CHAPTER II

THEORY OF COMPANDOR

2.1 Basic Principles

The compandor consists of two separate units, a compressor at the transmitting terminals of the circuit, and an expander at the receiving terminals.

The gain of both units is a function of signal level. The gain of the compressor at the sending end of the circuit is made to vary inversely as a level of transmitted signal i.e. the gain is greatest for the weakest signal. The weaker components of the signal are thus given a "compressor advantage" over the noise or crosstalk occurring after the compressor. The gain of the expander at the receiving end of the circuit is made to vary directly as the level of the received signal, i.e. the gain is least for the weakest signal. It follows that the gain of the expander is least when there is no signal and only noise or interference is present. This expander action reduces the noise in the silent periods given "expander advantage".

If the compressor and expander are complementary there is no net gain to an input signal, but the compressor advantage added to the expander advantage, produces a substantial subjective improvement in signal-to-noise ratio. The ideal steady state characteristic of the compandor is illustrated in Fig. 1.

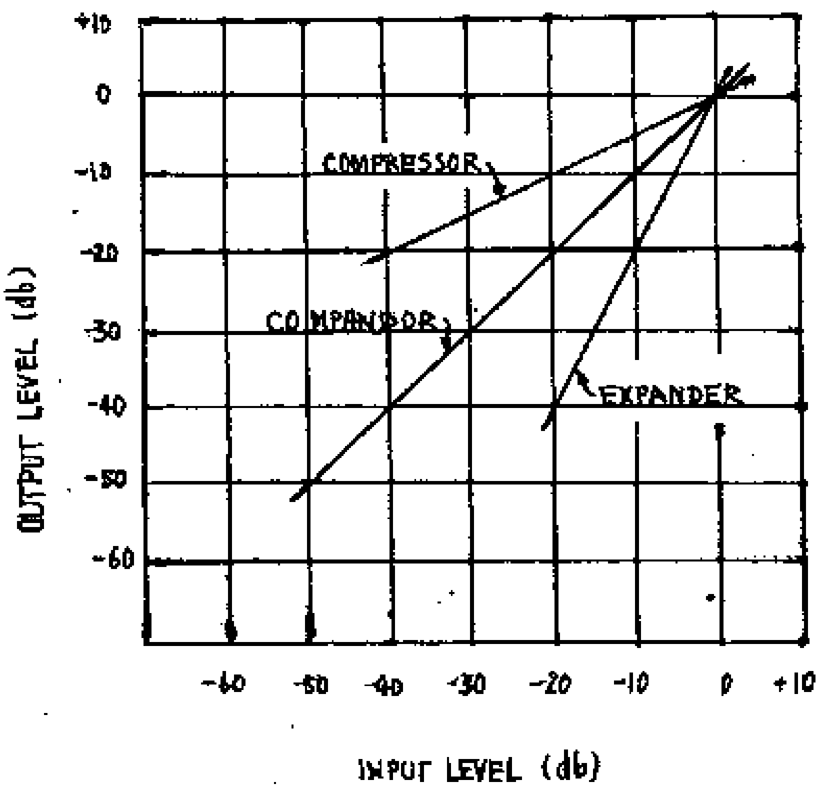


Fig. 1 -- Ideal steady state characteristic of compandor.

2.2 Basic Circuits of Compandor

2.2.1 Basic Circuit of the Compressor

The basic circuit of the compressor is shown in Fig 2. It consists of variable loss device having transfer function of G_{cv} ; followed by an amplifier having a high gain G_a, and providing a subsidiary output which after rectification,

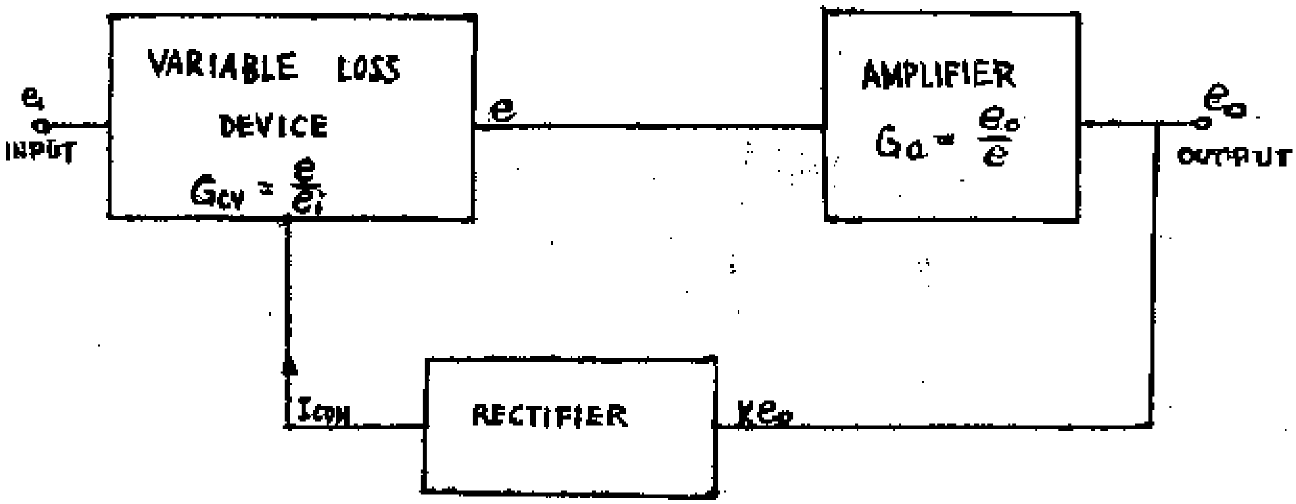


Fig. 2 -- Basic Circuit of Compressor

controls the loss of the variable-loss device. A 2:1 compression ratio is achieved if, over the required range of signal levels, the control current is directly proportional to the level of the signal at the input to the rectifier and transfer function of the variable loss device is inversely proportional to the control current.

2.2.2 <u>Basic Circuit of the Expander</u>

The basic circuit of the expander is shown in Fig 3. It consists of variable loss device having transfer function of G_{Ev}, and followed by an amplifier having a gain G_a. A fraction of the input is rectified to provide current to control the variable loss device. A 1:2 expansion ratio results if the control current is directly proportional to

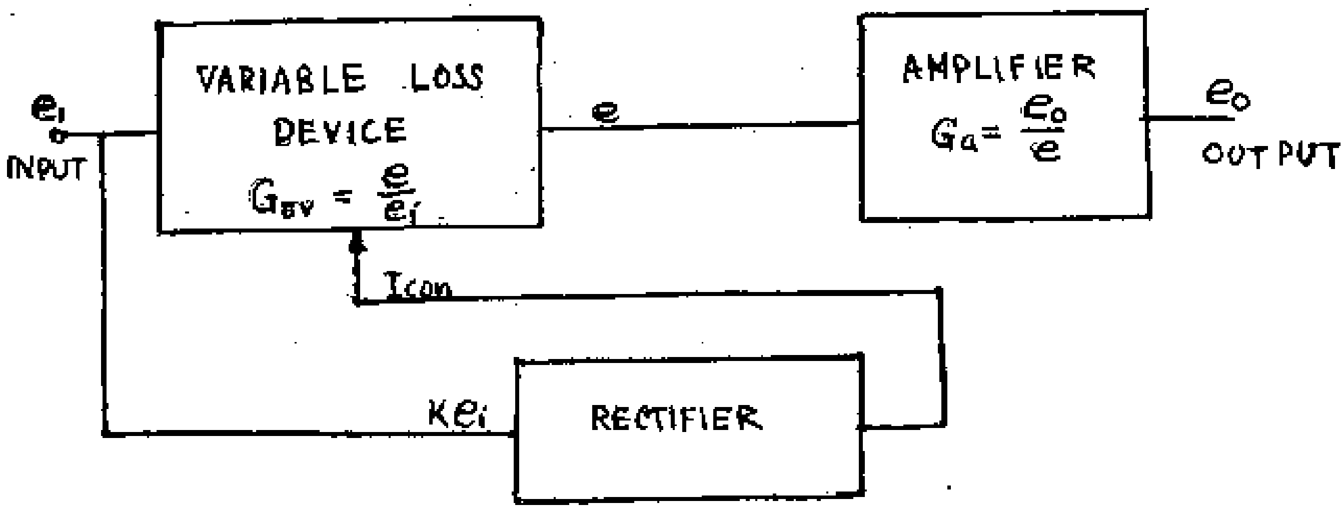


Fig. 3 — Basic Circuit of Expander

the level of the signal at the input to the rectifier and the transfer function of the variable-loss device is directly proportional to the control current.

2.3 Compressor Variable Loss Device

Refer to Fig 2. Let

e_i = input signal to the variable loss device,

e = output of the variable loss device or input of the amplifier.

e_o = output of the compressor, or output of the amplifier.

G_{cv} = transfer function of the variable loss device = $\frac{e}{e_i}$

G_a = Gain of the amplifier $= \frac{e_o}{e}$

I_{con} = Control current from rectifier

The design procedure of the compressor variable-loss device has been followed by the following steps:

<u>1st step</u>.

The variable loss device consists of transistors, requiring the 2:1 compression ratio of the compressor.

<u>2nd step</u>.

The compression ratio $\frac{20 \log e_1}{20 \log e_o} = 2:1$ ——(2.1)

therefore $e_1 = e_o^2$

also $e_1 = (e\ G_a)\ e_o$

hence $e/e_1 = \frac{1}{G_a\ e_o}$

that is $G_{ov} \propto \frac{1}{e_o}$ ——(2.2)

For a required compression ratio of 2, the transfer function of the variable loss device is inversely proportional to the output of the compressor.

<u>3rd step</u>. – Design of the rectifier

The control current is directly proportional to the level of the signal at the input to the rectifier.

$I_{con} \propto e_o$ ——(2.3)

therefore from (2.2) $G_{ov} \propto \frac{1}{I_{con}}$ ——(2.4)

<u>4th step</u>

The variable loss device can be considered as a four terminal active network as shown in Fig 4.

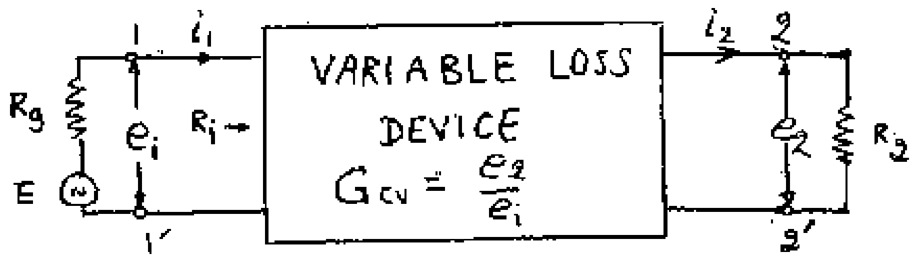


Fig 4 — Four terminal active network

The transfer function $G_{cv} = \frac{e_2}{e_1} = \frac{i_2 R_2}{i_1 R_1}$ —(2.5)

Where E = generator voltage

e_1 = voltage across terminals 11'

e_2 = voltage across terminals 22'

i_1 = input current

i_2 = output current

R_g = generator resistance

R_1 = input resistance looking into terminals 11'

R_2 = resistance across terminals 22'

From equations (2.4), (2.5) it can be seen that G_{cv} will vary inversely with input current when R_1 has small dependence on R_2 and this can be achieved by having a variable loss network with low input resistance connected in shunt with R_2.

This will result in i_2 to have small effect on i_1 as shown in Fig 5.

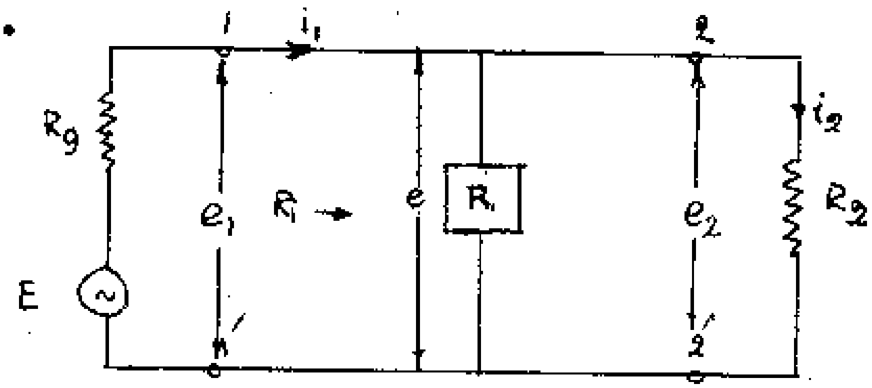


Fig 5 — Variable loss as shunt-element

Using a variable loss circuit as a shunt element, voltages across terminals 11 and 22 are equal to e.

Therefore $$G_{cv} = e/E \qquad \text{———(2.6)}$$

Let R = input resistance of variable loss device

then, $$R_1 = \frac{R\,R_2}{R + R_2} = \frac{1}{1 + R/R_2} \qquad \text{——(2.7a)}$$

If $R \ll R_2$,

$$R_1 \approx R \qquad \text{———(2.7b)}$$

therefore $$G_{ov} = \frac{e}{E} = \frac{i_1\,R}{E} = \frac{i_1\,R}{i_1\,(R_g + R)}$$

$$= \frac{R}{R_g + R} \qquad \text{———(2.8a)}$$

And if $R_g \gg R$, then

$$G_{ov} = \frac{R}{R_g} \qquad \text{———(2.8b)}$$

That is $$G_{ov} \propto R \qquad \text{———(2.9)}$$

The transfer function of the variable loss device varies as its input resistance .

5th step

From equations (2.4) and (2.9), it is necessary to select the transistor configuration that has very low input resistance and vary inversely as the control current.

From study of parameters relating with operating points, measurements of the various parameters which have been made on a typical diffusion-type p-n-p junction transistor yield the curves shown in Fig 6, 7, 8, 9, 10 and 11.

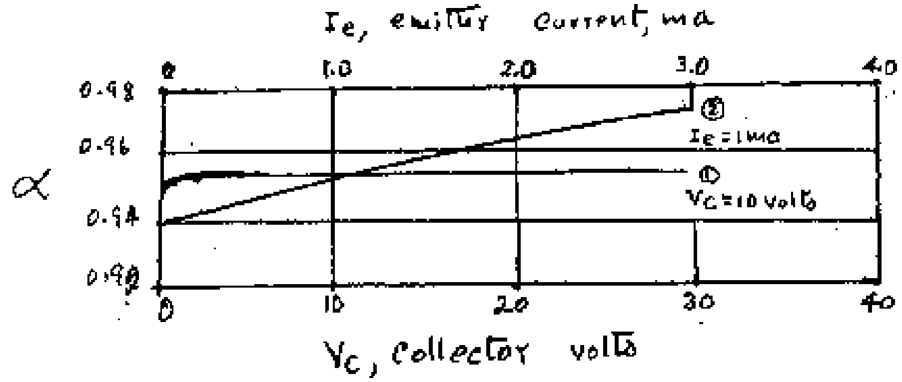


Fig 6 — Variation of α with emitter current and collector voltage.

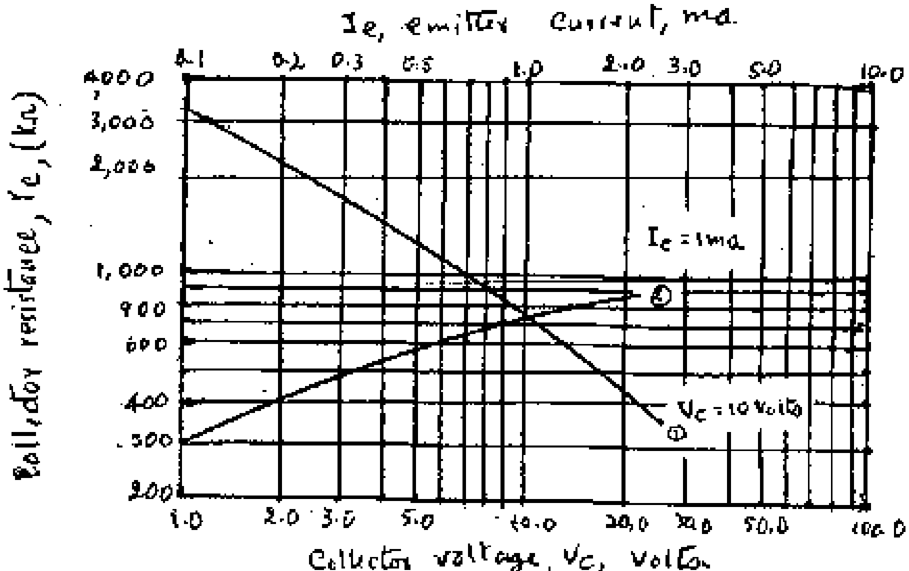


Fig 7 — Variation of collector resistance with emitter current and collector voltage.

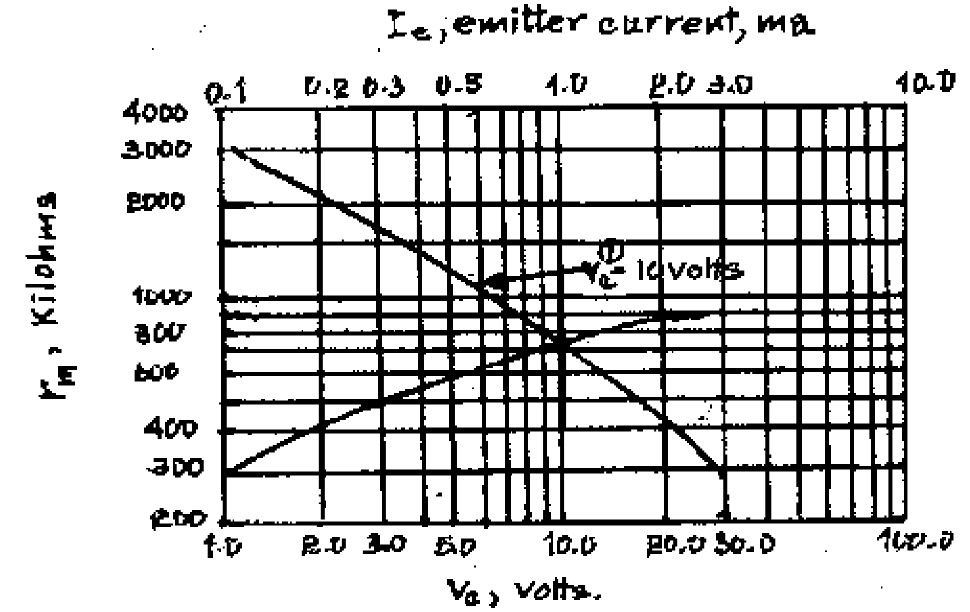


Fig.8 Variation of r_m with emitter current and collector voltage.

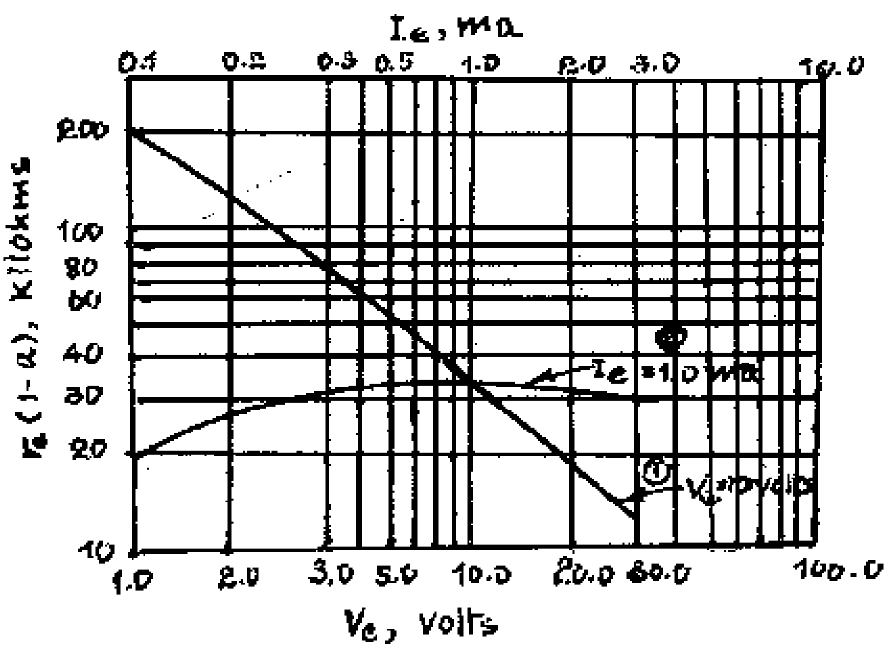
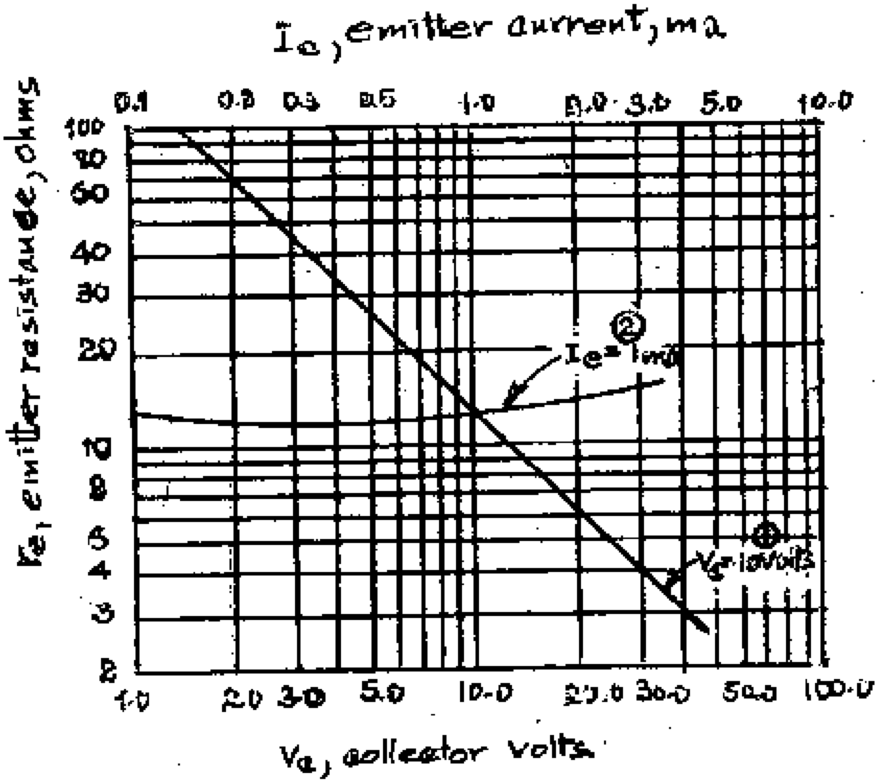
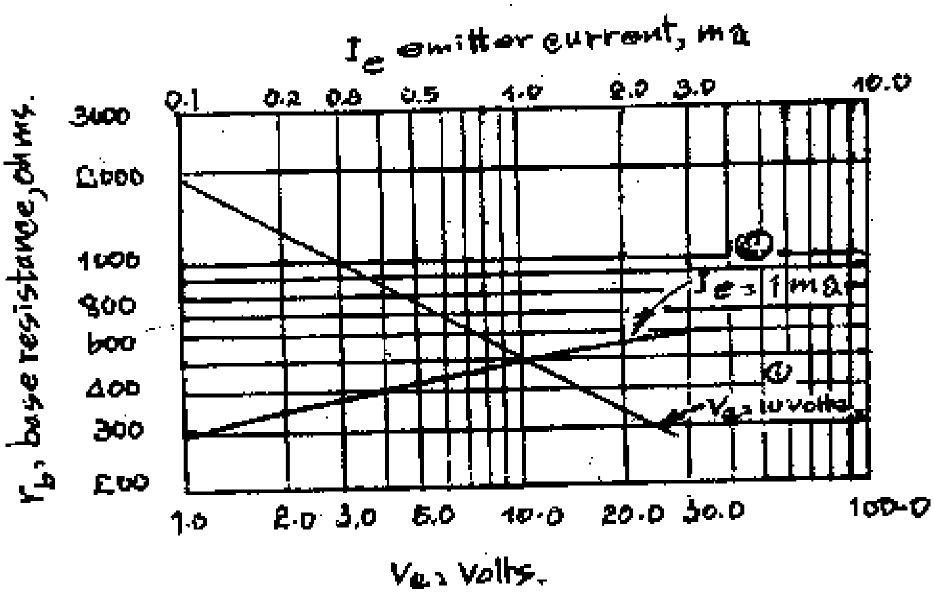


Fig9-Variation of $r_c(1 - a)$ with emitter current & collector voltage.



Fig—10 Variation of emitter resistance with emitter current and collector voltage.



Fig—11 Variation of base resistance with emitter current and collecter voltage.

It has been shown that the variation of in the range to be considered is small enough to be negligible, and r_e would be expected to vary, approximately, in inverse proportion to the emitter current, and can be expressed by the relation

$$r_e = r_{e1} / I_e \qquad \text{——(2.10)}$$

where r_{e1} is the emitter resistance measured at 1 milliampere emitter current,

r_e = emitter resistance in ohms, and

I_e = emitter current in milliampere.

It also appears that over the range of measurement, r_c, r_m, $r_c(1 - a)$, and r_b vary approximately inversely with emitter current I_e.

Therefore, the requirement that the input resistance varies inversely as control current can be achieved by using the control current as the emitter current. Another problem is the need of a very low input resistance of a transistor configuration.

Considering the input resistances of the three transistor configurations, they yield the following expressions:-

Common-base $$r_i = r_e + r_b \cdot \frac{r_c - r_m + r_1}{r_b + r_c + r_1} \qquad \text{——(2.11)}$$

Common-emitter $$r_i = r_b + r_e \cdot \frac{r_c + r_1}{r_c - r_m + r_e + r_1} \qquad \text{——(2.12)}$$

Common-collector $$r_i = r_b + r_c \cdot \frac{r_e + r_1}{r_c - r_m + r_e + r_1} \qquad \text{——(2.13)}$$

where

r_i = a.c. input resistance

r_e = equivalent emitter resistance

r_b = equivalent base resistance

r_c = equivalent collector resistance

r_m = equivalent emitter-collector trans-resistance

r_l = a.c. load resistance

From Fig 7,8,9,10,11 and equations (2.11), (2.12), (2.13), for a given operating point and a load resistance, the input resistance of common-base is the lowest and the common-collector is the highest. Therefore, the common-base is the most satisfactory configuration for the requirement.

6th step

Since a variable loss device is used as a shunt element, for good stability of the circuit, the collector-base voltage should be held constant. So we choose a very low r_l or a short-circuit load , thus the input resistance becomes

$$r_i = r_e + r_b(1 - a) \qquad \text{——(2.11a)}$$

where $a = r_m/r_c$ = current amplification factor

7th step

1) To obtain high load and generator resistances compared with input resistance of the variable loss device, a matching transformer will be used.

2) To eliminate direct energizing and even-order

harmonics and to split the input signal into two equal levels to ensure the small input amplitude, a push-pull connection as shown in Fig 12 is used as a variable-loss circuit for the compressor.

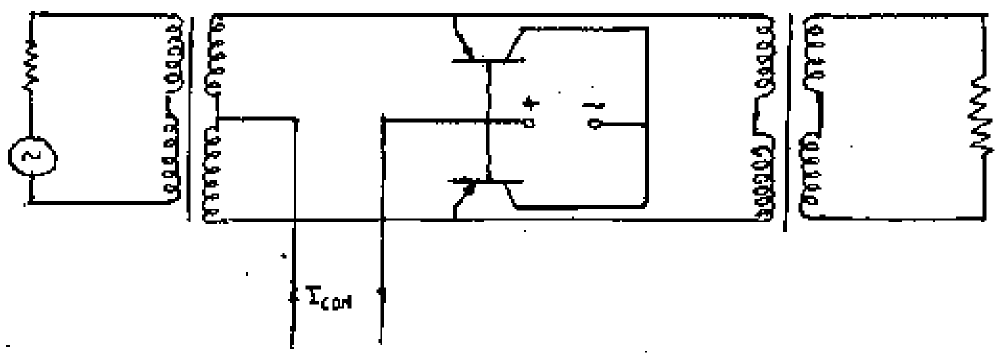


Fig 12 — Compressor variable loss device

2.4 <u>Expander Variable Loss Device</u> 907026

Refer to Fig 3, let

e_i = input signal to the variable loss device

e = output of the variable loss device

e_o = output of the expander

G_{Ev} = transfer function of variable loss device = $\frac{e}{e_i}$

G_a = gain of the amplifier = $\frac{e_o}{e}$

I_{con} = control current from rectifier

The procedure of the design of the expander variable-loss device is as follows:

<u>1st step</u>

It consists of a transistor circuit requiring the 2:1 expansion ratio of the expander.

2nd step

Since the expansion ratio $= \dfrac{20 \log e_o}{20 \log e_i} = 2:1$ ——(2.14)

therefore $e_o = e_i^2$,

Also $e_o = e\, G_a$,

hence $e/e_i = e_i/G_a$,

that is $G_{Ev} \propto e_i$ ——(2.15)

For a required expansion ratio 2:1, the transfer function of the variable-loss device varies directly as the received signal.

3rd step – Design of the rectifier.

The control current is directly proportional to the level of the signal at the input to the rectifier.

$$I_{con} \propto e_1 \qquad \text{——(2.16)}$$

Therefore from (2.15)

$$G_{Ev} \propto I_{con} \qquad \text{——(2.17)}$$

4th step – Refer to Fig 4.

The transfer function

$$G_{Ev} = \frac{e_2}{e_1} = \frac{i_2\, R_2}{i_1\, R_1} \qquad \text{——(2.5a)}$$

From equations (2.5a), (2.17) it can be seen that G_{Ev} will vary directly with input current if R_1 has small dependence on R_2 and i_2/i_1 is approximately constant. This can be done by using any configuration of transistor circuit connected between the generator and the load resistance. Since i_2/i_1 is the current amplification and it is nearly constant with load resistance,

therefore $$G_{Ev} \qquad \frac{1}{R_1} \qquad \text{———(2.18)}$$

5th step

From equations (2.17) and (2.18) it is necessary to select a transistor configuration of which its input resistance varies inversely as the control current. From Fig 6, 7,8,9,10,11, the common-base configuration is the most suitable one.

6th step

By the same reason as stated in consideration of the compressor variable-loss device, the expander variable-loss circuit is connected in push-pull as shown in Fig 13.

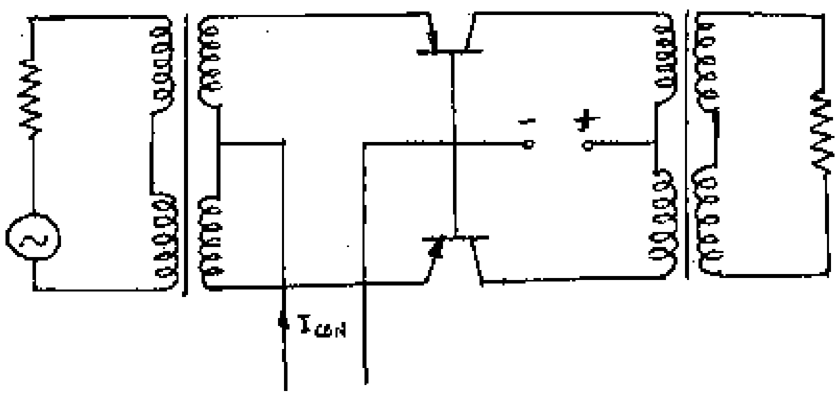


Fig 13 — Expander variable-loss device

From the 5th step of designs of the compressor and expander variable-loss devices, the a.c. input resistance of the emitter-base function varies inversely as the control current as shown in Fig 14.

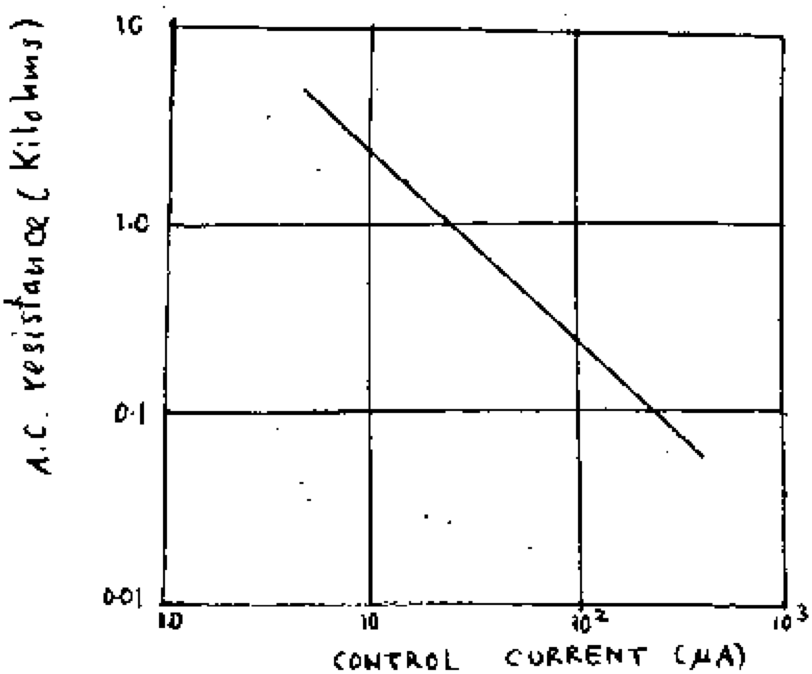


Fig 14 — A.C. input resistance/control current characteristic.

2.5 Control - Current Rectifier

From the third step of design of the compressor and expander variable-loss devices, the control current is directly proportional to the level of the signal at the input to the rectifier. Fig 15 shows the four terminal active-network of the rectifier.

Fig 15 — Rectifier regarded as four-terminal network.

Rectifier is any device which has a high resistance to current in one direction and a low resistance to current in the opposite direction possesses the ability to convert an a.c. current into a current which contains a d.c. components in addition to a.c. components. An ideal rectifier would be one with zero resistance in the forward direction and with an infinite resistance in the reverse direction. From the knowledge of rectifier principles it is seen to be necessary to use the transistor configuration having the lowest input resistance and the highest output resistance. Consider the three basic transistor configurations common-base configuration having the lowest input resistance and the highest output resistance, therefore the common-base type is selected. The rectification action can be achieved by connecting it as a class B amplifier as shown in Fig 16. (Please see in the conclusion, the reason why using the transistor as rectifier).

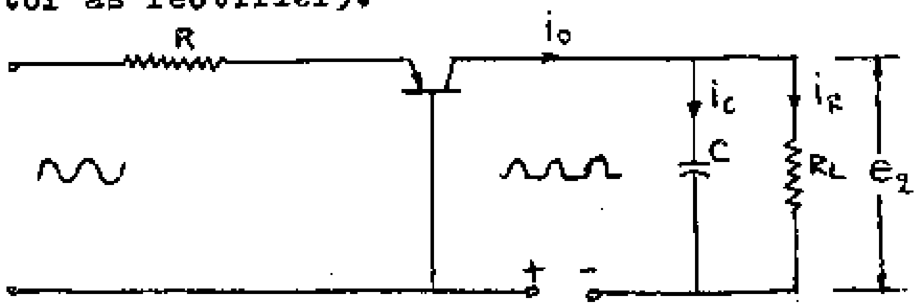


Fig 16 -- Control current rectifier

The resistance R connected in series with the emitter of the rectifier is to increase its output impedance. Filtering is frequently effected by shunting the load with a capacitor to improve the ripple factor.

2.6 Amplifier

2.6.1 Introduction

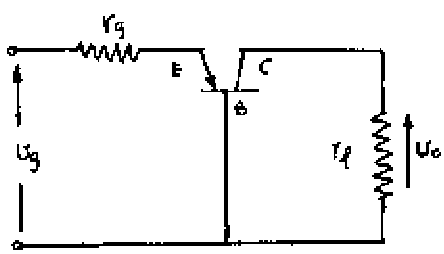
The three common configurations are known as

(1) grounded - base

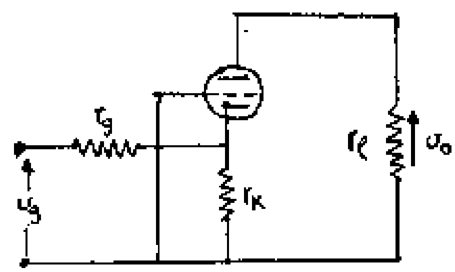
(2) grounded-emitter, and

(3) grounded-collector

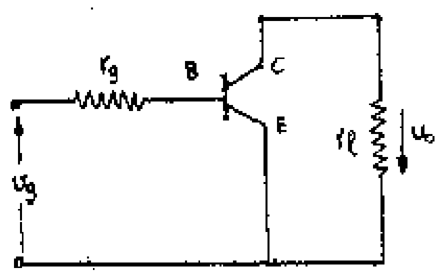
Comparing these configurations with vacuum tubes, they correspond roughly with (1) grounded-grid, (2) grounded cathode, and (3) grounded-plate. These are illustrated in Fig 17. As will be shown later, by adding a feed back path in the vacuum-tube circuits it is possible to obtain very close tube analogues to the transistor configurations.



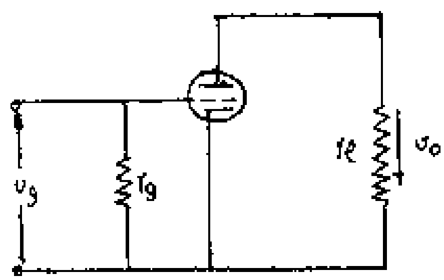
(a) Grounded-base



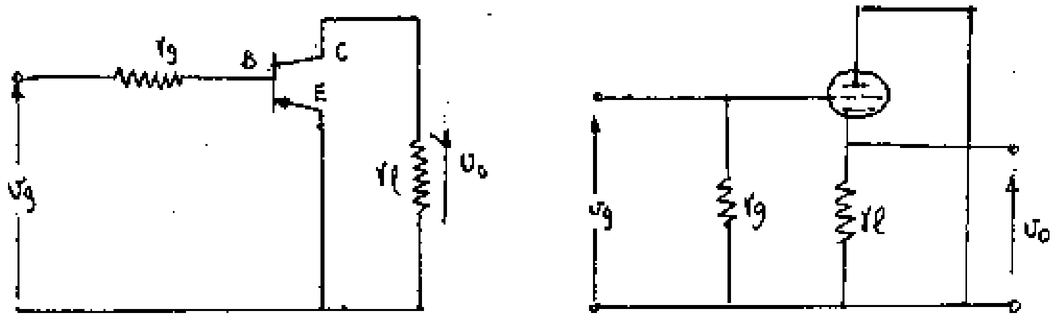
Grounded-grid



(b) Grounded-emitter



Grounded-cathode



(C) Grounded - Collector   Cathode - follower

Fig 17 — Corresponding configurations of transistor and vacuum tube.

Some general conclusione can be drawn from the study of three transistor configurations.

(1) The grounded-emitter configuration provides phase inversion; the grounded-base and grounded-collector do not.

(2) The input resistance for one transistor operating at a given fixed load will be lowest for grounded-base, intermediate for the grounded-emitter, and largest for the grounded-collector configuration.

(3) The output resistance for one transistor operating from a generator resistance r_g, the same for all configurations, will be highest, intermediate, and lowest for grounded-base, grounede-emitter and grounded-collector configurations respectively.

(4) The voltage amplification for a given load resistance will be almost identioal for grounded-base and

grounded-emitter, and will be some what less than unity for the grounded-collector.

(5) The current amplification approximates a for grounded-base, approximates b for the grounded-emitter (for a = 0.96, b = 24), and approximates b+1 for the grounded-collector.

(6) In almost all case of interest the power gain of the grounded-emitter is highest, the relative power gain of the other configurations depending on the size of the load resistance r_l.

The above comments will illustrate the single - stage performance; multi stage performance will be taken up as follow. In general, a junction transistor amplifier will consist of several stages coupled by passive networks. The coupling networks must be designed so that they will not materially reduce the gain of the amplifier in the desired frequency range.

It is customary to characterize the individual stages of an amplifier by their current and voltage amplification or by their power gain. Such a characterization is useful only if the amplification or gain is practically independent of the next stage. Fig 18 shows qualitatively the current amplification A_i and the voltage amplification A_v of the grounded-emitter single-stage transistor amplifier as a function of the load resistance r_l. Since in an RC or direct-coupled amplifier, the input resistance of the next stage will usually be considerably less than $r_o (1 - a)$, we will characterize individual stage in an RC or direct-coupled amplifier by their current amplification. The total gain of an amplifier with n stages is, then,

$$G = (r_1/r_i)\,(A_{i1} \cdot A_{i2} \cdots\cdots\cdots A_{in})^2 \qquad \text{———}(2.19)$$

Where A_{i1} is the current amplification of the first stage, A_{i2} that of the second stage, etc; r_i; the input resistance of the amplifier; and r_1 the load resistance.

In amplifiers with coupling by matching transformers; the load resistance of each stage is equal to its output resistance and is therefore independent of the next stage. Individual stages in such a transformer coupled amplifier may therefore be characterized by their voltage or current amplification or by their power gain.

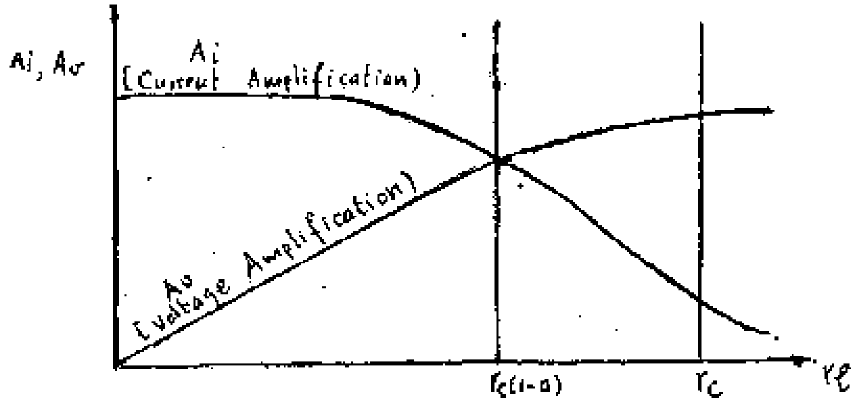


Fig 18 —— Current and Voltage Amplification

2.6.2 The Gain of an RC Coupled Amplifier

For the calculation of the multistage amplifiers, expressions are needed for the input resistance and current amplification of individual stages. We will consider only grounded-emitter here.

Input Resistance

$$r_i = r_b + r_e \cdot \frac{r_c + r_l}{r_c - r_m + r_e + r_l} \qquad \text{—(2.12a)}$$

Current Amplification

$$A_i \doteq \frac{a\, r_c - r_e}{r_c(1 - a) + r_e + r_l} \qquad \text{—(2.20)}$$

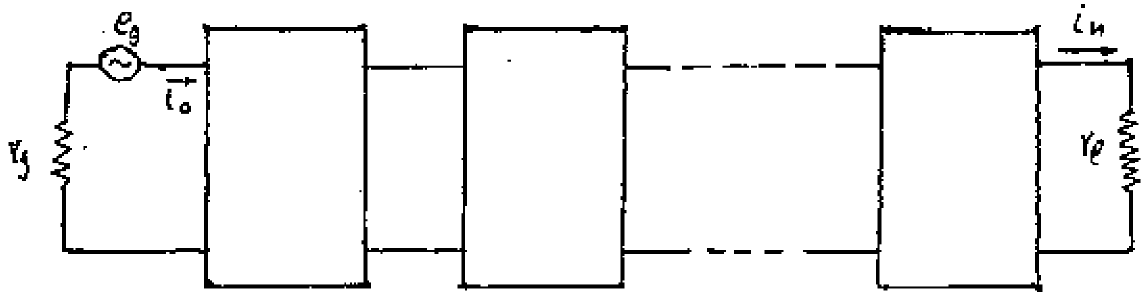


Fig 19 —— Multistage Amplifier

Fig 19 shows a block diagram of a multistage amplifier. The transducer gain G_t of this amplifier is

$$G_t = \frac{\text{Power delivered to the load}}{\text{the power available at the input}} \qquad \text{—(2.21a)}$$

$$= \frac{i_n^2 \cdot r_l}{e_g^2/4r_g} = \frac{4\, r_l \cdot r_g}{(r_g+r_i)^2} \frac{i_n^2}{i_o^2}$$

$$= 4\, \frac{r_g \cdot A_{i1}^2}{(r_g + r_i)^2} (A_{i2} \cdot \ldots \cdot A_{in-1})^2\, r_l \cdot A_{in}^2 \quad \text{—(2.21b)}$$

Where A_{i1}, A_{i2} and A_{in} are the current amplifications of the first, second and nth stages. r_i is the input resistance of the amplifier. In order to obtain a maximum transducer gain, the right-hand side of eq.(2.21b) must be a maximum; and therefore the following expressions must, individually, be as large as possible.

(a) First stage:

$$\frac{r_g}{(r_g+r_i)^2} \cdot Ai^2 = Y \text{ (Y has the form of a conductance)}$$

(b) Intermediate stages: Ai (2.22)

(c) Last stage: $r_l \cdot A_i^2 = R$ (an effective resistance)

Fig 20 shows an amplifier containing grounded-emitter stages only. All coupling network, etc. have been omitted.

The transducer gain of this amplifier is; with good approximation,

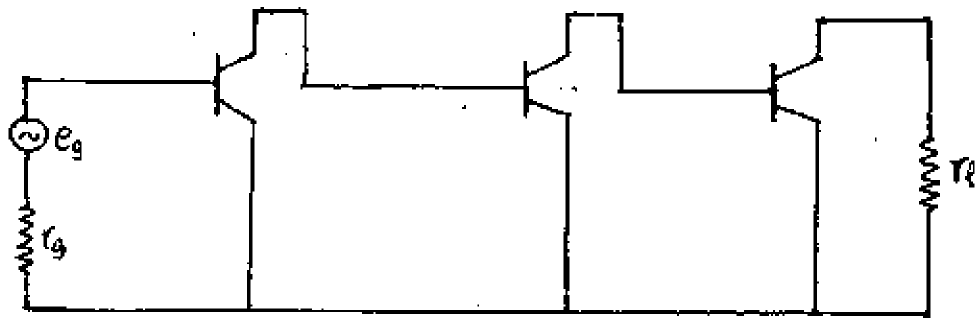


Fig 20 —— Three – stage amplifier

$$G_t = 4\frac{r_l r_g}{(r_g + r_{11})^2} \cdot \left(\frac{a_1}{1 - a_1} \cdot \frac{a_2}{1 - a_2} \cdot \frac{a_3 r_{c3}}{r_{c3}(1 - a_3) + r_l}\right)^2 \quad (2.23)$$

If both source and load are matched to the amplifier we have

$$r_g = r_{b1} + \frac{r_{e1}}{1 - a_1} \quad ;$$

$$r_l = r_{c3}(1 - a_3)$$

whence

$$G_t = \frac{a_1^2\, a_3^2\, r_{c3}}{4(1-a_1)(r_{e1} + r_{b1}(1-a_1))(1-a_3)} \left(\frac{a_2}{1-a_2}\right)^2 \quad (2.24)$$

This is the maximum available power gain that may be obtained with a three – stage transister amplifier with capacitive or direct coupling.2

2.6.3 The caloulation of coupling network

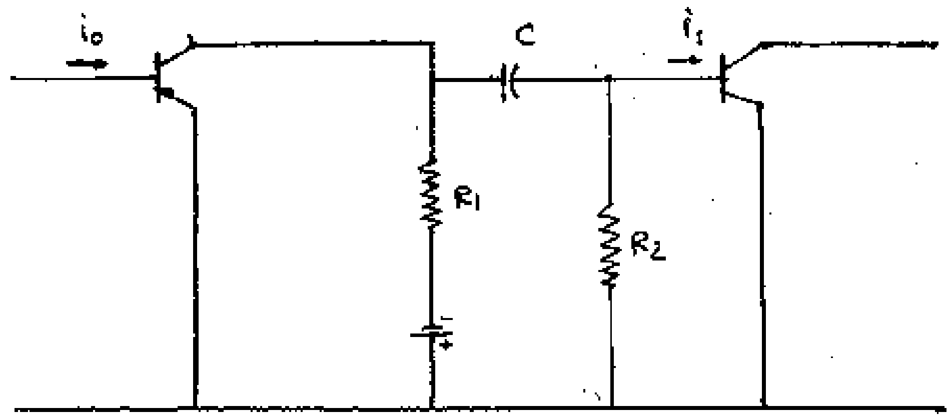


Fig 21 --- Simplified coupling network.

Fig 21 shows a simplified coupling network containing only two resister R_1 & R_2, and the capacitor C. It is first assumed that at the frequency of interest the impedance of the capacitor can be neglected. We have then for the current flowing into the second transistor:

$$i_1 = i_0 \frac{a}{1 - a} \cdot \frac{R}{R + r_i} \qquad \text{-------(2.25)}$$

Where $R = R_1R_2/(R_1 + R_2)$ is the resistance presented by R_1 and R_2 in parallel and r_i is the input resistance of the second stage. The reduction in current amplification due to R is

$$K = R/(R + r_1) \qquad \text{-------(2.26)}$$

Because of the capacitance of C, the amplification at low frequency is reduced. At low frequencies, the impedance of this capacitance is much larges than r_i, and we have therefore

$$A_i = \frac{i_1}{i_b} \cdot \frac{a\, r_c}{r_c(1-a) + r_1} \cdot \frac{R_1}{R_1 + (1/jwC)}$$

$$\approx \frac{a\, r_c}{r_c(1-a) + (R_1/1 + jwCR_1)} \cdot \frac{R_1}{R_1 + (1/jwC)}$$

$$\approx \frac{a\, r_c\, R_1}{r_c(1-a)R_1 + [R_1 + r_c(1-a)]/jwC} \qquad \text{-----(2.27)}$$

This equation shows that A_i is reduced by 3db if

$$\frac{1}{wC} = \frac{r_c(1-a)\, R_1}{R_i + r_c(1-a)} \qquad \text{\_\_\_\_(2.28)}$$

2.6.4 Feedback Amplifiers

Feed back in amplifiers improve such properties as stability under varying power supply conditions; stability to parameter variation, freedom from non-linearities and their attendant modulation effects. We will now show a few of the methods of applying feed back to transistor amplifiers, and analyze them very briefly. The grounded emitter circuit is considerably more useful for transistor applications, and emphasis will be placed on this type of configuration.

2.6.4a Shunt – Type Feedback (see Fig 22)

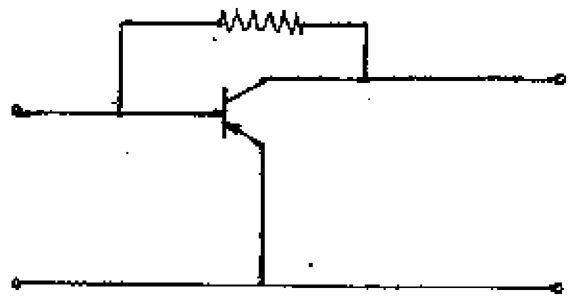


Fig 22 — Shunt feed back

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. The behavior of this circuit is very similar to that of the grounded-cathode vacuum tube with plate-to-grid feedback. For such a circuit we have, as a result of the feedback, a reduction in gain and input impedance as well as in output impedance.

Since the current ratio of a grounded-emitter amplifier is approximately

$$A_i \qquad \frac{a\, z_c}{z_c(1-a)+r_1} \qquad \text{———(2.29)}$$

its locus is of the same shape as that for $\alpha/1-\alpha$. The effect of r_1 is to lower the effective for the circuit. If the feedback is resistive, the $-A^{\circ}\beta$ diagram is, qualititatively, as shown in Fig 23. Where A° is vector current ratio;

β is the percentage of the output current of the amplifier that is fed back to the network, and less than 1.

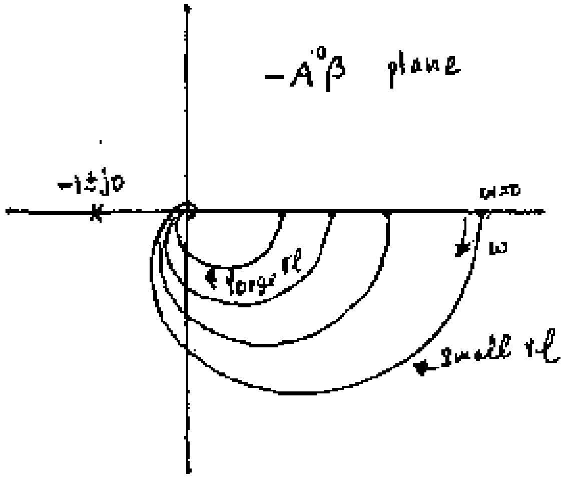


Fig 23 — $\mathring{A}\beta$ diagram for the circuit of Fig 22

2.6.4b Series - Type Feedback (see Fig 24)

This type of feedback is very similar to that employed with vacuum tube with unbypassed cathode resistor. It results in increased input and output impedances, and decreased gain. With such a circuit we cannot separate the $\mathring{A}$ and β terms, and so an elementary theory of feedback is insufficient in determining its stability.

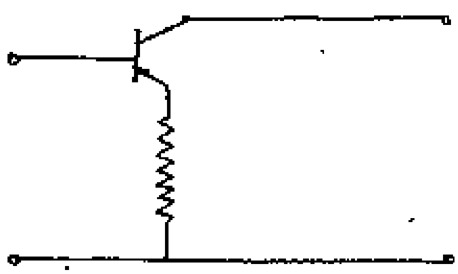


Fig 24 — Series Feedback

2.6.4c Multistage Series or Shunt Feedback

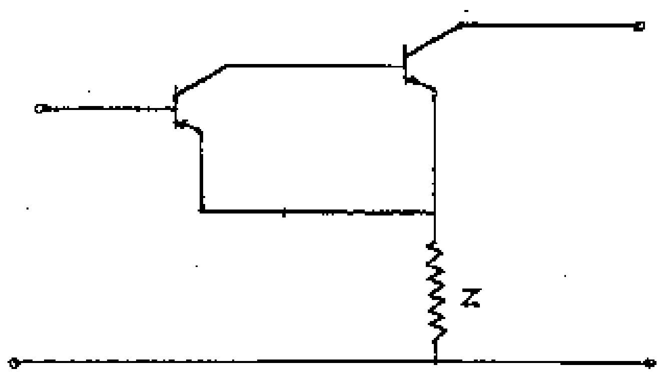


Fig 25 — Multistage Series Feedback

The number of combinations of the above circuits is legion, and it serves no purpose to sketch more than one of each type. Fig 25 shows a two - stage amplifier with series feedback, which resembles a cathode - coupled vacuum - tube amplifier in its behavior.

Fig 26 shows a multistage amplifier with shunt feedback. The behavior of such amplifier resembles that of a multistage vacuum - tube amplifier with the added complication of the phase shift of $\alpha/1-\alpha$ outside the mid band. The behavior of R_c is of secondary importance in that this term usually appears in both numerator and denominator, and its effects tend to cancel.

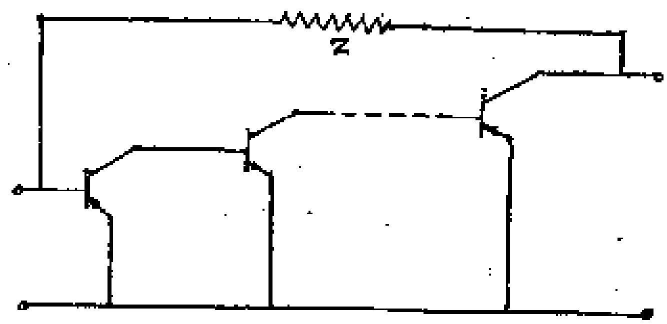


Fig 26 —— Multistage shunt feedback.

2.7 Compandor

2.7.1 Compressor

Fig 27 shows the circuit of the compressor. The transistor VT1 provide temperature compensation, The variable loss device consists of VT2, VT3. The high - gain amplifier, consisting of transistor VT4, VT5 and VT6, provides two outputs. The normal signal output is obtained via transformer T3, and an output to supply the control current rectifier circuit is provided via transformer T4. The primary windings of T3 and T4 are fed in series in the high - impedance collector circuit of the output transistor VT6, and this ensures that the amount of coupling between output and control circuits is sufficiently small.

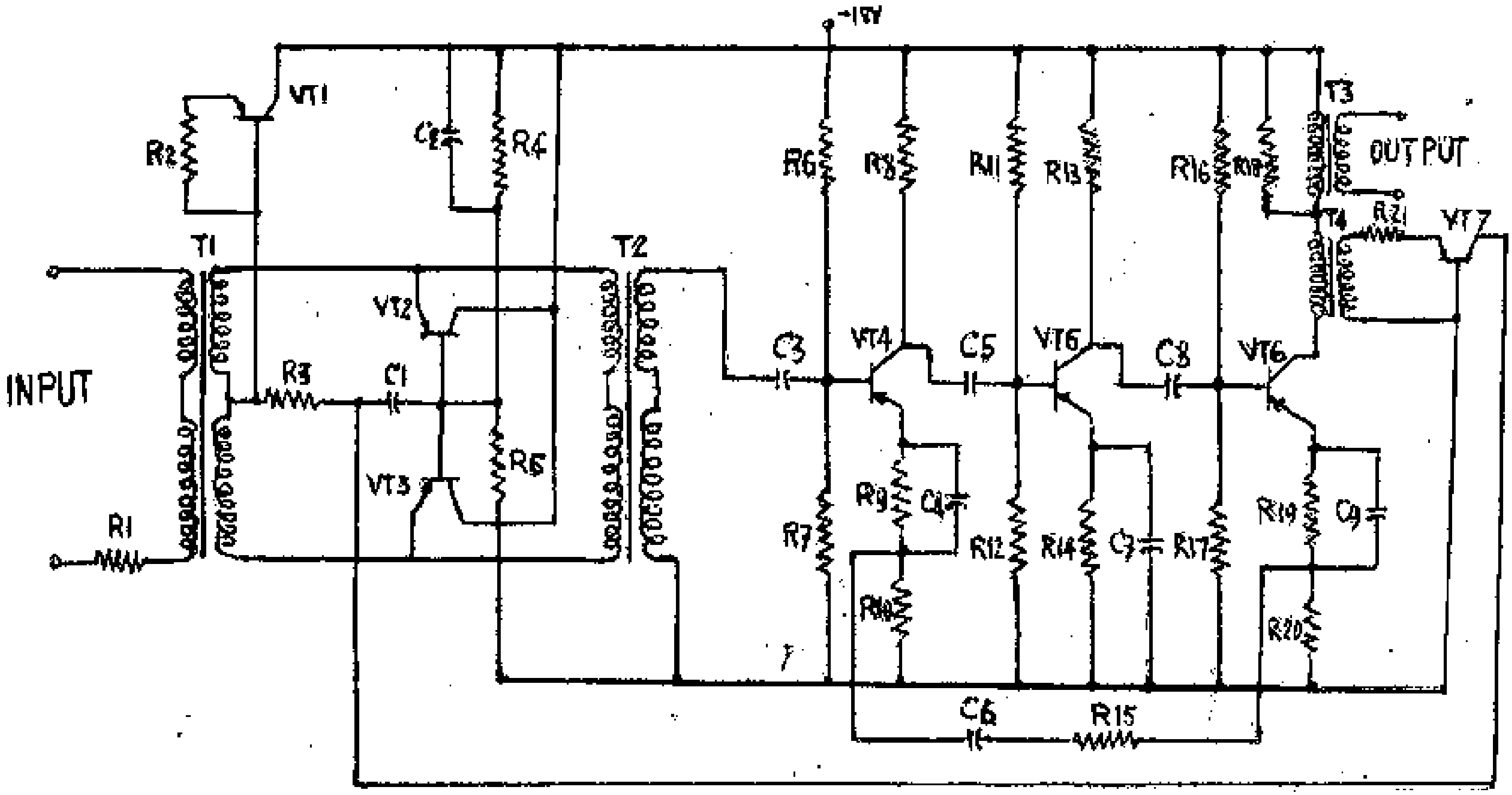


Fig.--27- Circuit diagram of compresser.

2.7.2 Expander

Fig 28 shows the circuit of the expander. The input signal is divided, by means of R_1, R_2 and R_6, to supply an input to the signal path via transformer T1, and an input to the control-current rectifier VT3 after amplification in the single-stage pre-amplifier VT1. The variable-loss device is followed by RC coupled amplifier. The negative feedback applied by means of R_{21}.

2.8 Noise Reduction

The compandor will improve the subjective signal/ noise ratio of a circuit only when the interfering signal is injected between the compressor output and the cxpander input. This is the condition when the compandor is situated in the terminal equipment of the carrier system and the interfering signal are line noise and crosstalk.

The gain of both the compressor and the expander is controlled by the rectified envelope of speech signal on the line, a reduction of gain in the oompressor being accompanied by an equal increase in the gain of the expander. Fig 29 is a simplified level diagram for a carrier system containing a oompandor. With no signal input, the compressor introduces 30 db gain and the expander 30 db loss relative to the condition at the test level. Thus, any line noise receive 30 db attenuation within the expander,

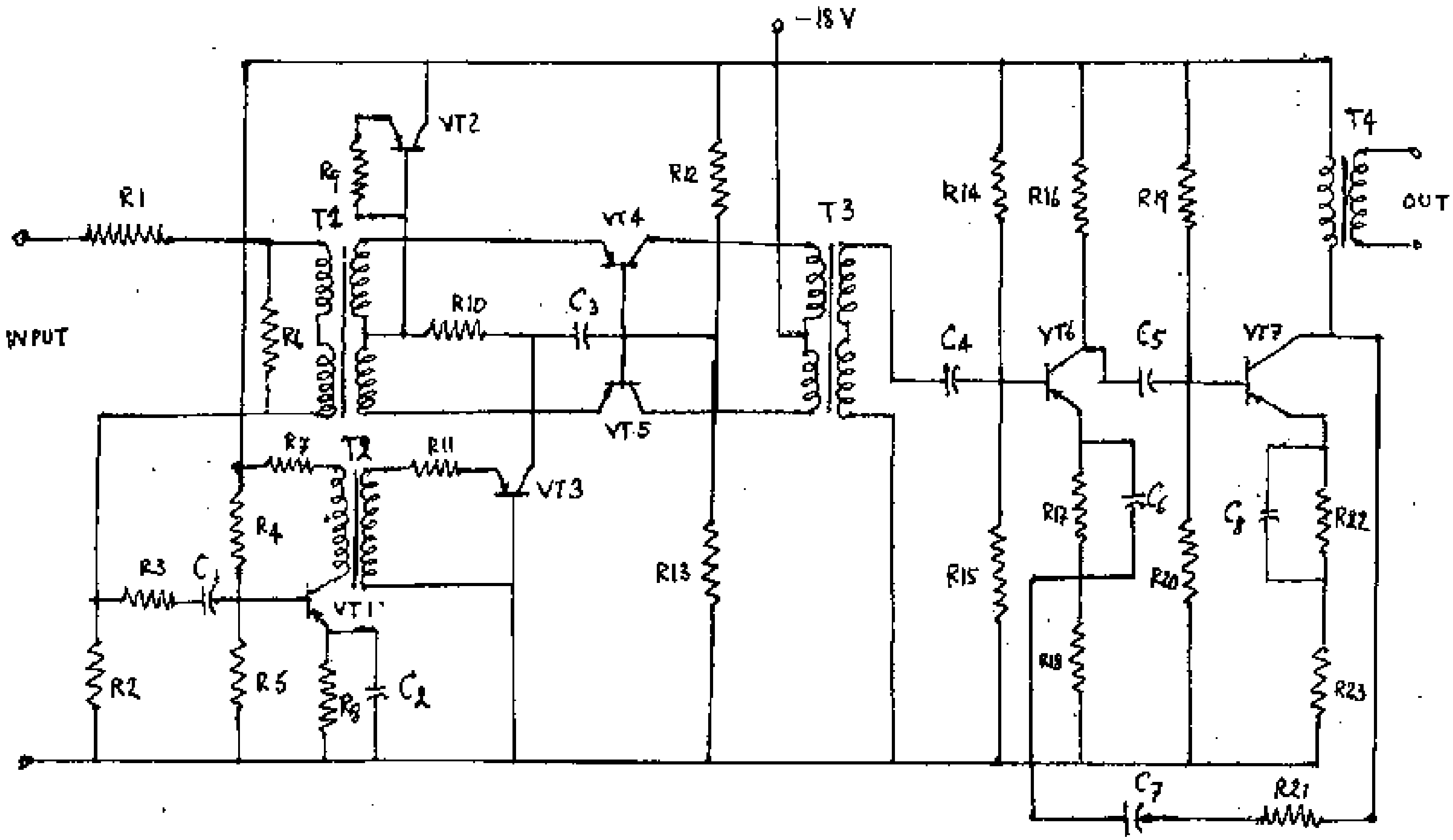


Fig. — 28 — Circuit diagram of expander.

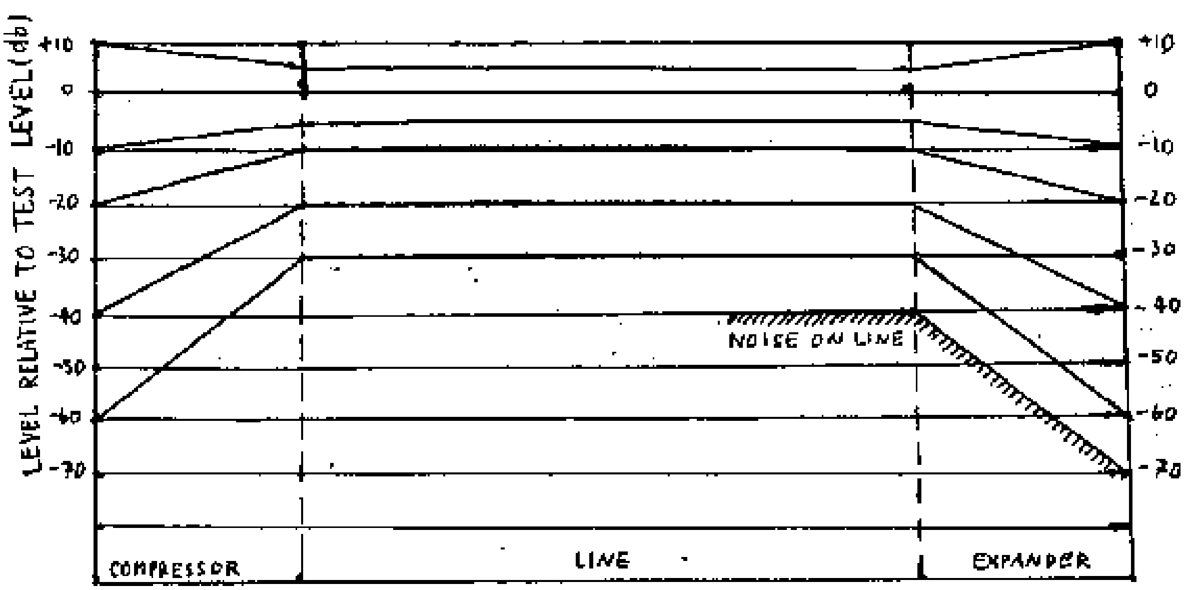


Fig 29 —— Level diagram for a carrier system containing a compandor.

provided that the level of the noise is 30 db or more below test level, as is normally the case with an input signal to the compressor of, say, -30 db relative to test level and a noise level of -40 db relative to the test level on the line, the compressor introduces 15 db gain and thus improves the signal/noise ratio on the line from 10 db to 25 db. The expander, which is controlled by the higher level signal, has a loss of 15 db, so that the noise now received only 15 db attention with the expander, but when the signal ceases, the loss to the noise rapidly increases to 30 db so that the noise at the output of the expander falls to-70 db relative to test level, and would be inaudible.

Compandor action may be illustrated as shown in Fig 30, where the input signal (a) represents discrete sound of various levels separated by silent intervals, e.g. separate words or syllables. The signals after compression are shown at (b), where the first aspect of compandor advantage may be seen in the improvement of signal/noise ratio on the line for all signals below test level. The effect of the expander is seen at (c), where the original signal level has been restored and the received noise is dependent on the signal level the noise is greatest when the signal is loudest, and least during the interval between words. It is this effect of a silent back-ground which greatly contributes to the subjective improvement produced by the compandor, this is particularly so if the interfering signal is intelligible crosstalk.

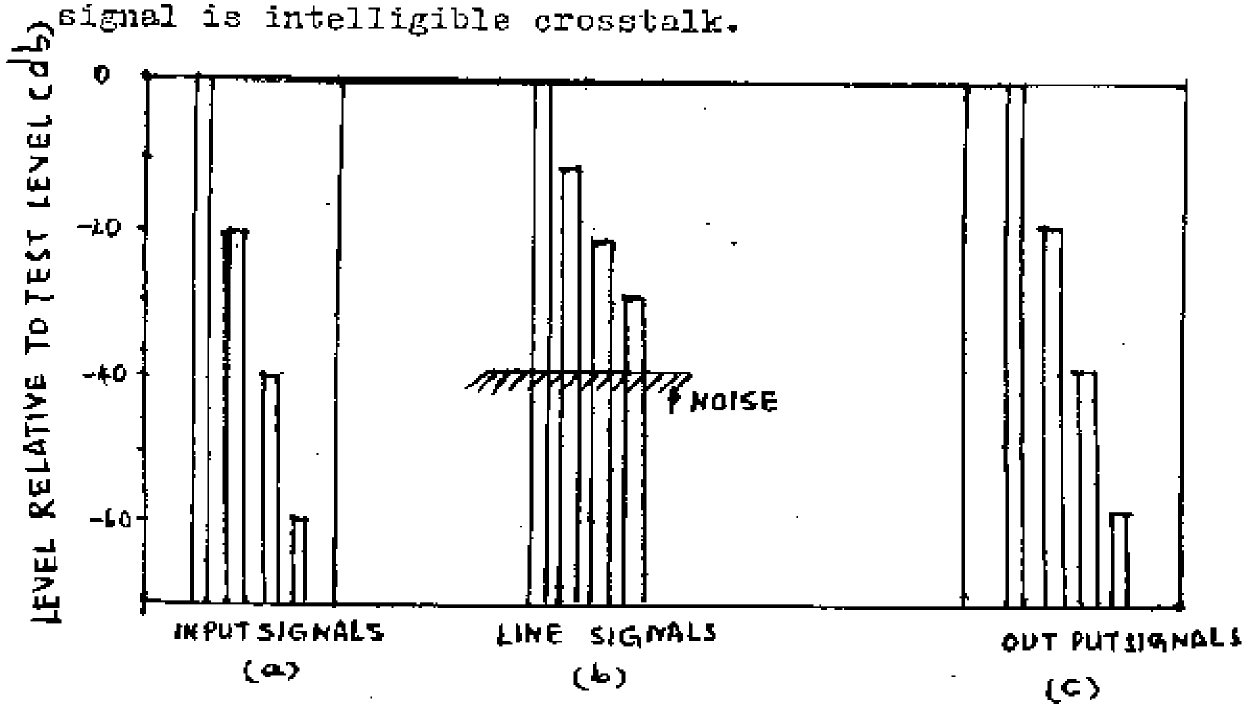


Fig 30 -- Compandor response to discrete signal

In practice the compandor is not instantaneous in action, as suggested by Fig 30, but both compressor and expander having time constants which are chosen from consideration of the characteristics of telephone speech.