Scherer Marc
2/01/69

INSTITUT D'AERONOMIE SPATIALE DE BELGIQUE

3, avenue Circulaire, UCCLE - BRUXELLES 18

AERONOMICA ACTA

A - N° 23 - 1963

Low and very low level DC amplifiers (Part I)
Theory (I)

by P. VILLE

BELGISCH INSTITUUT VOOR RUIMTE-AERONOMIE

3, Ringlaan, UKKEL - BRUSSEL 18

FOREWORD

In Ref.[61] it is stated that "The ability to process these low level d-c voltages to a range suitable for transmission is one of the major problems of modern telemetry".

This text is an attempt to bring together in a clear and orderly manner the basic information about the theory and the design of low level and very low level d-c amplifiers. Two such d-c amplifiers were built and their performance is discussed.

The text is subdivided into five parts :

I. Theory (I),
I.A.S, Aeronomica Acta A - N° 23 - 1963.

II. Theory (II),
I.A.S, Aeronomica Acta A - N° 24 - 1963.

III. Modulators and demodulators,
I.A.S, Aeronomica Acta A - N° 31 - 1964.

IV. A modulated d-c amplifier for microvolt signals,
I.A.S, Aeronomica Acta A - N° 32 - 1964.

V. Literature and References.
I.A.S, Aeronomica Acta A - N° 33 - 1964.

Part I and II deal with the basic theory of d-c amplifiers proper. The types of modulators and demodulators used in modulated d-c amplifiers are discussed in Part III. In Part IV we take up the design of a d-c amplifier with characteristics (performance, weight, size, power requirements,...) suitable for space applications. Finally Part V contains the abstracted references to which we refer in the text.

M. Nicolet.

AVANT-PROPOS

Dans la référence[61], on note que : "La possibilité d'adapter ces basses tensions continues à un domaine adéquat pour la transmission est un des principaux problèmes de la télémesure moderne".

Ce texte est un essai pour rassembler, sous une forme claire et ordonnée, les informations fondamentales concernant la théorie et l'utilisation des amplificateurs de tensions continues de faibles et de très faibles niveaux.

Le texte est divisé en cinq parties :

I. Theory (I),
I.A.S, Aeronomica Acta A - N° 23 - 1963.

II. Theory (II),
I.A.S, Aeronomica Acta A - N° 24 - 1963.

III. Modulators and demodulators,
I.A.S, Aeronomica Acta A - N° 31 - 1964.

IV. A modulated d-c amplifier for microvolt signals,
I.A.S, Aeronomica Acta A - N° 32 - 1964.

V. Literature and References.
I.A.S, Aeronomica Acta A - N° 33 - 1964.

Les deux premières parties se rapportent à la théorie fondamentale des amplificateurs d-c. Les types de modulateurs et de démodulateurs utilisés dans les amplificateurs d-c modulés sont discutés dans la partie III. L'utilisation d'un amplificateur d-c pour les applications spatiales ainsi que les caractéristiques (performance, poids, forme, puissance, exigences,...) sont discutées dans la partie IV. Finalement, la partie V contient les références citées dans le texte ainsi que leurs résumés.

M. Nicolet.

VOORWOORD

In Ref. [61] wordt gezegd dat "Het beheersen van de technieken die nodig zijn om deze zwakke gelijkspanningen om te zetten in signalen die kunnen overgeseind worden één van de grootste problemen is van de moderne telemeting".

Deze tekst is een poging om op een klare en ordelijke wijze de grondgegevens samen te brengen betreffende de theorie en het ontwerpen van gelijkstroomversterkers voor zwakke en zeer zwakke signalen. Twee zulke gelijkstroomversterkers werden gebouwd en hun eigenschappen worden besproken.

De tekst is onderverdeeld in vijf delen :

I. Theory (I),
I.A.S, Aeronomica Acta A - N° 23 - 1963.

II. Theory (II),
I.A.S, Aeronomica Acta A - N° 24 - 1963.

III. Modulators and demodulators,
I.A.S, Aeronomica Acta A - N° 31 - 1964.

IV. A modulated d-c amplifier for microvolt signals,
I.A.S, Aeronomica Acta A - N° 32 - 1964.

V. Literature and References.
I.A.S, Aeronomica Acta A - N° 33 - 1964.

Deel I en II behandelen de basistheorie van de eigenlijke gelijkstroomversterker. De types van modulatoren en demodulatoren, die gebruikt worden in gemoduleerde gelijkstroomversterkers, worden besproken in deel III. In deel IV handelen we over het ontwerpen van een gelijkstroomversterker met eigenschappen (gewicht, afmetingen, voedingsvereisten,...) die hem geschikt maken voor ruimtetoepassingen. Deel V eindelijk bevat de referentiën met korte inhoud, naar dewelke we in de tekst verwijzen.

M. Nicolet.

VORWORT

In Referenz[61] steht geschrieben dass : "Die Möglichkeit dieser schwachen d-c Spannungen zu einem Gebiet nützlich für die Übertragung zu verwenden, ist eines der wichtigsten Problemen der moderne Fernmessung".

Dieser Text ist ein Versuch, um die Grundinformationen über die Theorie und die Benützung der d-c Verstärker für schwachen und sehr schwachen Spannungen in einer klaren und geordneten Weise vorzustellen.

Der Text besteht aus fünf Teilen :

I. Theory (I),
   I.A.S, Aeronomica Acta A - N° 23 - 1963.

II. Theory (II),
   I.A.S, Aeronomica Acta A - N° 24 - 1963.

III. Modulators and demodulators,
   I.A.S, Aeronomica Acta A - N° 31 - 1964.

IV. A modulated d-c amplifier for microvolt signals,
   I.A.S, Aeronomica Acta A - N° 32 - 1964.

V. Literature and References.
   I.A.S, Aeronomica Acta A - N° 33 - 1964.

Die zwei ersten Teile haben Bezug auf die Grundtheorie der d-c Verstärker. Die verschiedenen Modulatoren und Demodulatoren die in modulierten d-c Verstärker gebraucht werden, sind im dritten Teil diskutiert. Die Verwendung eines d-c Verstärker für Raumforschung sowie die technischen Daten (Leistung, Gewicht, Form, Kraft, Anforderung,...) sind im vierten Teil diskutiert. Der fünfte Teil enthält die im Text angegebenen Referenzen sowie die Zusammenfassungen.

M. Nicolet.

CONTENTS OF PART I.

LOW AND VERY LOW LEVEL DC AMPLIFIERS (Part 1)

THEORY (1)

by

Paul VILLE

Centre National de Recherches de l'Espace,
3, Avenue Circulaire,
UCCLE - BRUXELLES 18.

CHAPTER 1

D-C AMPLIFICATION IN GENERAL.

1.- Generalities.

Sometimes people classify amplifiers into two general types : d-c amplifiers and a-c amplifiers. However this is only a convenient way to make some kind of distinction. Indeed (ideal) d-c amplifiers constitute a much broader class of amplifiers than do (ideal) a-c amplifiers. The purpose of an (ideal) d-c amplifier is to tell us at a certain time t_1 what the absolute level is at that time of the applied input signal above some arbitrary chosen but known zero level. It is evident therefore that (ideal) d-c amplifiers can also be used to inform us about the change of the input signal between two time instants t_1 and t_2. The term "d-c amplifier" is primarily used to denote that the amplifier is also able to amplify signals of zero frequency or "direct current signals". This does not inherently mean that such amplifiers cannot be used for higher frequency signals.

A-c amplifiers on the contrary are only meant to amplify changes in the input signal level : they will let us know whenever the input signal changes, by how much it changes and in which sense (positive or negative). Clearly also d-c amplifiers tell us all this. Why then ever use so-called a-c amplifiers? The answer is simple : a-c amplifiers do not tell us anything about the absolute (or d-c) level of the input signal, while it is the desire of including the accurate knowledge of this absolute level that causes a lot of particular problems in the design of d-c amplifiers; so, if we do not care about that d-c level, there is no reason why we should allow it to give us trouble.

In short then : if we want to know the absolute (or d-c) level of the input signal we have to face particular problems. From this came the desire to classify amplifiers in two broad classes :

- d-c amplifiers, in which special care is taken to solve as well as possible the d-c problems as mentioned above;
- a-c amplifiers, which simply ignore the absolute level of the input signal and so do not have to worry about the problems involved in reproducing it.

Let us note that purposely we have not spoken here about amplitude and phase characteristics of the amplifiers in consideration. Our main purpose was to distinguish between d-c and a-c amplifiers in a general way. It is true however that practical d-c amplifiers have a rather limited bandwidth. Most of the better procedures used to make d-c amplifiers indicate the absolute level of the input signal quite accurately also limit the useful bandwidth of the amplifier. We will come back to this point when we discuss the different types of d-c amplifiers. It may be useful however, to note that it is possible to design a wide-band d-c amplifier by using some special compensating methods as we will see later.

What are the particular problems we have to face in the design of d-c amplifiers ? Although we will later come back to these problems we will indicate here briefly what they consist of in order that the reader may clearly know what we are really speaking about.

The particular problems showing up in the design of d-c amplifiers are the ones related to zero stability. By "perfect zero stability" is meant that at any time zero input should give zero output. This however cannot be attained in practice: Even with zero input we measure a certain output. The input then that we have to apply in order to return the output to zero is called the "offset". If now this offset were constant in time the only problem would be to compensate for it. Unfortunately it is not constant in time and its change per unit time is called the "drift".

The reasons of zero offset and drift and several methods used to reduce them will be discussed later.

The d-c amplifier parameters whose values usually have to meet desired specifications are [75] :

(a) The input resistance to d-c.
(b) The overall d-c gain.
(c) The output resistance to d-c.
(d) The offset and the drift of the amplifier.

We shall therefore have to pay special attention to these parameters. A discussion of them will be taken up in chapter 3.

The four general classifications of d-c amplifiers are [5] :

(a) simple cascaded d-c amplifiers.
(b) modulated d-c amplifiers.
(c) compensated d-c amplifiers.
(d) bridge-balanced d-c amplifiers.

These will be discussed and compared with each other in the chapters 4, 5, 6 and 7.

In order to give the reader an idea of the performance attainable with actual d-c amplifiers, it may be useful to briefly mention what d-c level can be measured using present techniques.

D-c amplifiers not using modulation techniques are able to recognize d-c signals as low as 1 millivolt if the elements used are very good and if proper care is taken to reduce offset and drift : proper matching of transistors in the first stage of a d-c transistorized differential amplifier is an example.

D-c amplifiers using modulation methods have, up till now, still much better performance. In particular electromechanical choppers are very good in this respect and (as far as we know) all the d-c amplifiers recognizing as low as 1 to 10 microvolts are using them at their input. The range has recently been moved down to as low as 0.1 microvolt (or 100 nanovolt)[108] and some people even claim to have built a d-c amplifier with full-scale deflection of 0.1 microvolt*.

2.- Vacuum tubes versus transistors.

Before proceeding to the next chapter we will make a brief comparison between vacuum tubes and transistors in order to be able to determine which of them are better to use in a particular application.

1. First a basic difference between vacuum tube circuits and transistor circuits should be noted : vacuum tubes are essentially voltage sensing devices and transistors are current activated devices. Hence a vacuum tube essentially measures a voltage and a transistor measures a current. However it is possible to measure currents by means of vacuum tube devices if we measure the voltage drop across a known resistance inserted in the circuit of the current to be measured. Likewise a voltage can be measured by means of transistor devices if that voltage is allowed to give rise to a current which is measured by the device.

Since it is usually preferable to measure voltages and not currents we will emphasize voltage measurements as well for vacuum tube devices as for transistorized ones in this text.

2. Most vacuum tubes require power supplies with a voltage in the range between 100 and 500 volts whereas transistors work on much lower voltages (voltages from 6 to 30 volts are typical). The relatively high voltage that vacuum tubes require may hence be an objection against their use especially where the power has to be supplied by batteries.

---

* Keithley Instruments 1961-62 Catalog.

3. Vacuum tubes need a heater supply for the filaments whereas transistors do not need anything of this kind at all.

4. By virtue of 2. and 3. and also because their quiescent current is usually higher than for transistors, vacuum tubes need more power to work than transistors do.

5. Vacuum tubes are larger in size and are heavier than transistors.

6. The quiescent operating point of vacuum tubes depends upon the plate supply, the grid bias and the heater temperature.
The quiescent operating point of transistors depends upon the collector supply, the base bias (for common-emitter circuits), and the temperature of the transistor. It is noted that transistors are much more temperature dependent than vacuum tubes are : in particular the leakage current of transistors is very dependent on the temperature (it approximately doubles for each ten degrees C rise), and so is their current amplification factor and the base-to-emitter contact potential (which changes about 2.5 mV per degree C).
By comparing transistors with each other, however, we note that silicon transistors have much lower leakage current than germanium transistors, so that the performance of the former will be much better for temperature variations than the performance of the latter. Indeed for all applications over a more or less wide temperature range silicon transistors are preferred.

CHAPTER 2

NOISE IN D-C AMPLIFIERS.

Let us first note the difference between noise, offset and drift as it may perhaps not always be clear what is meant by each of them.

Noise is any signal that results in an unwanted output. Offset in a linear d-c amplifier is the amount of output, as referred to the input, that is due to sources other than the desired input signal (If the input signal is e_i, the output signal is e_o and the gain is A_1, then the offset as referred to the input is $e_i - (e_o / A_1)$. Note that without offset we have $e_o = A_1 e_i$). Drift is the change of the offset per unit time.

From the definitions above it is seen that noise results in offset but not all offset is due to noise : Indeed, some of it is due merely to the change of the characteristics of the elements used.

Noise then is any signal that results in an unwanted output. What now are the causes of noise in a d-c amplifier ? This question is not easy to answer as there are very many noise sources in every system and the noise generated is due to internal causes (Johnson noise, chopping noise, tube noise, transistor noise, ...) as well as to external ones (magnetic fields, electric fields, vibrations and shocks, ...). In d-c amplifiers the noise sources become worthy of consideration only when other causes of offset and drift (shift in operating point, variation in gain,...) are minimized because the latter are usually at least one order of magnitude higher than the former.

If them it is worthwhile to consider reduction of the noise itself then we can distinguish between :

1. Noise generated by external sources and by the particular set-up of the system. This noise can usually be reduced by taking special care when designing and building the circuit. It is due to electrical and magnetic fields, to mechanical vibration or shocks or to thermal effects. Since we have not spoken yet about the

circuits used for d-c amplification we will not take up here the discussion of the different types of this noise as mentioned above. We will however come back to it when we discuss the different d-c amplifier systems in use. In particular, we will thoroughly investigate the causes of such noise and the means to reduce it in chapter 5 where modulated d-c amplifiers will be dealt with. In these amplifiers the problem is notably rather acute because they are able to give very good performance (so that noise is to be avoided as much as possible) and because they inherently add some new sources of noise (modulator, driving source, ...).

2. Noise generated by internal sources. By this we mean noise originated in the resistors (Johnson noise), vacuum tubes (shot noise, flicker noise, partition noise, secondary emission noise, ...), transistors (partition noise, shot noise, 1/f noise,...), diodes (shot noise, flicker noise, ...) or other elements of the circuits. Obviously this noise should be reduced (if it is worthwhile to do so) by use of the proper elements or by filtering.
We are not going to give here a detailed discussion of this type of noise either. We pointed out above what it consists of and we will come back to the problem whenever necessary. Let us just point out here that worthwhile information about the causes of noise and the means to reduce it can be found for example in Ref.[1], [10], [20] for resistor noise, Ref.[11], [20], [35] for vacuum tube noise and Ref.[72], [77], [81], [89], [96], [98], [100], [107], [109] for diode and transistor noise.

CHAPTER 3.

GAIN STABILITY AND ZERO STABILITY IN D-C AMPLIFIERS.

INPUT RESISTANCE AND OUTPUT RESISTANCE OF D-C AMPLIFIERS.

As stated before all d-c amplifiers have some amount of zero offset and drift. Furthermore, due to variations in the characteristics of the elements used, the gain is not constant either. We will not discuss here the causes of zero offset and drift nor the causes of gain instability. These are indeed dependent upon the amplification system used (direct-coupled amplifiers, modulated amplifiers,...) and will be discussed when the different systems are discussed. We will however deal here with the general problems of zero and gain stability and at the same time we will point out some general means to reduce the two forms of instability.

1.- Gain Stability.

To stabilize the gain in d-c amplifiers negative feedback is used. The principle is as follows : We have a certain input and a certain output and we want the latter to be exactly A" times the former. Through an attenuation network we attenuate the output A" times and compare the resulting signal with the input. If the input is higher, it means that the output is less than A" times the input. If the input is lower, it means that the output is higher than A" times the input. In any case we use the difference of input and attenuated output to increase or decrease the latter until the input and the attenuated output are equal: in that case the output is exactly A" times the input.

How do we realize this? Let us for example consider the idealized system of Fig. 1 in which input and output are assumed to be both voltage signals. The input is in series with the opposite of the attenuated output. The input to the d-c amplifier is :

$$e_e = e_i - \frac{e_o}{A''}$$

Note that A" is assumed to be > 0 in order that we have negative (and not positive) feedback. Since the gain of the d-c amplifier is assumed to be positive and infinite any input to the amplifier will result in an infinite output. Let us assume that e_e is positive, hence $e_i > e_o/A''$. Then it follows that $e_o = +\infty$. In that case however $e_i > e_o/A''$ is impossible. Assuming e_e to be negative results in $e_i < e_o/A''$ and $e_o = -\infty$ that is not possible either. The only possibility is then :

$$e_e = e_i - \frac{e_o}{A''} = 0$$

wherefrom it follows that

$e_o = A''e_i$ as anticipated.

Let us now consider the actual real case where the gain of the amplifier is not infinite but is equal to A. Let us assume the attenuation to be given by β in order to use commonly used symbolism and let us denote the overall gain of the d-c amplifier with feedback by A'. Then we get Fig. 2. For the configuration shown it is assumed that the product Aβ is negative in order that we have negative or so-called degenerative feedback.

Note that by "Linear d-c amplifier" is meant a d-c amplifier for which the output is a linear function of the input. Or in our case :

$$e_o = A.\, e_e$$

Since

$e_e = e_i + \beta\, e_o$ we get

$$e_o = A(e_i + \beta\, e_o)$$

and after solving for e_o :

$$e_o = \frac{A}{1 - A\beta} \cdot e_i = \frac{1}{-\beta + \frac{1}{A}} \cdot e_i$$

If A = + ∞ as for Fig. 1 then it turns out that

$e_o = -\frac{1}{\beta} \cdot e_i$. But in Fig. 1

$\beta = -\frac{1}{A''}$ and we get our

formely found result that for A = + ∞

$$e_o = A'' . e_i$$

For A = - ∞ we find the same result :

$$e_o = -\frac{e_i}{\beta}$$

In this case however β should be positive in order that the product Aβ be negative. Also this result is evident for let us assume in Fig. 2 that $e_e = e_i + \beta\ e_o$ is positive. Hence $e_i > - \beta\ e_o$. Then e_o is - ∞ and $e_i > - \beta\ e_o$ is impossible since $- \beta\ e_o = + \infty$ as β is a positive quantity. Similarly for $e_e = e_i + \beta\ e_o < 0$ we get $e_o = + \infty$ but then $e_e = e_i + \beta\ e_o = e_i + \infty$ should be > 0. Hence the only possible solution is again : $e_e = 0$ wherefrom :

$e_o = -\frac{e_i}{\beta}$ which is equivalent to $e_o = A''. e_i$ in Fig. 1.

In general then we find for Fig. 2 :

$$e_o = \frac{A}{1 - A\beta} \cdot e_i = \frac{1}{-\beta + \frac{1}{A}} \cdot e_i$$

in the assumption that the feedback is negative (i.e. $A.\beta < 0$).

Let us now calculate the relative variation of e_o for variations of the forward gain A of the d-c amplifier. With :

$$e_o = \frac{A}{1 - \beta A} \cdot e_i$$

we find :

$$\Delta e_o = \frac{A + \Delta A}{1 - (A + \Delta A)\beta} \cdot e_i - \frac{A}{1 - \beta A} \cdot e_i$$

$$= \frac{\Delta A}{[1 - A\beta].[1 - (A + \Delta A)\beta]} \cdot e_i$$

and

$$\frac{\Delta e_o}{e_o} = \frac{1}{1 - (A + \Delta A).\beta} \cdot \frac{\Delta A}{A}$$

We see that the relative variation of the output is a factor $1 - (A + \Delta A).\beta$ less than the relative variation of the forward gain A.

Noting that in this case the total overall gain of the d-c amplifier is

$$A' = \frac{e_o}{e_i} = \frac{A}{1 - A.\beta}$$

we see that

$$\frac{\Delta A'}{A'} = \frac{\Delta e_o}{e_o}$$ (because e_i is supposed to be constant in these calculations)

$$= \frac{1}{1 - (A + \Delta A)\beta} \cdot \frac{\Delta A}{A}$$

The conclusion is that the overall gain A' of the feedback system is a factor $1 - A\beta$ less than the gain A of the system without feedback but at the same time the relative stability of the gain A' is better than the relative stability of the gain A by a factor $1 - (A + \Delta A).\beta$.

It is seen also that

$$\Delta A' = \Delta\left(\frac{A}{1 - A\beta}\right) = \frac{\Delta A}{(1 - A\beta).[1 - (A + \Delta A)\beta]}$$

We shall then want to make A as high as possible so that in first approximation

$$A' = \frac{A}{1 - A\beta} = \frac{1}{-\beta + \frac{1}{A}} \cong -\frac{1}{\beta}$$

$$\text{for } |\beta| \gg \left|\frac{1}{A}\right| \text{ or } |A\beta| \gg 1$$

and $\Delta A' \rightarrow 0$ for variations of A. As β usually consists of passive elements (resistors and sometimes capacitors) its numerical value can be made very stable.

We see that by use of negative feedback it is possible to make the overall gain of the d-c amplifier almost entirely independent of the forward gain of the amplifier, provided the latter is made very high such that $|A\beta| \gg 1$. This is a general result for feedback amplifiers.

2.- Zero stability.

First it is to be noticed that no amount of gain stability obtained through feedback will stabilize the zero. Feedback neither helps nor hinders zero stability. The general equation for any ideal amplifier is:

Input x gain = output.

For consideration of a disturbing input to the amplifier, this equation should be rewritten as follows :

(Input wanted + input disturbing) x gain = output

From this equation it is clear that in the presence of a disturbing input (noise, drift,...) and with the wanted input (signal) at zero, the output cannot be reduced to zero by stabilizing the gain.

Disturbances may enter at points other than at the input terminals (for example, after the first stage of amplification), but they are generally measured in terms of their equivalence at the input terminals. Let us for example consider Fig. 3. Let e_i be the input signal, e_o the output, and e_d and e_f two disturbing signals resp. at the input of the overall amplifier (e_d) and somewhere between input and output (e_f). $A = A_1.A_2$ is the overall forward gain of the amplifier. Let us calculate e_o as a function of e_i, e_d, e_f, A_1, A_2 and β :

It is seen that :

$$e_o = A_2 . e_g$$

$$e_g = e_f + A_1 . e_e$$

$$e_e = e_i + e_d + \beta . e_o$$

Hence

$$e_o = A_2 \left[e_f + A_1 \left(e_i + e_d + \beta \, e_o \right) \right]$$

wherefrom we can solve for e_o :

$$e_o = \frac{1}{1 - A_1 A_2 \beta} \cdot \left[A_2 e_f + A_1 A_2 \left(e_i + e_d \right) \right]$$

Since $A = A_1 A_2$ is the total forward gain of the amplifier we get :

$$e_o = \frac{A}{1 - A\beta} \cdot \left[e_i + e_d + \frac{e_f}{A_1} \right]$$

Obviously we see that the input disturbance e_d has exactly the same effect as the useful signal e_i. However the effect of e_f is the smaller according as it comes in farther away from the input.

For practical purposes now and in order that we may be able to compare amplifiers with each other we assume that all disturbances come in at the input. (This is only for measurement and comparison purposes). For the example considered above we then say that the equivalent disturbance as referred to the input is

$$e_d + \frac{e_f}{A_1}$$

The usefulness of this is that it tells us what the useful signal is that can be detected in the presence of the disturbing signals. If for example it turns out that $e_d + e_f/A_1$ is 1,000 times smaller than the signal to be amplified then the presence of the disturbing signals will not have much effect on the relative accuracy of the amplifier. If however $e_d + e_f/A_1$ is of the same order as the signal to be amplified then it can turn out that the amplifier is completely useless.

So for all practical purposes all the disturbances coming in anywhere in the amplifier are referred to the input.

From the equation :

$$e_o = \frac{A}{1 - A\beta} \left(e_i + e_d + \frac{e_f}{A_1} \right)$$

it is seen that the harmfulness of e_f is reduced by a factor A_1 as compared with a disturbance of the same value occuring at the input. This suggests that we should try to make A_1 as great as possible and reduce e_d as much as we can. In other words : the first stage or possibly the first stages of the amplifier are the most important ones whenever reduction of the influence of unwanted signals is desired and so we should make them as noiseless and unaffected by disturbing signals as possible (low e_d) while at the same time we should make their gain factor as high as possible (high A_1).

For simplicity it was assumed above that the nature of the disturbing signals e_d and e_f was similar to the desired signal e_i, so that the effect of the amplifier upon them is the same as its effect upon the signal e_i. If this is not true (for example if the gain of the amplifier is dependent on the frequency and if the frequencies of the signals considered are not the same) then the effect of the disturbances upon the output is assumed to result from disturbing signals which are of the same nature as the input signal e_i and hence the disturbing signal $e_d + e_f/A_1$ as dealt with above is in fact an equivalent one in this case : it has the same nature as the signal e_i and it affects the output in the same way as the real physical disturbances do.

3.- Input resistance and output resistance.

The input resistance to d-c of a d-c amplifier is defined to be the ratio of input voltage and input current to the amplifier. For vacuum tubes the input current is usually low (only the grid current) because they are voltage devices. Hence the input resistance of vacuum tubes for d-c signals is essentially high. Transistors however will usually have a much lower input resistance for d-c signals because their working is based on the input current so that the latter must have a non negligible value. If the amplifier input is the input to a vacuum tube or a transistor then the input resistance to the amplifier will, of course, be the input resistance to that vacuum tube or that transistor. (Note that for transistors the input resistance is not independent of the load of the transistor).

Sometimes (for current measurements) the input of the vacuum tube which constitutes the input of the amplifier is shunted by a resistor. In that case the input resistance to the amplifier is the parallel combination of the resistance of this resistor and the input resistance of the tube.

The output resistance to d-c of a d-c amplifier is defined to be the negative of the ratio of an infinitesimal change of the voltage across the load and the corresponding infinitesimal change of the current through the load when the latter is varied infinitesimally itself (Fig. 4):

$$R_{output} = -\frac{dE_{Load}}{dI_{Load}}$$

where the variations dE_{Load} and dI_{Load} are assumed to be due only to a load variation dR_{Load} (and not to an input or gain variation).

This output resistance will, of course, depend upon the particular output circuit used.

Note that the load resistance R_{Load} itself is defined by

$$R_{Load} = \frac{E_{Load}}{I_{Load}} \quad \text{and}$$

does not have to be a physical resistance. Any variation then in the external output circuitry that changes the ratio $\frac{E_{Load}}{I_{Load}}$ is considered to change the "load resistance".

<u>Effect of negative feedback upon the input and the output resistances.</u>

Let us use R_{id} and R_{od} for the d-c input and output resistances of the d-c amplifier without feedback while R_{if} and R_{of} will denote the corresponding d-c input and output resistances of the d-c amplifier with negative feedback.

Input resistance.

1) Circuit for voltage measurements

We assume as a simple example the circuit of Fig. 2 redrawn here as Fig. 5. It is seen from Fig. 5 that

$$R_{id} = \frac{e_e}{i_i}$$

What we want to know however is

$$R_{if} = \frac{e_i}{i_i} \text{ as a function of } R_{id}.$$

This is easily calculated :
Indeed

$$e_e = e_i + \beta . e_o = e_i + \beta . (Ae_e)$$

wherefrom :

$$e_e = \frac{1}{1 - \beta . A} \cdot e_i$$

It then follows that

$$R_{if} = \frac{e_i}{i_i} = \frac{(1 - \beta . A)e_e}{i_i} = (1 - \beta . A) . R_{id}$$

Since $1 - \beta.A$ is always rather large it is seen that the d-c input resistance with feedback is considerably increased as compared to the same input resistance without feedback. A high input resistance is highly desirable for voltage measurements so that this type of circuit is certainly recommandable for voltage measurements.

2) Circuit for current measurements.

We now consider for example Fig. 6.

Obviously

$$I = \frac{e_i - \beta . e_o}{R}$$

$$i_e = \frac{e_i}{R_{id}}$$

Hence

$$i_i = i_e + I$$

$$= \frac{e_i}{R_{id}} + \frac{e_i - \beta\ e_o}{R}$$

$$= \frac{e_i}{R_{id}} + \frac{e_i - A\beta\ e_i}{R}$$

$$= \left[\frac{1}{R_{id}} + \frac{1}{R} (1 - A\beta) \right] . e_i$$

Then the d-c input resistance of the d-c amplifier with feedback is :

$$R_{if} = \frac{e_i}{i_i} = \frac{1}{\frac{1}{R_{id}} + \frac{1}{R} (1 - A\beta)}$$

$$= \frac{R}{R + R_{id} (1 - A\beta)} . R_{id}$$

Since $R < [R + R_{id} (1 - A\beta)]$ it is seen that now R_{if} is smaller than R_{id} and can even be much smaller if $1 - A\beta$ is large as usually is the case.

Since current measurements require a low input resistance it is clear that circuits of the kind dealt with above will be appropriate for such measurements.

It is seen that $e_o = A.e_i$. This shows that the relative variation of e_o equals the relative variation of A if only the latter is supposed to be varying :

$$e_o = A.e_i$$

$$de_o = dA.e_i$$

wherefrom :

$$\frac{de_o}{e_o} = \frac{dA}{A}$$

However this is consistent because we do not want to measure the input voltage e_i but rather the input current i_i, i.e. e_o should be a measure of i_i. We see that

$$e_o = A.e_i$$

As we saw above we also have :

$$i_i = \left[\frac{1}{R_{id}} + \frac{1}{R} (1 - A\beta) \right] . e_i$$

wherefrom by combining the two equations we get :

$$e_o = \frac{A}{\frac{1}{R_{id}} + \frac{1}{R} (1 - A\beta)} \cdot i_i$$

Then for a variation dA of A :

$$\frac{de_o}{dA} = \frac{\frac{1}{R_{id}} + \frac{1}{R} (1 - A\beta) + A \frac{\beta}{R}}{\left[\frac{1}{R_{id}} + \frac{1}{R} (1 - A\beta) \right]^2} \cdot i_i$$

$$= \frac{\frac{1}{R_{id}} + \frac{1}{R}}{\left[\frac{1}{R_{id}} + \frac{1}{R}(1 - A\beta)\right]^2} \cdot i_i$$

wherefrom follows that

$$\frac{de_o}{e_o} = \frac{\frac{1}{R_{id}} + \frac{1}{R}}{\frac{1}{R_{id}} + \frac{1}{R}(1 - A\beta)} \cdot \frac{dA}{A}$$

Again we get a form which shows the stabilizing effect of the feedback loop :

from

$$\frac{de_o}{dA} = \frac{\frac{1}{R_{id}} + \frac{1}{R}}{\left[\frac{1}{R_{id}} + \frac{1}{R}(1 - A\beta)\right]^2} \cdot i_i$$

we see that $\frac{de_o}{dA}$ decreases as A increases and in the limit

$$\frac{de_o}{dA} \rightarrow 0 \text{ for } A \rightarrow \infty$$

In general feedback stabilizes the overall gain factor and

1) increases the d-c input resistance in the case of the voltage measurement configuration.
2) decreases the d-c input resistance in the case of the current measurement configuration.

Output resistance.

Although for the input resistance we considered high as well as low values according to whether we wanted to measure voltages or currents we will not do so for the output resistance because in applications such as the ones we are considering here a high output resistance is never wanted.

We again consider as a first example Fig. 5 which for convenience is redrawn as Fig. 7 wherein a load R_L is included.

We assume that the input resistance R_β of the feedback device is so high that the current the device drains is negligible with respect to the current through the load. This assumption only simplifies the calculations but if it is not satisfied the input resistance R_β can easily be included in the calculations as we will show later.

Furthermore it should be noted that the gain A is dependent upon the load. This is due to the fact that the output resistance of the d-c amplifier is not zero. Let us denote by A_∞ the d-c amplifier gain for $R_L = \infty$. Then it follows that

$$A = \frac{R_L}{R_L + R_{od}} \cdot A_\infty$$

where R_{od} is the output resistance of the amplifier without feedback.

It is also important to note that β is dependent on the source resistance R_s, the input resistance R_{id} of the d-c amplifier without feedback, and the output resistance of the β-device.

The equivalent Thevenin circuit of the d-c amplifier output for a variation de_o of the output voltage e_o is shown in Fig. 8. From this figure it is easy to calculate the corresponding variation di_o of the output current i_o : it turns out to be :

$$di_o = \frac{+ \beta . A_\infty \cdot \frac{R_{id}}{R_s + R_{id}} \cdot de_o - de_o}{R_{od}}$$

wherefrom the output resistance R_{of} for the d-c amplifier with feedback is found :

$$R_{of} = - \frac{de_o}{di_o} = \frac{R_{od}}{1 - \beta . A_\infty . \frac{R_{id}}{R_s + R_{id}}}$$

In many cases R_{id} is rather large and R_s is rather small, so that:

$$R_{of} \cong \frac{R_{od}}{1 - \beta . A_\infty}$$

Usually then the overall output resistance R_{of} will be very small if we make $|A_\infty . \beta|$ very large.

As a next example we consider Fig. 6 which is redrawn as Fig. 9. With the same notations as before and with again $R_\beta = \infty$ we will also here calculate the output resistance R_{of} of the d-c amplifier with feedback.

The equivalent Thevenin circuit of the d-c amplifier output for a variation de_o of the output voltage e_o is given in Fig. 10. It shows that the corresponding variation di_o of the output current is :

$$di_o = \frac{1}{R_{od}} \left(\beta . \frac{R'}{R' + R} . A_\infty . de_o - de_o \right)$$

where R' is the parallel combination of the resistances R_s and R_{id}.

It is seen that

$$R_{of} = - \frac{de_o}{di_o} = \frac{R_{od}}{1 - \beta . \frac{R'}{R' + R} . A_\infty}$$

$$= \frac{(R + R') R_{od}}{R + R' (1 - \beta A_\infty)}$$

Obviously the use of feedback again decreases the output resistance. This is very desirable.

What now if in any of the above examples the input resistance R_β to the feedback device is not infinite ? Then the real output resistance is easily found by again using Thevenin's theorem : First assume R_β to be infinite. Then the equivalent circuit of the output of the d-c amplifier with feedback is given in Fig. 11 where e_{of} is the output voltage for $R_L = \infty$. Since the resistance R_β is a resistance in parallel with R_L the equivalent circuit for the case when R_β is not very large with respect to R_L is as in Fig. 12 wherefrom using again Thevenin's theorem we get a new equivalent circuit as given in Fig. 13. From this figure we find the output resistance to be

$$\frac{R_{of} \cdot R_\beta}{R_{of} + R_\beta}$$

We shall also notice that now the gain will be given by :

$$A = \frac{R_L'}{R_L' + R_{od}} \cdot A_\infty$$

wherein R_L' is the parallel combination of R_L and R_β :

$$\frac{1}{R_L'} = \frac{1}{R_L} + \frac{1}{R_\beta}$$

and A_∞ is the gain for $R_\beta = R_L = \infty$.

The rather simple examples given above show that it is not very difficult to evaluate the effect of feedback upon the input and output resistances of the d-c amplifier. Although only two examples are discussed it is believed that they show the general way of dealing with feedback d-c amplifiers.

There are two other cases of feedback that we have not discussed :

1. The output current (instead of the output voltage) is used to get a voltage signal that is fed back to the input. (e.g. Fig. 14)

2. The output current (instead of the output voltage) is used to get a current signal that is fed back to the input (e.g. Fig. 15)

We will however not take up here a discussion of these cases because their characteristics can be found in a way analogous to the one we used in the discussion of the two other types of feedback.

4.- General conclusions concerning the use of negative feedback in d-c amplifiers.

In the sections above we have considered the influence of negative feedback upon the characteristics of the gain and the input and output resistances of d-c amplifiers in general. We have given complete calculations for some typical examples which are commonly used. It is true however that the results we obtained are in general true for all feedback circuits.

It turned out that the characteristics of the gain and the input and output resistances are much better with than without the use of negative feedback. This suggests that the forward gain of the amplifier be made as high as possible without disturbing the stability of the amplifier. Indeed the feedback makes the overall gain lower than the forward gain and the improvement in gain stability and input and output resistances is the better the greater the ratio is of the forward gain and the overall gain.

CHAPTER 4

SIMPLE CASCADED D-C AMPLIFIERS.

1.- Vacuum-tube circuits.

The simplest type of direct-coupled voltage amplifier[27] is shown in Fig. 16. This type of coupling has three major disadvantages :

1. The battery for coupling between stages has to be isolated and, if a large number of stages are required, the number of batteries for the coupling might be prohibitive.
2. The variations in the heater voltage will cause appreciable drift in the output with the input at a fixed voltage.
3. The variations in the anode supply voltage will also cause a drift.

The effects 2. and 3. can be reduced considerably by employing compensating methods as we will see in chapters 6 and 7 on compensated d-c amplifiers and bridge-balanced d-c amplifiers.

Disadvantage 1. can be eliminated by modifying the circuit a little so that succeeding amplifying stages [5] are stepped up a B supply bleeder (Fig. 17), or coupled with a potentiometer divider with consequent loss in gain (Fig. 18).

Although the circuits above are the simplest ones they are seldom used without some form of correcting or compensating system because (as stated above) they are essentially very notorious for drift.

2.- Transistor circuits.[69]

If it is assumed that all the drift originates in the input stage of the transistor amplifier, then this stage must have adequate voltage and current gain as we have seen in Chapter 3 under the heading "Zero stability". Of the three possible connections for a single transistor amplifier, common base, common collector, and common emitter, only the last gives both voltage and current gain, and so only this connection will be considered. As a matter of fact this connection is the most commonly used one for amplification purposes.

Figure 19 shows a simple common-emitter amplifying stage. A direct current, i_b, is extracted from the base by the negative input voltage through the source resistance R_s. The resulting collector current consists of $\beta\ i_b$, due to the base biasing current i_b, and $\beta i_{co} + i_{co}$, due to the collector leakage current i_{co}. Indeed by redrawing Fig. 19 using the current directions used there we get Fig. 20. Using the commonly used symbol α for the current gain in common base connection we see that $i_c = \alpha i_e + i_{co}$ as is explained in any elementary transistor textbook.

Furthermore

$$i_e = i_b + i_c$$

so that by elimination of i_e we get :

$$i_c = \alpha(i_b + i_c) + i_{co}$$

or when solving for i_c :

$$i_c = \frac{\alpha}{1 - \alpha}\ i_b + \frac{i_{co}}{1 - \alpha}$$

Now $\frac{\alpha}{1 - \alpha} = \beta$ so that

$$i_c = \beta i_b + (1 + \beta)\ i_{co}$$

$$= \beta(i_b + i_{co}) + i_{co}$$

as stated in Fig. 19

An increase in temperature will cause an increase in collector current i_c and hence a drift of the output voltage of the amplifier. This increase in collector current is primarily due to two main changes in the transistor characteristics :[78]

1. An increase Δi_{co} of i_{co}.
2. A decrease of the input resistance of the transistor and subsequently, if the input voltage is not changed, a decrease Δv_{eb} of the emitter-base voltage v_{eb}.

A third cause of drift is the variation of the d-c current transfer ratio.

Principally subsequent stages of a direct-coupled transistorized d-c amplifier are connected in a manner similar to the stages of vacuum-tube d-c amplifiers of which the basic connections are shown in Figs. 16, 17 and 18.

As stated for vacuum tubes, transistors are very rarely used in direct-coupled circuits without some form of correcting or compensating system. The reason for this is once more the notorious drift that uncompensated circuits exhibit. Compensating techniques will be discussed in chapters 6 and 7.

CHAPTER 5

MODULATED D-C AMPLIFIERS.

In Ref.[7] it is claimed that the idea of modulating D-C to A-C in order to amplify it was first suggested by R. Gunn in the "Review of Scientific Instruments" 9, 267 (1938) and by A.W. Sear in "Electronics" 13, 28 (January, 1940).

In this chapter we will deal with modulated d-c amplifiers as a whole, whereas the modulating systems are but a part of them and will be considered in Part III of this series on d-c amplifiers. So in this chapter the modulator itself will most of the time be considered as a building block whereas the emphasis will be on the amplifier using these blocks. In Part III however we will take up a detailed and comparative discussion of the several modulation systems themselves. Because of the importance of modulated d-c amplifiers such a separation seems to us to be highly justified.

The reason why we consider modulated d-c amplifiers before discussing the compensation methods used for direct-coupled d-c amplifiers is that some of the latter methods are based on the use of modulated d-c amplifiers. We therefore first discuss the modulated d-c amplifiers after which we will deal with the direct compensation methods for direct-coupled amplifiers as well as with the methods based upon modulation techniques.

A modulated d-c amplifier in its simplest form consists of an a-c amplifier preceded by some modulating system that converts d-c signals into a-c signals and followed by some other device that converts the amplified a-c signals back to d-c signals.

Let us consider Fig. 21, which gives the block diagram of a basic simple modulated d-c amplifier. The working is as follows : The input is a d-c signal but slow changes are not excluded (Later on in due time we will explain what "slow changes" mean). The input modulator then produces a signal that is periodic in time having an amplitude which is proportional to the amplitude of the d-c input signal. This signal behaves as a periodically changing signal, hence it is a so-called a-c signal and can therefore be amplified by the a-c amplifier. It follows that the output of the a-c amplifier consists of an a-c signal of which the amplitude is proportional to the amplitude of the d-c input signal. The detector then provides at its output a d-c signal of which the amplitude is proportional to the amplitude of the a-c signal at the output of the a-c amplifier. At the same time the set-up is such that the polarity of the output d-c signal corresponds with the polarity of the input d-c signal. For example the output polarity is always the same as or is always the opposite of the input polarity. Obviously the overall effect of the whole system is to give a d-c output of which the amplitude (or absolute level) is proportional to the amplitude (or absolute level) of the d-c input. This is exactly what we want to have because the whole system as a unit turns out to be nothing but a d-c amplifier.

The major advantage of this use of modulation techniques in order to amplify d-c signals is that we do not have to worry any longer about the inherent offset and drift of direct-coupled d-c amplifiers.

A question which will arise now is : what about the modulating and detecting systems ? Do they not introduce additional errors ? Indeed these devices introduce errors but the latter can be made so small as compared with the offset and drift errors which they help to eliminate that the overall performance of the system is considerably improved as compared with the performance of direct-coupled d-c amplifiers. (The errors of the modulating and detecting devices will be discussed in Part III in which the modulating systems themselves are considered).

1.- Modulator.

Although the general discussion of modulators will be given in Part III we will briefly consider here what they consist of in order to give the reader a clear understanding of what is happening.

Let us for example consider a switch as in Fig. 22, which is in turn connected to the input signal and to ground. The input signal and the corresponding signal at point A in Fig. 22 are shown in Fig. 23. It is assumed that one or the other of the two contacts is always closed but never the two at once. This is, of course, only a convenient assumption to make our basic analysis easier but cannot be attained in practice.

As seen in Fig. 23 the output of the modulator is a signal with periodic amplitude changes from zero to the actual value of the input signal and back to zero. The change of the chopped signal at the switching instants is therefore exactly the actual value of the amplitude of the input signal. This change is amplified by the a-c amplifier and appears at its output as an amplified signal change.

Other modulators perform the modulation in a somewhat different way but all of them produce some changing signal of which the amplitude is related to the amplitude of the d-c input signal.

2.- The A-C amplifier.

The purpose of the a-c amplifier is, as stated before, to amplify variations in its input signal. The wider the frequency bandwidth of the a-c amplifier is, the more nearly will the output resemble the input. Since a wider bandwidth means a lower gain a compromise should be considered between the resemblance of output to input and the available gain. In practice, however, it is not necessary that the output be a true replica of the input. The amplifier is therefore generally tuned to the modulating frequency or has a limited pass-band on either side of this frequency. This allows a high gain and has on the other hand the very desirable effect of reducing the noise of the amplifier. This noise is indeed spread out over a very large bandwidth and by tuning the amplifier so that it has a rather narrow pass-band (in practice a few cycles per second[27]) only the noise in this band is important and hence the equivalent noise of the amplifier is rather low (of the order of 1 microvolt or less).

3.- The output detector.

This device consists of a phase-sensitive rectifier in which the output voltage is negative or positive according to the sign of the input voltage.

This output detector can be either an ordinary diode rectifier (if the input signal is always of the same polarity) or a demodulator which is synchronized with the input modulator and may not be of the same type as the latter.

4.- Complete amplifier system.

Let us for example consider the system consisting of the chopping switch of Fig. 22, an ideal a-c amplifier of gain A for the entire frequency band (except for d-c) and a demodulator of the same type as the modulator and in synchronism with this one (Fig. 24). We assume that the upper contacts are closed simultaneously for the two switches. Note that the capacitor at the output of the system constitutes a so-called clamping device : if the upper contact of the output switch is not closed the voltage across the capacitor does not change. In Fig. 25 are given the signals that we find at the points labeled (1) (input), (2) (input of the a-c amplifier), (3) (output of the a-c amplifier) and (4) (output of the system). The dashed lines give the switching instants. Note that the signals at (2) and (3) are exactly proportional. This does not mean that the d-c component of the signal at (2) is transmitted through the a-c amplifier. Only differences are transmitted but since the signal at (3) is made zero by connecting (3) to the ground whenever (2) is connected to the ground then it turns out that differences at (2) with respect to the ground also appear as corresponding differences at (3) with respect to the ground. The latter are, however, A times larger than the former.

As we see the signal at (4) is not an exact replica of the signal at (1). This results from the fact that the assumed variations of the input signal are not slow enough with respect to the chopping frequency and also from the fact that we loose part of the signal by connecting input and output of the a-c amplifier to ground during one half of the chopping period.

In order to make the output look more like the input balanced input and output circuits can be used as for example in Fig. 26. The corresponding waveforms at several labeled points are shown in Fig. 27. We see that now the output is an exact replica of the input.

Through the action of the input switch the input is converted into an a-c signal that consists of a wave of amplitude equal to the amplitude of the input signal but with polarity reversal every other half period. During the rest of the half periods the polarity is opposite to that of the input signal. The output switch then flips the already flipped polarity back so that at the output a signal appears that, at every instant, is proportional to the input signal. In this case no clamping capacitor nor filter is necessary. However this is only a theoretical case and only occurs under the assumption that :

a) The switches take negligible time to change from one position to the other.
b) The switches are accurately synchronized.
c) The amplifier passes the wave without any distortion.

In practice, however, these ideal conditions cannot be obtained and a filter is required after the phase-sensitive rectifier.

5.- Gain stability by use of negative feedback.[10]

By "negative feedback" in modulated d-c amplifiers we mean overall negative feedback from the d-c output to the d-c input and do not include a-c feedback which may be used to stabilize the a-c amplifier.

As seen above the use of modulated d-c amplifiers is a good solution to the problem of zero stability. We do not have to worry any more about the offset and drift problems as far as the latter are due to direct coupling of d-c amplifier stages. This does not mean however that a modulated d-c amplifier as such is freed from all possible sources of errors. An other important source of error is left : the overall gain of the d-c amplifier. In order to stabilize the latter negative feedback is commonly used.

In the discussion of the simple modulated d-c amplifier we have seen that the d-c gain of the latter is either equal or in any case proportional to the gain of the a-c amplifier. Let us for the time being assume that the output of the overall system shown in Fig. 21 and redrawn in Fig. 28 is an exact replica of the input but A times greater. A is then called the overall gain of the d-c amplifier. It is dependent on the gain of the a-c amplifier and for all practical pruposes we may say that it is proportional to it. In any case we see that :

$$e_o = A.e_i \text{ (A = d-c gain)}$$

(assuming that offset and drift are zero). If now the gain of the a-c amplifier changes for any reason (e.g. because of a temperature change or because of aging of the amplifier elements) then the output also changes for a certain input :

$$\Delta e_o = e_i . \Delta A$$

(Δ denotes a change of the quantity which it precedes). The relative change of the output is then given by :

$$\frac{\Delta e_o}{e_o} = \frac{\Delta A . e_i}{A . e_i} = \frac{\Delta A}{A}$$

As seen the relative change of the output is equal to the relative change of the gain factor of the system. The latter can be rather high (10 °/₀ or even sometimes up to 100°/₀ or more -!) and such a relative variation of the output is certainly not allowed. To stabilize the gain negative feedback is used as discussed in chapter 3.

6.- Zero stability [10].

The use of d-c to a-c conversion so that a-c amplification may be used, is itself the greatest single factor in stabilizing the zero. This procedure avoids the conventional d-c amplifier, with its inherent zero offset and drift.

Unfortunately zero offset and drift due to direct coupling are not the only sources of zero disturbance. Others are :

1) Electric zero disturbances.
2) Magnetic zero disturbances.
3) Mechanical zero disturbances.
4) Thermal zero disturbances.

a) Electric zero disturbances [10].

Electric zero disturbances are defined as disturbances that are due to the action of voltages.

The changes in the tubes or transistors of a conventional d-c amplifier might be classified as electrical zero disturbances.

The signal as it passes through the a-c amplifier, is in the form an a-c signal of which the frequency is the chopping frequency of the modulators. Any signal then of this same frequency may disturb the zero. Of course, if it has an exact quadrature phase, it might not disturb the zero; but that possibility is remote, and the only safe practice is to exclude all disturbing voltages whose frequency is equal to the modulating frequency.

Grounds paths are very dangerous for disturbing d-c signals which may affect the d-c input as well as for a-c signals on modulating frequency. It is therefore a must to carefully design the grounding of the entire d-c amplifier. The disturbing voltages must especially be kept out of the meshes of the early amplifier stages. It is particularly dangerous to use a setup so that feedback can occur from the driving source of the modulators to the amplifier. An example to be avoided is given in Fig. 29. As seen the amplifier and the driving source use a common lead (which is grounded) to bring in their power from the external supplies. The driving source is very likely to produce, between the points a and b, an a-c voltage drop of which the frequency is exactly the modulating frequency. This will no doubt disturb the proper working of the amplifier. In order to avoid this but to keep amplifier and driving source on the same potential so as to prevent capacitive coupling between them we may wish to connect a single point of the amplifier and a single point of the driving source to a common ground by use of a very short wire while on the other hand feeding each of the two devices through completely isolated leads. It is understood that in that case the power supplies (here for example assumed to be batteries) should also be isolated from each other.

In passing we may also note that in the above example the situation will, of course, worsen if we use the same power supply for amplifier and driving source. We must not forget that no supply has zero internal resistance and so the latter will certainly give rise to unwanted feedback from driving source to amplifier. A capacitor may be used to reduce this feedback but if we want to measure signals as low as 1 to 10 microvolts then even with a large capacitor the influence of the feedback is still striking.

Another source of error due to ground path voltage drop is encountered when the external input circuit is grounded: a voltage drop resulting from the difference in potential of the two grounding points of the input circuit will give rise to an error in series with the input signal. (Note : One grounding point is the common ground of input and amplifier, the other is the added external grounding point.)

It is evident [10] also that if vacuum tubes are used and if the net frequency is the same as the modulating frequency then the power supplies for the electron tubes and the heater supplies should not contain net frequency components. If necessary these supplies shall be carefully filtered.

Since the input impedance of voltage d-c amplifiers is kept as high as possible in order not to drain an appreciable amount of current from the input source it is seen that the input of the amplifier should be protected against any low-frequency electrostatically induced voltage. The approach of any charged object (such as the operator's hand) will induce an error signal in the high impedance input circuit. This circuit should therefore be kept as small as possible and should also be electrostatically shielded against such effects.

Ahead[10] of the converter (= modulator) the signal is in the form of direct current; however the converter will also pass along a component of modulating frequency as a result of the lack of perfect balance in the input converter circuit. This may require the use of a filter to reduce this parasitic signal coming in on the wires with the d-c signal.

Although [10] a steady direct current from the first grid (or from the first base if transistors are used) cannot reach the input converter, variations of this current may reach the input converter, become modulated and cause trouble. Reduction of the absolute value of this current by the choice of the input tube (or transistor) should therefore be considered. It is also possible to use a filter between converter and first input stage in order to prevent the low-frequency grid - (or base -) current variations from reaching the converter, while the incoming signal at modulating frequency can freely pass. The filters to be used in the applications above usually consist of resistance-capacitance networks.

Radio frequency [10] disturbances are common because all forms of switching on power lines or in nearby electrical devices may generate considerable amounts of power, spread over the whole radio frequency spectrum. If this power reaches the amplifier or the contacts in the input converter a component at modulating frequency may be developed. The radio frequency path may be described as electric or electromagnetic but in any case, the protection against it is a tight, high- conductivity shield surrounding the critical tubes (or transistors) and the converter contacts. Radio frequency filters should then be used on all wires penetrating the shield.

b) Magnetic zero disturbances.[10]

Magnetic disturbances are disturbances which are generated by links of the d-c amplifier with varying magnetic fields.

Above we have seen that signals at modulating frequency appear on the input wires. These signals result from undesired electromagnetic coupling of the input wires with magnetic fields which are related to the driving of the converters (= modulator and demodulator). If filters are used in the input (as discussed in "Electrical zero disturbances" above) the external input circuit and early meshes of the input filter can tolerate larger induced voltages with frequency equal to the modulating frequency and from this point of view could tolerate larger loops. However, slowly varying magnetic fields may also be present and these, when loops are present will produce low-frequency voltages which can in no way be distinguished from the signal.

A means to reduce the influence of magnetic coupling upon the input wires is to twist the latter so that the voltages induced in each of them are approximately equal and so will tend to cancel.

To reduce [10] the effect of coupling with magnetic fields any part of the d-c amplifier system for which it is worth while to avoid this coupling may be shielded by use of mumetal or some other high-permeability material. This is particularly true for the entire input circuit and for the first amplifier stage(s).

Also [10] the radio frequency disturbances, as already mentioned before, are induced through the intermediary of electromagnetic coupling fields. Their effect can also be largely reduced by magnetic shielding of the input circuit and perhaps of the first amplifier stage(s).

If [10] the early stage amplifier tubes (or transistors) are subject to magnetic fields, the latter will have an unwanted effect on the electrons in the tubes (or on the electrons and holes in the transistors) and to protect against such fields magnetic shields may be used for the first tubes (or transistors).

If[10] an input transformer is used (as for example in Fig. 26) the latter must be well protected against magnetic fields. Since this transformer is operating at the lowest power level in the amplifier and since essentially it is a magnetic device it may very easily pick up any stray magnetic field if it is not thoroughly shielded by means of several high-permeability magnetic shields. (This "Input transformer problem" will also be considered apart in a later section of this chapter).

c) Mechanical zero disturbances.[10]

The item most susceptible to mechanical effect is the first stage amplifier tube. (The effect does not exist for transistors). Shocks and vibrational disturbances may have a component of modulating frequency which is high enough to show up in the output of the amplifier. Some tubes are more susceptible than others, but a rubber-mounted socket has proved adequate to protect against all ordinary shock and vibration. This socket is also of value in avoiding oscillations from mechanical feedback.

d) Thermal zero disturbances.[10]

The currents in the first amplifier stages are functions of temperatures (especially cathode temperature for tubes and overall temperature for transistors). However the variations of these currents, though they are the major cause of offset and drift in conventional d-c amplifiers, do almost not harm at all in modulated d-c amplifiers because these variations are attenuated by the conventional R-C networks between the several stages of the amplifier.

Thermoelectric voltages [10] arising from the contact of two non similar metals in the input circuit are the worst sources of disturbance due to thermal causes. There are two methods of attack :

1) To use metals giving the lowest obtainable thermoelectric power at all junctions.

2) To minimize temperature differences between thermo-electric junctions.

Since copper must be used in some places, notably the transformer primary, materials and special solder have been developed which have a low junction potential against copper and they have been found very helpful.[10]

To minimize [10] the temperature difference between thermoelectric junctions the complementary junctions may be placed in close thermal proximity, while at the same time the paths for heat flow to and from the junctions may be made thermally long and as nearly equal as possible.

7.- Input impedance and output impedance.

Since for voltage measurements a high input impedance is desirable for the overall d-c amplifier and since for current measurements a low input impedance is desired we will consider here the input impedance of modulated d-c amplifiers in order to determine whether or not the amplifier can be used for a particular application (either voltage or current measurement). Also the output impedance is worth mentioning as usually it is desired that it be as low as possible.

However, since the term "impedance" is usually reserved for applications where a-c signals are considered and we are dealing here with d-c input and output signals, we will speak here of "resistances" rather than of "impedances" and, if necessary, we will speak of instantaneous resistances to denote that the latter are variable in time.

a) Without feedback.

Let us first consider a simple modulated d-c amplifier without negative feedback (Fig. 30)

Input resistance.

By definition the instantaneous input resistance of the system of Fig. 30 is given by the ratio $\frac{e_i}{i_i}$. However we see that here this input resistance is infinite ($i_i = 0$) when the upper contact of the input switch is not closed and is equal to the input resistance of the a-c amplifier when the upper contact of the input switch is closed and the lower contact is open. The input resistance of the a-c amplifier to be considered here is the a-c input resistance since the input to the a-c amplifier is essentially in the form of a periodically changing signal. Let us for simplicity also assume that the band-width of the a-c amplifier is spread over the entire frequency band with the exception of direct current. This means in simple words that if the voltage applied to the amplifier remains constant over a small time interval then also the input current remains constant (in first approximation). In that case the input resistance during the time that the upper contact of the input switch is closed is also constant and equals the a-c input resistance of the a-c amplifier. Hence we see that in this case there are two different input resistances, one for the upper contact of the input being closed and one for this contact being open.

If the input circuit contains a parallel capacitor (low-pass filter) as in Fig. 31 for example and if the capacitor is very large then it turns out that the input current i_i is the average of the current i (Fig. 31). Hence in this case the input resistance is given by

$$R_{id} = \frac{e_i}{i_i} = \frac{e_i}{\frac{i_1 + i_2}{2}}$$

wherein as seen before i_1 is the current i when the upper contact of the input switch is open (hence $i_1 = 0$) and i_2 is the current i when the same contact is closed (hence $i_2 = \frac{e_i}{R_{ac}}$ where R_{ac} is the a-c input resistance of the a-c amplifier).

Then the overall input resistance is :

$$R_{id} = \frac{e_i}{\frac{e_i}{2R_{ac}}} = 2\ R_{ac}$$

If we use another input circuit, for example the one given in

Fig. 32 then evidently we may find another value for the overall input resistance. In the case of Fig. 32 we get (assuming the input transformer to be ideal and assuming zero time for the switch to change its position) :

1) when the upper contact of the input switch is closed and the lower one is open :

$$e = n.e_i$$

$$i = \frac{i_i}{n}$$

and the input resistance is :

$$R_{id} = \frac{e_i}{i_i} = \frac{e}{n.n.i} = \frac{R_{ac}}{n^2}$$

as expected from transformer theory.

2) when the upper contact of the input switch is open and the lower one is closed :

$$e = - n.e_i$$

$$i = \frac{-i_i}{n}$$

and the input resistance is again :

$$R_{id} = \frac{e_i}{i_i} = \frac{-e}{-n \;.\; ni} = \frac{R_{ac}}{n^2}$$

So the input resistance is always $\frac{R_{ac}}{n^2}$ in Fig. 32.

In any case (and this is true for any input circuit) we find the input resistance of the input circuit to be directly proportional to the d-c input resistance of the a-c amplifier (if we may assume ideal conditions to be present).

Output resistance.

We can consider the output of the a-c amplifier to be replaced by its equivalent Thevenin network with R_o being the output resistance of the a-c amplifier and E_o being the instantaneous output voltage if no load is present. Then the output circuit of the d-c amplifier of Fig. 30 can be represented as in Fig. 33. If now we load the output of the device then the output resistance of the latter is by definition the negative of the ratio of an infinitesimal change of the voltage across the load and the corresponding infinitesimal change of the current through the load when the latter is varied infinitesimally itself :

$$R_{od} = - \frac{dE_L}{dI_L} \quad \text{as in Fig. 34}$$

where R_{od} is the output resistance of the device. (Herein it is assumed that the voltage and current variations are due only to a load variation).

Obviously in Fig. 34 E_L is also the voltage across the output capacitor. If then the latter is very large the voltage E_L will be a constant in time if the load is not changed. What if we vary the load ? Then E_L will settle down to a new value which is such that the total charge which in one modulating period reaches the capacitor through the output circuit of the amplifier (the E-R_o circuit in Fig. 34) is equal to the total charge which in the same period flows away through the load. Now since the input and output converters are synchronized in such a way that during the half of the period that the upper contact of the input switch is closed the voltage E is constant (for constant d-c input !) we get for the total charge coming in via the a-c amplifier output circuit during the period T :

$$Q_T = \frac{E - E_L}{R_o} \cdot \frac{T}{2} .$$

The total charge flowing away through the load R_L for the same period is :

$$Q_{TL} = \frac{E_L}{R_L} \cdot T$$

Since for a steady-state situation it is required that the average charge of the capacitor remains constant we see that :

$$Q_T = Q_{TL}$$

which gives :

$$\frac{E - E_L}{R_o} \cdot \frac{T}{2} = \frac{E_L}{R_L} \cdot T$$

wherefrom by solving for E_L :

$$E_L = \frac{R_L}{R_L + 2R_o} \cdot E$$

The steady current in the load is :

$$I_L = \frac{E_L}{R_L} = \frac{E}{R_L + 2R_o}$$

The d-c output resistance of the device turns out to be :

$$R_{od} = - \frac{dE_L}{dI_L} = - \frac{dE_L}{dR_L} \cdot \frac{1}{dI_L/dR_L}$$

$$= - E \left[\frac{R_L + 2R_o - R_L}{(R_L + 2R_o)^2} \right] \cdot \left[\frac{(R_L + 2R_o)^2}{- E} \right]$$

$$= 2R_o \; .$$

In the case of a balanced output, for example the one shown in Fig. 26, of which the output is redrawn in Fig. 35, with equivalent Thevenin scheme given in Fig. 36, we find :

(Note that the output capacitor is not strictly necessary here if the arrangement consists of nothing but ideal elements).

1. For one half period :

Charge coming from the a-c amplifier output circuit =

$$Q_{T1} = I_{L1} \cdot \frac{T}{2} = \frac{I_o}{n} \cdot \frac{T}{2}$$

$$= \frac{E - E_o}{n.\ R_o} \cdot \frac{T}{2} = \frac{E - \frac{E_L}{n}}{n.\ R_o} \cdot \frac{T}{2}$$

$$= \frac{nE - E_L}{n^2 . R_o} \cdot \frac{T}{2}$$

Charge flowing through the load =

$$Q_{TL1} = \frac{E_L}{R_L} \cdot \frac{T}{2}$$

2. For the other half period :

$$Q_{T2} = I_{L2} \cdot \frac{T}{2} = - \frac{I_o}{n} \cdot \frac{T}{2}$$

$$= \frac{-(-E + \frac{E_L}{n})}{n.R_o} \cdot \frac{T}{2} = \frac{nE - E_L}{n^2 . R_o} \cdot \frac{T}{2}$$

$$= Q_{T1}$$

and

$$Q_{TL2} = \frac{E_L}{R_L} \cdot \frac{T}{2} = Q_{TL1}$$

Hence for a total period we get :

$$Q_T = Q_{T1} + Q_{T2} = \frac{n.E - EL}{n^2.R_o} \; . \; T$$

$$Q_{TL} = Q_{TL1} + Q_{TL2} = \frac{E_L}{R_L} \; . \; T$$

Since as in the first example it is required that

$$Q_T = Q_{TL}$$

we find :

$$\frac{n.E - E_L}{n^2.\,R_o} \; . \; T = \frac{E_L}{R_L} \; . \; T$$

wherefrom by solving for E_L :

$$E_L = \frac{n.\,R_L}{n^2.R_o + R_L} \; . \; E \; .$$

and subsequently :

$$I_L = \frac{E_L}{R_L} = \frac{n}{n^2.\,R_o + R_L} \; . \; E$$

The output resistance then turns out to be :

$$R_{od} = - \frac{dE_L}{dI_L} = - \frac{dE_L}{dR_L} \; . \; \frac{1}{dI_L / dR_L}$$

$$= - E \left[\frac{n(n^2R_o + R_L) - n\,R_L}{(n^2R_o + R_L)^2} \right] . \left[\frac{(n^2R_o + R_L)^2}{-nE} \right]$$

$$= n^2.\,R_o \; .$$

The two simple examples given above show that the d-c amplifier output resistance is in some way related to the a-c amplifier output resistance for a-c signals and will decrease or increase as the latter decreases or increases. This is also true in general and is not at all an unexpected result. In many cases however the set-up may be more complicated than the ones above so that the exact proportionality of d-c amplifier output resistance and a-c amplifier output resistance is lost. This is for example the case if the output capacitor is shunted by a resistor.

Obviously the input and output resistances of the overall d-c amplifier are in some way increasing functions of respectively the input and the output resistances of the a-c amplifier for a-c signals. If then we want for example the d-c input resistance to be high and the d-c output resistance to be low we should try to make the a-c input resistance of the a-c amplifier high and its a-c output resistance low.

b) With negative feedback.

If negative feedback is used in modulated d-c amplifiers we can simply refer to chapter 3 where the effect of negative feedback upon input and output resistance has been discussed in detail. The modulated d-c amplifier can indeed be considered as any other d-c amplifier as far as feedback is concerned because its input and output are both d-c. So all the conclusions made in chapter 3 are also valid here.

In the above discussion we made some simplifying assumptions (for example :

1. The times during which the input and output switches are closed (= the dwell times) are exactly equal to one half the modulation period.

2. Either of the two contacts of any switch was closed at a certain time instant, but never both at once).

The purpose of our discussion was to indicate the general way of treating problems related to d-c amplifiers. As a matter of fact it is not hard to find the performance characteristics of the d-c amplifier if the assumed simplifications are not allowed. We can however not give here all possible combinations and calculate the performance characteristics of all of them because this would really bring us too far. Good treatments of some of these problems can however be found in ref.[75], [68] and others. There calculations for gain and input and output resistances can be found based upon the assumption that the modulators are electromechanical choppers.

8.- Frequency response of modulated d-c amplifiers.

The frequency response of modulated d-c amplifiers is dependent upon the filters used in the circuitry and can also be dependent on the modulation frequency. Why do we say "can also be dependent on the modulation frequency" and not " is dependent on..."? We do say so because the frequency response is not necessarily dependent upon the modulation frequency. A popular misconception[68] is that no modulated amplifier can pass more than half its sampling frequency.

We think that it is clear that the frequency response of any amplifier is largely dependent upon the filtering (desired or not) in its circuitry. The modulated d-c amplifier, which is primarily intended to amplify d-c signals, often has low-pass filters at its input as well as at its output. These filters then severely limit the amplifier frequency response. It be noted however that at the same time the noise is also reduced considerably as we have already seen before. The latter effect is a very desirable one and constitutes the main reason for using filters in modulated d-c amplifiers.

Although usually it is believed that no modulated amplifier can pass more than half its frequency this is not necessarily true. Let us for example reconsider Fig. 26 and the corresponding wave form of Fig. 27. It was seen there that the output is an exact (but amplified) replica of the input, no matter what the modulating frequency is. This assumed however that the a-c amplifier was a perfect amplifier with constant gain A over the entire frequency spectrum and that the other elements used (transformers, choppers, etc) were also ideal. In practice however none of the elements used is ideal nor is the amplifier. Therefore some distortion of the output signal with respect to the input signal will occur. It is however not possible to speak of a useful bandwidth because the frequency response characteristic does not in general have the same form as that of a simple amplifier. The frequency response will depend upon the circuitry used and so will depend upon input and output circuits as well as upon the a-c amplifier circuit.

We will not take up here a discussion of the frequency response as a function of the circuitry used because our primary purpose is to discuss amplifiers for d-c signals rather than the use of such amplifiers for large bandwidths. This does not mean that the problem is not important. Indeed sometimes it can be very important but, as the reader will agree, it is not possible to discuss every complication of the problem of d-c amplification. We therefore think it is enough to point out that a very detailed and good discussion is given in part 6 of ref.[68] We cannot but refer to this reference if a more complete analysis of the dependence of the frequency response upon the circuitry used is desired.

9.- Input transformer problem.

Although we have already mentioned that the use of an input transformer in modulated d-c amplifiers causes particular problems we will look here a little deeper into these problems and investigate some possible means to reduce their effect.

The fact we have to pay attention to when designing the input circuit of a modulated d-c amplifier using an input transformer are :

1. Any transformer has some leakage flux. The latter should be made as small as possible because it reduces the primary inductance of the transformer and hence is the more objectionable the lower the operating frequency is. Furthermore, leakage is also to be avoided because it particularly affects low level application by making the flux density - input current characteristic very flat for low flux densities.

Leakage flux can be minimized by using a core made of material of very high permeability.

2. External magnetic fields will have a considerable effect upon the high-permeability transformer unless proper care has been taken to prevent their influence. This consists of using a transformer with the secondary and each half of the primary astatically wound on opposite limbs of the core (if the primary consists of only one winding the remark goes for this winding). Even so, a highly efficient magnetic shield is essential. In one design[53], [74] which permits the converter properties to be fully exploited, the primary inductance of the transformer is 50 Henry, and a double shield of 1/32 in. mumetal is used, with an air gap between the shields and another between the inner shield and the core. This gives a shielding efficiency of 60 dB, which could not be approached using a single shield, no matter how thick. In extreme cases it is necessary to demagnetize the whole unit to eliminate microphonic effects. Calculations about the shielding of input thransformers are given in ref.[2], [53].

3. The interwinding screen must be of full width to minimize interwinding capacitance and leads must be taken out as twisted pairs. This all is meant to reduce capacitive coupling between primary and secondary windings of the transformer.

4. Since the leakage and the parasitic capacitances in transformers can never be entirely eliminated, the transformer will have an optimal frequency range. This should be taken into account when an input transformer is chosen : the modulating frequency should be well into this optimum range for best performance.

5. Thermal e.m.f.'s may arise in the input transformer and in its connections to the rest of the circuitry. As these e.m.f.'s may be of the same level as the signal to be amplified they are highly objectionable. Such e.m.f.'s result from the fact that

the total length of the windings is rather great so that temperature gradients may easily arise along it, and also from the fact that the terminals of the transformer windings have to be soldered to circuit wiring so that thermoelectric voltages may certainly be expected. Circuit arrangements designed to avoid thermal e.m.f.'s in the input transformer are discussed in ref.[46].
Let us note also that we have already given some means of attack of the general problem of thermal e.m.f.'s arising from the contact of two non similar metals in the input circuit of modulated d-c amplifiers. This was done under the heading "thermal zero disturbances" of the section "Zero stability" of this chapter. We will not discuss it again here.

6. Fortunately[74], holes up to about 1/4-in. diameter in the shields of the input transformer have little effect on the shielding efficiency, even if those in both shields are in line (if two shields are used).

10.- Circuits for modulated d-c amplifiers.

Since the practical circuitry used for modulated d-c amplifiers is dependent on the type of modulators used, we shall give practical circuits in Part III where the different types of modulators are discussed and compared with each other. It will be seen there that the circuitry may differ depending on which type of modulator and demodulator is used. Although the general discussion of modulated d-c amplifiers does not depend upon the type of the modulating devices, the design of practical circuits does.

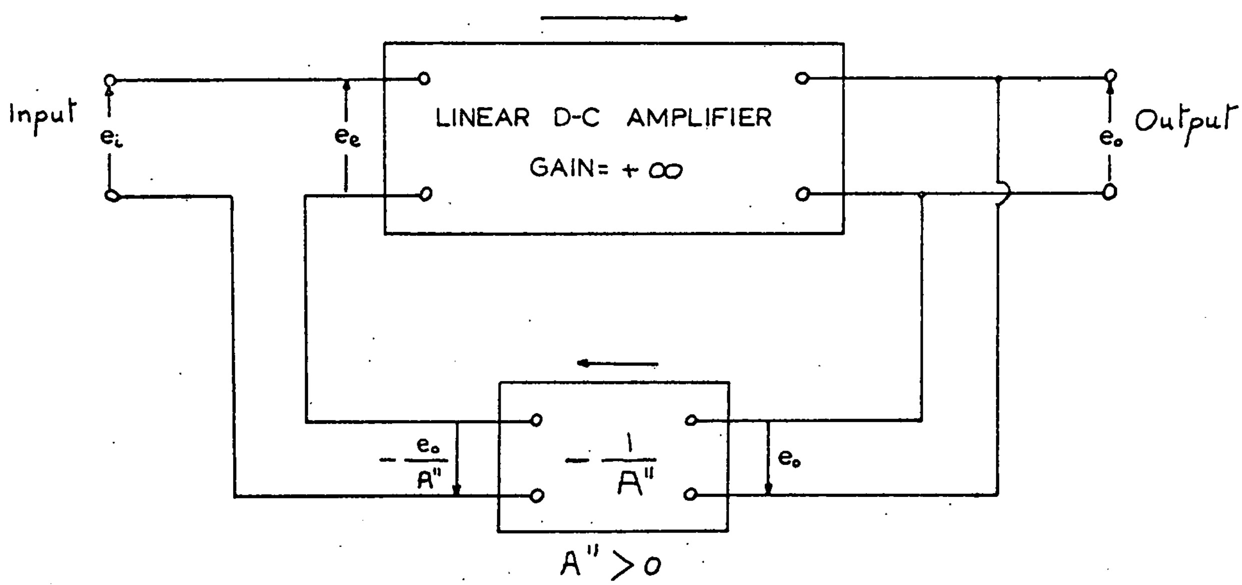

Input
e_i
e_e
LINEAR D-C AMPLIFIER
GAIN= + ∞
e_o
Output
$-\frac{e_o}{A''}$
$-\frac{1}{A''}$
e_o
$A'' > 0$

FIG. 1

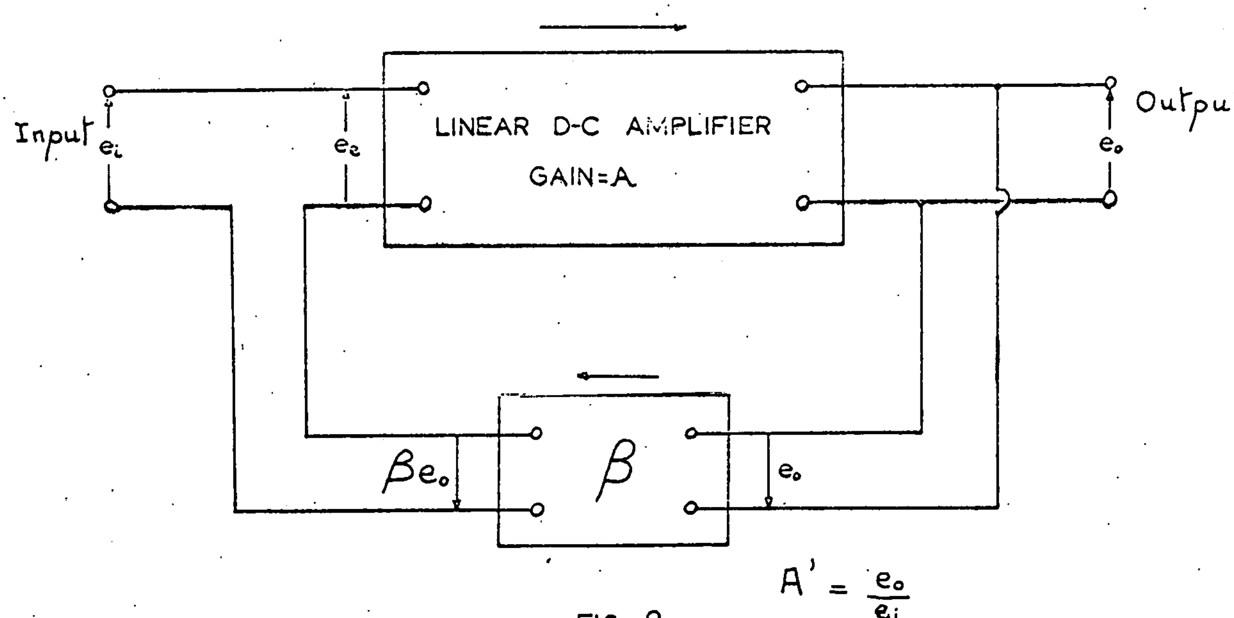
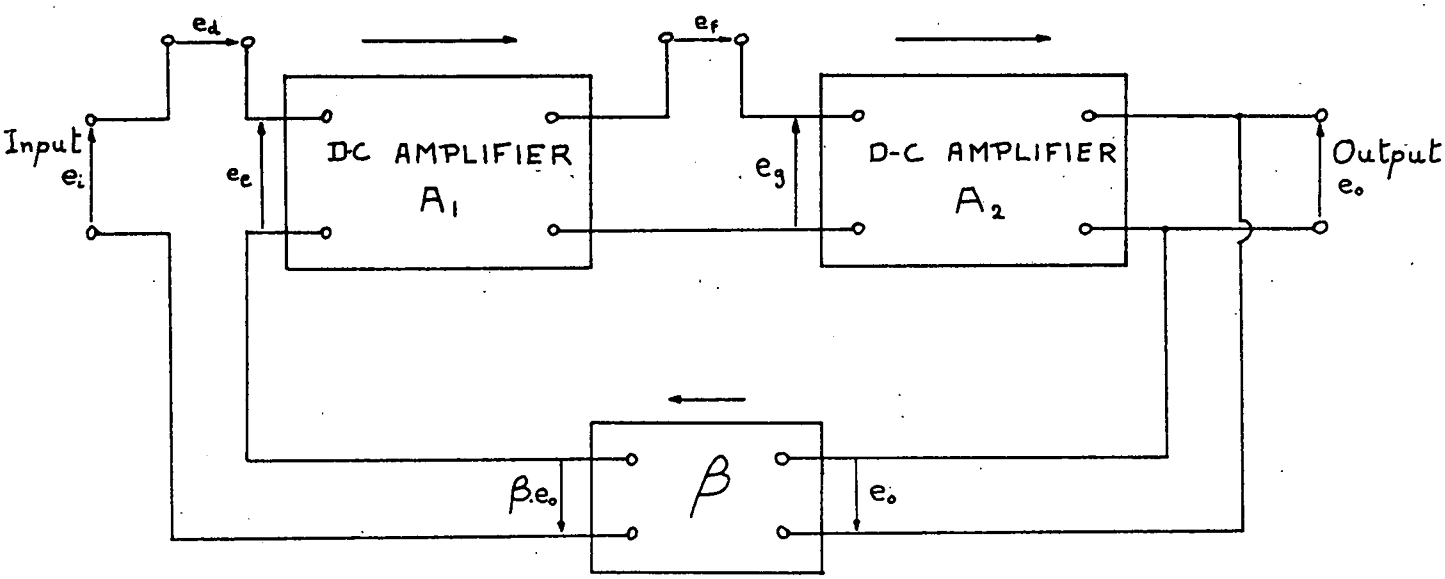
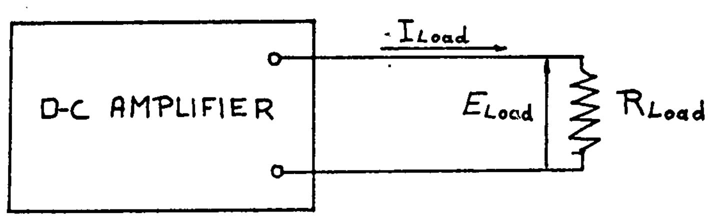

Input
e_i
e_e
LINEAR D-C AMPLIFIER
GAIN=A
e_o
Output
βe_o
β
e_o
$A' = \frac{e_o}{e_i}$

FIG. 2



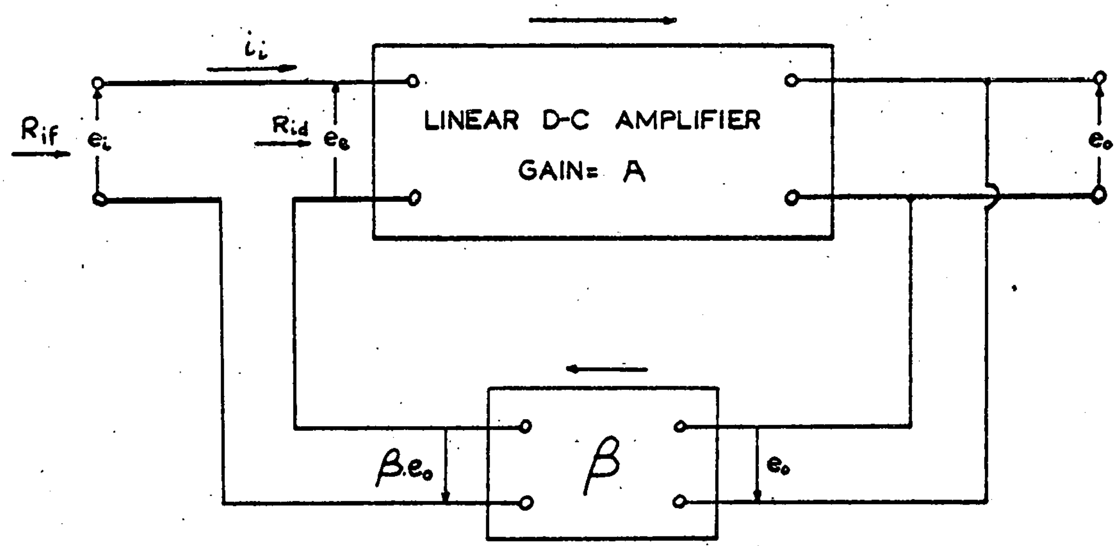
e_d
e_f
Input
e_i
e_e
D-C AMPLIFIER
A_1
e_g
D-C AMPLIFIER
A_2
Output
e_o
βe_o
β
e_o

FIG. 3



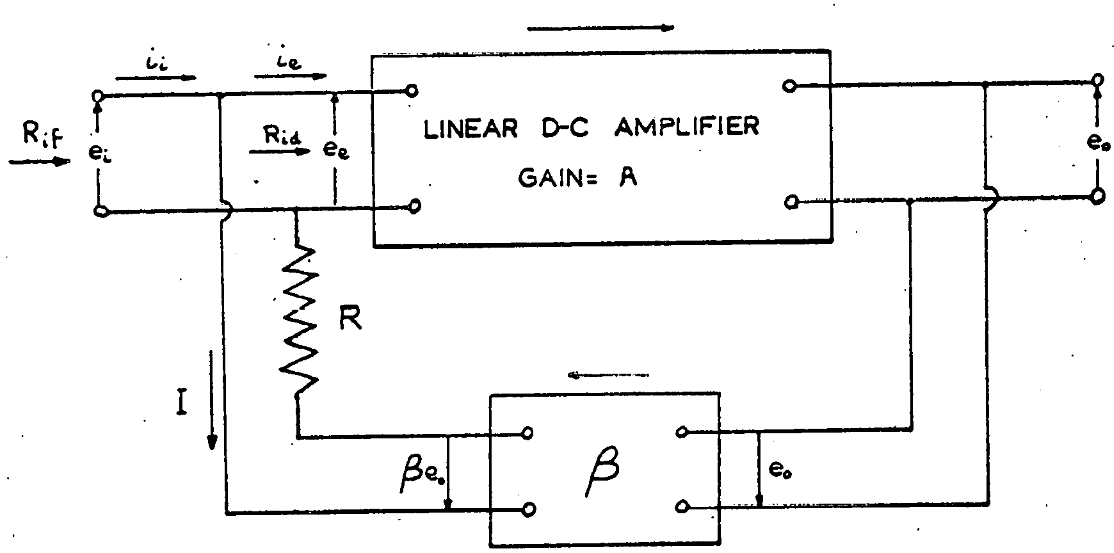
D-C AMPLIFIER
I_{Load}
E_{Load}
R_{Load}

FIG. 4



i_i
R_{if}
e_i
R_{id}
e_e
LINEAR D-C AMPLIFIER
GAIN= A
e_o
βe_o
β
e_o

FIG. 5



i_i
i_e
R_{if}
e_i
R_{id}
e_e
LINEAR D-C AMPLIFIER
GAIN= A
e_o
R
I
βe_o
β
e_o

FIG. 6

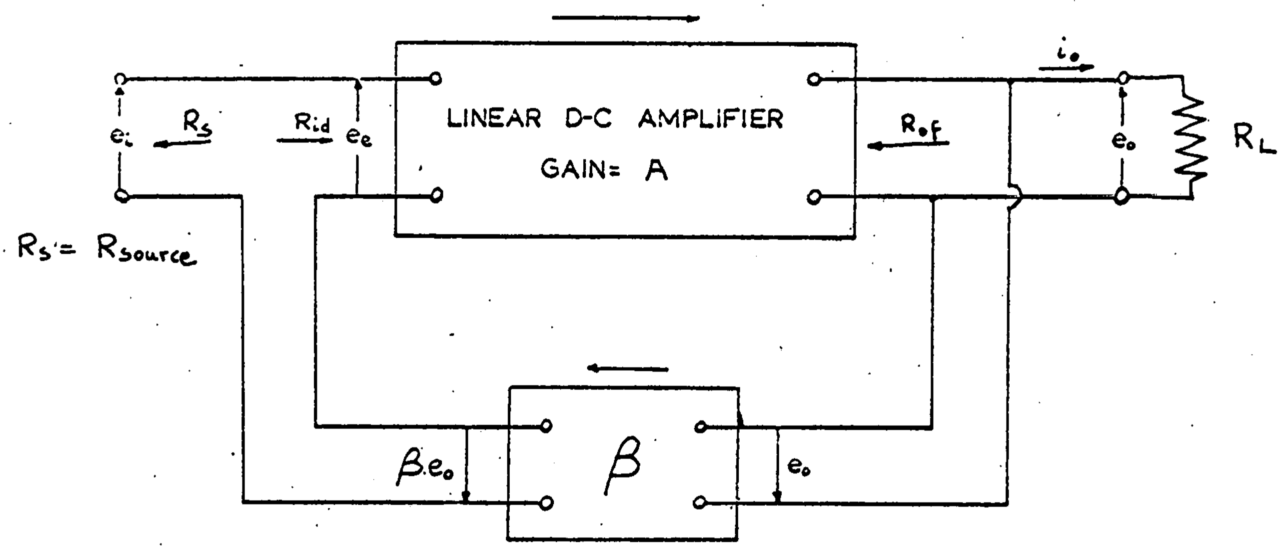
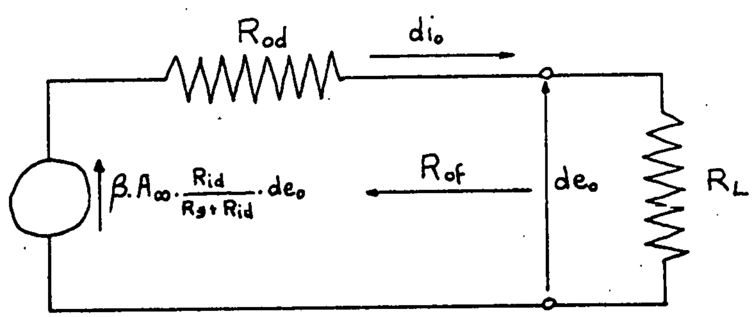
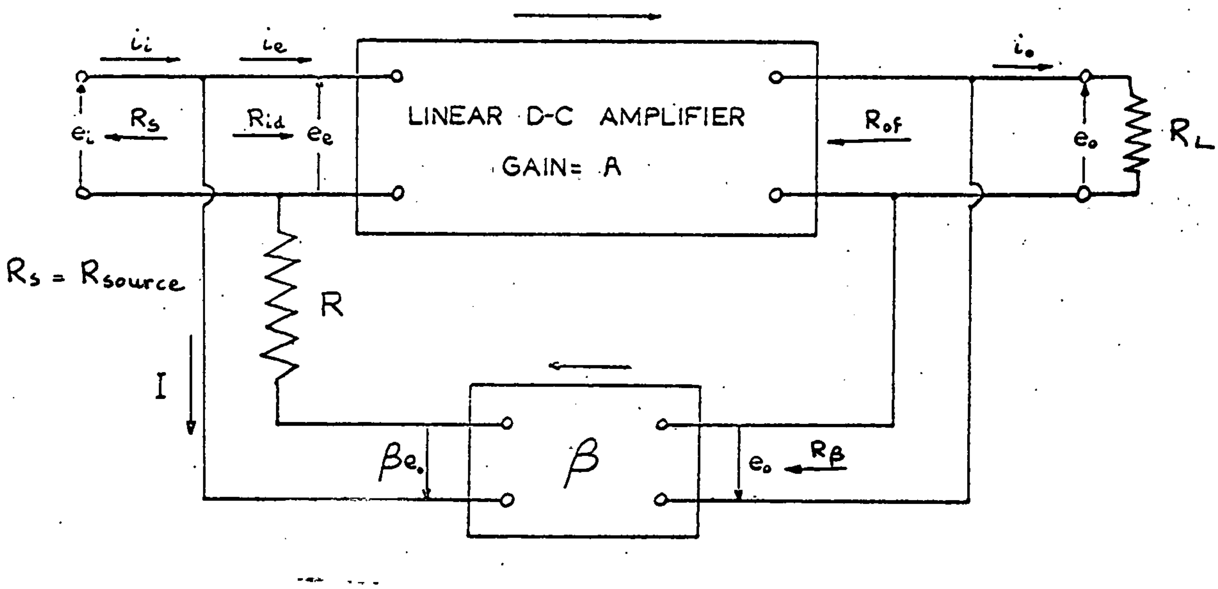


FIG. 7



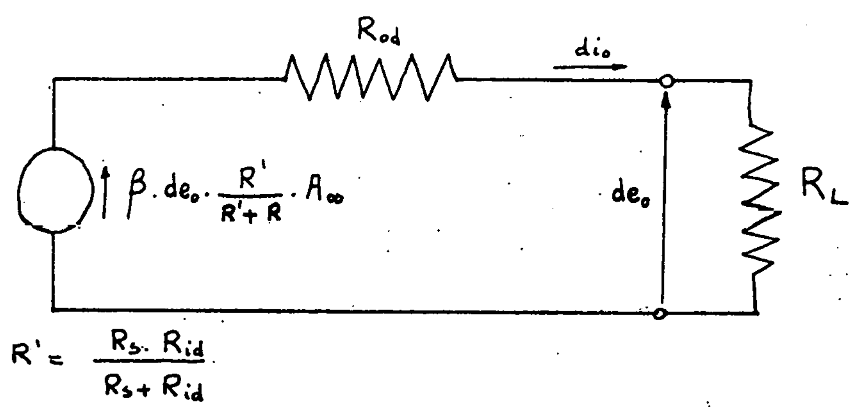
de_o = variation of output voltage
di_o = corresponding variation of output current

FIG. 8



i_i
i_e
i_o
e_i
R_s
R_{id}
e_e
LINEAR D-C AMPLIFIER
GAIN= A
R_{of}
e_o
R_L
$R_s = R_{source}$
R
I
βe_o
β
e_o
R_β

FIG. 9



R_{od}
di_o
$\beta . de_o . \frac{R'}{R'+R} . A_\infty$
de_o
R_L
$R' = \frac{R_s . R_{id}}{R_s + R_{id}}$

FIG. 10

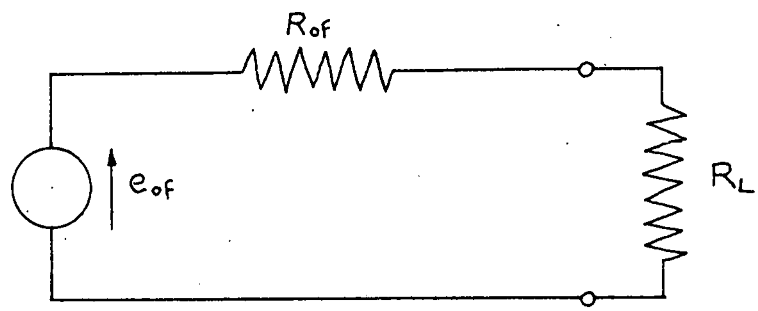


FIG. 11

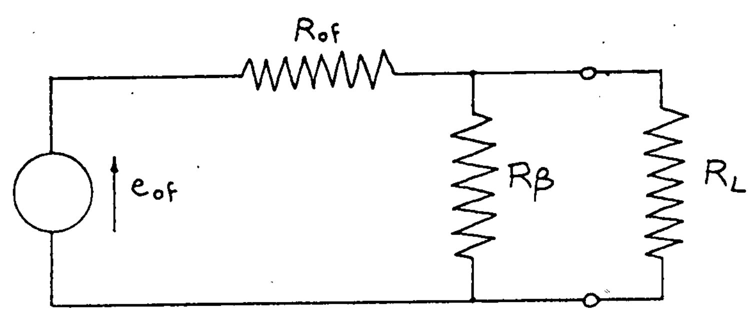


FIG. 12

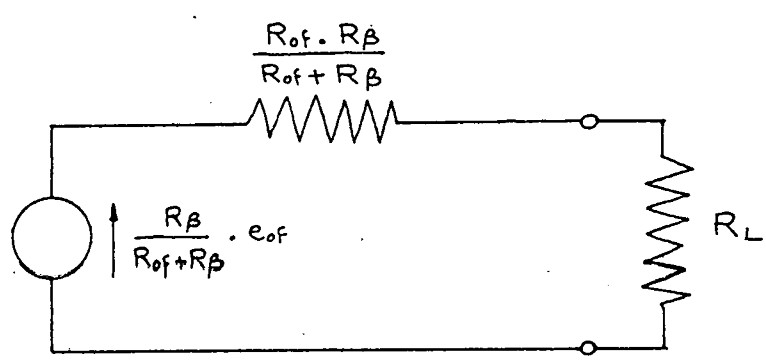
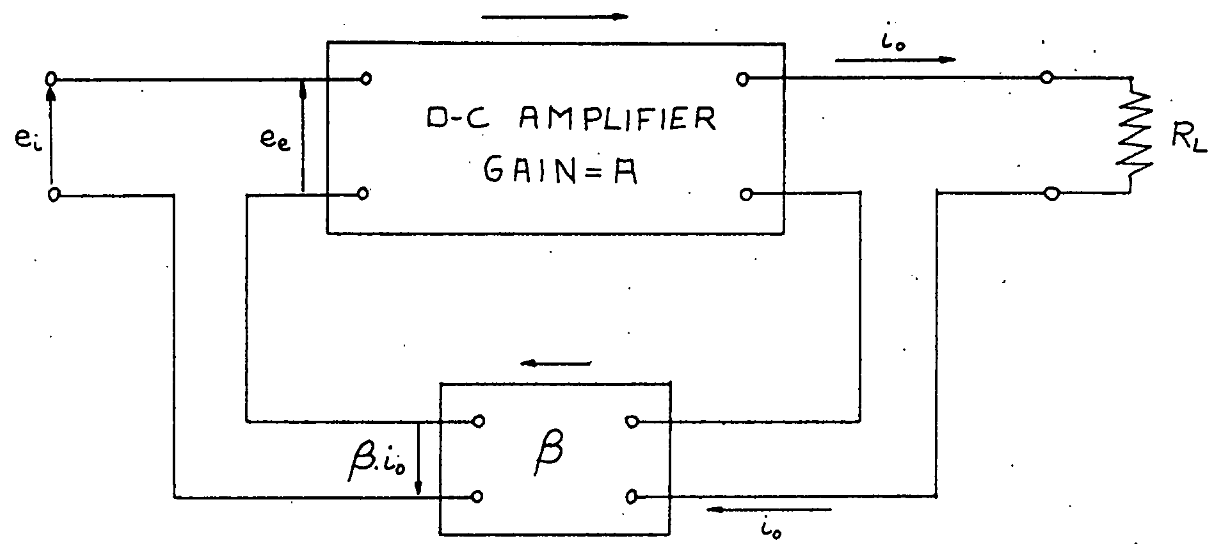
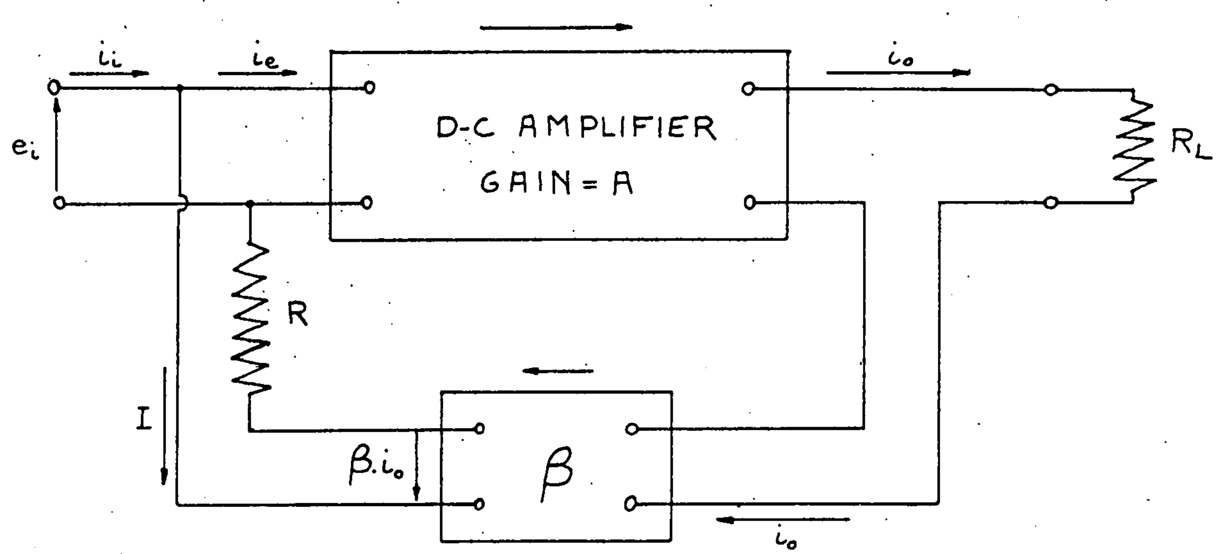


FIG. 13



D-C AMPLIFIER
GAIN = A
e_i
e_e
i_o
R_L
β
$\beta . i_o$

FIG. 14



D-C AMPLIFIER
GAIN = A
i_i
i_e
i_o
e_i
R_L
R
I
β
$\beta . i_o$

FIG. 15

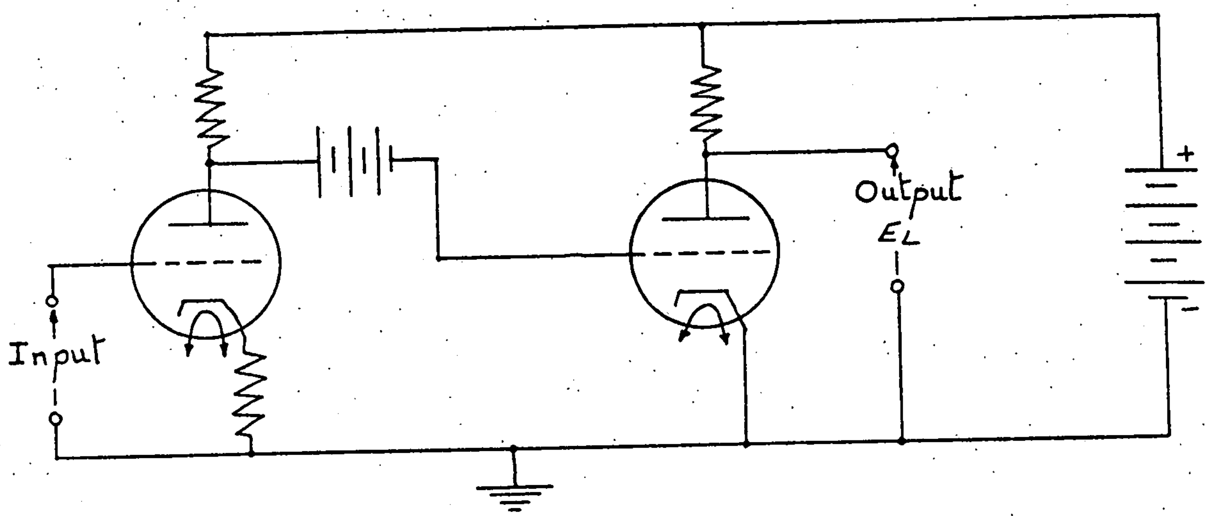
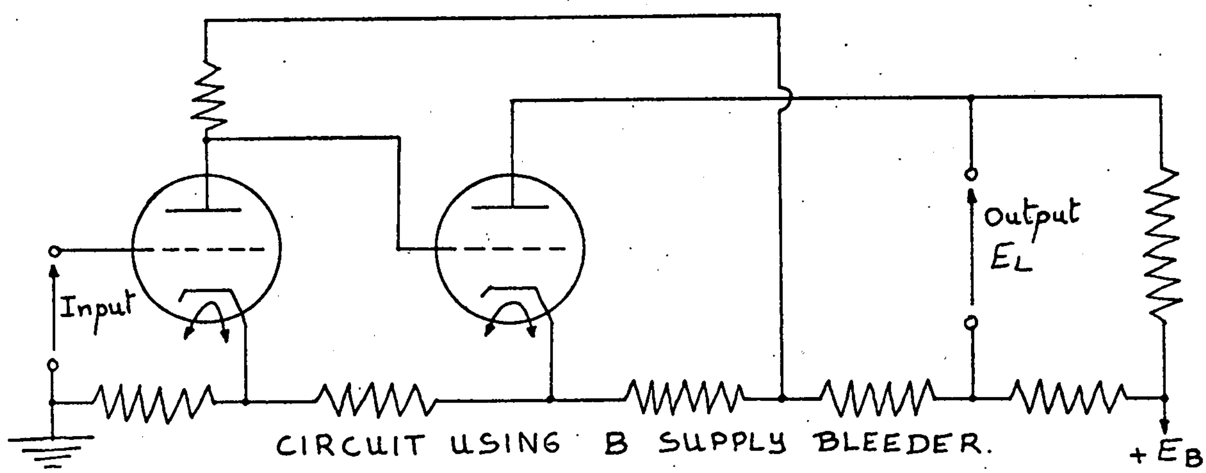
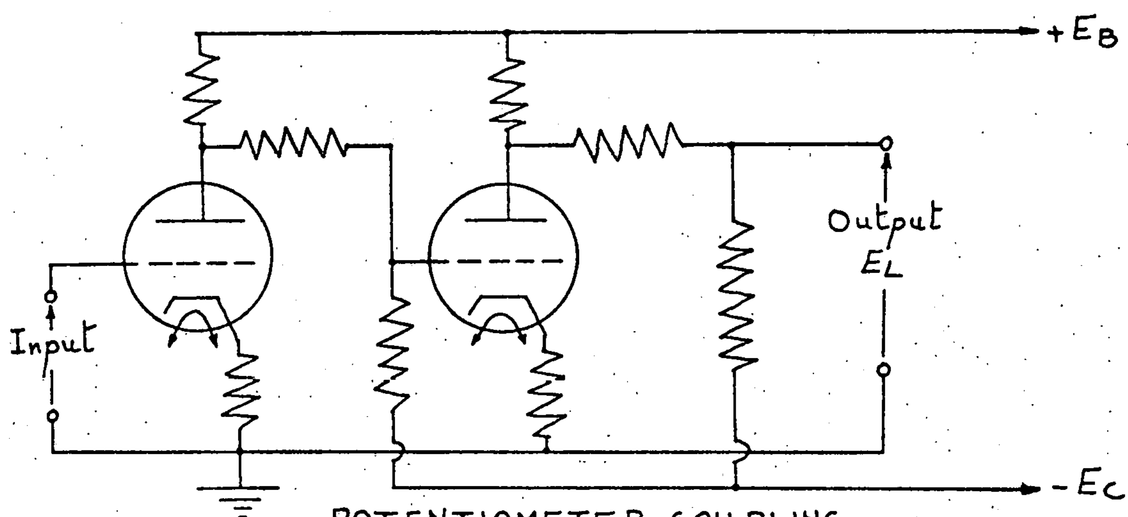


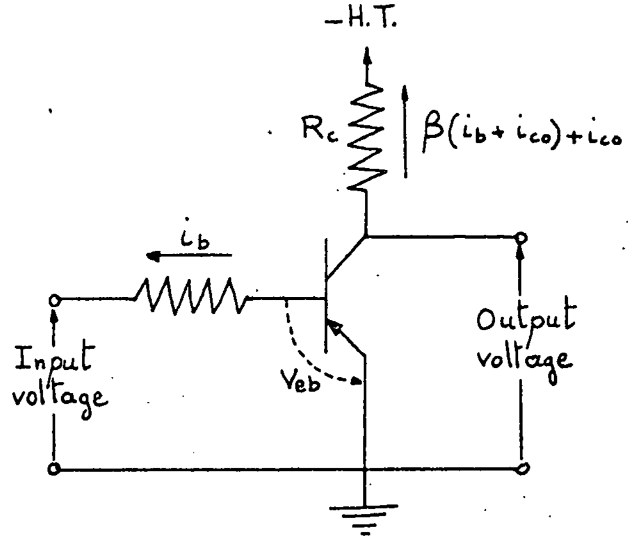
FIG. 16



CIRCUIT USING B SUPPLY BLEEDER.
FIG. 17



POTENTIOMETER COUPLING
FIG. 18



β = current gain for common emitter configuration

i_{co}, i_b, i_e, i_c = total instantaneous currents

V_{eb} = total instantaneous emitter-base voltage

Directly-coupled common emitter amplifying stage.
FIG. 19

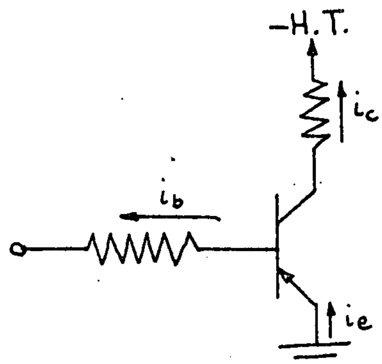
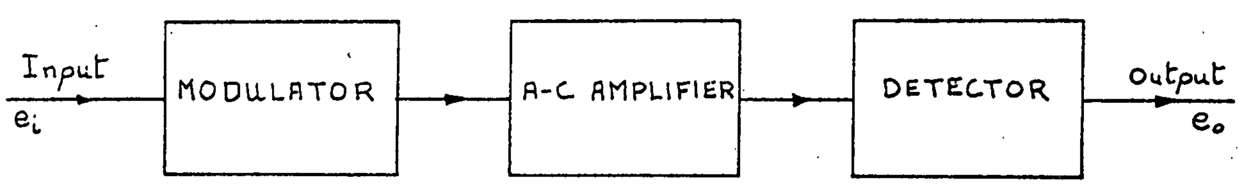


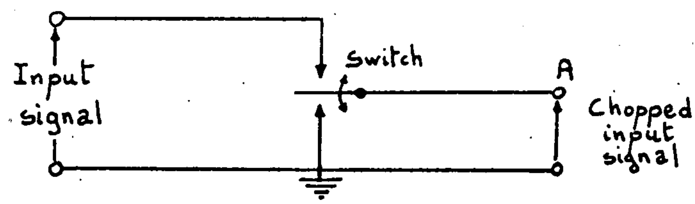
FIG. 20



BLOCK DIAGRAM OF A SIMPLE MODULATED

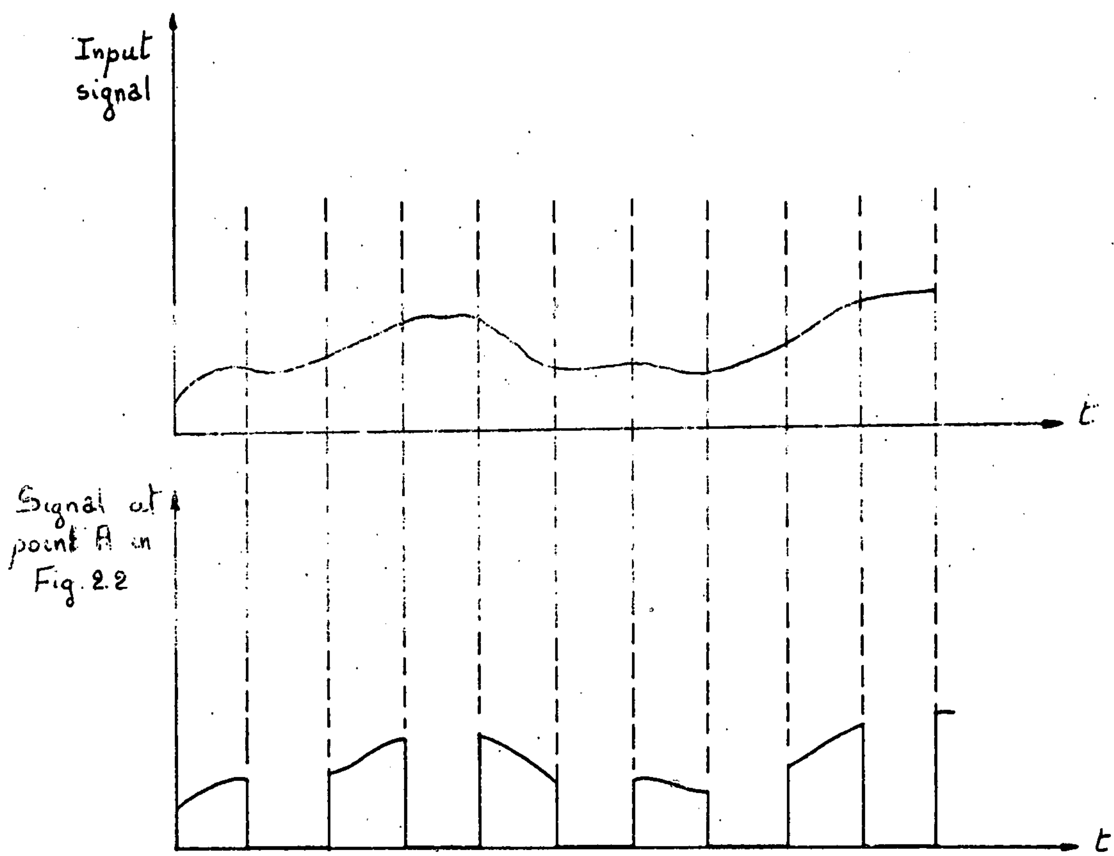
D-C AMPLIFIER

FIG. 21



CHOPPING SWITCH

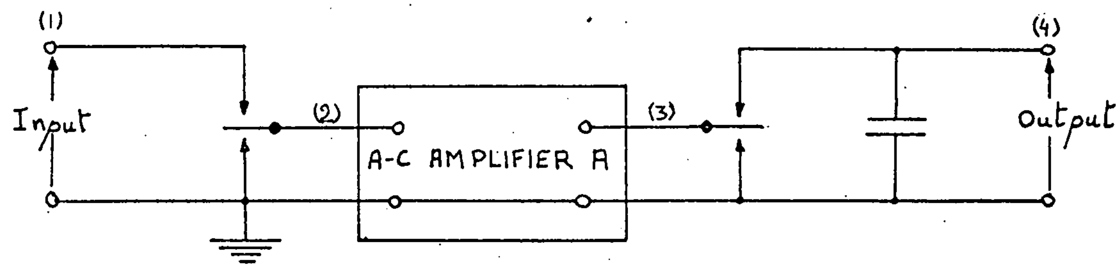
FIG. 22



INPUT SIGNAL AND CORRESPONDING CHOPPED INPUT SIGNAL

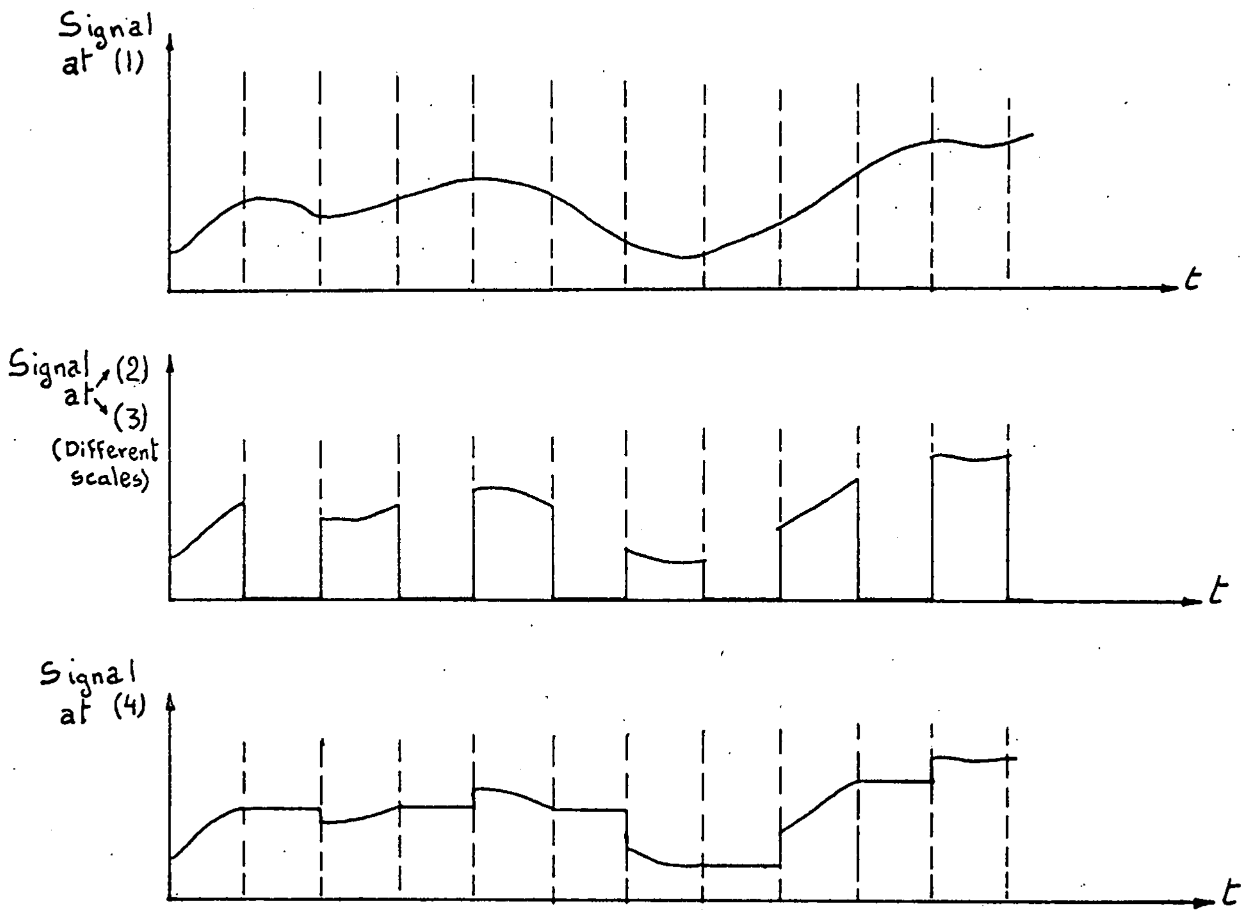
Dashed lines = switching instants

FIG. 23



COMPLETE SIMPLE MODULATED D-C AMPLIFIER

FIG. 24



Dashed lines = switching instants.

FIG. 25

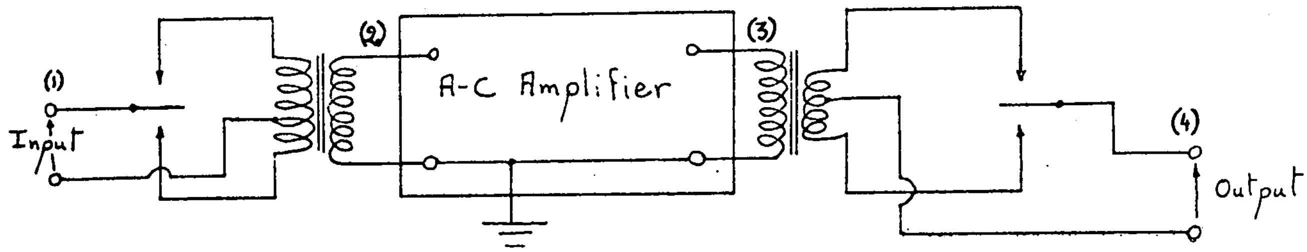

(1)
Input
(2)
A-C Amplifier
(3)
(4)
Output

FIG. 26

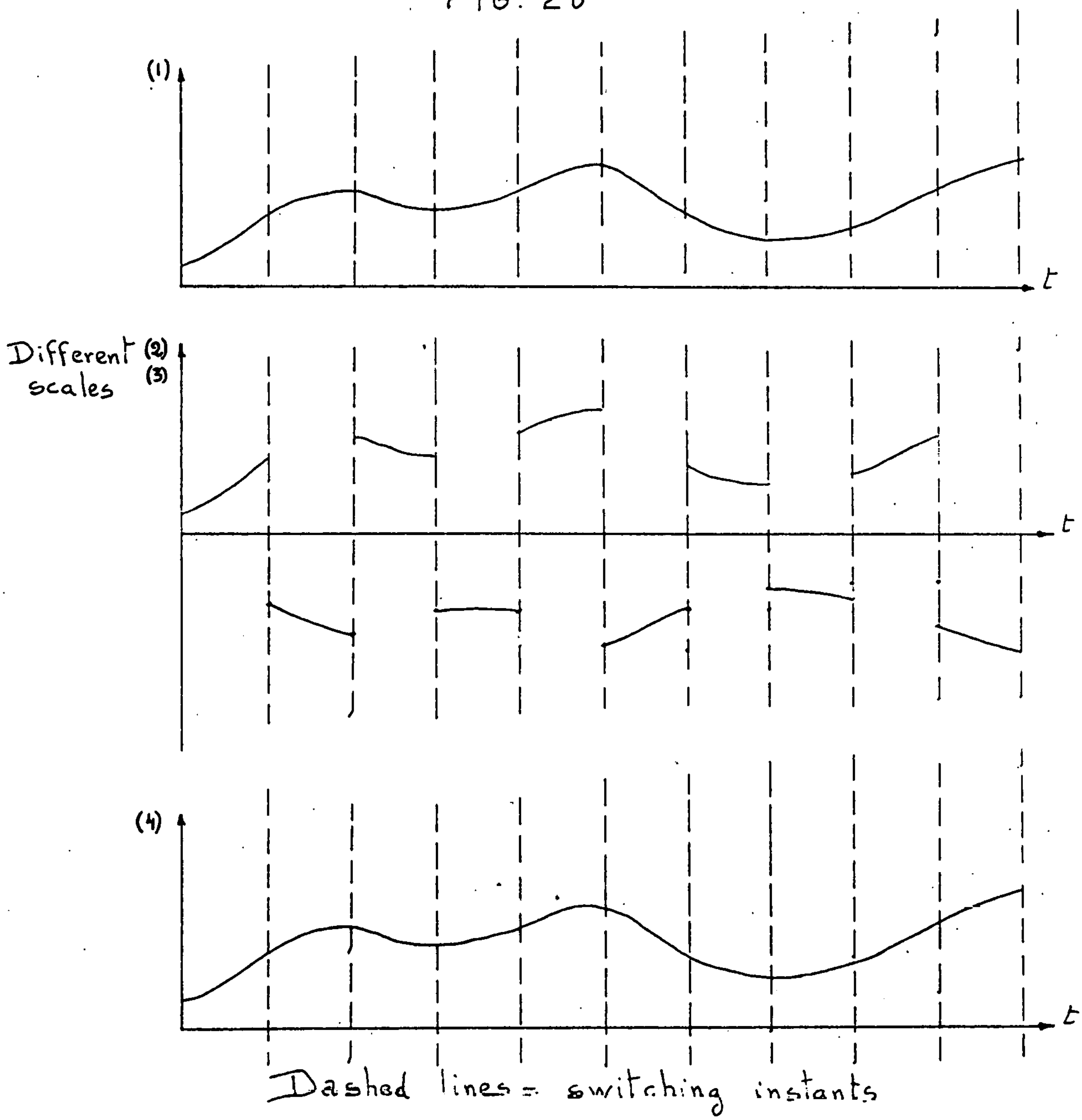
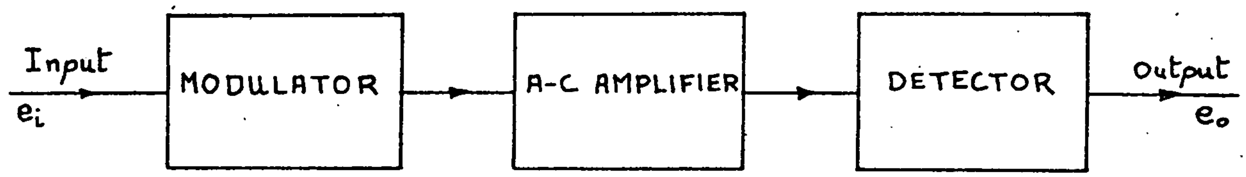
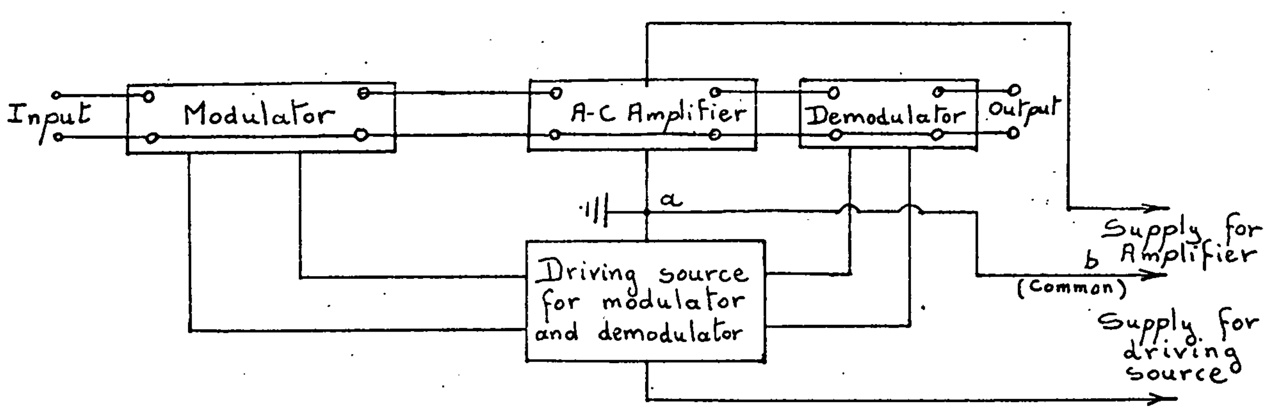

(1)
t
Different scales
(2)
(3)
t
(4)
t
Dashed lines = switching instants

FIG. 27



BLOCK DIAGRAM OF A SIMPLE MODULATED D-C AMPLIFIER

FIG. 28



Example of grounding connections to be avoided.

Fig. 29

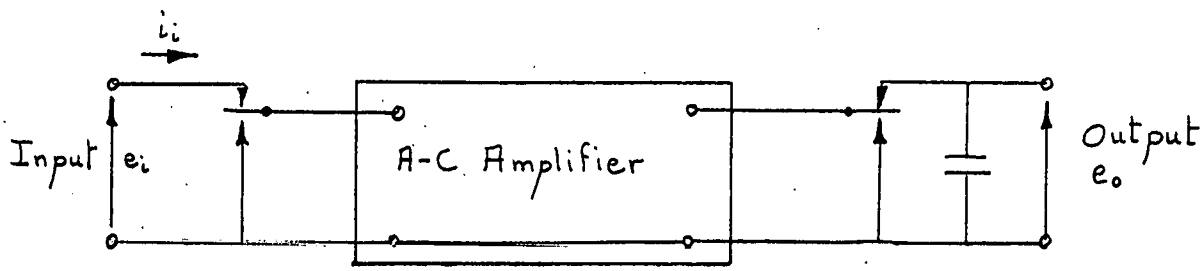


Fig. 30

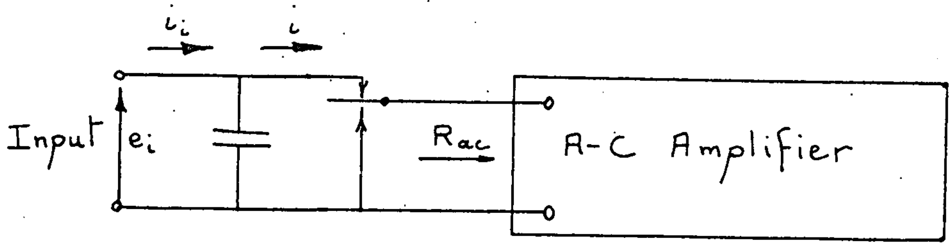


Fig. 31

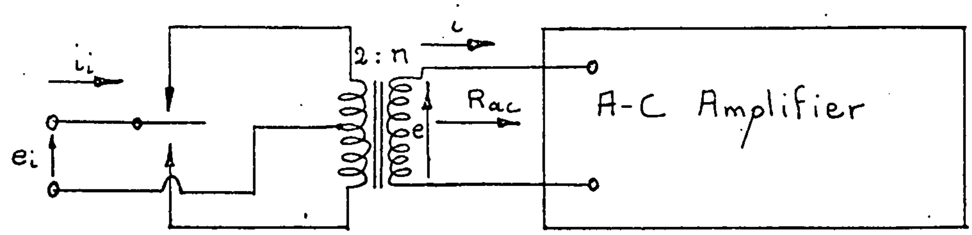


Fig. 32

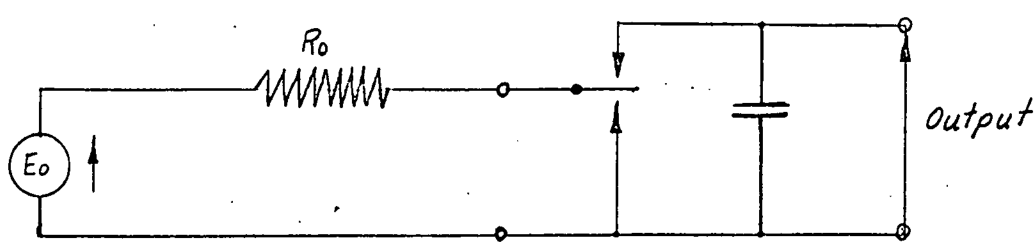


Fig. 33

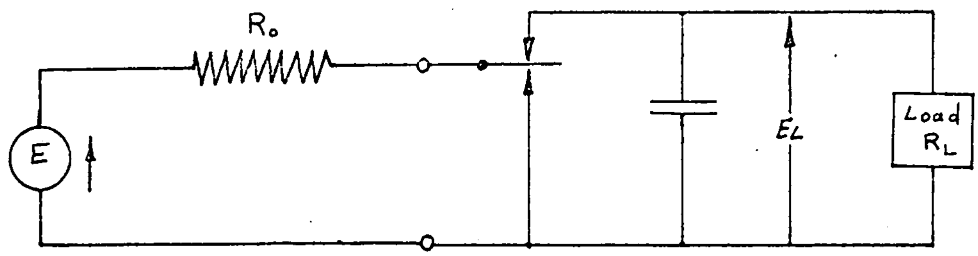


FIG. 34

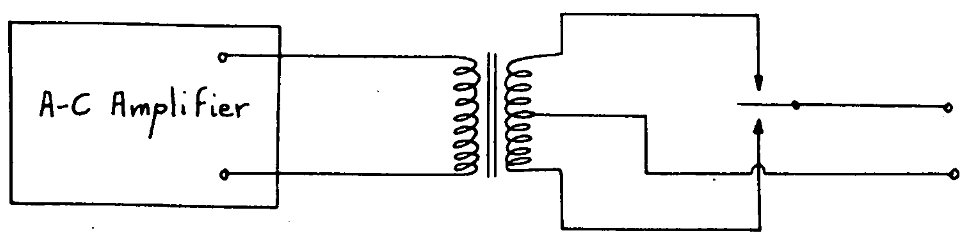


FIG. 35

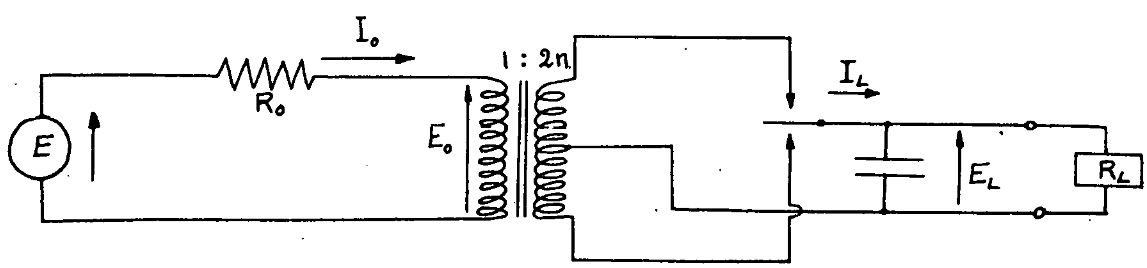


FIG. 36