CHAPTER II

General Considerations on Electronics for Neutron Monitor and Theory of Stabilized Transistor Circuits

2.1 Nature of the pulse input signal and requirements for pulse shaping

The interaction of the neutron with the BF_3 - counter gives particle which releases a charge at the input of the succeeding electronic circuit. This pulse of charge is considered to be the pulse input signal. The nature of the final signal after passage through the electronic circuit depends on both the characteristics of the pulse input signal and the electronic circuit.

The basic requirements for counting apparatus are (1) the ability to count accurately, often up to high counting rates, and (2) the ability to separate the desired pulses, which are called the signal, from the unwanted pulses.

The measuring apparatus for counting consists of three units. These units are the amplifier, the discriminator, and the scaler. In the amplifier, the pulses are amplified and shaped. The function of the discriminator is to pass only the desired pulses. Finally, the scaler counts the pulse passed by the discriminator.

2.2 Pulse-amplifier requirements

Important considerations for pulse-amplifier include pulse shaping, noise, gain, maximum output signal amplitude, polarity, linearity, overload tolerance, and physical arrangement.

The required gain depends on both the available input signal and the desired output signal. The maximum usable gain is set by the noise level, and in practice this maximum is found to be of the order of 10^6.

The shape of the output pulse of an amplifier is determined by the shape of the pulse input signal and the characteristics of the amplifier. The schematic diagram shown in Fig 2-1 is an idealized amplifier representation is useful for discussing pulse shaping.

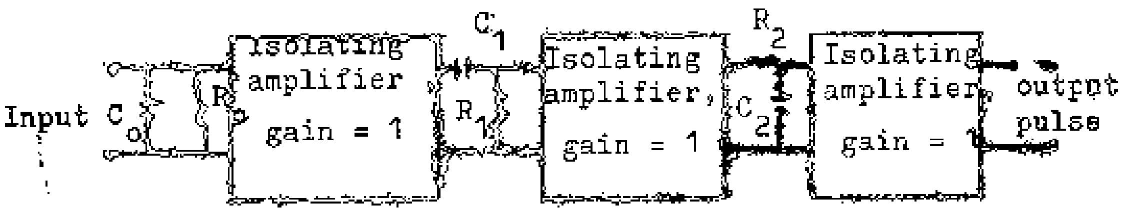


Fig 2-1 Equivalent circuit of a pulse input circuit and pulse amplifier

The circuit elements R_o and C_o include the characteristics of both the BF_3 - counter and the input of the amplifier. For example, consider Fig 2-2, which is a typical input circuit of a voltage-sensitive amplifier for use with a BF_3 - counter. The collector is at a high positive potential above ground, and a blocking capacitor C_1' is provided between it and the transistor base. The capacity C_2' represents the sum of the input capacity of the transitor and all wiring and stray capacities, in addition to any other capacity that may be placed at that position.

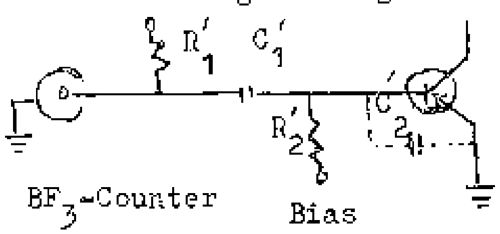


Fig 2-2 Input circuit for use with BF_3 - counter

Fig 2-3 is a simplification of the circuit in Fig 2-2 in which d-c potentials are disregarded (6).

C_3' = the capacity of the collector to ground

$C_1' \gg C_2', C_3'$

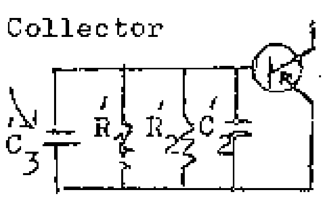


Fig 2-3 Input circuit for use in the analysis of pulse input signals.

As a further simplification, R_1' and R_2' can be replaced by the resistance R_6, which is equal to $R_1'R_2'/(R_1'+R_2')$, and the capacitance C_2' and C_3' can be represented by C_o , which is equal to $C_2'+C_3'$.

The circuit consisting of R_1 and C_1 , known as the differentiation network, determines the lower-frequency cut off and the clipping time of the amplifier. The time constant R_1C_1 , designated as Υ', is known as the differentiation, or clipping time-constant. The circuit R_2C_2 , determines the upper-frequency cut off of the amplifier and affect the rise and delay times of the pulses.

The pulse input signal may be either positive or negative, although the signals as produced by the circuit in Fig 2-2 are mostly negative.

Linearity between pulse input signal and output signal of 0.5 percent or better is often necessary. This is obtained by the choice of the operating points for the transistors within their linear range and by the use of negative feedback and other methods.

The use of the separate preamplifiers has the additional advantage that it makes it possible to provide the shortest possible leads to the counter and thereby, minimizing the input capacity. The gain of the preamplifier depends on the application, with common values varying from less than 1 to 100. Its output stage is always emitter follower, which can drive the long cable required to couple it to the main amplifier.

The gain control of a pulse amplifier is located several stages beyond the input. The location of the gain control is determined by balancing of the signal-noise ratio and the overload problem.

The position of the pulse-shaping circuit also needs special considerations. When low noise level is important, the short time constant cannot be placed in the early stages, since the low resistance associated with it would make the thermal noise intolerable. On the other hand, the pulse shaping cannot be left to the high-level stages because of the shift in the operating point of the transistors.

The shift occurs because of the pile-up of the pulses with the long decay constants. The usual position for the pulse-shaping network is after a gain of about 100, which usually falls between the preamplifier and the main amplifier. A further advantage of placing the pulse-shaping net work at the intermediate level rather than at a lower level is that it serves to reject the hum and other low-frequency noise generated at the low-gain levels of the amplifier.

Stability of the amplifier gain is quite important. In counting measurements a change in gain may result in a shift in the refraction of the pulses passing the discriminator. A change of a small fraction of 1 percent in the output voltage for a 1 percent change in line voltage is not an uncommon requirement. The primary methods employed for stabilization of the gain are the use of negative feedback in the amplifier section and the provision for good regulation in the power supply section.

2.3 Bias stability

One of the basic problems involved in the design of transistor amplifiers is establishing and maintaining biasing conditions (6). They must be maintained despite variations in ambient temperature and variations of gain and leakage current between transistors of the same type. This can readily be seen by referring to Fig 2-4(a) where the transistors are operated in the common emitter mode and are biased by a constant base current, I_B. Fig 2-4 (b) shows the common emitter collector characteristics of two different transisters with the same collector load line superimposed on them. For the transistor characteristic shown with solid lines and a base current I_B, the operating point is at A. On the other hand, if a higher gain transistor is used, or the original transistor's gain and leakage current are increased due to an increase in temperature, the transistor characteristic shown with dashed line could result.

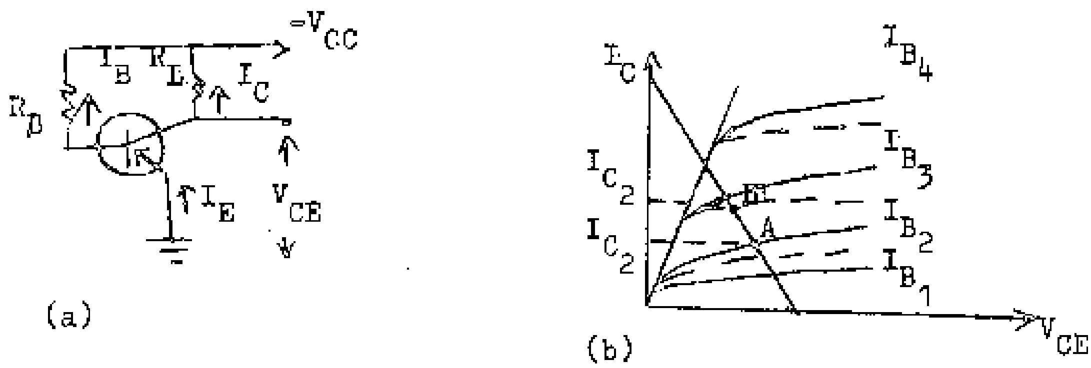


Fig 2-4 Shift of operating point.

2.4 The leakage current I_{CBO}

The leakage current I_{CBO} is the collector current when the collector is biased in the reverse (high resistance) direction with respect to the base, and the emitter is open-circuited. This current is made up of two components, one temperature-dependent and one voltage dependent (7).

I_{CBO} is of primary concern in transistor biasing. Because of its extreme temperature dependence it can be come an appreciable part of the base current in low-level applications, and it can cause self-heating and thermal runaway in large - signal applications.

The real collector current, including the leakage current, that flows in any common-emitter connection is (8)

$$I_C = h_{FE}I_B + (1+h_{FE})I_{CBO} \qquad (2.1)$$

instead of $I_C = h_{FE}I_B$

where $\quad h_{FE}$ = common-emitter current gain.

Since the I_{CBO} contribution is multiplied by h_{FE} ; it can be expected to affect drastically the magnitude of common-emitter collector current as temperature is raised.

2.5 Stability factors

Equation (2.1) shows that any variation of I_{CBO} due to temperature will manifest itself in a shift upwards in the collector characteristics of the transistor as shown in Fig 2-4. The degree to which any bias circuit actually maintains the operating point is measured by what is called stability factor. There are several stability factors, one is a measure of the effects of I_{CBO} in moving the operating point, the second is a measure of the effects of h_{FE} in moving the operating point. Mathematically the stability factors can be stated as

$$S' = \frac{\Delta I_C}{\Delta I_{CBO}} ,$$

and

$$S'' = \frac{\Delta I_C}{\Delta h_{FE}} ,$$

where S' = leakage stability factor

ΔI_C = change in collector current

ΔI_{CBO} = Change in leakage

S'' = current gain stability factor

ΔI_C = change in collector current

Δh_{FE} = change in current gain.

For any transistor where both leakage and current gain are changing simultaneously, the total incremental **change** in current is approximate by equal to the sum of the separate leakage and current-gain increments as predicted from the stability - factor calculations. The least stability factor is the best for the stability case.

2.6 Direct-coupled amplifier circuits

Direct-coupled amplifiers are commonly used in a wide variety of applications. In many of these applications the signal may contain a d-c component as well as low-frequency components. Signals of this type are encountered in many instruments and apparatus, such as in the electronic part of the neutron monitor. Consequently this type of amplifier requires direct coupling between the stages without the use of coupling capacitors.

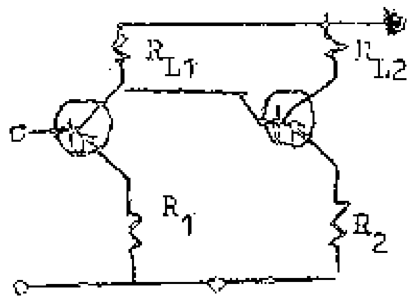


Fig 2-5 Simple d-c transistor amplifier with two common-emitter stages

Fig 2-5 shows one possible arrangement of this type consisting of two common-emitter stages in which the signal developed across the load resistance R_{L1} is applied directly to the base of the second transistor.

For proper operation of such an amplifier, the d-c voltage drop across resistance R_{L1} must be sufficient to maintain the emitter of the second transistor slightly positive with respect to the base.

2.6.1 Problem of drift

In d-c transistor amplifiers, drift is caused primarily by random change of temperature. All parameters of transistor vary with temperature. The current-gain is very sensitive to temperature change.

The current I_{CBO} may be expressed (9) approximately as

$$I_{CBO} = I_S \exp (0.05T)$$

Where I_S is the current at 0^o C, and T the temperature in degree centigrade. Hence

$$\frac{dI_{CBO}}{dT} = 0.05\, I_S \exp (0.05T)$$

$$= 0.05\, I_{CBO}$$

The incremental change of this current can be written as

$$i_{CBO} = 0.05\, I_{CBO} \Delta T$$

Where Δ T is the temperature change in degrees centigrade.

In general the temperature coefficient for I_{CBO} is about 5 - 10 percent per degree centigrade. In view of this, temperature compensation of the d-c transistor amplifier becomes a serious problem.

2.6.2 Multistage stability considerations.

The previous section dealt with the variation of collector current with I_{CBO}. For multistage direct-coupled transistor amplifiers, this problem becomes many times more severe, as the variation of collector current of one transistor is passed on to the next transistor, where the variation is amplified and combined with the similar variation due to changes in the second unit. As stages are added the effect is cascaded until finally a very small drift in the first unit may cause the whole chain to become inoperative.

2.7 Feedback amplifiers.

Negative feedback in amplifiers improve such properties as stability under varying power supply conditions, stability to parameter variation, freedom from non-linearities. We will now analyze a few of the methods of applying feedback to transistor amplifier very briefly. The common-emitter circuit is considerably more useful for transistor applications, and emphasis will be placed on this type of configuration.

Consider the block diagram shown in Fig 2-6. Here we are characterizing the active amplifier in terms of voltage gain A_v, voltage feedback is being applied to the input, through network B. In situations where the feedback network is passive we may define B as the voltage-feedback factor. The quantity represents the percentage of the output voltage that is fed back to the input.

Mathematically

$$B_v = \frac{v_f}{v_{out}} \qquad (2.2)$$

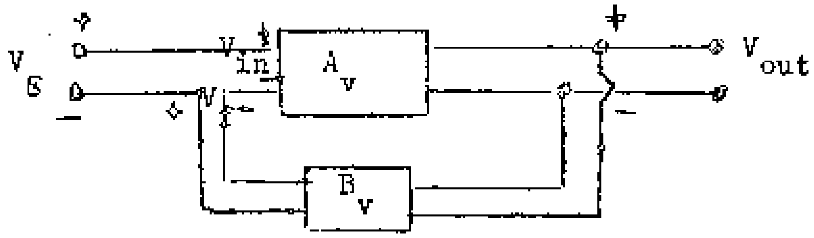


Fig 2-6 Block diagram of voltage feedback network

For the basic amplifier without feedback

$$A_v = \frac{v_{out}}{v_{in}} \qquad (2.3)$$

In the case of a power source that is a voltage source we can also say

$$A_v = \frac{v_{out}}{v_g} \qquad (2.4)$$

When feedback is applied, the toatl voltage condition at the input requires a summation of the individual voltages:

$$v_g = v_{in} - v_f \qquad (2.5)$$

To find the performance of the total amplifier with feedback, we substitute (2.5) into (2.4) to obtain

$$A'_v = \frac{v_{out}}{v_{in} - v_f}$$

where A'_v = voltage gain with feedback. Substituting (2.2) for v_f and dividing numerator and dominator by v_{in}, we obtain

$$A'_v = \frac{A_v}{1 - B_v A_v} \qquad (2.6)$$

For common-emitter amplifiers there is an inherent 180° phase shift between input and output ; that is, A_v is a negative quantity. The absolute value of (2.6) is

$$A_v' = \frac{A_v}{1+B_v A_v} \qquad (2.7)$$

Consider now the block diagram shown in Fig 2-7. Here we characterize the active amplifier in terms of current gain A. Current feedback is being applied to the input, out of phase with the input through network B_i , where B_i is the current-feedback factor.

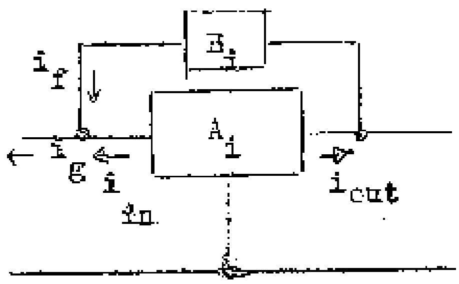


Fig 2-7 Block diagram of current feedback network.

In the same way as described for the voltage-feedback case, we shall get (8)

$$A_i' = \frac{A_i}{1+B_i A_i} \qquad (2.8)$$

Where A_i' = current gain with feedback.

The general case of feedback equation would be

$$A' = \frac{A}{1+BA}$$

Analysis of any given circuit will indicate the type of feedback that is most effective. When voltage sources are involved, voltage feedback is appropriate. Since the feedback voltage appears in series with the source voltage, this type of feedback is called series feedback. When current sources are involved, current feedback is appropriate. Since the feedback current parallels the source current in driving the input, this type of feedback is called shunt feedback.

Besides affecting gain, negative feedback increases the frequency response of an amplifier. The general-case equations applicable are (8)

$$f_L' = \frac{f_L}{1+BA}$$

$$f_H' = f_H\,(1+BA)$$

Where f_L' = low-frequency half-power point with feedback, and f_H' = high - frequency half-power point with feedback.

-Shunt-type Feedback

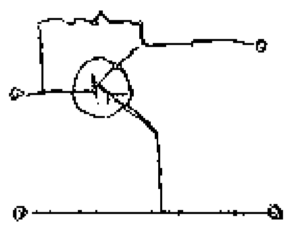


Fig 2-8 Shunt feedback

Since the grounded-emitter circuit has a 180° phase shift in the region where high-frequency effects are negligible, such a circuit supplies negative feedback. For such a circuit we have (9) , as a result of the feedback, a reduction in gain and input impedance as well as in output impedance.

-Series-type feedback

Fig 2-9 Series feedback

This type of feedback results in increased input and output impedances and decreased gain.

-Multistage series or shunt feedback

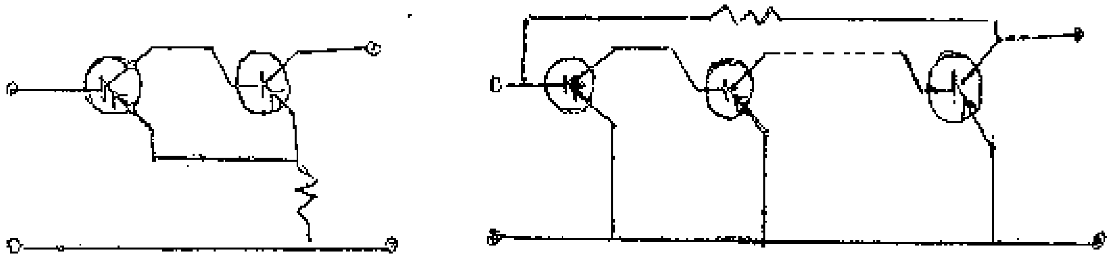


Fig 2-10 (a) Multistage series feedback

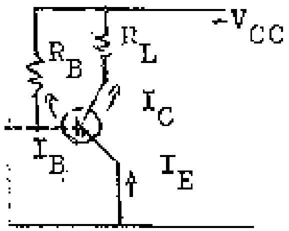
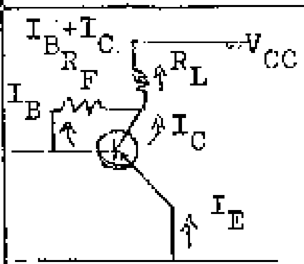
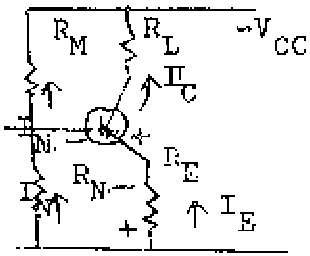
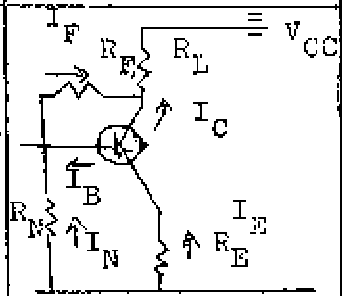
(b) Multistage shunt feedback

2.8 Comparison of stability factors involving feedback

The stability factors for fixed biasing, current-feedback biasing, voltage-feedback biasing and combination current-voltage-feedback biasing are tabulated in Table 2.1 (6).

2.8.1 Fixed biasing.

In spite of its simplicity, the fixed-bias circuit has a major drawback. The operating point is very responsive to change in I_{CBO} and h_{FE}. Since I_B is fixed in value, any changes in h_{FE} or I_{CBO} are factored directly into the collector-current expression. The shifts in collector current with changes in temperature and changes in transistors can be drastic.

type of biasing	circuit	leakage stability factor (S') $(S' = \frac{\Delta I_C}{\Delta I_{CBO}})$	current gain stability factor (S'') $(S'' = \frac{\Delta I_C}{\Delta h_{FE}})$
fixed-biasing		$1 + h_{FE}$	$I_B + I_{CBO}$
current-feedback-biasing		$\frac{(1+h_{FE})\,(R_L + R_F)}{R_F + R_L + h_{FE}\,R_L}$	$\frac{(R_L + R_F))(V_{CC}-V_B+R_F I_{CBO})}{(R_F+R_L+R_L h_{FE})^2}$
voltage-feedback-biasing		$\frac{(1+h_{FE})\,(1+\frac{R_E}{R_M}+\frac{R_E}{R_N})}{1+(1+h_{FE})\,(\frac{R_L}{R_M}+\frac{R_E}{R_N})}$	$\frac{(\frac{V_{CC}-V_B}{R_M} - \frac{V_B}{R_N}+I_{CBO})(1+\frac{R_E}{R_M}+\frac{R_E}{R_N})}{\left[1+(1+h_{FE})(\frac{R_E}{R_M}+\frac{R_E}{R_N})\right]^2}$
combination current-voltage feed-back-biasing		$\frac{(1+h_{FE})(1+\frac{R_E}{R_L+R_F}+\frac{R_E}{R_N})}{\frac{R_F}{R_L+R_F}+(\frac{R_E+R_L}{R_L+R_F}+\frac{R_E}{R_N})(1+h_{FE})}$	$\frac{(\frac{V_{CC}-V_B}{R_F+R_L} - \frac{V_B}{R_N}+\frac{R_F I_{CBO}}{R_L+R_F})(1+\frac{R_E}{R_L+R_F}+\frac{R_E}{R_N})}{\left[\frac{R_F}{R_L+R_F}+(1+h_{FE})(\frac{R_E+R_L}{R_L+R_F}+\frac{R_E}{R_N})\right]^2}$

2.8.2 Current-feedback biasing (See figure in Table 2.1)

In the fixed-bias approach, the variation of operating point is too excessive to be acceptable for most applications. It is necessary to improve the stability factors in order to minimize this variation.

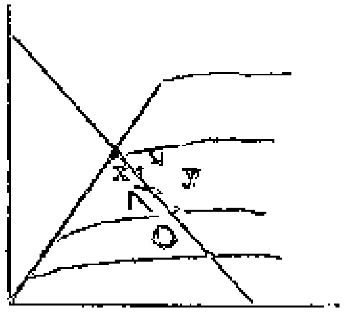


Fig 2-11 Graphical depiction of operating point control due to V_{CE} variation.

Consider the graphical depiction of the current-feedback circuit shown in Fig 2-11. Point O is the desired operating point for the circuit. As temperature increases, leakage increases, and the collector-current tends to increase towards point x. At this time, the effects of returning the bias resistor directly to the collector current causes a decrease in the collector-to-emitter voltage. Since this voltage determines the base current, there is a decrease in the base current. The operating point does not move to point x but rather ends up at an interim point, y.

2.8.3 Voltage-feedback biasing (See figure in Table 2.1)

Another type of feedback from the output is applied through a resistor in the emitter lead.

Any tendency for the collector current and hence the emitter current causes an increase in the IR drop in the emitter resistance. The polarity of this voltage is opposite to the normal forward-bias voltage between base and emitter. There is , then, some degree of compensation for operating-point excursion.

2.8.4 Combination current-voltage feedback.

The voltage-feedback circuit loses its effectiveness as R_L increases. At such a time, however, the introduction of current feedback is advantageous.

This method is generally capable of maintaining a very high degree of operating-point stability (8).

The principles discussed in conjunction with the d-c effects above can apply also to the a-c effects.

2.9 Bias compensation

In many cases, particularly in d-c amplifiers, feedback stabilization can be used only sparingly, since it reduces the d-c gain as stability factor is reduced. It is often possible to reduce drift of the operating point by using non-feedback or compensation method without reducing the d-c gain.

One of the simplest and most reliable compensation methods involves the use of semiconductor junction diode (10).

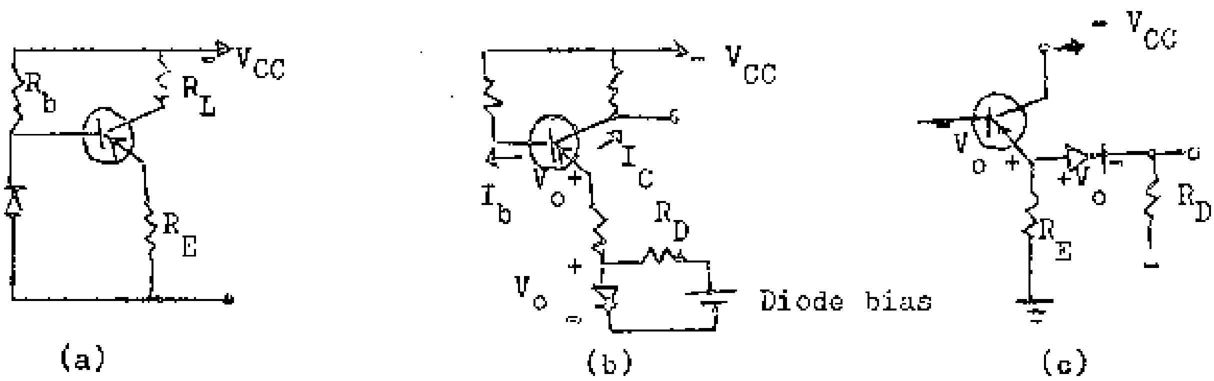


Fig 2-12 Diode Compensation

The significant thing in Fig 2-12 (a) is that as temperature increases, the forward drop in the bias diode decreases, cutting back on the forward bias of the transistor, compensating for what otherwise would have meant a shift in the operating point.

Fig 2-12 (b) shows a diode used to compensate for base-to-emitter voltage changes in a transistor. If the voltage across the diode and the base-to-emitter junction are equal the forward diode characteristic is very similar to the base-to-emitter transistor characteristic. This means that since these two voltages vary with temperature at the same rate, I_b is independent of V_o.

A series diode may also be used with the emitter follower circuit to keep the output voltage constant, as shown in Fig 2-12 (c).

Temperature sensitive resistors are also useful in compensating transistor drift. The most widely used of these is the thermistor, which has a negative temperature coefficient and a resistance that decreases exponentially with temperature.

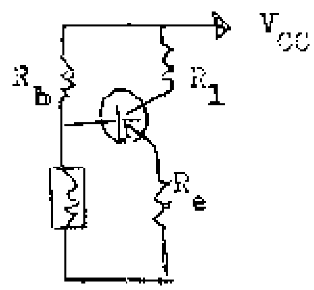


Fig 2-13 Thermistor compensation.

2.10 Temperature compensation of d-c amplifiers by suitable arrangement.

One of the methods which may be employed is based on proper selection of the characteristics of the various transistors employed in the d-c amplifier. If both transistors are of the same type, only the following five connections can be used for temperature compensation (9) of the amplifier:

1. Common-emitter common-emitter.
2. Common-base common-emitter.
3. Common-base common-collector.
4. Common-emitter common-collector.
5. Common-collector common-base.

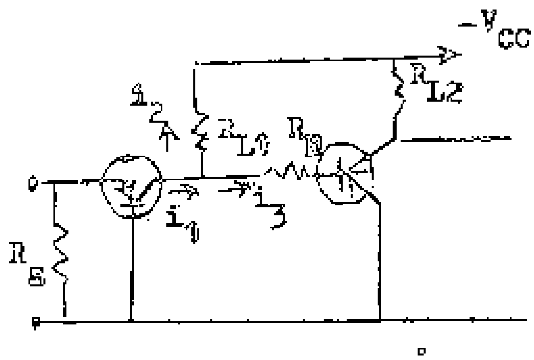


Fig 2-14 Circuit arrangement for temperature stabilization of a d-c amplifier.

Fig 2-14 shows another arrangement which may be used for temperature stabilization of the amplifier. The current increasment i_1 due to change of temperature of the first transistor is divided into currents i_2 and i_3 flowing through resistances R_{L1} and R_D respectively.

Owing to current increment i_3, a current increment will appear (with phase reversed) on the collector side of the second transistor. The ratio between R_{L1} and R_D may be so chosen that this current increment will compensate the current increment which is due to the change of temperature of the second transistor, resulting in the net increment i_1 becoming zero.

2.11 Half supply voltage principle.

When a transistor is used at high junction temperatures it is possible for regenerative heating to occur which will result in thermal run-away and possible destruction of the transistor. If the collector-emitter voltage is less than half the supply voltage in common-emitter mode the transistor will be stable from thermal run-away (14). This is the half supply voltage principle.