒ to **short circuit ac** signals to ground, while not effect the DC operating (or biasing) of a circuit

(RE ⇒ stabilizes the ac signals)

→ Bypass Capacitor



Graphical DC Bias Analysis



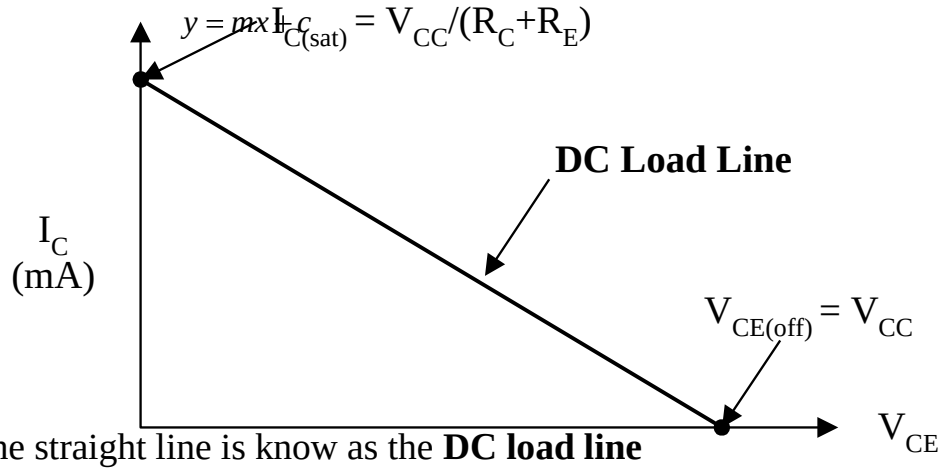
$$V_{CC} - I_C R_C - V_{CE} - I_E R_E = 0$$

$$\text{for } I_C \approx I_E$$

$$I_C = \frac{-1}{R_C + R_E} V_{CE} + \frac{V_{CC}}{R_C + R_E}$$

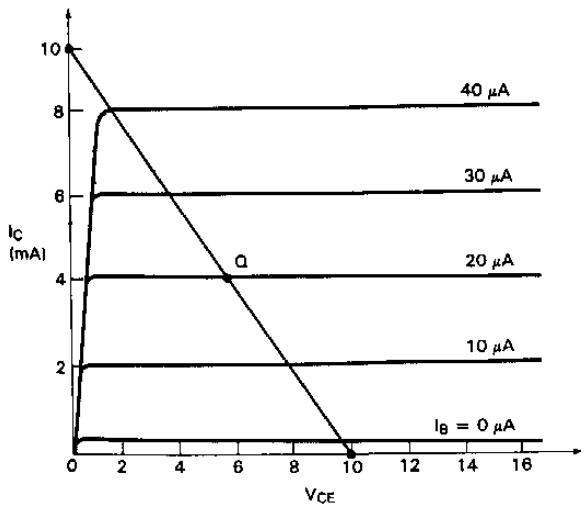
Point-slope form of straight line equation :

$y = mx + c$



- The straight line is know as the **DC load line**
- Its significance is that regardless of the behavior of the transistor, the collector current IC and the collector-emitter voltage VCE must always lie on the load line, depends ONLY on the VCC, RC and RE
- (i.e. The dc load line is a graph that **represents all the possible combinations of IC and VCE for a given amplifier**. For every possible value of *IC*, and amplifier will have a corresponding value of *VCE*.)
- It must be true at the same time as the transistor characteristic. Solve two condition using simultaneous equation

  → graphically → **Q-point !!**

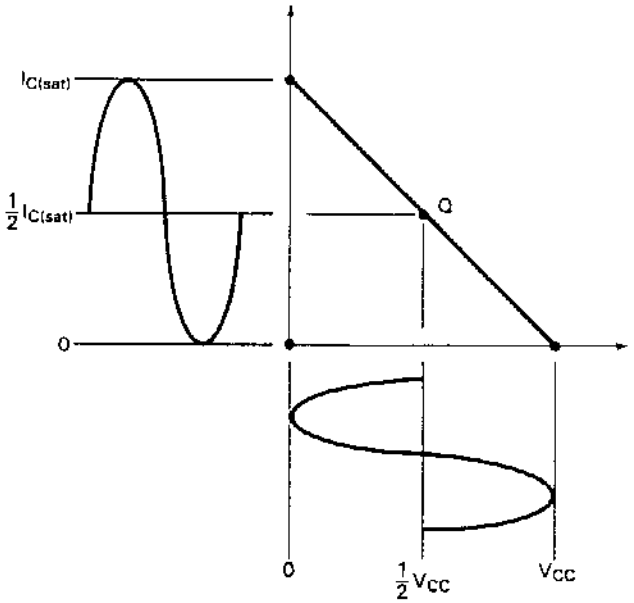


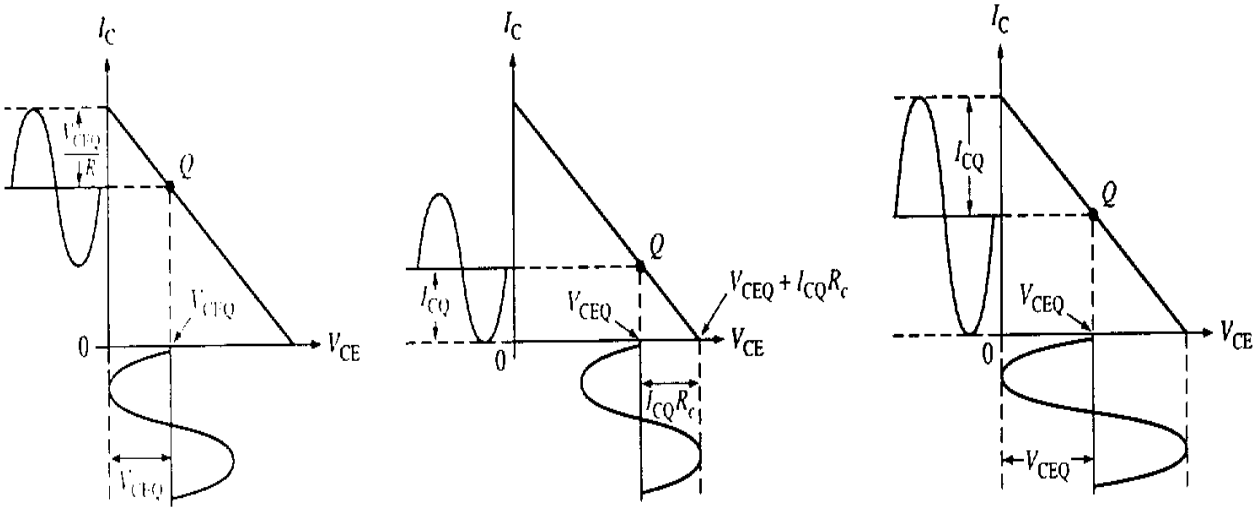
Q-Point (Static Operation Point)

- When a transistor does not have an **ac input**, it will have **specific dc values** of *IC* and *VCE*.
- These values correspond to a specific point on the **dc load line**. This point is called the ***Q-point***.
- The letter ***Q*** corresponds to the word (Latent) **quiescent**, meaning **at rest**.
- A quiescent amplifier is one that has no ac signal applied and therefore has constant dc values of *IC* and *VCE*.
- The intersection of the dc bias value of *IB* with the dc load line determines the *Q*-point.
- It is desirable to have the *Q*-point centered on the load line. Why?
- When a circuit is designed to have a centered *Q*-point, the amplifier is said to be midpoint biased.
- Midpoint biasing allows optimum ac operation of the amplifier.

DC Biasing + AC signal

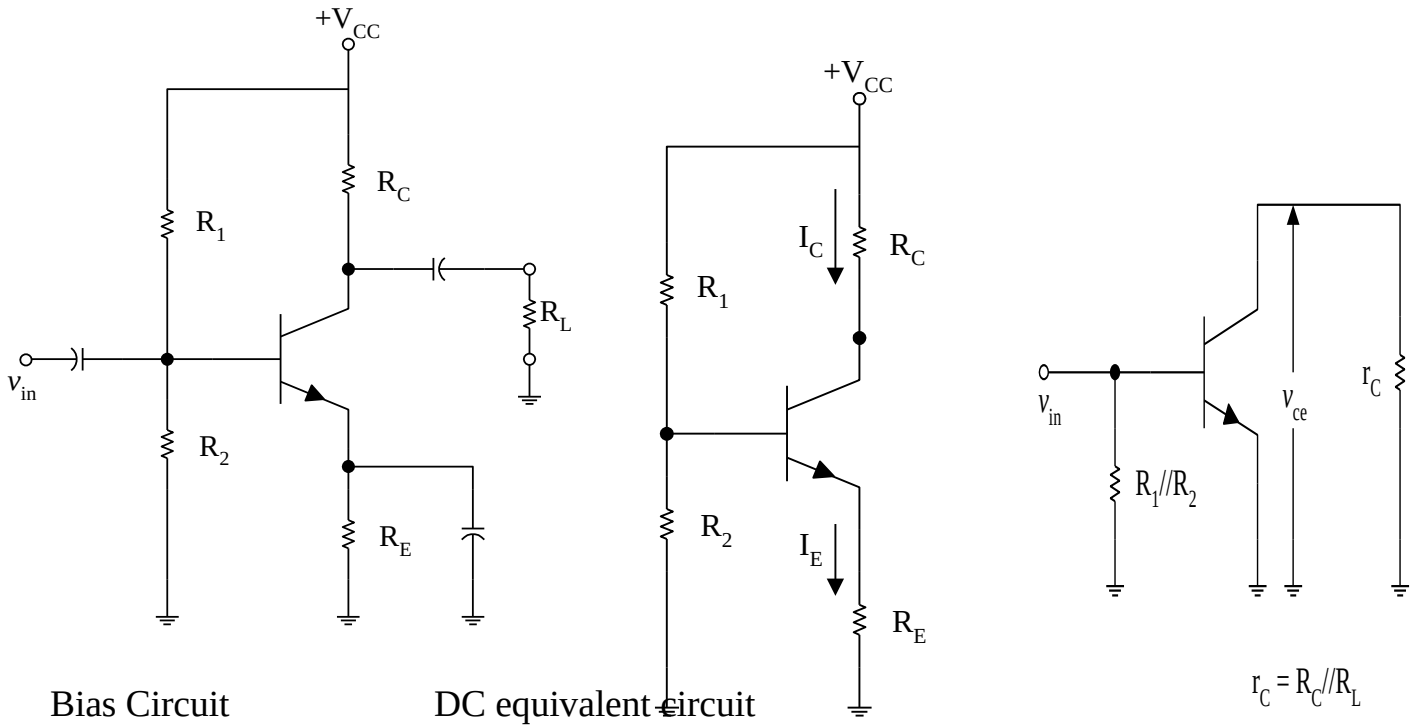
- When an **ac signal** is applied to the base of the transistor, *IC* and *VCE* will both vary around their *Q*-point values.
- When the *Q*-point is **centered**, *IC* and *VCE* can both make the **maximum** possible transitions above and below their initial dc values.
- When the *Q*-point is **above** the center on the load line, the input signal may cause the transistor to saturate. When this happens, a part of the output signal will be ***clipped*** off.
- When the *Q*-point is **below** midpoint on the load line, the input signal may cause the transistor to cutoff. This can also cause a portion of the output signal to be clipped.





(a) Limited by saturation (b) Limited by cutoff (c) Centered Q-point

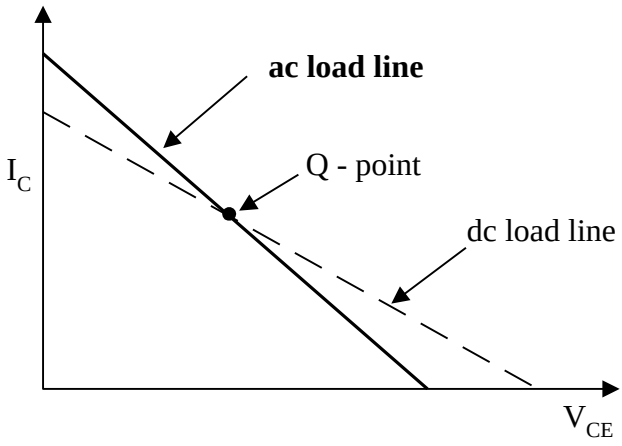
DC and AC Equivalent Circuits



Bias Circuit DC equivalent circuit

AC equ ckt

- The ac load line of a given amplifier will **not follow** the plot of the dc load line.
- This is due to the dc load of an amplifier is different from the ac load.



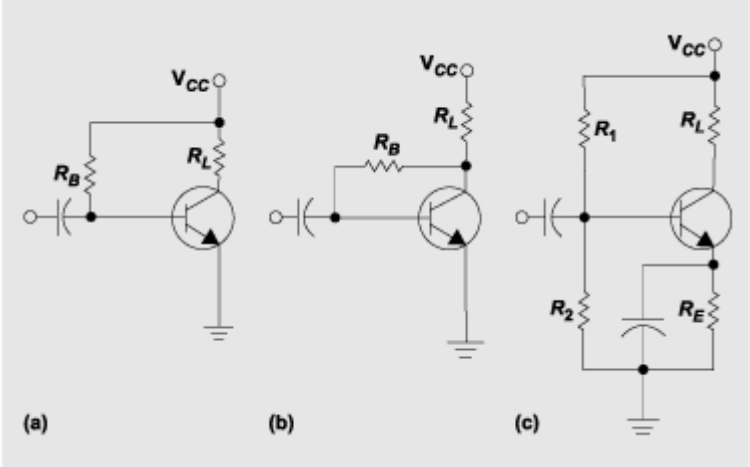
What does the ac load line tell you?

- The ac load line is used to tell you the maximum possible output voltage swing for a given common-emitter amplifier.
- In other words, the ac load line will tell you the maximum possible peak-to-peak output voltage (*Vpp*) from a given amplifier.
- This maximum *Vpp* is referred to as the **compliance** of the amplifier. (AC Saturation Current *Ic(sat)* , AC Cutoff Voltage *VCE(off)*)

Bias stabilization

The establishment of an operating point on the transistor volt-ampere characteristics by means of direct voltages and currents.

Since the transistor is a three-terminal device, any one of the three terminals may be used as a common terminal to both input and output. In most transistor circuits the emitter is used as the common terminal, and this common emitter, or grounded emitter, is indicated in illus. *a*. If the transistor is to used as a linear device, such as an audio amplifier, it must be biased to operate in the active region. In this region the collector is biased in the reverse direction and the emitter in the forward direction. The area in the common-emitter transistor characteristics to the right of the ordinate $V_{CE} = 0$ and above $I_C = 0$ is the active region. Two more biasing regions are of special interest for those cases in which the transistor is intended to operate as a switch. These are the saturation and cutoffregions. The saturation region may be defined as the region where the collector current is independent of base current for given values of V_{CC} and R_L. Thus, the onset of saturation can be considered to take place at the knee of the common-emitter transistor curves. *See also* Amplifier; Transistor.



Translator circuits. (a) Fixed-bias. (b) Collector-to-base bias. (c) Self-bias.

In saturation, the transistor current I_C is nominally V_{CC}/R_L. Since R_L is small, it may be necessary to keep V_{CC}correspondingly small in order to stay within the limitations imposed by the transistor on maximum-current and collector-power dissipation. In the cutoff region it is required that the emitter current I_E be zero, and to accomplish this it is necessary to reverse-bias the emitter junction so that the collector current is approximately equal to thereverse saturation current I_{CO}. A reverse-biasing voltage of the order of 0.1 V across the emitter junction willordinarily be adequate to cut off either a germanium or silicon transistor.

The particular method to be used in establishing an operating point on the transistor characteristics depends on whether the transistor is to operate in the active, saturation or cutoff regions; on the application under consideration; on the thermal stability of the circuit; and on other factors.

In a fixed-bias circuit, the operating point for the circuit of illus. *a* can be established by noting that the required current I_B is constant, independent of the quiescent collector current I_C, which is why this circuit is called the fixed-bias circuit. Transistor biasing circuits are frequently compared in terms of the value of the stability factor $S = \partial I_C/\partial I_{CO}$, which is the rate of change of collector current with respect to reverse saturation current. The smaller the value of S, the less likely the circuit will exhibit thermal runaway. S, as defined here, cannot be smaller than unity. Other stability factors are defined in terms of dc current gain h_{FE} as $\partial I_C/\partial h_{FE}$, and in terms of base-to-emitter voltage as $\partial I_C/\partial V_{BE}$. However, bias circuits with small values of S will also perform satisfactorily for transistors that have large variations of h_{FE} and V_{BE}. For the fixed-bias circuit it can be shown that $S = h_{FE} + 1$, and if $h_{FE} = 50$, then $S = 51$. Such a large value of S makes thermal runaway a definite possibility with this circuit.

In collector-to-base bias, an improvement in stability is obtained if the resistor R_B in illus. *a* is returned to the collector junction rather than to the battery terminal. Such a connection is shown in illus. *b*. In this bias circuit, if I_C tends to increase (either because of a rise in temperature or because the transistor has been replaced by another), then V_{CE} decreases. Hence I_B also decreases and, as a consequence of this lowered bias current, the collector current is not allowed to increase as much as it would if fixed bias were used. The stability factor S is shown in Eq. (1).

1. $$S = \frac{h_{FE} + 1}{1 + h_{FE} R_L/(R_L + R_B)}$$

This value is smaller than $h_{FE} + 1$, which is the value obtained for the fixed-bias case.

If the load resistance R_L is very small, as in a transformer-coupled circuit, then the previous expression for S shows that there would be no improvement in the stabilization in the collector-to-base bias circuit over the fixed-bias circuit. A circuit that can be used even if there is zero dc resistance in series with the collector terminal is the self-biasing configuration of illus. *c*. The current in the resistance R_E in the emitter lead causes a voltage drop which is in the direction to reverse-bias the emitter junction. Since this junction must be forward-biased (for active region bias), the bleeder R_1-R_2 has been added to the circuit.

If I_C tends to increase, the current in R_E increases. As a consequence of the increase in voltage drop across R_E, the base current is decreased. Hence I_C will increase less than it would have had there been no self-biasing resistor R_E. The stabilization factor for the self-bias circuit is shown by Eq. (2), where $R_B = R_1R_2/(R_1 + R_2)$. The smaller the value of R_B, the

2. $$S = (1 + h_{FE})\frac{1 + R_B/R_E}{1 + h_{FE} + R_B/R_E}$$

better the stabilization. Even if R_B approaches zero, the value of S cannot be reduced below unity.

In order to avoid the loss of signal gain because of the degeneration caused by R_E, this resistor is often bypassed by a very large capacitance, so that its reactance at the frequencies under consideration is very small.

The selection of an appropriate operating point (I_D, V_{GS}, V_{DS}) for a field-effect transistor (FET) amplifier stage is determined by considerations similar to those given to transistors, as discussed previously. These considerations are output-voltage swing, distortion, power dissipation,

voltage gain, and drift of drain current. In most cases it is not possible to satisfy all desired specifications simultaneously.

Question Bank

PART –A

1.what is barrier potential?

2. Define biasing.

PART-B

1.(a) Explain the working of varactor diode.

(b) What does a DC load line represent? May this line be used when an AC signal is applied? When it is necessary to draw an AC load line?

2.(a) Explain the input and output characteristics of a transistor in common emitter configuration with neat diagram.

(b) Explain how the zener diode acts as a voltage regulator with relevant diagram.

UNIT II

FET, UJT AND SCR

JFET characteristics and parameters – JFET biasing – Self bias – Voltage divider bias – Q point – Stability over temperature – MOSFET – D-MOSFET and E-MOSFET – MOSFET characteristics and parameters – MOSFET biasing – Zero bias – Voltage divider bias – Drain feedback bias – Characteristics and applications of UJT, SCR, DIAC, TRIAC.

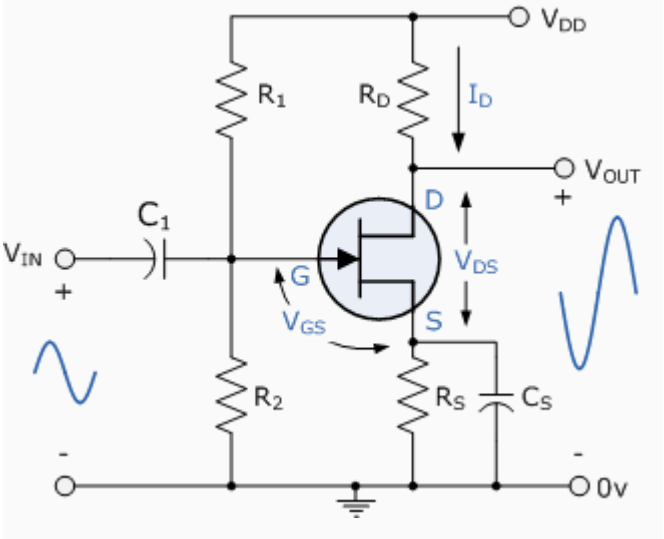
UNIT II

FET, UJT AND SCR

The Common Source JFET Amplifier

So far we have looked at the bipolar type transistor amplifier and especially the common emitter amplifier, but small signal amplifiers can also be made using **Field Effect Transistors** or **FET's** for short. These devices have the advantage over bipolar transistors of having an extremely high input impedance along with a low noise output making them ideal for use in amplifier circuits that have very small input signals. The design of an amplifier circuit based around a junction field effect transistor or "JFET", (n-channel FET for this tutorial) or even a metal oxide silicon FET or "MOSFET" is exactly the same principle as that for the bipolar transistor circuit used for a Class A amplifier circuit we looked at in the previous tutorial. Firstly, a suitable quiescent point or "Q-point" needs to be found for the correct biasing of the JFET amplifier circuit with single amplifier configurations of Common-source (CS), Common-drain (CD) or Source-follower (SF) and the Common-gate (CG) available for most FET devices. These three JFET amplifier configurations correspond to the common-emitter, emitter-follower and the common-base configurations using bipolar transistors. In this tutorial we will look at the **Common Source JFET Amplifier** as this is the most widely used JFET amplifier design. Then consider the common source JFET amplifier circuit below.

Common Source JFET Amplifier



The amplifier circuit consists of an N-channel JFET, but the device could also be an equivalent N-channel depletion-mode MOSFET as the circuit diagram would be the same just a change in the FET, connected in a common source configuration. The JFET gate voltage Vg is biased through the potential divider network set up by resistors R1 and R2 and is biased to operate within its saturation region which is equivalent to the active region of the bipolar junction transistor. Unlike a bipolar transistor circuit, the junction FET takes virtually no input gate current allowing the gate to be treated as an open circuit. Then no input characteristics curves are required. We can compare the JFET to the bipolar junction transistor (BJT) in the following table.

JFET to BJT Comparison

JFET	BJT
Gate, (*G*)	Base, (*B*)
Drain, (*D*)	Collector, (*C*)
Source, (*S*)	Emitter, (*E*)

Gate Supply, (V_g)	Base Supply, (V_B)
Drain Supply, (V_{DD})	Collector Supply, (V_{CC})
Drain Current, (i_D)	Collector Current, (i_C)

Since the N-Channel JFET is a depletion mode device and is normally "ON", a negative gate voltage with respect to the source is required to modulate or control the drain current. This negative voltage can be provided by biasing from a separate power supply voltage or by a self biasing arrangement as long as a steady current flows through the JFET even when there is no input signal present and Vg maintains a reverse bias of the gate-source pn junction. In this example the biasing is provided from a potential divider network allowing the input signal to produce a voltage fall at the gate as well as voltage rise at the gate with a sinusoidal signal. Any suitable pair of resistor values in the correct proportions would produce the correct biasing voltage so the DC gate biasing voltage Vg is given as:

$$V_G = \frac{V_{DD}\,R_2}{R_1 + R_2} = V_{DD}\left(\frac{R_2}{R_1 + R_2}\right)$$

Note that this equation only determines the ratio of the resistors R1 and R2, but in order to take advantage of the very high input impedance of the JFET as well as reducing the power dissipation within the circuit, we need to make these resistor values as high as possible, with values in the order of 1 to 10MΩ being common.

The input signal, (Vin) of the common source JFET amplifier is applied between the Gate terminal and the zero volts rail, (0v). With a constant value of gate voltage Vg applied the JFET operates within its "Ohmic region" acting like a linear resistive device. The drain circuit contains the load resistor, Rd. The output voltage, Vout is developed across this load resistance. The efficiency of the common source JFET amplifier can be improved by the addition of a resistor, Rs included in the source lead with the same drain current flowing through this resistor. Resistor, Rs is also used to set the JFET amplifiers "Q-point".

When the JFET is switched fully "ON" a voltage drop equal to Rs x Id is developed across this resistor raising the potential of the source terminal above 0v or ground level. This voltage drop across Rs due to the drain current provides the necessary reverse biasing condition across the gate resistor, R2 effectively generating negative feedback. In order to keep the gate-source junction reverse biased, the source voltage, Vs needs to be higher than the gate voltage, Vg. This source voltage is therefore given as:

$$V_S = I_D \times R_S = V_G - V_{GS}$$

Then the Drain current, Id is also equal to the Source current, Is as "No Current" enters the Gate terminal and this can be given as:

$$I_D = \frac{V_S}{R_S} = \frac{V_{DD}}{R_D + R_S}$$

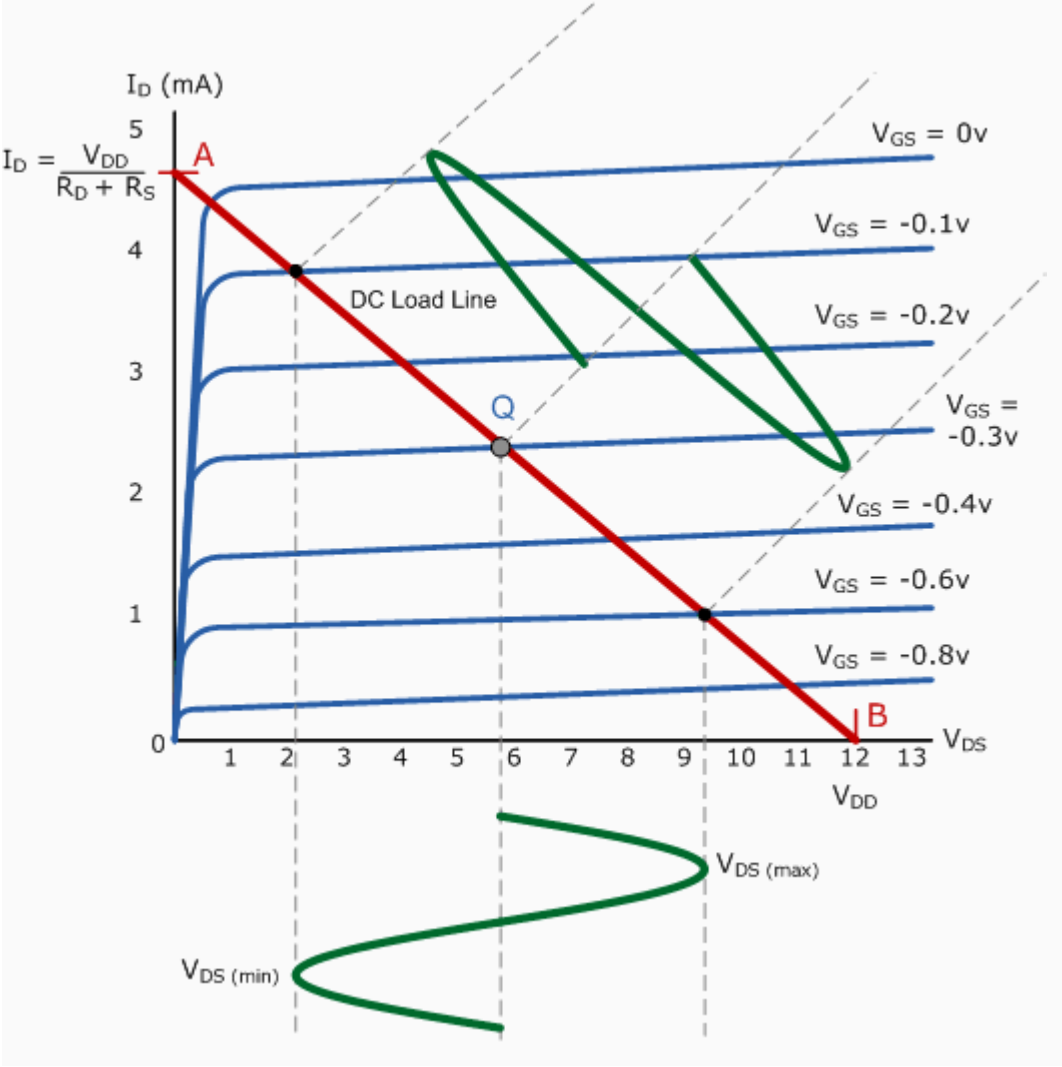
This potential divider biasing circuit improves the stability of the common source JFET amplifier circuit when being fed from a single DC supply

compared to that of a fixed voltage biasing circuit. Both resistor, Rs and the source by-pass capacitor, Cs serve basically the same function as the emitter resistor and capacitor in the common emitter bipolar transistor amplifier circuit, namely to provide good stability and prevent a reduction in the loss of the voltage gain. However, the price paid for a stabilized quiescent gate voltage is that more of the supply voltage is dropped across Rs.

The the value in farads of the source by-pass capacitor is generally fairly high above 100uF and will be polarized. This gives the capacitor an impedance value much smaller, less than 10% of the transconductance, gm (the transfer coefficient representing gain) value of the device. At high frequencies the by-pass capacitor acts essentially as a short-circuit and the source will be effectively connected directly to ground.

The basic circuit and characteristics of a **Common Source JFET Amplifier** are very similar to that of the common emitter amplifier. A DC load line is constructed by joining the two points relating to the drain current, Id and the supply voltage, Vdd remembering that when Id = 0: (Vdd = Vds) and when Vds = 0: (Id = Vdd/R_L). The load line is therefore the intersection of the curves at the Q-point as follows.

Common Source JFET Amplifier Characteristics Curves



As with the common emitter bipolar circuit, the DC load line for the common source JFET amplifier produces a straight line equation whose gradient is given as: -1/(Rd + Rs) and that it crosses the vertical Id axis at point A equal to Vdd/(Rd + Rs). The other end of the load line crosses the horizontal axis at point B which is equal to the supply voltage, Vdd. The actual position of the Q-point on the DC load line is generally positioned at the mid centre point of the load line (for class-A operation) and is determined by the mean value of Vg which is biased negatively as the JFET is a depletion-mode device. Like the bipolar common emitter amplifier the

output of the **Common Source JFET Amplifier** is 180^{o} out of phase with the input signal.

One of the main disadvantages of using Depletion-mode JFET is that they need to be negatively biased. Should this bias fail for any reason the gate-source voltage may rise and become positive causing an increase in drain current resulting in failure of the drain voltage, Vd. Also the high channel resistance, Rds(on) of the junction FET, coupled with high quiescent steady state drain current makes these devices run hot so additional heatsink is required. However, most of the problems associated with using JFET's can be greatly reduced by using enhancement-mode MOSFET devices instead.

We know that controlling the Q point of our JFET is more difficult than it was with our junction transistor. This is because IDSS varies widely from one JFET to the next. In order to stabilize I to a constant level from one JFET to another, we need a circuit that will vary
VGSwidely.

Gate bias

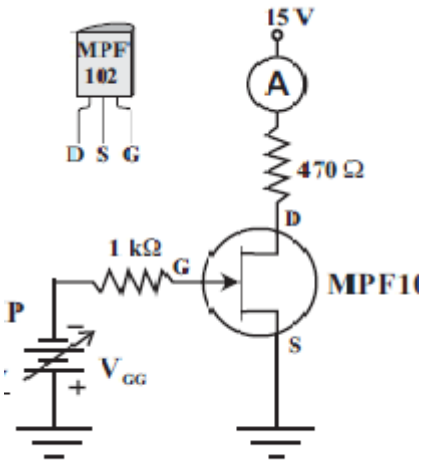
is the simplest but worst way to control the drain current.With gate bias, we supply a constant value of V , and the resulting drain current will vary widely from device to device.

Self bias

offers some improvement because the source resistor produces local feedback. Here, the value
of V varies somewhat with the value of drain current. This helps to control the drain current.

Voltage Divider bias

results in a relatively stable Q point, however, it requires a large supply voltage.



Gate Bias

Gate Bias

1) With gate bias, you apply a fixed gate voltage that reverse biases the gate of the JFET. This produces a drain current that is less than I .

The problem is that you cannot accurately predict the drain current in mass production because of the variation in the required V .

The following will illustrate this point.

2) Build the circuit shown in Figure 3. Apply a V of -1.5V.

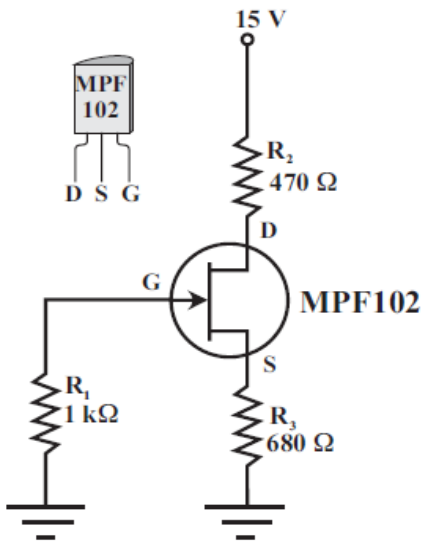
Measure V , I , and V . and record the data in Table 2 for each JFET.

Self Bias

Build the circuit shown in Figure 4. Measure and record the 3 values shown in Table 3. Repeat the measurements for the other JFETs.

The drain current variation for the self- bias circuit should be less than the

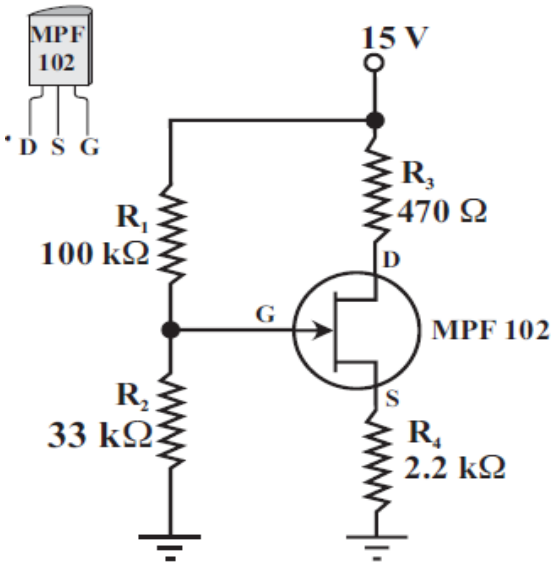
variation than the gate-biased circuit.



Voltage Divider Bias

Build the circuit shown in Figure 5. Measure and record the 3 values shown in Table 4. Repeat the measurements for the other JFET.

The drain current variation for the voltage divider bias circuit should continue to become more stable. It should show less variation than either of the circuits above.



Voltage Divider Bias

Q-Point

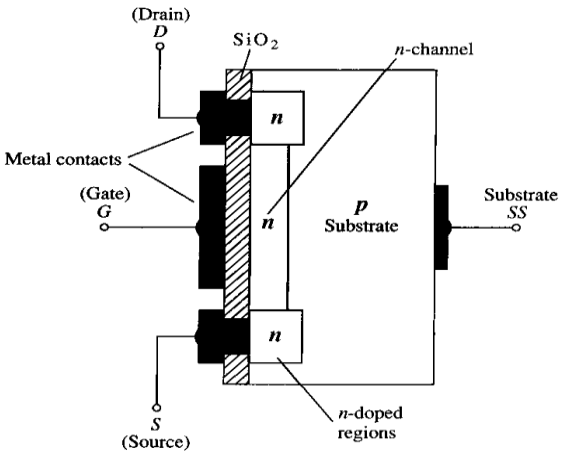
the bias point is chosen to keep the transistor operating in the active mode, using a variety of circuit techniques, establishing the Q-point DC voltage and current. A small signal is then applied on top of the Q-point bias voltage, thereby either modulating or switching the current, depending on the purpose of the circuit.

The quiescent point of operation is typically near the middle of DC load line. The process of obtaining certain DC collector current at a certain DC collector voltage by setting up operating point is called biasing.

After establishing the operating point, when input signal is applied, the output signal should not move the transistor either to saturation or to cut-off. However, this unwanted shift still might occur, due to the following reasons:

1. Parameters of transistors depend on junction temperature. As junction temperature increases, leakage current due to minority charge carriers (I_{CBO}) increases. As I_{CBO} increases, I_{CEO} also increases, causing an increase in collector current I_C. This produces heat at the collector junction. This process repeats, and, finally, Q-point may shift into the saturation region. Sometimes, the excess heat produced at the junction may even burn the transistor. This is known as thermal runaway.
2. When a transistor is replaced by another of the same type, the Q-point may shift, due to changes in parameters of the transistor, such as *current gain* (β) which varies slightly for each unique transistor.

To avoid a shift of Q-point, bias-stabilization is necessary. Various biasing circuits can be used for this purpose.



MOSFETs or Metal Oxide Semiconductor FET's have much higher input impedances and low channel resistances compared to the equivalent JFET. Also the biasing arrangements for MOSFETs are different and unless we bias them positively for N-channel devices and negatively for P-channel devices no drain current will flow, then we have in effect a fail safe transistor.

A **Power MOSFET** is a specific type of metal oxide semiconductor field-effect transistor (MOSFET) designed to handle large amounts of power. Compared to the other power semiconductor devices (IGBT, Thyristor...), its main advantages are high commutation speed and good efficiency at low voltages. It shares with the IGBT an isolated gate that makes it easy to drive.

It was made possible by the evolution of CMOS technology, developed for manufacturing Integrated circuits in the late 1970s. The power MOSFET

shares its operating principle with its low-power counterpart, the lateral MOSFET.

- The power MOSFET is the most widely used low-voltage (i.e. less than 200 V) switch. It can be found in most power supplies, DC to DC converters, and low voltage motor controllers.

Basic structure

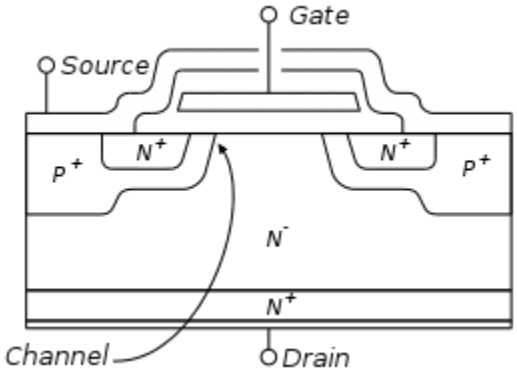


Fig. Cross section of a VDMOS, showing an elementary cell. Note that a cell is very small (some micrometres to some tens of micrometres wide), and that a power MOSFET is composed of several thousand of them.

Several structures had been explored at the beginning of the 1980s, when the first Power MOSFETs were introduced. However, most of them have been abandoned (at least until recently) in favour of the Vertical Diffused MOS (VDMOS) structure (also called Double-Diffused MOS or simply DMOS).

The cross section of a VDMOS (see figure 1) shows the "verticality" of the device: It can be seen that the source electrode is placed over the drain,

resulting in a current mainly vertical when the transistor is in the on-state. The "diffusion" in VDMOS refers to the manufacturing process: the P wells are obtained by a diffusion process (actually a double diffusion process to get the P and N^+ regions, hence the name double diffused).

Power MOSFETs have a different structure than the lateral MOSFET: as with all power devices, their structure is vertical and not planar. In a planar structure, the current and breakdown voltage ratings are both functions of the channel dimensions (respectively width and length of the channel), resulting in inefficient use of the "silicon estate". With a vertical structure, the voltage rating of the transistor is a function of the doping and thickness of the N epitaxial layer (see cross section), while the current rating is a function of the channel width. This makes possible for the transistor to sustain both high blocking voltage and high current within a compact piece of silicon.

It is worth noting that power MOSFETs with lateral structure exist. They are mainly used in high-end audio amplifiers. Their advantage is a better behaviour in the saturated region (corresponding to the linear region of a bipolar transistor) than the vertical MOSFETs. Vertical MOSFETs are designed for switching applications, so they are only used in On or Off states.

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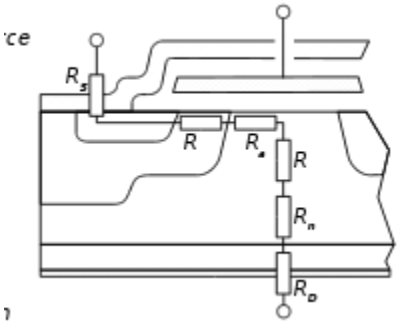
State characteristics

On-state resistance



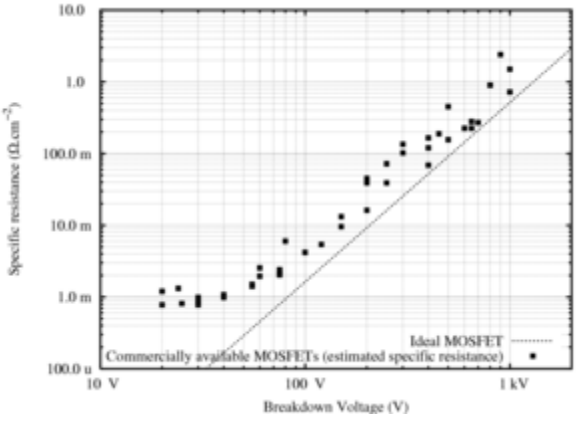
Contribution of the different parts of the MOSFET to the on-state resistance.

When the power MOSFET is in the on-state (see MOSFET for a discussion on operation modes), it exhibits a resistive behaviour between the drain and source terminals. It can be seen in figure 2 that this resistance (called R_{DSon} for "drain to source resistance in on-state") is the sum of many elementary contributions:

- R_S is the source resistance. It represents all resistances between the source terminal of the package to the channel of the MOSFET: resistance of the wire bonds, of the source metallisation, and of the N^+ wells;
- R_{ch}. This is the channel resistance. It is inversely proportional to the channel width, and for a given die size, to the channel density. The channel resistance is one of the main contributors to the R_{DSon} of low-voltage MOSFETs, and intensive work has been carried out to reduce their cell size in order to increase the channel density;

- R_a is the *access* resistance. It represents the resistance of the epitaxial zone directly under the gate electrode, where the direction of the current changes from horizontal (in the channel) to vertical (to the drain contact);
- R_{JFET} is the detrimental effect of the cell size reduction mentioned above: the P implantations form the gates of a parasitic JFET transistor that tend to reduce the width of the current flow;
- R_n is the resistance of the epitaxial layer. As the role of this layer is to sustain the blocking voltage, R_n is directly related to the voltage rating of the device. A high voltage MOSFET requires a thick, low-doped layer (i.e. highly resistive), whereas a low-voltage transistor only requires a thin layer with a higher doping level (i.e. less resistive). As a result, R_n is the main factor responsible for the resistance of high-voltage MOSFETs;
- R_D is the equivalent of R_S for the drain. It represents the resistance of the transistor substrate (note that the cross section in figure is not at scale, the bottom N^+ layer is actually the thickest) and of the package connections.

Breakdown voltage/on-state resistance trade-off



The R_{DSon} of the MOSFETs increase with their Voltage rating.

When in the OFF-state, the power MOSFET is equivalent to a PIN diode (constituted by the P^+ diffusion, the N^- epitaxial layer and the N^+ substrate). When this highly non-symmetrical structure is reverse-biased, the space-charge region extends principally on the light-doped side, i.e over the N^- layer. This means that this layer has to withstand most of the MOSFET's OFF-state drain-to-source voltage.

However, when the MOSFET is in the ON-state, this N^- layer has no function. Furthermore, as it is a lightly-doped region, its intrinsic resistivity is non-negligible and adds to the MOSFET's ON-state Drain-to-Source Resistance (R_{DSon}) (this is the R_n resistance in figure).

Two main parameters govern both the breakdown voltage and the R_{DSon} of the transistor: the doping level and the thickness of the N^- epitaxial layer. The thicker the layer and the lower its doping level, the higher the breakdown voltage. On the contrary, the thinner the layer and the higher the doping level, the lower the R_{DSon} (and therefore the lower the conduction losses of the MOSFET). Therefore, it can be seen that there is a trade-off in the design of a MOSFET, between its voltage rating and its ON-state resistance. This is demonstrated by the plot in figure.

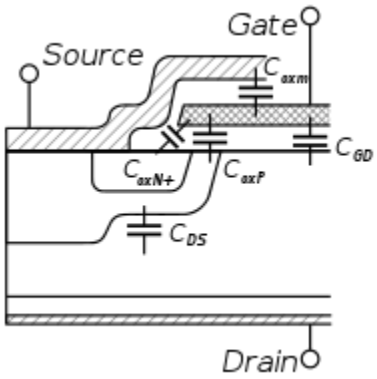
Body diode

It can be seen in figure 1 that the source metallization connects both the N^+ and P implantations, although the operating principle of the MOSFET only requires the source to be connected to the N^+ zone. However, if it were, this

would result in a floating P zone between the N-doped source and drain, which is equivalent to a NPN transistor with a non-connected base. Under certain conditions (under high drain current, when the on-state drain to source voltage is in the order of some volts), this parasitic NPN transistor would be triggered, making the MOSFET uncontrollable. The connection of the P implantation to the source metallization shorts the base of the parasitic transistor to its emitter (the source of the MOSFET) and thus prevents spurious latching.

This solution, however, creates a diode between the drain (cathode) and the source (anode) of the MOSFET, making it able to block current in only one direction.

Switching operation



Location of the intrinsic capacitances of a power MOSFET.

Because of their unipolar nature, the power MOSFET can switch at very high speed. Indeed, there is no need to remove minority carriers as with bipolar devices.

The only intrinsic limitation in commutation speed is due to the internal capacitances of the MOSFET (see figure). These capacitances must be charged or discharged when the transistor switches. This can be a relatively slow process because the current that flows through the gate capacitances is limited by the external driver circuit. This circuit will actually dictate the commutation speed of the transistor (assuming the power circuit has sufficiently low inductance).

Capacitances

In the MOSFETs datasheets, the capacitances are often named C_{iss} (input capacitance, drain and source terminal shorted), C_{oss} (output capacitance, gate and source shorted), and C_{rss} (reverse transfer capacitance, gate and source shorted). The relationship between these capacitances and those described below is:

$$\begin{aligned} C_{iss} &= C_{GS} + C_{GD} \\ C_{oss} &= C_{GD} + C_{DS} \\ C_{rss} &= C_{GD} \end{aligned}$$

Where C_{GS}, C_{GD} and C_{DS} are respectively the gate-to-source, gate-to-drain and drain-to-source capacitances (see below). Manufacturers prefer to quote C_{iss}, C_{oss} and C_{rss} because they can be directly measured on the transistor. However, as C_{GS}, C_{GD} and C_{DS} are closer to the physical meaning, they will be used in the remaining of this article.

Gate to source capacitance

The C_{GS} capacitance is constituted by the parallel connection of C_{oxN+}, C_{oxP} and C_{oxm} (see figure 4). As the N^+ and P regions are highly doped, the two

former capacitances can be considered as constant. C_{oxm} is the capacitance between the (polysilicon) gate and the (metal) source electrode, so it is also constant. Therefore, it is common practice to consider C_{GS} as a constant capacitance, i.e its value does not depend on the transistor state.

Gate to drain capacitance

The C_{GD} capacitance can be seen as the connection in series of two elementary capacitances. The first one is the oxide capacitance (C_{oxD}), constituted by the gate electrode, the silicon dioxide and the top of the N epitaxial layer. It has a constant value. The second capacitance (C_{CDj}) is caused by the extension of the space-charge zone when the MOSFET is in off-state (see the section Blocking Voltage). Therefore, it is dependent upon the drain to source voltage. From this, the value of C_{GD} is:

$$C_{GD} = \frac{C_{oxD} \times C_{GDj}\left(V_{GD}\right)}{C_{oxD} + C_{GDj}\left(V_{GD}\right)}$$

The width of the space-charge region is given by

$$w_{GDj} = \sqrt{\frac{2\epsilon_{Si} V_{GD}}{qN}}$$

where ε_{Si} is the permittivity of the Silicon, q is the electron charge, and N is the doping level. The value of C_{GDj} can be approximated using the expression of the plane capacitor:

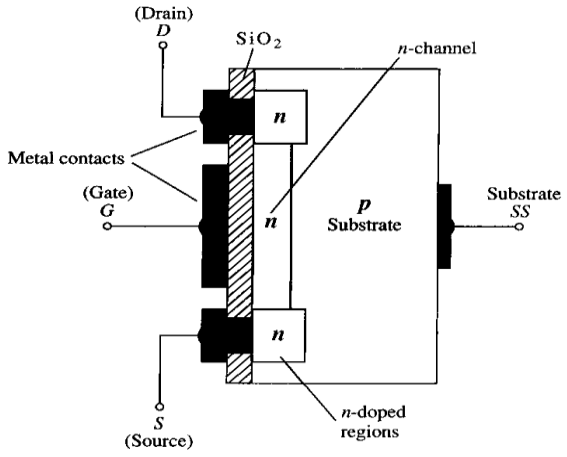
$$C_{GDj} = A_{GD} \frac{\epsilon_{Si}}{w_{GDj}}$$

Where A_{GD} is the surface area of the gate-drain overlap. Therefore, it comes:

$$C_{GDj}\left(V_{GD}\right) = A_{GD}\sqrt{\frac{q\epsilon_{Si}N}{2V_{GD}}}$$

It can be seen that C_{GDj} (and thus C_{GD}) is a capacitance which value is dependent upon the gate to drain voltage. As this voltage increases, the capacitance decreases. When the MOSFET is in on-state, C_{GDj} is shunted, so the gate to drain capacitance remains equal to C_{oxD}, a constant value.

D-MOSFET

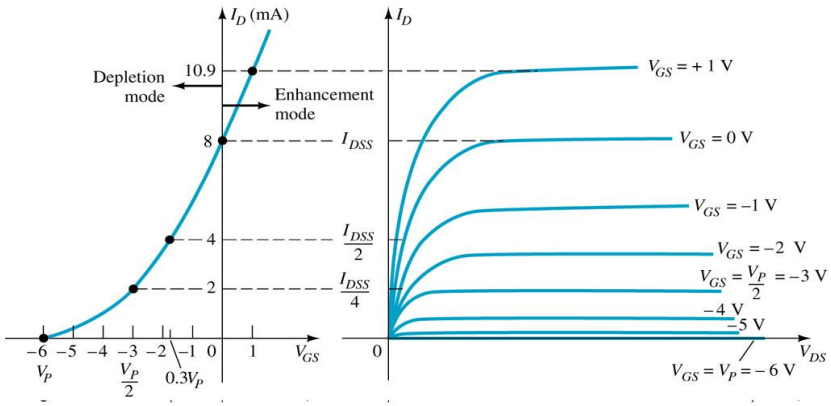


The Drain (D) and Source (S) leads connect to the to n-doped regions

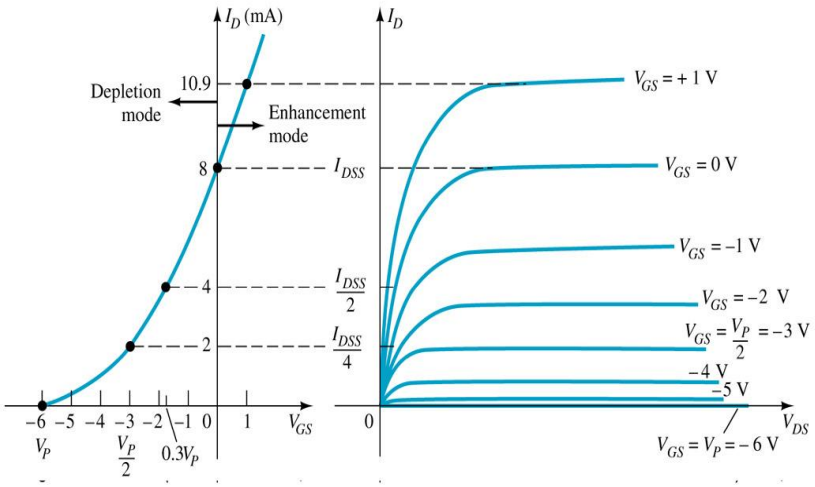
These N-doped regions are connected via an n-channel.This n-channel is connected to the Gate (G) via a thin insulating layer of SiO_2.The n-doped material lies on a p-doped substrate that may have an additional terminal connection called SS.

A D-MOSFET may be biased to operate in two modes:

the **Depletion** mode or the **Enhancement** mode



D-MOSFET Depletion Mode Operation



The transfer characteristics are similar to the JFET

In Depletion Mode operation:

When VGS = 0V, ID = IDSS

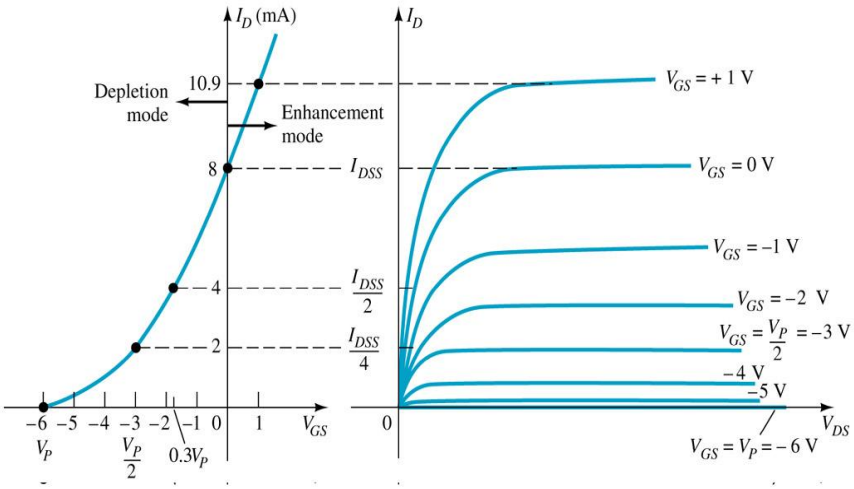
When VGS < 0V, ID < IDSS

When VGS > 0V, ID > IDSS

The formula used to plot the Transfer Curve, is:

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

D-MOSFET Enhancement Mode Operation

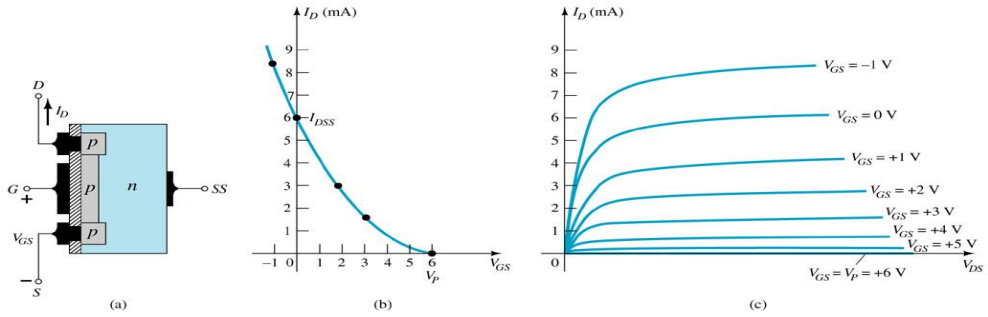


In this mode, the transistor operates with $V_{GS} > 0V$, and I_D increases above I_{DSS}

Shockley's equation, the formula used to plot the Transfer Curve, still applies but VGS is positive:

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

p-Channel Depletion Mode MOSFET



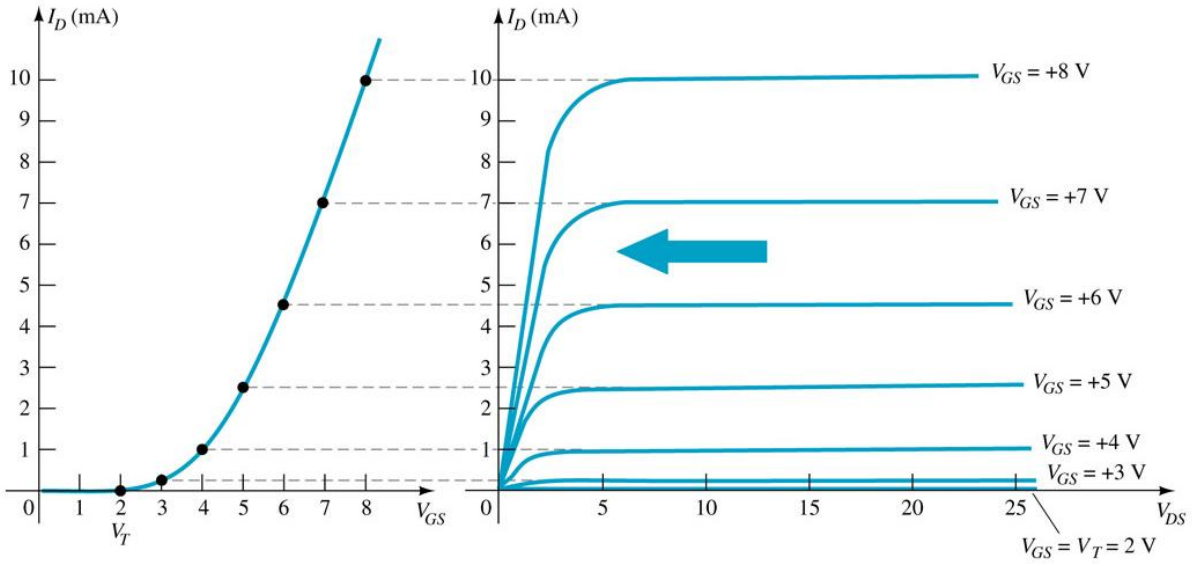
The p-channel Depletion mode MOSFET is similar to the n-channel except that the voltage polarities and current directions are reversed

The Drain (D) and Source (S) connect to the to n-doped regions

These n-doped regions are not connected via an n-channel without an external voltage

The Gate (G) connects to the p-doped substrate via a thin insulating layer of SiO_2

The n-doped material lies on a p-doped substrate that may have an additional terminal connection called SS



VGS is always positive

IDSS = 0 when VGS < VT

As VGS increases above VT, ID increases

If VGS is kept constant and VDS is increased, then ID saturates (IDSS)

The saturation level, VDSsat is reached.

To determine ID given VGS: where VT = threshold voltage or voltage at which the MOSFET turns on.

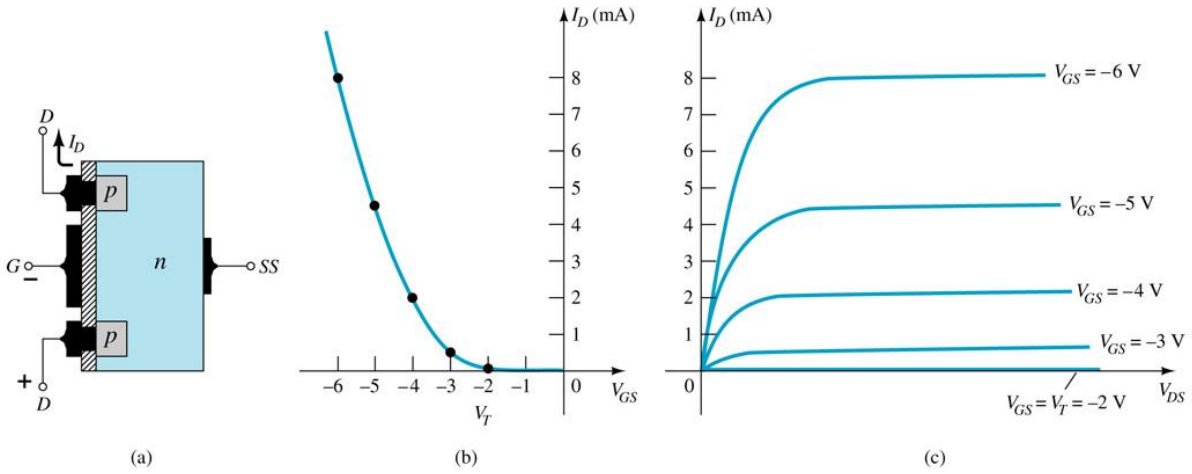
k = constant found in the specification sheet

The PSpice determination of k is based on the geometry of the device:

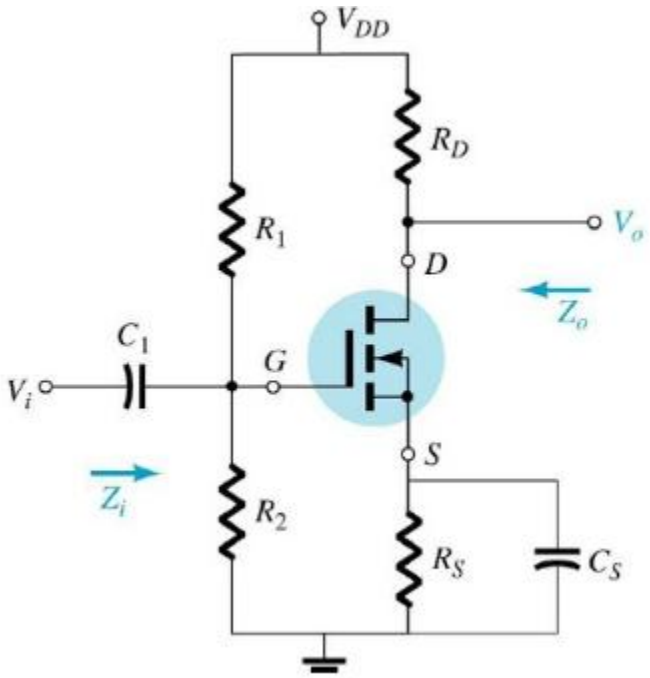
$$\mathbf{k} = \left(\frac{\mathbf{W}}{\mathbf{L}}\right)\left(\frac{\mathbf{KP}}{\mathbf{2}}\right) \quad \textbf{where } \mathbf{KP} = \mu_N C_{ox}$$

p-Channel Enhancement Mode MOSFET

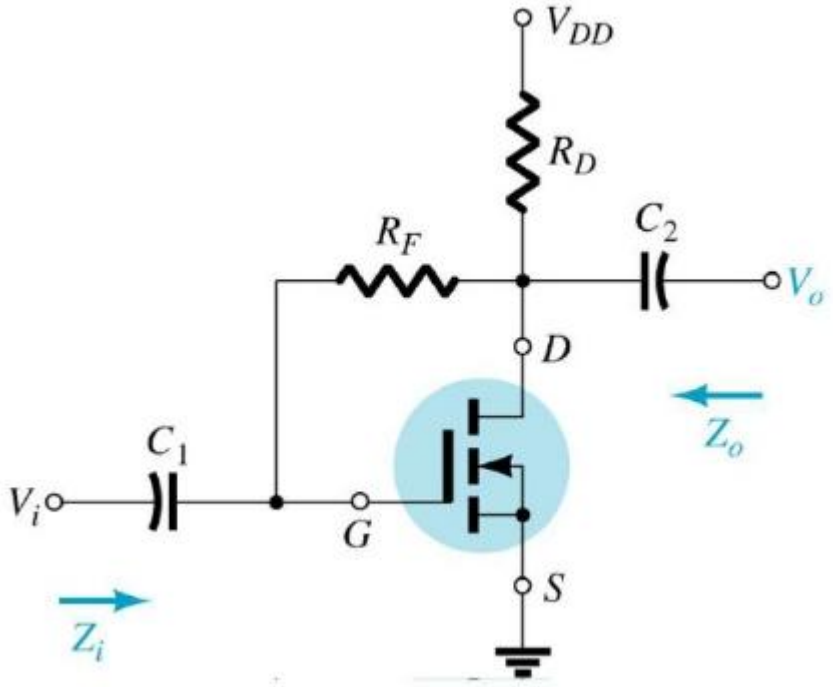
The p-channel Enhancement mode MOSFET is similar to the n-channel except that the voltage polarities and current directions are reversed.



Voltage divider bias

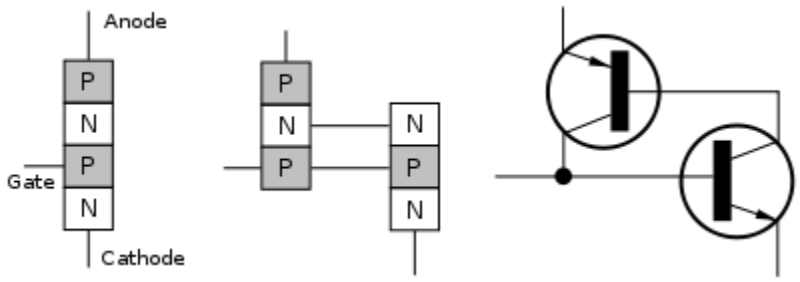


Drain feedback bias



SCR

The thyristor is a four-layer, three terminal semiconducting device, with each layer consisting of alternately N-type or P-type material, for example P-N-P-N. The main terminals, labeled anode and cathode, are across the full four layers, and the control terminal, called the gate, is attached to p-type material near to the cathode. (A variant called an SCS—Silicon Controlled Switch—brings all four layers out to terminals.) The operation of a thyristor can be understood in terms of a pair of tightly coupled bipolar junction transistors, arranged to cause the self-latching action:

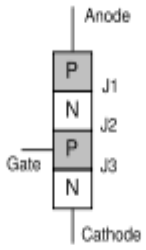


Thyristors have three states:

1. Reverse blocking mode — Voltage is applied in the direction that would be blocked by a diode
2. Forward blocking mode — Voltage is applied in the direction that would cause a diode to conduct, but the thyristor has not yet been triggered into conduction
3. Forward conducting mode — The thyristor has been triggered into conduction and will remain conducting until the forward current drops below a threshold value known as the "holding current"

Function of the gate terminal

The thyristor has three p-n junctions (serially named J_1, J_2, J_3 from the anode).



Layer diagram of thyristor.

When the anode is at a positive potential V_{AK} with respect to the cathode with no voltage applied at the gate, junctions J_1 and J_3 are forward biased, while junction J_2 is reverse biased. As J_2 is reverse biased, no conduction takes place (Off state). Now if V_{AK} is increased beyond the breakdown voltage V_{BO} of the thyristor, avalanche breakdown of J_2 takes place and the thyristor starts conducting (On state).

If a positive potential V_G is applied at the gate terminal with respect to the cathode, the breakdown of the junction J_2 occurs at a lower value of V_{AK}. By selecting an appropriate value of V_G, the thyristor can be switched into the on state suddenly.

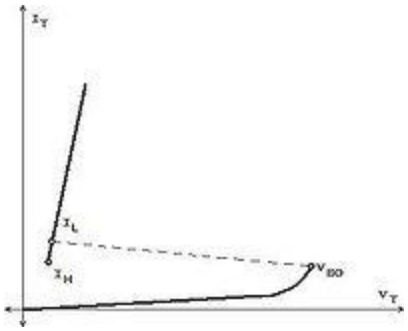
Once avalanche breakdown has occurred, the thyristor continues to conduct, irrespective of the gate voltage, until both: (a) the potential V_G is removed and (b) the current through the device (anode−cathode) is less than the holding current specified by the manufacturer. Hence V_G can be a voltage pulse, such as the voltage output from a UJT relaxation oscillator.

These gate pulses are characterized in terms of gate trigger voltage (V_{GT}) and gate trigger current (I_{GT}). Gate trigger current varies inversely with

gate pulse width in such a way that it is evident that there is a minimum gate charge required to trigger the thyristor.

Switching characteristics

in a conventional thyristor, once it has been switched on by the gate terminal, the device remains latched in the on-state (*i.e.* does not need a continuous supply of gate current to conduct), providing the anode current has exceeded the latching current (I_L). As long as the anode remains positively biased, it cannot be switched off until the anode current falls below the holding current (I_H).



V - *I* characteristics.

A thyristor can be switched off if the external circuit causes the anode to become negatively biased. In some applications this is done by switching a second thyristor to discharge a capacitor into the cathode of the first thyristor. This method is called forced commutation.

After a thyristor has been switched off by forced commutation, a finite time delay must have elapsed before the anode can again be positively biased *and* retain the thyristor in the off-state. This minimum delay is called the circuit commutated turn off time (t_Q). Attempting to positively bias the anode within this time causes the thyristor to be self-triggered

by the remaining charge carriers (holes and electrons) that have not yet recombined.

For applications with frequencies higher than the domestic AC mains supply (e.g. 50 Hz or 60 Hz), thyristors with lower values of t_Q are required. Such fast thyristors are made by diffusing into the silicon heavy metals ions such as gold or platinum which act as charge combination centres. Alternatively, fast thyristors may be made by neutron irradiation of the silicon.

The **SIDAC** is a less common electrically similar device, the difference in naming being determined by the manufacturer. In general, SIDACs have higher breakover voltages and current handling.

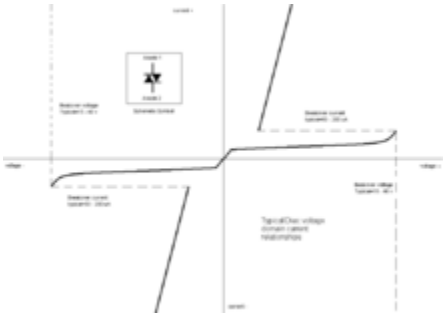
The SIDAC, or *Silicon Diode for Alternating Current*, is another member of the thyristor family. Also referred to as a SYDAC (Silicon thyristor for Alternating Current), bi-directional thyristor breakover diode, or more simply a bi-directional thyristor diode, it is technically specified as a bilateral voltage triggered switch. Its operation is similar to that of the DIAC, but SIDAC is always a five-layer device with low-voltage drop in latched conducting state, more like a voltage triggered TRIAC without a gate. In general, SIDACs have higher breakover voltages and current handling capacities than DIACs, so they can be directly used for switching and not just for triggering of another switching device.

The operation of the SIDAC is functionally similar to that of a spark gap. The SIDAC remains nonconducting until the applied voltage meets or exceeds its rated breakover voltage. Once entering this conductive state going through the negative dynamic resistance region, the SIDAC continues to conduct, regardless of voltage, until the applied current falls below its

rated holding current. At this point, the SIDAC returns to its initial nonconductive state to begin the cycle once again.

Somewhat uncommon in most electronics, the SIDAC is relegated to the status of a special purpose device. However, where part-counts are to be kept low, simple relaxation oscillators are needed, and when the voltages are too low for practical operation of a spark gap, the SIDAC is an indispensable component.

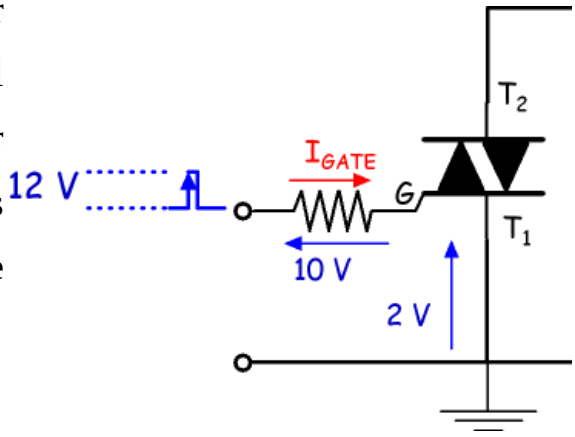
Similar devices, though usually not functionally interchangeable with SIDACs, are the Thyristor Surge Protection Devices (TSPD), Trisil, SID Actor, or the now-obsolete Surgector. These are designed to tolerate large surge currents for the suppression of overvoltage transients.



Idealized breakover diode voltage and current relationships. Once the voltage exceeds the turn-on threshold, the device turns on and the voltage rapidly falls while the current increases.

Gate Trigger Voltage and Gate Trigger Current: they are the input signal conditions needed to trigger the TRIAC. We must choose the input voltage and limiting resistor in order to get a several times higher current than the required for the triggering. Let's analize just that part of the circuit.

In our example, we want to use a 12 Volt pulse, but we cannot apply it directly to the gate, we need a limiting resistor. We know that the gate trigger voltage is 2 Volt as a maximum. The gate trigger current is 10mA, so we will choose a bigger value in order to assure the triggering, let's take 30 mA. The voltage across the resistor is:



$$V_R \simeq 12V - 2V = 10V$$

So if we want a gate trigger current of approximately 30 mA, we will need a limiting resistor of 330 ohm, this way the gate current is:

$$I_G = I_R = \frac{V_R}{R} = \frac{10V}{330\Omega} = 30mA$$ (by Ohm's law)

Latching and Holding Current: these parameters refer to the current between T2 and T1 terminals, which is the same that the load current. As we said before, once triggered, the TRIAC will remain conducting. Well this is

true whenever the current on the load at the moment of the triggering exceeds the **Latching Current**. And when we said "It will switch off when the current becomes too low", this threshhold is specified by the **Holding Current** value, which is lower than the latching current.
From a practical point of view, we just need to verify that the load current is far bigger than the latching and holding currents (five to ten times at least). In our case, the latching current is 30mA, so the circuit will work perfectly with currents above 150mA rms.
Let's assume that we are going to drive a pure resistive load, as a solderer, as the rms voltage is 110 Volt, the solderer's minimum power should be:

$$P_{min} = 110V \cdot 0.3A = 33Watt$$

This assures that the TRIAC will be able to conduce near the 100% of the hemicycle when needed.

> In the previous explanations, we have used a pulse to trigger the TRIAC, and this is an usual practice, but, is it mandatory? No, it is not, in fact you can switch on the TRIAC with a DC voltage and keep it active for an indefinite period of time.
> The importance of the TRIACs is that they allow us to drive a high voltage AC load from a low-voltage electronic circuit, as we will see next.

The basic structure of a unijunction transistor (UJT) is shown in Fig.1. It is essentially a bar of N type semiconductor material into which P type material has been diffused somewhere along its length. Contacts are then

made to the device as shown; these are referred to as the emitter, base 1 and base 2 respectively. Fig.2 shows the schematic symbol used to denote a UJT in circuit diagrams. For ease of manufacture alternative methods of making contact with the bar have been developed, giving rise to the two types of structure - bar and cube - shown in Fig.3

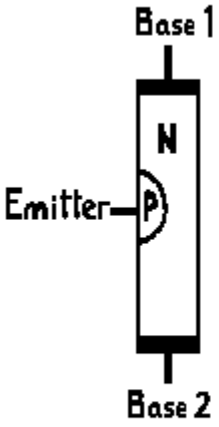
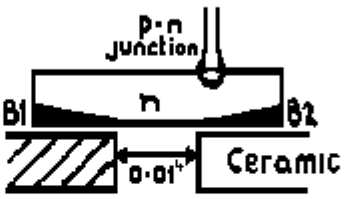
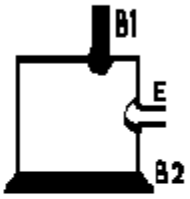


Fig.1

Fig.2



Bar structure



Cube structure

Fig.3

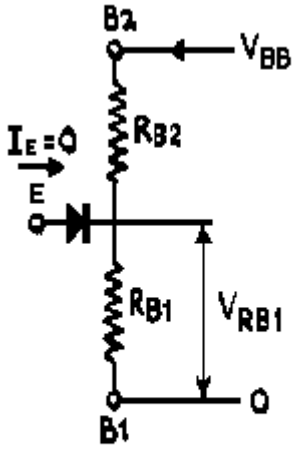


Fig.4.

The equivalent circuit shown in Fig.4 has been developed to explain how the device works, and it is necessary to define the terms used in this explanation.

R_{BB} is known as the interbase resistance, and is the sum of R_{B1} and R_{B2}:

$$R_{BB} = R_{B1} + R_{B2} \quad (1)$$

N.B. This is only true when the emitter is open circuit.

V_{RB1} is the voltage developed across R_{B1}; this is given by the voltage divider rule:

$$V_{RB1} = \frac{R_{B1}}{R_{B1} + R_{B2}} \quad (2)$$

Since the denominator of equation 2 is equal to equation 1, the former can be rewritten as:

$$VRB1 = \frac{R_{B1}}{R_{BB}} \times V_{BB} \quad (3)$$

The ratio R_{B1} / R_{BB} is referred to as the intrinsic standoff ratio and is denoted by η(the Greek letter eta).

If an external voltage V_e is connected to the emitter, the equivalent circuit can be redrawn as shown in Fig.5.

If Ve is less than V_{RB1}, the diode is reverse biased and the circuit behaves as though the emitter was open circuit. If however V_e is increased so that it exceeds V_{RB1} by at least 0.7V, the diode becomes forward biased and emitter current Ie flows into the base 1 region. Because of this, the value of R_{B1}

decreases. It has been suggested that this is due to the presence of additional charge carriers (holes) in the bar. Further increase in V_e causes the emitter current to increase which in turn reduces R_{B1} and this causes a further increase in current. This runaway effect is termed regeneration. The value of emitter voltage at which this occurs is known as the peak voltage V_P and is given by: $V_P = \eta_{AV} V_{BB} + V_D$ (4)

The characteristics of the UJT are illustrated by the graph of emitter voltage against emitter current (Fig.6).

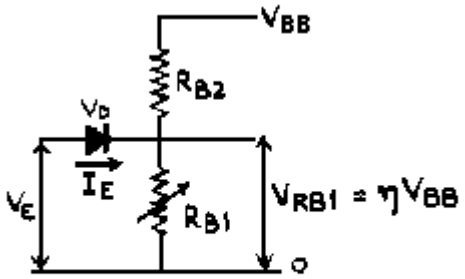


Fig.5

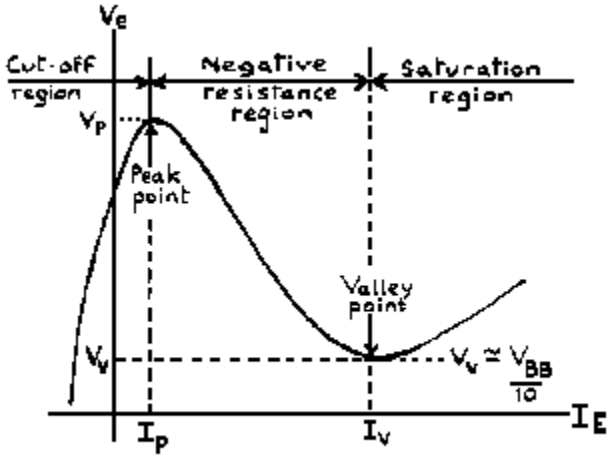


Fig.6

As the emitter voltage is increased, the current is very small - just a few microamps. When the peak point is reached, the current rises rapidly, until at

the valley point the device runs into saturation. At this point R_{B1} is at its lowest value, which is known as the saturation resistance.

The simplest application of a UJT is as a relaxation oscillator, which is defined as one in which a capacitor is charged gradually and then discharged rapidly. The basic circuit is shown in Fig.7; in the practical circuit of Fig.8 R3 limits the emitter current and provides a voltage pulse, while R2 provides a measure of temperature compensation. Fig. 9 shows the waveforms occurring at the emitter and base 1; the first is an approximation to a sawtooth and the second is a pulse of short duration.

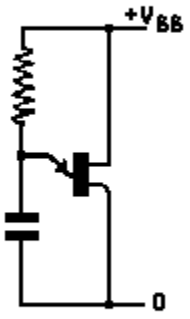


Fig.7

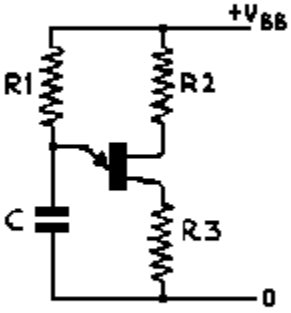


Fig.8

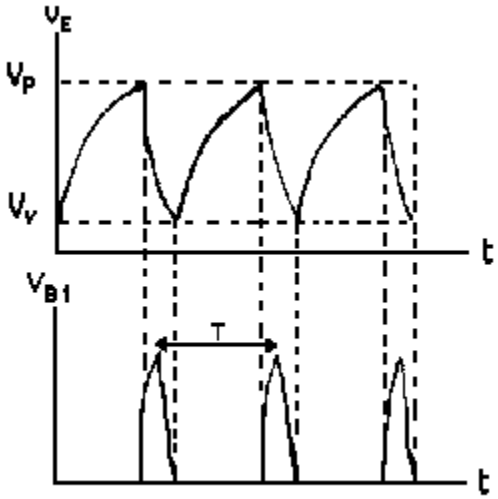


Fig.9

The operation of the circuit is as follows: C1 charges through R1 until the voltage across it reaches the peak point. The emitter current then rises rapidly, discharging C1 through the base 1 region and R3. The sudden rise of current through R3 produces the voltage pulse. When the current falls to I_V the UJT switches off and the cycle is repeated.

It can be shown that the time t between successive pulses is given by:

$$t + R1C \ln \frac{V_{BB} - V_V}{V_{BB} - V_P} \text{ secs (5) N.B. R measured in Megaohms. C in µF.}$$

Question Bank

PART –A

1.Define pinch-off voltage.

2. What is intrinsic stand-off ratio of UJT?

PART-B

1.(a) describe the different manufacturing techniques adopted to reduce the magnitude of gate source threshold voltage Vj in enhancement MOSFET.

(b) Compare the performance of JFET with MOSFET.

(c) Discuss the relative performance of p-channel MOSFET and n-channel MOSFET.

2.(a) Explain the working and characteristics of DIAC.

(b) Write short notes on MOSFET biasing.

UNIT III AMPLIFIERS

CE, CC and CB amplifiers – Small-signal low frequency transistor amplifier circuits –h-parameter representation of a transistor – Analysis of single stage transistor amplifier circuits – Voltage gain – Current gain – Input impedance and output impedance – Frequency response – RC coupled amplifier – Classification of Power amplifiers – Class A, B, AB and C Power amplifiers – Push-Pull and Complementary-Symmetry amplifiers – Design of power output, efficiency and cross-over distortion.

UNIT III

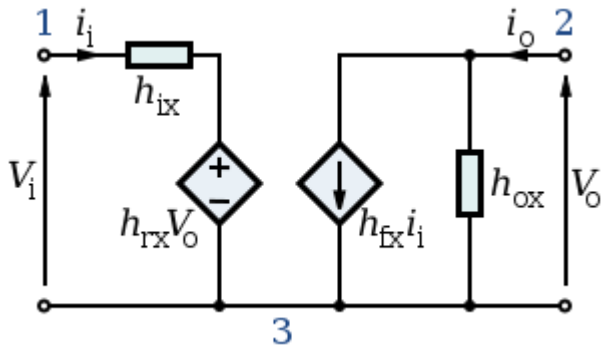
AMPLIFIERS

CE, CC and CB amplifiers

Small-signal models

hybrid-pi model

h-parameter model



Another model commonly used to analyze BJT circuits is the "h-parameter" model, closely related to the hybrid-pi model and the y-parameter two-port, but using input current and output voltage as independent variables, rather than input and output voltages. This two-port network is particularly suited to BJTs as it lends itself easily to the analysis of circuit behaviour, and may be used to develop further accurate models. As shown, the term *"x"* in the model represents a different BJT lead depending on the topology used. For common-emitter mode the various symbols take on the specific values as:

- *x* = 'e' because it is a common-emitter topology

- Terminal 1 = Base
- Terminal 2 = Collector
- Terminal 3 = Emitter
- i_i = Base current (i_b)
- i_o = Collector current (i_c)
- V_{in} = Base-to-emitter voltage (V_{BE})
- V_o = Collector-to-emitter voltage (V_{CE})

and the h-parameters are given by:

- $h_{ix} = h_{ie}$ – The input impedance of the transistor (corresponding to the emitter resistance r_e).
- $h_{rx} = h_{re}$ – Represents the dependence of the transistor's I_B–V_{BE} curve on the value of V_{CE}. It is usually very small and is often neglected (assumed to be zero).
- $h_{fx} = h_{fe}$ – The current-gain of the transistor. This parameter is often specified as h_{FE} or the DC current-gain (β_{DC}) in datasheets.
- $h_{ox} = h_{oe}$ – The output impedance of transistor. This term is usually specified as an admittance and has to be inverted to convert it to an impedance.

As shown, the h-parameters have lower-case subscripts and hence signify AC conditions or analyses. For DC conditions they are specified in upper-case. For the CE topology, an approximate h-parameter model is commonly used which further simplifies the circuit analysis. For this the h_{oe} and h_{re} parameters are neglected (that is, they are set to infinity and zero, respectively). It should also be noted that the h-parameter model as shown is suited to low-frequency, small-signal analysis. For high-frequency

analyses the inter-electrode capacitances that are important at high frequencies must be added.

Applications

The BJT remains a device that excels in some applications, such as discrete circuit design, due to the very wide selection of BJT types available, and because of its high transconductance and output resistance compared to MOSFETs. The BJT is also the choice for demanding analog circuits, especially for very-high-frequency applications, such as radio-frequency circuits for wireless systems. Bipolar transistors can be combined with MOSFETs in an integrated circuit by using a BiCMOS process of wafer fabrication to create circuits that take advantage of the application strengths of both types of transistor.

Temperature sensors

Because of the known temperature and current dependence of the forward-biased base–emitter junction voltage, the BJT can be used to measure temperature by subtracting two voltages at two different bias currents in a known ratio [3].

Logarithmic converters

Because base–emitter voltage varies as the log of the base–emitter and collector–emitter currents, a BJT can also be used to compute logarithms and anti-logarithms. A diode can also perform these nonlinear functions, but the transistor provides more circuit flexibility.

Common base

it is usually employed for amplifiers that require an unusually low input impedance, for example to act as a preamplifier for moving-coil microphones. However, it is popular in high-frequency amplifiers, for example for VHF and UHF, because its input capacitance does not suffer from the Miller effect, which degrades the bandwidth of the common-emitter configuration, and because of the relatively high isolation between the input and output. This high isolation means that there is little feedback from the output back to the input, leading to high stability.

This configuration is also useful as a current buffer since it has a current gain of approximately unity (see formulas below). Often a common base is used in this manner, preceded by a common-emitter stage. The combination of these two form the cascode configuration, which possesses several of the benefits of each configuration, such as high input impedance and isolation.

Low frequency characteristics

At low frequencies and under small-signal conditions, the circuit in Figure 1 can be represented by that in Figure 2, where the hybrid-pi model for the BJT has been employed. The input signal is represented by a Thévenin voltage source, v_s, with a series resistance R_s and the load is a resistor R_L. This circuit can be used to derive the following characteristics of the common-base amplifier.

	Definition	**Expression**	**Approximate expression**	**Conditions**

Open-circuit voltage gain	$A_v = \frac{v_o}{v_i}\Big\vert_{I}$	$\frac{(g_m r_O+1)R_C}{R_C+r_O}$	$g_m R_C$	$r_O \gg R_C$
Short-circuit current gain	$A_i = \frac{i_o}{i_i}\Big\vert_{I}$	$\frac{r_\pi+\beta r_O}{r_\pi+(\beta+1)r_O}$	1	$\beta \gg 1$
Input resistance	$R_{\text{in}} = \frac{v_i}{i_i}$	$\frac{(r_O+R_C\|R_L)r_E}{r_O+r_E+\frac{R_C\|R_L}{\beta+1}}$	$r_E \left(\approx \frac{1}{g_m}\right)$	$r_O \gg R_C\|R_L$
Output resistance	$R_{\text{out}} = \frac{v}{-}$	$R_C\|\{[1+g_m(r_\pi\|R_S)]$	$R_C \,\vert\vert\, r_O$ $R_C\vert\vert\,[(r_\pi\vert\vert R_S)($	$R_S \ll r_E$ $R_S \gg r_E$

Note: Parallel lines (||) indicate components in parallel.

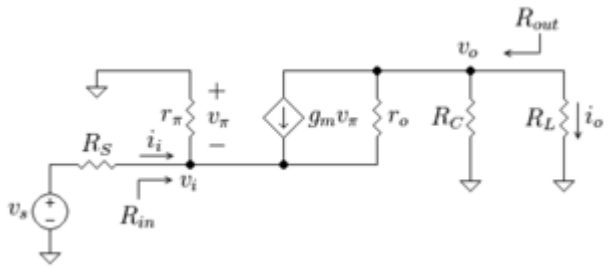
In general the overall voltage/current gain may be substantially less than the open/short circuit gains listed above (depending on the source and load resistances) due to the loading effect.

Active loads

For voltage amplification, the range of allowed output voltage swing in this amplifier is tied to its voltage gain when a resistor load R_C is employed, as in Figure 1. That is, large voltage gain requires largeR_C, and that in turn implies a large DC voltage drop across R_C. For a given supply voltage, the larger this drop, the smaller the transistor V_{CB} and the less output swing is allowed before saturation of the transistor occurs, with resultant distortion of the output signal. To avoid this situation, an active load can be used, for example, a current mirror. If this choice is made, the value of R_C in the table above is replaced by the small-signal output resistance of the active load, which is generally at least as large as the r_O of the active transistor in Figure 1. On the other hand, the DC voltage drop across the active load is a fixed low value (the **compliance voltage** of the active load), much less than the DC voltage drop incurred for comparable gain using a resistor R_C. That is, an active load imposes less restriction on the output voltage swing. Notice that active load or not, large AC gain still is coupled to large AC output resistance, which leads to poor voltage division at the output except for large loads $R_L >> R_{out}$.

For use as a current buffer, gain is not affected by R_C, but output resistance is. Because of the current division at the output, it is desirable to have an output resistance for the buffer much larger than the load R_L being driven so large signal currents can be delivered to a load. If a resistor R_C is used, as in Figure 1, a large output resistance is coupled to a large R_C, again limiting the signal swing at the output. (Even though current is delivered to the load, usually a large current signal into the load implies a large voltage swing across the load as well.) An active load provides high AC output resistance with much less serious impact upon the amplitude of output signal swing.

Voltage amplifier



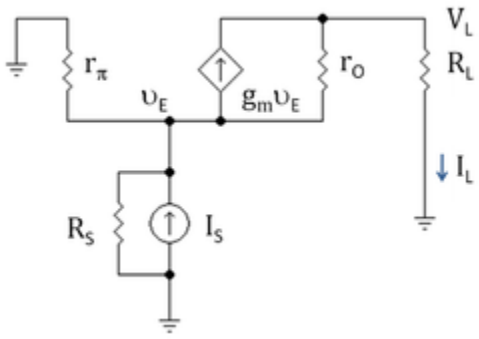
Small-signal model for calculating various parameters; Thévenin voltage source as signal.

For the case when the common-base circuit is used as a voltage amplifier, the circuit is shown in Figure

The output resistance is large, at least $R_C \parallel r_O$, the value which arises with low source impedance ($R_S << r_E$). A large output resistance is undesirable in a voltage amplifier, as it leads to poor voltage division at the output. Nonetheless, the voltage gain is appreciable even for small loads: according to the table, with $R_S = r_E$ the gain is $A_v = g_m R_L / 2$. For larger source impedances, the gain is determined by the resistor ratio R_L / R_S, and not by the transistor properties, which can be an advantage where insensitivity to temperature or transistor variations is important.

An alternative to the use of the hybrid-pi model for these calculations is a general technique based upon two-port networks. For example, in an application like this one where voltage is the output, a g-equivalent two-port could be selected for simplicity, as it uses a voltage amplifier in the output port.

For R_S values in the vicinity of r_E the amplifier is transitional between voltage amplifier and current buffer. For $R_S >> r_E$ the driver representation as a Thévenin source should be replaced by representation with a Norton source. The common-base circuit stops behaving like a voltage amplifier and behaves like a current follower, as discussed next.



Common-base circuit with Norton driver; R_C is omitted because an active load is assumed with infinite small-signal output resistance

Figure shows the common-base amplifier used as a current follower. The circuit signal is provided by an AC Norton source (current I_S, Norton resistance R_S) at the input, and the circuit has a resistor load R_L at the output.

As mentioned earlier, this amplifier is **bilateral** as a consequence of the output resistance r_O, which connects the output to the input. In this case the output resistance is large even in the worst case (it is at least $r_O \,||\, R_C$ and can become $(\beta + 1)\, r_O \,||\, R_C$ for large R_S). Large output resistance is a desirable attribute of a current source because favorable current division sends most of the current to the load. The current gain is very nearly unity as long as $R_S >> r_E$.

An alternative analysis technique is based upon two-port networks. For example, in an application like this one where current is the output, an h-equivalent two-port is selected because it uses a current amplifier in the output port.

A *common-collector* amplifier (also known as an *emitter follower* or *BJT voltage follower*) is one of three basic single-stage bipolar junction transistor (BJT) amplifier topologies, typically used as a voltage buffer. In this circuit the base terminal of the transistor serves as the input, the emitter the output, and the collector is common to both (for example, it may be tied to ground reference or a power supply rail), hence its name. The analogous field-effect transistor circuit is the common-drain amplifier.

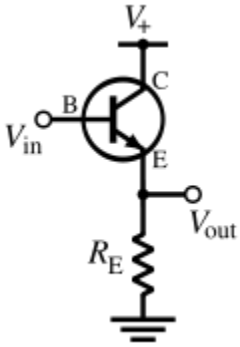
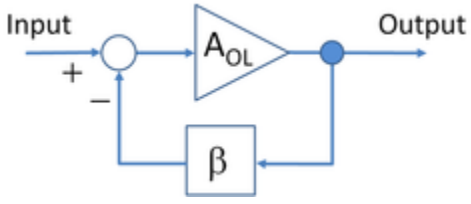


Figure 1: Basic NPN common-collector circuit (neglecting biasing details).

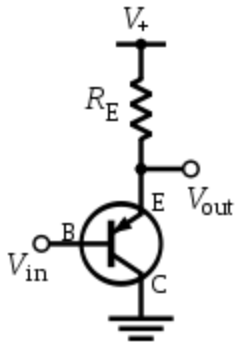
Basic circuit



The circuit can be explained by viewing the transistor as being under the control of negative feedback. From this viewpoint, a common-collector stage (Fig. 1) is an amplifier with full series negative feedback. In this configuration (Fig. 2 with $\beta = 1$), the entire output voltage V_{OUT} is placed contrary and in series with the input voltage V_{IN}. Thus the two voltages are subtracted according to KVL (the subtractor from the function block diagram is implemented just by the input loop) and their difference Vdiff = V_{IN} - V_{OUT} is applied to the base-emitter junction. The transistor monitors continuously Vdiff and adjusts its emitter voltage almost equal (less V_{BEO}) to the input voltage by passing the according collector current through the emitter resistor R_E. As a result, the output voltage follows the input voltage variations from V_{BEO} up to V_+; hence the name, emitter follower. Intuitively, this behavior can be also understood by realizing that the base-emitter voltage in the bipolar transistor is very insensitive to bias changes, so any change in base voltage is transmitted (to good approximation) directly to the emitter. It depends slightly on various disturbances (transistor tolerances, temperature variations, load resistance, collector resistor if it is added, etc.) since the transistor reacts to these disturbances and restores the equilibrium. It never saturates even if the input voltage reaches the positive rail.

The common collector circuit can be shown mathematically to have a voltage gain of almost unity:

$$A_\mathrm{v} = \frac{v_\mathrm{out}}{v_\mathrm{in}} \approx 1$$



: PNP version of the emitter follower circuit, all polarities are reversed.

A small voltage change on the input terminal will be replicated at the output (depending slightly on the transistor's gain and the value of the load resistance; see gain formula below). This circuit is useful because it has a large input impedance, so it will not load down the previous circuit:

$$r_{\mathrm{in}} \approx \beta_0 R_{\mathrm{E}}$$

and a small output impedance, so it can drive low-resistance loads:

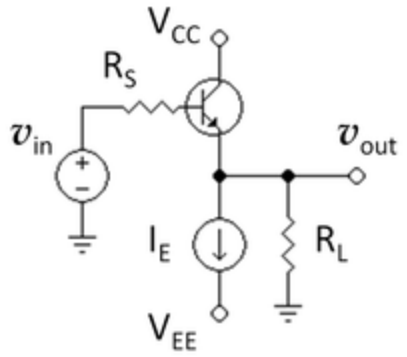
$$r_{\mathrm{out}} \approx R_{\mathrm{E}} \| \frac{R_{\mathrm{source}}}{\beta_0}$$

Typically, the emitter resistor is significantly larger and can be removed from the equation:

$$r_{\mathrm{out}} \approx \frac{R_{\mathrm{source}}}{\beta_0}$$

One aspect of buffer action is transformation of impedances. For example, the Thévenin resistance of a combination of a voltage follower driven by a voltage

Application



NPN voltage follower withcurrent source biasing suitable for integrated circuits

The low output impedance allows a source with a large output impedance to drive a small load impedance; it functions as a voltage buffer. In other words, the circuit has current gain (which depends largely on the h_{FE} of the transistor) instead of voltage gain. A small change to the input current results in much larger change in the output current supplied to the output load.

One aspect of buffer action is transformation of impedances. For example, the Thévenin resistance of a combination of a voltage follower driven by a voltage source with high Thevenin resistance is reduced to only the output resistance of the voltage follower, a small resistance. That resistance reduction makes the combination a more ideal voltage source. Conversely, a voltage follower inserted between a small load resistance and a driving stage presents a large load to the driving stage, an advantage in coupling a voltage signal to a small load.

This configuration is commonly used in the output stages of class-B and class-ABamplifier — the base circuit is modified to operate the transistor in class-B or AB mode. In class-A mode, sometimes an

active current source is used instead of R_E (Fig. 4) to improve linearity and/or efficiency.[1]

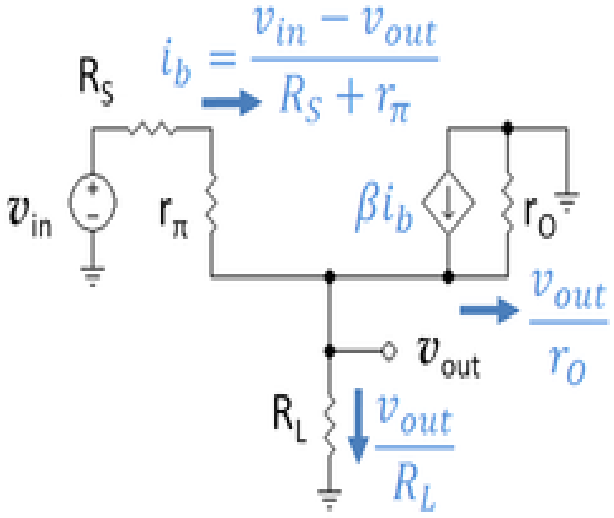
Characteristics

At low frequencies and using a simplified hybrid-pi model, the following small-signal characteristics can be derived. (Parameter $\beta = g_m r_\pi$ and the parallel lines indicate components in parallel.)

	Definition	Expression	Approximate expression	Conditions
Current gain	$A_\mathrm{i} = \frac{i_\mathrm{out}}{i_\mathrm{in}}$	$\beta_0 + 1$	$\approx \beta_0$	$\beta_0 \gg 1$
Voltage gain	$A_\mathrm{v} = \frac{v_\mathrm{ou}}{v_\mathrm{in}}$	$\frac{g_m R_\mathrm{E}}{g_m R_\mathrm{E} + 1}$	≈ 1	$g_m R_\mathrm{E} \gg 1$
Input resistance	$r_\mathrm{in} = \frac{v_{in}}{i_{in}}$	$r_\pi + (\beta_0 + 1) R_\mathrm{E}$	$\approx \beta_0 R_\mathrm{E}$	$(g_m R_\mathrm{E} \gg 1) \wedge (\beta_0 \gg$
Output resistance	$r_\mathrm{out} = \frac{v_o}{i_o}$	$R_\mathrm{E} \mid\mid \left(\frac{r_\pi + R_\mathrm{sou}}{\beta_0 + 1}$	$\approx \frac{1}{g_m} + \frac{R_\mathrm{so}}{\beta}$	$(\beta_0 \gg 1) \wedge (r_\mathrm{in} \gg R_$

Where R_source is the Thévenin equivalent source resistance.

Derivations



Small-signal circuit corresponding to Figure 3 using the hybrid-pi model for the bipolar transistor at frequencies low enough to allow neglect of bipolar device capacitances

Figure shows a low-frequency hybrid-pi model for the circuit of Figure 3. Using Ohm's law various currents have been determined and these results are shown on the diagram. Applying Kirchhoff's current law at the emitter one finds:

$$(\beta+1)\frac{v_{in}-v_{out}}{R_S+r_\pi} = v_{out}\left(\frac{1}{R_L}+\frac{1}{r_O}\right) .$$

Define the following resistance values:

$$\frac{1}{R_E} = \frac{1}{R_L}+\frac{1}{r_O}$$

$$R = \frac{R_S+r_\pi}{\beta+1} .$$

Then collecting terms the voltage gain is found as:

$$A_v = \frac{v_{out}}{v_{in}} = \frac{1}{1+\frac{R}{R_E}} .$$

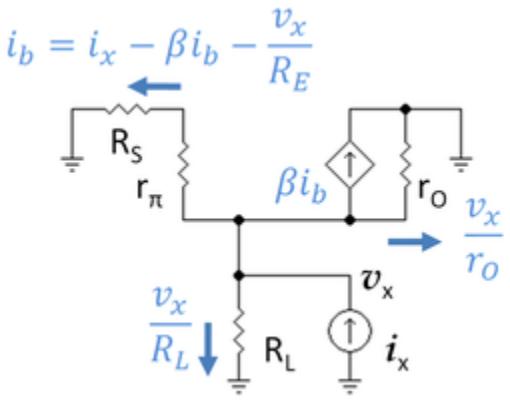
From this result the gain approaches unity (as expected for a buffer amplifier) if the resistance ratio in the denominator is small. This ratio decreases with larger values of current gain β and with larger vales of R_E. The input resistance is found as:

$$R_{in} = \frac{v_{in}}{i_b} = \frac{R_S + r_\pi}{1 - A_v}$$

$$= (R_S + r_\pi)\left(1 + \frac{R_E}{R}\right)$$

$$= R_S + r_\pi + (\beta + 1)R_E \ .$$

The transistor output resistance r_O ordinarily is large compared to the load R_L and therefore R_L dominates R_E. From this result, the input resistance of the amplifier is much larger than the output load resistance R_L for large current gain β. That is, placing the amplifier between the load and the source presents a smaller (high-resistive) load to the source than direct coupling to R_L, which results in less signal attenuation in the source impedance R_S as a consequence of voltage division



Low-frequency small-signal circuit for bipolar voltage follower with test current at output for finding output resistance. Resistor $R_E = R_L \,||\, r_O$.

Figure shows the small-signal circuit of Figure 5 with the input short-circuited and a test current placed at its output. The output resistance is found using this circuit as:

$$R_{out} = \frac{v_x}{i_x} \ .$$

Using Ohm's law, various currents have been found, as indicated on the diagram. Collecting the terms for the base current, the base current is found as:

$$(\beta + 1) i_b = i_x - \frac{v_x}{R_E} \ ,$$

where R_E is defined above. Using this value for base current, Ohm's law provides v_x as:

$$v_x = i_b (R_S + r_\pi) \ .$$

Substituting for the base current, and collecting terms,

$$R_{out} = \frac{v_x}{i_x} = R || R_E \ ,$$

where || denotes a parallel connection and R is defined above. Because R generally is a small resistance when the current gain β is large, R dominates the output impedance which therefore also is small. A small output impedance means the series combination of the original voltage source and the voltage follower presents a Thévenin voltage source with a lower Thévenin resistance at its output node; that is, the combination of voltage source with voltage follower makes a more ideal voltage source than the original one.

Single stage amplifier

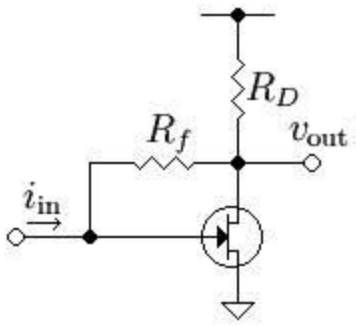
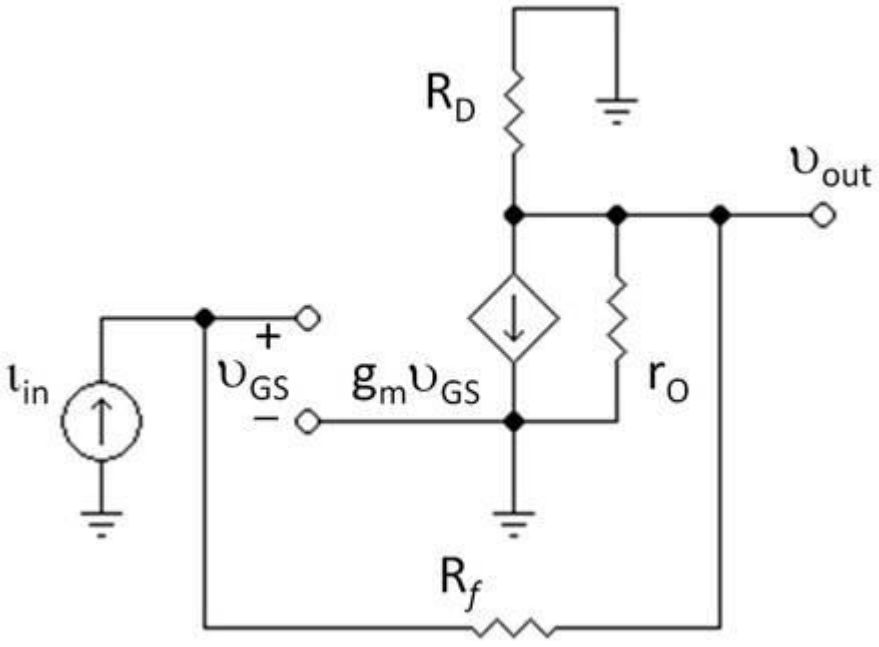
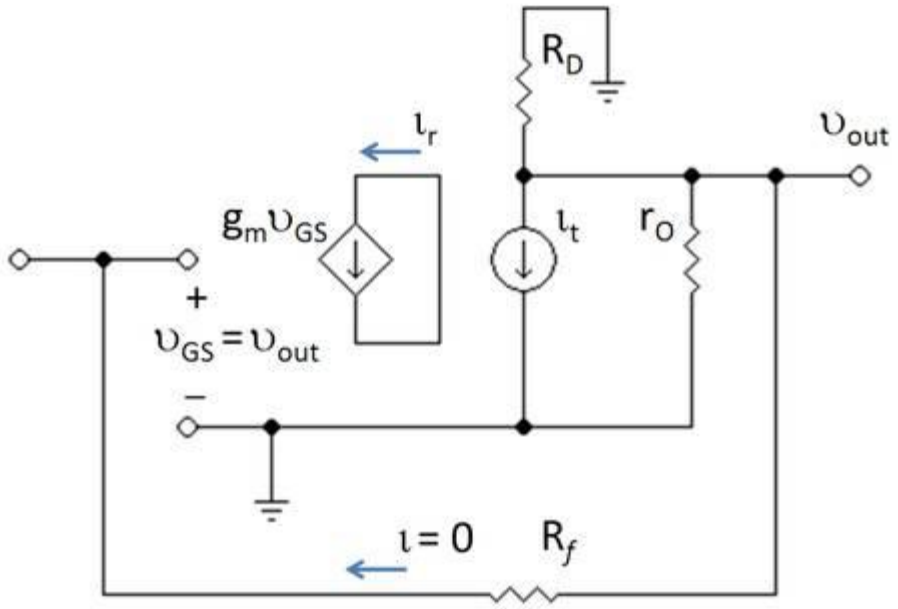


Figure 3: FET feedback amplifier

Consider the simple FET feedback amplifier in Figure 3. The aim is to find the low-frequency, open-circuit, transresistance gain of this circuit $G = v_{out} / i_{in}$ using the asymptotic gain model.

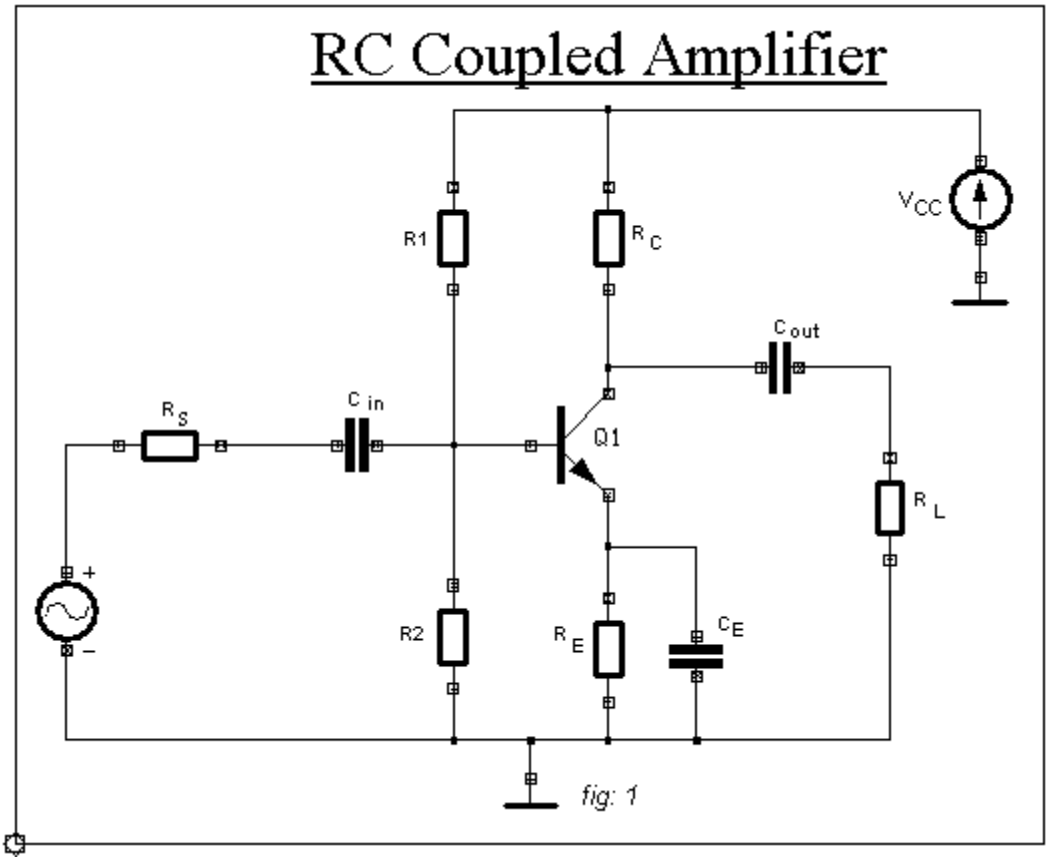


Small-signal circuit for transresistance amplifier; the feedback resistor R_f is placed below the amplifier to resemble the standard topology



Small-signal circuit with return path broken and test voltage driving amplifier at the break

The small-signal equivalent circuit is shown in Figure 4, where the transistor is replaced by its hybrid-pi model.



DC Analysis

A transistor amplifier is designed to work in the active region. A single stage amplifier is designed using voltage divider bias. In a voltage divider bias the base voltage, V_B is given by $V_B = V_{CC}\frac{R2}{R1+R2}$

$V_{CE} = 0.5V_{CC} - (1)$

$V_{RE} = 0.1V_{CC} - (2)$

$V_{RC} = 0.4V_{CC} - (3)$ First we bias the circuit in the active region with dc conditions.

$I_{R1} \cong 21_B - (4)$ i.e. Current through R1, I_{R1} is a large current of the order of 20 or 30 times that of I_B.

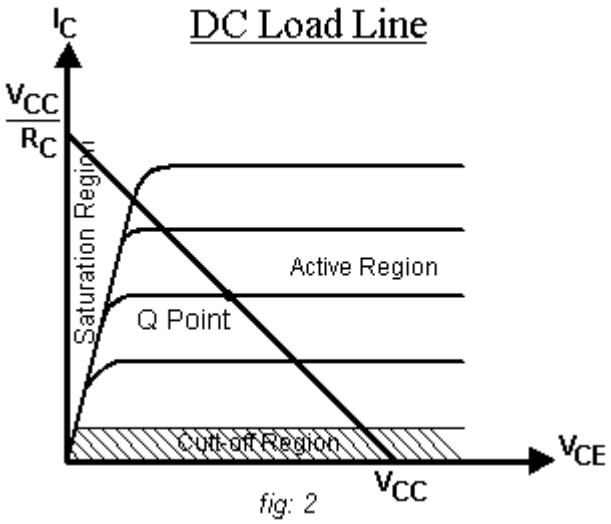


fig: 2

Current through R2, I_{R2} is given by $I_{R2} = 20I_B$ – (5)

AC Analysis

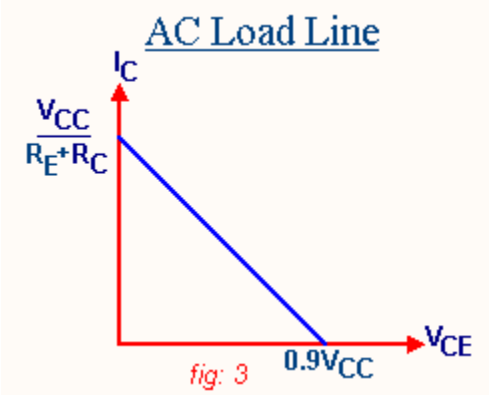


fig: 3

Modifications required from the dc circuit are

1. Coupling capacitors (C_{in}, C_{out})
2. Bypass Capacitor C_E.

The function of the coupling capacitor is to isolate the amplifier input circuit from the source. Since capacitor blocks dc it does not allow the dc

components from the input circuit to get to the base of transistor and to change the Q point.

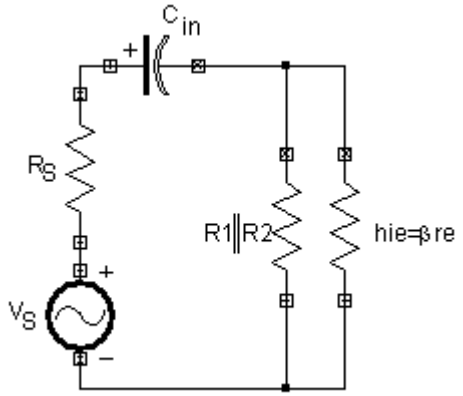


fig: 5

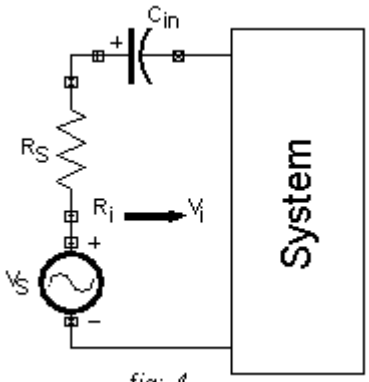


fig: 4

Additional requirement if C_{in} is that it must pass the input ac signal unattenuated. Hence the lower cut-off frequency (LCF) of the amplifier is determined by the coupling capacitor. The coupling capacitor Cin along with resistance combinations shown in fig forms a high pass filter whose cut-off frequency; $f = \frac{1}{2\pi RC} - (13)$.

The LCF due to Cin; $f_{in} = \frac{1}{2\pi(R_i + R_S)C_{in}} - (14)$

At mid or high frequencies capacitor almost acts as a short circuit.

Then V_i is related to V_S by; $V_i|_{mid} = \frac{R_i.V_S}{R_i + R_S}$

$$R_i = R_1 \| R_2 \| \beta r_e - (15)$$

$$V_i = R_i . \frac{V_S}{R_S + R_i - jXC_{in}} - (16)$$

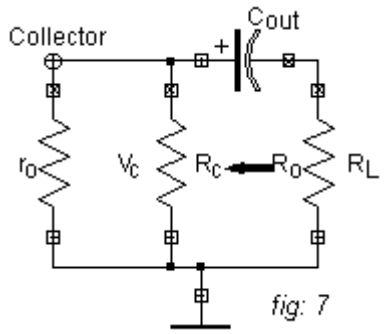


fig: 7

Effect of C_{out}

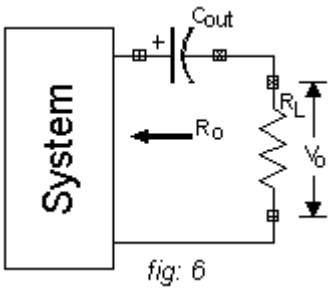


fig: 6

The effect of C_{out} on LCF is given by; $f_{Lout} = \frac{1}{2\pi(R_o + R_L)C_{out}}$

where $R_o = R_C \| r_o$.

Effect of C_E

C_E is designed such that impedance of the capacitor must be less than $^{1}/_{10}{}^{th}$ of R_E across which it is connected at LCF. The function of C_E is to bypass the emitter resistor during ac operation other wise during amplification with increase or decrease in current, the drop across R_E will increase or decrease and hence V_{BE} decreases or increases opposing the change in current. So the function of R_E is limited to dc where the capacitor will be open and R_E will stabilize the circuit. In the case of ac the capacitor will take over current.

Power amplifier classes

Angle of flow or conduction angle

Power amplifier circuits (output stages) are classified as A, B, AB and C for analog designs, and class D and E for switching designs based upon the conduction angle or *angle of flow*, Θ, of the input signal through the (or each) output amplifying device, that is, the portion of the input signal cycle during which the amplifying device conducts. The image of the conduction angle is derived from amplifying a sinusoidal signal. (If the device is always on, Θ = 360°.) The angle of flow is closely related to the amplifier power efficiency. The various classes are introduced below, followed by more detailed discussion under individual headings later on.

Class A

> 100% of the input signal is used (conduction angle Θ = 360° or 2π); i.e., the active element remains conducting[5] (works in its "linear" range) all of the time. Where efficiency is not a consideration, most small signal linear amplifiers are designed as Class A. Class A amplifiers are typically more linear and less complex than other types, but are very inefficient. This type of amplifier is most commonly used in small-signal stages or for low-power applications (such as driving headphones). Subclass A2 is sometimes used to refer to vacuum tube Class A stages where the grid is allowed to be driven slightly positive on signal peaks, resulting in slightly more power than normal Class A (A1; where the grid is always negative[6]), but incurring more distortion.

Class B

50% of the input signal is used ($\Theta = 180°$ or π; i.e., the active element works in its linear range half of the time and is more or less turned off for the other half). In most Class B, there are two output devices (or sets of output devices), each of which conducts alternately (push–pull) for exactly 180° (or half cycle) of the input signal; selective RF amplifiers can also be implemented using a single active element.

These amplifiers are subject to *crossover distortion* if the transition from one active element to the other is not perfect, as when two complementary transistors (i.e., one PNP, one NPN) are connected as two emitter followers with their base and emitter terminals in common, requiring the base voltage to slew across the region where both devices are turned off.[7]

Class AB

Here the two active elements conduct more than half of the time as a means to reduce the cross-over distortions of Class B amplifiers. In the example of the complementary emitter followers a bias network allows for more or less quiescent current thus providing an operating point somewhere between Class A and Class B. Sometimes a figure is added (e.g., AB_1 or AB_2) for vacuum tube stages where the grid voltage is always negative with respect to the cathode (Class AB_1) or may be slightly positive (hence drawing grid current, adding more distortion, but giving slightly higher output power) on signal peaks (Class AB_2); another interpretation being higher figures implying a higher quiescent current and therefore more of the properties of Class A.

Class C

Less than 50% of the input signal is used (conduction angle $\Theta < 180°$). The advantage is potentially high efficiency, but a disadvantage is high distortion.

Class D

Main article: Switching amplifier

These use switching to achieve a very high power efficiency (more than 90% in modern designs). By allowing each output device to be either fully on or off, losses are minimized. The analog output is created by pulse-width modulation; i.e., the active element is switched on for shorter or longer intervals instead of modifying its resistance. There are more complicated switching schemes like sigma-delta modulation, to improve some performance aspects like lower distortions or better efficiency.

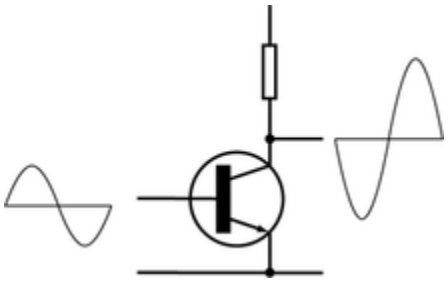
Additional classes

There are several other amplifier classes, although they are mainly variations of the previous classes. For example, Class G and Class H amplifiers are marked by variation of the supply rails (in discrete steps or in a continuous fashion, respectively) following the input signal. Wasted heat on the output devices can be reduced as excess voltage is kept to a minimum. The amplifier that is fed with these rails itself can be of any class. These kinds of amplifiers are more complex, and are mainly used for specialized applications, such as very high-power units. Also, Class E and Class F amplifiers are commonly described in literature for radio frequencies applications where efficiency of the traditional classes in are important, yet several aspects not covered elsewhere (e.g.: amplifiers often simply said to have a gain of x dB - so what power gain?) deviate substantially from their ideal values.

These classes use harmonic tuning of their output networks to achieve higher efficiency and can be considered a subset of Class C due to their conduction angle characteristics.

The classes can be most easily understood using the diagrams in each section below. For the sake of illustration, a bipolar junction transistor is shown as the amplifying device, but in practice this could be a MOSFET or vacuum tube device. In an analog amplifier (the most common kind), the signal is applied to the input terminal of the device (base, gate or grid), and this causes a proportional output drive current to flow out of the output terminal. The output drive current comes from the power supply.

Class A



Class A amplifier

Amplifying devices operating in Class A conduct over the whole of the input cycle such that the output signal is an exact scaled-up replica of the input with no clipping. A *Class A amplifier* (or operational amplifier) is distinguished by the *output stage* (and perhaps the driver) device(s) being biased into Class A; even Class AB and B amplifiers normally have early stages operating in Class A. Class A is the usual means of implementing small-signal amplifiers, so the term *Class A design* applied to equipment

such as preamplifiers (for example, in recording studios) implies not so much their *use* of Class A, but that their sound is top quality - good enough to be matched with top quality Class A power amplifiers.

Advantages of Class A Amplifiers

- Class A designs are simpler than other classes; for example Class AB and B designs require two devices (push-pull output) to handle both halves of the waveform, and circuitry to keep the quiescent bias optimal during temperature changes; Class A can use either single-ended or push-pull and bias is usually less critical.
- The amplifying element is biased so the device is always conducting to some extent, normally implying the quiescent (small-signal) collector current (for transistors; drain current for FETs or anode/plate current for vacuum tubes) is close to the most linear portion (sometimes called the "sweet spot") of its characteristic curve (known as its transfer characteristic or transconductance curve), giving the least audio distortion.
- Because the device is never shut off completely there is no "turn on" time, little problem with charge storage, and generally better high frequency performance and feedback loop stability (and usually fewer high-order harmonics).
- The point at which the device comes closest to being cut off (and so significant change in gain, hence non-linearity) is not close to zero signal, so the problem of crossover distortion associated with Class AB and B designs is avoided, even in Class A double-ended stages.

Disadvantage of Class A Amplifiers

- They are very inefficient; a theoretical maximum of 50% is obtainable with inductive output coupling and only 25% with capacitive coupling, unless Square law output stages are used. In a power amplifier this not only wastes power and limits battery operation, it may place restrictions on the output devices that can be used (for example: ruling out some audio triodes if modern low-efficiency loudspeakers are to be used), and will increase costs. Inefficiency comes not just from the fact that the device is always conducting to some extent (that happens even with Class AB, yet its efficiency can be close to that of Class B); it is that the standing current is roughly half the maximum output current (although this can be less with Square law output stage), together with the problem that a large part of the power supply voltage is developed across the output device at low signal levels (as with Classes AB and B, but unlike output stages such as Class D). If high output powers are needed from a Class A circuit, the power waste (and the accompanying heat) will become significant. For every watt delivered to the load, the amplifier itself will, *at best*, dissipate another watt. For large powers this means very large and expensive power supplies and heat sinking.

Class A designs have largely been superseded by the more efficient designs for power amplifiers, though they remain popular with some hobbyists, mostly for their simplicity. Also, many audiophiles believe that Class A gives the best sound quality (for their absence of crossover distortion and reduced odd-harmonic and high-order harmonic distortion) which provides a small market for expensive **high fidelity** Class A amps.

Single-Ended and Triode Class A Amplifiers

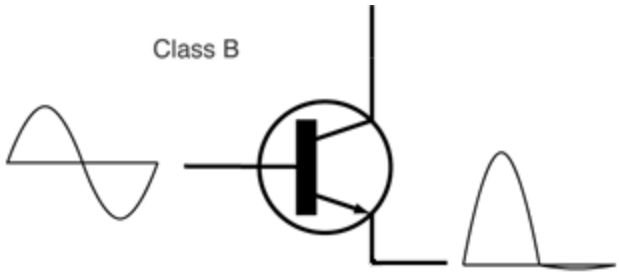
Some aficionados who prefer Class A amplifiers also prefer the use of thermionic valve (or "tube") designs instead of transistors, especially in Single-ended triode output configurations for several claimed reasons:

- Single-ended output stages (be they tube or transistor) have an asymmetrical transfer function, meaning that even harmonics in the created distortion tend not to be canceled (as they are in push-pull output stages); by using tubes OR FETs most of the distortion is from the square law transfer characteristic and so second-order, which some consider to be "warmer" and more pleasant.[8][9]
- For those who prefer low distortion figures, the use of tubes with Class A (generating little odd-harmonic distortion, as mentioned above) together with symmetrical circuits (such as push-pull output stages, or balanced low-level stages) results in the cancellation of most of the even distortion harmonics, hence the removal of most of the distortion.
- Though good amplifier design can reduce harmonic distortion patterns to almost nothing, distortion is essential to the sound of electric guitar amplifiers, for example, and is held by recording engineers to offer more flattering microphones and to enhance "clinical-sounding" digital technology.
- Historically, valve amplifiers often used a Class A power amplifier simply because valves are large and expensive; many Class A designs use only a single device.

Transistors are much cheaper, and so more elaborate designs that give greater efficiency but use more parts are still cost-effective. A classic application for a pair of class A devices is the long-tailed pair, which is exceptionally linear, and forms the basis of many more complex circuits, including many audio amplifiers and almost all op-amps. Class A amplifiers are often used in output stages of high quality op-amps (although the accuracy of the bias in low cost op-amps such as the **741** may result in Class A or Class AB or Class B, varying from device to device or with temperature). They are sometimes used as medium-power, low-efficiency, and high-cost audio amplifiers. The power consumption is unrelated to the output power. At idle (no input), the power consumption is essentially the same as at high output volume. The result is low efficiency and high heat dissipation.

Class B and AB

Class B amplifiers only amplify half of the input wave cycle, thus creating a large amount of distortion, but their efficiency is greatly improved and is much better than Class A. Class B has a maximum theoretical efficiency of 78.5% (i.e., $\pi/4$). This is because the amplifying element is switched off altogether half of the time, and so cannot dissipate power. A single Class B element is rarely found in practice, though it has been used for driving the loudspeaker in the early IBM Personal Computers with beeps, and it can be used in RF power amplifier where the distortion levels are less important. However Class C is more commonly used for this.



Class B Amplifier

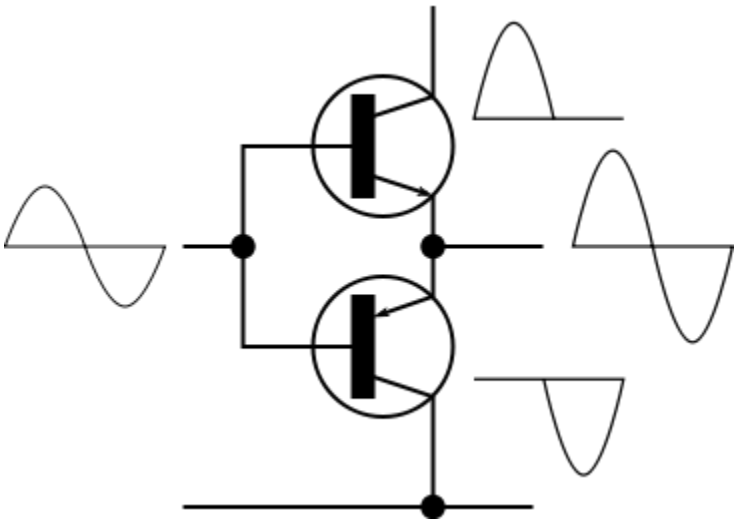
A practical circuit using Class B elements is the push-pull stage, such as the very simplified complementary pair arrangement shown below. Here, complementary or quasi-complementary devices are each used for amplifying the opposite halves of the input signal, which is then recombined at the output. This arrangement gives excellent efficiency, but can suffer from the drawback that there is a small mismatch in the cross-over region - at the "joins" between the two halves of the signal, as one output device has to take over supplying power exactly as the other finishes. This is called crossover distortion. An improvement is to bias the devices so they are not completely off when they're not in use. This approach is called *Class AB* operation.

In Class AB operation, each device operates the same way as in Class B over half the waveform, but also conducts a small amount on the other half. As a result, the region where both devices simultaneously are nearly off (the "dead zone") is reduced. The result is that when the waveforms from the two devices are combined, the crossover is greatly minimized or eliminated altogether. The exact choice of **quiescent current**, the standing current through both devices when there is no signal, makes a large difference to the level of distortion (and to the risk of thermal runaway, that may damage the

devices); often the bias voltage applied to set this quiescent current has to be adjusted with the temperature of the output transistors (for example in the circuit at the beginning of the article the diodes would be mounted physically close to the output transistors, and chosen to have a matched temperature coefficient). Another approach (often used as well as thermally-tracking bias voltages) is to include small value resistors in series with the emitters.

Push-Pull and Complementary-Symmetry amplifiers

Class AB sacrifices some efficiency over class B in favor of linearity, thus is less efficient (below 78.5% for full-amplitude sinewaves in transistor amplifiers, typically; much less is common in Class AB vacuum tube amplifiers). It is typically much more efficient than class A.



Class B push–pull amplifier

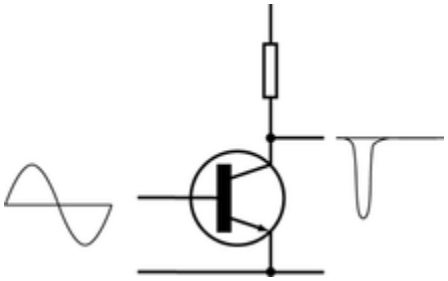
Class B or AB push–pull circuits are the most common design type found in audio power amplifiers. Class AB is widely considered a good compromise for audio amplifiers, since much of the time the music is quiet enough that

the signal stays in the "class A" region, where it is amplified with good fidelity, and by definition if passing out of this region, is large enough that the distortion products typical of class B are relatively small. The crossover distortion can be reduced further by using negative feedback. Class B and AB amplifiers are sometimes used for RF linear amplifiers as well. Class B amplifiers are also favored in battery-operated devices, such as transistor radios.

Digital Class B

A limited power output Class-B amplifier with a single-ended supply rail of 5±0.5 V.

Class C



Class C amplifier

Class C amplifiers conduct less than 50% of the input signal and the distortion at the output is high, but high efficiencies (up to 90%) are possible. Some applications (for example, megaphones) can tolerate the distortion. A much more common application for Class C amplifiers is in RF transmitters, where the distortion can be vastly reduced by using tuned loads on the amplifier stage. The input signal is used to roughly switch the

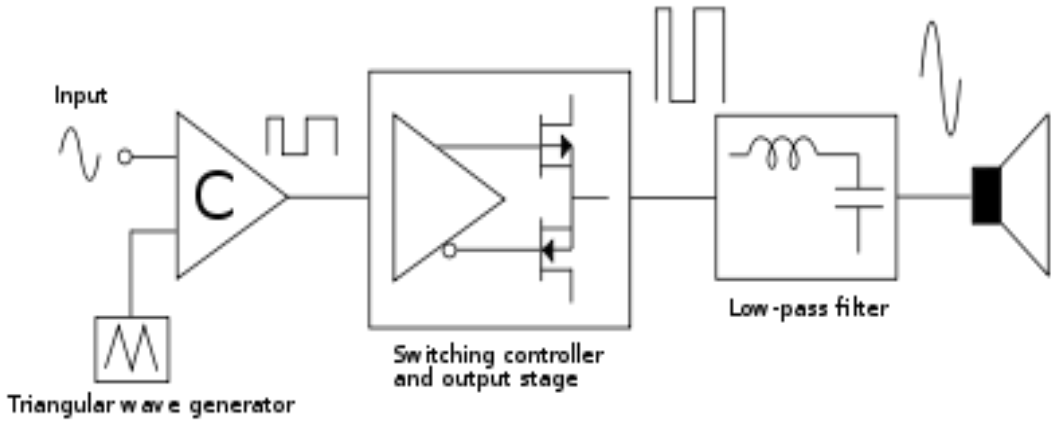
amplifying device on and off, which causes pulses of current to flow through a tuned circuit.

The Class C amplifier has two modes of operation: tuned and untuned.[10] The diagram shows a waveform from a simple class C circuit without the tuned load. This is called untuned operation, and the analysis of the waveforms shows the massive distortion that appears in the signal. When the proper load (e.g., a pure inductive-capacitive filter) is used, two things happen. The first is that the output's bias level is clamped, so that the output variation is centered at one-half of the supply voltage. This is why tuned operation is sometimes called a *clamper*. This action of elevating bias level allows the waveform to be restored to its proper shape, allowing a complete waveform to be re-established despite having only a one-polarity supply. This is directly related to the second phenomenon: the waveform on the center frequency becomes much less distorted. The distortion that is present is dependent upon the bandwidth of the tuned load, with the center frequency seeing very little distortion, but greater attenuation the farther from the tuned frequency that the signal gets.

The tuned circuit will only resonate at particular frequencies, and so the unwanted frequencies are dramatically suppressed, and the wanted full signal (sine wave) will be extracted by the tuned load (e.g., a high-quality bell will ring at a particular frequency when it is hit periodically with a hammer). Provided the transmitter is not required to operate over a very wide band of frequencies, this arrangement works extremely well. Other residual harmonics can be removed using a filter.

Class D

Class D Amplifier



Boss Audio Class D mono car audio amplifier with a low pass filter for powering subwoofers

In the Class D amplifier the input signal is converted to a sequence of higher voltage output pulses. The averaged-over-time power values of these pulses are directly proportional to the instantaneous amplitude of the input signal. The frequency of the output pulses is typically ten or more times the highest frequency in the input signal to be amplified. The output pulses contain inaccurate spectral components (that is, the pulse frequency and its harmonics) which must be removed by a low-pass passive filter. The resulting filtered signal is then an amplified replica of the input.

These amplifiers use pulse width modulation, pulse density modulation (sometimes referred to as pulse frequency modulation) or more advanced form of modulation such as Delta-sigma modulation (for example, in the Analog Devices AD1990 Class-D audio power amplifier). Output stages such as those used in pulse generators are examples of Class D amplifiers.

The term Class D is usually applied to devices intended to reproduce signals with a bandwidth well below the switching frequency.

Class D amplifiers can be controlled by either analog or digital circuits. The digital control introduces additional distortion called *quantization error* caused by its conversion of the input signal to a digital value.

The main advantage of a Class D amplifier is power efficiency. Because the output pulses have a fixed amplitude, the switching elements (usually MOSFETs, but valves and bipolar transistors were once used) are switched either completely on or completely off, rather than operated in linear mode. A MOSFET operates with the lowest resistance when fully-on and thus has the lowest power dissipation when in that condition, except when fully off. When operated in a linear mode the MOSFET has variable amounts of resistance that vary linearly with the input voltage and the resistance is something other than the minimum possible, therefore more electrical energy is dissipated as heat. Compared to Class A/B operation, Class D's lower losses permit the use of a smaller heat sink for the MOSFETS while also reducing the amount of AC power supply power required. Thus, Class D amplifiers do not need as large or as heavy power supply transformers or heatsinks, so they are smaller and more compact in size than an equivalent Class AB amplifier.

Class D amplifiers have been widely used to control motors, and almost exclusively for small DC motors, but they are now also used as audio amplifiers, with some extra circuitry to allow analogue to be converted to a much higher frequency pulse width modulated signal. The relative difficulty of achieving good audio quality means that nearly all are used in

applications where quality is not a factor, such as modestly-priced bookshelf audio systems and "DVD-receivers" in mid-price home theater systems.

High quality Class D audio amplifiers are now, however, starting to appear in the market:

- Tripath have called their revised Class D designs Class T.
- Bang and Olufsen's ICEPower Class D system has been used in the Alpine PDX range and some of Pioneer's PRS range and for other manufacturers' equipment.

These revised designs have been said to rival good traditional AB amplifiers in terms of quality.

Before these higher quality designs existed an earlier use of Class D amplifiers and prolific area of application was high-powered, subwoofer amplifiers in cars. Because subwoofers are generally limited to a bandwidth of no higher than 150 Hz, the switching speed for the amplifier does not have to be as high as for a full range amplifier. The drawback with Class D designs being used to power subwoofers is that their output filters (typically inductors that convert the pulse width signal back into an analogue waveform) lower the damping factor of the amplifier.

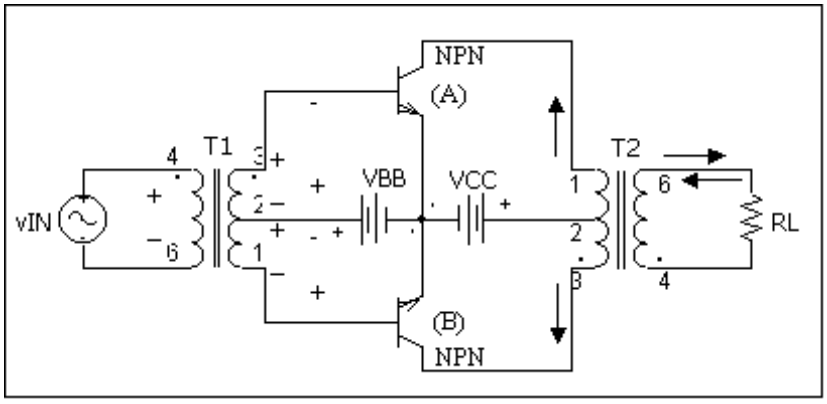
This means that the amplifier cannot prevent the subwoofer's reactive nature from lessening the impact of low bass sounds (as explained in the feedback part of the Class AB section). Class D amplifiers for driving subwoofers are relatively inexpensive, in comparison to Class AB amplifiers. A 1000 W Class D subwoofer amplifier that can operate at about 80% to 95%

efficiency costs about US$250, much less than a Class AB amplifier of this power, which would cost several thousand dollars.

The letter D used to designate this amplifier class is simply the next letter after *C*, and does not stand for digital. Class D and Class E amplifiers are sometimes mistakenly described as "digital" because the output waveform superficially resembles a pulse-train of digital symbols, but a Class D amplifier merely converts an input waveform into a continuously pulse-width modulated (square wave) analog signal. (A digital waveform would be pulse-code modulated.)

POWER OUTPUT CIRCUITS

PUSH-PULL AMPLIFIERS



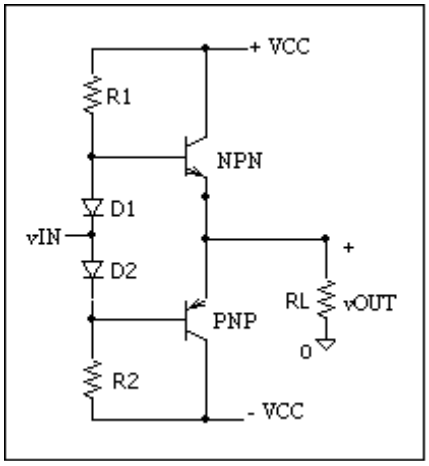
*V*BB determines whether the stage will operate in Class-A, B, AB, or C . Typically all audio amplifiers are biased to operate in Class-AB in order to minimize the effects of cross-over distortion at low drive levels and to benefit from the high efficiency of Class-B operation at medium and high volume. In this push-pull amplifier, *QA* will turn on and amplify when *v*IN is positive; *QB* will turn on and amplify when *v*IN is negative. The center-

tapped transformer, T1 makes sure the transistors get the right polarity to turn on (both require positive voltage on their bases to turn on) but receive alternate halves of *v*IN . The center-tapped transformer, T2 superimposes the amplified halves by combining them with opposite phases, thereby correcting for the phase reversal introduced by T1. *RL* receives the amplified halves superimposed with the correct polarity for each, therefore the whole waveform is reproduced on it at amplified levels. Advantages: (1) No need for large capacitors, (2) No need for complementary devices, (3) Suitable for any type power

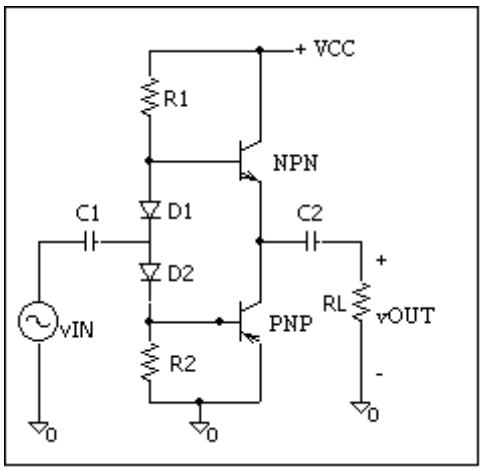
device, including vacuum tubes, (4) tranformer coupling at the input allows matching of amplifier input impedanceto the source/driver impedance and (5) transformer coupling at the output allows optimization of power coupling to any load impedance.

Disadvantages: (1) Transformer coupling: heavy, bulky, costly, and limited frequency response.

COMPLEMENTARY SYMMETRY AMPLIFIERS



(a)



(b)

Figure (a) Dual supply DC-coupled Complementary Symmetry Emitter Follower Output Stage

(b) Single Supply AC-coupled Complementary Symmetry Emitter Follower Output Stage

Complementary output stages rely on two complementary and "matched" (i.e. symmetrical) transistors to split and share the task of amplifying an AC signal, positive half by one and the negative half of the signal by its complementary pair. Figure (a) shows a complementary symmetry emitter follower circuit biased with two diodes to operate in class- AB mode. Note the symmetry (anti-symmetry) with respect to a line drawn at the center from the input to the output terminal. Upper half of the circuit responds to the positive half of an input, *v*IN . A positive *v*IN raises the voltage on the base of the NPN, forcing its emitter to follow and deliver a positive current into the load *RL* from the positive +*V*CC through the NPN. The NPN transistor cannot supply a negative current into *RL* and turns off when *v*IN becomes negative. For the negative half cycle of *v*IN , the PNP half of the circuit takes over and helps to create a reproduction of *v*IN on *RL* by

delivering the negative current to *RL* from the negative supply *V*EE = - *V*CC . This circuit can supply undistorted peak swings to *RL* by as much as (*V*CC - *V*CESAT) . For sinusoidal waveforms the ac power delivered to the load can be as high as H *V*CC- *V*CESAT L2 Figure (b) gives the single supply version of the complementary symmetry circuit. Note that both the input and the output terminals of this circuit are expected to be at *V*CC /2 , therefore creating a need to ac-couple the input as well as the output. The coupling capacitor needed at the output may have to be very large if high ac powers are to be deliveredwhile operating from a small *V*CC voltage power supply.

Question Bank

PART –A

1.What is emitter follower?

2. What are class AB Amplifiers?

PART-B

1.Obtain the current gain, voltage gain, output admittance and power gain of AC commojn emitter amplifier using h-parameter analysis. Also give the relation between AVs and AIs.

2. A CC amplifier (emitter follower) is fed from a voltage source Vs of internal resistance Rs=800ohm. The load impedance is a resistor Ro=1600ohm.The CE h-parameters are hie=1000ohm, hre=2.2X10-4, hfe=55 and hoe=23μὺ .Compute the current gain A1,input resistance Ri,voltage gain Av, output resistance Ro and output terminal resistance Rot using (a) exact analysis and (b) approximate analysis.

3. Explain the working of a transistorized Hartley oscillator. Derive expressions for frequency of oscillation and condition for sustained oscillation.

UNIT IV

FEEDBACK AMPLIFIERS AND OSCILLATORS

Advantages of negative feedback – Voltage/current, series/shunt feedback – Positive feedback – Conditions for oscillation – Phase shift – Wein Bridge – Hartley – Colpitts and Crystal oscillators.

UNIT IV

FEEDBACK AMPLIFIERS AND OSCILLATORS

Feedback is a process of injecting some energy from the output and then return it to the input .The amplifier which uses this is called feedback amplifier.

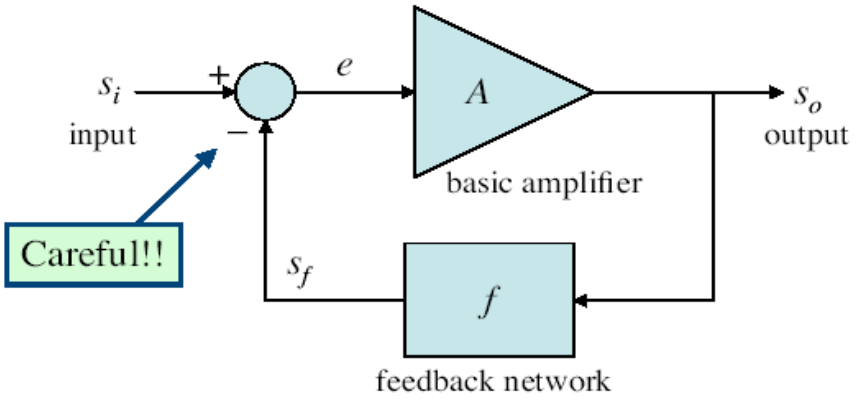
Classification of feedback amplifiers

1. Positive Feedback amplifiers
2. Negative Feedback amplifiers

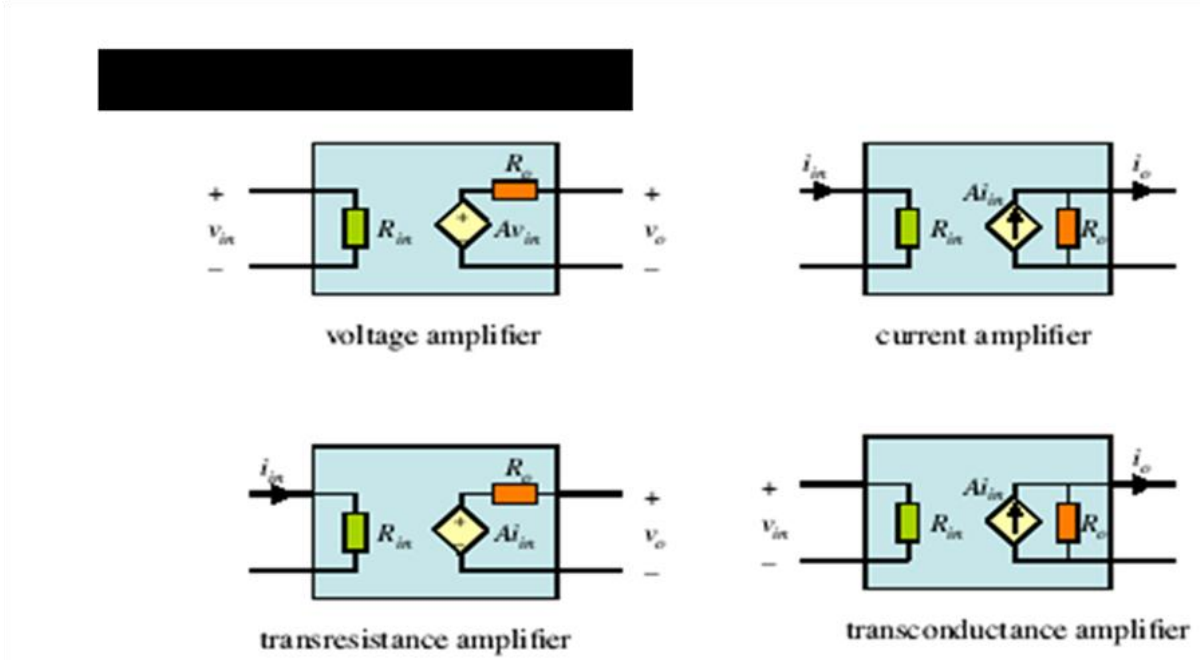
Contents

Basic feedback configuration

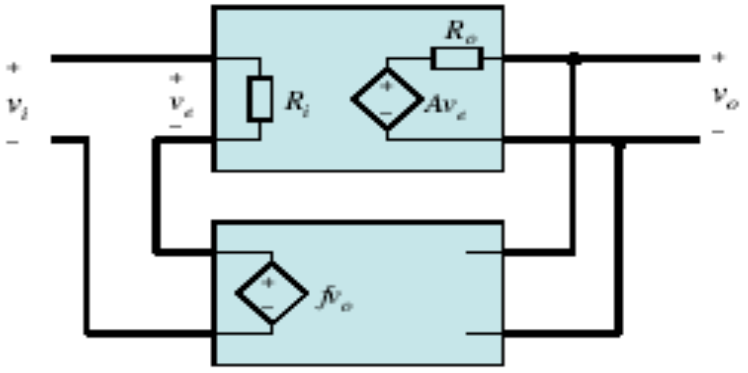
The basic feedback amplifier consists of a *basic amplifier* and a *feedback network.*



A = basic amplifier gain
f = feedback gain



Series-shunt feedback (for voltage amplifier)

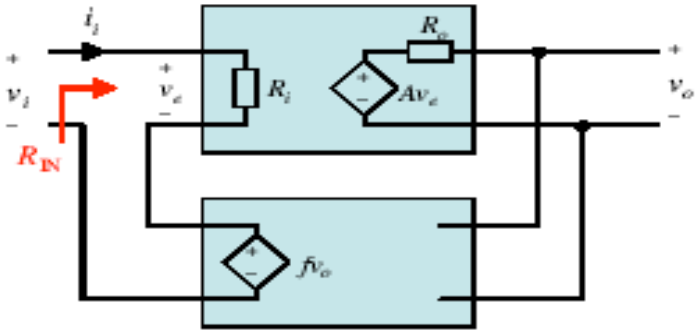


Overall gain (closed-loop gain) : $A_o = \frac{v_o}{v_i} = \frac{A}{1 + Af}$

Series-shunt feedback (for voltage amplifier)

To find the input resistance, we consider the ratio of v_i and i_i, with output opened.

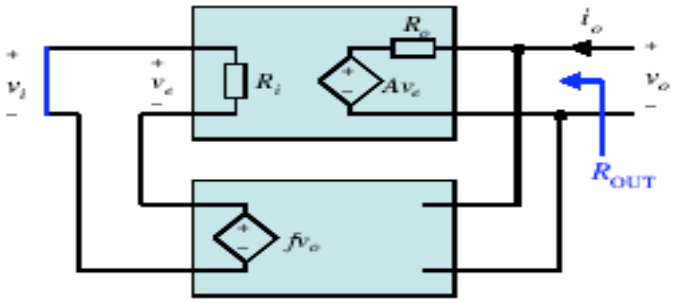
$$R_{IN} = \frac{v_i}{i_i} = \frac{v_i}{v_e / R_i}$$
$$= R_i \frac{v_e + f v_o}{v_e}$$
$$= R_i (1 + Af)$$



The input resistance has been enlarged by $(1+Af)$. This is a desirable feature for voltage amplifier as a large input resistance minimizes loading effect to the previous stage.

Series-shunt feedback (for voltage amplifier)

To find the output resistance, we consider shorting the input source and calculate the ratio of v_o and i_o.



First, we have $v_e = -fv_o$.
Also,

$$i_o = \frac{v_o - Av_e}{R_o} = \frac{v_o + Afv_o}{R_o}$$

Hence,

$$R_{OUT} = \frac{v_o}{i_o} = \frac{R_o}{1 + Af}$$

The output resistance has been reduced by $(1+Af)$. This is a desirable feature for voltage amplifier as a small output resistance emulates a better voltage source for the load.

Series-shunt feedback (for voltage amplifier)

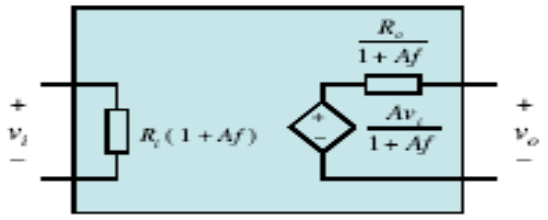
Summary of features

Closed-loop gain = $\frac{A}{1+Af} \approx \frac{1}{f}$

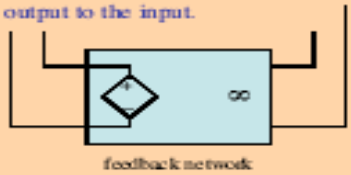
Input resistance = $R_i(1+Af)$

Output resistance = $\frac{R_o}{1+Af}$

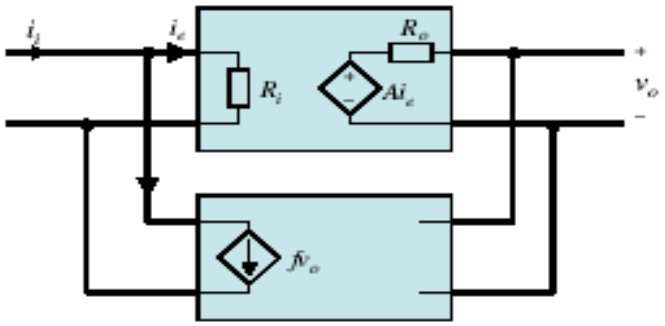
Equivalent model



NOTE: We did not consider loading effect of the feedback network, i.e., we assume that the feedback network is an ideal amplifier which feeds a scaled-down copy of the output to the input.



Shunt-shunt feedback (for transresistance amplifier)

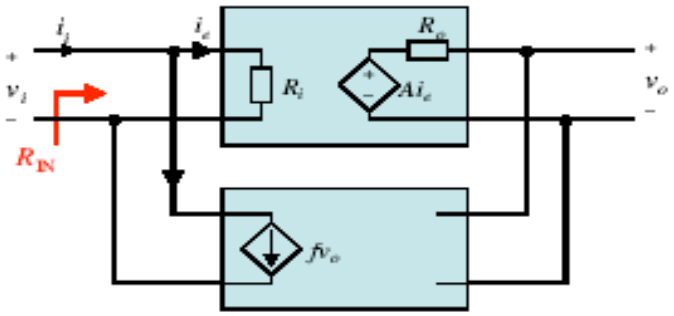


Overall gain (closed-loop gain) : $A_o = \frac{v_o}{i_i} = \frac{A}{1+Af}$

Shunt-shunt feedback (for transresistance amplifier)

To find the input resistance, we consider the ratio of v_i and i_i, with output opened.

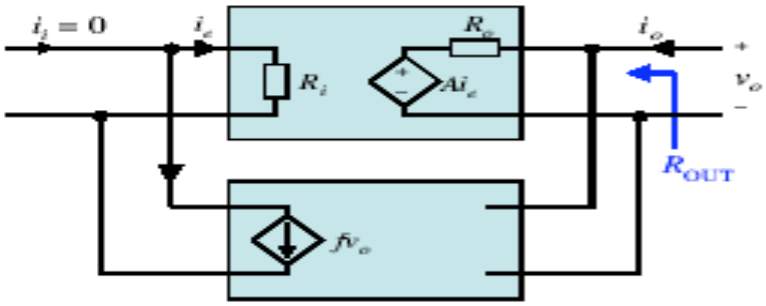
$$R_{IN} = \frac{v_i}{i_i} = \frac{R_i i_e}{i_i}$$
$$= R_i \frac{i_e}{i_e + fv_o}$$
$$= \frac{R_i}{1 + Af}$$



The input resistance has been reduced by $(1+Af)$. This is a desirable feature for transresistance amplifier as a small input resistance ensures better current sensing from the previous stage.

Shunt-shunt feedback (for transresistance amplifier)

To find the output resistance, we consider opening the input source (putting $i_i = 0$) and calculate the ratio of v_o and i_o.



First, we have $i_e = -fv_o$.
Also,

$$i_o = \frac{v_o - Ai_e}{R_o} = \frac{v_o + Afv_o}{R_o}$$

Hence,

$$R_{OUT} = \frac{v_o}{i_o} = \frac{R_o}{1 + Af}$$

The output resistance has been reduced by $(1+Af)$. This is a desirable feature for transresistance amplifier as a large small resistance emulates a better voltage source for the load.

Shunt-shunt feedback (for transresistance amplifier)

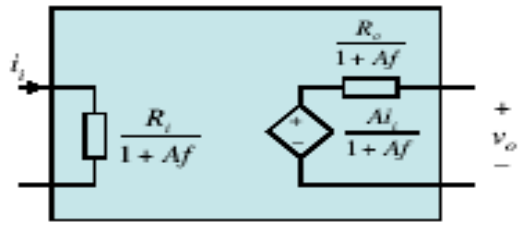
Summary of features

Closed-loop gain = $\frac{A}{1+Af} \approx \frac{1}{f}$

Input resistance = $\frac{R_i}{1+Af}$

Output resistance = $\frac{R_o}{1+Af}$

Equivalent model

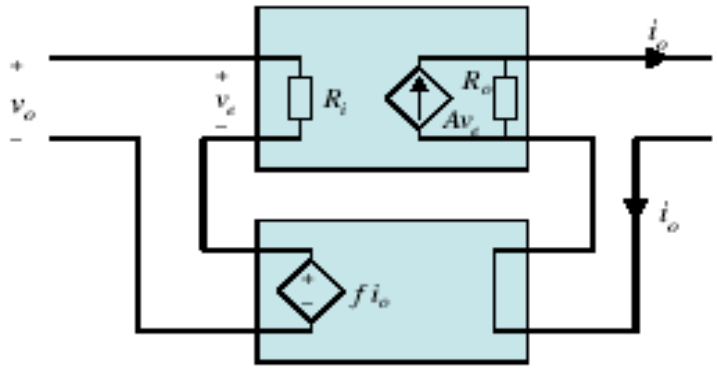


Similar, we can develop the feedback configurations for transconductance amplifier and current amplifier.

Transconductance amplifier: series-series feedback

Current amplifier: shunt-series feedback

Series-series feedback (for transconductance amplifier)

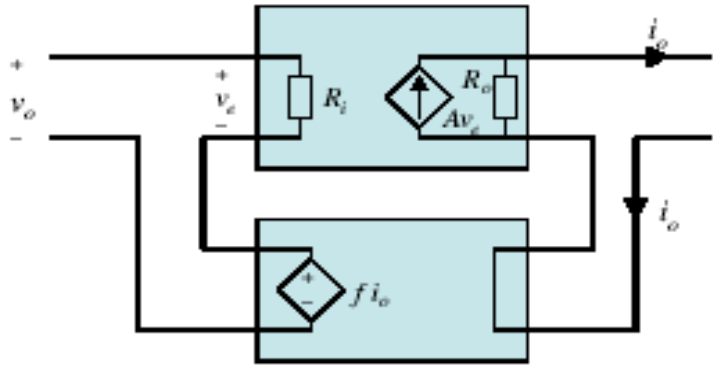


Overall gain (closed-loop gain) : $A_o = \frac{i_o}{v_i} = \frac{A}{1+Af}$

Input resistance: $R_{IN} = R_i(1+Af)$ Desirable!

Output resistance: $R_{OUT} = R_o(1+Af)$ Desirable!

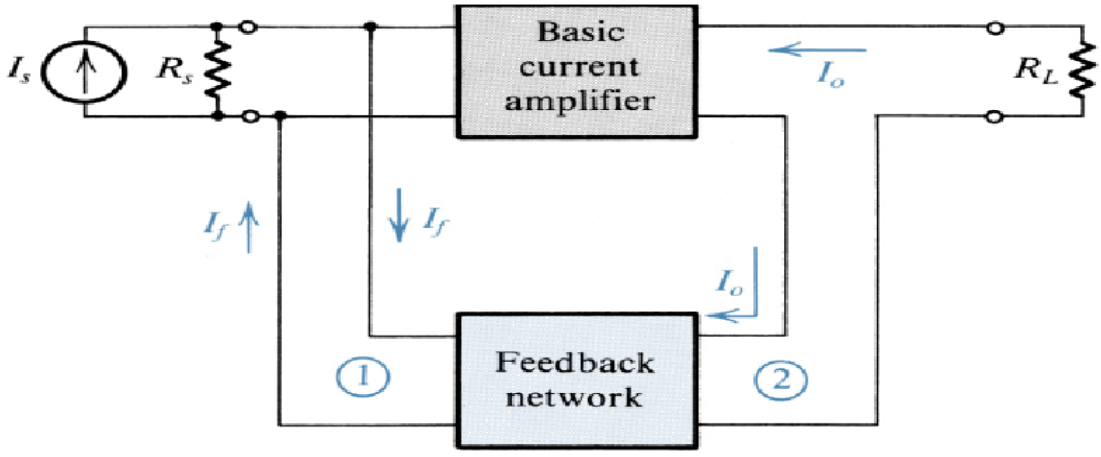
Series-series feedback (for transconductance amplifier)



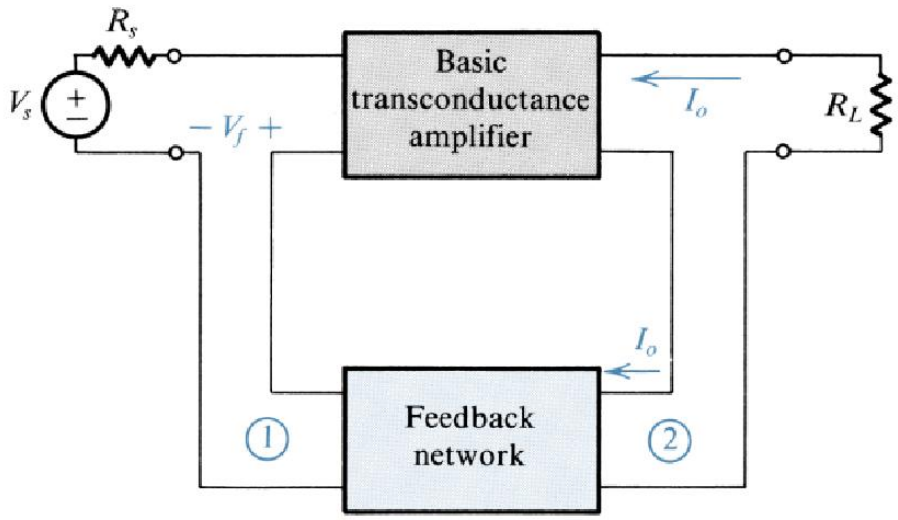
Overall gain (closed-loop gain) : $A_o = \dfrac{i_o}{v_i} = \dfrac{A}{1+Af}$

Input resistance: $R_{IN} = R_i(1+Af)$ Desirable!

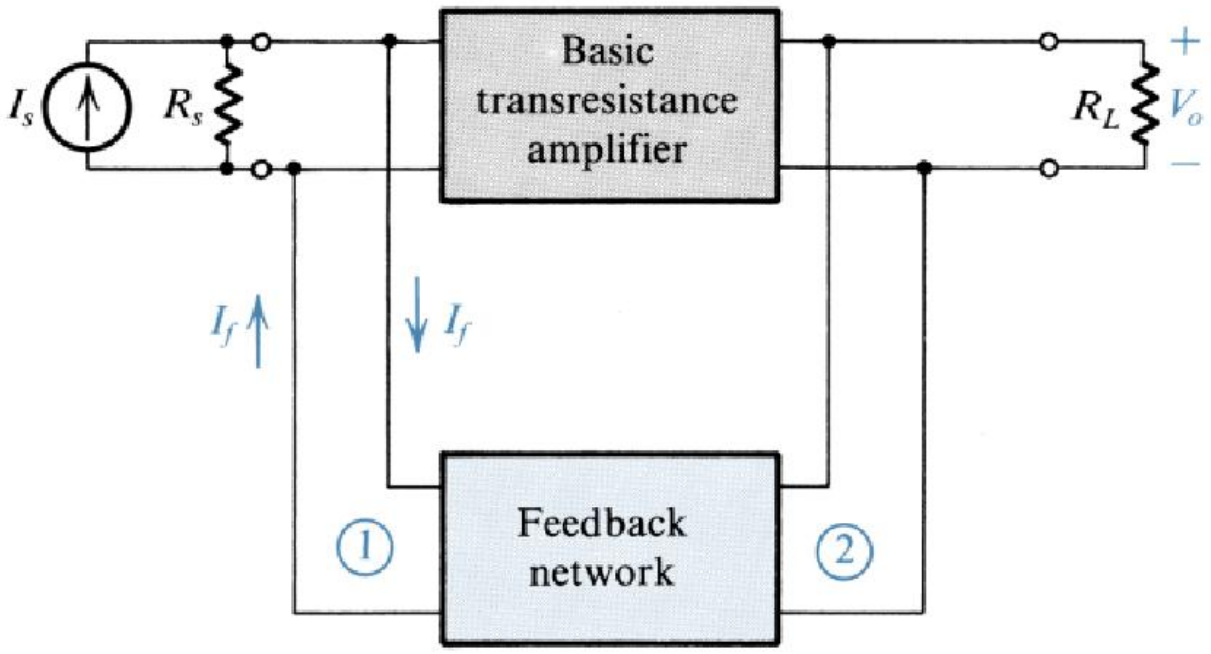
Output resistance: $R_{OUT} = R_o(1+Af)$ Desirable!



he four basic feedback topologies: current-sampling shunt-mixing (shunt-series) topology



The four basic feedback topologies: current-sampling series-mixing (series-series) topology



The four basic feedback topologies: voltage-sampling shunt-mixing (shunt-shunt) topology

Oscillator

- Introduction of Oscillator
- Linear Oscillator
  - Wien Bridge Oscillator
  - RC Phase-Shift Oscillator
  - LC Oscillator

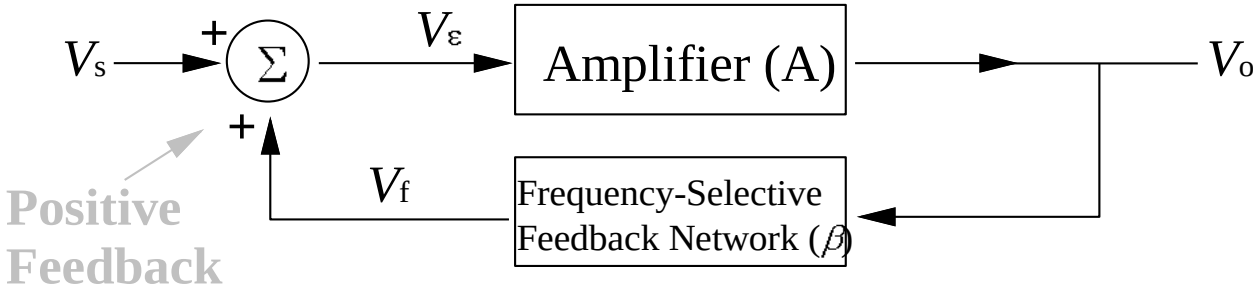
Stability Oscillation: an effect that repeatedly and regularly fluctuates about the mean value Oscillator: circuit that produces oscillation
Characteristics: wave-shape, frequency, amplitude, distortion, stability

- Oscillators are used to generate signals, e.g.
  - Used as a local oscillator to transform the RF signals to IF signals in a receiver;
  - Used to generate RF carrier in a transmitter
  - Used to generate clocks in digital systems;
  - Used as sweep circuits in TV sets and CRO.

Linear Oscillators

1. Wien Bridge Oscillators
2. RC Phase-Shift Oscillators
3. LC Oscillators
4. Stability

Integrant of Linear Oscillators



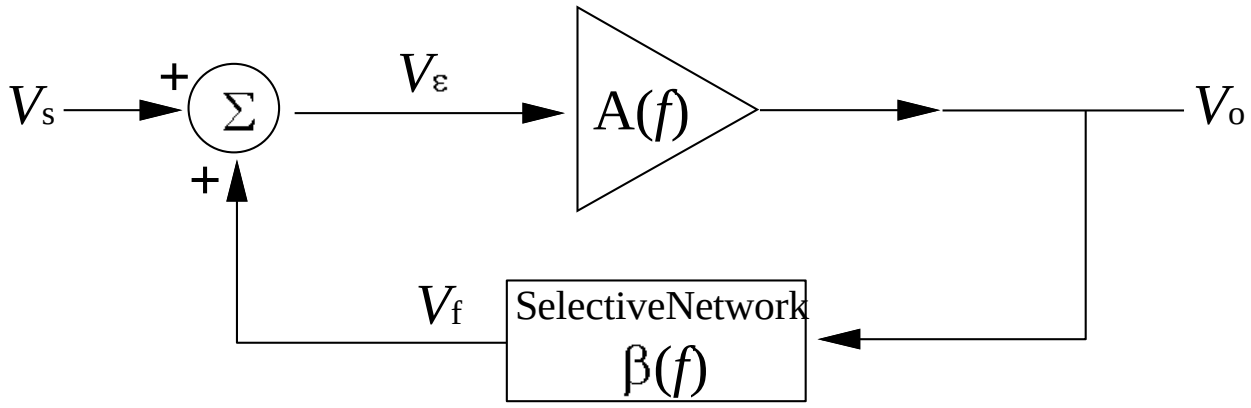
For sinusoidal input is connected

"Linear" because the output is approximately sinusoidal

A linear oscillator contains:

- a frequency selection feedback network
- an amplifier to maintain the loop gain at **unity**

Basic Linear Oscillator



$$V_o = AV_\varepsilon = A(V_s + V_f) \quad \text{and} \quad V_f = \beta V_o$$

$$\Rightarrow \frac{V_o}{V_s} = \frac{A}{1 - A\beta}$$

If $V_s = 0$, the only way that V_o can be nonzero is that loop gain $A\beta=1$ which implies that

$$|A\beta| = 1$$

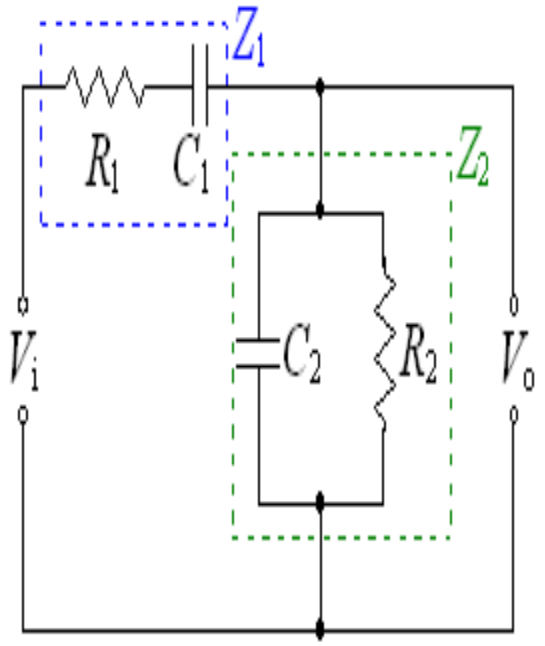
$$\angle A\beta = 0$$

(Barkhausen Criterion)

Wien Bridge Oscillator

Let $X_{C1} = \dfrac{1}{\omega C_1}$ and $X_{C2} = \dfrac{1}{\omega C_2}$

Frequency Selection Network



$$Z_1 = R_1 - jX_{C1}$$

$$Z_2 = \left[\frac{1}{R_2} + \frac{1}{-jX_{C2}}\right]^{-1} = \frac{-jR_2X_{C2}}{R_2 - jX_{C2}}$$

Therefore, the feedback factor,

$$\beta = \frac{V_o}{V_i} = \frac{Z_2}{Z_1 + Z_2} = \frac{(-jR_2X_{C2}/R_2 - jX_{C2})}{(R_1 - jX_{C1}) + (-jR_2X_{C2}/R_2 - jX_{C2})}$$

$$\beta = \frac{-jR_2X_{C2}}{(R_1 - jX_{C1})(R_2 - jX_{C2}) - jR_2X_{C2}}$$

β can be rewritten as:

$$\beta = \frac{R_2 X_{C2}}{R_1 X_{C2} + R_2 X_{C1} + R_2 X_{C2} + j(R_1 R_2 - X_{C1} X_{C2})}$$

For ***Barkhausen Criterion***, imaginary part = 0, i.e.,

$$R_1 R_2 - X_{C1} X_{C2} = 0$$

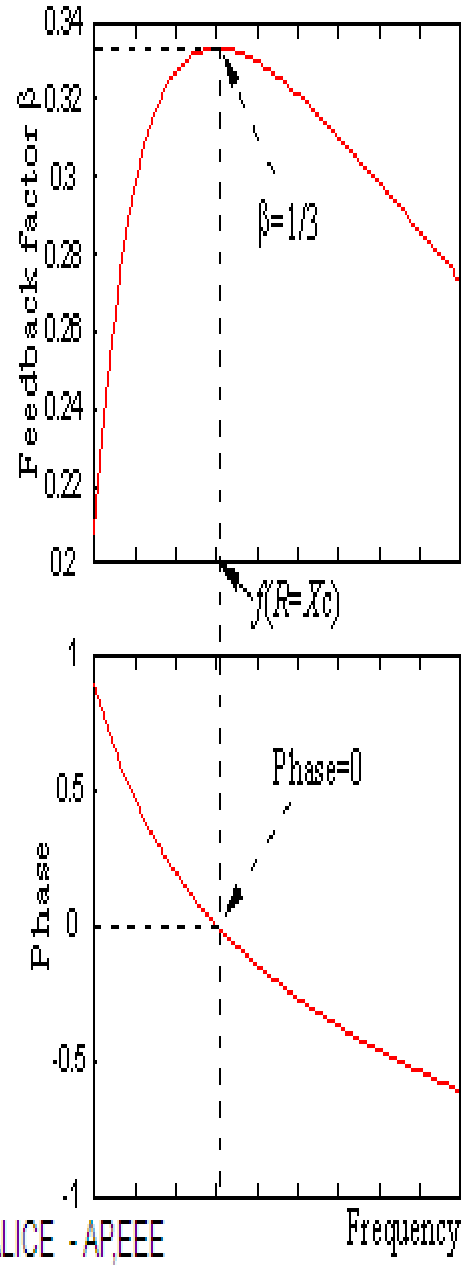
$$\text{or} \quad R_1 R_2 = \frac{1}{\omega C_1} \frac{1}{\omega C_2}$$

$$\Rightarrow \omega = 1/\sqrt{R_1 R_2 C_1 C_2}$$

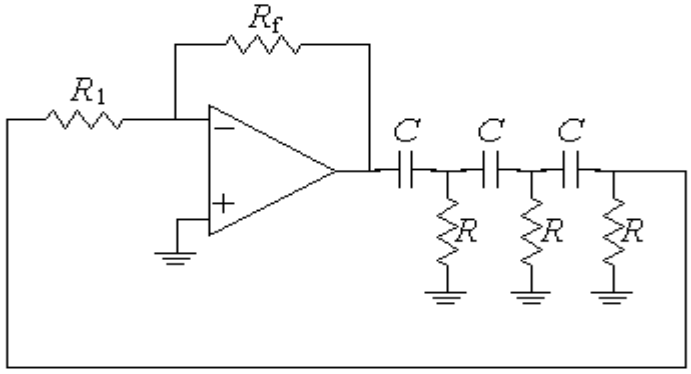
Supposing,

$R_1 = R_2 = R$ and $X_{C1} = X_{C2} = X_C$,

$$\beta = \frac{R X_C}{3R X_C + j(R^2 - X_C^2)}$$



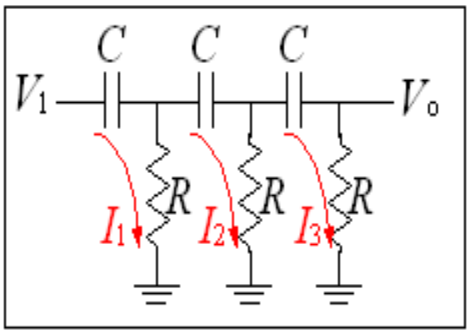
RC Phase-Shift Oscillator



- Using an inverting amplifier
- The additional 180° phase shift is provided by an RC phase-shift network

Applying KVL to the phase-shift network, we have

$$V_1 = I_1(R - jX_C) - I_2 R$$
$$0 = -I_1 R + I_2(2R - jX_C) - I_3 R$$
$$0 = -I_2 R + I_3(2R - jX_C)$$



Solve for I_3, we get

$$I_3 = \frac{\begin{vmatrix} R - jX_C & -R & V_1 \\ -R & 2R - jX_C & 0 \\ 0 & -R & 0 \end{vmatrix}}{\begin{vmatrix} R - jX_C & -R & 0 \\ -R & 2R - jX_C & -R \\ 0 & -R & 2R - jX_C \end{vmatrix}}$$

Or
$$I_3 = \frac{V_1 R^2}{(R - jX_C)[(2R - jX_C)^2 - R^2] - R^2(2R - jX_C)}$$

The output voltage,

$$V_o = I_3 R = \frac{V_1 R^3}{(R - jX_C)[(2R - jX_C)^2 - R^2] - R^2(2R - jX_C)}$$

Hence the transfer function of the phase-shift network is given by,

$$\beta = \frac{V_o}{V_1} = \frac{R^3}{(R^3 - 5RX_C^2) + j(X_C^3 - 6R^2X_C)}$$

For 180° phase shift, the imaginary part = 0, i.e.,

$$X_C^3 - 6R^2X_C = 0 \quad \text{or} \quad X_C = 0 \text{ (Rejected)}$$

$$\Rightarrow X_C^2 = 6R^2$$

$$\omega = \frac{1}{\sqrt{6}RC}$$

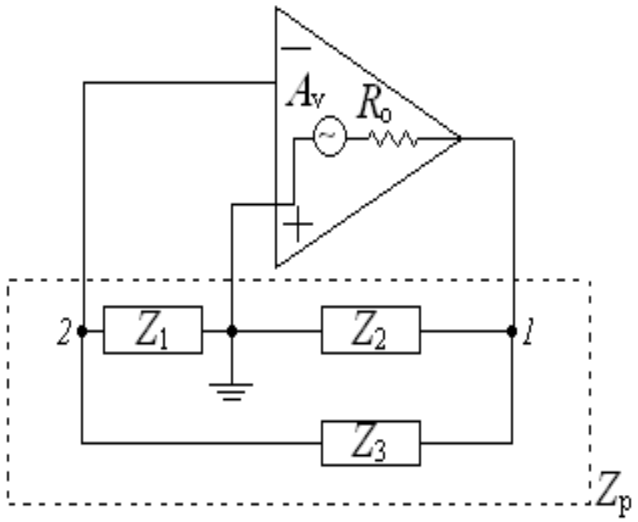
and,

$$\beta = -\frac{1}{29}$$

Note: The –ve sign mean the phase inversion from the voltage

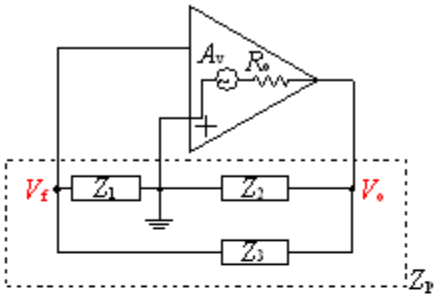
LC Oscillators

- The frequency selection network (Z_1, Z_2 and Z_3) provides a phase shift of 180°
- The amplifier provides an addition shift of 180°



Two well-known Oscillators:

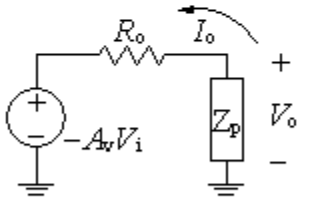
- Colpitts Oscillator
- Harley Oscillator



$$V_f = \beta V_o = \frac{Z_1}{Z_1 + Z_3} V_o$$

$$Z_p = Z_2 \,//\, (Z_1 + Z_3) = \frac{Z_2(Z_1 + Z_3)}{Z_1 + Z_2 + Z_3}$$

For the equivalent circuit from the output



$$\frac{-A_vV_i}{R_o + Z_p} = \frac{V_o}{Z_p} \quad \text{or} \quad \frac{V_o}{V_i} = \frac{-A_vZ_p}{R_o + Z_p}$$

Therefore, the amplifier gain is obtained,

$$A = \frac{V_o}{V_i} = \frac{-A_vZ_2(Z_1 + Z_3)}{R_o(Z_1 + Z_2 + Z_3) + Z_2(Z_1 + Z_3)}$$

The loop gain,

$$A\beta = \frac{-A_vZ_1Z_2}{R_o(Z_1 + Z_2 + Z_3) + Z_2(Z_1 + Z_3)}$$

If the impedance are all pure reactances, i.e.,

$Z_1 = jX_1$, $Z_2 = jX_2$ and $Z_3 = jX_3$

The loop gain becomes, $A\beta = \dfrac{A_vX_1X_2}{jR_o(X_1 + X_2 + X_3) - X_2(X_1 + X_3)}$

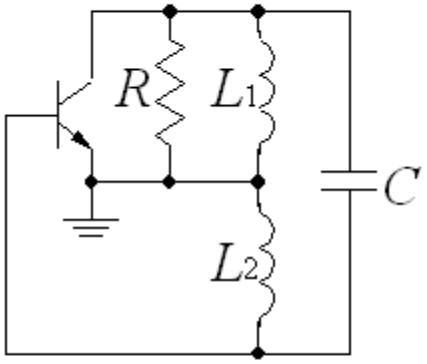
The imaginary part = 0 only when $X_1 + X_2 + X_3 = 0$

- It indicates that at least one reactance must be –ve (capacitor)
- X_1 and X_2 must be of same type and X_3 must be of opposite type

With imaginary part = 0, $A\beta = \dfrac{-A_vX_1}{X_1 + X_3} = \dfrac{A_vX_1}{X_2}$

For Unit Gain & 180° Phase-shift, $A\beta = 1 \Rightarrow A_v = \dfrac{X_2}{X_1}$

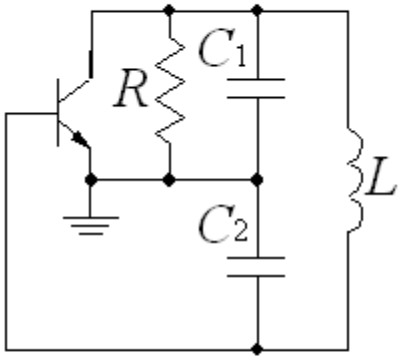
Hartley Oscillator



$$\omega_o = \frac{1}{\sqrt{(L_1 + L_2)C}}$$

$$g_m = \frac{L_1}{RL_2}$$

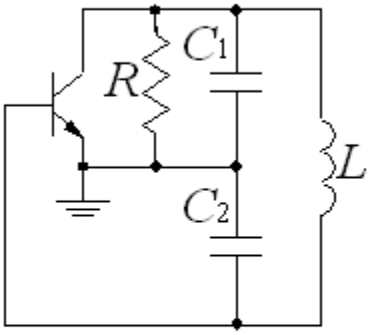
Colpitts Oscillator



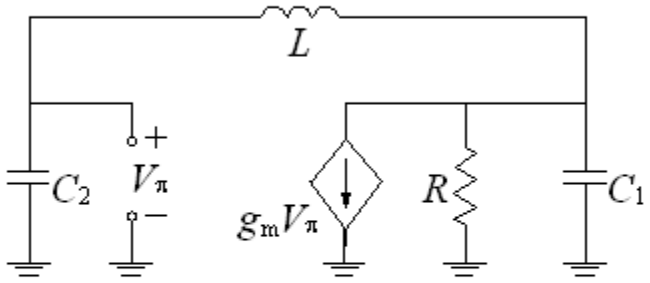
$$\omega_o = \frac{1}{\sqrt{LC_T}} \qquad C_T = \frac{C_1 C_2}{C_1 + C_2}$$

$$g_m = \frac{C_2}{RC_1}$$

Colpitts Oscillator



Equivalent circuit



In the equivalent circuit, it is assumed that:

- Linear small signal model of transistor is used
- The transistor capacitances are neglected
- Input resistance of the transistor is large enough

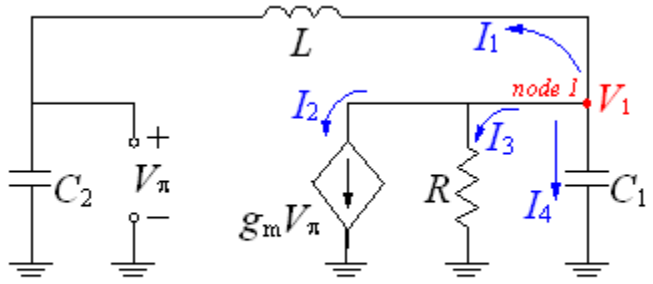
At node 1,

$$V_1 = V_\pi + i_1(j\omega L)$$

where,

$$i_1 = j\omega C_2 V_\pi$$

$$\Rightarrow \quad V_1 = V_\pi(1 - \omega^2 LC_2)$$



Apply KCL at node 1, we have

$$j\omega C_2 V_\pi + g_m V_\pi + \frac{V_1}{R} + j\omega C_1 V_1 = 0$$

$$j\omega C_2 V_\pi + g_m V_\pi + V_\pi(1 - \omega^2 LC_2)\left(\frac{1}{R} + j\omega C_1\right) = 0$$

For Oscillator V_π must not be zero, therefore it enforces,

$$\left(g_m + \frac{1}{R} - \frac{\omega^2 LC_2}{R}\right) + j\left[\omega(C_1 + C_2) - \omega^3 LC_1C_2\right] = 0$$

$$\left(g_m + \frac{1}{R} - \frac{\omega^2 LC_2}{R}\right) + j\left[\omega(C_1 + C_2) - \omega^3 LC_1C_2\right] = 0$$

Imaginary part = 0, we have

$$\omega_o = \frac{1}{\sqrt{LC_T}} \qquad C_T = \frac{C_1C_2}{C_1 + C_2}$$

Real part = 0, yields

$$g_m = \frac{C_2}{RC_1}$$

Question Bank

PART –A

1.List the two important types of RC oscillators.

2. What is the main advantage of crystal oscillator?

PART-B

- Explain the working of a transistorized Hartley oscillator. Derive expressions for frequency of oscillation and condition for sustained oscillation.

- Draw the equivalent circuit of current amplifier with current shunt feedback and derive expressions for (a) the output resistance with feedback Rof (b) output terminal resistance with feedback Roft.

UNIT V

PULSE CIRCUITS AND POWER SUPPLY

RC wave shaping circuits – Diode clampers and clippers – Multivibrators – Schmitt triggers – UJT saw-tooth oscillators – Single and poly-phase rectifiers and analysis of filter circuits – Design of zener and transistor series voltage regulators – Switched mode power supplies.

UNIT V

PULSE CIRCUITS AND POWER SUPPLY

Wave Shaping

The process whereby the form of a non sinusoidal signal is altered by transmission through a linear network is called “linear wave shaping”. An ideal low pass circuit is one that allows all the input frequencies below a frequency called cutoff frequency fc and attenuates all those above this frequency. For practical low pass circuit (Fig.1) cutoff is set to occur at a frequency where the gain of the circuit falls by 3 dB from its maximum at very high frequencies the capacitive reactance is very small, so the output is almost equal to the input and hence the gain is equal to 1. Since circuit attenuates low frequency signals and allows high frequency signals with little or no attenuation, it is called a high pass circuit.

Circuit Diagram:

Low Pass RC Circuit :

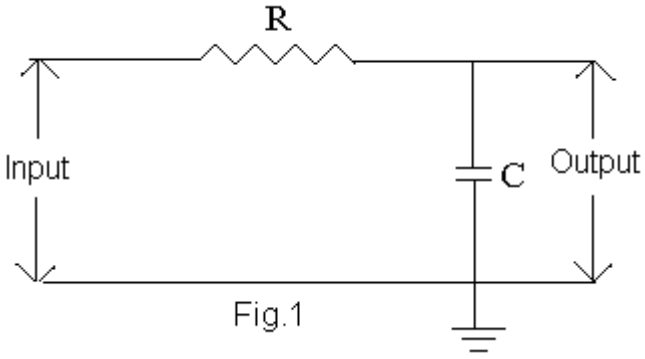


Fig.1

High Pass RC Circuit :

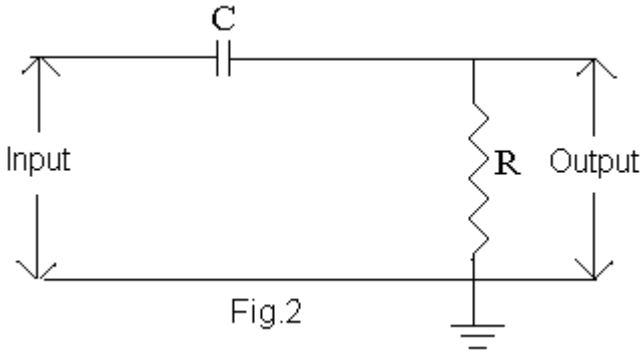
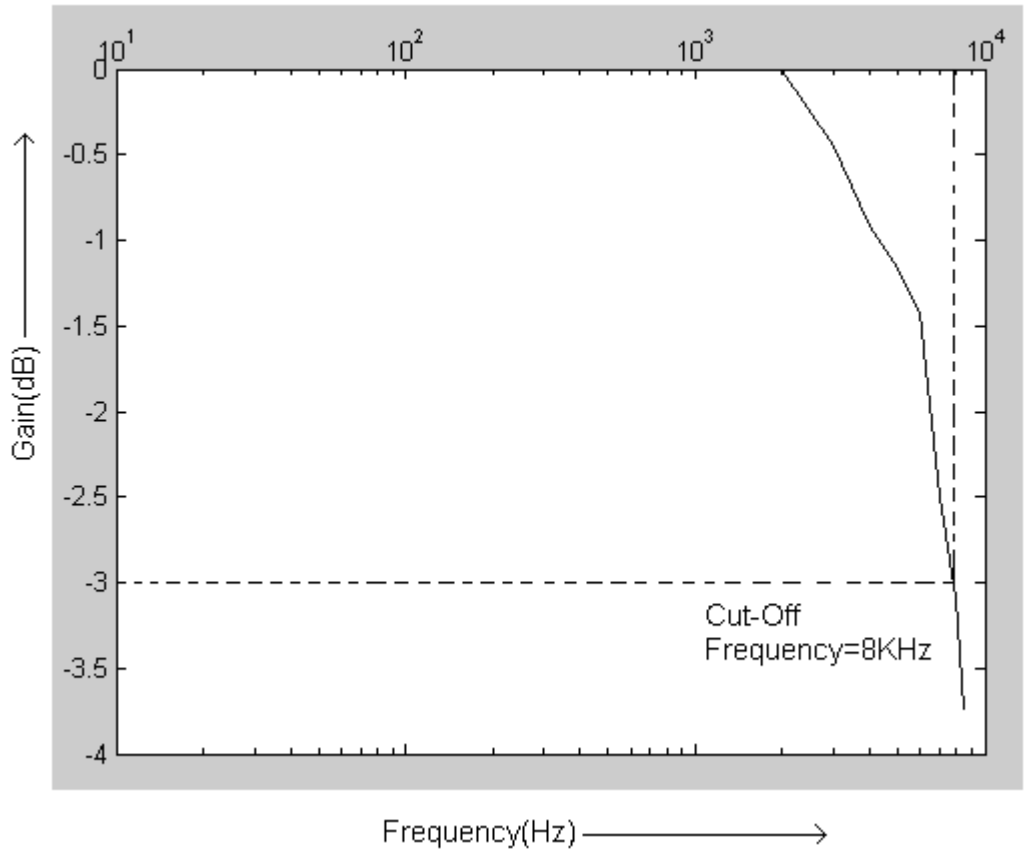
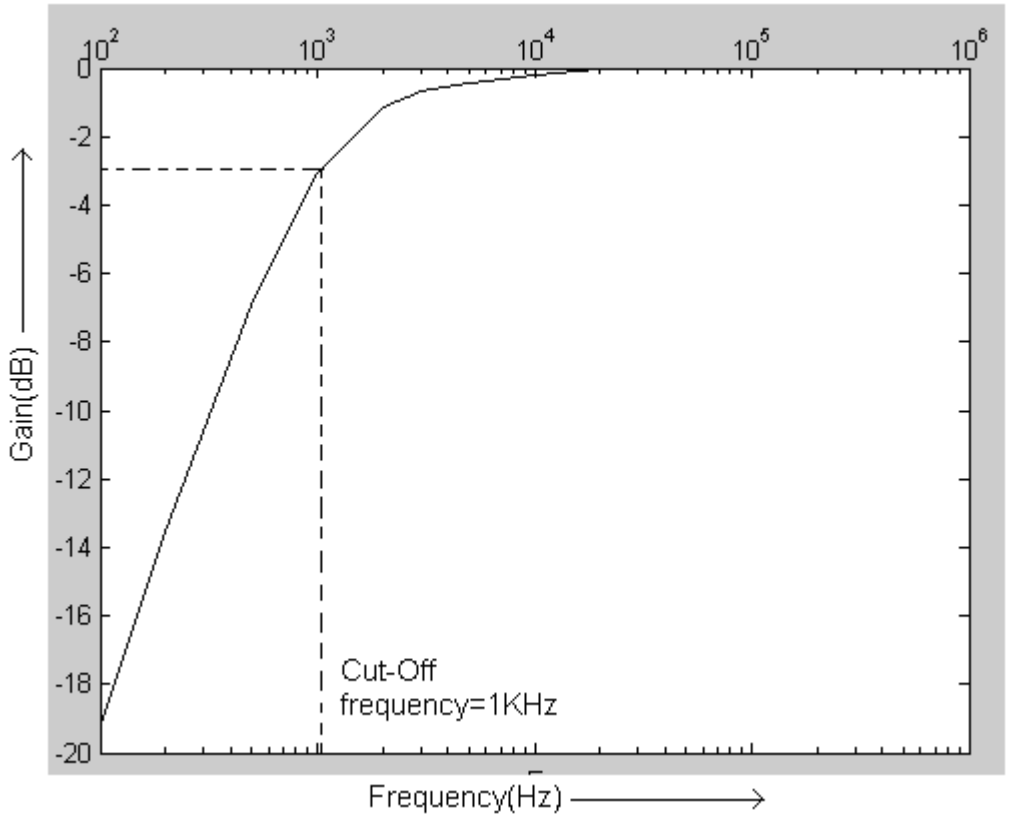


Fig.2

Low Pass RC circuit frequency response

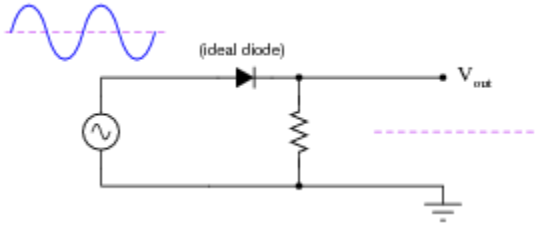


High Pass RC circuit frequency response:



CLIPPER CIRCUIT

an ideal diode with no forward voltage drop:



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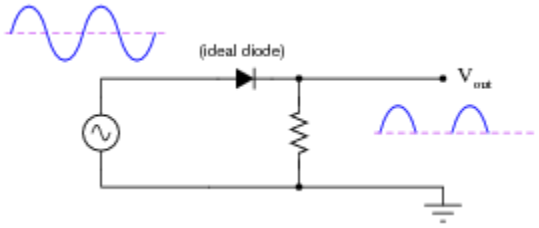

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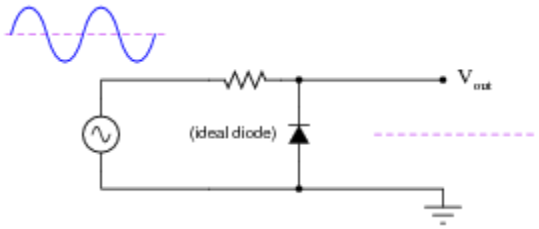

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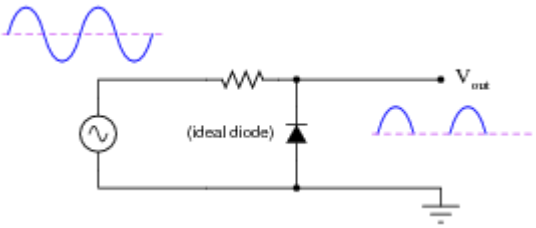

AC voltage source, one at a time.

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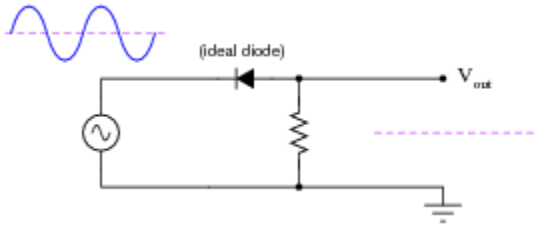

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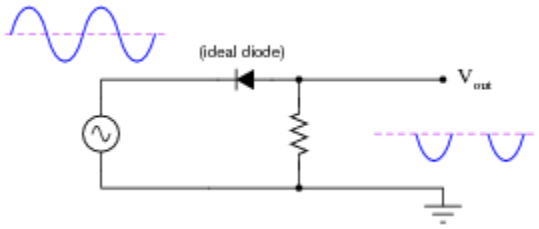

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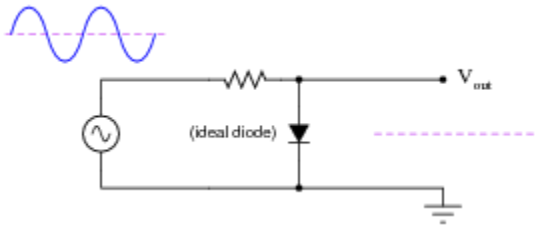

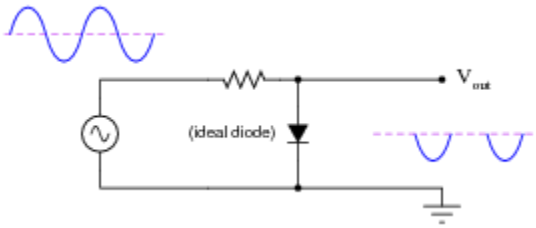

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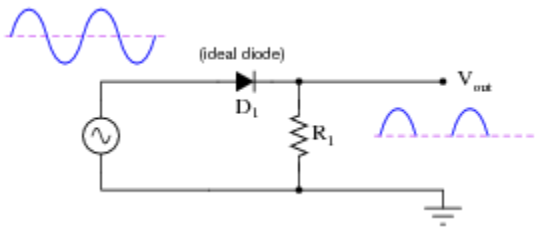

Diode D_1 fails open:

Diode D_1 fails shorted:

Resistor R_1 fails open:

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For each of these conditions, explain *why* the resulting effects will occur.

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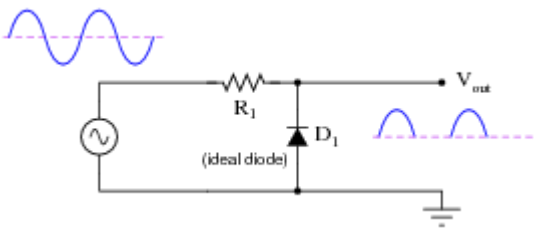

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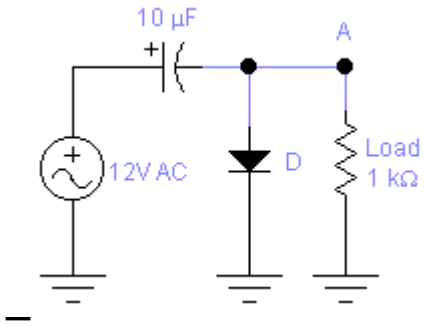
CLAMPPING CIRCUIT



Positive voltage clamping in point A

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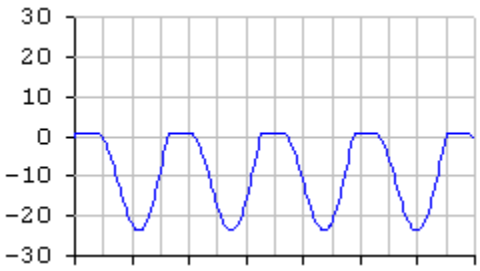

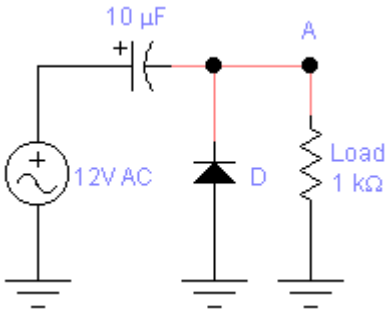

Negative voltage clamping in point A

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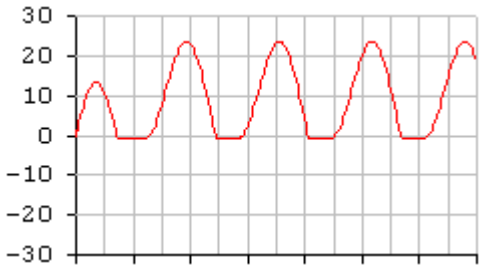
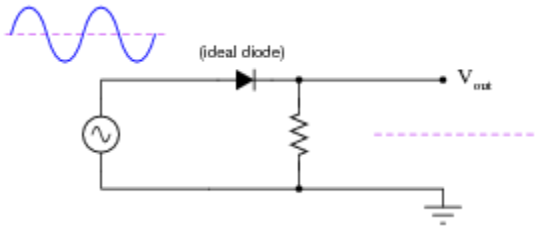


CLIPPER CIRCUIT

Sketch the shape of the output voltage waveform for this "clipper" circuit, assuming an ideal diode with no forward voltage drop:



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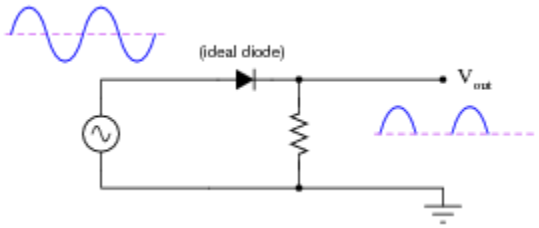

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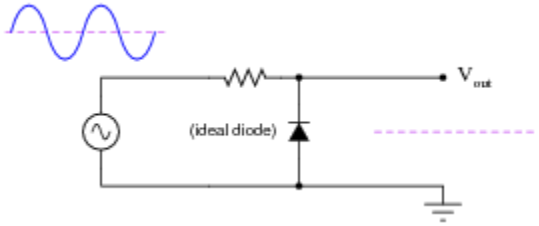

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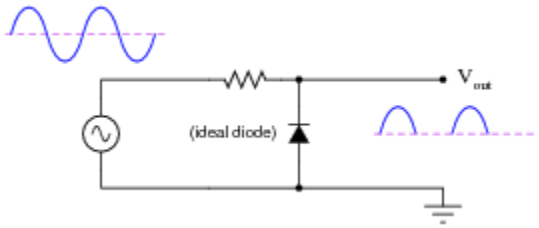

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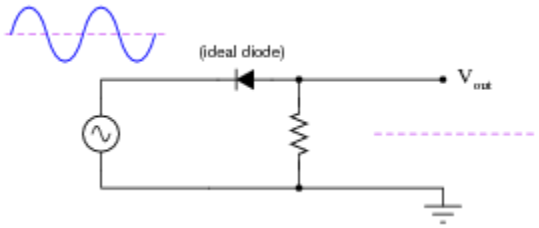

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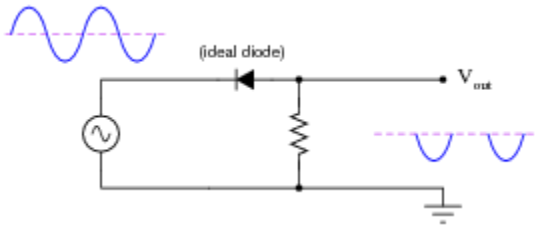

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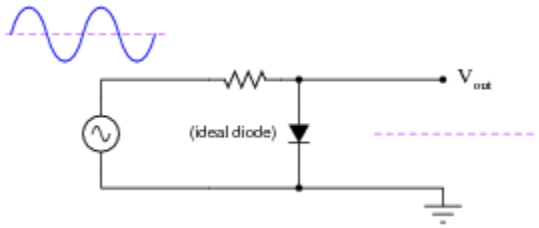

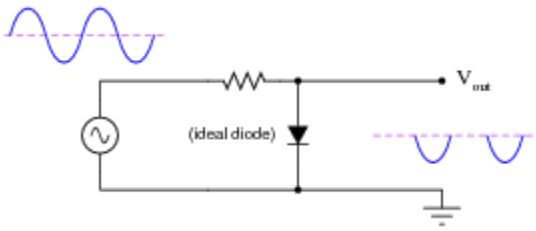

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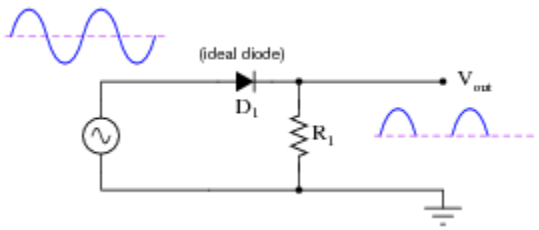

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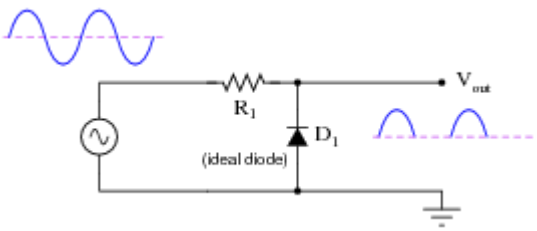

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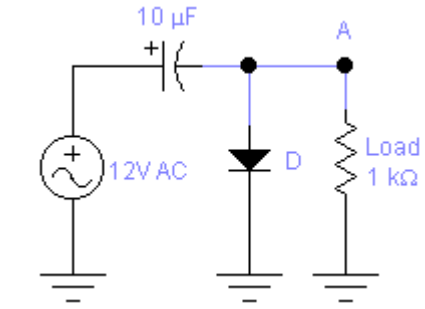

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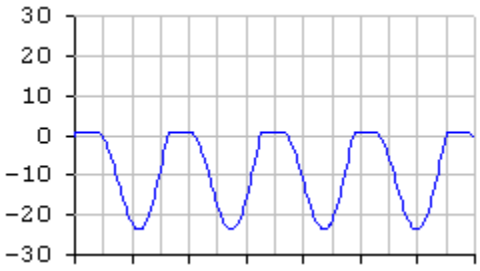

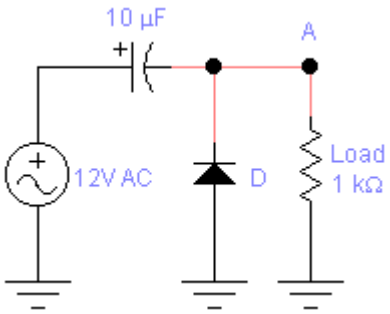

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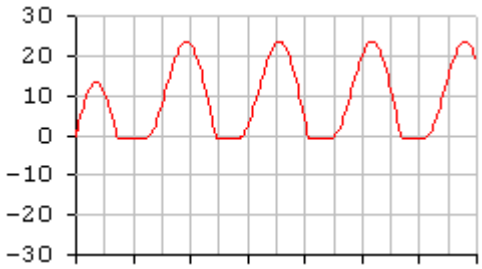


MULTIVIBRATOR

A **multivibrator** is an electronic circuit used to implement a variety of simple two-state systems such as oscillators, timers and flip-flops. It is characterized by two amplifying devices (transistors, electron tubes or other devices) cross-coupled by resistors and capacitors.

There are three types of multivibrator circuit:

- **Astable**, in which the circuit is not stable in either state—it continuously oscillates from one state to the other. Due to this, it does not require an input (Clock pulse or other).
- **Monostable**, in which one of the states is stable, but the other is not—the circuit will flip into the unstable state for a determined period, but will eventually return to the stable state. Such a circuit is useful for

creating a timing period of fixed duration in response to some external event. This circuit is also known as a **one shot**. A common application is in eliminating switch bounce.

- **Bistable**, in which the circuit will remain in either state indefinitely. The circuit can be flipped from one state to the other by an external event or trigger. Such a circuit is important as the fundamental building block of a register or memory device. This circuit is also known as a latch or a flip-flop.

In its simplest form the multivibrator circuit consists of two cross-coupled transistors. Using resistor-capacitor networks within the circuit to define the time periods of the unstable states, the various types may be implemented. Multivibrators find applications in a variety of systems where square waves or timed intervals are required. Simple circuits tend to be inaccurate since many factors affect their timing, so they are rarely used where very high precision is required.

Before the advent of low-cost integrated circuits, chains of multivibrators found use as frequency dividers. A free-running multivibrator with a frequency of one-half to one-tenth of the reference frequency would accurately lock to the reference frequency. This technique was used in early electronic organs, to keep notes of different octaves accurately in tune. Other applications included early television systems, where the various line and frame frequencies were kept synchronized by pulses included in the video signal. Astable multivibrator circuit

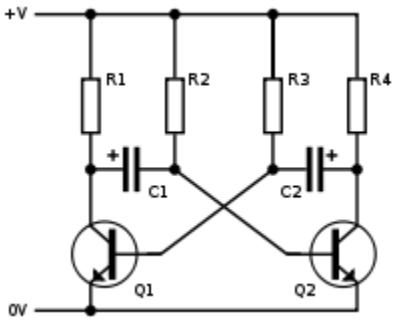


Figure 1: Basic BJT astable multivibrator

This circuit shows a typical simple astable circuit, with an output from the collector of Q1, and an inverted output from the collector of Q2.

Basic mode of operation

The circuit keeps one transistor switched on and the other switched off. Suppose that initially, Q1 is switched on and Q2 is switched off.

State 1:

- Q1 is on and holds the bottom of R1 near 0 V.
- The right side of C1 (and the base of Q2) is being charged by R2 from below ground to 0.6 V.
- R3 is pulling the base of Q1 up, but its base-emitter diode prevents the voltage from rising above 0.6 .
- R4 is charging the right side of C2 up to the power supply voltage (+*V*). Because R4 is less than R2, C2 charges faster than C1.

When the base of Q2 reaches 0.6 V, Q2 turns on, and the following positive feedback loop occurs:

- Q2 abruptly pulls the right side of C2 down to near 0 V.
- Because the voltage across a capacitor cannot suddenly change, this causes the left side of C2 to suddenly fall to almost $-V$, well below 0 V.
- Q1 switches off due to the sudden disappearance of its base voltage.
- R1 and R2 work to pull both ends of C1 toward $+V$, completing Q2's turn on. The process is stopped by the B-E diode of Q2, which will not let the right side of C1 rise very far.

This now takes us to **State 2**, the mirror image of the initial state, where Q1 is switched off and Q2 is switched on. Then R1 rapidly pulls C1's left side toward $+V$, while R3 more slowly pulls C2's left side toward +0.6 V. When C2's left side reaches 0.6 V, the cycle repeats.

Multivibrator frequency

The period of each *half* of the multivibrator is given by $t = \ln(2)RC$. The total period of oscillation is given by:

$T = t_1 + t_2 = \ln(2)R_2\,C_1 + \ln(2)R_3\,C_2$

$$f = \frac{1}{T} = \frac{1}{\ln(2)\cdot(R_2C_1 + R_3C_2)} \approx \frac{1}{0.693\cdot(R_2C_1 + R_3C_2)}$$

Where...

- f is frequency in hertz.
- R_2 and R_3 are resistor values in ohms.
- C_1 and C_2 are capacitor values in farads.
- T is period time (In this case, the sum of two period durations).

For the special case where

- $t_1 = t_2$ (50% duty cycle)
- $R_2 = R_3$
- $C_1 = C_2$

$$f = \frac{1}{T} = \frac{1}{\ln(2) \cdot 2RC} \approx \frac{0.721}{RC}$$

Initial power-up

When the circuit is first powered up, neither transistor will be switched on. However, this means that at this stage they will both have high base voltages and therefore a tendency to switch on, and inevitable slight asymmetries will mean that one of the transistors is first to switch on. This will quickly put the circuit into one of the above states, and oscillation will ensue. In practice, oscillation always occurs for practical values of *R* and *C*.

However, if the circuit is temporarily held with both bases high, for longer than it takes for both capacitors to charge fully, then the circuit will remain in this stable state, with both bases at 0.6 V, both collectors at 0 V, and both capacitors charged backwards to −0.6 V. This can occur at startup without external intervention, if *R* and *C* are both very small. For example, a 10 MHz oscillator of this type will often be unreliable. (Different oscillator designs, such as relaxation oscillators, are required at high frequencies.)

Period of oscillation

Very roughly, the duration of state 1 (low output) will be related to the time constant R_2C_1 as it depends on the charging of C1, and the duration of state 2

(high output) will be related to the time constant R_3C_2 as it depends on the charging of C2. Because they do not need to be the same, an asymmetric duty cycle is easily achieved.

However, the duration of each state also depends on the initial state of charge of the capacitor in question, and this in turn will depend on the amount of discharge during the previous state, which will also depend on the resistors used during discharge (R1 and R4) and also on the duration of the previous state, *etc*. The result is that when first powered up, the period will be quite long as the capacitors are initially fully discharged, but the period will quickly shorten and stabilize.

The period will also depend on any current drawn from the output and on the supply voltage.

Protective components

While not fundamental to circuit operation, diodes connected in series with the base or emitter of the transistors are required to prevent the base-emitter junction being driven into reverse breakdown when the supply voltage is in excess of the V_{eb} breakdown voltage, typically around 5-10 volts for general purpose silicon transistors. In the monostable configuration, only one of the transistors requires protection.

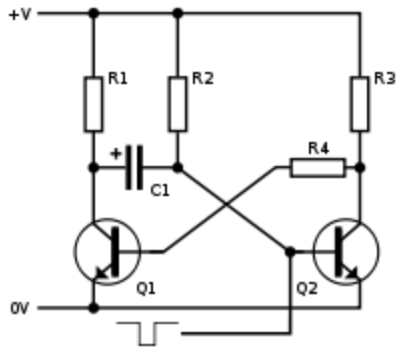


Figure 2: Basic BJT monostable multivibrator.

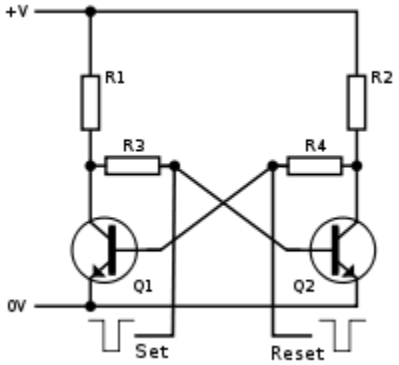


Figure 3: Basic BJT bistable multivibrator.

Non-electronic astables

Astable oscillators are usually thought of as electronic circuits, but need not be. Bimetallic strips are used which switch an electric current on as they cool and off as they heat—flashing Christmas and car indicator lights may use this mechanism.

Monostable multivibrator circuit

When triggered by an input pulse, a monostable multivibrator will switch to its unstable position for a period of time, and then return to its stable state. The time period monostable multivibrator remains in unstable state is given by $t = \ln(2)R_2C_1$. If repeated application of the input pulse maintains the circuit in the unstable state, it is called a *retriggerable* monostable. If further trigger pulses do not affect the period, the circuit is a *non-retriggerable* multivibrator.

Bistable multivibrator circuit

Suggested values:

- **R1, R2** = 10 kΩ
- **R3, R4** = 10 kΩ

This latch circuit is similar to an astable multivibrator, except that there is no charge or discharge time, due to the absence of capacitors. Hence, when the circuit is switched on, if Q1 is on, its collector is at 0 V. As a result, Q2 gets switched off. This results in more than half $+V$ volts being applied to R4 causing current into the base of Q1, thus keeping it on. Thus, the circuit remains stable in a single state continuously. Similarly, Q2 remains on continuously, if it happens to get switched on first.

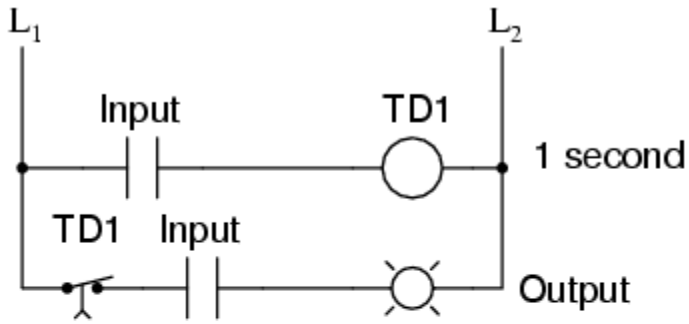
Switching of state can be done via Set and Reset terminals connected to the bases. For example, if Q2 is on and Set is grounded momentarily, this switches Q2 off, and makes Q1 on. Thus, Set is used to "set" Q1 on, and Reset is used to "reset" it to off state.

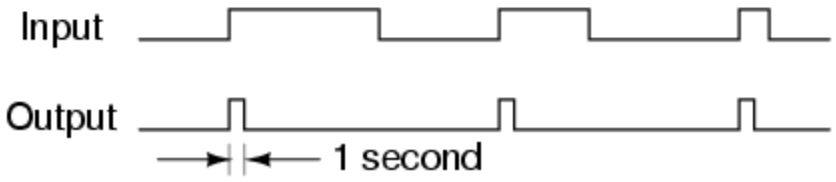
We've already seen one example of a monostable multivibrator in use: the pulse detector used within the circuitry of flip-flops, to enable the latch portion for a brief time when the clock input signal transitions from either low to high or high to low. The pulse detector is classified as a monostable multivibrator because it has only *one* stable state. By *stable,* I mean a state of output where the device is able to latch or hold to forever, without external prodding. A latch or flip-flop, being a bistable device, can hold in either the "set" or "reset" state for an indefinite period of time. Once its set or reset, it will continue to latch in that state unless prompted to change by an external input. A monostable device, on the other hand, is only able to hold in one particular state indefinitely. Its other state can only be held momentarily when triggered by an external input.

A mechanical analogy of a monostable device would be a momentary contact pushbutton switch, which spring-returns to its normal (stable) position when pressure is removed from its button actuator. Likewise, a standard wall (toggle) switch, such as the type used to turn lights on and off in a house, is a bistable device. It can latch in one of two modes: on or off.

All monostable multivibrators are *timed* devices. That is, their unstable output state will hold only for a certain minimum amount of time before returning to its stable state. With semiconductor monostable circuits, this timing function is typically accomplished through the use of resistors and capacitors, making use of the exponential charging rates of RC circuits. A comparator is often used to compare the voltage across the charging (or discharging) capacitor with a steady reference voltage, and the on/off output of the comparator used for a logic signal. With ladder logic, time delays are accomplished with time-delay relays, which can be constructed with

semiconductor/RC circuits like that just mentioned, or mechanical delay devices which impede the immediate motion of the relay's armature. Note the design and operation of the pulse detector circuit in ladder logic:

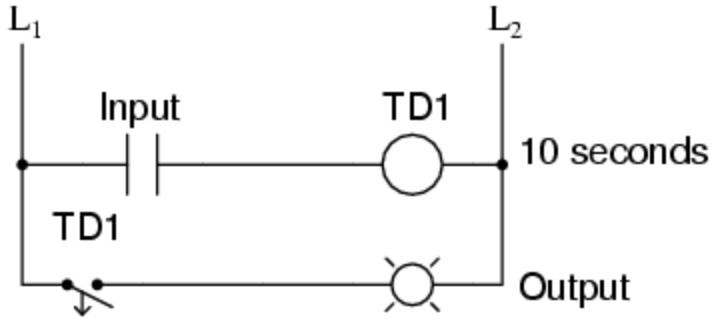


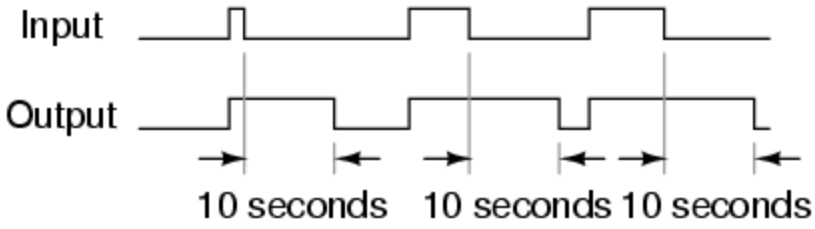


No matter how long the input signal stays high (1), the output remains high for just 1 second of time, then returns to its normal (stable) low state.

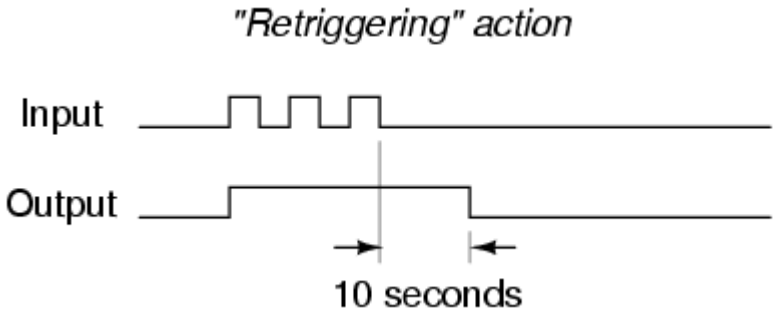
For some applications, it is necessary to have a monostable device that outputs a longer pulse than the input pulse which triggers it. Consider the

following ladder logic circuit:

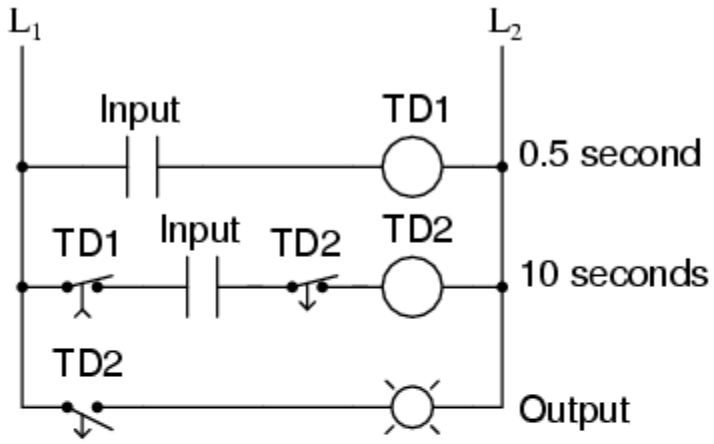


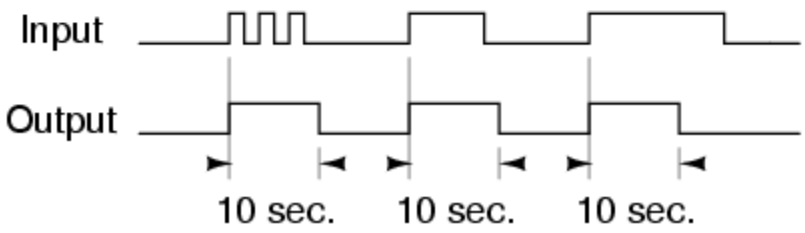


When the input contact closes, TD_1 contact immediately closes, and stays closed for 10 seconds after the input contact opens. No matter how short the input pulse is, the output stays high (1) for exactly 10 seconds after the input drops low again. This kind of monostable multivibrator is called a *one-shot*. More specifically, it is a *retriggerable* one-shot, because the timing begins after the input drops to a low state, meaning that multiple input pulses within 10 seconds of each other will maintain a continuous high output:



What if we only wanted a 10 second timed pulse output from a relay logic circuit, *regardless* of how many input pulses we received or how long-lived they may be? In that case, we'd have to couple a pulse-detector circuit to the retriggerable one-shot time delay circuit, like this:





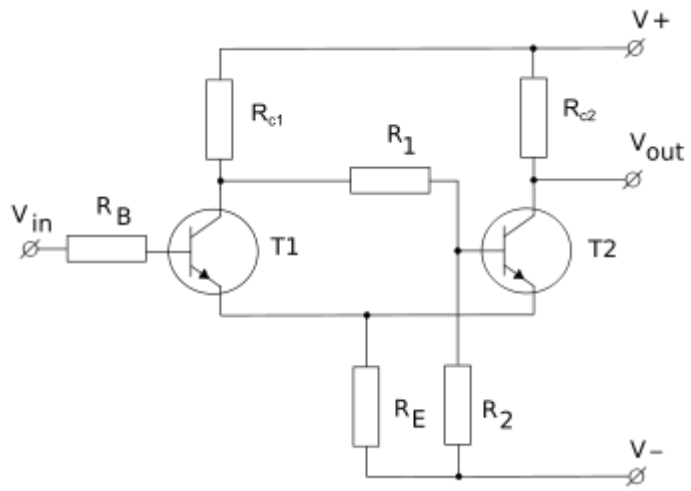
Time delay relay TD_1 provides an "on" pulse to time delay relay coil TD_2 for an arbitrarily short moment (in this circuit, for at least 0.5 second each time the input contact is actuated). As soon as TD_2 is energized, the normally-closed, timed-closed TD_2 contact in series with it prevents coil TD_2 from being re-energized as long as its timing out (10 seconds). This effectively makes it unresponsive to any more actuations of the input switch during that 10 second period.

Only after TD_2 times out does the normally-closed, timed-closed TD_2 contact in series with it allow coil TD_2 to be energized again. This type of one-shot is called a *nonretriggerable* one-shot.

One-shot multivibrators of both the retriggerable and nonretriggerable variety find wide application in industry for siren actuation and machine sequencing, where an intermittent input signal produces an output signal of a set time.

Schmitt trigger with emitter-coupled transistors

Structure



Schmitt trigger implemented by two emitter-coupled transistor stages

Ii is implemented as a transistor non-inverting amplifier with positive feedback consisting of two emitter-coupled (cascaded) transistor stages (common-collector T2 and common-base T1) connected in a loop. There is also a smaller negative feedback introduced by the emitter resistor R_E. This

configuration can be also considered as a differential amplifier with positive feedback between its non-inverting input (T2 base) and output (T1 collector). The chain R_{c1}-R_1-R_2 sets the base voltage for transistor T2. To make the positive feedback dominate over the negative one and to obtain a hysteresis, the proportion between the two collector resistors is chosen $R_{c1} > R_{c2}$. Thus less current flows through and less voltage drop is across R_E when T1 is switched than in the case when T2 is switched on. As a result, the circuit has two different thresholds in regard to ground (V_- in the picture).

Operation

Initial state. For NPN transistors as shown, imagine the input voltage is below the shared emitter voltage (high threshold for concreteness) so that T1 base-emitter junction is backward-biased and T1 does not conduct. T2 base voltage is determined by the mentioned divider so that T2 is conducting and the trigger output is in the low state. The two resistors R_{c2} and R_E form another voltage divider that determines the high threshold. Neglecting V_{BE}, the high threshold value is approximately

$$V_{HT} = \frac{R_E}{R_E + R_{c2}} V_+.$$

The output voltage is low but well above the ground. It is approximately equal to the high threshold and may not be low enough to be a logical zero for next digital circuits. This may require additional forming circuit (an amplifier) following the trigger circuit.

Crossing up the high threshold. When the input voltage (T1 base voltage) rises slightly above the voltage across the emitter resistor R_E (the high

threshold), T1 begins conducting. Its collector voltage goes down and T2 begins going cut-off, because the voltage divider now provides lower T2 base voltage. The common emitter voltage follows this change and goes down thus making T1 conduct more. The current begins steering from the right leg of the circuit to the left one. Although T1 is more conducting, it passes less current through R_E (since $R_{c1} > R_{c2}$); the emitter voltage continues dropping and the effective T1 base-emitter voltage continuously increases. This avalanche-like process continues until T1 becomes completely turned on (saturated) and T2 turned off. The trigger is transitioned to the high state and the output (T2 collector) voltage is close to V+. Now, the two resistors R_{c1} and R_E form a voltage divider that determines the low threshold. Its value is approximately

$$V_{LT} = \frac{R_E}{R_E + R_{c1}} V_+ .$$

Crossing down the low threshold. With the trigger now in the high state, if the input voltage lowers enough (below the low threshold), T1 begins to cut-off. Its collector current reduces; as a result, the shared emitter voltage lowers slightly and T1 collector voltage rises significantly. R_1-R_2 voltage divider conveys this change to T2 base voltage and it begins conducting. The voltage across R_E rises, further reducing the T1 base-emitter potential in the same avalanche-like manner, and T1 ceases to conduct. T2 becomes completely turned-on (saturated) and the output voltage is low again.

Generalization

The circuit can be generalized to see the basic dynamic threshold idea behind it. The two collector resistors R_{c1} and R_{c2} represent the upper legs of a switchable voltage divider with a common lower leg (R_E). T1 acts as a comparator with a differential input (T1 base-emitter junction) consisting of an inverting (T1 base) and a non-inverting (T1 emitter) inputs. The input voltage is applied to the inverting input; the output voltage of the voltage divider is applied to the non-inverting input thus determining its threshold. The comparator output drives an electronic double throw switch (the transistors T1 and T2) that switches over the upper legs of the voltage divider and changes the threshold in different to the input voltage direction. The series positive feedback forces and finishes this process.

UJT

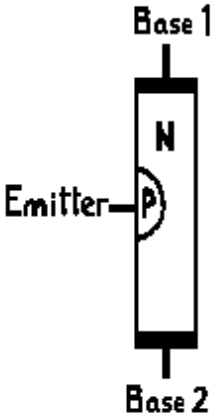
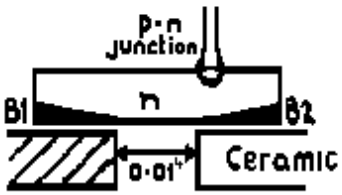
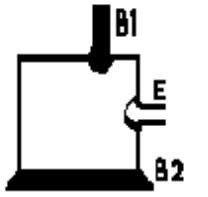


Fig.1

Fig.2



Bar structure



Cube structure

Fig.3

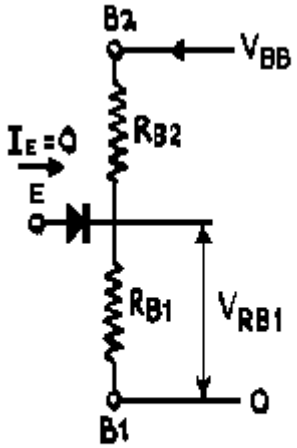


Fig.4.

The equivalent circuit shown in Fig.4 has been developed to explain how the device works, and it is necessary to define the terms used in this explanation.

R_{BB} is known as the interbase resistance, and is the sum of R_{B1} and R_{B2}:

$$R_{BB} = R_{B1} + R_{B2} \quad (1)$$

N.B. This is only true when the emitter is open circuit.

V_{RB1} is the voltage developed across R_{B1}; this is given by the voltage divider rule:

$$V_{RB1} = \frac{R_{B1}}{R_{B1} + R_{B2}} \quad (2)$$

Since the denominator of equation 2 is equal to equation 1, the former can be rewritten as:

$$\frac{R_{B1}}{}$$

$VRB1 = \_ \times V_{BB}$ (3)

R_{BB}

The ratio R_{B1} / R_{BB} is referred to as the intrinsic standoff ratio and is denoted by η(the Greek letter eta).

If an external voltage V_e is connected to the emitter, the equivalent circuit can be redrawn as shown in Fig.5.

If Ve is less than V_{RB1}, the diode is reverse biased and the circuit behaves as though the emitter was open circuit. If however V_e is increased so that it exceeds V_{RB1} by at least 0.7V, the diode becomes forward biased and emitter current Ie flows into the base 1 region. Because of this, the value of R_{B1} decreases. It has been suggested that this is due to the presence of additional charge carriers (holes) in the bar. Further increase in V_e causes the emitter current to increase which in turn reduces R_{B1} and this causes a further increase in current. This runaway effect is termed regeneration. The value of emitter voltage at which this occurs is known as the peak voltage V_P and is given by: $V_P = \eta_{AV} V_{BB} + V_D$ (4)

The characteristics of the UJT are illustrated by the graph of emitter voltage

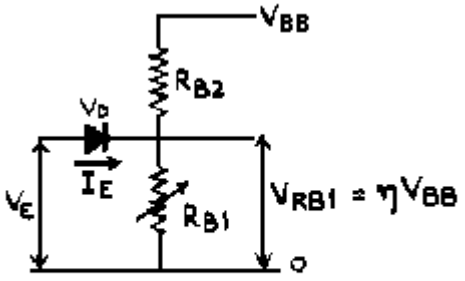


Fig.5

against

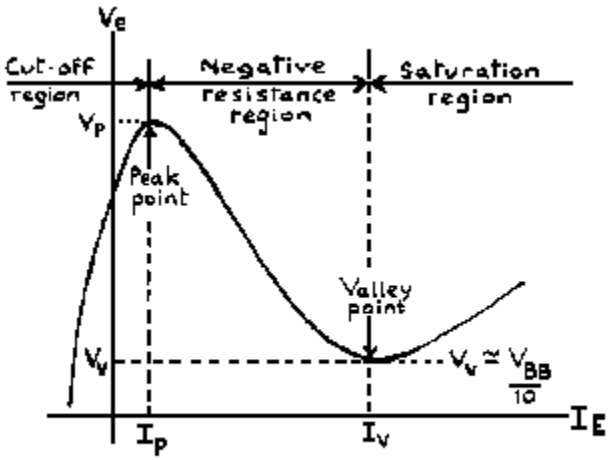


Fig.6

As the emitter voltage is increased, the current is very small - just a few microamps. When the peak point is reached, the current rises rapidly, until at the valley point the device runs into saturation. At this point R_{B1} is at its lowest value, which is known as the saturation resistance.

The simplest application of a UJT is as a relaxation oscillator, which is defined as one in which a capacitor is charged gradually and then discharged rapidly. The basic circuit is shown in Fig.7; in the practical circuit of Fig.8 R3 limits the emitter current and provides a voltage pulse, while R2 provides a measure of temperature compensation. Fig. 9 shows the waveforms occurring at the emitter and base 1; the first is an approximation to a sawtooth and the second is a pulse of short duration.

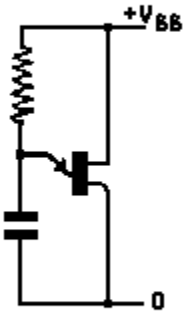


Fig.7

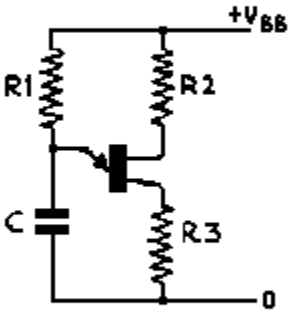


Fig.8

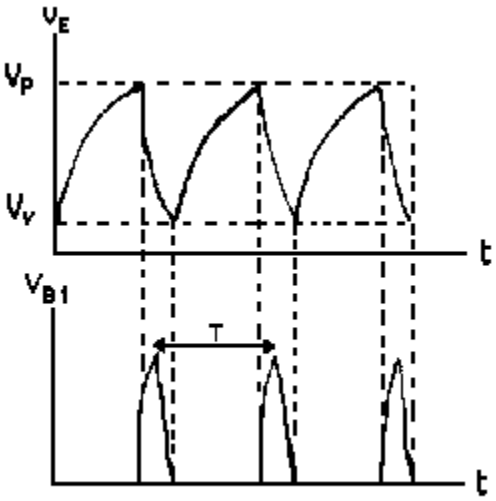


Fig.9

The operation of the circuit is as follows: C1 charges through R1 until the voltage across it reaches the peak point. The emitter current then rises rapidly, discharging C1 through the base 1 region and R3. The sudden rise of current through R3 produces the voltage pulse. When the current falls to I_V the UJT switches off and the cycle is repeated.

It can be shown that the time t between successive pulses is given by:

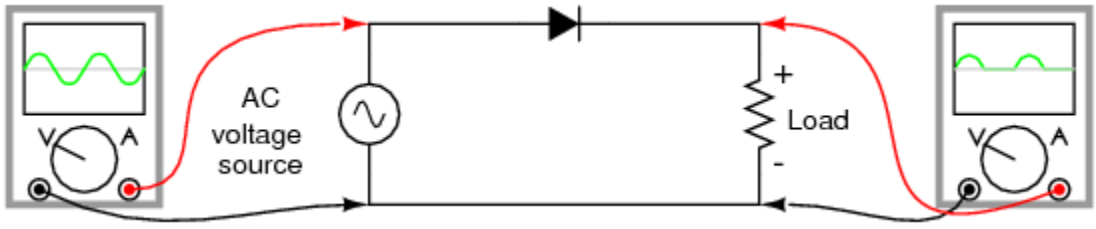
$$t + R1C \ln \frac{V_{BB} - V_V}{V_{BB} - V_P} \text{ secs (5)}$$

N.B. R measured in Megaohms. C in µF.

Single and poly-phase rectifiers and analysis of filter circuits

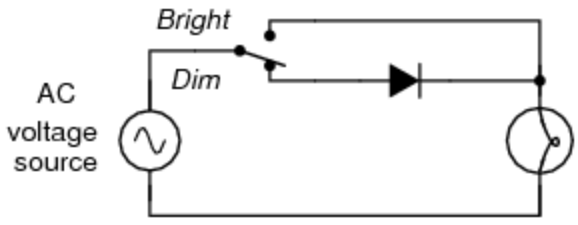
Rectification is the conversion of alternating current (AC) to direct current (DC). This involves a device that only allows one-way flow of electrons. As we have seen, this is exactly what a semiconductor diode does. The simplest

kind of rectifier circuit is the *half-wave* rectifier. It only allows one half of an AC waveform to pass through to the load. (Figure below)



Half-wave rectifier circuit.

For most power applications, half-wave rectification is insufficient for the task. The harmonic content of the rectifier's output waveform is very large and consequently difficult to filter. Furthermore, the AC power source only supplies power to the load one half every full cycle, meaning that half of its capacity is unused. Half-wave rectification is, however, a very simple way to reduce power to a resistive load. Some two-position lamp dimmer switches apply full AC power to the lamp filament for "full" brightness and then half-wave rectify it for a lesser light output. (Figure below)

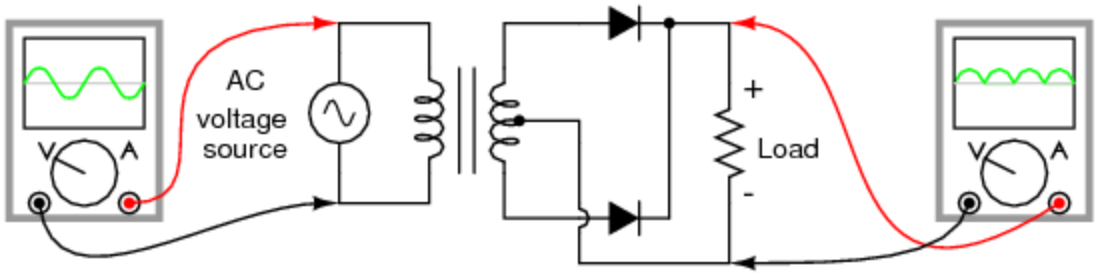


Half-wave rectifier application: Two level lamp dimmer.

In the "Dim" switch position, the incandescent lamp receives approximately one-half the power it would normally receive operating on full-wave AC.

Because the half-wave rectified power pulses far more rapidly than the filament has time to heat up and cool down, the lamp does not blink. Instead, its filament merely operates at a lesser temperature than normal, providing less light output. This principle of "pulsing" power rapidly to a slow-responding load device to control the electrical power sent to it is common in the world of industrial electronics. Since the controlling device (the diode, in this case) is either fully conducting or fully nonconducting at any given time, it dissipates little heat energy while controlling load power, making this method of power control very energy-efficient. This circuit is perhaps the crudest possible method of pulsing power to a load, but it suffices as a proof-of-concept application.

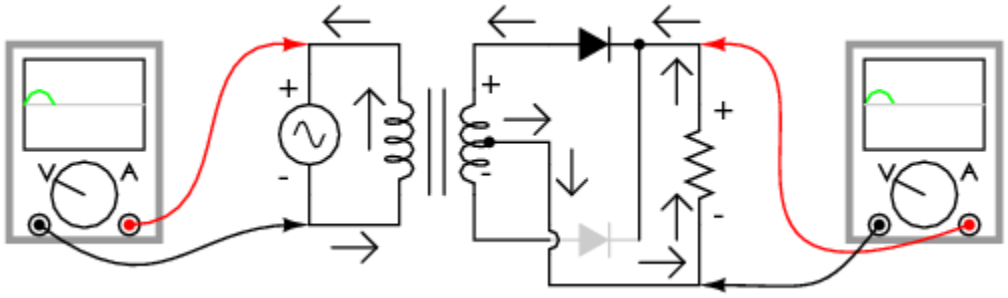
If we need to rectify AC power to obtain the full use of *both* half-cycles of the sine wave, a different rectifier circuit configuration must be used. Such a circuit is called a *full-wave* rectifier. One kind of full-wave rectifier, called the *center-tap* design, uses a transformer with a center-tapped secondary winding and two diodes, as in Figure below.



Full-wave rectifier, center-tapped design.

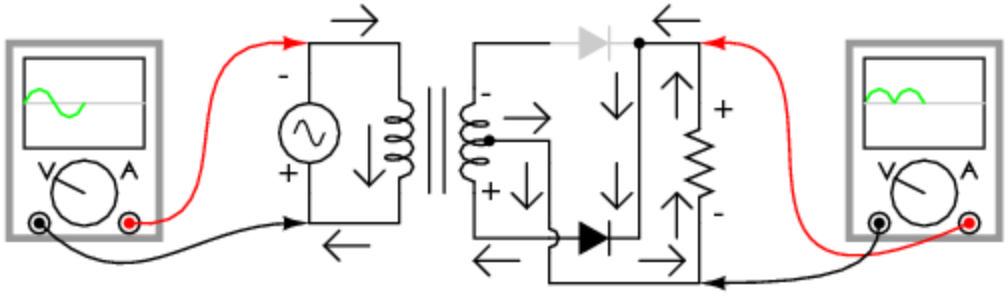
This circuit's operation is easily understood one half-cycle at a time. Consider the first half-cycle, when the source voltage polarity is positive (+)

on top and negative (-) on bottom. At this time, only the top diode is conducting; the bottom diode is blocking current, and the load "sees" the first half of the sine wave, positive on top and negative on bottom. Only the top half of the transformer's secondary winding carries current during this half-cycle as in Figure below.



Full-wave center-tap rectifier: Top half of secondary winding conducts during positive half-cycle of input, delivering positive half-cycle to load..

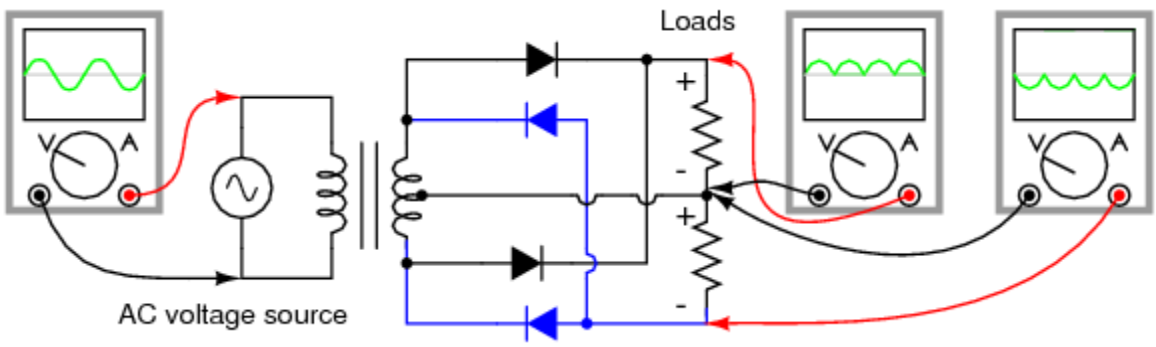
During the next half-cycle, the AC polarity reverses. Now, the other diode and the other half of the transformer's secondary winding carry current while the portions of the circuit formerly carrying current during the last half-cycle sit idle. The load still "sees" half of a sine wave, of the same polarity as before: positive on top and negative on bottom. (Figure below)



Full-wave center-tap rectifier: During negative input half-cycle, bottom half of secondary winding conducts, delivering a positive half-cycle to the load.

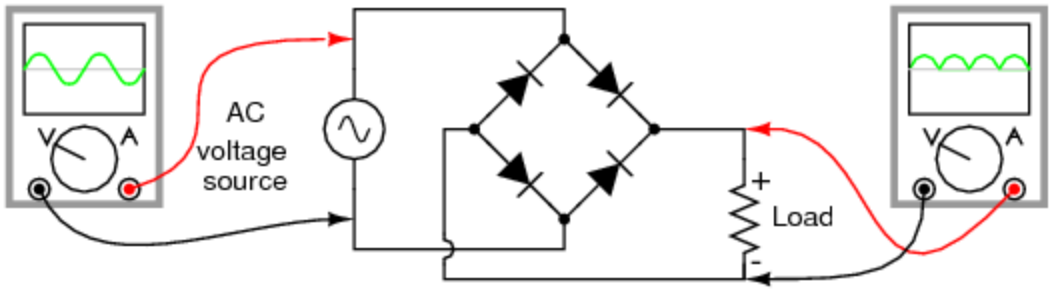
One disadvantage of this full-wave rectifier design is the necessity of a transformer with a center-tapped secondary winding. If the circuit in question is one of high power, the size and expense of a suitable transformer is significant. Consequently, the center-tap rectifier design is only seen in low-power applications.

The full-wave center-tapped rectifier polarity at the load may be reversed by changing the direction of the diodes. Furthermore, the reversed diodes can be paralleled with an existing positive-output rectifier. The result is dual-polarity full-wave center-tapped rectifier in Figure below. Note that the connectivity of the diodes themselves is the same configuration as a bridge.



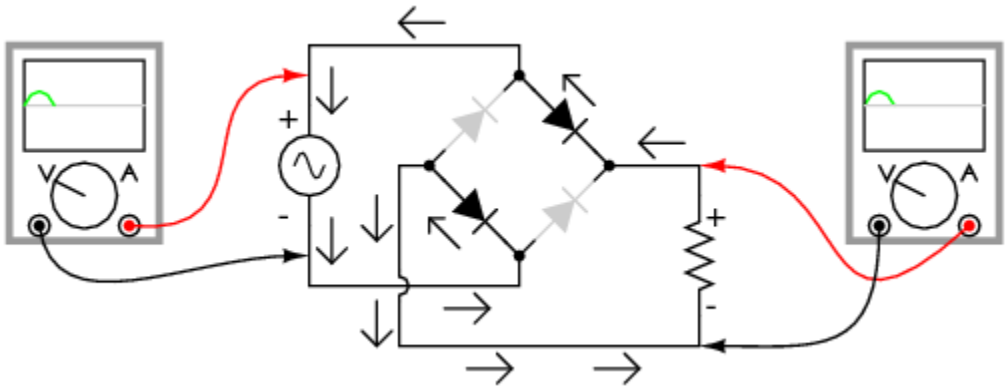
Dual polarity full-wave center tap rectifier

Another, more popular full-wave rectifier design exists, and it is built around a four-diode bridge configuration. For obvious reasons, this design is called a *full-wave bridge*. (Figure below)

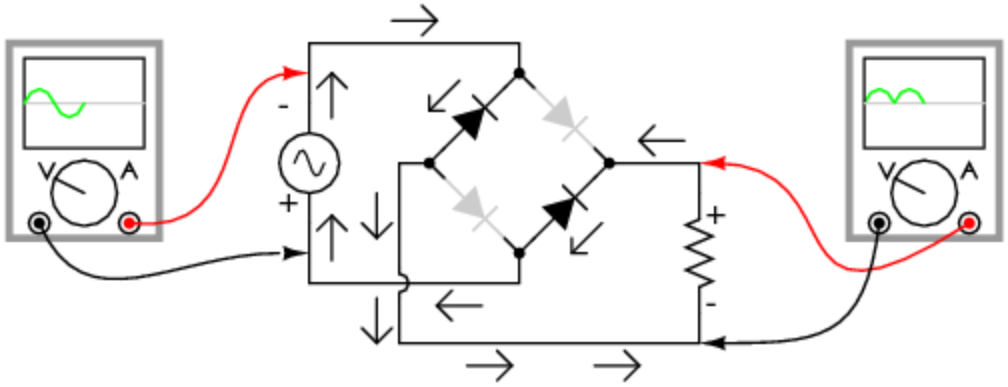


Full-wave bridge rectifier.

Current directions for the full-wave bridge rectifier circuit are as shown in Figure below for positive half-cycle and Figure below for negative half-cycles of the AC source waveform. Note that regardless of the polarity of the input, the current flows in the same direction through the load. That is, the negative half-cycle of source is a positive half-cycle at the load. The current flow is through two diodes in series for both polarities. Thus, two diode drops of the source voltage are lost (0.7·2=1.4 V for Si) in the diodes. This is a disadvantage compared with a full-wave center-tap design. This disadvantage is only a problem in very low voltage power supplies.

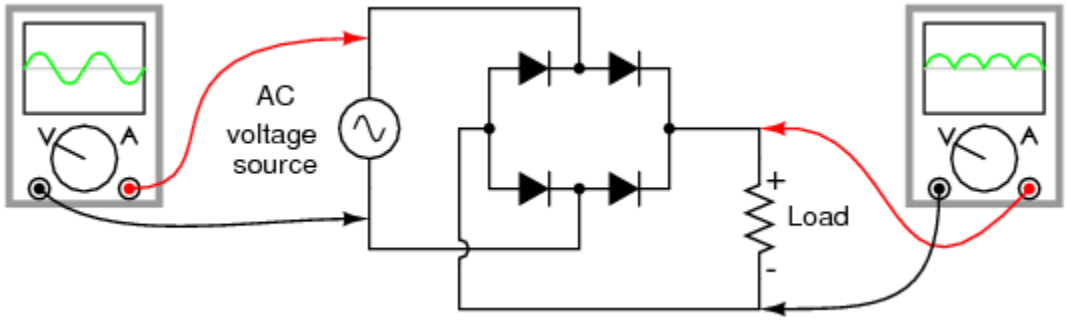


Full-wave bridge rectifier: Electron flow for positive half-cycles.



Full-wave bridge rectifier: Electron flow for negative half=cycles.

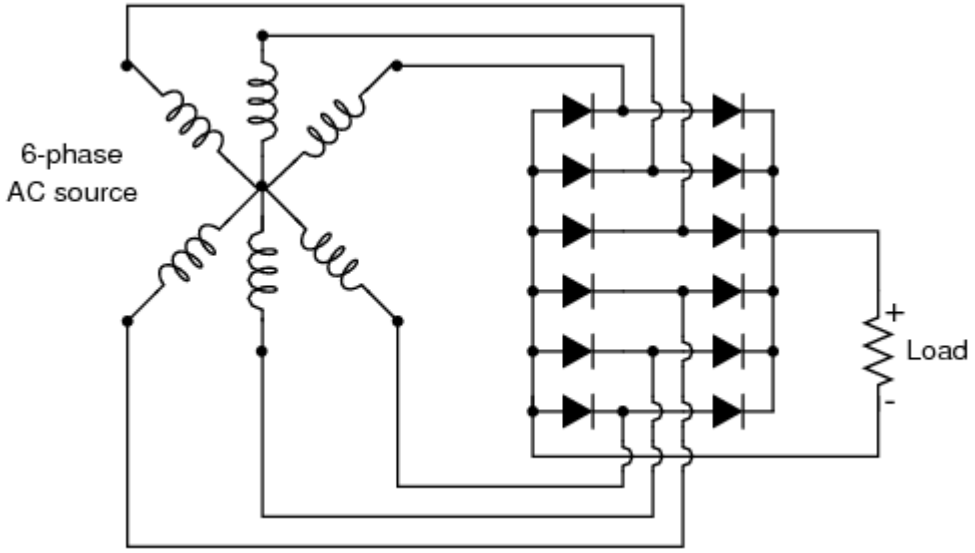
Remembering the proper layout of diodes in a full-wave bridge rectifier circuit can often be frustrating to the new student of electronics. I've found that an alternative representation of this circuit is easier both to remember and to comprehend. It's the exact same circuit, except all diodes are drawn in a horizontal attitude, all "pointing" the same direction. (Figure below)



Alternative layout style for Full-wave bridge rectifier.

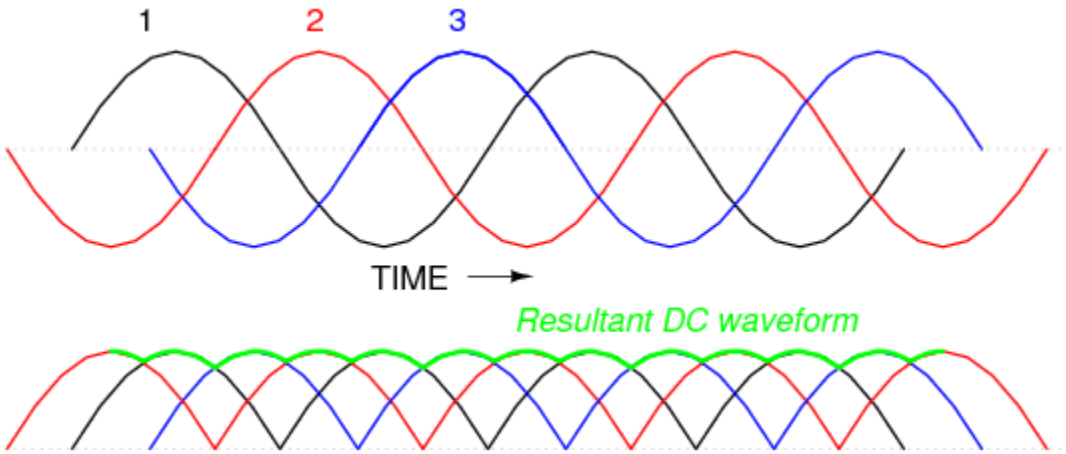
hree-phase full-wave bridge rectifier circuit.

Each three-phase line connects between a pair of diodes: one to route power to the positive (+) side of the load, and the other to route power to the negative (-) side of the load. Polyphase systems with more than three phases are easily accommodated into a bridge rectifier scheme. Take for instance the six-phase bridge rectifier circuit in Figure below.



Six-phase full-wave bridge rectifier circuit.

When polyphase AC is rectified, the phase-shifted pulses overlap each other to produce a DC output that is much "smoother" (has less AC content) than that produced by the rectification of single-phase AC. This is a decided advantage in high-power rectifier circuits, where the sheer physical size of filtering components would be prohibitive but low-noise DC power must be obtained. The diagram in Figure below shows the full-wave rectification of three-phase AC.



Three-phase AC and 3-phase full-wave rectifier output.

In any case of rectification — single-phase or polyphase — the amount of AC voltage mixed with the rectifier's DC output is called *ripple voltage*. In most cases, since "pure" DC is the desired goal, ripple voltage is undesirable. If the power levels are not too great, filtering networks may be employed to reduce the amount of ripple in the output voltage.

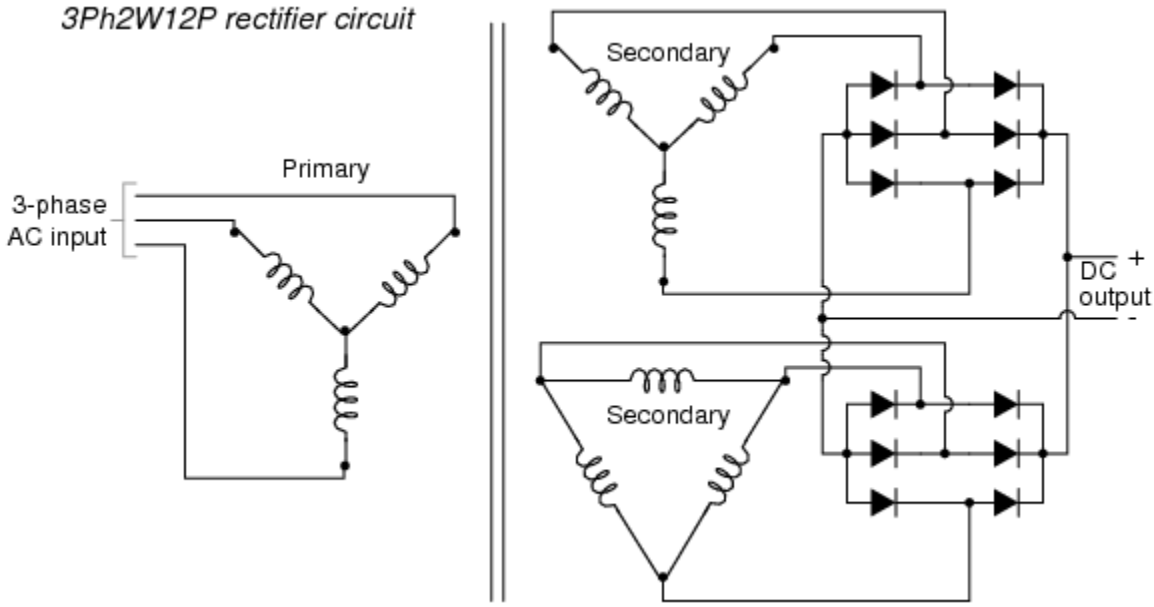
Sometimes, the method of rectification is referred to by counting the number of DC "pulses" output for every 360^{o} of electrical "rotation." A single-phase, half-wave rectifier circuit, then, would be called a *1-pulse* rectifier, because it produces a single pulse during the time of one complete cycle (360^{o}) of the AC waveform. A single-phase, full-wave rectifier (regardless of design, center-tap or bridge) would be called a *2-pulse* rectifier, because it outputs two pulses of DC during one AC cycle's worth of time. A three-phase full-wave rectifier would be called a *6-pulse* unit.

Modern electrical engineering convention further describes the function of a rectifier circuit by using a three-field notation of *phases*, *ways*, and number

of *pulses*. A single-phase, half-wave rectifier circuit is given the somewhat cryptic designation of 1Ph1W1P (1 phase, 1 way, 1 pulse), meaning that the AC supply voltage is single-phase, that current on each phase of the AC supply lines moves in only one direction (way), and that there is a single pulse of DC produced for every 360^{o} of electrical rotation. A single-phase, full-wave, center-tap rectifier circuit would be designated as 1Ph1W2P in this notational system: 1 phase, 1 way or direction of current in each winding half, and 2 pulses or output voltage per cycle. A single-phase, full-wave, bridge rectifier would be designated as 1Ph2W2P: the same as for the center-tap design, except current can go *both* ways through the AC lines instead of just one way. The three-phase bridge rectifier circuit shown earlier would be called a 3Ph2W6P rectifier.

Is it possible to obtain more pulses than twice the number of phases in a rectifier circuit? The answer to this question is yes: especially in polyphase circuits. Through the creative use of transformers, sets of full-wave rectifiers may be paralleled in such a way that more than six pulses of DC are produced for three phases of AC. A 30^{o} phase shift is introduced from primary to secondary of a three-phase transformer when the winding configurations are not of the same type. In other words, a transformer connected either Y-? or ?-Y will exhibit this 30^{o} phase shift, while a transformer connected Y-Y or ?-? will not. This phenomenon may be exploited by having one transformer connected Y-Y feed a bridge rectifier, and have another transformer connected Y-? feed a second bridge rectifier, then parallel the DC outputs of both rectifiers. (Figure below) Since the ripple voltage waveforms of the two rectifiers' outputs are phase-shifted

30°from one another, their superposition results in less ripple than either rectifier output considered separately: 12 pulses per 360° instead of just six:



Polyphase rectifier circuit: 3-phase 2-way 12-pulse (3Ph2W12P)

Design of zener and transistor series voltage regulators

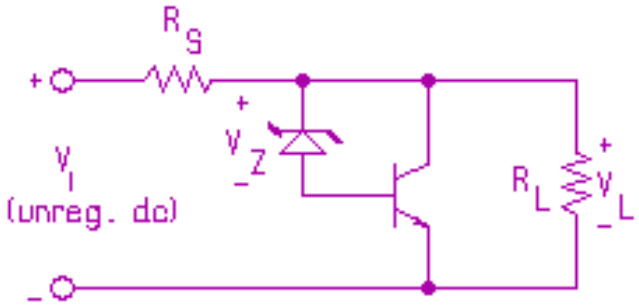
TYPES OF VOLTAGE REGULATORS:

1) Shunt voltage regulator
2) Series voltage regulator

1) Shunt voltage regulator :

The block diagram of the shunt voltage regulator is as shown in figure. The control element is connected in parallel with the load therefore it is called as shunt voltage regulator. Vin is a unregulated voltage which is

obtained from rectifier o/p. The total current is shared by the load & control element .The current flowing through control element is Ish.



Summary of regulatory action of a shunt regulator:

- If output voltage increases above regulated value

▼

Feedback signal increases

▼

Control signal increases

▼

Control element draws more current

▼

Load current decreases as I is constant

▼

Output voltage is brought back to its regulated value

- If output voltage decreases below regulated value

▼

Feedback signal decreases

▼

Control signal decreases

▼

Control element draws less current

▼

Load current increases as I is constant

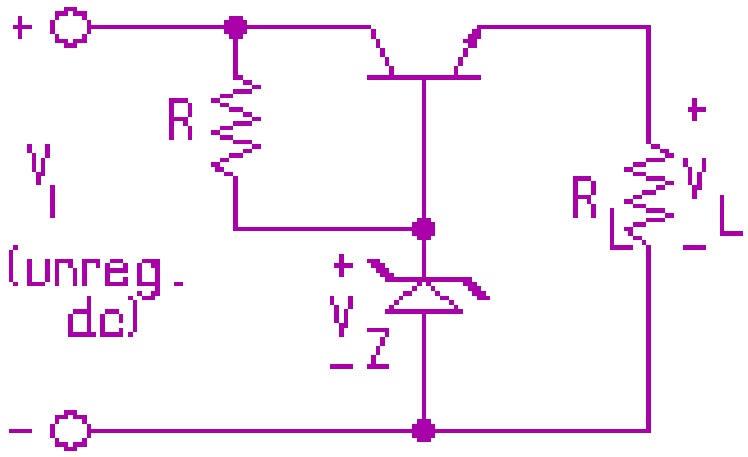
▼

Output voltage is brought back to its regulated value

2) Series voltage regulator:

The block diagram of the series voltage regulator is shown in figure. In this the control element is connected in series with the load therefore it is called as series voltage regulator. As control element is in series with the load the expression for the output voltage is given by:

Vo = Vin –



Vs

Fig: Transistor Series Voltage Regulator

Summary of regulatory action of a series regulator:

- If output voltage increases above regulated value

▼

Feedback signal increases

▼

Control signal decreases

▼

Voltage across control element increases

▼

Output voltage is brought back to its regulated value

- If output voltage decreases below regulated value

▼

Feedback signal decreases

▼

Control signal increases

▼

Voltage across control element decreases

▼

Output voltage is brought back to its regulated

ZENER DIODE AS A VOLTAGE REGULATOR

The voltage across zener diode is remains constant when it is operated in zener region of reverse characteristics. This fact is utilized in applications as

a voltage regulator. Zener diode regulator is a shunt type regulator. The reverse characteristics are shown in fig

The regulatory action can be summarized as follows:

- If output voltage increases above regulated value

▼

Then I increases

▼

but IL is constant

▼

Hence Iz increases

▼

If Iz< Izmax then zener operates in the

Zener region & output voltage remains constant.

- Regulating action with varying Load (constant Vin):

If RL increases then IL will decrease

▼

But I is constant & Iz = I- IL

▼

Hence Iz increases

▼

Vo is constant as long as Iz < Izmax &

Vice versa action if RL decreases.

.

ZENER DIODE AS A VOLTAGE REGULATOR

The voltage across zener diode is remains constant when it is operated in zener region of reverse characteristics. This fact is utilized in applications as a voltage regulator. Zener diode regulator is a shunt type regulator. The reverse characteristics are shown in fig

Regulating action with varying input voltage(constant IL):

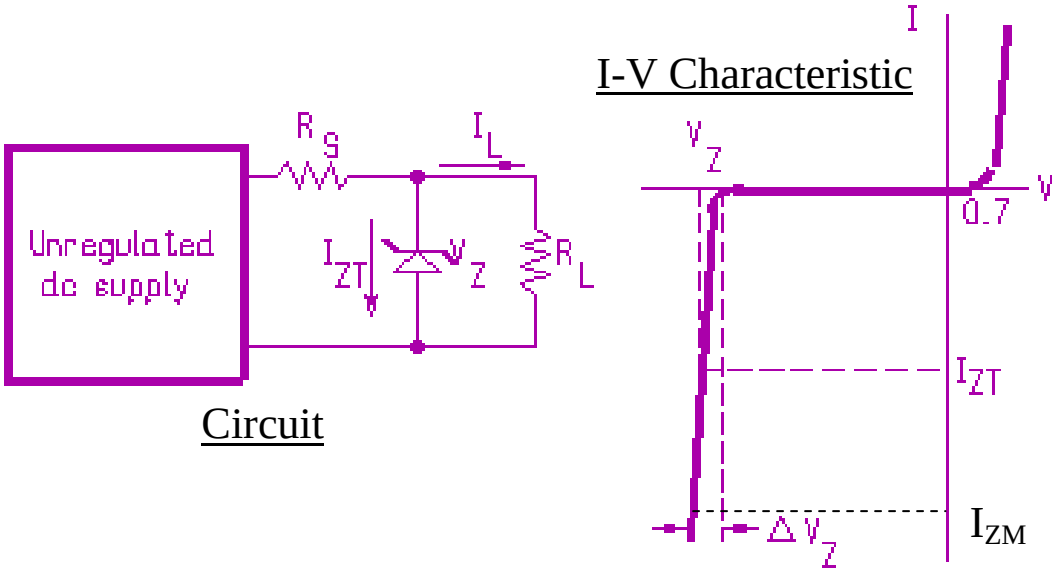
We assume that RL is constant & Vin is varying. IL is also constant because IL = Vz/ RL but supply current keeps changing due to change in Vs.

$$I = Vin - Vz/Rs$$

*NOTES ON ZENER DIODE REGULATOR:

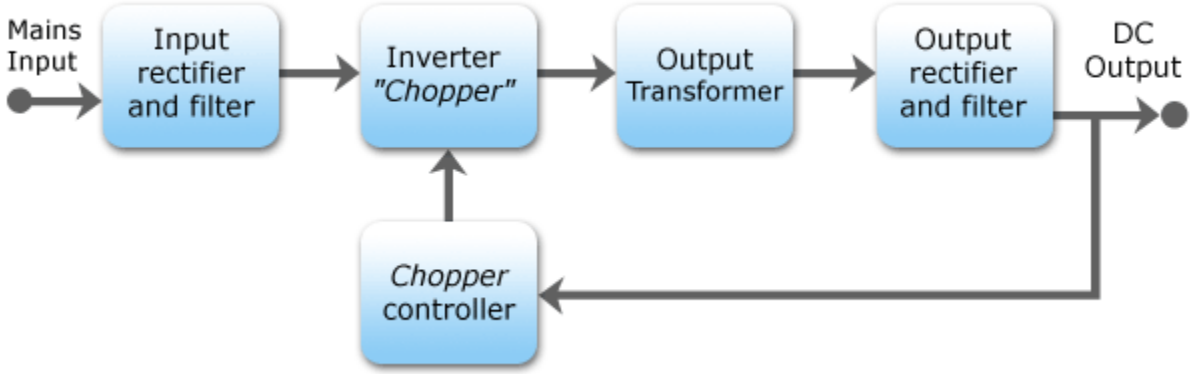
- V_Z depends on I and temperature.
- Zener diodes with rated voltage < 6 V have negative temperature coefficient; those rated > 6 V have positive temperature coefficient.
- In order to maintain a constant V_o, I_{ZT} varies in response to a change of either I_L or V_i. For example, when R_L increases, I_L decreases, then I_{ZT} has to increase to keep the current through R_s constant. Since the voltage drop across R_s is constant, V_o stays constant.

Zener Diode Voltage Regulator



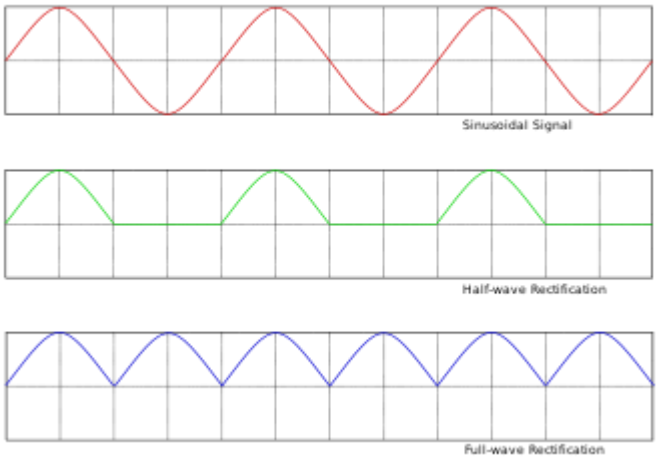
SWITCHED MODE POWER SUPPLY

Theory of operation



Block diagram of a mains operated AC–DC SMPS with output voltage regulation

Input rectifier stage



AC, half-wave and full-wave rectified signals.

If the SMPS has an AC input, then the first stage is to convert the input to DC. This is called *rectification*. The rectifier circuit can be configured as a voltage doubler by the addition of a switch operated either manually or automatically. This is a feature of larger supplies to permit operation from nominally 120 V or 240 V supplies. The rectifier produces an unregulated DC voltage which is then sent to a large filter capacitor. The current drawn from the mains supply by this rectifier circuit occurs in short pulses around the AC voltage peaks. These pulses have significant high frequency energy which reduces the power factor. Special control techniques can be employed

by the following SMPS to force the average input current to follow the sinusoidal shape of the AC input voltage thus the designer should try correcting the power factor. An SMPS with a DC input does not require this stage. An SMPS designed for AC input can often be run from a DC supply (for 230 V AC this would be 330 V DC), as the DC passes through the rectifier stage unchanged. It's however advisable to consult the manual before trying this, though most supplies are quite capable of such operation even though nothing is mentioned in the documentation. However, this type of use may be harmful to the rectifier stage as it will only use half of diodes in the rectifier for the full load. This may result in overheating of these components, and cause them to fail prematurely. [4]

If an input range switch is used, the rectifier stage is usually configured to operate as a voltage doubler when operating on the low voltage (~120 V AC) range and as a straight rectifier when operating on the high voltage (~240 V AC) range. If an input range switch is not used, then a full-wave rectifier is usually used and the downstream inverter stage is simply designed to be flexible enough to accept the wide range of DC voltages that will be produced by the rectifier stage. In higher-power SMPSs, some form of automatic range switching may be used.

Inverter stage

This section refers to the block marked *chopper* in the block diagram.

The inverter stage converts DC, whether directly from the input or from the rectifier stage described above, to AC by running it through a power

oscillator, whose output transformer is very small with few windings at a frequency of tens or hundreds of kilohertz (kHz). The frequency is usually chosen to be above 20 kHz, to make it inaudible to humans. The output voltage is optically coupled to the input and thus very tightly controlled. The switching is implemented as a multistage (to achieve high gain) MOSFET amplifier. MOSFETs are a type of transistor with a low on-resistance and a high current-handling capacity.

Voltage converter and output rectifier

If the output is required to be isolated from the input, as is usually the case in mains power supplies, the inverted AC is used to drive the primary winding of a high-frequency transformer. This converts the voltage up or down to the required output level on its secondary winding. The output transformer in the block diagram serves this purpose.

If a DC output is required, the AC output from the transformer is rectified. For output voltages above ten volts or so, ordinary silicon diodes are commonly used. For lower voltages, Schottky diodes are commonly used as the rectifier elements; they have the advantages of faster recovery times than silicon diodes (allowing low-loss operation at higher frequencies) and a lower voltage drop when conducting. For even lower output voltages, MOSFETs may be used as synchronous rectifiers; compared to Schottky diodes, these have even lower conducting state voltage drops.

The rectified output is then smoothed by a filter consisting of inductors and capacitors. For higher switching frequencies, components with lower capacitance and inductance are needed.

Simpler, non-isolated power supplies contain an inductor instead of a transformer. This type includes *boost converters*, *buck converters*, and the so called *buck-boost converters*. These belong to the simplest class of single input, single output converters which use one inductor and one active switch. The buck converter reduces the input voltage in direct proportion to the ratio of conductive time to the total switching period, called the duty cycle. For example an ideal buck converter with a 10 V input operating at a 50% duty cycle will produce an average output voltage of 5 V. A feedback control loop is employed to regulate the output voltage by varying the duty cycle to compensate for variations in input voltage. The output voltage of a boost converter is always greater than the input voltage and the buck-boost output voltage is inverted but can be greater than, equal to, or less than the magnitude of its input voltage. There are many variations and extensions to this class of converters but these three form the basis of almost all isolated and non-isolated DC to DC converters. By adding a second inductor the Cuk and SEPIC converters can be implemented, or, by adding additional active switches, various bridge converters can be realised.

Other types of SMPSs use a capacitor-diode voltage multiplier instead of inductors and transformers. These are mostly used for generating high voltages at low currents (*Cockcroft-Walton generator*). The low voltage variant is called charge pump.

Regulation

A feedback circuit monitors the output voltage and compares it with a reference voltage, which shown in the block diagram serves this purpose. Depending on design/safety requirements, the controller may contain an

isolation mechanism (such as opto-couplers) to isolate it from the DC output. Switching supplies in computers, TVs and VCRs have these opto-couplers to tightly control the output voltage.

Open-loop regulators do not have a feedback circuit. Instead, they rely on feeding a constant voltage to the input of the transformer or inductor, and assume that the output will be correct. Regulated designs compensate for the impedance of the transformer or coil. Monopolar designs also compensate for the magnetic hysteresis of the core.

The feedback circuit needs power to run before it can generate power, so an additional non-switching power-supply for stand-by is added.

Transformer design

SMPS transformers run at high frequency. Most of the cost savings (and space savings) in off-line power supplies come from the fact that a high frequency transformer is much smaller than the 50/60 Hz transformers formerly used. There are additional design tradeoffs.

Transformer size

The higher the switching frequency, the lesser the amount of energy that needs to be stored intermediately during the time of a single switching cycle. Because this energy is stored in form of magnetic energy in the transformer core material (like ferrite), less of such material is needed.

However, higher frequency also means more energy lost during transitions of the switching semiconductor. Furthermore, much more attention to the

physical layout of the circuit board is required, and the amount of electromagnetic interference will be more pronounced.

Core loss

There are several differences in the design of transformers for 50 Hz vs 500 kHz. Firstly a low frequency transformer usually transfers energy through its core (soft iron), while the (usually ferrite) core of a high frequency transformer limits leakage.

Copper loss

Main article: Copper loss

At low frequencies (such as the line frequency of 50 or 60 Hz), designers can usually ignore the skin effect. At line frequencies, the skin effect becomes important when the conductors have a diameter larger than about 0.3 inches (7.6 mm).

Switching power supplies must pay more attention to the skin effect because it is a source of power loss. At 500 kHz, the skin depth is about 0.003 inches (0.076 mm) – a dimension smaller than the typical wires used in a power supply.

The skin effect is exacerbated by the harmonics present in the switching waveforms. The appropriate skin depth is not just the depth at the fundamental, but also the skin depths at the harmonics.[5]

Since the waveforms in a SMPS are generally high speed (PWM square waves), the wiring must be capable of supporting high harmonics of the base frequency due to skin effect.

In addition to the skin effect, there is also a proximity effect, which is another source of power loss.

Power factor

Main article: power factor

Simple off-line switched mode power supplies incorporate a simple full wave rectifier connected to a large energy storing capacitor. Such SMPSs draw current from the AC line in short pulses when the mains instantaneous voltage exceeds the voltage across this capacitor. During the remaining portion of the AC cycle the capacitor provides energy to the power supply.

As a result, the input current of such basic switched mode power supplies has high harmonic content and relatively low power factor. This creates extra load on utility lines, increases heating of the utility transformers and standard AC electric motors, and may cause stability problems in some applications such as in emergency generator systems or aircraft generators. Harmonics can be removed through the use of filter banks but the filtering is expensive, and the power utility may require a business with a very low power factor to purchase and install the filtering onsite.

Unlike displacement power factor created by linear inductive or capacitive loads, this distortion cannot be corrected by addition of a single linear component. Additional circuits are required to counteract the effect of the brief current pulses.

In 2001, the European Union put into effect the standard IEC/EN61000-3-2 to set limits on the harmonics of the AC input current up to the 40th harmonic for equipment above 75 W. The standard defines four classes of equipment depending on its type and current waveform. The most rigorous limits (class D) are established for personal computers, computer monitors, and TV receivers. In order to comply with these requirements modern switched-mode power supplies normally include an additional power factor correction (PFC) stage.

Putting a current regulated boost chopper stage after the off-line rectifier (to charge the storage capacitor) can correct the power factor, but increases the complexity (and any cost).

Question Bank

PART –A

1.What is clipper?

2. What is a Schmitt trigger?

PART-B

1.(a) Explain the working of a mono-stable multi-vibrator. Draw the waveforms at the collector and base of the transistors.

(b) Write short notes on UJT based saw-tooth oscillator.

2. (a) Explain the working of a seres voltage regulator and and perform a simplified analysis of the same. (b) Write short notes on SMPS.

NPR COLLEGE OF ENGINEERING AND TECHNOLOGY

NATHAM

DEPT:EEE **SEM:III**

EC1209 – ELECTRONIC DEVICES AND CIRCUITS

PART-A **(10*2=20)**

1. How a PN junction can be formed?
2. Define avalanche and zener breakdown?
3. Draw and explain the VI characteristics of pn junction diode
4. Define Reverse saturation current.
5. Define Q point.
6. what is mean by 'Thermal runaway'?
7. Define Stability factor.
8. Give the relation between the a & β .
9. Define "punch through" in early effect
10. Draw the CE configuration.

PART-B

11a ..Explain the PN diode

(or)

b.Discuss about Zener

12.a. Explain the BIAS STABILITY

a.DC Load line b. AC Load line.

(or)

b. Explain the BIAS Compensation

a.Diode compensation b. Thermistor compensation

NPR COLLEGE OF ENGINEERING AND TECHNOLOGY-NATHAM

II-Internal Test

Subject Name: Electronic Devices & circuits **Sem : III**

Subject Code: EC1209 **Date : 27:08:2010**

PART-A **(2*10=20)**

1) What is mean by FET?
2) Define Pinch off Voltage.
3) What is mean by Ohmic region?
4) Define latching current.
5) Define Holding current.
6) Define Gate Current.
7) Define Voltage safety factor.
8) What is mean Relaxation Capacitor?
9) Distinguish between JFET and SCR.
10) Give advantages of SCR and MOSFET.

PART-B **(2*15=30)**

1) Draw and explain the Characteristics of MOSFET.(OR)
2) Draw and explain the Characteristics of TRIAC and DIAC.
3) Explain the Half wave and full wave Rectifier by using SCR. (OR)
4) Draw and explain the Characteristics of UJT.

N.P.R. COLLEGE OF ENGINEERING & TECHNOLOGY

Natham - 624 401,

Subject Name: Electronic devices and Circuits | **Dept/Year/Sem**: II EEE / III sem

Marks: 50 | **Time**: 1 hour 30minutes

Third Internal

PART A

Answer all questions: (10*2=20)

1.Write any two advantages of Negative feedback ?

2.What are the conditions of Oscillator?

3.What are the Limitation of RC Oscillator?

4.Write the frequency response of the Hartley Oscillator?

5.Draw the circuit diagram of the Colpitts Oscillator?

6.What are the Difference between Positive feedback and Negative feedback?

7. What are the Difference between Voltage amplifier & Power amplifier?

8.What are the Classification of Power amplifier?

9.Draw the circuit diagram of RC amplifier?

10.What are advantages of the Push Pull amplifier?

PART B

Answer the following Questions : (2*15=30)

11.Explain the Voltage series feedback.(OR)

12. Explain the Current shunt feedback.

13. Explain the Wien Bridge Oscillator.(OR)

14. Explain the Push Pull Amplifier.

NPR COLLEGE OF ENGINEERING AND TECHNOLOGY

THIRD SEMESTER

EC1209-ELECTRON DEVICES AND CIRCUITS

ELECCTRICAL AND ELECTRONICS ENGINEERING

Time: Three hours Maximum:100

Answer ALL questions.

PART A-(10*2=20)

1.Which diode is used for rectification? Why?

2.What is a condition for Thermal stability?

3. Differentiate between JFET and MOSFET.

4. What are the advantages of SCR.

5. Why the gain of the amplifier decreases in the low frequency and high frequency range?

6.Which amplifiers are classified as power amplifier?

7. What is the effect of feedback on amplifier bandwidth?

8.What is the principle of operation of Hartley Oscillator?

9.What is a multivibrator?

10. How a shunt regulator is differentiated from series regulator?

PART B-(5*16=80)

11. (a) Write short notes on varactor diode and schottky diode.(8)

(b) Explain the characteristics of Zener diode. (8)

Or

12. (a) Design a collector to base bias circuit for the specified conditions Vcc=15V,Vcc=5V,Ic=5mA and β=100. (8)

(b) What is mean by bias stability? What are the factors that affect BJT biasing(8)?

13. (a) With the help of a neat diagram ,explain the voltage divider biasing metho for JFET. (8)

(b) Explain the characteristics of MOSFET. Explain the types. (8)

Or

14. (a) Explain the characteristics of UJT and Explain the one application of UJT. (8)

(b)Explain why the input resistance of the FET is so very high. (8)

15. (a)With small signal equivalent circuit how to analyze the performance in CE amplifier with bypass resistance if $h_{oe} R_L < 0.1$. (8)

(b) Consider a single stage CE amplifier with $R_1 = 1k\Omega$ $R_1 = 50k\Omega$ $R_2 = 2\ k\Omega$, $R_C = 2\ k\Omega$, $h_{fe} = 50$, $h_{ie} = 1.1\ k\Omega$, $h_{oe} = 25\mu A/V$ and $h_{re} = 2.5*10^{-4}$ Find the Voltage & current gain,input & output resistance. (8)

Or

16.(a) What is crossover distortion ? How to eliminate it in Audio frequency power Amplifier? (8)

(b)Draw a neat circuit diagram of Push Pull class B amplifier& & Explain it operation. (8)

17. What is the general characteristics of Negative feedback amplifier . Derive the characteristics of current series feedback.(16)

Or

18. Explain in detail the principle of Wien & RC Oscillators(16)

19.(a) Write short notes on Clippers & Clampers. (8)

(b) Explain the operation of Astable multivibrators. (8)

Or

20.(a)With a capacitor filterexplain the working of a full wave rectifier and Obtain is ripple factor. Compare it with π filter. (8)

(b) Explain in detail about Switch mode power supply& explain the types(8)

E 0130

B.E./B.Tech. DEGREE EXAMINATION, NOVEMBER/DECEMBER 2009

THIRD SEMESTER

ELECTRICAL AND ELECTRONICS ENGINEERING

EC1209 ELECTRON DEVICES AND CIRCUITS

(Common to EIE and ICE)

(REGULATION 2008)

Time : Three hours Maximum : 100 marks

Answer ALL questions.

PART A — ($10 \times 2 = 20$ marks)

1. What are the current components in a diode?

2. What do you understand by Q-point? What is its significance?

3. Why is thermal runaway not present in FET?

4. What is TRIAC?

5. Draw the small signal low frequency hybrid model of common base configuration driven by an ideal voltage source.

6. What is crossover distortion?

7. What are the advantages of negative feedback?

8. State the conditions for oscillation.

9. What is a clamper?

10. What is the need for filter circuit?

PART B — (5 × 16 = 80 marks)

11. Explain the Tunneling phenomenon in a Tunnel Diode with transfer characteristic curves and state the advantages. (16)

Or

12. (a) Explain how the self biasing stabilizes the operating point with associated diagrams. (8)

(b) Design a collector to base bias circuit to have an operating point of (10 V, 4mA). The circuit is supplied with 20 V and uses a silicon transistor of h_{fe} 250. (8)

13. (a) Explain the fixed biasing of n-channel JFET circuits. (8)

(b) Explain in detail the biasing circuits of EMOSFET. (8)

Or

14. (a) Draw the characteristics of UJT and explain its operation in detail. (8)

(b) With the characteristics curve explain the operation of a SCR. (8)

15. A common collector circuit has the following components: R1 = 27KΩ; R2 = 27KΩ $R_E = 5.6K\Omega$, $R_L = 47K\Omega$, $Rs = 600\Omega$. The transistor parameters are $h_{ie} = 1k\Omega$, $h_{fe} = 85$ $h_{oe} = 2\mu A/V$. Calculate Ai, Ri, Av, Ro, Avs=Vo/Vs and Ai=Io/Is. (16)

Or

16. (a) Explain the frequency response of an RC coupled amplifier. (8)

(b) Draw a Quasi complementary symmetry power amplifier and explain its operations. (8)

17. Explain the operation of a current series and current shunt feedback amplifier. (16)

Or

18. With the Equivalent circuit diagram explain the operation of RC phase shift Oscillator. (16)

19. (a) Explain the operation of Schmitt trigger with neat sketch. (8)

(b) Explain how a diode can be used as a clipper. (8)

Or

20. (a) Explain the operation of single phase rectifier with LC filter. (6)

(b) Explain the operation of SMPS with neat block diagram. (10)

F 0454

B.E./B.Tech. DEGREE EXAMINATION, APRIL/MAY 2010

THIRD SEMESTER

ELECTRICAL AND ELECTRONICS ENGINEERING

EC1209 — ELECTRON DEVICES AND CIRCUITS

(Common to EIE, ICE)

(REGULATION 2008)

Time : Three hours Maximum : 100 marks

Answer ALL questions.

PART A — (10 × 2 = 20 marks)

1. Which diode can be used for regulation? Why?

2. What is a condition for Thermal stability?

3. Draw two different circuits that bias a JFET amplifier.

4. State the differences between UJT and FET.

5. Why the gain of the amplifier decreases in the low frequency and high frequency range?

6. Which amplifiers are classified as power amplifiers?

7. What is the effect of feedback on amplifier bandwidth?

8. What is the principle of operation of crystal oscillator?

9. What is a multivibrator?

10. How a shunt regulator is differentiated from series regulator?

PART B — (5 × 16 = 80 marks)

11. (a) Write short notes on varacter diode and schottky diode. (8)

(b) Explain the characteristics of pn junction diode. (8)

Or

12. (a) Design a collector to base bias circuit for the specified conditions $V_{cc} = 15\ V$, $V_{BE} = 5V$, $I_C = 5\,mA$ and $\beta = 100$. (8)

(b) What is meant by bias stability? What are the factors that affect BJT biasing? (8)

13. (a) With the help of a neat diagram, explain the voltage divider biasing method for JFET. (8)

(b) With the characteristic curves explain the principle of operations of DMOSFET and EMOSFET. (8)

Or

14. (a) Explain any one applications of UJT with circuit diagram. (8)

(b) Derive the expression to find the operating point of JFET with fixed bias. (8)

15. (a) With small signal equivalent circuit how to analyze the performance in CE amplifier with bypass resistance if $h_{oe}R_L < 0.1$. (8)

(b) Consider a single stage CE amplifier with $R_s = 1\ k\Omega$, $R_1 = 50\ K\Omega$, $R_2 = 2\ K\Omega$, $R_c = 2\ K\Omega$, $h_{fe} = 50$, $h_{ie} = 1.1\ k\Omega$, $h_{oe} = 25\ \mu\ A/V$ and $h_{re} = 2.5 \times 10^{-4}$. Find the voltage and current gain, Input and output resistance. (8)

Or

16. (a) What is cross over Distortion? How to eliminate it in Audio frequency power amplifier? (8)

(b) Draw a neat circuit diagram of push pull class B amplifier and explain its operations. (8)

17. What is the general characteristics of negative feedback amplifier. Derive the characteristics of Voltage series feedback. (16)

Or

18. Explain in detail the principle of operations of Hartley and Colpitts oscillators. (16)

19. (a) Write short notes on Clippers and Clampers. (8)s

(b) Explain the operations of Bistable multivibrators. (8)

Or

20. (a) With a neat circuit diagram explain the operations of transistorized series type voltage regulator. (8)

(b) With a capacitor filter explain the working of a half wave rectifier and obtain its ripple factor. Compare it with π filter. (8)

G 0398

B.E./B.Tech. DEGREE EXAMINATION, NOVEMBER/DECEMBER 2010

THIRD SEMESTER

ELECTRICAL AND ELECTRONICS ENGINEERING

EC1209 ELECTRONIC DEVICES AND CIRCUITS

(Common to EIE, ICE)

(REGULATION 2008)

Time : Three hours Maximum : 100 marks

Answer ALL questions.

PART A — ($10 \times 2 = 20$ marks)

1. What is barrier potential?

2. Define biasing

3. Define Pinch-off voltage.

4. What is intrinsic stand-off ratio of UJT?

5. What is emitter follower?

6. What are Class AB Amplifiers?

7. List the two important types of RC oscillators.

8. What is the main advantage of crystal oscillator?

9. What is clipper?

10. What is a Schmitt trigger?

PART B — (5 × 16 = 80 marks)

11. (a) Explain the working of varactor diode. (8)

(b) What does a dc load line represent? May this line be used when an ac signal is applied? When it is necessary to draw an ac load line? (8)

Or

12. (a) Explain the input and output characteristics of a transistor in common emitter configuration with neat diagram. (8)

(b) Explain how the zener diode acts as a voltage regulator with relevant diagram. (8)

13. (a) Describe the different manufacturing techniques adopted to reduce the magnitude of gate source threshold voltage V_j in enhancement MOSFET. (8)

(b) Compare the performance of JFET with MOSFET. (4)

(c) Discuss the relative performance of p-channel MOSFET and n-channel MOSFET. (4)

Or

14. (a) Explain the working and characteristics of DIAC. (6)

(b) Write short notes on MOSFET biasing. (10)

15. Obtain the current gain, voltage gain, output admittance and power gain of ac common emitter amplifier using h-parameter analysis. Also give the relation between AVs and AIs. (16)

Or

16. A CC amplifier (emitter follower) is fed from a voltage source Vs of internal resistance $R_s = 800\ \Omega$. The load impedance is a resistor $R_L = 1600\ \Omega$. The CE h-parameters are $h_{ie} = 1000\ \Omega$, $h_{re} = 2.2 \times 10^{-4}$, $h_{fe} = 55$ and $h_{oe} = 23\ \mu\mho$. Compute the current gain A_i, input resistance R_i, voltage gain Av, output resistance R_o and output terminal resistance R_{os} using (a) exact analysis and (b) approximate analysis. (16)

17. Explain the working of a transistorized Hartley oscillator. Derive expressions for frequency of oscillation and condition for sustained oscillation. (16)

Or

18. Draw the equivalent circuit of current amplifier with current shunt feedback and derive expressions for (a) the output resistance with feedback R_{of} (b) Output terminal resistance with feedback R_{it}. (16)

19. (a) Explain the working of a mono-stable multi-vibrator. Draw the waveforms at the collector and base of both the transistors. (10)

(b) Write short notes on UJT based saw-tooth oscillator. (6)

Or

20. (a) Explain the working of a series voltage regulator and perform a simplified analysis of the same. (10)

(b) Write short notes on SMPS. (6)