5.1 ACTIVE RF COMPONENTS:

SEMICONDUCTOR BASCS IN RF:

PHYSISCAL PROPERTIES OF SEMICONDUCTORS:

The operation of semiconductor devices is naturally dependent on the physic behavior of the semiconductors themselves. This section presents a brief introduction to the basic building blocks of semiconductor device modeling, particularly the operation of the p n -junction. In our discussion we will concentrate on the three most commonly used semiconductors: germanium (Ge), silicon (Si), and gallium arsenide (Ga As). Fig 5.1.1 (a) schematically shows the bonding structure of pure silicon: Each silicon atom shares its four valence electrons with the four neighboring atoms, forming four covalent bonds.

In the absence of thermal energy (i.e., when the temperature is equal to zero degree Kelvin [T°K = 0 or F°C = —273.15, where T'° K = 273.15 + T'°C)) all electrons are bonded to the corresponding atoms and the semiconductor is not conductive. However, when the temperature increases, some of the electrons obtain sufficient energy to break up the covalent bond and cross the energy gap W = *W* - *W* , as shown in Figure 5.1.1(b) (at room temperature *T* -- 300°K the band gap energy is equal to 1.12 eV for Si, 0.62 eV for Gee, and 1.42 eV for Gas). These free electrons form negative charge carriers that allow electric current conduction. The concentrator of the conduction electrons in the semiconductor is denoted as n. When an elector breaks the covalent bond it leaves behind a positively charged vacancy, which can be occupied by another free electron. These types of vacancies are called holes and their concentration is denoted by *p.*

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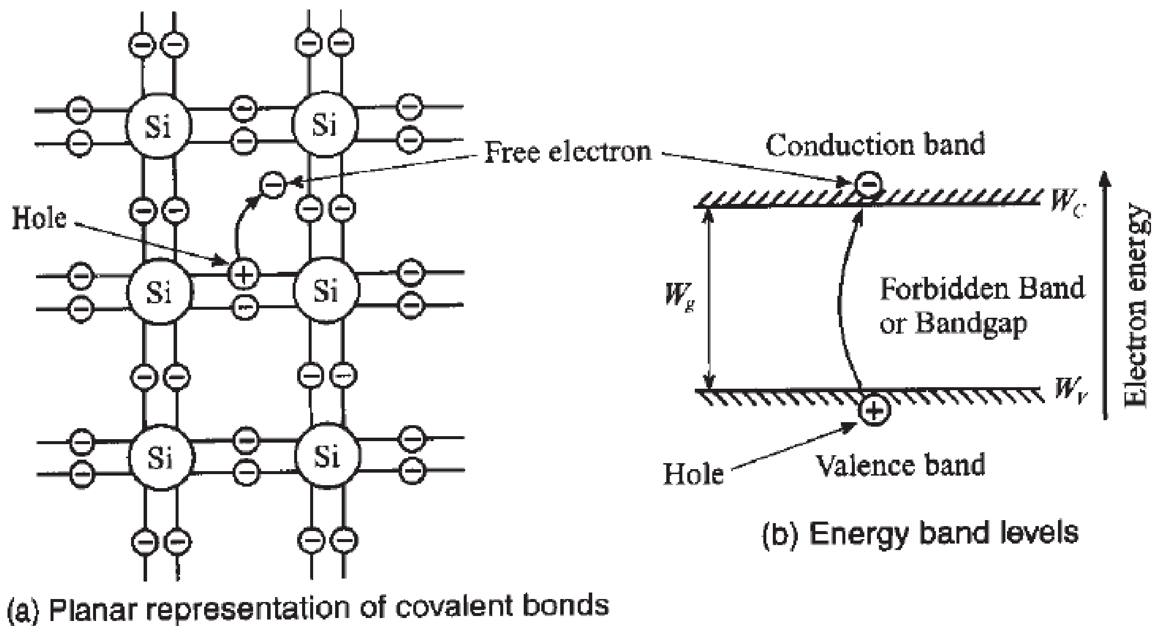

Fig: 5.1.1 Lattice structure and energy levels of silicon

[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory and Applications, Page-273]

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Role, they recombine and both charge carriers disappear. In thermal equilibrium we love equal number of re combinations and generations of holes and electrons. The con- centrations obey the Fermi statistics according to

$$n = N_C \text{ Expel} \left[\frac{W_C - W_F}{kT}\right]$$

$$p = N_V \text{ expel} \left[\frac{W_F - W_V}{kT}\right]$$

$$N_{C,} = 2\,(2\,m^*_{n,p}\,\pi\,k\,T/\,h^2)^{3/2}$$

Are the **effective carrier concentration** in the conduction $N_{C)}$ and valence $Nq)$ bands, respectively. The terms W_C and W_V denote the energy levels associated with the conduction and valence bands and we is the **Fermi energy level,** which indicates

The energy level that has a 50% probability of being occupied by an electron. For

Intrinsic (i.e., pure) semiconductors at room temperature the Fermi level is very close to the middle of the band gap. In (6.2), m_n and *MP* refer to the effective mass of elect- Trans and holes in the semiconductor that are different from the free electron rest mass due to interaction with the crystal lattice; k is Boltzmann's constant; *h* is Planck's con-

Stint; and P is the absolute temperature measured in Kelvin.

In an intrinsic semiconductor the number of free electrons produced by thermal excitation is equal to the number of holes (i.e. n = *p* -- *Ni*). Therefore, electron and hole concentrations are described by the concentration law

$$\text{No} = n_i{}^2$$

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Where a; is the intrinsic concentration. Equation (6.3) is true not only for intrinsic but also for doped semiconductors, which are discussed later in this section.

$$n_i = \sqrt{N_C N_V} \exp\left[-\frac{W_C - W_V}{2kT}\right] = \sqrt{N_C N_V} \exp\left[-\frac{W_g}{2kT}\right]$$

Classical electromagnetic theory specifies the electrical conductivity in a mattered to be *a* = *JZE,* where is the current density and ñ is the applied electric field. The conductivity in the classical model (Drudge model) can be found through the carrier con- centration N, the associated elementary charge q, the drift velocity v_d, and the applied

Electric field ñ:

$$\sigma = \frac{q\ N\ v_d}{E}$$

In semiconductors, we have both electrons and holes contributing to the conductivity of the material. At low electric fields the drift velocity *VD* of the carriers is proportional to the applied field strength through a proportionality constant known as **mobility** q.

$$\sigma = qn\mu_n + qp\mu_p$$

Where q, *qi* are the mobility's of electrons and holes, respectively. For intrinsic semi-conductors we can simplify (6.6) further by recalling that n = *p* - *n,* that is,

$$\sigma = qn_i(\mu_n + \mu_p) = q\sqrt{N_C N_V}\exp\left[-\frac{W_g}{2kT}\right](\mu_n + \mu_p)$$

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It is desired to find the conductivities for the intrinsic materials of Si, Gee, and Gas as a function of temperature. To make the computations not too difficult, we assume that the band gap energy and the mobility's for holes and electrons are temperature independent over the range of interest —50°C ñ *T* ñ 200°C in Fig 5.1.2.

Solution: As a first step it is convenient to combine into one parameter O_0 (F) all factors without the exponential term is given by,

$$\sigma_0(T) \quad q\sqrt{N_C N_V}(\mu_n + \mu_p)$$

Where electron and hole mobility's are found from Table E-1: Jim = 1350(Si), 3900(Gee), 8500(Gas)

Up 480(Si), 1900(Gee), 400(Gas)

$$N_{C,V}(T) = N_{C,V}(300°\text{K})\left(\frac{T}{300}\right)^{3/2}$$

This leads to the form

$$\sigma = \sigma_0(T)\exp\left(-\frac{W_g}{2kT}\right) = q(\mu_n + \mu_p)\sqrt{N_C N_V}\left(\frac{T}{300}\right)^{3/2}\exp\left(-\frac{W_g}{2kT}\right)$$

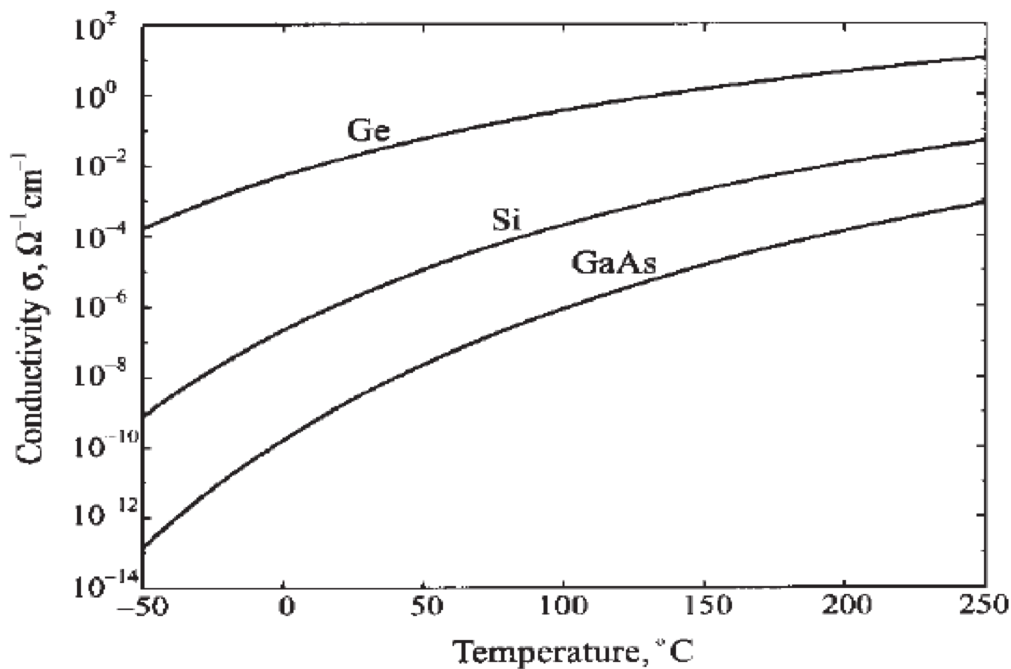

Fig: 5.1.2 Conductivity of Si, Gee, Gas in the range from -50° C to 250° C

[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory And Applications, Page-276]

A major change in the electrical properties of a semiconductor can be initiated by introducing impurity atoms. This process is called doping, as shown in Figure 5.3(a). To achieve o-type doping (which supplies additional electrons to the conduction band) we introduce atoms with a larger number of valence electrons than the atoms in the intrinsic semiconductor lattice that they

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Substitute. For instance, the implantation of phosphorous (P) atoms into Si intro - dukes loosely bound electrons into the neutral crystal lattice, as shown in Figure 5.3(b).

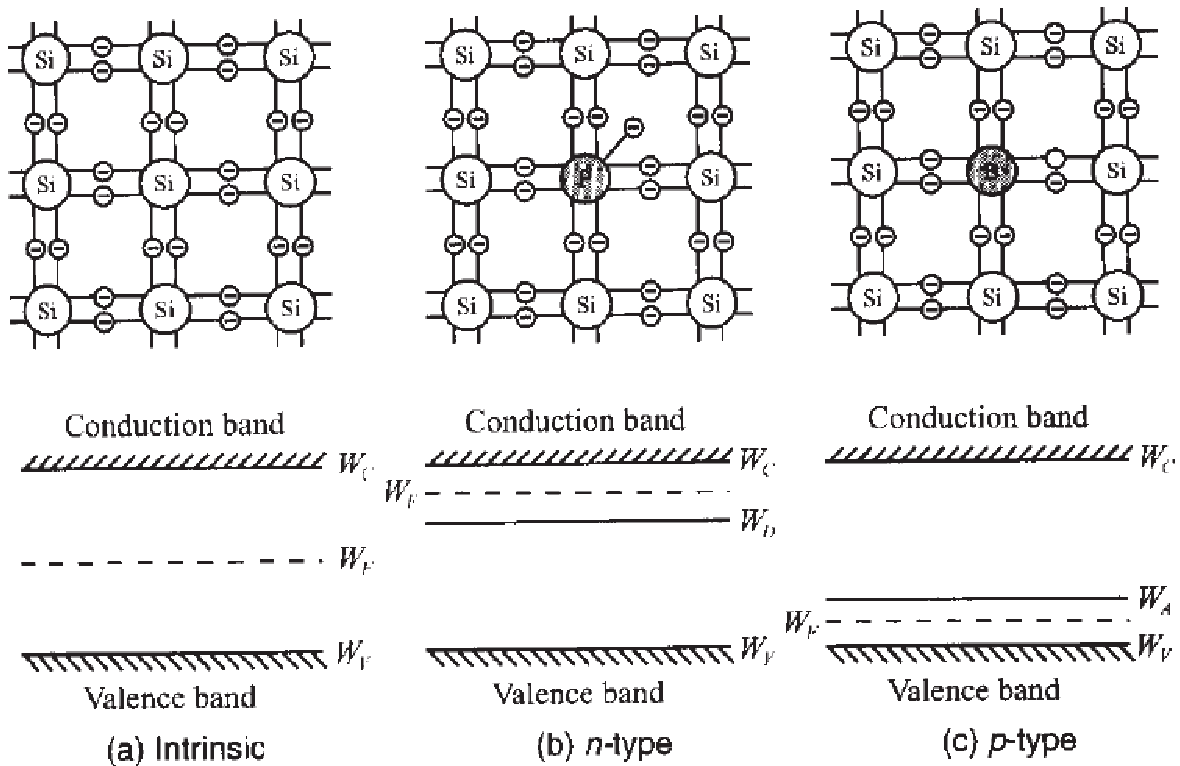

Fig: 5.3 Lattice structure and energy bang model for (a) intrinsic (b) n-type (c) p-type

[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory and Applications, Page-277]

It is intuitively apparent that the energy level of this "extra" electron is closer to the conduction band than the energy of the remaining four valence electrons. When the temperature is increased above absolute zero, the loosely bound electron separates from the atom, forming a free negative charge and leaving behind the fixed positive ion of phosphorous, as shown in Figure 5.3(c). Thus, while still maintaining charge neutrality, the atom

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Has donated an electron to the conduction band without creating a hole in the valence band. This results in an increase in the Fermi level since more electrons are located in the conduction band. Contrary to the intrinsic semiconductor (n_i, p_i) we now have an n-type silicon- doctor in which the electron concentration is related to the whole concentration as

$$No = N_D + p$$

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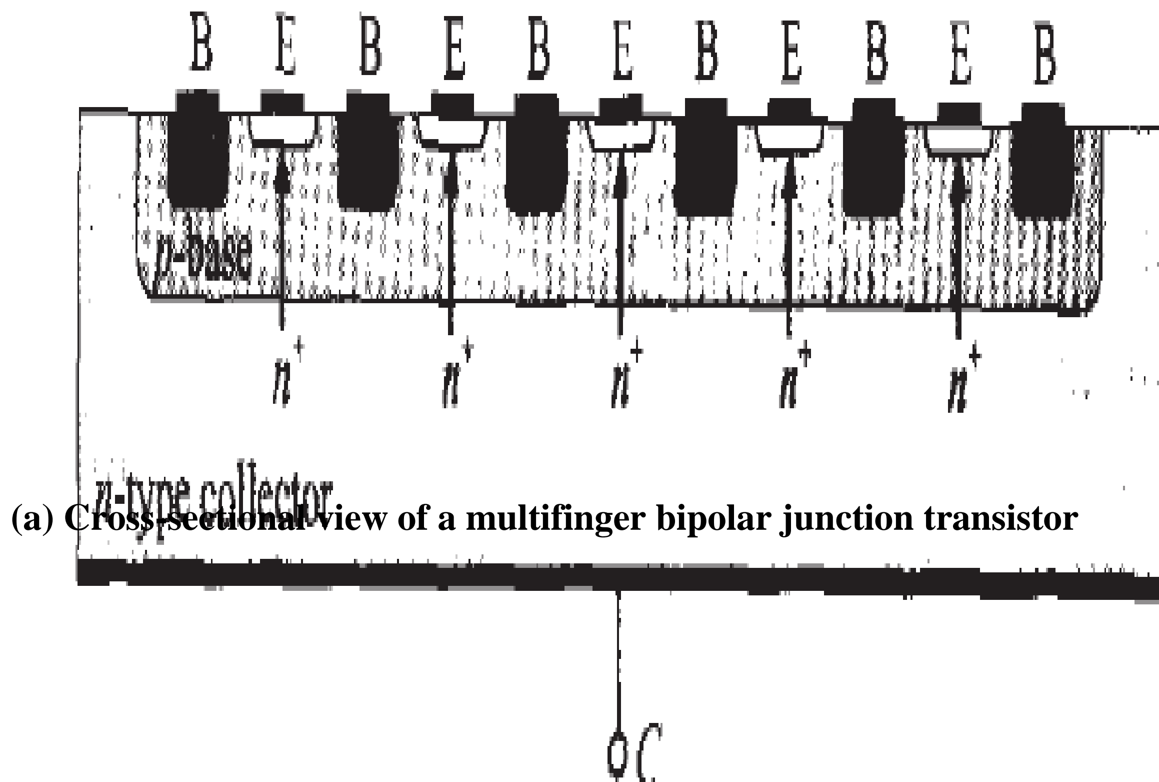
5.6 BIPOLAR JUNCTION TRANSISTORS:

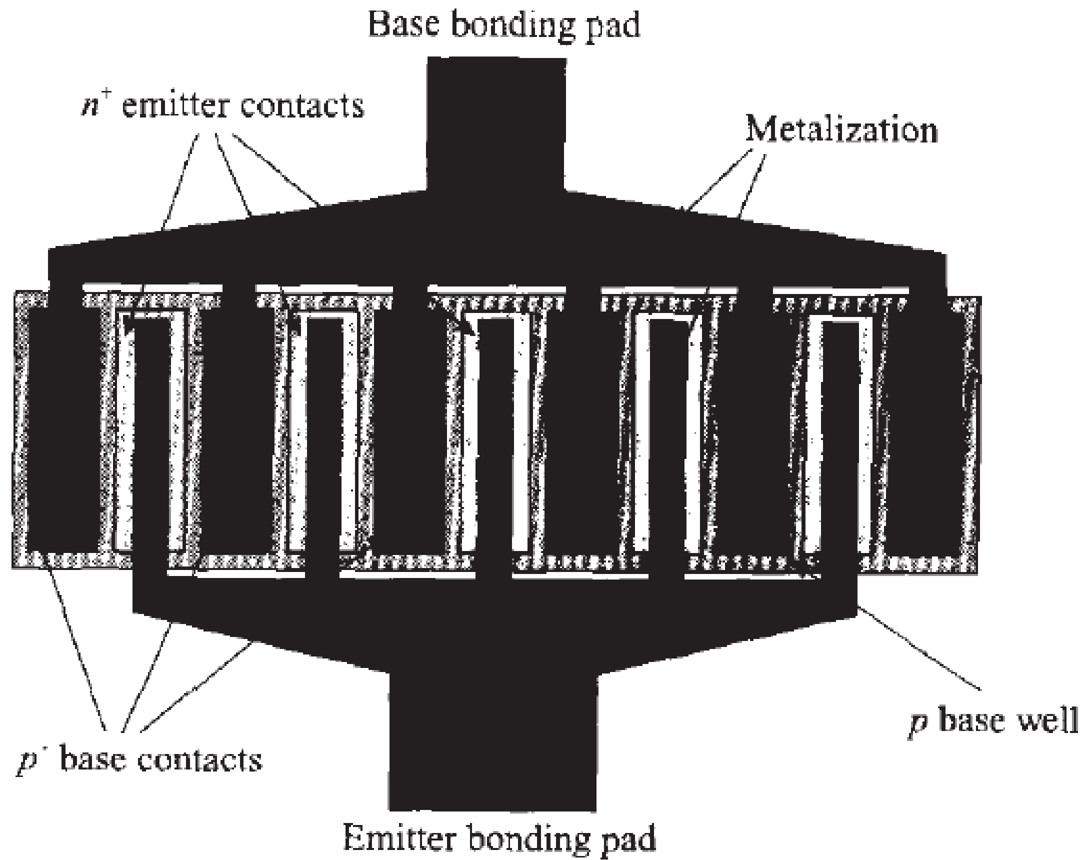
The constituents of a bipolar junction transistor are three alternatively doped semiconductors, PNP in on NPN configuration. As the word bipolar means, the internal current is due to both minority and majority carriers.

The BJT is one of the widely used active RF elements due to its low-cost construction, relatively high operating frequency, low noise performance, and high-power handing capacity. The high-power capacity is achieved through a special inter-digital emitter-base construction as a part of planer structure. Figure 5.6.1(a), shows both the cross-sectional planer

Shown in Figure 5.6.1 (b) the base-emitter resistance is kept at a minimum while not compromising the gain performance. A low base resistance directly improves the signal-to-noise ratio by reducing the current density through the base emitter junction and by reducing the random thermal motion in the base.



(a) Cross-sectional view of a multifinger bipolar junction transistor



(b) Top view of a MultiFinder bipolar junction

Figure 5.6.1 Inter digitized structure of high-frequency BJT

[Source: Reinhold Ludwig and Powel Bretchko, RF Circuit Design – Theory and Applications, Page-313]

The BJT is still the preferred device in very demanding analog circuit applications, both integrated and discrete. This is especially true in very-high-frequency applications, such as radio frequency (RF) circuits for wireless systems. A very-high-speed digital logic-circuit family based on bipolar transistors, namely emitter-coupled logic, is still in use. Finally, bipolar transistors can be combined with MOSFETs to create innovative circuits that take advantage of the high-input-impedance and low-power operation of MOSFETs and the very-high-frequency operation and high-current-driving capability of bipolar transistors. The resulting

Technology is known as Bios or Becomes. Fig 5.6.2 shoes a cross-sectional view of such a structure.

The bipolar transistor enjoyed nearly three decades as the device of choice in the design of both discrete and integrated circuits. Although the MOSFET had been known very early on, it was not until the 1970s and 1980s that it became a serious competitor to the BJT. At the time of this writing (2003), the MOSFET is undoubtedly the most widely used electronic device, and CMOS technology is the technology of choice in the design of integrated circuits. Nevertheless, the BJT remains a significant device that excels in certain applications.

A terminal is connected to each of the three semiconductor regions of a transistor, with the terminals labeled emitter (E), base (B), and collector (C). The transistor consists of two PN junctions, the emitter–base junction (EBJ) and the collector–base junction (CBJ). Depending on the bias condition (forward or reverse) of each of these junctions, different modes of operation of the BJT are obtained, as shown in Table 5.1. The active mode, which is also called forward active mode, is the one used if the transistor is to operate as an amplifier. Switching applications (e.g., logic circuits) utilize both the cutoff and the saturation modes.

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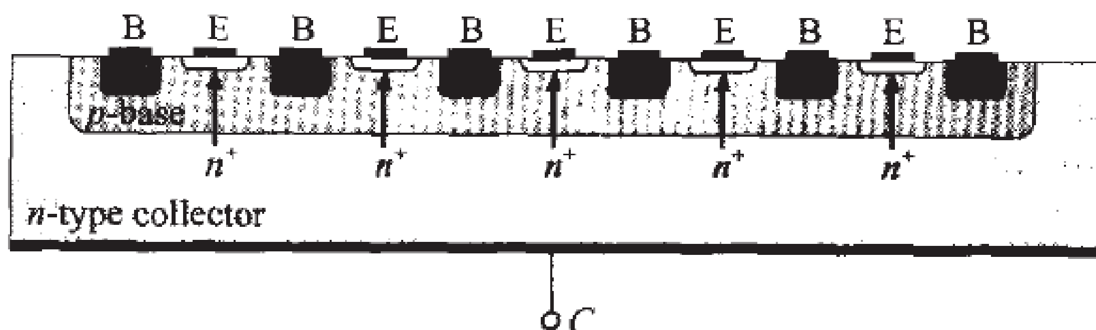

Figure: 5.6.2 Cross-sectional view of a Gas Hero junction bipolar junction involving a Gaius- Gas interface

[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory and Applications, Page-314]

The reverse active (or inverse active) mode has very limited application but is conceptually important. As we will see shortly, charge carriers of both polarities— that is, electrons and holes— participate in the current conduction process in a bipolar transistor, which is the reason for the name bipolar. Fig 5.12 shows the cross-sectional view of such a structure.

The forward bias on the emitter–base junction will cause current to flow across this junction. Current will consist of two components: electrons injected from the emitter into the base, and holes injected from the base into the emitter. As will become apparent shortly, it is highly desirable to have the first component (electrons from emitter to base) at a much higher level than the second component (holes from base to emitter). This can be accomplished by fabricating the device with a heavily doped emitter and a lightly doped base; that is, the device is designed to have a high density of electrons in the emitter and a low density of holes in the base. The BJT is a current controlled device that is best explained by referring to Fig 5.6.3(a) which shows the structure, electrical symbol, and diode model with associated voltage and current convention for the NPN structure.

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[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory and Applications, Page-315]

Let us now consider the electrons injected from the emitter into the base. These electrons will be minority carriers in the p-type base region. Because the base is usually very thin, in the steady state the excess minority carrier (electron) concentration in the base will have an almost straight-line profile, as indicated by

The solid straight line in Fig. 5.6.3(b). The electron concentration will be highest [denoted by NP(0)] at the emitter side and lowest (zero) at the collector side.2 As in the case of any forward-biased pn junction (Section 3.X), the concentration will be

Pro portional to e^{V_{BE}/V_τ} in Fig 5.6.3(c).

$$n_P(0) = n_{P0} e^{V_{BE}/V_\tau}$$

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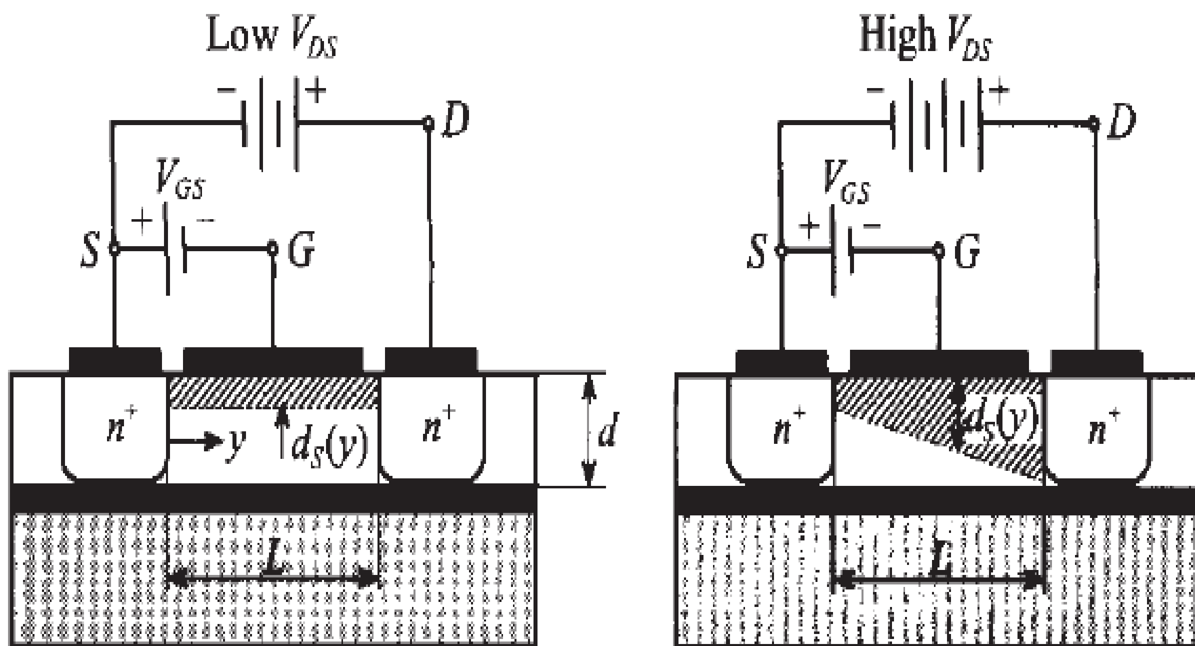

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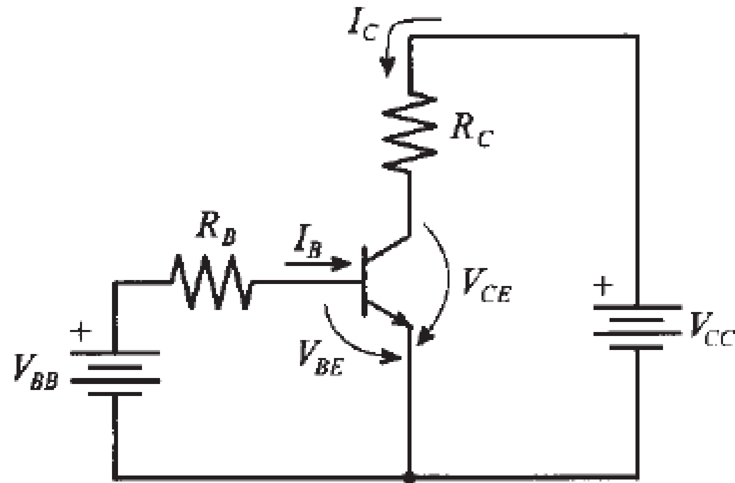
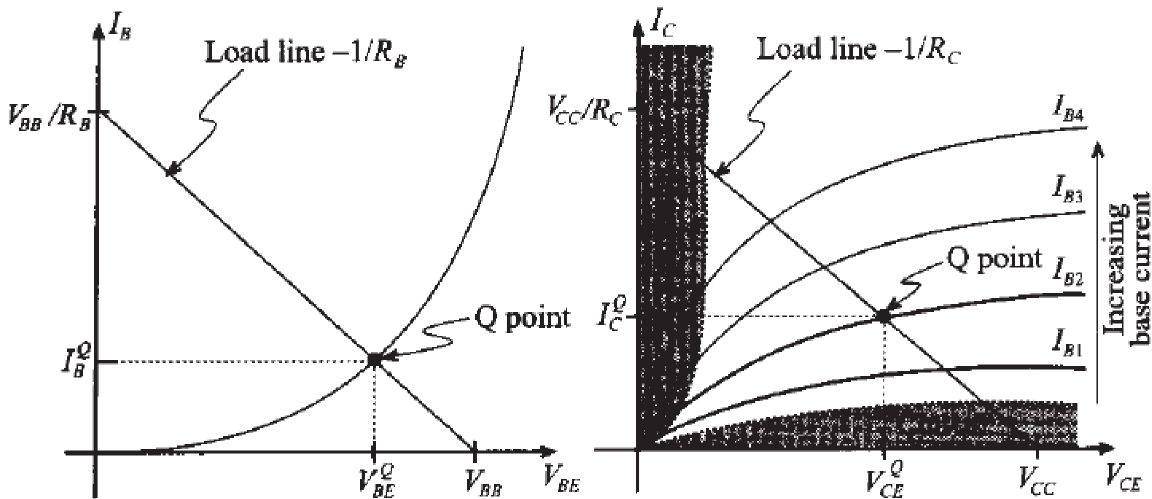


(a) Biasing circuit for npn BJT in common-emitter configuration



(b) Input characteristic of transistor **(c) Output characteristics of transistor**

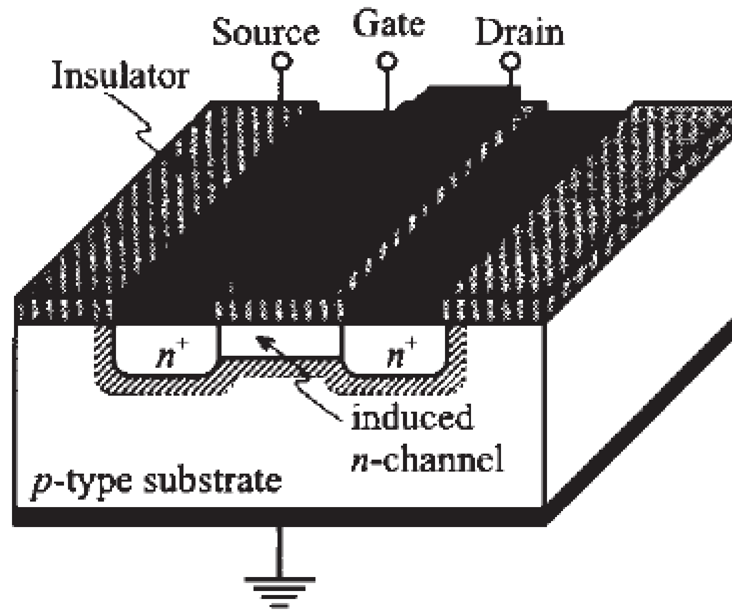
Fig: 5.6.4 Biasing of input, output characteristics of an npn BJT.

[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory and Applications, Page-316]

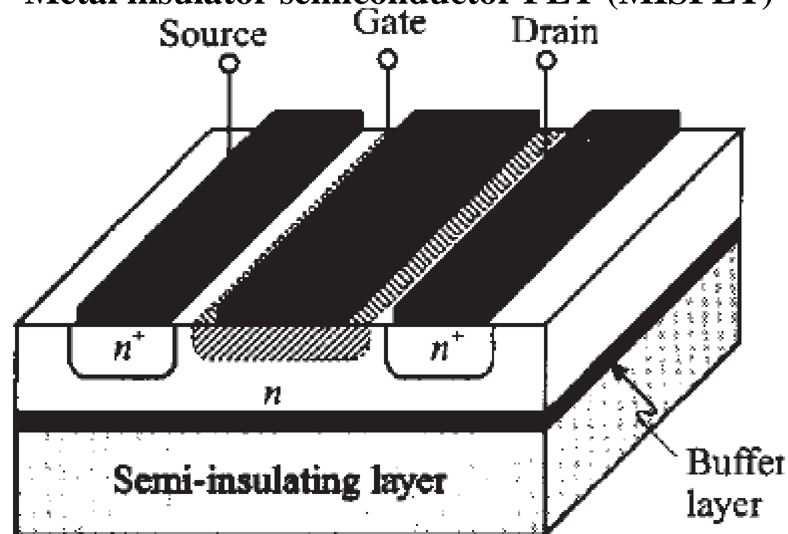
In Fig 5.6.4(a), the saturation current IS inversely proportional to the base width W and is directly proportional to the area of the EBJ. Typically IS in the range of 10−12 A to 10−15 a (depending on the size of the device). Because IS proportional to it is a strong function of temperature, approximately doubling for every 5°C rise in temperature shown in Fig 5.6.4(b). (For the dependence of on temperature) Since IS directly proportional to the junction area (i.e., the device size), it will also be referred to as the scale current. Two transistors that are identical except that one has an EBJ area, say, twice that of the other will have saturation currents with that same ratio (i.e., 2). Thus for the same value of vibe the larger device will have a collector current twice that in the smaller device. This concept is frequently employed in integrated-circuit design shown in Fig 5.6.4 (c).

RF FIELD EFFECT TRANSISTORS:

There are two types of field-effect transistors, the Junction Field-Effect Transistor (JFET) and the "Metal-Oxide Semiconductor" Field-Effect Transistor (MOSFET), or Insulated-Gate Field-Effect Transistor (IGFET). The principles on which these devices operate (current controlled by an electric field) are very similar — the primary difference being in the methods by which the control element is made shown in Fig 5.6.5(a)



(a) Metal insulator semiconductor FET (MISFET)



(b) Junction field effect transistor (JFET)

[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory and Applications, Page-330]

MOSFET CAPACITANCES The physical structure of a MOSFET results in capacitors between the terminals. The metal anode gate structure determines the capacitors from gate–to–drain (Cod), and gate– to–source (Cogs) shown in Fig 5.6.5 (b)

. The PN junction formed during the fabrication of the MOSFET results in a junction capacitance from drain–to–source (Cods). These capacitances are characterized as input (Cass), output (Cuss) and reverse transfer (Cress) capacitances on data sheets. The relationships between the inter–terminal capacitances and those given on data sheets are shown in Fig 5.6.5(c). The Cass can be specified in two ways: 1. Drain shorted to source and positive voltage at the gate. 2. Positive voltage of the drain in respect to source and zero volts at the gate. In the latter case the numbers are lower.

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However, neither method represents the actual operating conditions in RF applications.

DRAIN CHARACTERISTICS One figure of merit for a FET is its static resistance in the full–on condition. This on–resistance, VDS (on), occurs in the linear region of the output characteristic and is specified under specific test conditions for gate–source voltage and drain current. For MOSFETs, VDS (on) has a positive temperature coefficient and constitutes an important design consideration at high temperatures, because it contributes to the power dissipation within the device.

GATE CHARACTERISTICS The gate of the MOSFET is a poly silicon material, and is electrically isolated from the source by a layer of oxide. The input resistance is very high — on the order of 109 ohms — resulting in a leakage current of a few Nano amperes. Gate control is achieved by applying a positive voltage slightly in excess of the gate–to–source threshold voltage, VGS (THz)

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Gate Voltage Rating — never exceed the gate voltage rating. Exceeding the rated VGS can result in permanent damage to the oxide layer in the gate region. Gate Termination — the gate of this device is essentially capacitor. Circuits that leave the gate open–circuited or float in should be avoided. These conditions can result in turn– on of the device due to voltage build–up on the input capacitor due to leakage currents or pickup. Gate Protection — this device does not have an internal monolithic sneer diode from gate–to–source shown in in Fig 5.6.5(b).

If gate protection is required, an external sneer diode is recommended. Using a resistor to keep the gate–to–source impedance low also helps damp transients and serves another important function. Voltage transients on the drain can be coupled to the gate through the parasitic gate–drain capacitance. If the gate–to–source impedance and the rate of voltage change on the drain are both high, then the signal

Coupled to the gate may be large enough to exceed the gate–threshold voltage and turn the device on. HANDLING CONSIDERATIONS when shipping, the devices should be transported only in antistatic bags or conductive foam. Upon removal from the packaging, careful handling procedures should be adhered to. Those handling the devices should wear grounding straps and devices not in the antistatic packaging should be kept in metal tote bins. MOSFETs should be handled by the case and not by the leads, and when testing the device, all leads should make good electrical contact before voltage is applied. As a final note, when placing the FET into the system it is designed for, soldering should be done with a grounded iron. DESIGN CONSIDERATIONS The MRF141G is an RF Power, MOS, N–channel enhancement mode field–effect transistor (FET) designed for HF and VHF power amplifier application Motorola Application Note AN211A, FETs in Theory and Practice, is suggested reading for those not familiar with the construction and characteristics of FETs. The major advantages of RF power MOSFETs include high gain, low noise, simple bias systems, relative immunity from thermal runaway, and the ability to withstand severely mismatched loads without suffering damage. Power output can be varied over a wide range with a low power dc control signal. DC BIAS The MRF141G is an enhancement mode FET and, therefore, does not conduct when drain voltage is applied. Drain current flows when a positive voltage is applied to the gate. RF power FETs require forward bias for optimum performance. The value of quiescent drain current (IDQ) is not critical for many applications. The MRF141G was characterized at IDQ = 250 mA, each side, which is the suggested minimum value of IDQ. For special applications such as linear amplification, IDQ may have to be selected to optimize the critical parameters. The gate is a dc open circuit and draws no current. Therefore, the gate bias circuit may be just a simple resistive divider network. Some applications may require a more elaborate bias system. GAIN CONTROL Power output of the MRF141G may be controlled from its rated value

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Down to zero (negative gain) by varying the dc gate voltage. This feature facilitates the design of manual gain control, AGC/ALC and modulation systems.

FUCTIONALITY:

The input resistance of the MOSFET is exceptionally high because the gate behaves as a capacitor with very low leakage (ran ≈ 1014 Ω). The output impedance is a function of rds. (which is related to the gate voltage) and the drain and source bulk resistances (RD and RS). To turn the MOSFET "on", the gate-channel capacitance, Cg (chi), and the Miller capacitance, Cod, must be charged. In turning "on", the drain- substrate capacitance, Cd (sub), must be discharged. The resistance of the substrate determines the peak discharge current for this capacitance. The FET just described is called an enhancement-type MOSFET. A depletion-type MOSFET can be made in the following manner: Starting with the basic structure of Figure 4, a moderate resistivity n-channel is diffused between the source and drain so that drain current can flow when the gate potential is at zero volts (Figure 5.6. In this manner, the MOSFET can be made to exhibit depletion characteristics. For positive gate voltages, the structure enhances in the same manner as the device of Figure 5.6.5(b). With negative gate voltage, the enhancement process is reversed and the channel begins to deplete of carriers as seen in Figure 5.6. As with the JFET, drain-current.

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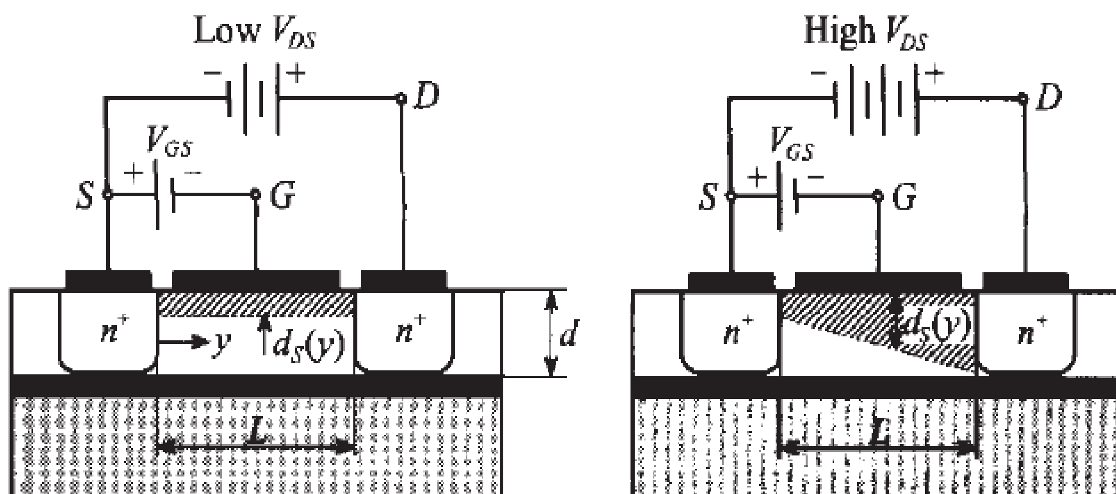

(A) Operation in the linear region **(b) Operation in the saturation region**

Fig: 5.17 Functionality of MSEFET for different drain source voltages

[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory and Applications, Page-331]

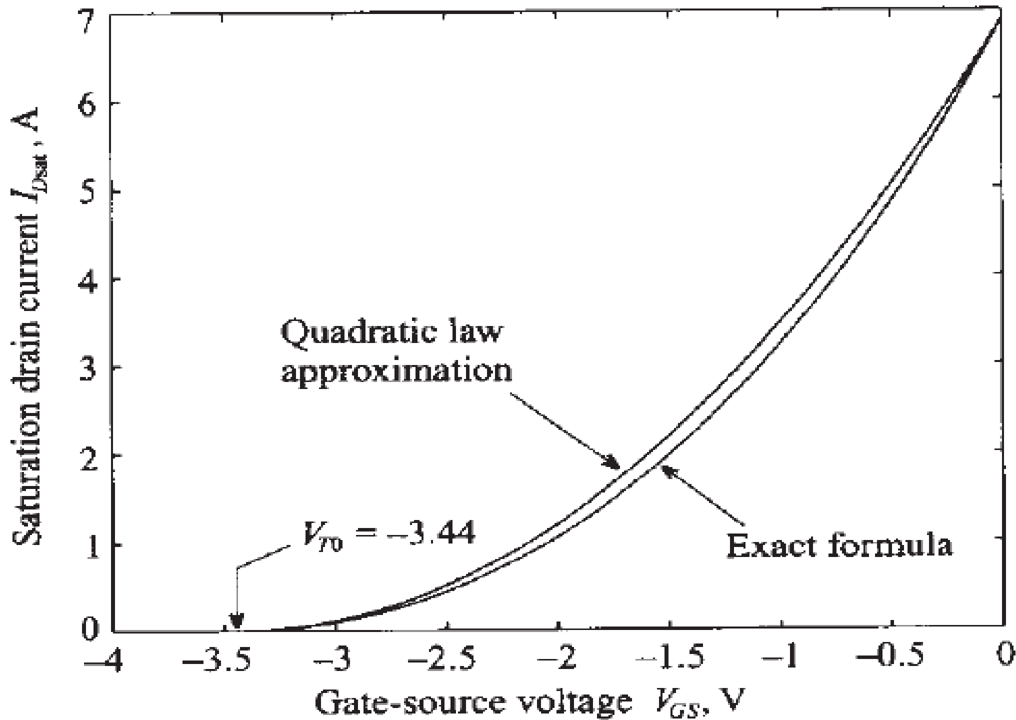

Fig: 5.18 Drain current vs V_{GS}

[Source: Reinhold Ludwig and Powel Bretchko,| RF Circuit Design – Theory and Applications, Page-334]

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5.5 PN JUNCTION:

PN junction diode is a 2 terminal polarity sensitive device. The diode conducts when forward bias is applied and it will introduce zero resistance in the circuit. The diode does not conduct when reverse bias is applied and it will introduce infinite resistance in the circuit, is shown in Fig 5.5.1.

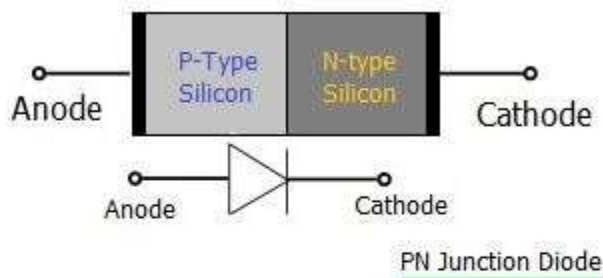

Fig : 5.5.1 PN Diode symbol

[Source: Reinhold Ludwig and Powel Bretchko,‖ RF Circuit Design – Theory and Applications, Page-279]

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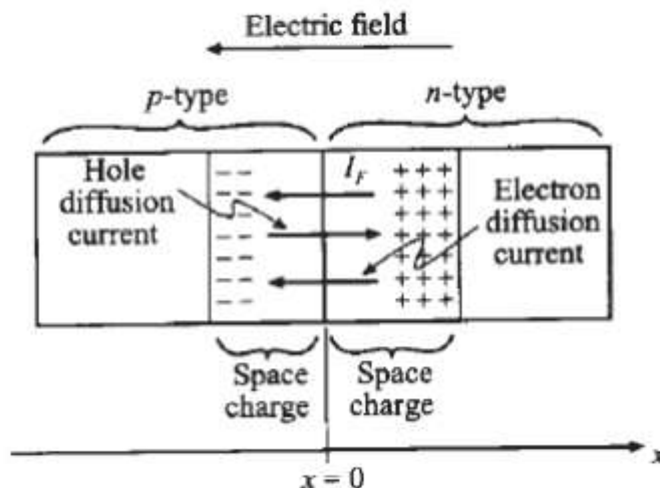
current is commonly known as a diffusion current and is composed of electrons End holes in Fig



It is made up of P-type silicon and N-type silicon semiconductor materials. The

Fig : 5.5.2 Current flow in the PN-junction

[Source: Reinhold Ludwig and Powel Bretchko,‖ RF Circuit Design – Theory and Applications, Page-279]

For applications where high currents and high voltages are used, neither the alloy nor the grown junction diodes were satisfactory because of voltage and current limitations respectively. A new type of diode was invented for high, power applications. This diode is made from a die of very nearly pure intrinsic silicon which has a very high resistivity. A P type and an N type impurity are diffused into the intrinsic material on either side of the die. This makes the PIN or P+ it N+ configuration sketched in Figure 10a. Figure 10b shows the variation of doping level for this diode which is directly proportional to the conductivity as a function of distance through the diode. Thus the N and P regions are high conductivity or low resistivity regions, and the I region is a low conductivity region.

For the purposes of this section, an ideal diode is defined as a diode with a voltage current characteristic as given by equation 4. This equation relates the current through a semiconductor diode to the voltage externally applied to it.

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$$I = I_S[e^{(\frac{qV}{kT})} - 1]$$

where I is the temperature sensitive saturation current given in equation 3, kT is a voltage equivalent of temperature which is 0.026 volt at room q temperature, and V and I are the voltage across the diode and the current through it respectively with polarities and current directions defined in Figure 8. Die symbol V is substituted for VA for simplification. Equation 5 and 6 represent the V-I characteristic for the forward biased and the reverse biased situations. These equations hold for values of applied voltage greater than 0.1 volt at useful temperatures.

$$I = I_S e^{(\frac{qV}{kT})}$$

$$I = -\ I_S$$

The ideal diode presents a purely resistive component of impedance to an a-c signal since equation 4 allows for no frequency effects. For a small signal situation the instantaneous voltage applied to the diode

v = V_0 + v costωt

Where V_O is the d-c bias voltage and v is the peak amplitude of the small a-c signal. For small signal applications v <•< V.

I = $I_S + \frac{dI}{dV}$ costωt

Where I is the instantaneous current, I_0 is equal to I_B $(\frac{qV}{kT})$ and $\frac{dI}{dV}$ a conductance Evaluated at V equal to V . Equation is valid for small a-c voltages only since the higher order terms of the Taylor series expansion are neglected. The current I can be written as

I = I_0 + I costωt

Where I is the peak value of the a-c current, and equation 10 gives the relationship between i and v.

i = $\frac{dI}{dV}$ v = give

The conductance g is a small signal a-c conductance relating the a-c current and voltage. The approach outlined above allows the analysis of ideal semiconductor diodes to be performed in two distinct steps. First the d-c bias conditions are analyzed followed by the computation of the a-c operating conditions using the small signal conductance g.

SCHOTTKY DIODE:

The diode is constructed on a thin silicon (n+ type) substrate by growing epitaxial on n-type active layer of about 2 micron thickness. A thin SiO2 layer is grown thermally over this active layer. Metal semiconductor junction is formed by depositing metal over SiO2 in Fig 5.5.3.

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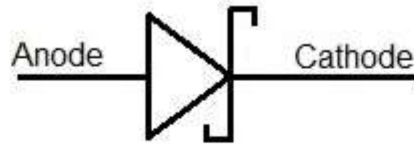

Fig: 5.5.3 Scotty diode symbol

[Source: Reinhold Ludwig and Powel Bretchko, ‖ RF Circuit Design – Theory and Applications, Page-289]

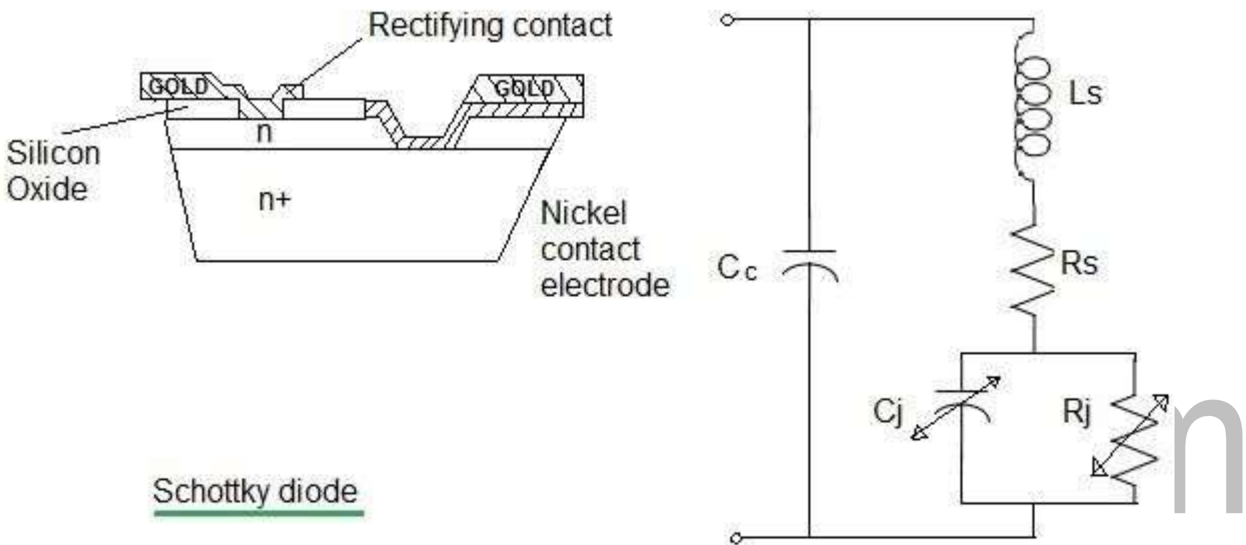

Fig: 5. 5.4 Scotty diode contact

[Source: Reinhold Ludwig and Powel Bretchko,‖ RF Circuit Design – Theory and Applications, Page-290]

When the Scotty diode is forward biased, the major carriers (i.e. electrons) can be easily injected from highly doped n-semiconductor material into the metal.

When it is reverse biased, the barrier height becomes too high for the electrons to cross and no conduction will happen. These applications are very useful.
In Fig 5.5.41, RF power flow in the device is limited by power dissipation in Rs. It is shorted across Cu, Cc and Less. These will produce RF mismatch and can be matched by external circuit.

5.2 RF AMPLIFIERS:

CHARACTERISTICS OF AMPLIFIERS:

The most important and complex task is the amplification of an input signal through either a single or multistage transistor circuit. A generic single-stage amplifier configuration embedded between input and output matching networks, as shown in Fig 5.2.1.

Input and Output matching networks, are needed to reduce undesired reflections and thus improve the power flow capabilities. Inters of performance specifications, the following list constitutes a set of key amplifier parameters.

- Gain and gain flatness
- Operating frequency and bandwidth
- Output power
- Power supply requirements
- Input and output power reflections
- Noise figure Γ_S

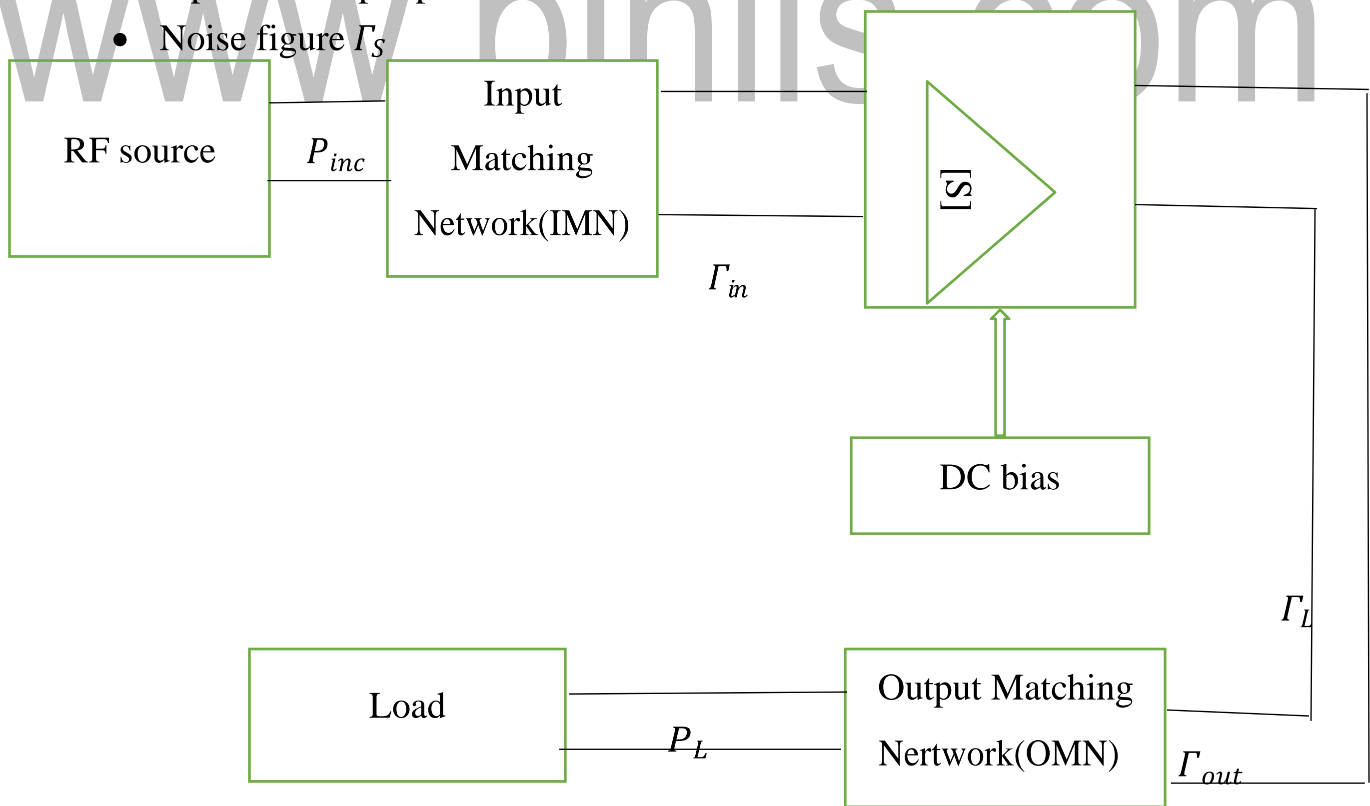

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Fig: 5.2.1 Generic amplifier system

i. Reflection Coefficient seen looking toward the load

$$\Gamma_L = \frac{Z_L - Z_O}{Z_L + Z_O}$$

ii. Reflection Coefficient seen looking toward the source

$$\Gamma_S = \frac{Z_S - Z_O}{Z_S + Z_O}$$

iii. Input reflection Coefficient

$$\Gamma_{in} = S_{11} + \frac{S_{12} S_{21} \Gamma_L}{1 - S_{22} \Gamma_L}$$

iv. Output reflection Coefficient

$$\Gamma_{out} = S_{22} + \frac{S_{12} S_{21} \Gamma_S}{1 - S_{11} \Gamma_s}$$

From signal flow graph in figure,

$$b_s + a_1{}' \Gamma_S = b_1{}'$$

$$b_s = b_1{}' - a_1{}' \Gamma_S$$

$$b_s = b_1{}' [1 - \frac{a_1{}'}{b_1{}'} \Gamma_S]$$

$$b_s = b_1{}' [1 - \frac{b_1}{a_1} \Gamma_S]$$

$$b_s = b_1{}' [1 - \Gamma_{in} \Gamma_S] \qquad \text{......... (1)}$$

$$b_1{}' = \frac{b_s}{1 - \Gamma_{in} \Gamma_S} \qquad \text{...... (2)}$$

Incident power wave (P_{inc}) associated with $b_1{}'$ is given as,

$$P_{inc} = \frac{|b_1{}'|^2}{2}$$

$$P_{inc} = \frac{1}{2} \frac{|b_s|^2}{|1 - \Gamma_{in} \Gamma_S|^2} \qquad \text{...... (3)}$$

Which is the power launched towards the amplifier,

$$\Gamma_{in} = 0 \ \& \ \Gamma_S \neq 0$$

$$P_{inc} = \frac{|b_s|^2}{2} \qquad \text{....... (4)}$$

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The actual input power P_{in} is observed at the input terminal of the amplifier is composed of the incident & reflected power waves.

$$P_{in} = P_{inc}\,(1 - |\Gamma_{in}|^2)$$

$$P_{in} = \frac{1}{2}\frac{|b_s|^2}{|1 - \Gamma_{in}\Gamma_S|^2}(1 - |\Gamma_{in}|^2) \qquad \ldots..(5)$$

Maximum power transfer from the source to the amplifier is achieved if the input Impedance is complex conjugate matched ($Z_{in} = Z_{v)}$ or, in terms of the reflection coefficients if $\Gamma_{in} = \Gamma_S^*$

Under maximum power transfer condition, the available power P_A as,

$$P_A = P_{in}|_{\Gamma_{in} = \Gamma_S^*}$$

$$P_A = \frac{1}{2}\frac{|b_s|^2}{|1 - \Gamma_{in}\Gamma_S|^2}(1 - |\Gamma_{in}|^2)\Big|_{\Gamma_{in} = \Gamma_S^*}$$

$$P_A = \frac{1}{2}\frac{|b_s|^2}{|1 - |\Gamma_S|^2|^2}(1 - |\Gamma_s|^2)$$

$$P_A = \frac{1}{2}\frac{|b_s|^2}{(1 - |\Gamma_s|^2)} \qquad \ldots\ldots (6)$$

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Power delivered to the load is given by,

$$P_L = \frac{1}{2}|b_2|^2(1 - |\Gamma_L|^2) \qquad \ldots\ldots (7)$$

TRANDUCER POWER GAIN

$$G_T = \frac{power\ delivered\ to\ the\ load}{available\ power\ from\ the\ source} \qquad \ldots..\ (1)$$

$$G_T = \frac{P_L}{P_{av(s)}} = \frac{P_L}{P_A} \qquad \ldots\ldots \quad (2)$$

$$P_L = \frac{1}{2}|b_2|^2(1\text{-}|\Gamma_L|^2) \qquad \ldots\ldots (3)$$

$$P_A = \frac{|b_s|^2}{(1 - |\Gamma_s|^2)} \qquad \ldots..\ (4)$$

Sub equal (3) and (4) in equal (2),

$$G_T = \frac{\frac{1}{2}|b_2|^2(1 - |\Gamma_L|^2)}{\frac{|b_s|^2}{(1 - |\Gamma_s|^2)}}$$

(

$$G_T = \left|\frac{b_Z}{b_S}\right| (1 - |\Gamma_L|)^2 (1 - |\Gamma_S|)............(5)$$

The reflected wave at port 2 is b_2,

$$b_2 = \frac{a_1 S_{21}}{1 - S_{22}\Gamma_L} \quad .... (6)$$

By rearranging the above equation (6),

$$b_2 (1 - S_{22}\Gamma_L) = a_1 S_{21}$$

$$a_1 = \frac{b_2 (1 - S_{22}\Gamma)}{S_{21}} \quad .... (7)$$

The source reflection coefficient b_s is given by,

$$b_s = (1 - \Gamma_S \Gamma_{in})\, a_1 \quad ...... (8)$$

The input reflection coefficient Γ_{in} is given by,

$$\Gamma_{in} = S_{11} + \frac{S_{12} S_{21} \Gamma_L}{1 - S_{22}\Gamma_L} \quad .... (9)$$

By rearranging the equal (11) by,

$$\Gamma_{in} = \frac{S_{11}(1 - S_{22}\Gamma_L) + S_{12} S_{21} \Gamma_L}{1 - S_{22}\Gamma_L}$$

To find 1- $\Gamma_S \Gamma_{in}$:

$$\Gamma_S \Gamma_{in} = \frac{S_{11}(1 - S_{22}\Gamma_L)_S + S_{12} S_{21} \Gamma_L \Gamma_S}{1 - S_{22}\Gamma_L}$$

$$1 - \Gamma_S \Gamma_{in} = 1 - \frac{S_{11}(1 - S_{22}\Gamma_L)_S + S_{12} S_{21} \Gamma_L \Gamma_S}{1 - S_{22}\Gamma_L}$$

$$= 1 - \frac{S_{11}\Gamma_S - S_{11} S_{22}\Gamma_L \Gamma_S + S_{12} S_{21} \Gamma_L \Gamma_S}{1 - S_{22}\Gamma_L}$$

$$= \frac{(1 - S_{22}\Gamma_L) - S_{11}\Gamma_S - S_{11} S_{22}\Gamma_L \Gamma_S - S_{12} S_{21} \Gamma_L \Gamma_S}{1 - S_{22}\Gamma_L}$$

$$= \frac{(1 - S_{22}\Gamma_L) - S_{11}\Gamma_S(1 - S_{22}\Gamma_L) - S_{12} S_{21} \Gamma_L \Gamma_S}{1 - S_{22}\Gamma_L}$$

$$= \frac{(1 - S_{22}\Gamma_L)(1 - S_{11}\Gamma_S) - S_{12} S_{21} \Gamma_L \Gamma_S}{1 - S_{22}\Gamma_L} \quad .... (12)$$

To find b_s:

From equ (8),

$$b_s = (1 - \Gamma_S \Gamma_{in})\, a_1$$

$$b_s = \left[\frac{(1- S_{22}\,\Gamma_L)(1- S_{11}\,\Gamma_S)- S_{12}\,S_{21}\,\Gamma_L\,\Gamma_S}{1- S_{22}\,\Gamma_L}\right] a_1 \quad .....(13)$$

Sub equ (7) in equ (13),

$$b_s = \left[\frac{(1- S_{22}\,\Gamma_L)(1- S_{11}\,\Gamma_S)- S_{12}\,S_{21}\,\Gamma_L\,\Gamma_S}{1- S_{22}\,\Gamma_L}\right] \text{x}\ \frac{b_2(1-S_{22}\Gamma_L)}{S_{21}}$$

$$\frac{b_s}{b_2} = \frac{(1- S_{22}\,\Gamma_L)(1- S_{11}\,\Gamma_S)- S_{12}\,S_{21}\,\Gamma_L\,\Gamma_S}{S_{21}} \quad ..... (14)$$

To find $\left|\frac{b_s}{b_2}\right|^2$:

$$\left|\frac{b_s}{b_2}\right|^2 = \frac{|S_{21}|^2}{|(1- S_{22}\,\Gamma_L)(1- S_{11}\,\Gamma_S)- S_{12}\,S_{21}\,\Gamma_L\,\Gamma_S|^2} \quad .....(15)$$

To find G_T,

from equal (5),

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$$G_T = \left|\frac{b_2}{b_s}\right|^2 (1- |\Gamma_L|^2)\,(1 - |\Gamma_s|^2)$$

$$G_T = \frac{|S_{21}|^2\,(1- |\Gamma_L|^2)\,(1- |\Gamma_s|^2)}{|(1- S_{22}\,\Gamma_L)(1- S_{11}\,\Gamma_S)- S_{12}\,S_{21}\,\Gamma_L\,\Gamma_S|^2}$$

$$G_T = \frac{|S_{21}|^2\,(1- |\Gamma_L|^2)\,(1- |\Gamma_s|^2)}{|1- S_{22}\,\Gamma_L|^2|1-\Gamma_S\Gamma_{in}|^2}$$

Problem 1:

An amplifier has the following S-parameters: S_{11}=0.3∟-70, S_{12}=0.2∟-10, S_{21}=3.5∟85 and S_{22}=0.4∟-45. Furthermore, input side of the amplifier is connected to a voltage source with V_S = 5V∟0 and the source impedance Z_S = 40ohm. The output is utilized to drive the antenna which has an impedance of Z_L = 73ohm. Assuming that the S-parameters of the amplifier are measured with reference to a Z_O = 50ohm characteristic impedance, Calculate

a) Transducer gain, unilateral transducer gain, available gain, operating power gain.

b) Power delivered to the load, available power and incident power to the amplifier.

Given data:

$S_{11}=0.3\angle -70$

$S_{12}=0.2\angle -10$

$S_{21}=3.5\angle 85$

$S_{22}=0.4\angle -45$

$Z_S = 40ohm$

$Z_L = 73ohm$

$Z_O = 50ohm$

$V_S = 5V\angle 0$

$$\Gamma_L = \frac{Z_L - Z_O}{Z_L + Z_O}$$

$$\Gamma_L = \frac{73-50}{73+50}$$

$\boldsymbol{\Gamma_L = 0.187}$

$$\Gamma_S = \frac{Z_S - Z_O}{Z_S + Z_O}$$

$$\Gamma_S = \frac{40-50}{40+50}$$

$\boldsymbol{\Gamma_S = -0.111}$

$$b_s = \frac{\sqrt{Z_O}}{Z_S + Z_O} V_S$$

$$b_s = \frac{\sqrt{50}}{40+50} 5$$

$\boldsymbol{b_s = 0.392}$

$$\Gamma_{in} = S_{11} + \frac{S_{12} S_{21} \Gamma_L}{1 - S_{22} \Gamma_L}$$

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$\Gamma_{in} = 0.3\angle{-70} + \frac{(0.2\angle{-10})(3.5\angle{85})(0.187)}{1-(0.4\angle{-45})(0.187)}$

$\Gamma_{in} = 0.3\angle{-70} + \frac{(0.7\angle{75})(0.187)}{1-(0.0748\angle{-45})}$

$\Gamma_{in} = 0.3\angle{-70} + \frac{(0.1309\angle{75})}{1-(0.0748\angle{-45})}$

$\Gamma_{in} = 0.3\angle{-70} + 0.138\angle{71.81}$

$\boldsymbol{\Gamma_{in} = 0.209\angle{-45.98}}$

$\boldsymbol{\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_s}}$

$\Gamma_{out} = 0.4\angle{-45} + \frac{(0.2\angle{-10})(3.5\angle{85})(-0.111)}{1-(0.3\angle{-70})(-0.111)}$

$\Gamma_{out} = 0.4\angle{-45} + \frac{(0.7\angle{75})(-0.111)}{1-0.333\angle{-70}}$

$\Gamma_{out} = 0.4\angle{-45} + \frac{(0.0777\angle{75})}{1-0.333\angle{-70}}$

$\Gamma_{out} = 0.4\angle{-45} + 0.0767\angle{76.77}$

$\boldsymbol{\Gamma_{out} = 0.445\angle{-53.42}}$

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a) Transducer gain, unilateral transducer gain, available gain, operating power gain.

$\boldsymbol{G_T = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_s|^2)}{|1-S_{11}\Gamma_S|^2|1-\Gamma_{out}\Gamma_L|^2}}$

$G_T = \frac{(3.5)^2(1-(0.187)^2)(1-(-0.111)^2)}{|1-(0.187)(0.445\angle{-53.42})|^2|1-(0.3\angle{-70})(-0.111)|^2}$

$\boldsymbol{G_T = 12.556}$

$\boldsymbol{G(\text{dB}) = 10\log_{10}(G_T)}$

$G(\text{dB}) = 10\log_{10}(12.556)$

$\boldsymbol{G(\text{dB}) = 10.988\text{ dB}}$

$\boldsymbol{G_{TU} = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_s|^2)}{|1-S_{22}\Gamma_L|^2|1-S_{11}\Gamma_S|^2}}$

$\boldsymbol{G_{TU} = 12.68}$

$G_T(\text{dB}) = 10\ log_{10}\,(G_{TU})$

$G_T(\text{dB}) = 10\ log_{10}\,(12.68)$

$\boldsymbol{G_T(\text{dB}) = 11.03\ \text{Db.}}$

$$G_A = \frac{|S_{21}|^2\,(1-|\Gamma_s|^2)}{|1-S_{11}\,\Gamma_S|^2\,(1-|\Gamma_{out}|^2)}$$

$\boldsymbol{G_A = 14.74}$

$\boldsymbol{G(\text{dB}) = 10\ log_{10}\,(G_A)}$

$G(\text{dB}) = 10\ log_{10}\,(14.74)$

$\boldsymbol{G(\text{dB}) = 11.68\ \text{dB}}$

$$G = \frac{|S_{21}|^2\,(1-|\Gamma_L|^2)}{(1-|\Gamma_{in}|^2)|1-S_{22}\,\Gamma_L|^2}$$

G = 13.75

$\boldsymbol{G\ (\text{dB}) = 10\ log_{10}\,(G)}$

$G\ (\text{dB}) = 10\ log_{10}\,(13.75)$

G (dB) = 11.38 dB

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b) Power delivered to the load, available power and incident power to the amplifier.

$$P_{inc} = \frac{1}{2}\,\frac{|b_s|^2}{|1-\Gamma_{in}\,\Gamma_S|^2}$$

$\boldsymbol{P_{inc} = 0.0733\text{W}}$

$P_A\ (\text{dBm}) = 10\ log_{10}\left(\frac{P_A\ (\text{watts})}{1mW}\right)$

$P_A\ (\text{dBm}) = 18.909\ \text{dBm}$

$G_T = \frac{P_L}{P_A}$

$P_L = P_A\ G_T$

$P_L = 0.976\ \text{W}$

$P_L\ (\text{dBm}) = 10\ log_{10}\left(\frac{P_L\ (\text{watts})}{1mW}\right)$

$P_L\ (\text{dBm}) = 29.89\ \text{dBm}$

$$P_A \text{ (dBm)} = 10 \, log_{10} \left(\frac{P_A \text{ (watts)}}{1mW}\right)$$

$$P_A \text{ (dBm)} = 18.909 \text{ dBm}$$

$$G_T = \frac{P_L}{P_A}$$

$$P_L = P_A \, G_T$$

$$P_L = 0.976 \text{ W}$$

$$P_L \text{ (dBm)} = 10 \, log_{10} \left(\frac{P_L \text{ (watts)}}{1mW}\right)$$

$$P_L \text{ (dBm)} = 29.89 \text{ dBm}$$

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5.3 STABILITY CONSIDERATIONS:

STABILITY:

Is the ability of an amplifier to maintain effectiveness in its nominal operating characteristics such as physical temperature, signal frequency, source & load coefficient etc.

STABILITY REQUIREMENT:

Stability then implies that the magnitudes of the reflection coefficients are less than unity.

$|\Gamma_L| < 1, |\Gamma_S| < 1$

$|\Gamma_{in}| = S_{11} + |\frac{S_{12} S_{21} \Gamma_L}{1 - S_{22} \Gamma_L}| < 1$

$|\Gamma_{out}| = S_{22} + |\frac{S_{12} S_{21} \Gamma_S}{1 - S_{11} \Gamma_s}| < 1$

TYPES OF STABILITY:

i) UNCONDITIONAL STABILITY:

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A network is said to be "unconditionally stable" in a frequency range if:

$|\Gamma_{in}| < 1$ & $|\Gamma_{out}| < 1$

$|\Gamma_L| < 1$ & $|\Gamma_S| < 1$

ii) CONDITIONAL STABILITY:

A network is said to be "conditionally stable" or "potentially unstable" in a frequency range if:

$|\Gamma_{in}| < 1$ & $|\Gamma_{out}| < 1$

STABILITY CIRCLES:

i. Input stability circle
ii. Output stability circle

OUTPUT STABILITY CIRCLE:

In Fig 5.31, for output stability circle $|\Gamma_{in}| = 1$

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$|\Gamma_{in}| = |S_{11} + \frac{S_{12} S_{21} \Gamma_L}{1 - S_{22}\Gamma_L}| = 1$

$|S_{11}(1 - S_{22}\Gamma_L) + S_{12}S_{21}\Gamma_L| = |1 - S_{22}\Gamma_L|$

$|S_{11} - S_{11}S_{22}\Gamma_L + S_{12}S_{21}\Gamma_L| = |1 - S_{22}\Gamma_L|$

$|S_{11} - (S_{11}S_{22} - S_{12}S_{21})\Gamma_L| = |1 - S_{22}\Gamma_L|$

$$S_{11}S_{22} - S_{12}S_{21} = \Delta$$

$|S_{11} - \Delta\Gamma_L| = \quad |1 - S_{22}\Gamma_L|$

$|S_{11}| - \Delta|\Gamma_L| = \quad 1 - |S_{22}\Gamma_L|$

$|S_{22}\Gamma_L| - \Delta|\Gamma_L| = 1 - |S_{11}|$

$(S_{22} - \Delta)\,|\Gamma_L| = 1 - |S_{11}|$

Multiply $(S_{22}{}^* + \Delta^*)$ both sides

$|\Gamma_L|\,\{(S_{22} - \Delta)\}\,\{(S_{22}{}^* + \Delta^*)\} = \{1 - |S_{11}|\}\,\{(S_{22}{}^* + \Delta^*)\}$

$|\Gamma_L|\,\{|S_{22}|^2 - |\Delta|^2\} = |S_{22}{}^*| + |\Delta^*| - |S_{11}S_{22}{}^*| - |S_{11}\Delta^*|$

$|\Gamma_L|\{|S_{22}|^2 - |\Delta|^2\} = |S_{22}{}^*| + |S_{11}{}^*S_{22}{}^*| - |S_{12}{}^*S_{21}{}^*| - |S_{11}S_{22}{}^*| - |S_{11}\Delta^*|$

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$|\Gamma_L| = \frac{|S_{22}{}^*| + |S_{11}{}^*S_{22}{}^*| - |S_{12}{}^*S_{21}{}^*| - |S_{11}S_{22}{}^*| - |S_{11}\Delta^*|}{|S_{22}|^2 - |\Delta|^2}$

$\frac{|S_{22}{}^*| - |S_{12}{}^*S_{21}{}^*| - |S_{11}\Delta^*|}{|S_{22}|^2 - |\Delta|^2}$

$- \quad |\frac{|S_{22}{}^* - S_{11}\Delta^*|}{|S_{22}|^2 - |\Delta|^2}| = |\frac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2}|$

The equation for the output stability circle ($|\Gamma_{in}| = 1$) drawn in the Γ_L plane cab be written as,

$|\Gamma_L - C_L| = R_L$

$C_L = C_{out} = \frac{(S_{22} - S_{11}{}^*\Delta)^*}{|S_{22}|^2 - |\Delta|^2}$

$R_L = r_{out} = |\frac{S_{12}\,S_{21}}{|S_{22}|^2 - |\Delta|^2}|$

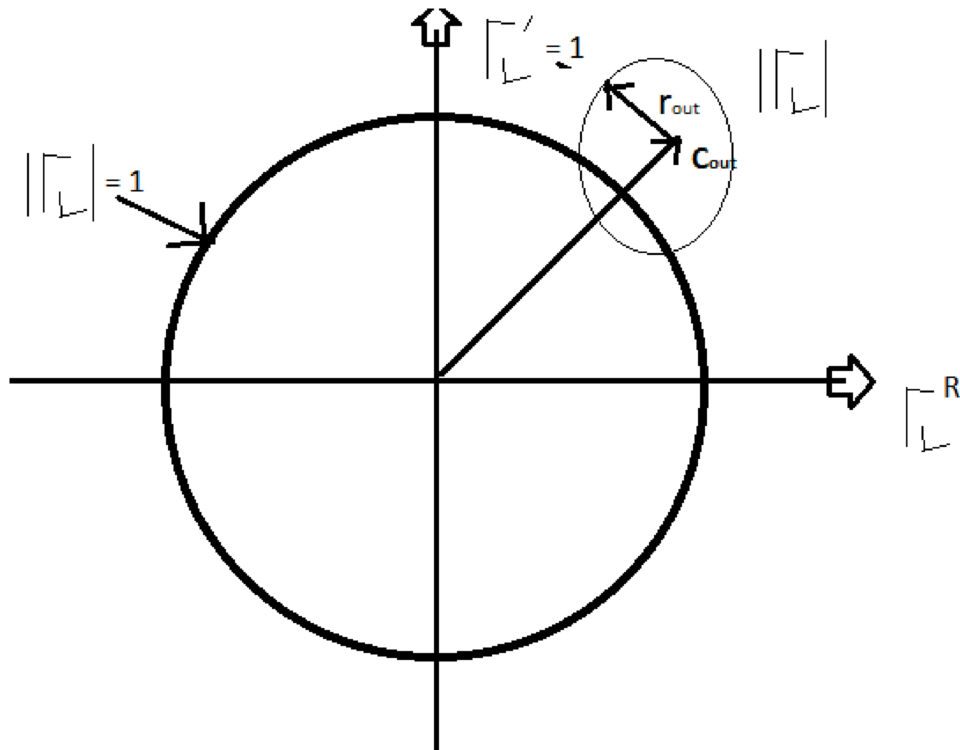

Fig: 5.3.1 Output stability circle

INPUT STABILITY CIRCLE:

In Fig 5.3.2, the input stability circle ($|\Gamma_{out}|$ =1) drawn in the Γ_S plane cab be written as,

$$|\Gamma_S - C_S| = R_S$$

$$C_S = C_{in} = \frac{(S_{11} - S_{11}{}^{*}\Delta)^{*}}{|S_{11}|^2 - |\Delta|^2}$$

$$R_S = r_{in} = \left|\frac{S_{12}\, S_{21}}{|S_{11}|^2 - |\Delta|^2}\right|$$

Special case: unconditional stability:

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2}{2|S_{12}||S_{21}|} > 1$$

$$|\Delta| = |S_{11}\, S_{22} - S_{12}\, S_{21}| < 1$$

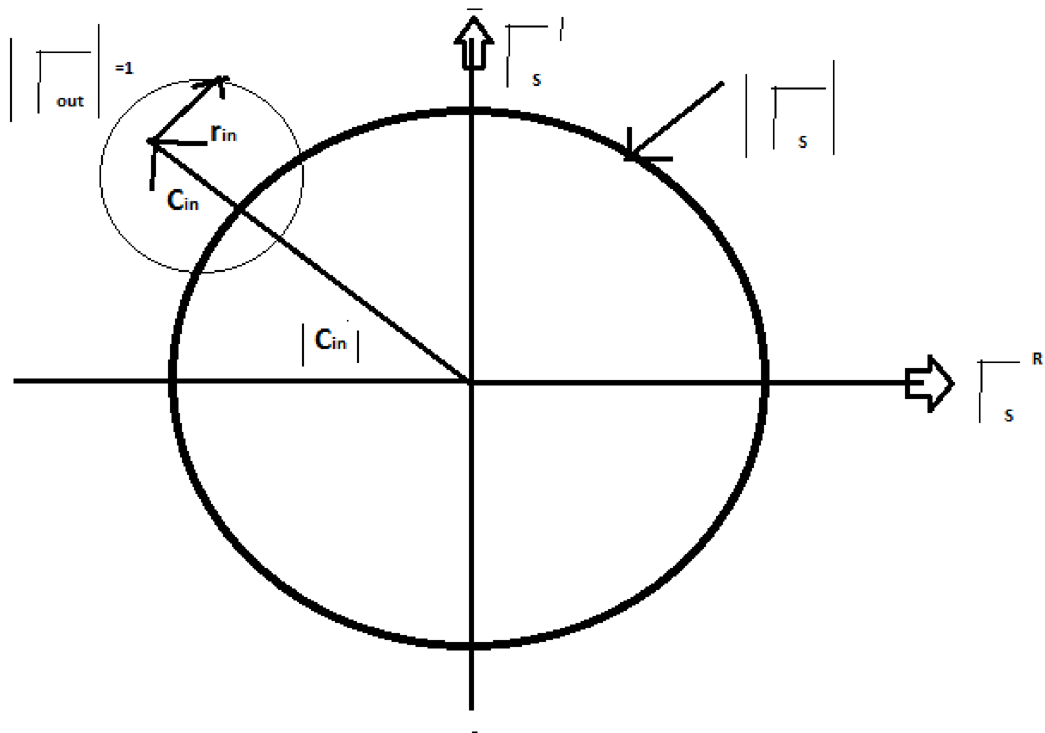

Fig: 5.3.2 Input stability circle

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5.4 VOLTAGE CONTROLLED OSILLATOR:

A Reactor-tuned oscillator uses a reactor diode to achieve the desired frequency tuning, as shown in Fig 5.4.1. A reactor diode is a two-terminal semiconductor device that utilizes the voltage-sensitive property of a PN junction. In these diodes, unlike regular diodes, the PN junction capacitance under reverse-bias conditions is accentuated by proper choice of diode profiles. (e.g., use of hyper-abrupt junctions).

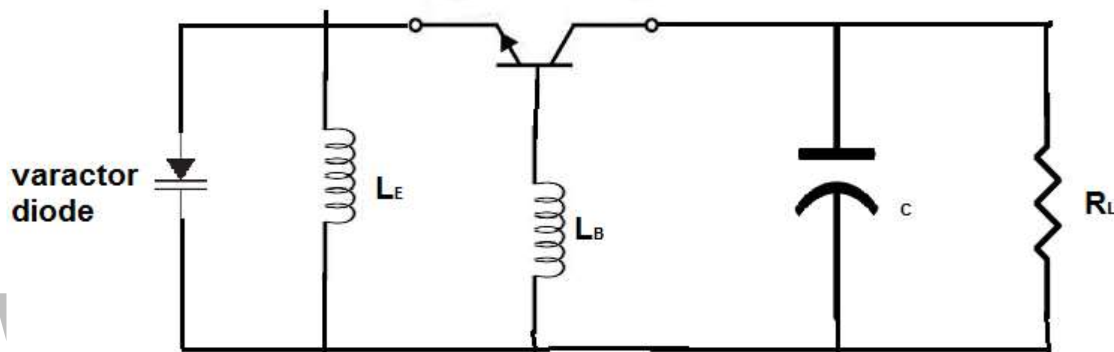

Fig: 5. 4.1 Reactor-tuned BJT oscillator

Assuming a one-sided junction, the junction capacitance due to diode's junction capacitance (C_j) cab be shown to be a junction of the applied bias voltage as follows:

$$C_j = \frac{C_{jo}}{(1+\frac{V_a}{V_{bi}})^s} \qquad \ldots\ldots (1)$$

Where,

C_{jo} - The junction capacitance at zero bias voltage

V_a - The applied bias voltage across the junction ($V_a = -V_r$)

V_{bi} – The built in voltage

and

S = $\frac{1}{m+2}$

"M" is the power4 parameter in the doping distribution function (N) of the lighter side, defined as:

N(x) = $N_B\, x^m$

And N_B is an arbitrary constant established by the doping profile.

Equal (1) is plotted in Fig 5.4.2. From this figure, it is seen that the capacitor value becomes very high for $V_a = V_{bi,}$ which is impractical.

A reactor is usually operated in the reverse bias region that is $V_a = -\ V_r$ <0.

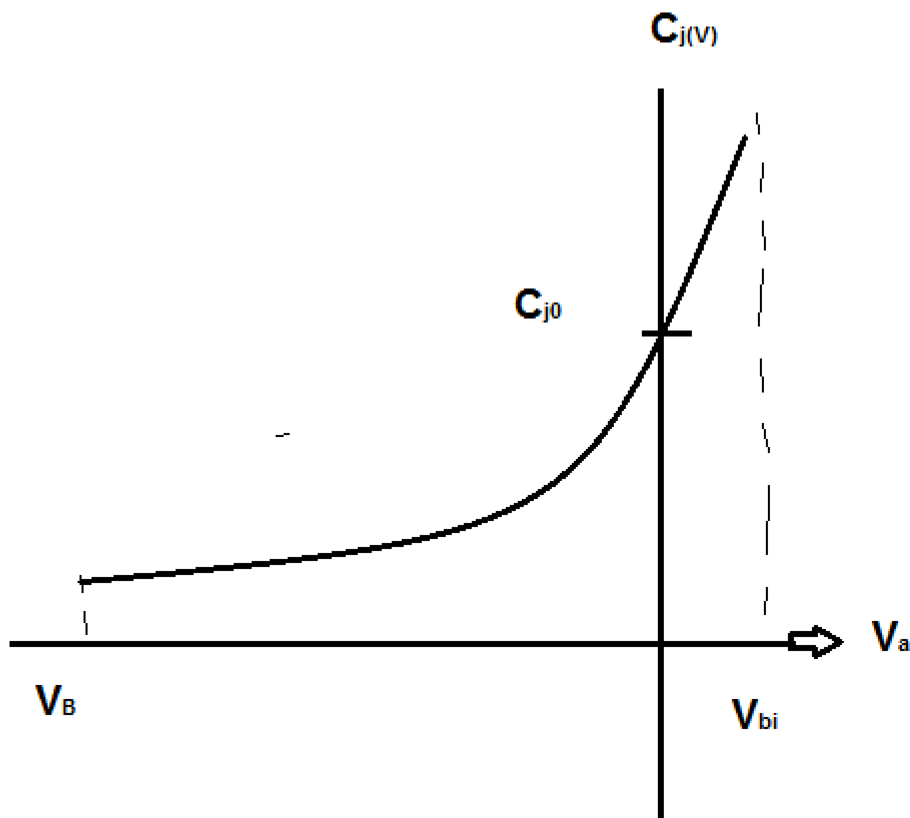

Fig: 5.4.2 reactor characteristic ($C_j$vs. V_a) with V_B as the breakdown voltage.

Then m=0

S=$\frac{1}{2}$

The equal (1) becomes,

$$C_j = \frac{c_{jo}}{\sqrt{1+\frac{v_a}{v_{bd}}}} \ldots\ldots\ldots\ldots.(2)$$

MIXERS:

A mixer is a nonlinear device that allows the translation of frequencies. It is used to bring either the modulated signal to the RF carrier or translate the RF signal to IF. When mixing two signals, a mixer generates both higher and lower translations (i.e., the sum and difference of input signals). A low pass or a high pass filter is required to select the desired frequency product. Generally, a LO is required as input to allow the signal down- or up-conversion. When the mixer is used as an up-converter (e.q., in a transmitter), the "sum frequency" is utilized and the "difference frequency" is rejected, as shown in figure 5.4.3.

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Various technologies are used to build mixers: A mixer can be based on diodes or transistors. In both cases, the device is nonlinear. Its nonlinear behavior has to be monitored (e.g., its 1dB-compression point and intermodulation distortion, etc.). Refer to references 20 and 29 for more information about mixers.

The **RF mixer** can be considered ON when the LO voltage switches it on and OFF when the local oscillator signal switches it off. This then acts upon the incoming signal on the **RF** port to enable the two signals to **mix** and provide the two output Signals

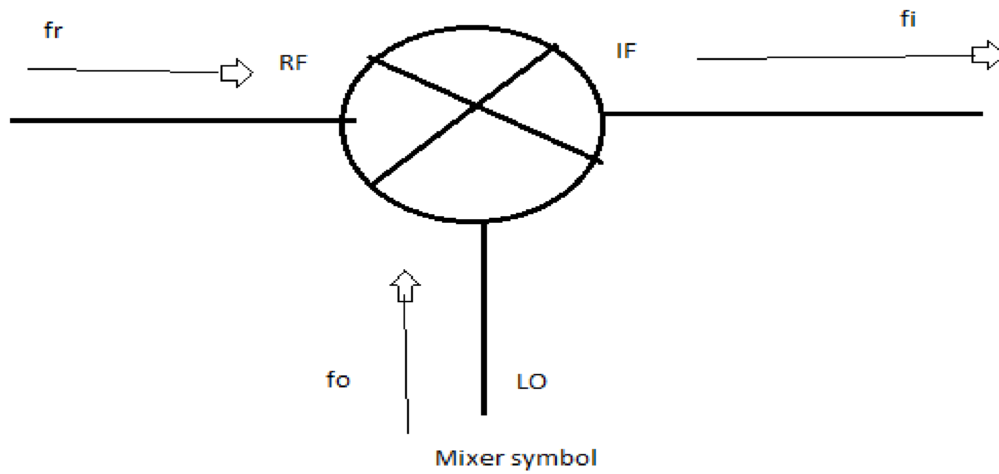

Fig: 5.4.3 Mixer Symbol

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