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

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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.

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I.A.S, Aeronomica Acta A - N° 24 - 1963.

III. Modulators and demodulators,
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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.

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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,
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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 II.

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LOW AND VERY LOW LEVEL DC AMPLIFIERS (Part II)

THEORY (II)

by

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CHAPTER 6

COMPENSATED D-C AMPLIFIERS

In chapter 5 we have seen the excellent performance we can get when using modulated d-c amplifiers. However it was also mentioned there that the frequency response of the amplifier can be adversely affected by the use of modulating techniques. This does not occur in conventional direct-coupled d-c amplifiers. The latter however have considerable offset and drift. The question is then : is it not possible to improve the the performance of direct-coupled d-c amplifiers by using compensating techniques ?
The answer is yes.

A possible solution [5] is to build the amplifier in the form of a bridge wherein changing parameters compensate each other: this solution will be examined in champter 7.

Another solution[5] is to build the amplifier so that some variable characteristic of one element (e.g. a vacuum tube) is balanced against the same variation in another element or against a different characteristic of the same element. Note that this does not require the elements to be in the arms of a bridge. We shall discuss this solution in this chapter.

In chapter 4 we considered the basic circuits for simple cascaded direct-coupled d-c amplifiers. We shall now indicate how the circuit has to be changed in order to reduce the inherent offset and drift of these simple d-c amplifiers.

1. Vacuum-tube circuits.

There are two primary causes[5] of zero or steady-state drift in the conventional d-c amplifier circuit:

1) Drift due to changes in the plate supply voltage.
   This can be balances by
   a) either complete regulation of the supply
   b) or by designing tube circuits in which the change of plate current is exactly proportional to the change of supply voltage.

   The second method is to be preferred wherever possible, for it permits the use of an unregulated voltage supply. If the plate current variation of a tube is exactly proportional to the supply voltage variation, then the voltage developed between the plate of this tube and a proper tap on a supply bleeder will remain constant as the supply voltage varies (Fig.37).

   Small letters represent variations of a quantity from the quiescent conditions which are given by the corresponding capital letters.

   It is seen from Fig.37 that if $i = K.e_B$ we get for the variation of E_L :

$$e_L = (e_B - R.i) - \frac{R_2}{R_1 + R_2} e_B$$

$$= e_B (1 - KR) - \frac{R_2}{R_1 + R_2} e_B$$

$$= \left(1 - KR - \frac{R_2}{R_1 + R_2}\right) e_B$$

If $KR + \frac{R_2}{R_1 + R_2} = 1$

i.e. if $K = \frac{1}{R} \cdot \frac{R_1}{R_1 + R_2}$

then $e_L = 0$ which means that E_L is independent of E_B at least about the quiescent point for which the assumption $i = K.\ e_B$ is valid.

Practically R_1 and R_2 are so chosen that

$$\frac{R_1}{R_1 + R_2} = K.R$$

2) Drift due to changes in cathode temperature which can be subdivided into
   a) drift due to contact potentials which are a function of the cathode temperature. This drift is the most troublesome and has an effect similar to a change in grid bias as the cathode temperature changes.
   b) drift due to a variation in equivalent plate impedance of the tube as the latter is also a function of the cathode temperature.

These two forms of drift due to cathode temperature changes do not change by a definitely fixed amount with cathode temperature. They both have rather large components interrelated with plate current and plate voltage.

Where the tube is used only in the center of its linear range, and excursions in plate current due to signals are small, then the equivalent plate resistance r_p of the tube can be considered constant and the effect of the two forms of drift due to cathode temperature variations can be lumped together and treated as a shift in grid bias with cathode temperature.

A) A first modification to the circuit of a direct-coupled d-c amplifier is to use a differential amplifier circuit as the first stage (or stages) of the amplifier.

Let us consider Fig. 38 which represents a cathode-compensated differential amplifier stage [3],[5] which was first introduced by S. E. Miller in 1941 [3]. This circuit can compensate for contact potential drifts, e_f, due to variations of the cathode temperature of the tube. Indeed, let us for example, assume that the grid bias of both tubes changes over a voltage $-e_f$ due to a change of their cathode temperature. What will be the effect upon the output E_L ?

Let us denote by small letters the variations of the quiescent quantities which in Fig. 38 are given by capital letters. Futhermore

r_p = internal dynamic resistance of the tubes (assumed identical).

μ = amplification factor of the tubes.

e_f = change of the cathode voltage of both tubes due to a change in temperature of the cathode (Fig. 39).

For generality we also include an input voltage variation e_1. However we assume the supply voltage E_B to be constant. We now apply the general vacuum tube formula :

$$r_p.i = e_b + \mu.\, e_g$$ to

both tubes :

tube T_1 :

$$r_p.i_1 = - R_3 \,.\, i_1 - e_f - e_K$$

$$+ \mu\,(e_1 - e_f - e_K)$$

$$= - R_3 \,.\, i_1 - (1 + \mu)\,(e_f + e_K) + \mu e_1$$

tube T_2 :

$$r_p.i_2 = - e_f - e_K$$

$$+ \mu(e_2 - e_f - e_K)$$

$$= - (1 + \mu)\,(e_f + e_K) + \mu e_2$$

Adding the two equations we get:

$$r_p.(i_1 + i_2) = -2\ (1 + \mu)\ (e_f + e_K) - R_3.i_1$$
$$+ \mu\ (e_1 + e_2)$$

But $e_K = (R_1 + R_2)\ (i_1 + i_2)$

$e_2 = R_2\ (i_1 + i_2)$

Hence :

$$r_p.\ (i_1 + i_2) = -2(1 + \mu)\ e_f - 2(1 + \mu)(R_1 + R_2)(i_1 + i_2)$$
$$-R_3.i_1 + \mu e_1 + \mu R_2\ (i_1 + i_2)$$

or

$$[\,r_p + 2(1 + \mu)\ (R_1 + R_2) - \mu\ R_2\,]\ (i_1 + i_2)$$
$$= -2(1 + \mu)\ e_f + \mu e_1 - R_3\ .\ i_1$$

Now letting

$$R_K \equiv r_p + 2(1 + \mu)\ (R_1 + R_2) - \mu R_2$$

we find:

$$R_K\ (i_1 + i_2) = -\ (1 + \mu)\ e_f + \mu e_1 - R_3\ .\ i_1$$

Eliminating $(i_1 + i_2)$ between the latter equation and the equation for tube T_1 yields:

$$r_p.i_1 = -R_3.\ i_1 - (1 + \mu)\ e_f + \mu e_1$$
$$- \frac{(1+\mu)\quad (R_1 + R_2)}{R_K}[-2\ (1+\mu)\ e_f + \mu e_1 - R_3\ i_1]$$

or

$$(r_p + R_3).R_K.i_1 = -(1+\mu)R_K\ e_f + \mu.\ R_K\ .\ e_1$$
$$+ (1+\mu)\ (R_1 + R_2)\ [\,2(1 + \mu)e_f - \mu e_1 + R_3 i_1\,]$$

wherefrom we find after bringing all the terms containing i_1 together:

$$i_1 \left[(r_p + R_3) R_K - (1 + \mu)(R_1 + R_2) R_3 \right]$$

$$= -(1 + \mu) \left[2(1 + \mu)(R_1 + R_2) - R_K \right] e_f$$

$$+ \mu \left[R_K - (1 + \mu)(R_1 + R_2) \right] e_1$$

or by substituting for R_K:

$$i_1 \left\{ (r_p + R_3) \left[r_p + 2(1 + \mu)(R_1 + R_2) - \mu R_2 \right] - (1+\mu)(R_1 + R_2) R_3 \right\}$$

$$= -(1 + \mu) \left[2(1 + \mu)(R_1 + R_2) - r_p - 2(1+ \mu)(R_1 + R_2) + \mu R_2 \right] e_f$$

$$+ \mu e_1 \cdot \left[r_p + 2(1 + \mu)(R_1 + R_2) - \mu R_2 - (1 + \mu)(R_1 + R_2) \right]$$

or

$$i_1 R_c = (1 + \mu)(r_p - \mu R_2) e_f + \mu R_b e_1$$

where

$$R_c \equiv (r_p + R_3) \left[r_p + 2(1 + \mu)(R_1 + R_2) - \mu R_2 \right]$$

$$- (1 + \mu)(R_1 + R_2) R_3$$

$$R_b \equiv r_p + (R_1 + R_2) + \mu R_1$$

It will be seen that the output is given by:

$$e_L = e_3 = - R_3 \cdot i_1$$

In order then for the disturbance e_f not to appear in the output we choose R_2 such that $\mu R_2 = r_p$: in that case the term containing e_f drops out of the equation which determines i_1 and will hence not show up in the output either.

Then
$$i_1 = \frac{\mu R_b}{R_c} \, e_1$$

But in that case
$$R_K = \mu R_2 + 2\,(1+\mu)\,(R_1 + R_2) - \mu\, R_2$$
$$= 2\,(1+\mu)\,(R_1 + R_2)$$

and
$$R_c = 2(1+\mu)\,(R_1 + R_2)\,(R_3 + \mu R_2)$$
$$- (1+\mu)\,(R_1 + R_2)\,R_3$$
$$= (1+\mu)\,(R_1 + R_2)\,(R_3 + 2\mu R_2)$$

and
$$R_b = \mu R_2 + (R_1 + R_2) + \mu R_1 = (1+\mu)\,(R_1 + R_2)$$

wherefrom:
$$i_1 = \mu \frac{(1+\mu)\,(R_1 + R_2)}{(1+\mu)\,(R_1 + R_2)\,(R_3 + 2\mu R_2)} \cdot e_1$$
$$= \frac{\mu}{R_3 + 2\mu R_2} \cdot e_1 = \frac{\mu}{R_3 + 2\,r_p} \, e_1$$

Obviously (Fig. 39) the variation of the anode voltage of tube T_1 is:

$$e_3 = - R_3\, i_1 = - \frac{\mu R_3}{R_3 + 2\,r_p} \cdot e_1$$

provided the condition $\mu R_2 = r_p$ is satisfied and the supply voltage E_B doe not change. (Changes of this voltage will be considered later.) Since the voltage of the common point of the resistors R_4 and R_5 doesnot change if the supply voltage E_B is constant the variation e_L of the output will be equal to the variation of the anode voltage of T_1 :

$$e_L = e_3 = - \frac{R_3}{R_3 + 2\,r_p} \cdot e_1$$

if the conditions : $\mu R_2 = r_p$

and E_B= constant

are satisfied.

This output variation is independent of e_f so that we get complete balance for e_f provided all the assumptions we have made are valid. In any case we see that the effect of e_f will be considerably reduced by this circuit.

Note that the condition which is necessary to make the output independent of e_f can also very easily be derived on physical grounds: Let us assume that in Fig. 39 $i_1 = 0$ (small letters are variations from the quiescent conditions) for $e_1 = 0$ in order that e_3 be = 0 for $e_1 = 0$. The condition $i_1 = 0$ means that $e_K = -e_f$ in order that the cathode voltage of tube T_1 remains independent of e_f as is required if $i_1 = 0$.
If the cathode voltage of T_1 remains constant, then the cathode voltage of T_2 must also remain constant, for the disturbing voltages e_f are assumed to be identical for both tubes. It is clear that the general vacuum tube equation for T_2 becomes in this case:

$$r_p \cdot i_2 = e_b + \mu e_2$$

where $e_b = 0$

$$e_2 = R_2 \cdot i_2 \text{ (note that } i_K = i_1 + i_2 = 0 + i_2 = i_2 \text{)}$$

for since the cathode voltage of T_2 does not change, the variation of the grid voltage of T_2 with respect to its cathode is the same as the variation of that same grid voltage with respect to any other constant voltage (here for convenience the ground voltage).

Consequently the tube equation yields:

$$r_p \cdot i_2 = \mu \cdot R_2 \cdot i_2$$

wherefrom the condition $r_p = \mu R_2$ is obtained.

Also other possible schemes for balancing effects due to cathode temperature variations were proposed[65] but they turned out not to be as good as the one described above.

B) The cathode-compensated amplifier described above can be considerably improved[5] and still geater freedom from drift obtained if the input voltage includes a tap on the supply bleeder at the value of E_F as shown in Fig. 40. (However, in a specific case this voltage E_f can be zero.) When the output is returned to the proper point on the bleeder, the effect of changing E_B is eliminated and the plate supply need not be regulated. There is a definite relationship between the points on the bleeder to which the input and the output have to be returned in order to give the desired elimination of the influence of plate supply changes upon the output. Note that (Fig.40) if $n \neq 0$ the input has to be floating.

Let us in Fig. 40 calculate what the voltages E_D and E_F have to be in order that the output be independent of supply voltage variations. We will not again as in the previous case develop here a complete formal derivation of the output but we will rather find the values of E_D and E_F by partially using physical reasoning as we did in the last part of the previous case.

Again let small letters be the variations of the quantities which in Fig. 40 are given by capital letters.
In Fig. 40 the condition $\mu R_2 = r_p$ derived in case A is assumed to be satisfied. We now assume a variation e_B of the supply voltage E_B and we will derive the necessary conditions which make e_L zero for such a variation.

Let $E_D = m. E_B$ and $E_F = n. E_B$. If E_B varies by an amount e_B then it is necessary that

$$e_3 = e_D \quad \text{in order that}$$

$$e_L = e_3 - e_D \quad \text{be zero.}$$

Now $\qquad e_D = m. e_B.$

In order that $e_3 = e_D = m. e_B$ it is necessary that

$$e_3 = e_B - R_3.i_1 = e_D = m. e_B$$

wherefrom :

$$i_1 = \frac{(1 - m) e_B}{R_3}$$

Considering tube T_2 we have, using the general vacuum tube formula:

$$r_p.i_2 = e_{b2} + \mu.e_{g2}$$

or

$$r_p.i_2 = e_B - e_K + \mu(e_2 - e_K)$$

$$= e_B + \mu e_2 - (1 + \mu) e_K$$

Since $e_2 = R_2 (i_1 + i_2)$ we have

$$r_p.i_2 = e_B + \mu R_2 i_1 + \mu R_2 i_2 - (1 + \mu) e_K$$

or

$$0 = e_B + \mu R_2 i_1 - (1 + \mu) e_K$$

because $r_p = \mu R_2$

Using the general vacuum tube formula for tube T_1 yields:

$$r_p.i_1 = e_{b1} + \mu.e_{g1}$$

$$r_p.i_1 = e_3 - e_K + \mu (e_1 - e_K)$$

or

$$r_p.i_1 = e_3 + \mu e_1 - (1 + \mu) e_K$$

Eliminating $(1 + \mu) e_K$ between this formula and

$$0 = e_B + \mu R_2 i_1 - (1 + \mu) e_K$$

found above we get:

$$r_p \cdot i_1 = e_3 + \mu\, e_1 - e_B - \mu R_2\, i_1$$

or
from $r_p = \mu R_2$

$$2r_p \cdot i_1 = e_3 + \mu e_1 - e_B$$

We also see (Fig.40) that

$$e_1 = e_F = n \cdot e_B$$

(because only variations of the voltage supply are considered)

This then gives when substituted in the last equation above:

$$2r_p \cdot i_1 = e_3 + \mu n e_B - e_B$$

Furthermore $e_3 = e_B - R_3 \cdot i_1$

which yields

$$2r_p \cdot i_1 = e_B - R_3 \cdot i_1 + \mu n e_B - e_B$$

or

$$(2r_p + R_3)\, i_1 = \mu n e_B$$

Above we found:

$$i_1 = \frac{(1 - m)\, e_B}{R_3}$$

From the two last equations we get the desired condition by eliminating i_1 :

$$(2r_p + R_3)\ \frac{(1 - m)}{R_3}\ e_B = \mu n e_B$$

or

$$(2r_p + R_3)\,(1 - m) = \mu n R_3$$

or

$$2(1 - m)\, r_p = R_3\,(m + \mu n - 1)$$

This final result is the condition for which variations of the B supply E_B in Fig. 40 do not affect the output voltage E_L provided $r_p = \mu R_2$. This condition states a relationship between m and n.

Note that m and n should be such that

$$0 \leq m \leq 1$$

$$0 \leq n \leq 1$$

for the configuration of Fig. 40.

If the input voltage has a terminal on ground potential then n = 0 (that is $E_F = 0$). This requires m = 1 which means that $E_D = E_B$. In other words if the input voltage is not returned to a plate supply bleeder but has a terminal on ground potential, the voltage difference between E_3 (anode voltage of tube T_1) and E_B (plate supply voltage) is independent of variations of the plate supply voltage.

It may be interesting to calculate to which point on the supply bleeder the input voltage has to be connected in order that the anode voltage of tube T_1 be independent of supply changes. In that case indeed the two output terminals are the anode of tube T_1 and the ground. (It is sometimes very desired that one of the output terminals is the ground).

We find:

$m = 0$ (because $E_D = m.\ E_B$ has to equal zero)

and the condition gives:

$$2\,r_p = R_3\,(\mu n - 1)$$

$$\text{or}\quad n = \frac{2r_p + R_3}{\mu R_3}$$

provided $r_p = \mu R_2$

Note also that the discussions in sections A and B above do assume the variations to be small in order that the vacuum tube formula may be used: the balance hence holds over only a limited range of filament supply and E_B supply variations.

C.) Instead of using two tubes as in sections A and B the compensation for variations in the filament supply and the plate supply can also be obtained by only using one tube with more than one control grid.

A possible circuit[5] which is claimed to have low drift and to have accurate balance for E_B and tube internal resistance for short-time use is shown in Fig.41 and is called the "Electrometer Tube circuit".

D.) Compensation for changes in emission whether due to heater voltage change or to a random effect can be obtained[5] in a pentagrid type of tube as shown in Fig. 42. The value of R_1 can be adjusted to make grids 1 and 4 have equal but opposite transconductance to the plate. Therefore the compensation includes all circuit elements common to both grids, but does not include the plate supply. In this circuit, as in the cathode-compensated circuits, with the two control grids having equal but opposite effects on the plate current, signals can be applied to either grid, or to both at the same time to obtain a differential amplifier.

If grid 4 is returned to a tap on a bleeder across the plate supply, correction for changes in E_B will be obtained in a manner similar to that discussed in section B.

E.) Another problem arises when currents are to be measured: an error occurs because the grid current is not exactly zero. Furthermore the grid current is dependent upon the grid voltage. How then can we account for this error ?

The simplest way to reduce the error is by using negative feedback as in Fig. 6 which is shown in Fig. 43 in an actual circuit (27). This circuit has been discussed earlier (Fig.6) and it has been found that the output voltage and the input current where related by:

$$e_o = \frac{A}{\frac{1}{R_{id}} + \frac{1}{R_1}\,(1 - A\beta)} \cdot i_i$$

where R_{id} = input resistance of the first vacuum tube.

Obviously by making $|A\beta|$ very high we get:

$$e_o \cong - \frac{R_1 \cdot i_1}{\beta}$$ which is independent of any

characteristic of the input tube. It was also seen that the input resistance of the amplifier with feedback is :

$$R_{if} = \frac{R_1}{R_1 + R_{id}(1 - A\beta)} \cdot R_{id}$$

which is very low if $|A\beta|$ is very high.

To make the magnitude of the product $|\beta A|$ as high as possible we may wish to set $\beta = 1$. However if the magnitude of $|\beta A|$ is very high even when $\beta < 1$, then it may be desired to use a value of β which is smaller than one: in that case the output e_o for an input i_i will be higher than when $\beta = 1$.

It may be interesting to note that the circuit of Fig. 43 closely resembles an operational amplifier. Particularly when $\beta = 1$ the circuit is an operational amplifier with current input. The grid of the first amplifier tube is a virtual ground (its voltage = 0) and hence the grid current will be very small. This is really what we want.

In Ref.(27) are given the following precautions which have to be taken to keep the drift of the zero in direct-coupled tube amplifiers to a minimum :

(a) Heater and plate supplies schould be stabilized.

(b) Push-pull (i.e. differential) amplification is necessary in the early stages of the amplifier.

(c) Normal receiving tubes of medium transconductance generally give improved stability with slightly reduced heater voltage after aging for about 100 hours.

(d) Ambient-temperature changes should be kept to a minimum.

(e) Resilient mounting of the first tube is desirable.
(f) Overloading of the first tube should be avoided.
(g) The cathode should be allowed to reach operating temperature before other voltages are applied and it should not be allowed to fall below this temperature until these voltages are removed.

The above compensating circuits are characteristic for vacuum tubes and constitute direct-compensating circuits. In section 3 of this chapter we will show a few compensating methods which can be used for vacuum tube as well as for transistor circuits. The latter methods will be seen to be based upon an indirect principle : instead of eliminating the error by proper circuit design the error is sensed and a signal is injected to compensate for this error.

2. Transistor circuits

Drift in direct-coupled transistorized d-c amplifiers is mainly due to the following changes in the transistor characteristics[69] : (Fig.44)

(a) An increase in i_{co}. If an increase Δi_{co} occurs in the value of i_{co}, the base current must be reduced by an amount $\Delta i_{co}/\alpha$ to keep the collector current constant. (small letters for quantities are quiescent values here).
(b) A decrease in the emitter-base voltage, v_{eb}. A decrease of Δv_{eb} in the value of v_{eb} causes an extra current $\Delta v_{eb}/R_s$ in R_s, and a similar current must be supplied to the base to oppose this effect.

A third cause of drift[78] is the variation of the d-c current transfer ratio α . However the influence of the two other causes is much greater than the one due to α .

The total compensating current then is (by neglecting changes of α):

$$\frac{\Delta i_{co}}{\alpha} + \frac{\Delta v_{eb}}{R_s} \qquad (69)$$

By definition this is the magnitude of the drift. The relative importance of the two main sources of drift depends on R_s. An example[69] is a typical OC 71 transistor with

$$\Delta i_{co} \cong 50\ \mu A \text{ (for } 20 - 50^{\circ}C)$$

$$\Delta v_{eb} = 100\ mV$$

Thus if R_s = 2 kilohms the two components of drift are approximately equal.

Let us show now that

$$\frac{\Delta i_{co}}{\alpha} + \frac{\Delta v_{eb}}{R_s}$$ is really the

value of the compensating current. We refer to Fig.45. Note that the arrows are in the same direction as the ones of Fig.44. Although these directions are not all the conventional ones we use them because:

(1) they are the directions of positive current
(2) they were used in the original reference[69].
(Also the direction of v_{eb}!)

Obviously (small letters give quiescent values.):

$$R_s . i_b = v_{eb} - e_i$$

Furthermore $$i_c = i_{co} + \alpha . i_e$$

and also $$i_e = i_c + i_b - i_p$$

where i_p is the compensating current (Fig. 45).
Eliminating i_e and i_b between the three equations we get :

$$i_c = i_{co} + \alpha \left(i_c - i_p + \frac{v_{eb} - e_i}{R_s} \right)$$

wherefrom:

$$i_c = \frac{i_{co}}{1-\alpha} - \frac{\alpha}{1-\alpha} i_p + \frac{\alpha}{1-\alpha} \frac{v_{eb} - e_i}{R_s}$$

Let us now assume a small variation Δi_{co} of i_{co} and a small variation Δv_{eb} of v_{eb}: then, in order to keep $\Delta i_c = 0$, we have to apply a compensating current Δi_p such that this current will counterbalance the effects due to Δi_{co} and Δv_{eb}. By definition the compensating current Δi_p to be applied gives the magnitude of the drift.

We get from the last equation above:

$$\Delta i_c = 0 = \frac{\Delta i_{co}}{1-\alpha} - \frac{\alpha}{1-\alpha} \Delta i_p + \frac{\alpha}{1-\alpha} \frac{\Delta v_{eb}}{R_s}$$

wherefrom the drift magnitude is found to be

$$\Delta i_p = \frac{\Delta i_{co}}{\alpha} + \frac{\Delta v_{eb}}{R_s} \quad \text{as stated above.}$$

The third cause of drift, variations of α (78), has not had much attention here because the drift due to variations of v_{eb} and i_{co} turns out to be much larger than the one due to variations of α (88).

A.) Now the drift compensating current required by

$$\frac{\Delta i_{co}}{\alpha} + \frac{\Delta v_{eb}}{R_s}$$

could be supplied (69) automatically by a circuit element having a resistance varying appropriately with temperature. For accurate compensation throughout the required temperature range, the compensating element would have to follow closely a particular resistance/temperature law, and to possess a thermal time-constant similar to the transistor. Thermistors and semi-conductor diodes have been used for this purpose, but perhaps the most obvious choice is another similar transistor. The use of a transistor for the purpose allows the changes in i_{co} and v_{eb} to be compensated separately, and also provides the correct thermal time-constant.

The best way to use a second transistor to compensate for changes of i_{co} and v_{eb} in the first transistor is to build the circuit in the form of a bridge with the two transistors in adjacent arms of the bridge. This will be shown in chapter 7 where we will discuss bridge-balanced d-c amplifiers.

B). A more fundamental way of preventing the drift[69] caused by the change of i_{co} would be to eliminate i_{co}, and this can be achieved by reducing to zero the collector-base voltage which causes it. The collector slope resistance however decreases rapidly as v_{cb} is reduced towards zero, and this effect is magnified when the temperature is increased[69]. With low-frequency germanium transistors the slope resistance (= dynamic resistance) becomes so low at the higher temperatures that the stage gain is reduced to a value which precludes their use in this manner[69].
The collector slope resistance on the other hand is a function of i_{co}. With transistors having lower values of i_{co} and hence higher collector resistances, it should be possible to operate direct-coupled amplifiers in this region of zero i_{co}[69]. This is particularly true for the better silicon transistors.

C.) A differential amplifier featuring excellent drift characteristics (due to well-matched transistors) has been presented by Slaughter[47]. The basic circuit is given in Fig. 46.

The first stage of the amplifier consists of a differential amplifier with well-matched transistors. The output of this first stage is fed into an ordinary d-c amplifier the output of which is fed back to the second side of the differential amplifier in order to stabilize the gain.

Hence :
(1) The matched transistors improve the drift characteristic.
(2) The feedback circuit improves the gain stability.
The use of silicon transistors gives good working conditions despite high ambient temperatures.

Let us assume that the gain of the first stage (the differential stage) of the amplifier is A'. Then (Fig.46) the input to the ordinary amplifier is

$$A' (e_1 - e_2)$$ and its output

is $$e_o = A\,A' (e_1 - e_2)$$

However denoting e_2/e_o by β

we find: $e_o = A\,A' \,(e_1 - \beta\, e_o)$

wherefrom:

$$e_o = \frac{A\,A' \; e_1}{1 + A\,A'\beta}$$

where the gain A has to be positive in order that we have negative feedback (A' and β are essentially positive).

If $A\,A'\beta \gg 1$, then

$$e_o \cong \frac{e_1}{\beta}$$

Let us assume next that the ordinary d-c amplifier has some drift e_d referred to its input. Then the input to this amplifier is :

$$A' \,(e_1 - e_2) + e_d$$

and its output:

$$e_o = A\,A' \,(e_1 - e_2) + A .\, e_d$$

but with

$$e_2 = \beta\, e_o$$

we get:

$$e_o = A\,A' \,(e_1 - \beta\, e_o) + A .\, e_d$$

wherefrom :

$$e_o = \frac{A\,A'}{1 + A\,A'\,\beta} \cdot e_1 + \frac{A}{1 + A\,A'\,\beta} \cdot e_d$$

The effect of e_d is seen to be A' times less than the effect of e_1, as expected.

We clearly have here an example where the first stage of the amplifier is very well taken care of as is required for a good drift-free characteristic.

D.) Other combinations are possible but they all result in a d-c amplifier consisting of a first stage which is a d-c differential amplifier with matched transistors (usually silicon transistors to improve the temperature performance). Only the part of the amplifier following the first stage can be slightly different from one device to another:

(1) If some additional stabilization is desired, then the second stage is connected so that it helps to reduce remaining influence of the temperature on the first one. (An example will be seen in D of section 2 of chapter 7).

(2) If gain stabilization is desired then some kind of feedback is used (as for example in C above).

It is evident that use of additional zero stabilization and gain stabilization can be made at the same time.

In the discussion above it has been seen that careful matching and selection of transistors was important. This however does not lend itself to production. Also, variations with age are difficult to predict, which adds to the delicateness of the technique based on matching and selecting.

3. Vacuum-tube or transistor circuits.

The compensation methods that we will discuss now can be used for vacuum-tube devices as well as for transistorized ones.

As a matter of fact the methods to be dealt with now sense the drift from zero (no matter by which phenomenon it is caused) and inject a signal that will reduce this drift.

a. Goldberg circuit[16],[65]

This circuit is based upon a method for automatically stabilizing direct-coupled d-c amplifiers against zero offset voltage and its drift. Stabilization is obtained through the use of a modulated d-c amplifier which senses the error, amplifies it and reinjects it in the main circuit at the proper place. The circuit is such that the stabilization device does not alter the high-frequency response characteristics of the amplifier.

The basic circuit is given in Fig. 47. It consists of a conventional direct-coupled d-c amplifier and a modulated d-c amplifier, the latter being of the type described in chapter 5. The first stage of the direct-coupled d-c amplifier usually consists of a differential amplifier with inputs A and C. The modulated d-c amplifier senses the d-c potential which appears at point A. This d-c voltage is amplified and applied to point C.

The gain A_d of the main amplifier as given in Fig. 47 is the gain from A to the output. Normally A_d has a negative value. The gain from C to the output is equal in magnitude to the gain from A to the output but has the opposite sign. Hence normally the gain from C to the output is positive.

Let e_K = zero offset voltage of the main amplifier as referred to the input.

e_i = input voltage to the amplifier.

e_o = output voltage of the amplifier.

ε = voltage at point A.

e_d = voltage at point C.

A_d = gain of the main amplifier from point A to the output. (The gain of the main amplifier from point C to the output is $-A_d$). Note that A_d is normally negative and hence $-A_d$ is normally positive.

A_m = gain of the modulated d-c amplifier.

Z_i = input impedance (approximately).

Z_f = feedback impedance (Fig. 47).

Note that we consider the impedances to be impedances for complex frequencies (s in Laplace transform theory).

It is seen that

$$e_o = (\varepsilon - e_d + e_K) \cdot A_d$$

and

$$e_d = \varepsilon A_m$$

Also

$$\frac{e_i - \varepsilon}{Z_i} = \frac{\varepsilon - e_o}{Z_f}$$ if we assume

the input impedances of the two d-c amplifiers to be infinite.

Later, however, we will see how to account for these impedances.

Solving the last equation above for ε we get:

$$\varepsilon = \frac{Z_f}{Z_f + Z_i} \circ e_i + \frac{Z_i}{Z_f + Z_i} \cdot e_o$$

Then from:

$$e_d = \varepsilon \cdot A_m$$

we see that

$$\varepsilon - e_d = \varepsilon (1 - A_m)$$

Bringing this into

$$e_o = (\varepsilon - e_d + e_K) \cdot A_d$$

we get:

$$e_o = A_d \cdot \left[\frac{Z_f}{Z_f + Z_i} \cdot (1 - A_m) \cdot e_i + \frac{Z_i}{Z_f + Z_i} \circ (1 - A_m) \circ e_o + e_K \right]$$

from which e_o is found to be :

$$e_o = \frac{\frac{Z_f}{Z_f + Z_i} \cdot (1 - A_m) \cdot A_d \cdot e_i + A_d \circ e_K}{1 - (1 - A_m) \cdot A_d \cdot \frac{Z_i}{Z_f + Z_i}}$$

If $\left| (1 - A_m) \cdot A_d \cdot \frac{Z_i}{Z_f + Z_i} \right| \gg 1$

then

$$e_o \cong - e_i \frac{Z_f}{Z_i} + e_K \frac{Z_f + Z_i}{Z_i (A_m - 1)}$$

The first term on the right-hand side is a desired one, but the second term is the zero offset voltage term. Note that in order to reduce the zero offset voltage, it is only necessary to make A_m large at d-c. A_m may be made large enough so that the second term of the last equation above is negligible, in which case the gain is:

$$\frac{e_o}{e_i} = - \frac{Z_f}{Z_i}$$

which states that the gain is the opposite of the ratio of the feedback impedance and the input impedance and that there is no d-c offset. Note that this corresponds to an ideal operational amplifier. Indeed this circuit is the basic form of the d-c amplifiers used in many analog computers.

If the error sensed is positive it means that the value of the output is too positive (or not enough negative) and hence the circuitry must be such that a positive error makes the output less positive (or more negative); also a negative error shall make the value of the output more positive (or less negative). This means that the overall gain of the error via G_2 and G_1 should be negative. Hence for stability

$$A_m \cdot (- A_d) \text{ must be } < 0$$

or

$$A_m \cdot A_d > 0$$

Usually A_m and A_d are both negative.

A practical Goldberg circuit is given in Fig. 48.

A cause of trouble remaining is the input impedance into the main d-c amplifier (G_1 in Fig. 48). In the fundamental theory this impedance was assumed to be infinite but this is, of course, not true in practice. For transistorized amplifiers this impedance can even be rather low, whereas for vacuum tubes it is usually very high but not always high enough. (for vacuum tubes this input impedance or resistance is the ratio of the applied grid voltage and the grid current).

The circuit can be modified a little to take account of this undesired effect. We get Fig. 49 as an example. Now the auxiliary d-c amplifier G_2 senses the potential of point B. The ratio of impedances Z'_i/Z'_f is equal to the ratio of impedances Z_i/Z_f. If now the voltage ε_B at point B is not zero then a correction voltage will appear at point C and change the output so as to make ε_B = zero (or anyway very near zero). In that case:

$$\frac{e_i}{Z'_i} = -\frac{e_o}{Z'_f}$$

and

$$e_o = -\frac{Z'_f}{Z'_i} \cdot e_i = -\frac{Z_f}{Z_i} \cdot e_i$$

as desired.

What we really do here is the following: we amplify e_i by means of d-c amplifier G_1. Then we compare output against input by means of the separate network consisting of Z'_i and Z'_f. If the output is not what we want it to be, a correcting signal is introduced at C. Obviously we no longer have to worry about errors due to G_1 for it is the circuit consisting of Z'_i and Z'_f which is the sensed one now. It is clear however that in this case also it is required that the gain factors of G_1 and G_2 be as high as possible.

Noteworthy is that the input impedance of G_2 does not matter much, provided the gain of the latter is high enough. Indeed whether or not there is an error does not depend upon the input impedance of G_2, only its magnitude does and since we assume G_2 to have a very high gain the magnitude of the error turns out not to be very important.

As in any compensating method, also in the Goldberg circuit, residual errors are possible. For example, the auxiliary d-c amplifier will certainly have a certain offset, no matter how small, so that its output will never be exactly its gain factor times the input as was assumed in the ideal case. It is true however(88) that a very good stabilization is obtained by this circuit. If transistors are used then by a combination of silicon transistor differential amplifier and chopper stabilization, a d-c amplifier can be built with the use of transistors which have normal production spreads of v_{be} temperature variations, and resulting in a drift referred to the input of less than

500 microvolts[88]. By selecting matched v_{be} temperature characteristics the drift can be reduced much farther.

The input impedance of the circuit is approximately equal to the parallel combination of Z_i and Z'_i. (Note that the points A and B are virtual grounds). Hence if the input impedance has to be high the values of the impedances Z_i and Z'_i should also be high. However there may be other considerations (e.g.impedance level, gain factor,...) which call for low values of Z_i and Z'_i. A compromise is then necessary.

b. Owen-Prinz circuit.[9],[19],[65],[71]

A block diagram of the basic Owen-Prinz circuit is shown in Fig.50, in which A represents the internal gain (=gain if the load were not present) of the direct-coupled d-c amplifier[71]. For proper operation it is essential that the amplifier to be drift-corrected causes a net phase-reversal of the signal; in other words, if normal methods of coupling are used, the amplifier must contain an odd number of stages. Another requirement of the amplifier is that the available output-voltage swing shall embrace zero; this will usually require potentiometer coupling to the output circuit, using a negative as well as positive d-c supply.

Let us now examine the properties of the circuit. We assume some offset to be present which we denote by e_d. We then get Fig.51 where we assume the direct-coupled d-c amplifier to be ideal but having at its input a voltage source giving the equivalent offset.

Let the voltage across capacitor C be given by e_c as shown in Fig.51. The input and output switches are syn - chronous. In one part of their period (the one shown in Fig. 51) they provide a path for drift correction of the amplifier. In the other part they provide a path for amplification of the input signal.

The principle of the circuit is to make e_c equal to e_d so that these two voltages will cancel when the input is measured. How is this obtained? In the position shown in Fig.51 the output of the amplifier must be $-e_c$. Its input is $e_d - e_c$. It then follows that:

$$-e_c = A\,(e_d - e_c)$$

or

$$e_c\,(A - 1) = A.\,e_d$$

wherefrom:

$$e_c = \frac{A}{A - 1}\,.\,e_d$$

We see here why the amplifier has to have an output-voltage swing which embraces zero: the drift can be either positive or negative so that the output of the amplifier should also have the possibility to be either negative or positive.

If A is infinite then it turns out that $e_c = e_d$, which is exactly what we want. However, A cannot be infinite so that the remaining equivalent drift is now:

$$e_d - e_c = e_d\,\left(1 - \frac{A}{A - 1}\right)$$

$$= e_d\,\left(\frac{-1}{A - 1}\right) = \frac{e_d}{1 - A}$$

Obviously the drift is reduced by a factor $1 - A$.

The physical reasoning behind the circuit is simple. Suppose that $e_c = 0$. We sense the drift e_d. Since in the position shown in Fig.51 the output is connected to the capacitor C the capacitor will tend to be charged to $A.\,e_d$. If A were positive this would enhance e_d and the equivalent drift would increase. We do not want this. Therefore A has to be negative. In that case we will see the capacitor charged up to a voltage which opposes the original drift e_d. It is clear that if the capacitor is charged up to a certain voltage e'_c the output of the amplifier will tend to charge the capacitor to a voltage $A(e_d - e'_c)$. The total equivalent drift when a steady-state condition for e_c is reached is given by :

$$e_d - \frac{A}{A - 1} e_d = \frac{e_d}{1 - A}$$

as seen above.
Since A has to be negative this remaining equivalent drift is:

$$\frac{e_d}{1 + |A|}$$

for note that A is real because it is the gain of a d-c amplifier.

In the next part of the switching period the input e_i in series with the remaining equivalent drift is amplified so that:

$$e_o = A \left(e_i + \frac{e_d}{1 + |A|} \right)$$

$$= - |A| \; e_i - \frac{|A| \; e_d}{1 + |A|}$$

Obviously the effect of e_d is considerably reduced if $|A|$ is very large.

To maintain the correct capacitor charge against leakage and against changing of the drift, frequent operation of the switches is necessary. This necessitates frequent interruption of the signal channel which is very undesirable. As a matter of fact the necessity of frequently interrupting the signal channel is the main objection to the Owen-Prinz system.

Usually(71) the Owen-Prinz method is only applied to the first stage of a direct-coupled d-c amplifier which consists in its basic form of several identical (or almost identical) amplifier stages connected in cascade in the normal fashion. Let us as an example consider here an amplifier consisting of only two identical stages. In the absence of correction, the effects of supply voltage variations and temperature changes would be approximately the same in each stage; but referred to a fixed point in the amplifier, such as the input terminal, variations in the first stage would predominate by a factor equal to the stage gain, since they suffer a further stage of amplification. However, the Owen-Prinz method is applied to the first stage only. The zero variations in that stage are thus reduced by a factor equal to the stage gain (about!); in their ultimate effect, therefore, they are now approximately equal to the variations in the second stage. Because of the phase reversal in each stage, the two variations are also opposite in sign; they therefore tend to cancel out.

In order to improve the Owen-Prinz method two modifications have been proposed[71]:

(1) The cascade-balance system.

(2) The reflex-monitor system.

1. The cascade-balance system.[71]

Since as seen above the drift reduction is usually confined to the first stage of the d-c amplifier, the disadvantage of the periodic interruption of the signal channel may be overcome simply by duplicating the first stage of the amplifier, it being arranged that the two counterparts are taken out of service for correction alternately.

A block diagram of the system is shown in Fig.52, where A and B are identical stages. The automatic switching must be so arranged that neither of stages A and B is taken[71] out of service for correction until the other is restored to service; in other words, there must be a short "overlap" period in which both stages are operating in parallel.

This system is such that overall d-c negative feedback in order to stabilize the gain is possible.

A complete practical cascade-balance amplifier is discussed in Ref.[71]

Although the cascade-balance system has the major advantages of high stability and wide frequency response, the possible drawbacks to the system are[71]:

(1) it involves moving parts.
(2) the primary balance is a double one, somewhat complicating initial installation.
(3) the first stages A and B should be similar in design to the first stage of C (to get optimal balancing of drifts as seen before) possibly restricting the scope of the system.
(4) the output may contain a small square-wave component (up to 50 μV[71]) at the switching frequency, as the result of a residual unbalance between the input stages.
(5) if a differential input is required the switching tends to become cumbersome.

The first of these is generally accepted as being unavoidable where maximum stability is required. The others have been oliminatod by applying the basic cascade-balance principle to another form of drift-corrected amplifier: the reflex-monitorsystem.

2. The reflex-monitor system[71].

A block diagram of this circuit is given in Fig.53. The essential difference in this system as compared with the cascade-balance system is that the twin input stages are not operated on a time-shared basis. Instead, one stage (A) is permanently allocated to the signal channel, which is now never diverted. The other (B) serves as a monitor, and alternates between two conditions: in the first of these, which may be termed the "self correcting" condition (the opposite of the one shown in Fig. 53) stage B is isolated from the rest of the amplifier and its drift is checked and corrected by the Owen-Prinz capacitor-storage method (capacitor C_2). In the other, the "monitoring" condition (the condition shown in Fig. 53), stage B is used to check and correct the drift in the main amplifier A, again with the use of a storage capacitor (capacitor C_1).

Let us consider the physical working of the whole system. Let us first assume B to be in the "self-correcting" condition which gives Fig. 54 for the circuit of B. Obviously the two inputs are at ground potential and, if the proper connections are made, any drift will charge capacitor C_2 to a voltage which will reduce the equivalent drift to a low value (as seen in the basic discussion of the Owen-Prinz method). In the "monitoring" condition, which is the one shown in Fig. 53, we then assume B to be a drift-free d-c amplifier. Amplifiers A and B are set up so that their third input (labeled 3) adds to their first input (labeled 1) (Fig. 53). Let us call the six input voltages to the amplifiers:

to	A_1	: e_{A1}
	A_2	: e_{A2}
	A_3	: e_{A3}
	B_1	: e_{B1}
	B_2	: e_{B2}
	B_3	: e_{B3}

Then we see that

$$e_{A1} = e_{B2}$$

$$e_{A2} = e_{B1}$$

We really want an output e_o which is proportional to the input difference $e_{A1} - e_{A2}$. If now there is some drift in amplifier A or amplifier C, then e_o will not be what we want it to be and the output of amplifier B will therefore not be zero. The output of B conditions the voltage of capacitor C_1 and this voltage will tend to reduce the effect of the drift.

In the main amplifier it is the first stage only which is important from considerations of drift, and it is this stage only which need be identical to the monitor stage if best performance is wanted (we show the truth of this statement below). Whatever form the rest of the amplifier may take, the additional drift introduced by it will be small by comparison. Hence A and B shall be identical conventional one-stage amplifiers.

The correct sense relationships of the system require that the input signal as applied to the monitor stage shall be reversed in polarity with respect to that applied to the main amplifier. That is the reason why

$$e_{A1} = e_{B2}$$

$$e_{A2} = e_{B1}$$

To achieve this without destroying the essential similarity of the input stages A and B, a differential input is necessary in each stage; by simple transposition of the input connections the necessary phase-reversal is effected. The entire amplifier is then differential by nature, though either input terminal may, of course, be earthed to suit single-sided input signals.

Let us now consider a simple mathematical analysis of this circuit. We will not assume that amplifiers A and B are identical, but we will show that they should be for best performance. We assume amplifier A to have a gain A and a drift e_{dA} and amplifier B to have a gain B and a drift e_{dB}. The inputs to the amplifiers are as stated above:

$$e_{A1} - e_{A2} \text{ to amplifier A}$$

and

$$-(e_{A1} - e_{A2}) \text{ to amplifier B.}$$

Let the drift of amplifier C be denoted by e_{dC} and its gain by C. Then we get the equivalent circuit of Fig.55 where A, B and C are assumed to be driftless d-c amplifiers.

In Fig. 55 it is seen that a "self-correcting" circuit exists when the circuit is in the "self-correcting" condition. In that case we find for amplifier B :

$$-e_{C2} = B \left[(e_{dB} - 0) - e_{C2} \right]$$

wherefrom :

$$e_{C2} = \frac{B}{B - 1} \cdot e_{dB}$$

When then the switches are thrown over to the "monitoring" condition we get :

Output of $B = e_{C1} =$

$$B \left[e_{A2} + e_{dB} - e_{A1} - e_{C2} + \beta \cdot e_o \right]$$

with

$$\beta = \frac{R_1}{R_1 + R_2}$$

Hence

$$e_{C1} = B \left[e_{A2} - e_{A1} + e_{dB} (1 - \frac{B}{B - 1} + \beta \cdot e_o \right]$$

$$= B \left[e_{A2} - e_{A1} + \frac{e_{dB}}{1 - B} + \beta \cdot e_o \right]$$

The output of A (Fig.55) is then :

$$e_{oA} = A \left[e_{A1} + e_{dA} - e_{A2} + e_{C1} \right]$$

and the output of C is :

$$e_o = C \quad (e_{oA} + e_{dC})$$

Hence eliminating e_{C1} and e_{oA} from

$$e_{C1} = B \quad (e_{A2} - e_{A1} + \frac{e_{dB}}{1 - B} + \beta \cdot e_o)$$

$$e_{oA} = A \quad (e_{A1} + e_{dA} - e_{A2} + e_{C1})$$

and

$$e_o = C \ (e_{oA} + e_{dC})$$

we should be able to express e_o in terms of $e_{A1} - e_{A2}$ (which, for convenience, we will denote by e_i), A, B, β, e_{dA}, e_{dB} and e_{dC}.

We have

$$e_{C1} = B \left[- e_i + \frac{e_{dB}}{1 - B} + \beta \cdot e_o \right]$$

$$e_{oA} = A \left[e_i + e_{dA} + e_{C1} \right]$$

$$e_o = C \left[e_{oA} + e_{dC} \right]$$

Hence :

$$e_o = C \left\{ e_{dC} + A \left[e_i + e_{dA} + B \quad (-e_i + \frac{e_{dB}}{1 - B} + \beta \cdot e_o) \right] \right\}$$

wherefrom :

$$e_o = Ce_{dC} + CA\ e_i + CA\ e_{dA} - CA\ Be_i$$

$$+ \frac{CAB}{1 - B} \cdot e_{dB} + CAB\beta e_o$$

or by solving for e_o :

$$e_o = \frac{(1 - B)\ AC}{1 - A\ B\ C\ \beta} \cdot e_i + \frac{AC}{1 - A\ B\ C\ \beta} \cdot e_{dA}$$

$$+ \frac{ABC}{(1 - B)\ (1 - ABC\beta)} \cdot e_{dB} + \frac{C}{1 - ABC\beta} \cdot e_{dC}$$

Since for stability the gain B of amplifier B as used here must have a negative value as seen in the basic theory of the Owen-Prinz system we get :

$$e_o = \frac{(1 + |B|\ )\ \ AC}{1 + A|B|C\ \beta} \cdot e_i + \frac{AC}{1 + A|B|C\ \beta} \cdot e_{dA}$$

$$- \frac{A\,|B|\,C}{(1 + |B|\ )(1 + A|B|C\ \beta)} \cdot e_{dB} + \frac{C}{1 + A|B|C\ \beta} \cdot e_{dC}$$

for B is real because it is the gain of a d-c amplifier.

Let us now assume that the drifts of A and C are such that e_o is too positive(or not enough negative) as compared with the driftless case. Then this will make the input to B (= differential input + feedback signal) positive (if we assume the drift of B to be zero). The voltage e_{C1} becomes therefore negative (since the gain of B is negative). This means that if we want this voltage e_{C1} to counteract the effects of drift in A and C the total amplification over A and C, i.e. AC must be positive. Indeed in this case a negative e_{C1} will tend to make the output less positive (or more negative). This shows that the product AC has to be positive.

Let us now consider the formula which gives the output :

$$e_o = \frac{(1 + |B|\ )\ AC}{1 + |B|\,.\ AC\ \beta} \cdot e_i + \frac{AC}{1 + |B|\ AC\ \beta} \cdot e_{dA}$$

$$- \frac{|B|\ \ AC}{(1 + |B|\ )\ (1 + |B|\ AC\ \beta)} \cdot e_{dB} + \frac{C}{1 + |B|\ AC\ \beta} \cdot e_{dC}$$

If $|B| \gg 1$

and $|B|\,.\,AC\ \beta \gg 1$ as usually is the case then :

$$e_o \cong \frac{1}{\beta} \cdot e_i + \frac{1}{|B| \cdot \beta} (e_{dA} - e_{dB}) + \frac{1}{|B| \beta A} \cdot e_{dC}$$

We now see why we want to make the amplifiers A and B identical. Indeed in that case

$$e_{dA} \cong e_{dB}$$

and

$$e_o \cong \frac{1}{\beta} e_i + \frac{1}{|B| \beta A} \cdot e_{dC}$$

The contribution of e_{dC} to the output is a factor $|B| A$ less than the contribution of e_i. Since the product $|B| A$ is made as high as possible we do not have to worry much about e_{dC}.

A simple basic circuit which has three input terminals and one output terminal as required for the reflex-monitor system is shown in Fig. 56. This input circuit is discussed in Ref.(71) and is essentially that of a cascade amplifier with the lower triode triplicated. We will however not discuss this circuit here because our purpose was to explain the basic working of the compensating system and not to discuss every single detail involved in the circuit itself.

4. Circuit using fluid-state tetrodes(105)

Recently(105) a new approach has been taken to the problem of amplifying low-level and very-low-level (microvolt range) d-c voltages by introducing a new type of amplifying device: the fluid-state solion tetrode*.
It is claimed that a pair of such devices can replace the vacuum tubes or the transistors in the first stage of a low-level or a very-low-level direct-coupled d-c amplifier.
The performance is said to be much better than that of vacuum tubes or transistors.

* Developed by Texas Research and Electronic Corp.

Let us consider Fig. 57. The device consists of a glass tube filled with an aqueous solution of a small amount of iodine (I_3^-) and a larger amount of iodide(I^-). If voltage E is zero a chemical equilibrium is obtained between the iodide and the iodine:

$$3I^- \rightleftharpoons I_3^- + 2e^-$$

(e = electron)

This equilibrium is determined by the ionization product of the chemical reaction :

$$I.\ P. = \frac{[I_3^-]\,[e^-]^2}{[I^-]^3}$$

where [ ] means "concentration of". If now the voltage E is added then the electrons and the negative ions will all be driven toward the anode which is the electrode connected to the positive side of the voltage source E.

At the anode iodide (I^-) ions will give up part of their negative charge and will become iodine ions :

$$3I^- \rightarrow I^-_3 + 2e$$

This gives rise to a current i. At the same instant also a current i must exist from the cathode toward the negative side of the voltage source. This cannot happen unless at the cathode the reaction

$$I^-_3 + 2e \rightarrow 3I^-$$

is existing.

For reasons of continuity in the electrical circuit the number of electrons coming in at the cathode must be equal to the number of electrons flowing away at the anode because the currents to and from the voltage source must be equal. Thus it is seen that by virtue of the reaction

$$3I^- \rightarrow I^-_3 + 2e \quad \text{(at the anode)}$$

the same amount of I^-_3 is formed as the amount that disappears by virtue of the reaction $I^-_3 + 2e \rightarrow 3I^-$ (at the cathode.) Hence the net amount of I^-_3 and I^- ions does not change in the glass tube. However let us notice the following : as we stated above the aqueous solution contains a small amount of iodine and a larger amount of iodide. Therefore the concentration of iodide at the anode will permanently be higher than the concentration of iodine at the cathode. This means that if we want to increase the overall current we should increase the iodine concentration. Increase or slight decrease of the iodide concentration will not have much effect for the latter is high as compared with the iodine concentration.

Obviously then the rate of the above reactions and hence the current in the external circuit will be controlled by the amount of iodine ions reaching the cathode.

The characteristics of the solion tetrode then are determined by the diffusion velocity of the iodine ions toward the cathode. If the voltage E (Fig. 57) is low the amount of iodine ions reaching the cathode is such that we get a piling up of ions there. This is equivalent to a space charge in vacuum tubes or a base charge in transistors. If E is increased the current i will increase until this increase is stopped by the fact that the ions cannot diffuse fast enough to allow more increase of current.

In short, it is seen that for low voltages E the current is dependent on the latter; for high values of E however, the current becomes independent of E. This gives characteristic curves similar to the vacuum tube pentode or to the ordinary transistor[105]. For every concentration of iodine ions we get another curve.

The practical set-up of a circuit containing one solion tetrode is as shown in Fig. 58. The[105] output circuit is such that the output current depends upon the concentration of iodine ions in the compartment between R and C (called "integral compartment"). The latter concentration however is made dependent upon the input voltage e_i (or eventually the input current i_i) by means of the circuit consisting of C, S and I. E_s is so chosen that[105] the concentration of iodine ions close to the common electrode C is not independent of the voltage e_i : The input circuit operates in the region where the input current through the device (hence also the amount of iodine ions produced at the electrode C) is still dependent on the input voltage. The output current I_o then is dependent upon the input voltage e_i because it is dependent upon the iodine concentration at C. This current, however, is independent of the voltage between the electrodes R and C because the output circuit is operated in the region (seen above) where the current I_o is only dependent upon the diffusion rate of the iodine ions in the integral compartment (compartment between R and C) and not upon the applied voltage.

Obviously the device yields an output current which is proportional to the input voltage. Note that in the case that the input signal is a voltage the input current is necessary only to maintain the proper concentration of iodine at electrode C. This current is low. If the input is a current then the iodine concentration at C and hence the output current will be proportional to the integral of the input current. In that case the device acts as an integrator.[105]

A practical simple example of a direct-coupled d-c amplifier using a pair of fluid-state solion tetrodes in the first stage is given in Fig. 59[105]. If desired the input of the second solion tetrode (which is connected to ground in Fig. 59) can be used to apply a feedback signal to improve the overall characteristics of the amplifier.[105]

Characteristic curves and the complete scheme of an amplifier built using solion tetrodes are given in Ref.[105]

Typical characteristics that are claimed to apply to amplifiers using solion tetrodes are [105]:

(1) Voltage gain of 500 for the amplifier of Fig.59 if the supply voltages (+ and -) are 10 volts and if R_1 = 8 kilohms

$R_2 = R_3$ = 10 kilohms

(2) Response for the same amplifier is flat from d-c to 0.005 cps. (Note that feedback may be used to increase the bandwidth.)

For a feedback amplifier described in Ref.[105] the following characteristics are said to apply :

(1) An output capability greater than ± 1 ma into a 2 kilohm load is obtained.

(2) Drift is less than 10 microvolts over a 10-day interval.

(3) The temperature coefficient of input zero is approximately 1 microvolt /°C.

(4) Equivalent input noise is 3 microvolts rms.

(5) Spurious input current is approximately 10^{-9} amp.

(6) The input impedance depends on the frequency of the signal and varies from a few megohms at zero frequency to a few kilohms at 3 cps.

(7) The voltage gain is 2700.

(8) The bandwidth is 3 cps.

CHAPTER 7

BRIDGE-BALANCED D-C AMPLIFIERS

A distinction[5] is made between compensated and bridge-balanced d-c amplifiers, for in bridge-balanced circuits all variables in the amplifying arm, including tube (or transistor) characteristics and supply voltages, are balanced against similar variables in a second arm of the bridge. Thus if the parameters of the nonamplifying arm vary in the same manner as those in the amplifying arm, a complete balance is obtained. While it may be argued that no two tubes (or transistors) will balance for all of their changes, it has been found[5] experimentally that such a balance is not so difficult to obtain as usually is thought.

1. Vacuum-tube circuits[5].

The advantages of the use of bridge-balance in vacuum tube direct-coupled d-c amplifiers are that regulation of plate and filament supplies becomes unnecessary except where extreme precision is required, and the balance holds for any part of the tube curves. Operation becomes less critical and more consistent.

A.) The conventional type of balanced d-c amplifier[5] is shown in Fig. 60. It is the equivalent of returning the output voltage to a bleeder in which the bleeder proportioning changes in the same manner as the amplifying tube characteristic. This type will balance perfectly[5] if the two tubes remain identical in characteristics as supply voltages change. It is often used in tube voltmeter circuits[5].

B.) A modification of the above circuit is given in Fig.61. This circuit is found in the Volt-Ohmyst circuit for voltage and resistance measurements. While having some of the features of the cathode-coupled circuits, the main balance is still between the two tubes in parallel, with the common coupling resistor R_1 assisting in maintaining a constant zero position.[5]

C.) Neither of the above circuits can be cascaded, and both are most useful as meter amplifiers for example. Where higher gain is required than one stage will give, the circuit of Fig. 60 can be modified to that of Fig. 62 to allow cascading. The output then is returned to the plate supply bleeder and additional stages can be stepped up the bleeder as seen in chapter 4 in the case of simple cascaded d-c amplifiers. For proper balance the circuit requires that four pairs of resistors (R_1, R_2, R_3 and R_4) as well as the tubes remain constant and it will therefore have slightly more drift than the circuit of Fig.60 under similar conditions.

D.) Improvement can be gotten by shifting T_2 and its cathode resistor (Fig. 62) over to replace the plate resistor of T_1 as shown in Fig. 63 and Fig. 64. Here the balance is no longer for tubes in parallel, but rather in series, and both pass the same plate current at the zero-signal position.[5] The balance for plate supply voltage changes and changes in the internal dynamic resistance of the tubes is more nearly perfect than in Fig. 60, 61 and 62. A differential change in contact potential between the two tubes will shift the zero position as in any other balanced arrangement, but with the same plate current in each tube this effect is much less[5] pronounced than in the parallel systems.

If the value of load resistor R_2 is low, somewhat higher amplification[5] will be obtained with the circuit in Fig. 63 than with the simplified version in Fig. 64. In Fig.63 the load current gives rise to a regenerative (= positive feedback) action across the cathode resistor of T_2. However, for all practical purposes, if a load resistor equal to or greater than the internal dynamic resistance r_p of the tubes is used, there is little difference and the simpler arrangement of Fig. 64 with its fewer parts is preferred.

In both fig. 63 and 64 the cathode resistor[5] of T_1 is usually made variable as a balance adjustment. The range of adjustment is purposely made small: $\pm$ 15 percent from the value of the cathode resistor of the upper tube.
Under these conditions it has been found that if R_1 can be set within this range so that $E_o = 0$ with no input to the grid of T_1, the two tubes are sufficiently well matched for all practical purposes, and hold this balance over very wide changes of filament and plate voltages. When properly balanced in this manner a sensitive voltmeter can be connected across E_o and will show no deflection even when the line voltage supplying filament and plate power is switched off and on[5].

E.) A third arrangement of the series-balanced system is shown in Fig. 65. Here the upper grid is returned to the bleeder instead of to the plate of T_1, and the voltage gain is reduced to slightly less than unity. This circuit is, for example, useful as a current amplifier for meters. Its drift is slightly more than the one of the true series systems of Fig.63 and 64, but is still very low.[5]

In all the circuits we have discussed above it has been assumed that the characteristics of the tubes were the same. This would require selecting of tubes. However, it is said[67] that this selecting and matching of tubes is not necessary, if a circuit is used that matches them artificially.

A possibility of matching two vacuum tubes for filament changes is given in Ref.[67] and is reproduced here in Fig.66. Let us only mention that it is claimed that [67], [48] by adjusting the potentiometer R_3 for equality of anode currents a satisfactory all-round balance (in anode-currents, slopes and heater-supply dependence) can be obtained in a pair of tubes, the other circuit parameters (resistors, etc) being equalized.

2. Transistor circuits

A.) Basically the principle of balancing transistor circuits for common changes in the characteristics of the transistors used is the same as the one used for vacuum-tube circuits: the two transistors are connected in bridge form such that common variations do not appear in the output.

Although transistorized differential d-c amplifiers are not free from drift, it has been pointed out by several authors[78],[88],[69],[47],[101],[102],[110] that such amplifiers can be made to have very low drift if matched pairs of transistors are used . The matching should be made on the basis of equality of the magnitudes of the base-to-emitter voltages V_{BE} of the transistors. Temperature coefficients for I_{co} and α are not considered sufficiently uniform to warrant matching. Effects of I_{co} are also minimized by selecting units with low I_{co} (silicon transistors) at the operating temperature. The base-to-emitter voltage, V_{BE}, decreases almost linearly with temperature at a rate of 2.6 mV/deg. C in silicon transistors and 2.3 mV/deg. C in germanium transistors[78].

Basically a balanced transistorized direct-coupled d-c differential amplifier has a circuit as, for example, given in Fig.67[110]. The transistors used are assumed to have negligible I_{co} values (surface-passivated silicon devices). However drift is still possible[88] due to differences in the variations of V_{BE2} and V_{BE1} and also to differences in the variations of β_1 and β_2 (Complete calculation can be found in Ref[88]). However, if R_E is made very large[110] the main portion of the drift will be due to $V_{BE2} - V_{BE1}$. If then we match the transistors for V_{BE} the drift can be made rather low (down to 50µV/°C is claimed.)

B.) If adjusting possibility is required the circuit can be provided with adjusting potentiometers as for example[69] in Fig.68.

C.) It is not impossible to use negative feedback in circuits as those above. A part of the output is then fed back to the input as for example[69] in Fig. 69.

D.) A rather stable balanced differential amplifier without overall negative feedback is shown in Fig.70[110]. It consists of a pair of matched transistors (T_1 and T_2) and two other transistors the latter providing some additional stabilization of the zero. The only purpose of the lower transistor in the circuit is to provide an almost constant current source .

An analysis of this circuit of which a basic half part is shown in Fig. 71 is given in Ref.[110] and is reproduced here :

If I_{B1} is the base current of T_1, then $h_{FE1} \cdot I_{B1}$ is the current from the base of T_3. Therefore, the collector current of T_3 is $h_{FE3} \cdot h_{FE1} \cdot I_{B1}$. Then, I_2, the current through R_2 , is made up of the emitter current of T_1 plus the collector current of T_3. Therefore :

$$I_2 = (1 + h_{FE1}) \cdot I_{B1} + h_{FE1} \cdot h_{FE3} \cdot I_{B1}$$

$$\cong h_{FE1} \cdot h_{FE3} \cdot I_{B1}$$

assuming $1 + h_{FE1} \ll h_{FE1} \cdot h_{FE3}$

Now $$I_{B1} = \frac{e_1 - V_{B1}}{R_s}$$

$$= \frac{1}{R_s} \left[e_1 - (V_{BE1} + V_{E1}) \right]$$

$$= \frac{1}{R_s} (e_1 - V_{BE1} - h_{FE1} \cdot h_{FE3} \cdot I_{B1} \cdot R_2 - V')$$

Solving for I_{B1} gives :

$$I_{B1} = \frac{e_1 - V_{BE1} - V'}{R_s + h_{FE1} \cdot h_{FE3} \cdot R_2}$$

Therefore

$$e_{o3} = V_{cc} - I_{E3} \cdot R_1$$

$$= V_{cc} - h_{FE1} \cdot (1+h_{FE3}) \cdot I_{B1} \cdot R_1$$

$$= V_{cc} - h_{FE1} \cdot (1+h_{FE3}) \cdot R_1 \cdot \left(\frac{e_1 - V_{BE1} - V'}{R_s + h_{FE1} \cdot h_{FE3} \cdot R_2} \right)$$

Assuming $h_{FE1} \cdot (1+h_{FE3}) \cong h_{FE1} \cdot h_{FE3}$

and

$$R_s \ll h_{FE1} \cdot h_{FE3} \cdot R_2 \quad ,$$

then

$$e_{o3} \cong V_{cc} - \frac{R_1}{R_2} (e_1 - V_{BE1} - V')$$

In the same manner we get for the other half of the differential amplifier :

$$e_{o4} \cong V_{cc} - \frac{R_1}{R_2} (e_2 - V_{BE2} - V')$$

and

$$e_0 = e_{03} - e_{04} \cong - \frac{R_1}{R_2} \left[(e_1 - e_2) + (V_{BE2} - V_{BE1}) \right]$$

The pair consisting of T_1 and T_2 should be matched so that $V_{BE2} - V_{BE1} = 0$ over the desired temperature range. Then

$$e_o \cong - \frac{R_1}{R_2} (e_1 - e_2)$$

Note that here again it is assumed that the leakage currents I_{co1} and I_{co2} were negligible. This is usually true within a few percent for surface passivated units as the ones used here.

It is claimed that this circuit has an equivalent drift referred to the input of about $10\mu V/^{o}C$.

We have built this amplifier and have investigated it. Practical results will be discussed in Chaper 9.

E.) It is, of course, also possible to combine balancing by use of bridge form circuits with other compensating techniques(such as for example thermal compensation as seen in chapter 6) to improve the overall working of d-c amplifiers.

Here again we may point out that as far as the matching and selecting of the transistors are concerned, variations with age are difficult to predict, which makes the teohnique based on such matching and selecting rather delicate.

3. Circuit using fluid-state tetrodes[105].

The circuit using fluid-state tetrodes that we discussed in chapter 6 can be considered as containing a combination of two different compensating techniques :

(1) We used the tetrodes to have stable, reliable circuit elements. This means individual compensation of the elements by reducing the drift itself.

(2) The circuit of Fig. 59 (reproduced here as Fig. 72) is in bridge form such that any residual drift, which is the same for both input tetrodes will not show up in the output.

Obviously this circuit benefits by using both compensating techniques and thus may be able to exhibit very good performance.

CHAPTER 8

UNUSUAL CIRCUITS AND TECHNIQUES IN D-C AMPLIFIER DESIGN

Apart from the circuits discussed in earlier parts of this text several special systems have been developed which either measure d-c signals in a somewhat different way than usual d-c amplifiers do or which include a special feature in the usual d-c amplification process. For this reason we have brought these systems together under the heading " unusual circuits". We will discuss them briefly and indicate where more information can be found.

At the end of this chapter we will also mention a special recently developed technique of building circuits in semiconductor blocks and the application of it to d-c amplifiers.

1. Amplifiers using reflecting galvanometers.[27]

A. Photovoltaic - cell type[27].

The schematic of a simple d-c amplifier of this type is shown in Fig. 73. G_1 is a sensitive reflecting galvanometer which acts as the input element and G_2 is the indicating galvanometer. A and B are two similar photovoltaic cells connected in series opposition and placed side by side so that, in the zero position, the light from the mirror of G_1 falls equally on both cells. There is then no current through G_2. When a voltage is applied at the input, G_1 tends to be deflected, illuminating one cell more than the other and, therefore, causing a current through G_2 and the feedback resistor R. The voltage developed across R opposes the input voltage and thus limits the deflection of G_1. This type of feedback is suitable when the instrument is used for voltage measurements because it makes the apparent input resistance rather high. When the instrument is used to measure currents, it is desirable to make the apparent input resistance as small as possible. A circuit like the one is Fig. 74 is then appropriate.

One of the advantages of these systems is that with[27] only a small amount of feedback the deflection sensitivity of G_2 for a given current in G_1 can be greater than that of a much more sensitive single galvanometer. A sensitivity[27] of 10^5 mm/μA was obtained with one instrument.

Although these systems possess the virtue of extreme simplicity, they suffer from the disadvantage of requiring two delicate galvanometers in order to obtain high sensitivity.

It is interesting to note that a very high degree of power amplification is obtained in the system, the source of power being the lamp illuminating G_1.

B. Photo-emissive cell type[22].

By the use of photo-emissive cells and thermionic tubes it is possible to get such a high degree of amplification that the indicating meter can be a robust milliammeter. The schematic of a simple system using photo-emissive cells is shown in Fig. 75. The light from the reflecting galvanometer G falls on two photocells A and B, the voltage at the grid of tube V_4 being dependent on the relative illumination of the cells. The potential difference between the cathode of V_6 and that of the junction of the neon stabilizers V_2 and V_3 is indicated on the meter M. R is the feedback resistor. With this system it is possible to obtain a current amplification[27] greater than 10^5 in the absence of feedback. The upper limit, set partly by the characteristics of the photocells and partly by the excessive power consumption of the lamp, is rarely reached.

A simpler system[27] uses a gas-filled triode and an a-c supply to the amplifier. This makes it possible to produce a unit operated from the mains with very few components.

In all the systems using reflecting galvanometers the limit of sensitivity is set by the galvanometer. When photo-emissive cells are used in conjunction with thermionic tubes, the output current is large enough to operate conventional recorders. The response time of these systems can be made very much shorter than that of the galvanometer, but all systems using a galvanometer are prone to mechanical disturbances and cannot be used at the limit of sensitivity unless a specially constructed supporting bench is provided.

2. Modulated d-c amplifier with special output circuit[80],[90]

In ref.[80],[90] F. Offner discusses the use of a special output circuit for modulated d-c amplifiers. The circuit is shown in Fig. 76.

Because of the two capacitors in the output circuit effects of non-constancy and inequality of the chopper dwell times will not matter. Indeed, let us first consider the conventional chopper circuit of Fig. 77. If the chopper dwell times are not equal, the output will contain a square wave ripple of modulating frequency. This results from the fact that the d-c voltage component of the secondary of the output transformer must be zero. The a-c voltage is then given in Fig. 78. T_1 and T_2 are the dwell times. Since the d-c component must be zero we get :

$$\frac{E2}{E1} = \frac{T1}{T2}$$

At the output capacitor in Fig. 77 we then get a voltage of the form shown in Fig. 79.

Consider now again Fig. 76 and assume that the waveform of Fig. 78 also appears at the output transformer terminals in this case. Since now the voltage across C_1 is say E_1 and the one across C_2 is E_2, the total voltage at the output is $E_1 + E_2$ and this sum is independent of the dwell times, but is determined only by the input signal because it is the peak-to-peak voltage of the output signal of the transformer (see Fig. 78).

3. Circuit utilized to compensate for transistor parameter variations in modulated d-c amplifiers using choppers[90]

The basic (patented) circuit is given in Fig. 80. It consists of a direct-coupled transistor amplifier with transformer-coupled chopper input and output circuits. The amplifier stability is maintained by two feedback circuits. Operating point feedback (patented) is obtained from the voltage drop across R_1. The d-c voltage across R_1 is fed to the base of Q_1 holding the voltage across the resistor, and thus the current through the output stage Q_2, almost constant. This substantially eliminates the effect of transistor parameter changes on amplifier operating point.

The gain is stabilized by the use of a large amount of inverse feedback, taken from the tertiary winding of the output transformer. Sufficient feedback is employed to hold the gain within the desired accuracy limits.

4. Circuit to increase the input impedance of chopper d-c amplifiers[80]

Let us consider the circuit shown in Fig. 81. It is seen that the input transformer is brought within the feedback loop. Two additional windings are added to the output transformer. These windings are used alternately for feedback, as the chopper changes from one side of the input circuit to the other and this serves to maintain the feedback signal in proper phase relationship.

Input impedances of several megohms are usually obtained[80] with this arrangement.

5. Circuit to compare two electrically isolated small voltages[32],[62]

When two small voltages are to be compared with no electrical connection between them allowed (for example, if their absolute potentials are different) then a circuit like the one shown in Fig. 82 can be utilized. The circuit uses two synchronized switches which alternately connect capacitor C across e_1 and across e_2.

When C is connected across e_1 it is charged to this voltage. Then when C is connected across e_2 its charge is changed so as to make its voltage now equal to e_2. This change of charge gives rise to a current in galvanometer G and this current is a measure of the voltage difference $e_1 - e_2$.

It is seen that the absolute voltage difference V between the sources e_1 and e_2 does not matter, as the latter are never in direct electrical contact.

A complete discussion and the rigorous mathematical analysis of this circuit can be found in ref.[32], [62].

6. Circuits to measure d-c voltages by null-detecting methods.

A d-c voltage can be measured by comparing it to another accurately known voltage. The difference of the two is sensed by some instrument or by the human eye and the known voltage is adjusted so as to make this difference equal to zero. The adjusting device then will allow us to determine in one way or another what the value of the known voltage is that just cancels the unknown voltage, hence it will also give us a measure of this unknown voltage.

The most commonly used type of d-c measuring system based on this principle uses a slidewire potentiometer as, for example, in Fig. 83. The slider is moved until the detector reads zero. The position of the slider can then be calibrated directly in terms of the unknown d-c input.

Many recorders are based on this principle: the detector signal is amplified and applied to a motor which moves the slider to the detector null.

The circuit shown in Fig. 83 is a simple one. Usually a standard cell is inserted somewhere in the circuit to allow for calibration. A circuit with a standard cell is shown in Fig. 84. To calibrate the device the standardizing switch is thrown over so as to disconnect the input and to insert the standard cell into the circuit, whereafter R is adjusted so as to give zero current through the detector with the slider in a certain position.

Practical circuits and more discussion can be found in Ref.[8]

7. Circuit using two modulated d-c amplifiers in series[34].

In ref.[34] a d-c amplifier is discussed which is claimed to precisely measure low-level d-c voltages. The principle of the circuit is shown in Fig. 85 wherein the three choppers are accurately synchronized, the one labeled 3, however, being 180 deg. out of phase with the two others as seen in Fig. 85.

The first modulated d-c amplifier consists of the first part of the circuit up to the capacitors C_1 and C_2.

It is seen that across the capacitors d-c voltages E_{C1} and E_{C2} appear, the difference of which is proportional to the difference of the input voltages E_K and E_i. However, the capacitors C_1 and C_2 are very large so that all the noise at frequencies other than the exact modulating frequency vanish in the output signal of the first amplifier (if this output is defined to be the difference $E_{C1} - E_{C2}$). This differential output signal now is amplified again via a second modulated d-c amplifier. The noise does not matter now because its level is much lower than the level of the differential input signal, the latter being the output of the first amplifier.

This circuit can easily be used as a null-detector[34] wherein the known input voltage E_K is adjusted such that the overall output is zero.

It is claimed[34] that the performance of this circuit is excellent.

The input resistance can be up to 1 megohm.

The circuit can recognize voltage differences of the order of 1 microvolt.

Complete mathematical treatment and discussion of possible applications (voltage measurements, current measurements, resistance measurements,......) are given in Ref.[34]

8. Building of d-c amplifiers in semiconductor blocks.

A recently developed method of building circuits consists of using a semiconductor block in which the different circuit elements are produced simply by treating the appropriate parts of the block in a specific manner so as to give them the properties of the elements they are assumed to be a substitute for in the actual circuit. In this way resistors, transistors and other elements can be built directly into the block.

An example is given in Ref.[104] where the building of a direct-coupled d-c amplifier in a silicon block is discussed. However, the technique is not advanced enough yet to allow highly accurate d-c amplifiers to be built in this way.

CHAPTER 9

A TRANSISTORIZED DIRECT-COUPLED LOW-LEVEL D-C AMPLIFIER

In D. of section 2. of chapter 7 a transistorized differential amplifier was discussed which[110] is claimed to have high stability and can be used to detect d-c signals in the sub-millivolt (= about 100 microvolt) range. The circuit is given in Fig. 70, but for convenience reproduced in Fig. 86.

The amplifier consists of a modified differential amplifier. The purpose of the modification is to yield compensation for temperature variations of the gain of the transistors used (as we have seen in chapter 7).

We have built an experimental circuit and the purpose of this chapter is to discuss the results obtained and to investigate the possibility of using this amplifier as a practical unit.

First it is to be noted that for the first amplifier stage consisting of the transistors T_1 and T_2 we used a single package containing the two transistors: these units constituted a pair matched for V_{BE} variations as explained in D. of section 2 of chapter 7 of this text. It was seen indeed that the only remaining disturbing factor in the circuit was the difference $V_{BE1} - V_{BE2}$ between the base-emitter voltages of the transistors T_1 and T_2 (in first approximation). The output was found to be

$$e_o \cong - \frac{R_1}{R_2} \left[(e_1 - e_2) + (V_{BE2} - V_{BE1}) \right]$$

where R_1 and R_2 are two circuit resistors and e_1 and e_2 are the input voltages. Obviously if the two transistors T_1 and

T_2 are matched for V_{BE} then the output is

$$e_o \cong - \frac{R_1}{R_2} (e_1 - e_2)$$

and is seen to be independent of any transistor characteristic (in first approximation) . This, of course, assumes the temperatures of the two transistors to be equal, which is a good assumption because they are in very close proximity in the package. Let us also notice that it is assumed that the leakage currents I_{co1} and I_{co2} of the transistors T_1 and T_2 are negligibly small. This is approximately true for surface-passivated silicon devices such as the 2N 1613 transistors that are recommended for this circuit (110).

The practical circuit we built is given in Fig.87 where it is seen that we connected the second input to ground so that $e_2 = o$.

The input e_1 was obtained by means of a stabilized voltage which was passed through an attenuator. The stabilized voltage was 28.00 volts and the attenuator allowed attenuation in steps of 1 decibel from 0 to 110 db. The attenuator's output resistance (hence, the " internal resistance of the source") was about 300 ohms for all values of attenuation.

The potentiometer in the circuit was adjusted for zero output voltage for the input voltage being zero at room temperature.

Amplifying characteristics, noise, offset and drift

We measured the output voltage as a function of the input voltage, the latter being varied between zero volt(when the input of the attenuator is short-circuited) and 4.44 millivolts (attenuation of 28 Volt by 76 db.)
The results are given in Table 10.1 and plotted in Fig.88 under the heading " July 19th, 1962" the date that we made

this measurement. From these numerical values and the corresponding plot we learn :

a. that for input values below about 100 microvolts the output values are not worth much

b. the curve representing output versus input may be represented by a straight line with the experimental points forming a scatter diagram around it.

On July 23, 1962, we again measured output versus input and these values are given in Table 10.1, but we have not plotted them in Fig.88 because they almost coincide with the values measured the previous time. In Table 10.1 is, however, also given the difference between corresponding values on the two different days.

We see that the difference in the measured outputs for the same inputs never exceeds 2 millivolts. Since the gain is about 41 as derived from the plots, it is seen that this difference of 2 millivolts is about 50 microvolts when referred to the input.

The next obvious step was to investigate the behavior of the amplifier as a function of the temperature. In order to do this the amplifier was enclosed in an aluminum box and put in a temperature chamber. Since this is an important test we have taken data under identical conditions on four different days notably August 2, 3, 6 and 13, 1962. Every time we waited for the temperature in the chamber to get stabilized respectively at -20°C, -10°C, 0°C, 10°C, 20°C, 30°C, 40°C and 50°C, whereafter we let the input vary from zero up to 62,58 millivolts (by using the attenuator) and measured the corresponding outputs. The results for the four days are given in Tables 10.2, 10.3, 10.4 and 10.5 and contain 59 measurements for each of the 8 temperatures in each case. As an example of what the plots look like we plotted the values of the first day (Table 10.2) in Fig.89 for values of the input up to

9 millivolts. A complete plot of one set (first day, 20°C) is given in Fig.90. This was done in order to show the linearity of the curve over the whole range. The other sets are not given, however, in Fig.90 because they would just overload the diagram.

We see that the curves plotted resemble very well straight lines. Assuming that every particular group of values gives a straight line we then propose to find the line which is most likely to be represented by each group of values, that is: the regression line (= the line for which the total least mean square error for the experimental values of the particular group is minimum). We also want to know what the value of this least mean square error is in order to have an idea of the approaching of the regression line by the experimental values. Since we see that the differences between the output values of different groups are of the same order for low as for high input values, we find it reasonable to consider the least mean square error as the criterion and not some other error as, for example, the relative one.

By use of an IBM 1620 computing machine we find the table 10.6 where

X = input value in millivolts

Y = corresponding output value in millivolts

Coefficient of X = Gain of the amplifier

Value of Y for X zero = Output when the input is zero.

The RMS error (= least mean square error) is given as referred to the input of the amplifier. The results shown in Table 10.6 are quite good. This allows us to draw the following conclusions :

a. The gain of the amplifier is about 41.5. However, the gain is not constant with temperature but varies about linearly from 42.0 for -20°C to 41.3 for +50°C.

b. The root mean square error (giving the scattering of the data about the assumed straight line) is of the order of 50μV, but is lower for the high tempera-

tures and higher for the low temperatures.
Hence this 50μV also gives a measure of the noise in the amplifier.

c. The null offset as referred to the input is clearly dependent upon the temperature. The drift is not a linear function of the temperature, but is minimum at room temperature (about 25μV/°C when referred to the imput) and increases for lower or higher temperatures (about 80μV/°C at -15°C and about 50μV/°C at + 45°C). The reasons for this large drift with temperature are that we have not used any precision resistors and that we have not used any heat sink to equalize the temperatures of the different parts of the circuit either. In ref.[110] it is claimed that by using a heat sink consisting of a bath of silicon oil for the transistors it is possible to get drifts as low as 10μV/°C (when referred to the input.)

If precision resistors are used for R_1 and R_2 of Fig. 86 (Note that the output

$$e_o \cong - \frac{R1}{R2} (e_1 - e_2) \quad \text{is}$$

approximately only dependent on these resistors).
then for R_1 = 14 kilohm and

R_2 = 255 ohm

all the other elements being the same as in Fig. 87, we get numerical values as shown in Table 10.7. It is seen there that the temperature drift is certainly reduced as compared with the case without precision resistors : It is now (on average and when referred to the input) about 60μV/°C for -15°C and about 20μV/°C for -5°C, but for temperatures higher than 0°C it turns out to be negligible.

As for the drift in time, it is rather low so that no change on an output millivoltmeter is observed (even after several hours) if the working conditions are not altered.

Influence of supply voltage changes upon the output.

We varied respectively the positive and the negative supply voltage in steps of 0.5 Volts over 2 Volts at either side of the nominal voltage in order to find the influence of such changes upon the output. Table 10.8 gives the results for changes of the positive voltage while the negative one is kept constant at its nominal value and Table 10.9 for changes of the negative voltage with the positive one kept constant at its nominal value. The influence is measured for several input values. We find that the output changes approximately 1 mV (=25μV when referred to the input) per 0.5 Volt change of the positive supply voltage and 4mV (= 100μV when referred to the input) per 0.5 Volt change of the negative supply voltage.

Final conclusions

The amplifier as it stands now is better than many other available transistorized direct-coupled d-c amplifiers. However, it is obvious that it cannot yet compete with modulated amplifiers which can have stability levels in the micro-volt range, whereas this direct-coupled amplifier is only capable of useful operation from 100 microvolts up. We do not say that it is not possible to make the device work better by using temperature compensating networks and by using more precision resistors than we did, but even then it is not conceivable that the stability would reach the one currently obtained in modulated d-c amplifiers.

An advantage of the amplifier over modulated ones is that the bandwidth is not adversely affected by modulation techniques.

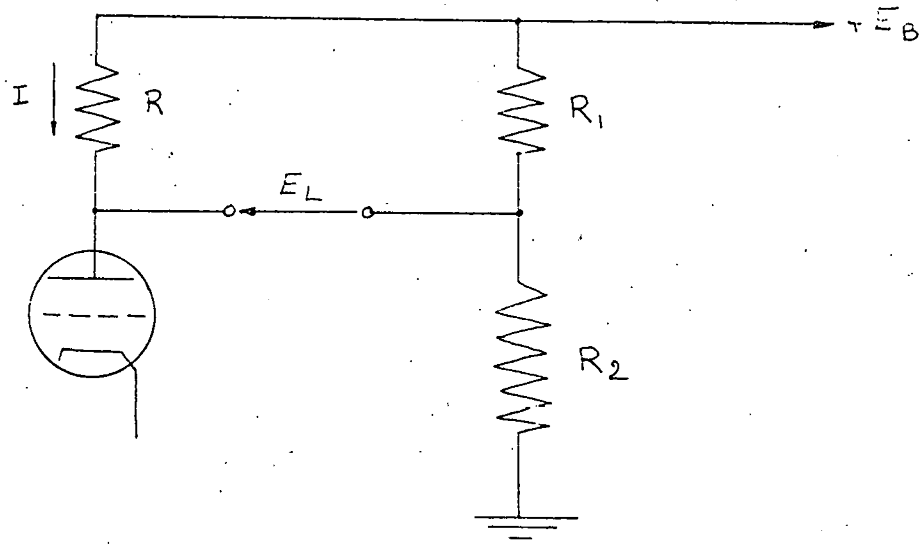
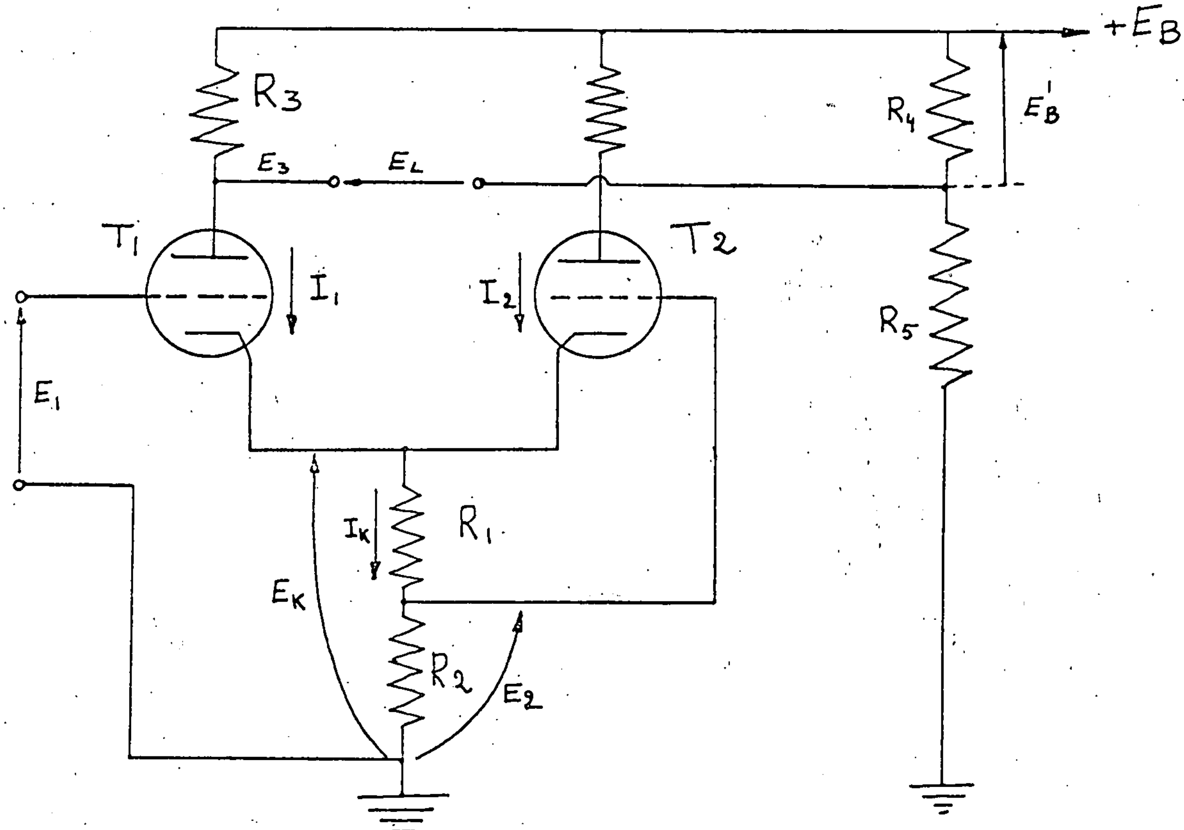


FIG. 37



Miller compensating circuit

$$E_L = E_B' - \frac{\mu R_3}{2r_p + R_3} \quad \text{when } \mu R_2 = r_p$$

FIG. 38

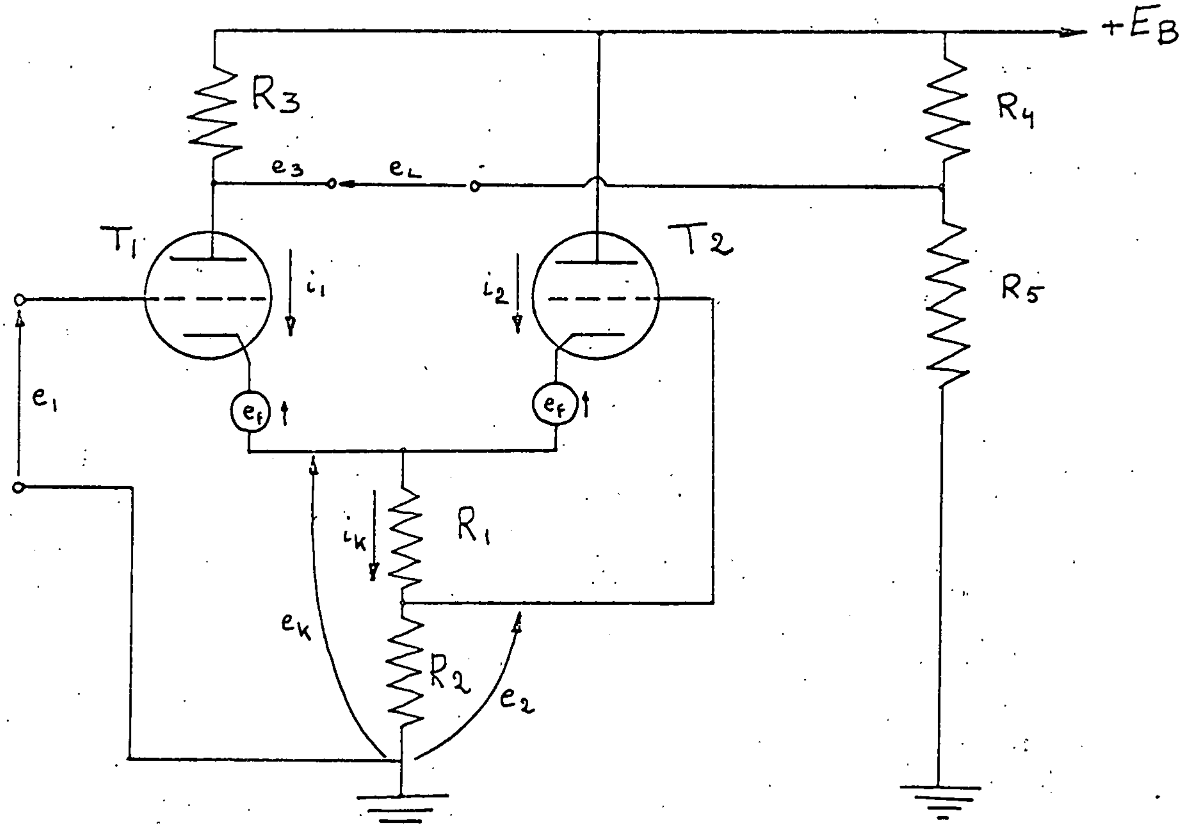
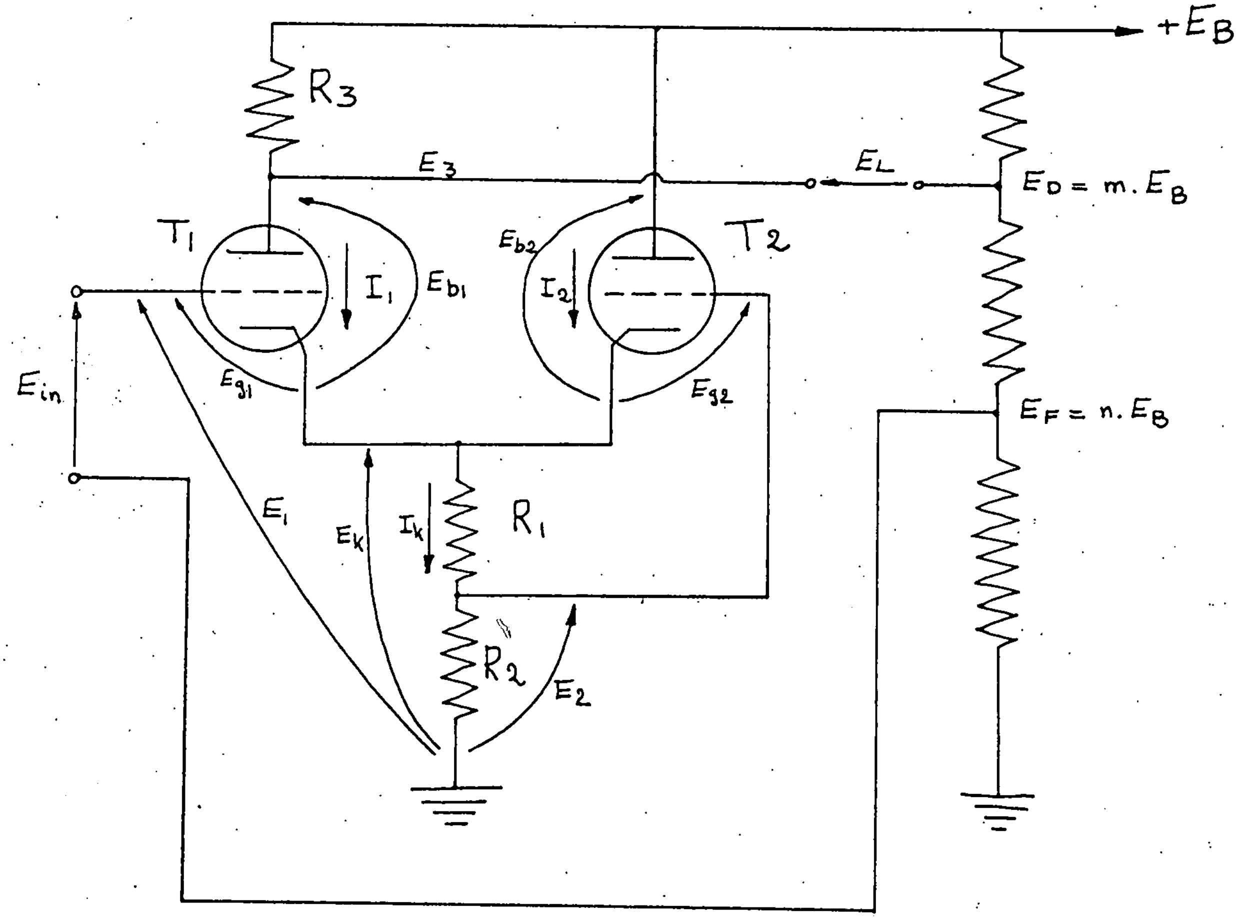


FIG. 39

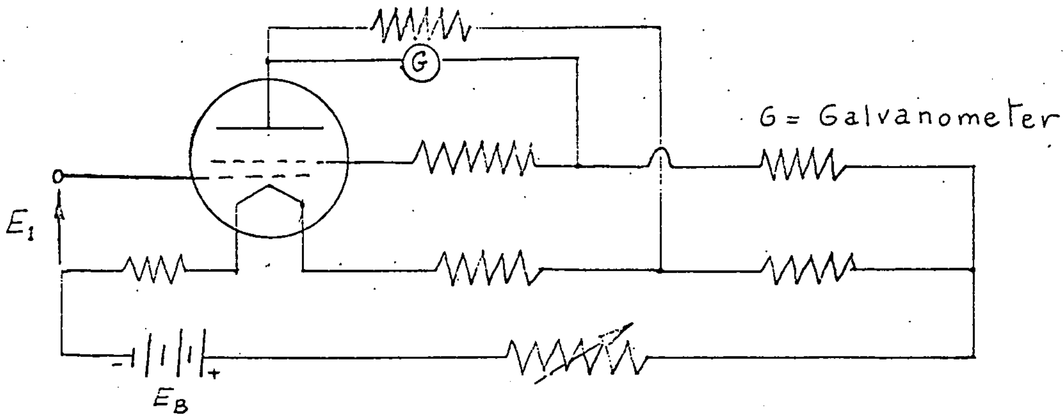


$$\frac{E_L}{E_1} = \frac{-\mu R_3}{2 r_p + R_3} \qquad \text{when } \mu R_2 = r_p$$

and m and n are related by:

$$(2 r_p + R_3)(1 - m) = \mu n R_3$$

FIG. 40



Electrometer Tube Circuit

FIG. 41

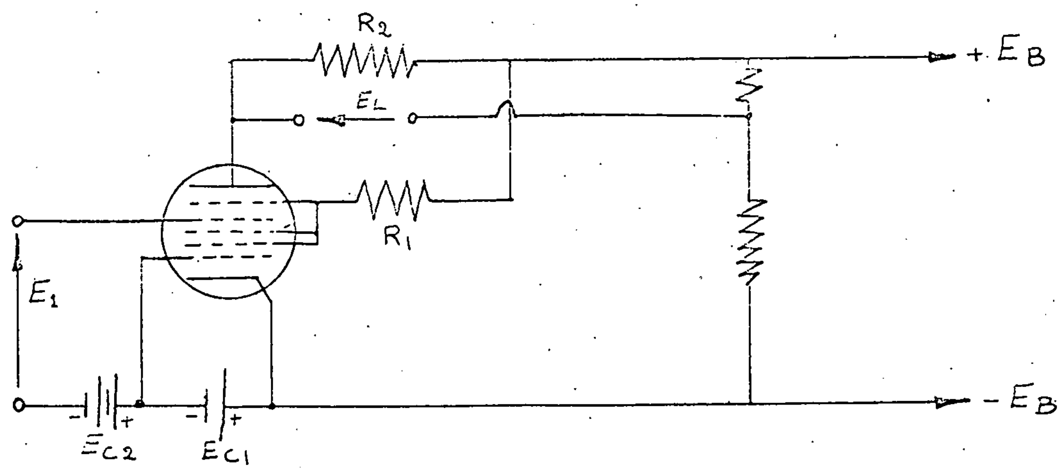


FIG. 42

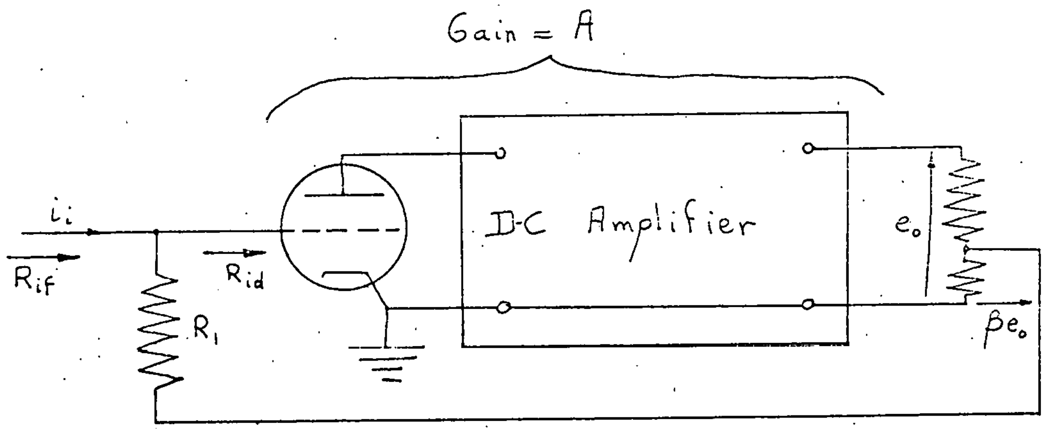


FIG. 43

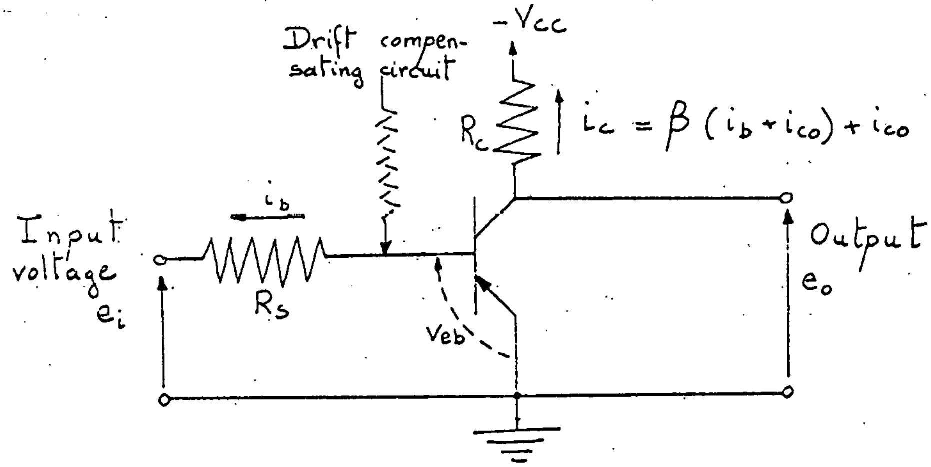

Drift compensating circuit
-Vcc
R_c
$i_c = \beta (i_b + i_{co}) + i_{co}$
i_b
Input voltage e_i
R_s
V_{eb}
Output e_o

FIG. 44

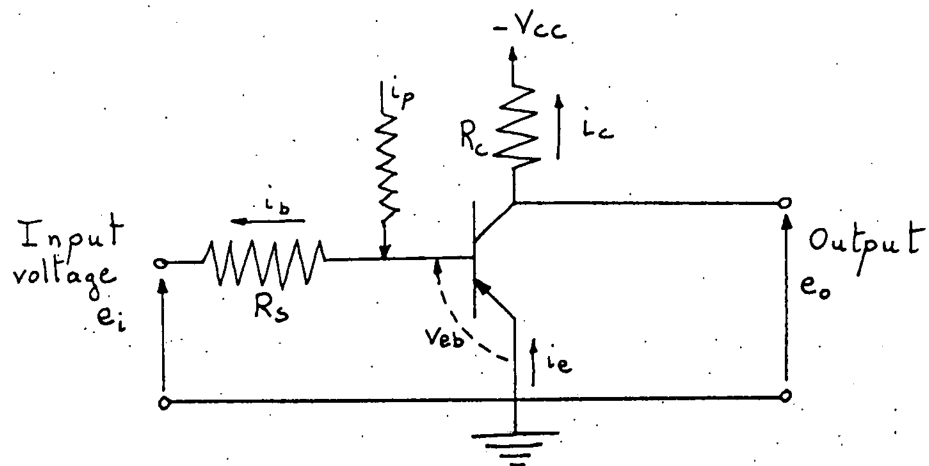
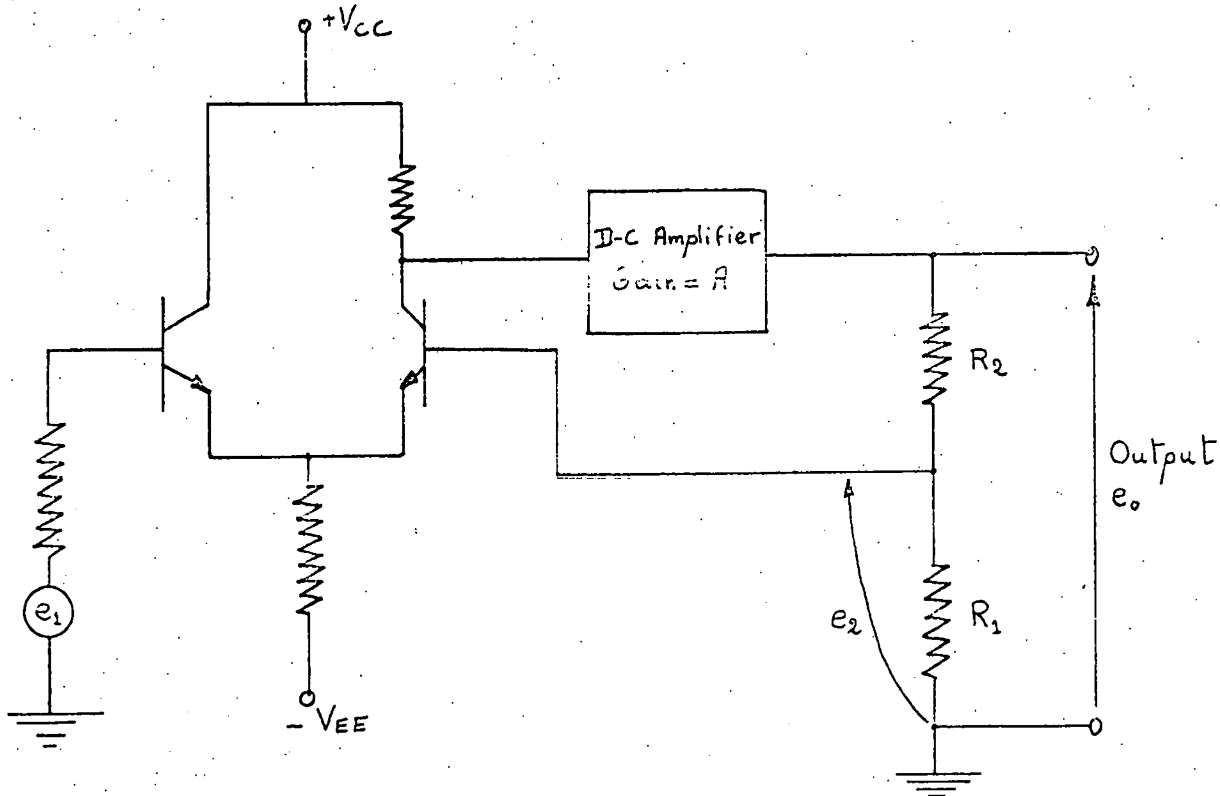
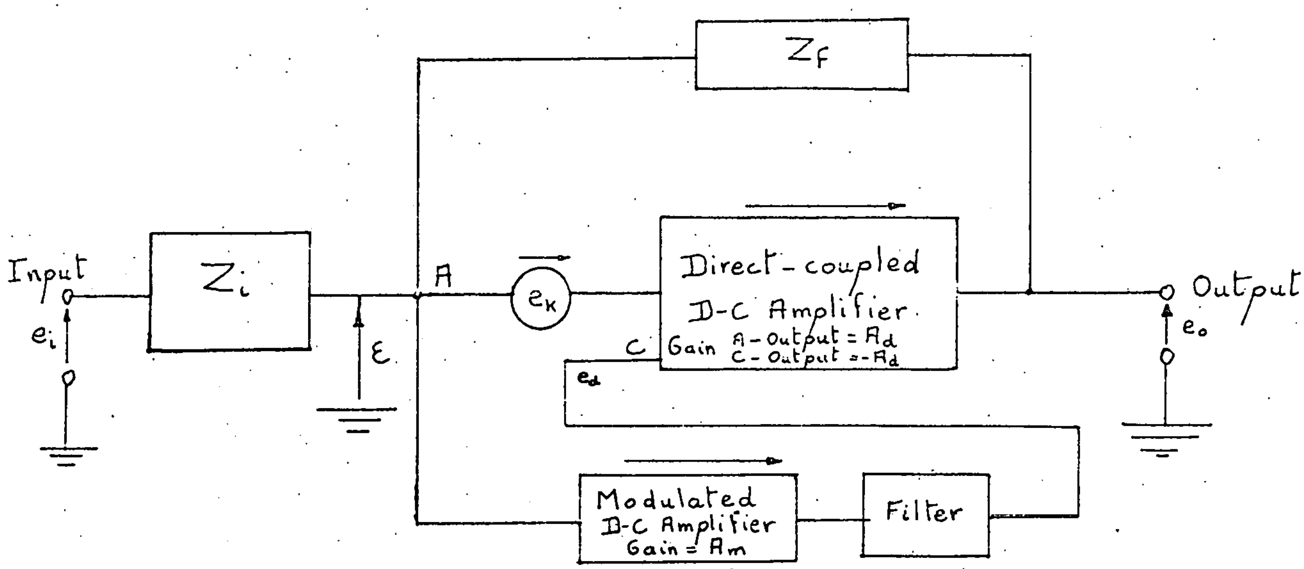

-Vcc
i_p
R_c
i_c
i_b
Input voltage e_i
R_s
V_{eb}
i_e
Output e_o

FIG. 45



Slaughter Circuit

FIG. 46

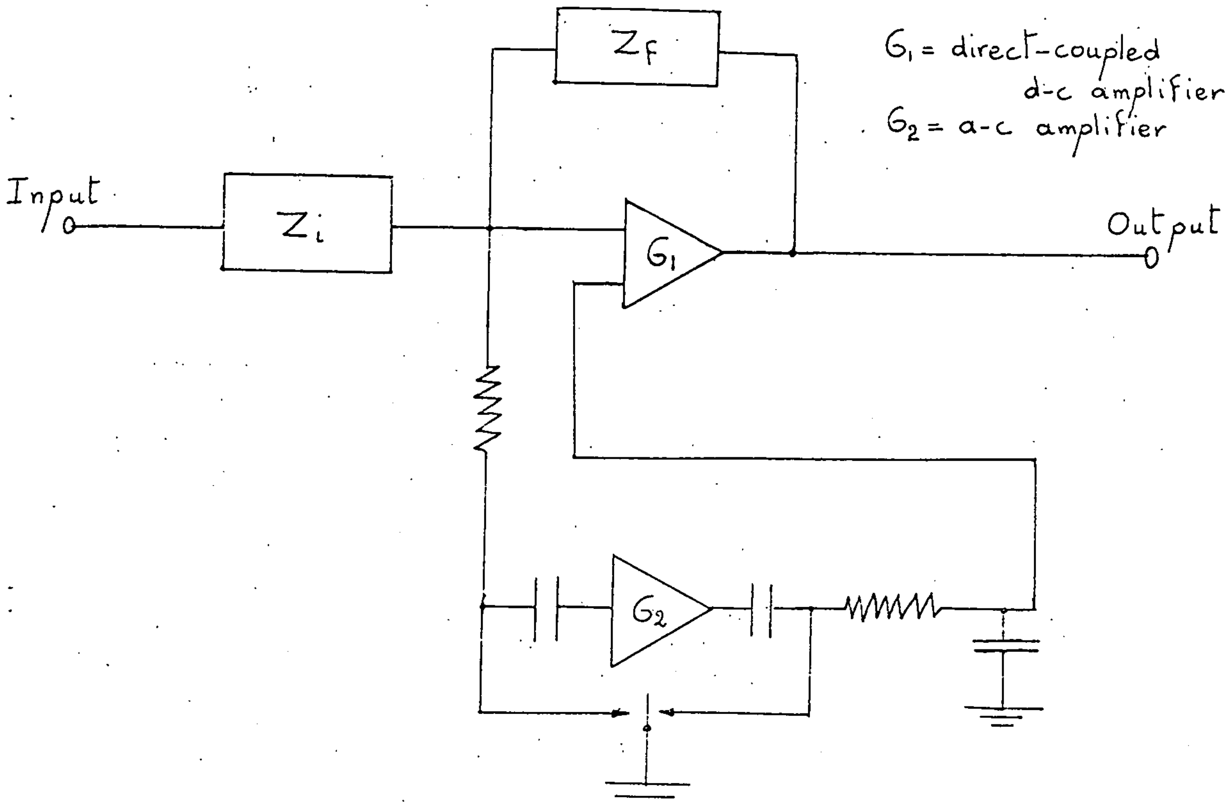


e_K = equivalent offset

Forward Gain { Gain from A to output = A_d
Gain from C to output = $-A_d$

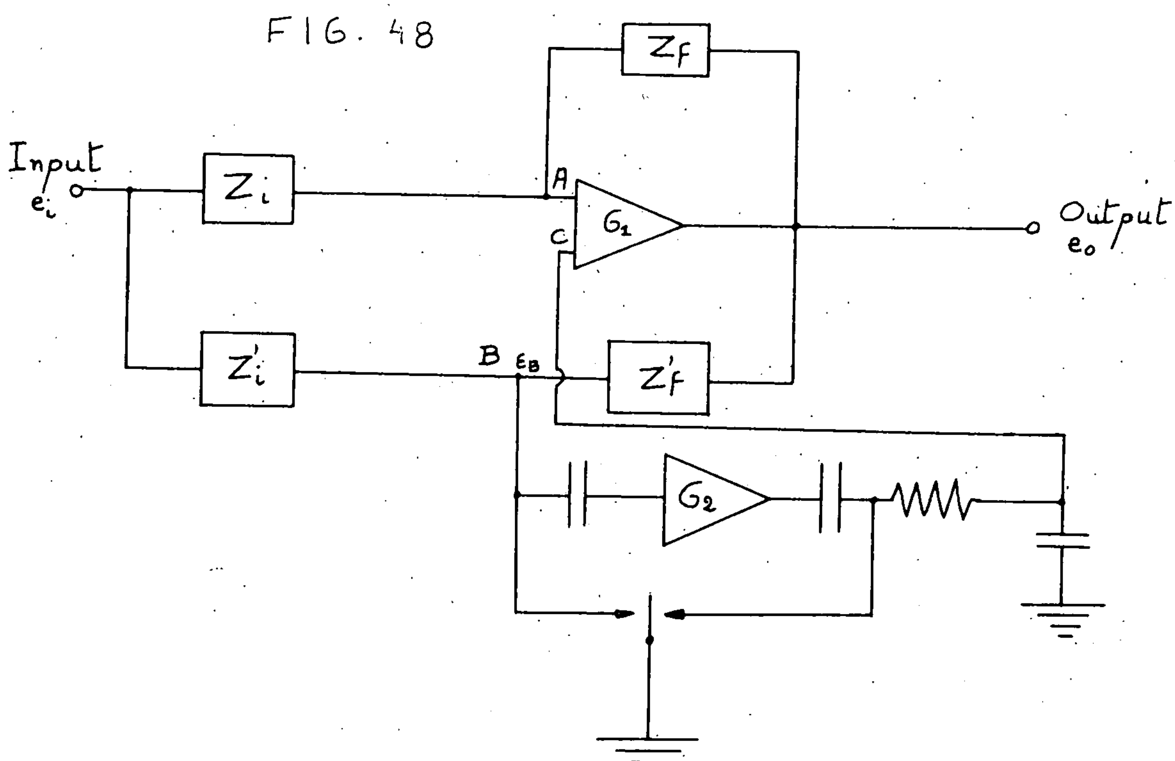
Basic Goldberg Circuit

FIG. 47



Practical Goldberg Circuit

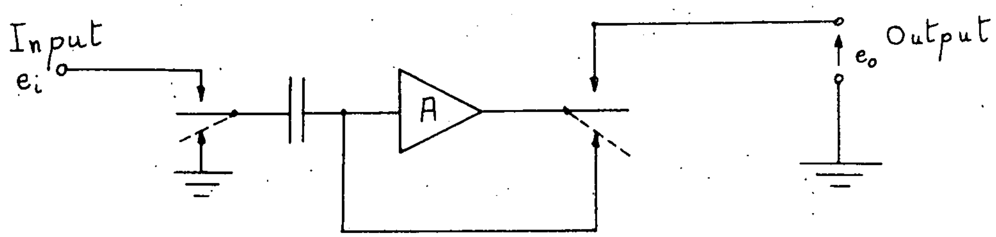
FIG. 48



G_1 = direct-coupled d-c amplifier
G_2 = a-c amplifier

Modified Goldberg Circuit

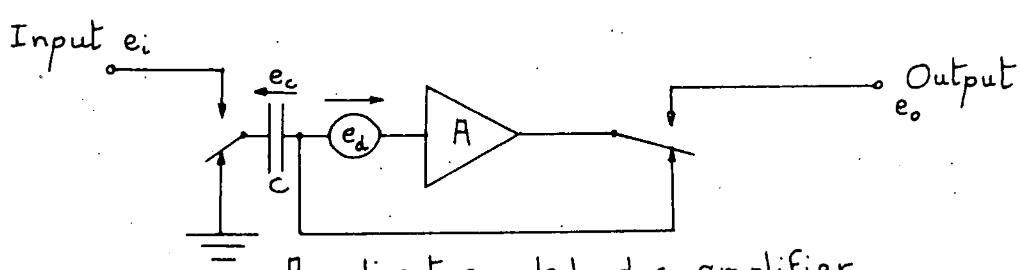
FIG. 49



A = direct-coupled d-c amplifier

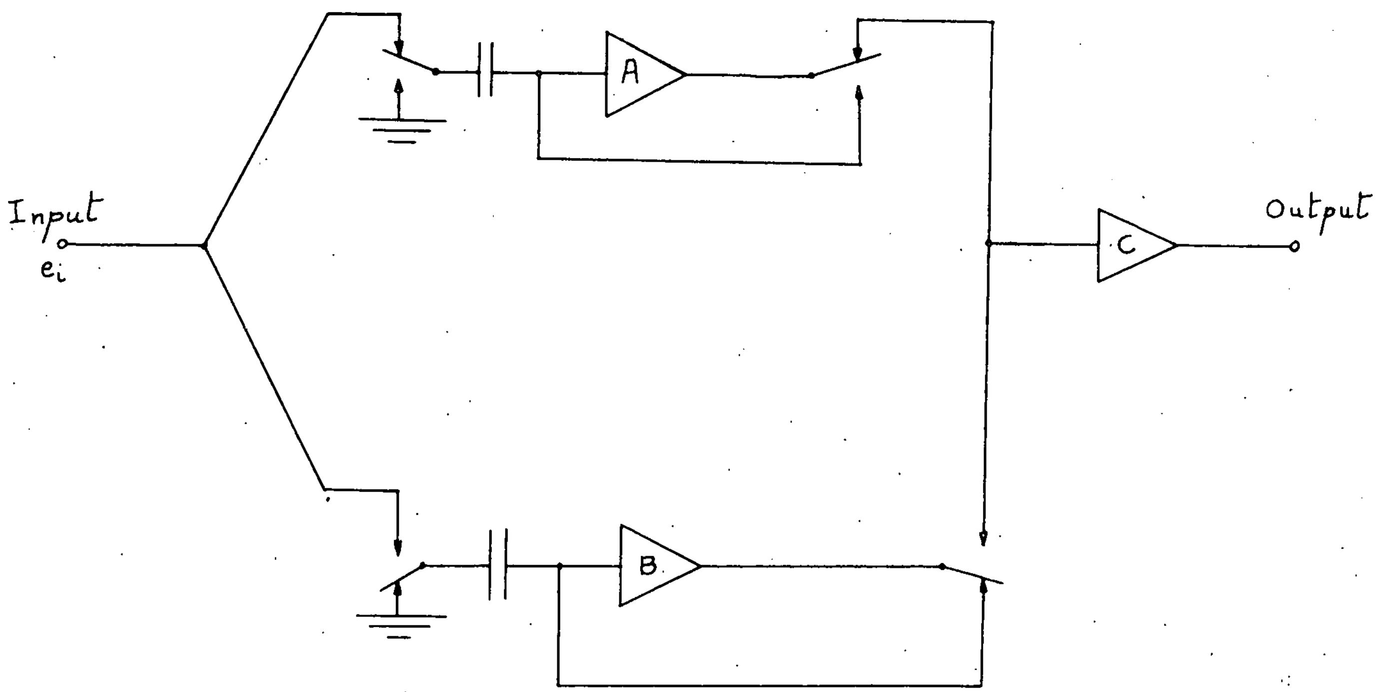
Basic Owen-Prinz Circuit

FIG. 50



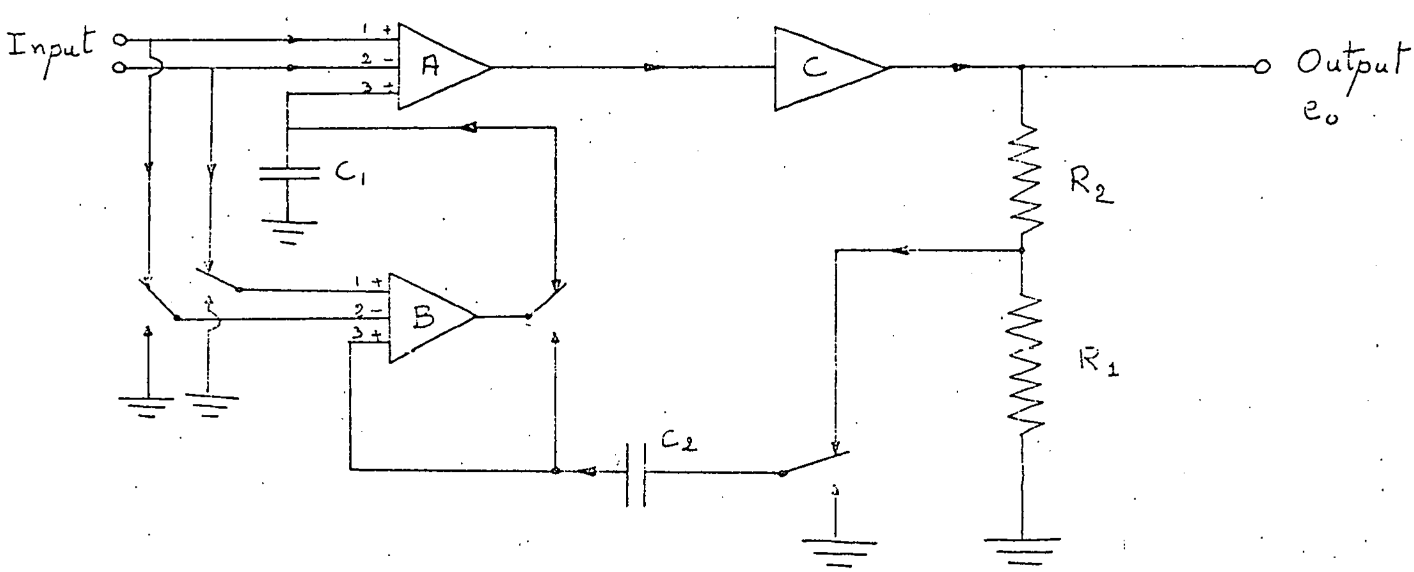
A = direct-coupled d-c amplifier

FIG. 51



Cascade-balance Circuit

FIG. 52



Reflex-monitor Circuit

FIG. 53

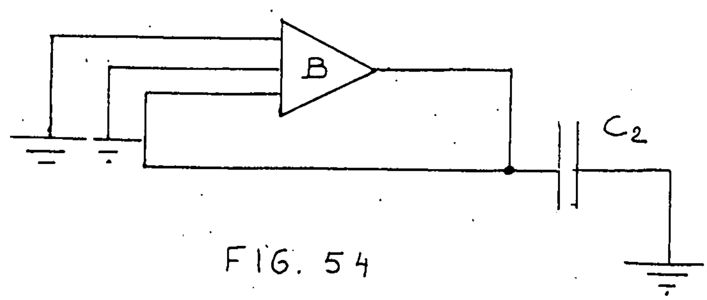
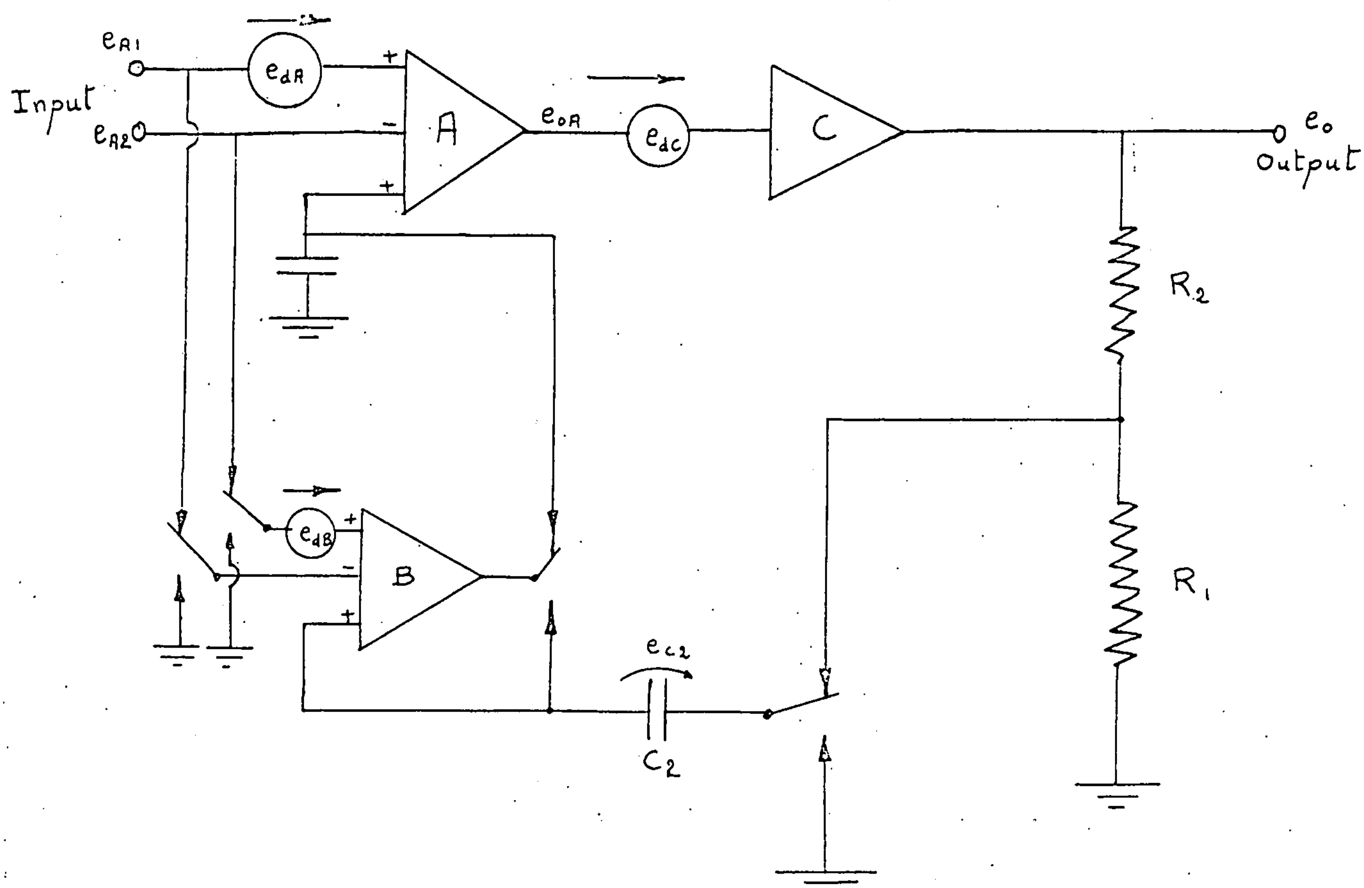


FIG. 54



(Note: The amplifier input impedances are all assumed to be very high.)

Reflex-monitor Circuit

FIG. 55

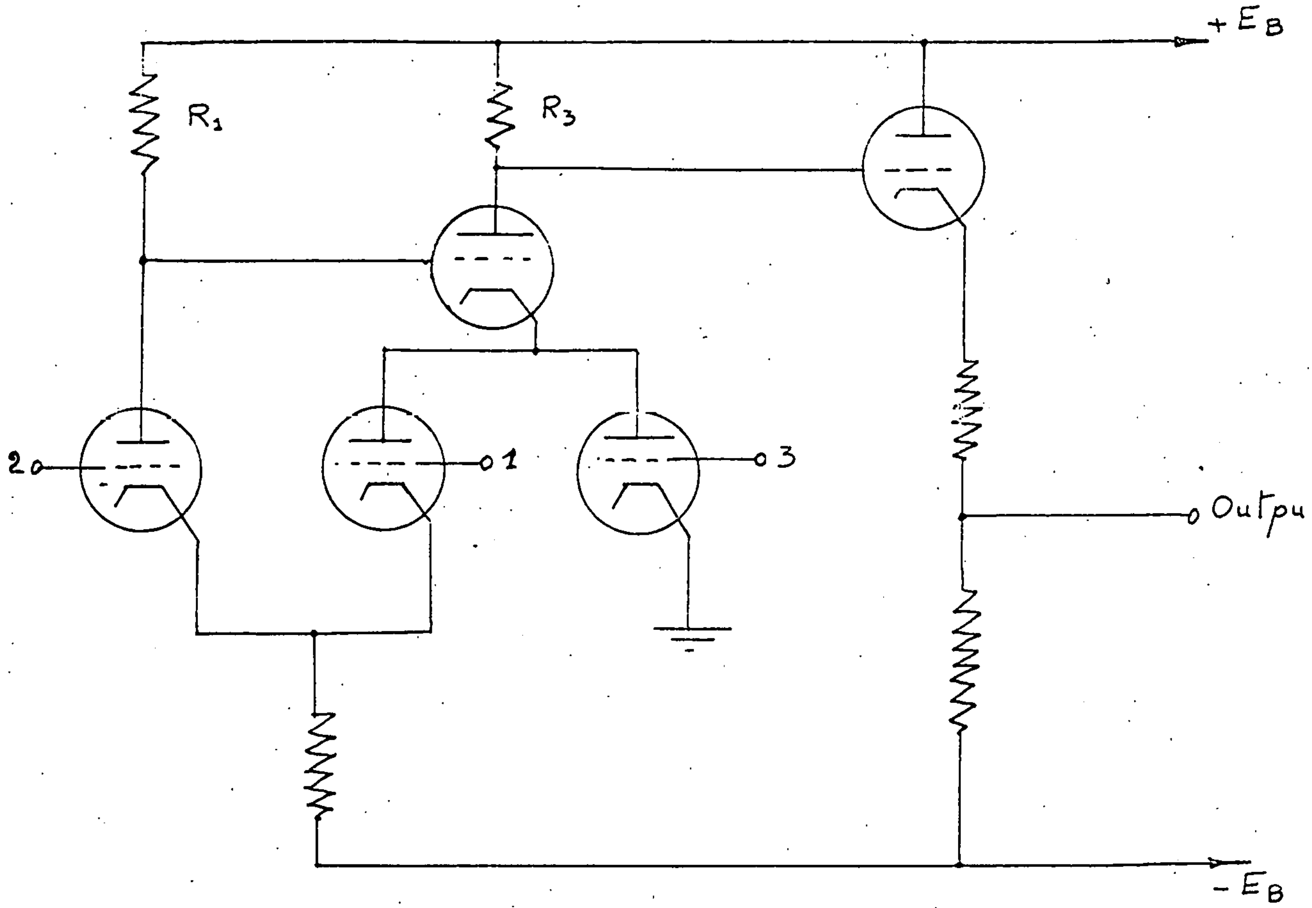
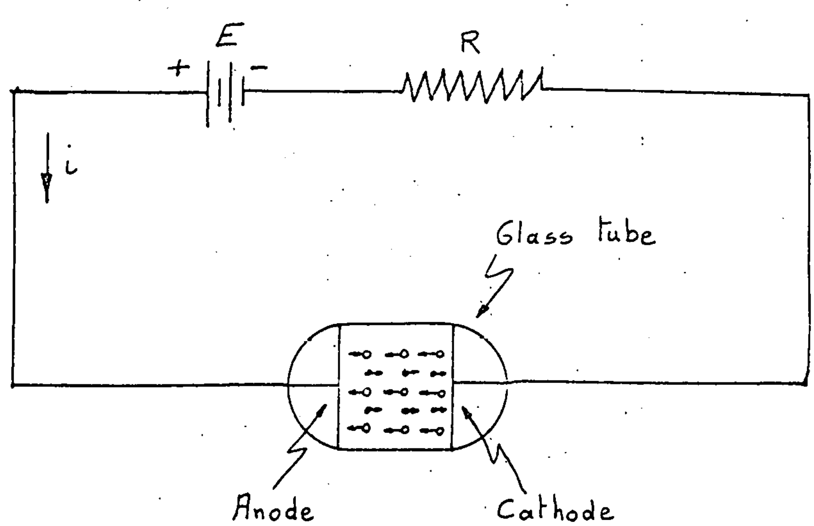


FIG. 56



o = Iodide (I^-)
• = Iodine (I_3^-)

FIG. 57

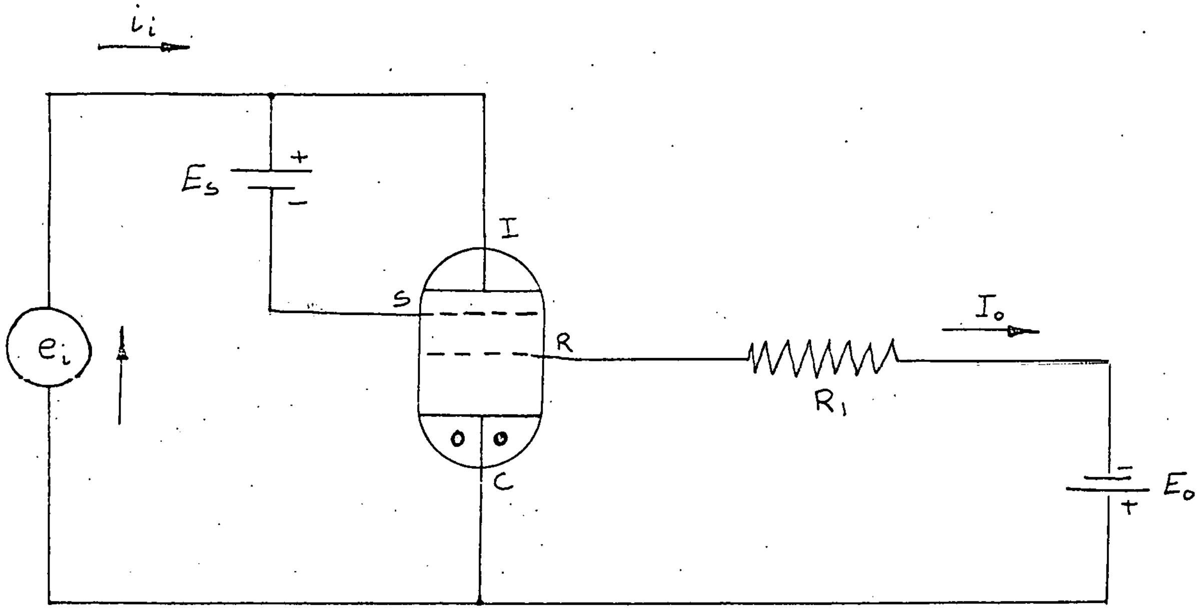
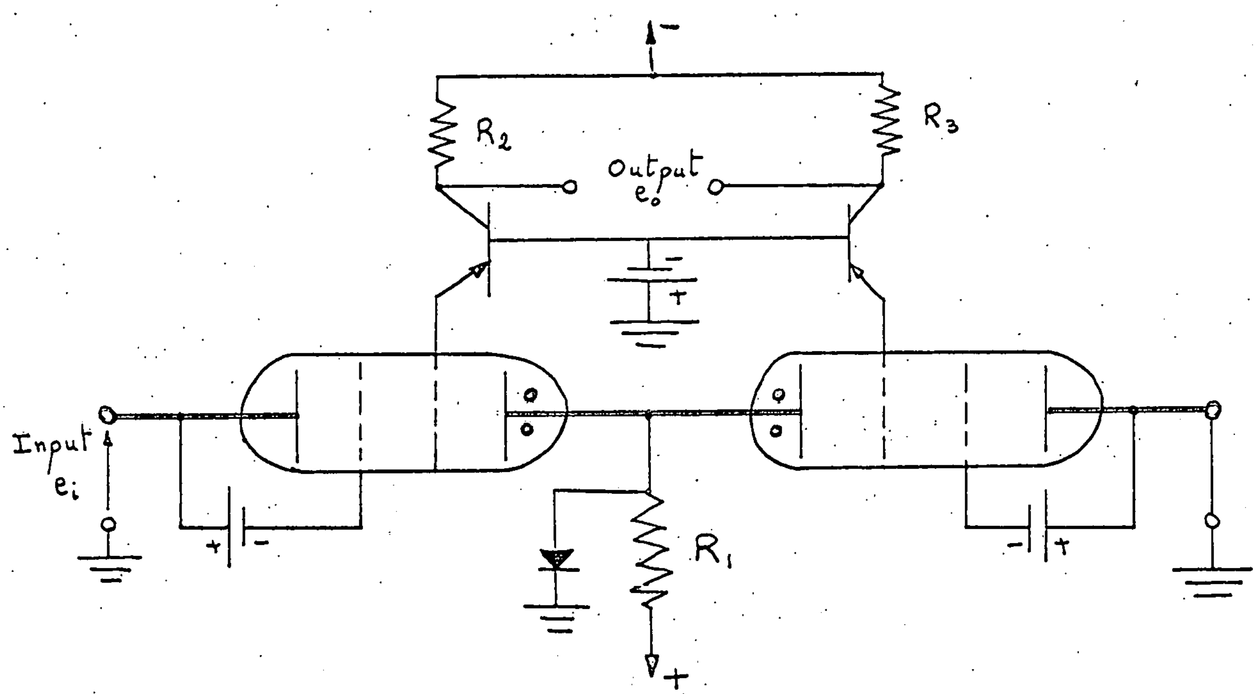
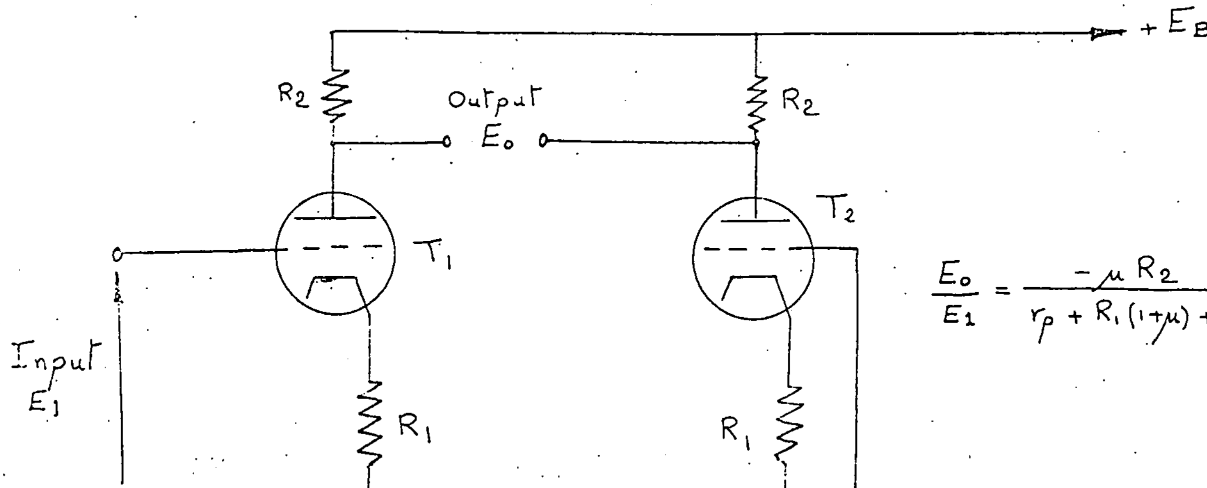


FIG 58



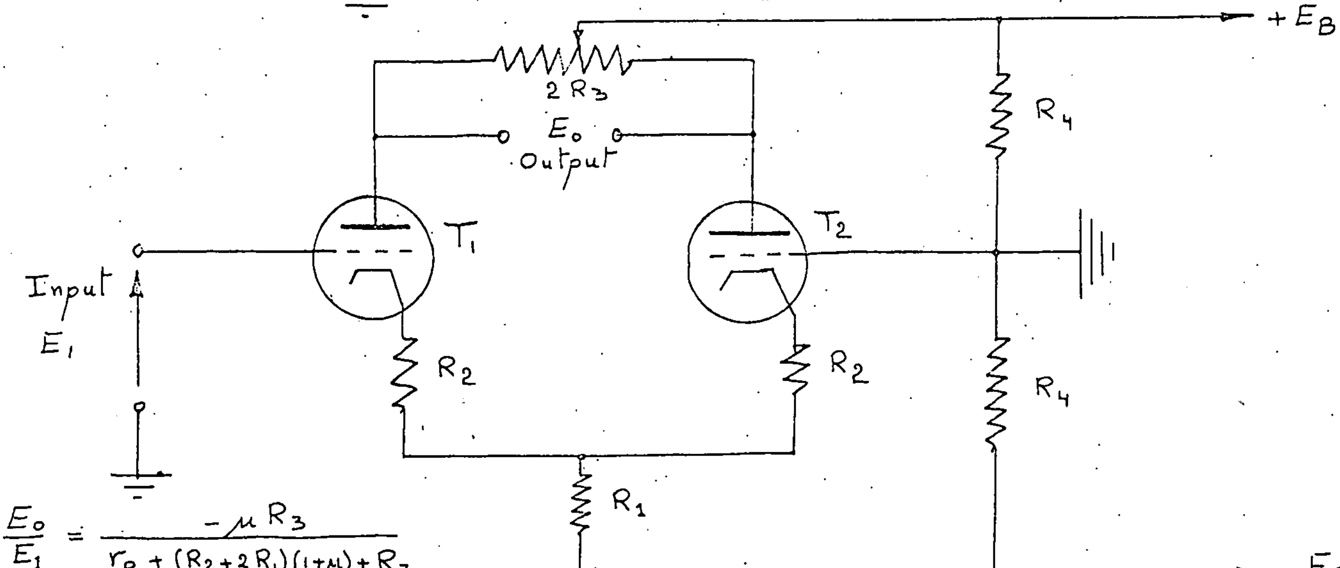
Practical solion-tetrode circuit

FIG. 59



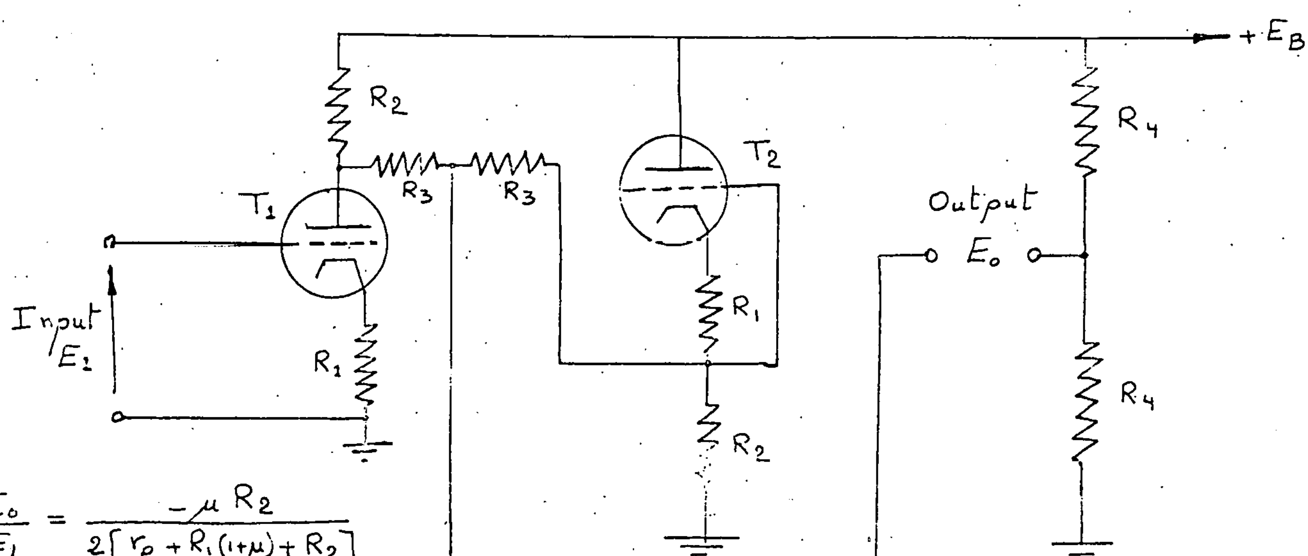
$$\frac{E_o}{E_1} = \frac{-\mu R_2}{r_p + R_1(1+\mu) + R_2}$$

FIG. 60



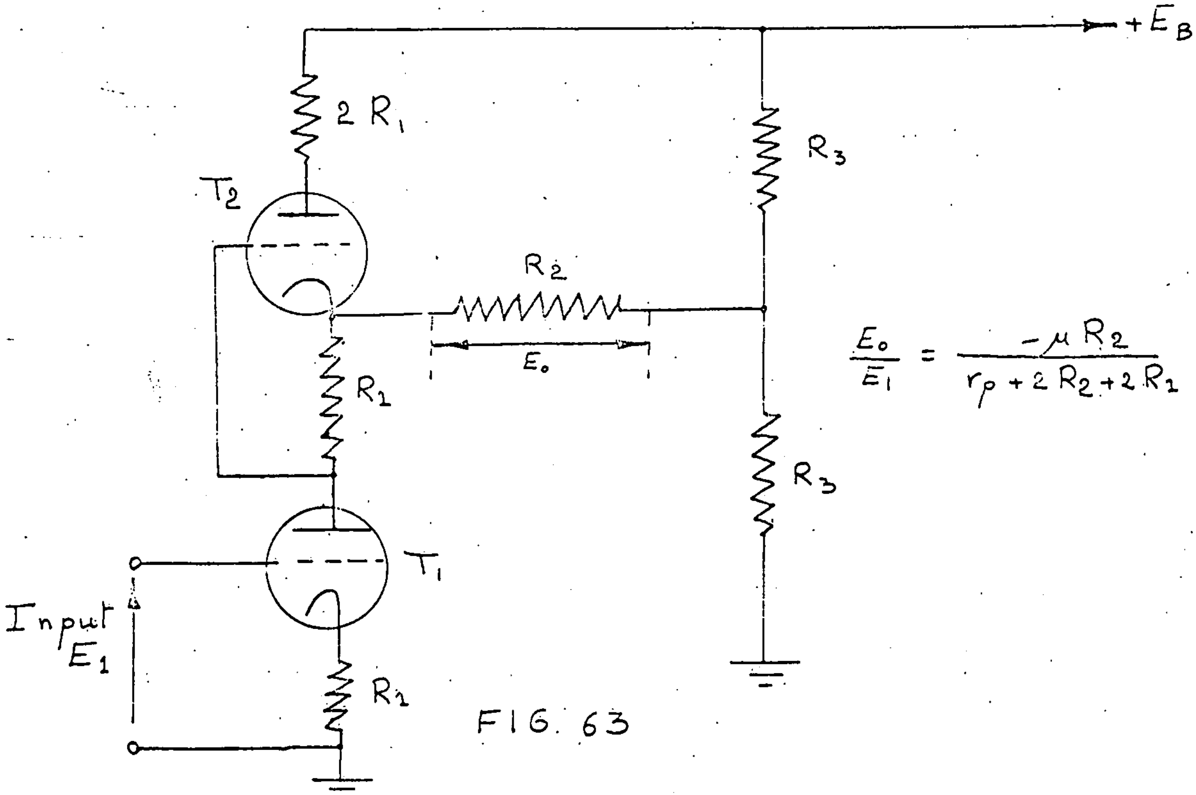
$$\frac{E_o}{E_1} = \frac{-\mu R_3}{r_p + (R_2 + 2R_1)(1+\mu) + R_3}$$

FIG. 61



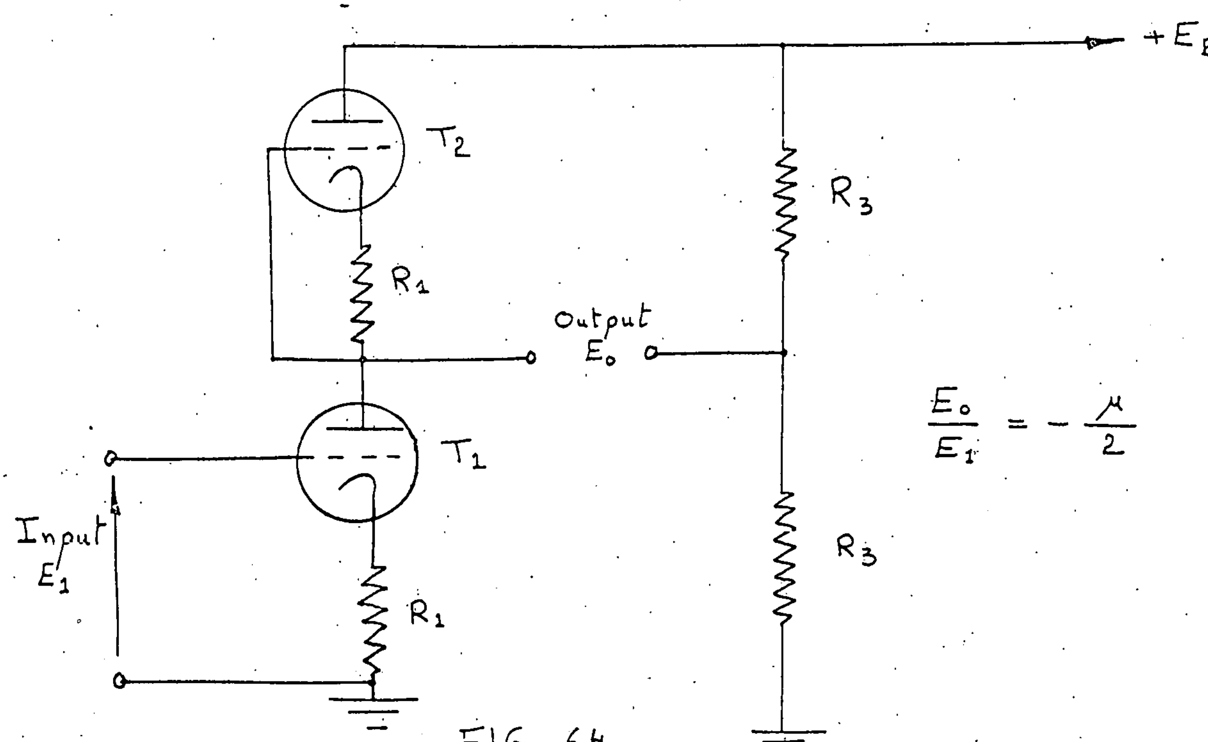
$$\frac{E_o}{E_1} = \frac{-\mu R_2}{2[r_p + R_1(1+\mu) + R_2]}$$

FIG. 62



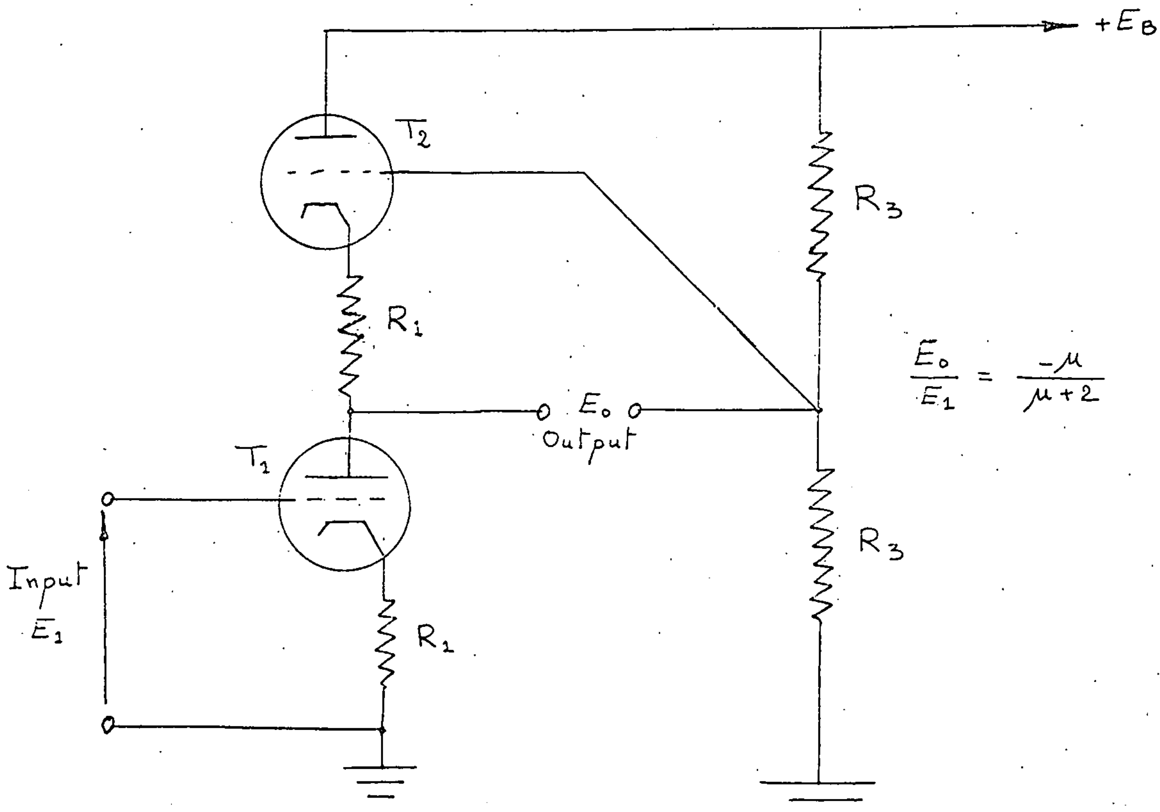
$$\frac{E_0}{E_1} = \frac{-\mu R_2}{r_p + 2R_2 + 2R_1}$$

FIG. 63



$$\frac{E_0}{E_1} = -\frac{\mu}{2}$$

FIG. 64



$$\frac{E_o}{E_1} = \frac{-\mu}{\mu+2}$$

FIG. 65

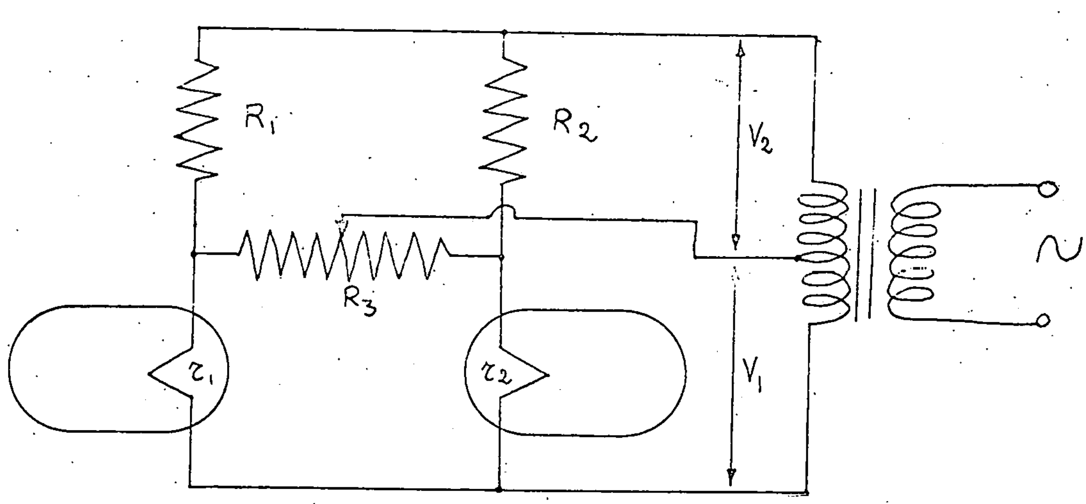


FIG. 66

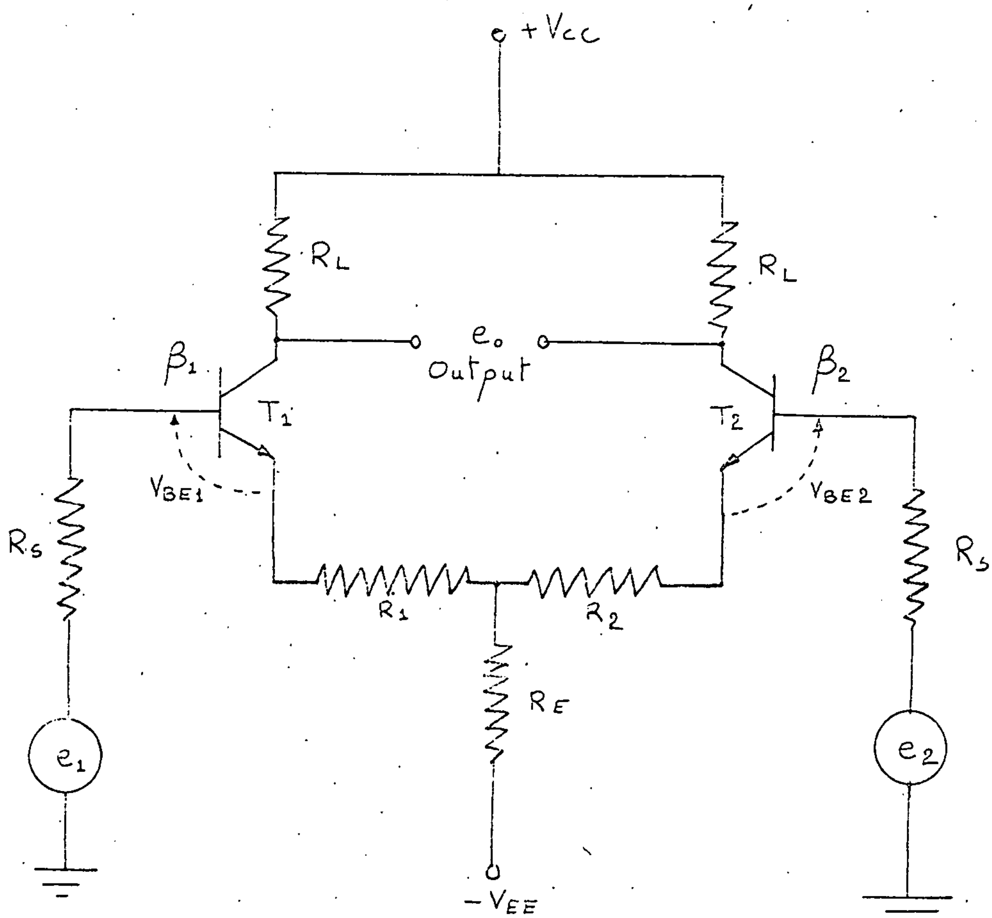


FIG. 67

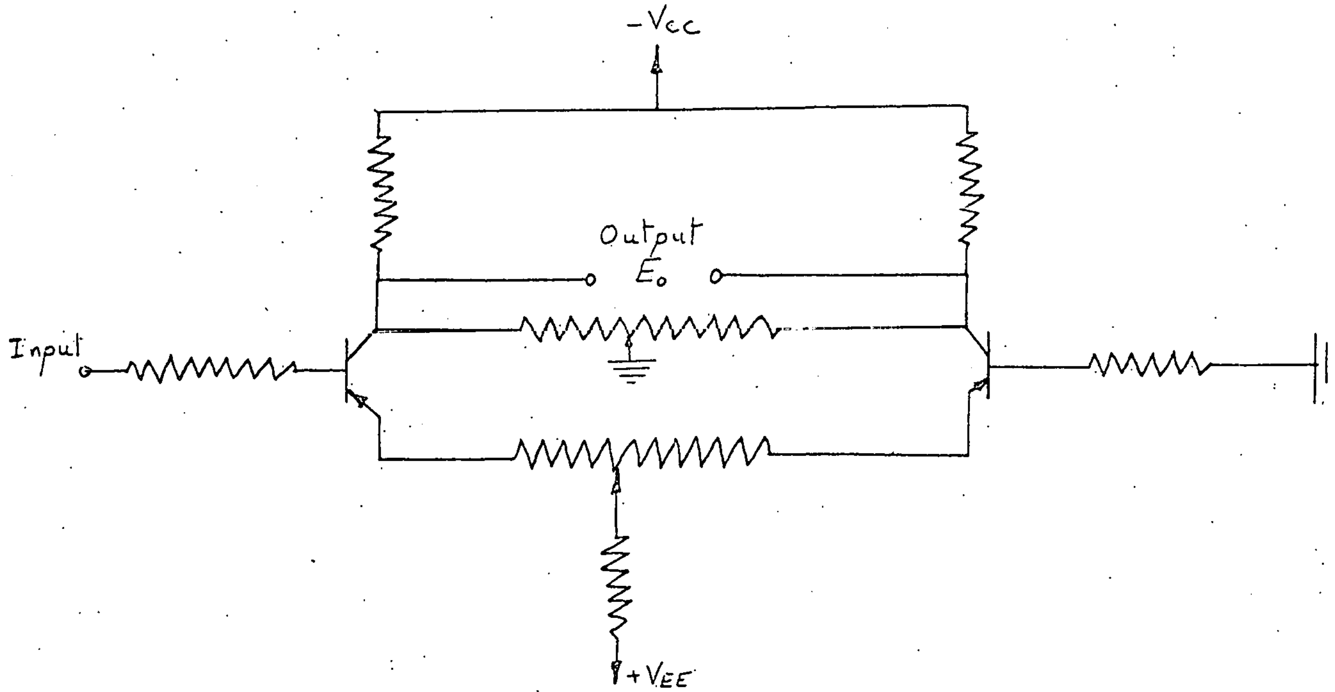


FIG. 68

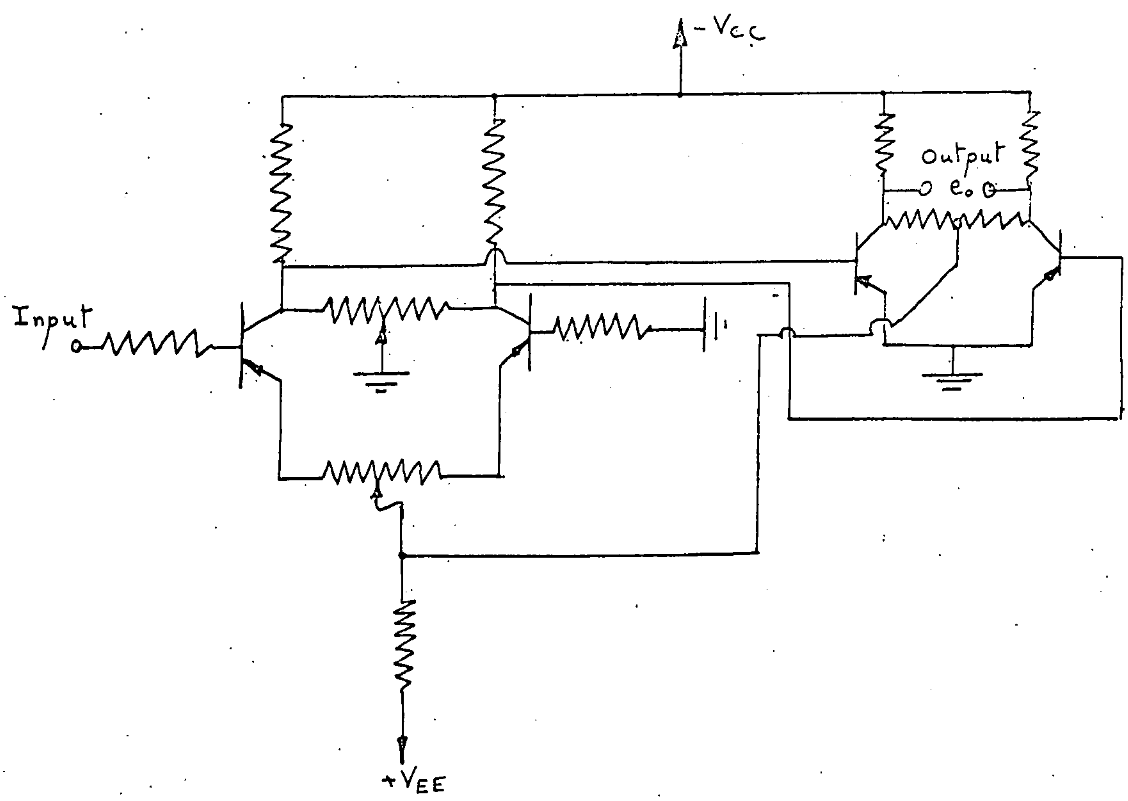


FIG. 69

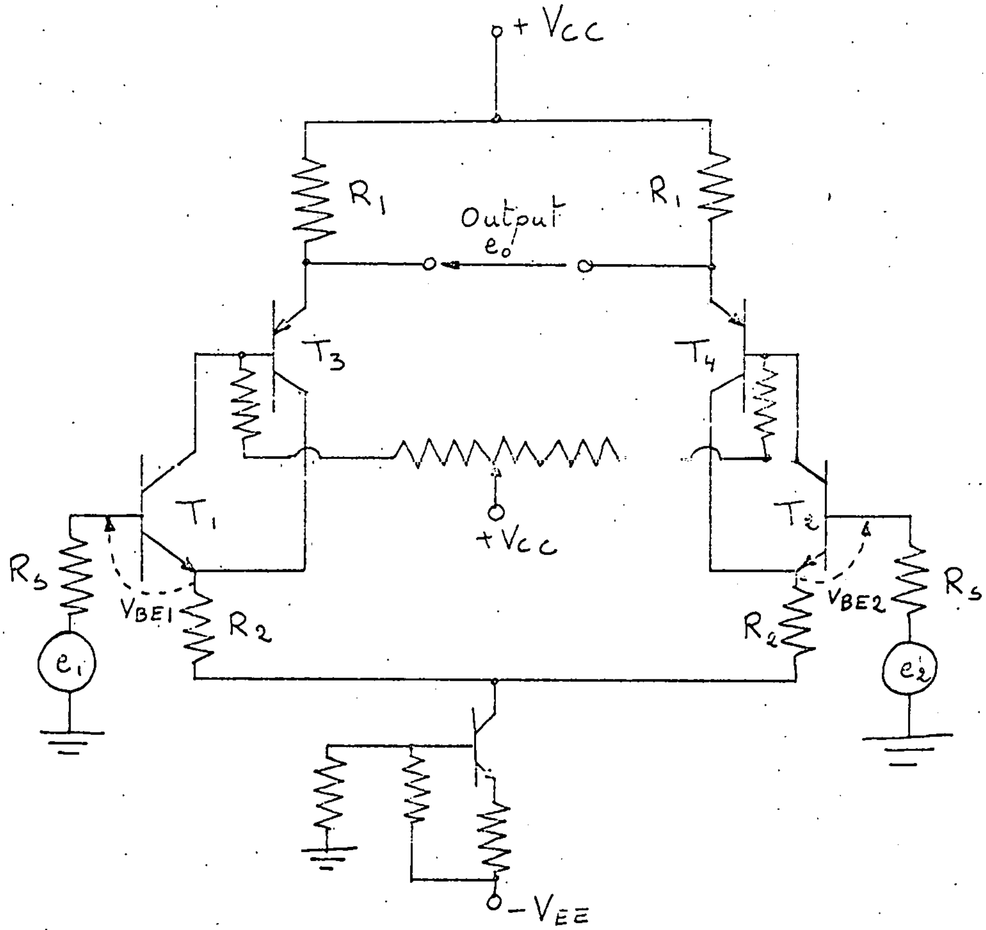

+Vcc
R1
R1
Output
e0
T3
T4
T1
T2
+Vcc
Rs
Rs
VBE1
VBE2
R2
R2
e1
e2
-VEE

FIG. 70

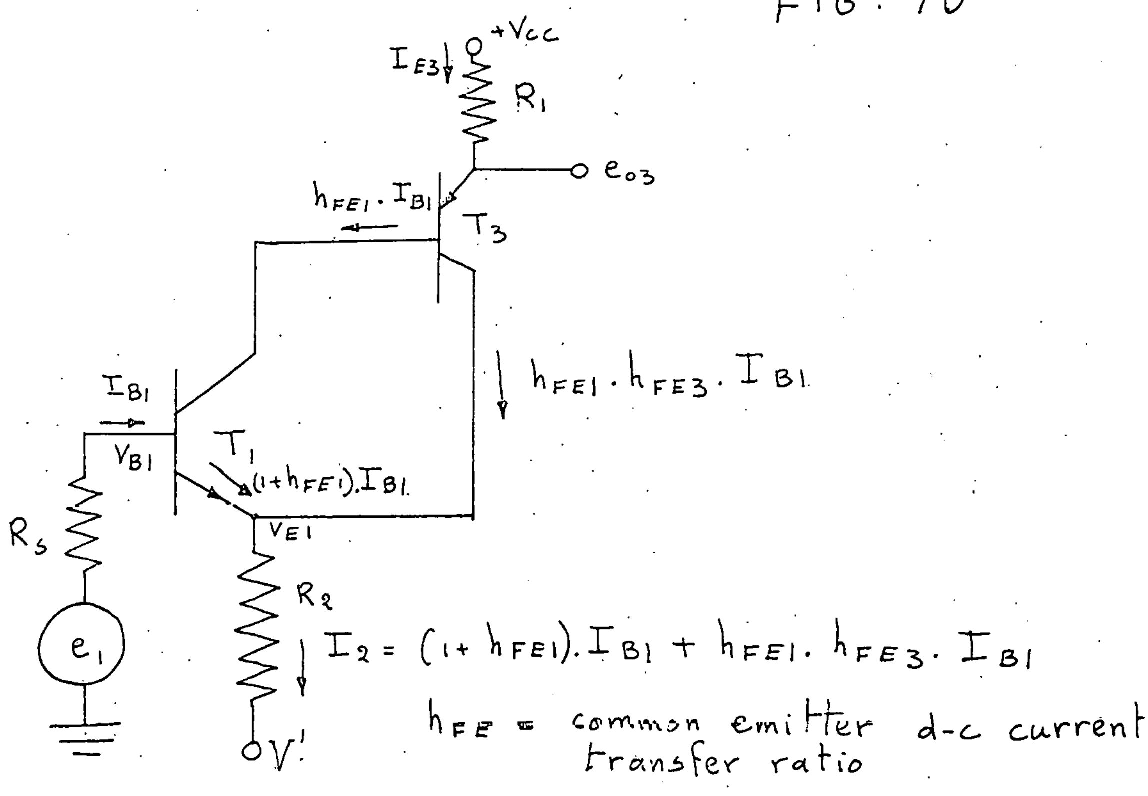
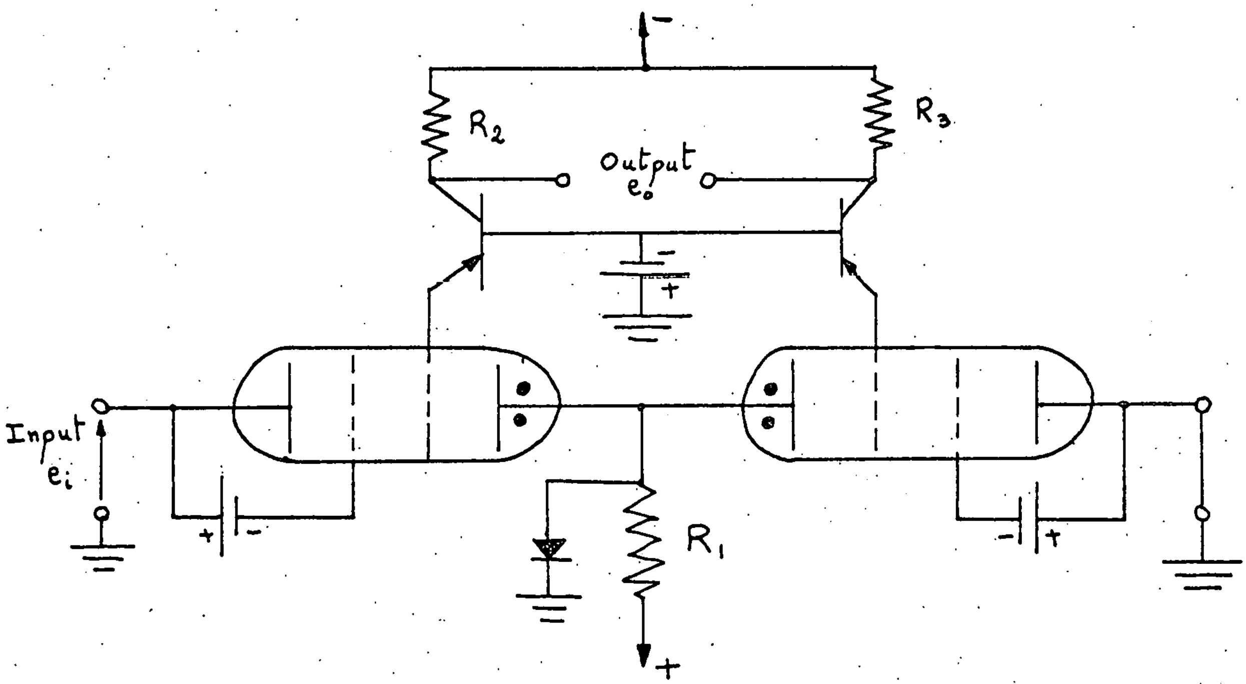
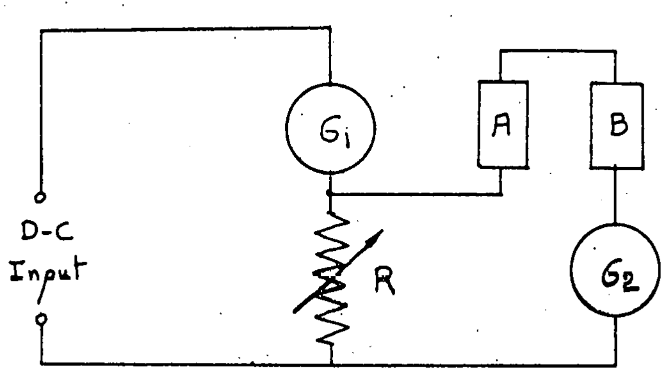

IE3
+Vcc
R1
e03
hFE1 · IB1
T3
hFE1 · hFE3 · IB1
IB1
VB1
T1
(1+hFE1).IB1
VE1
Rs
R2
e1
I2 = (1+ hFE1). IB1 + hFE1 . hFE3 . IB1
hFE = common emitter d-c current transfer ratio
V'

FIG. 71



Practical solion-tetrode circuit

FIG. 72



A simple d-c amplifier using photo-voltaic cells.

FIG. 73

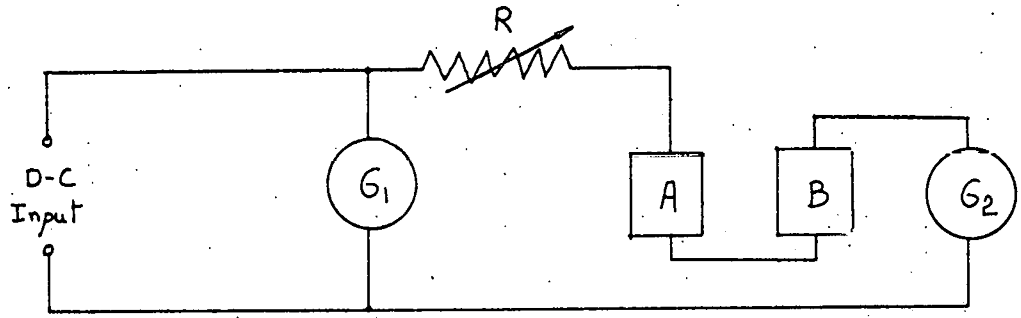
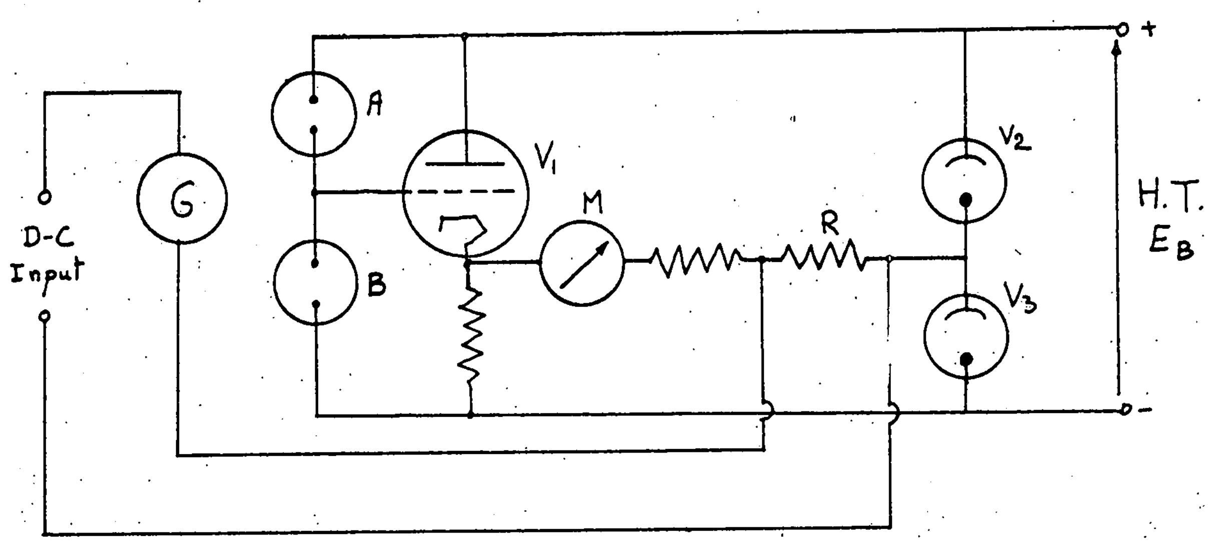


FIG. 74



D-C Amplifier using photo-emissive cells.

FIG. 75

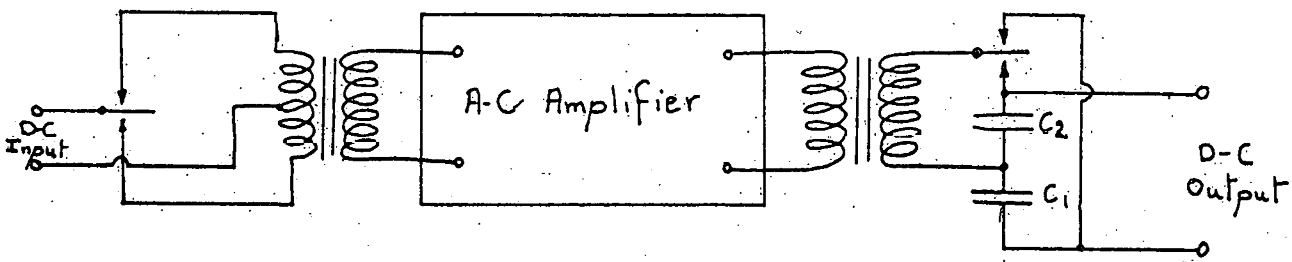


FIG. 76

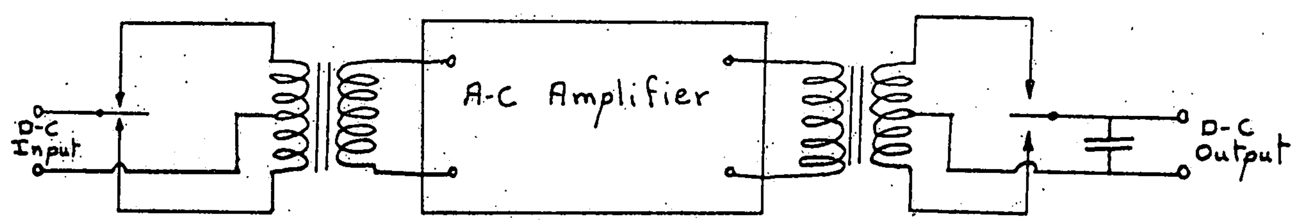


FIG. 77

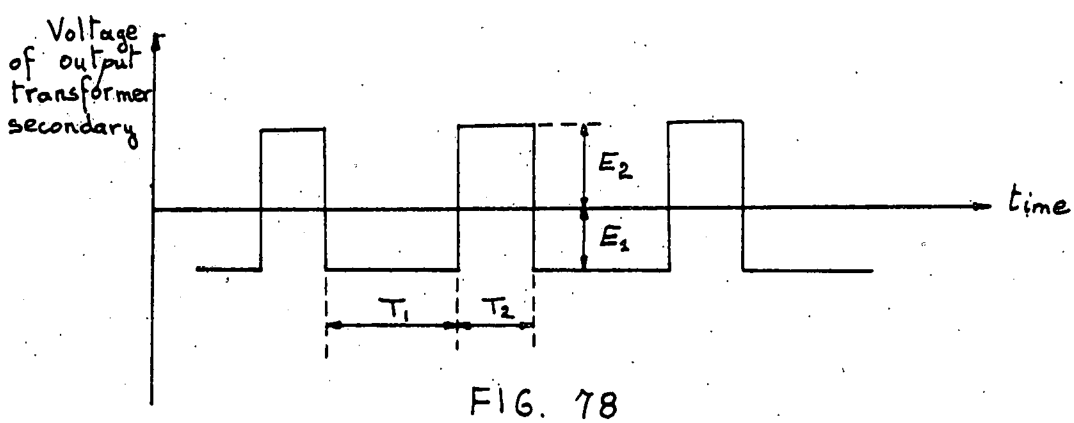
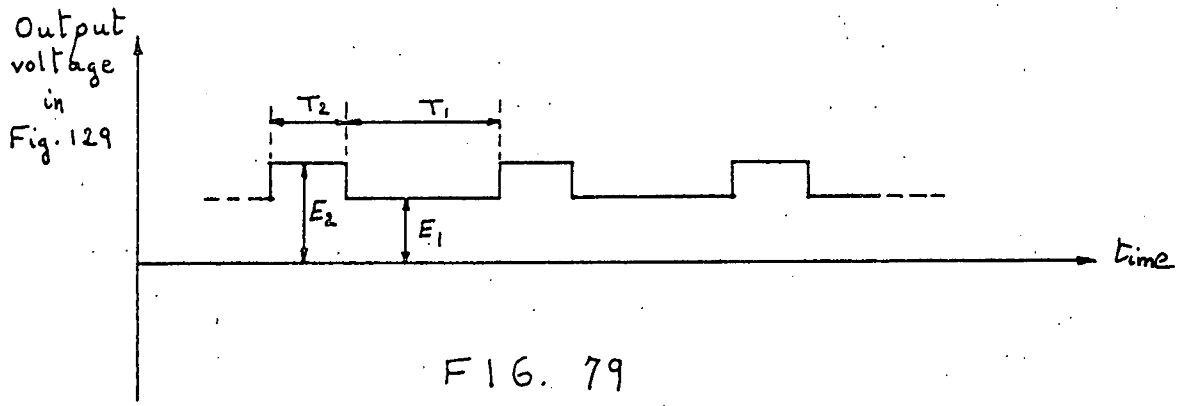
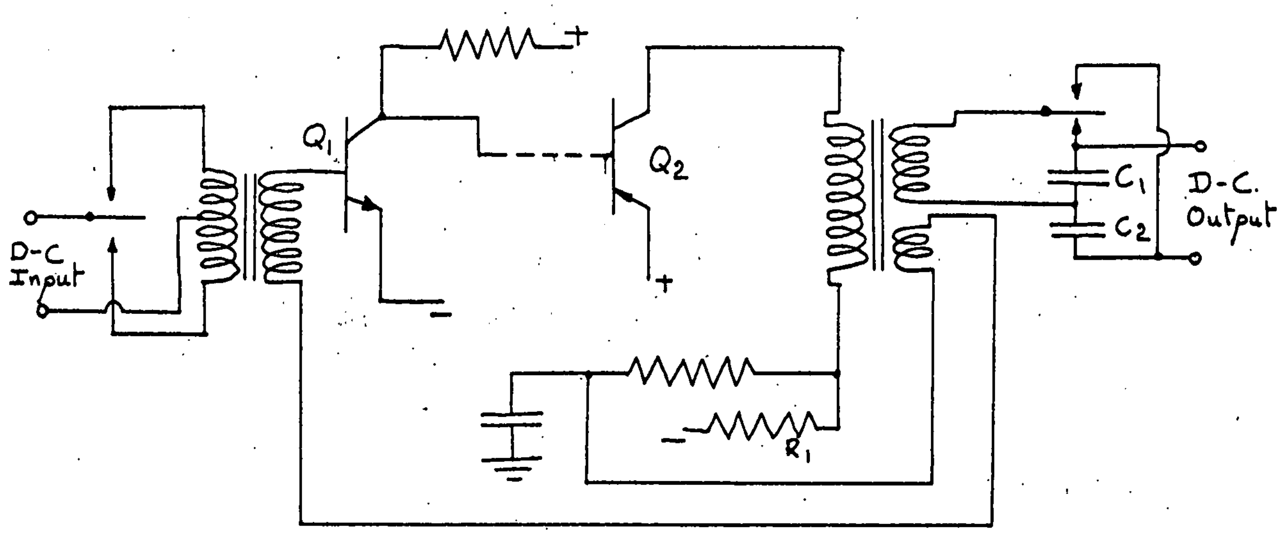


FIG. 78


Output voltage in Fig. 129
T2
T1
E2
E1
time
FIG. 79


Q1
Q2
+
+
-
-
C1
C2
D-C Input
D-C. Output
R1
FIG. 80

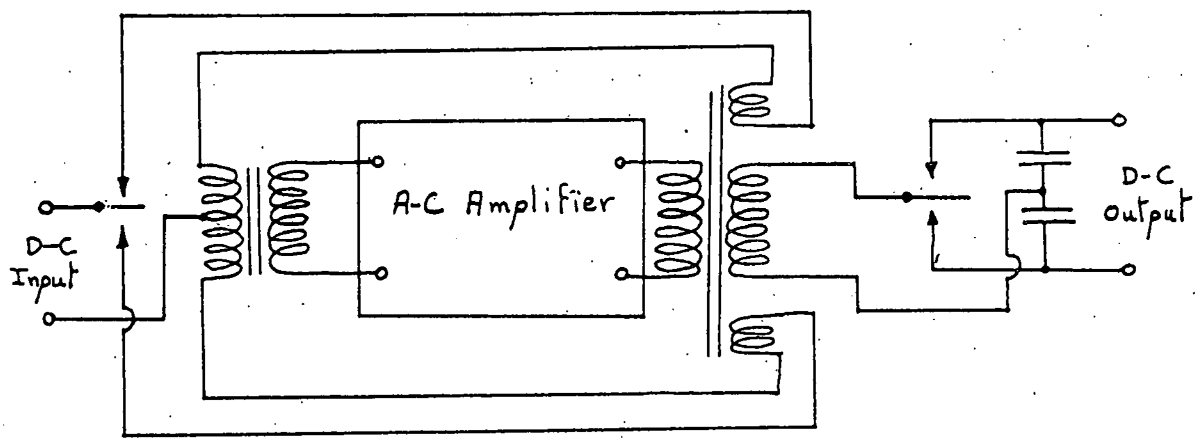
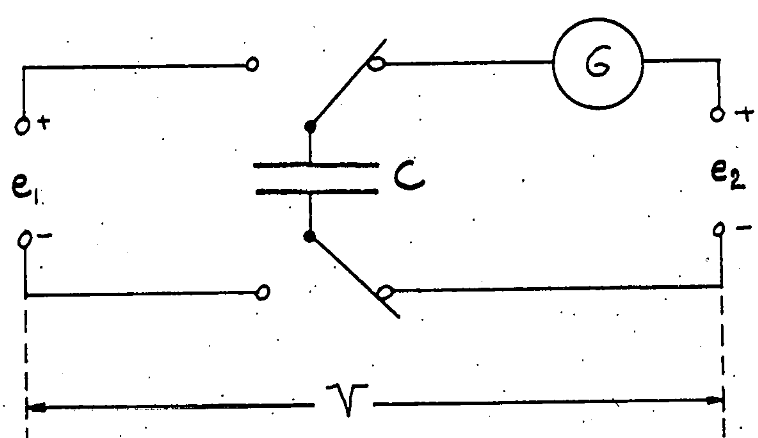


FIG. 81



Potential comparator.

FIG. 82

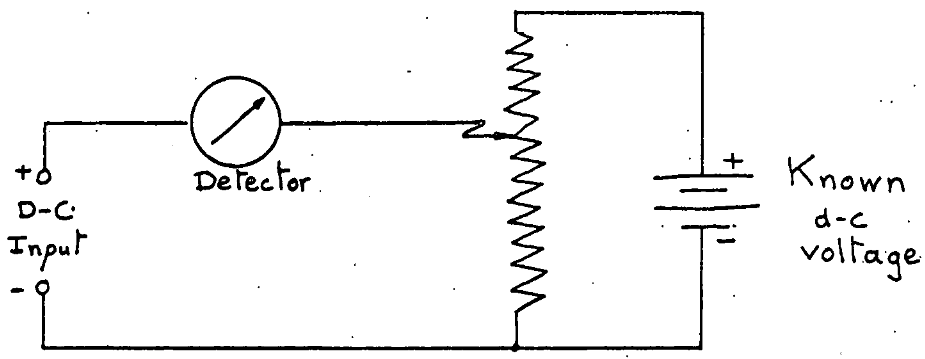

Detector
+
D-C
Input
-
+
-
Known
d-c
voltage

FIG. 83

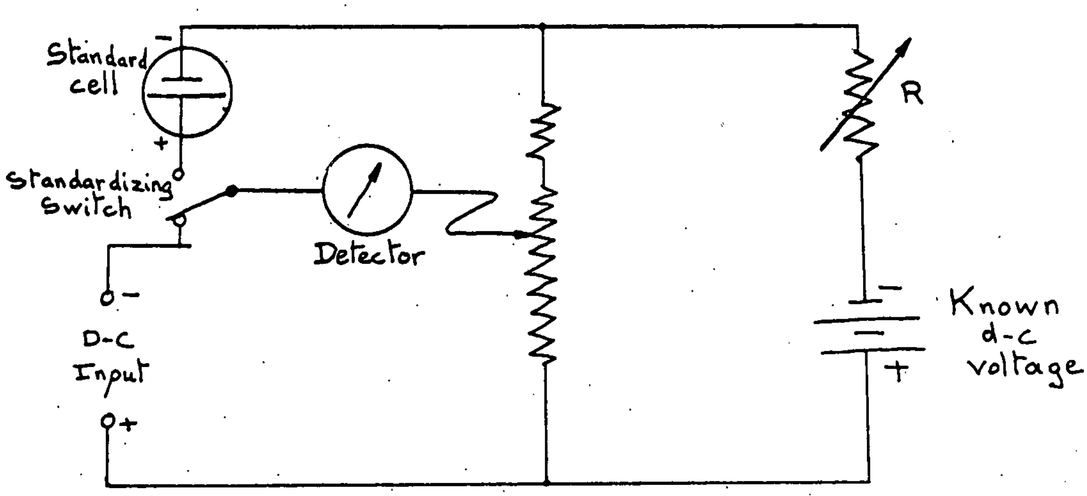

Standard
cell
-
+
Standardizing
Switch
Detector
R
D-C
Input
-
+
-
+
Known
d-c
voltage

FIG. 84

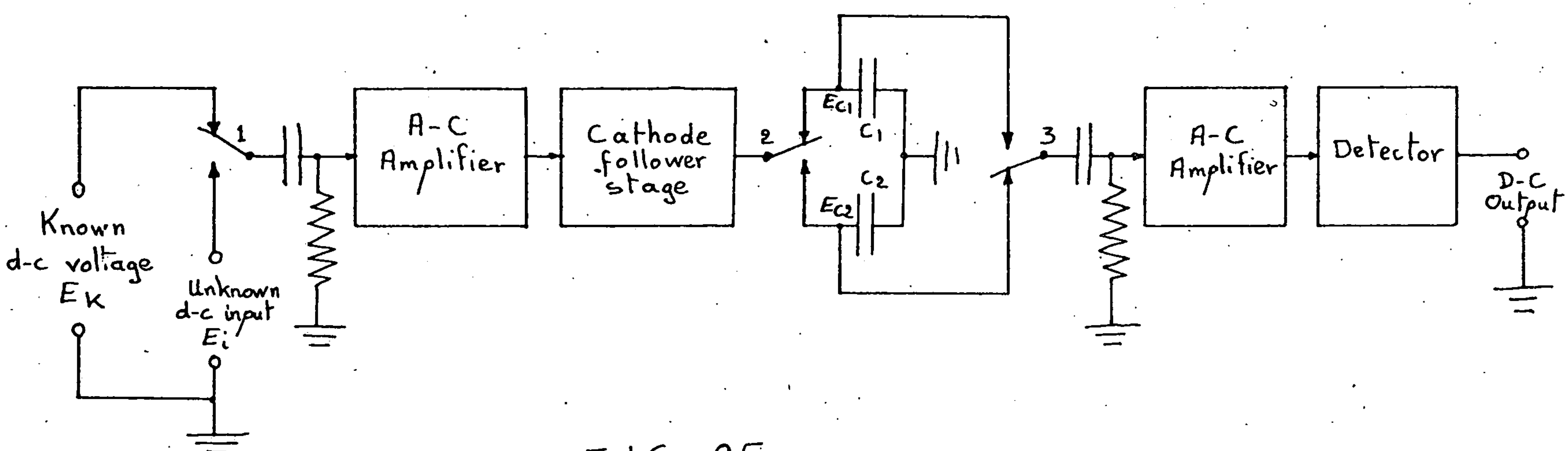

Known
d-c voltage
EK
Unknown
d-c input
Ei
1
A-C
Amplifier
Cathode
follower
stage
2
EC1
C1
C2
EC2
3
A-C
Amplifier
Detector
D-C
Output

FIG. 85

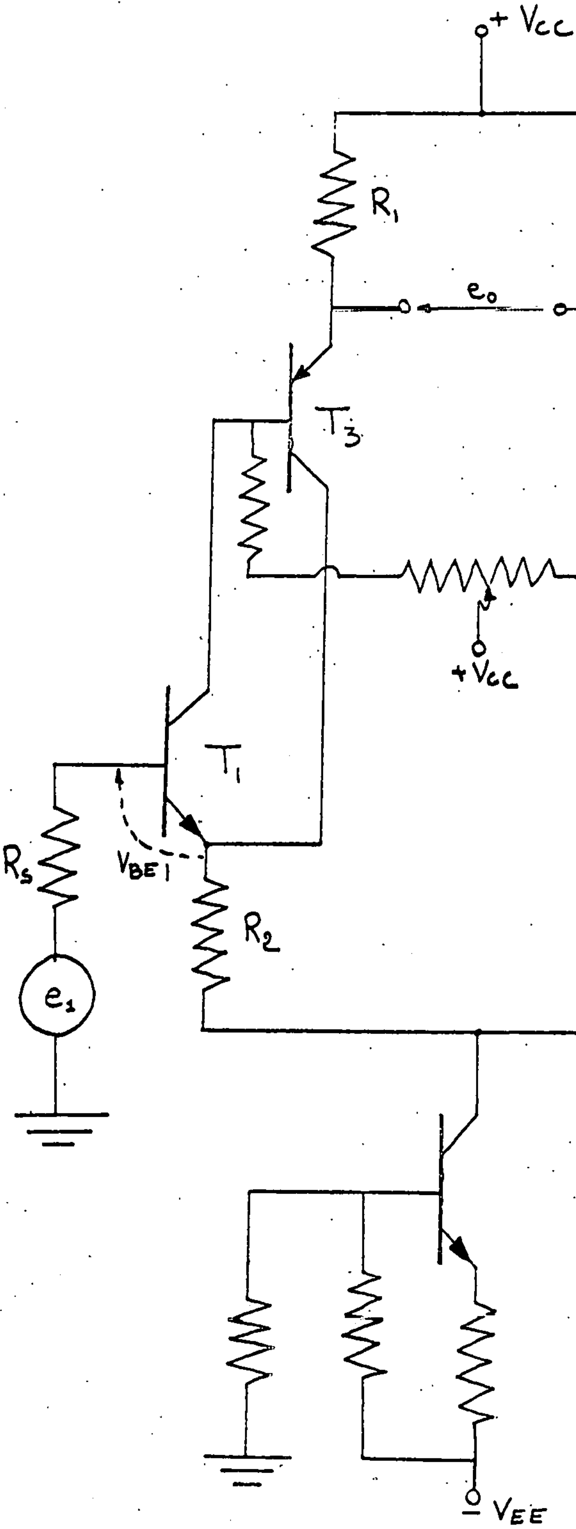


FIG. 86

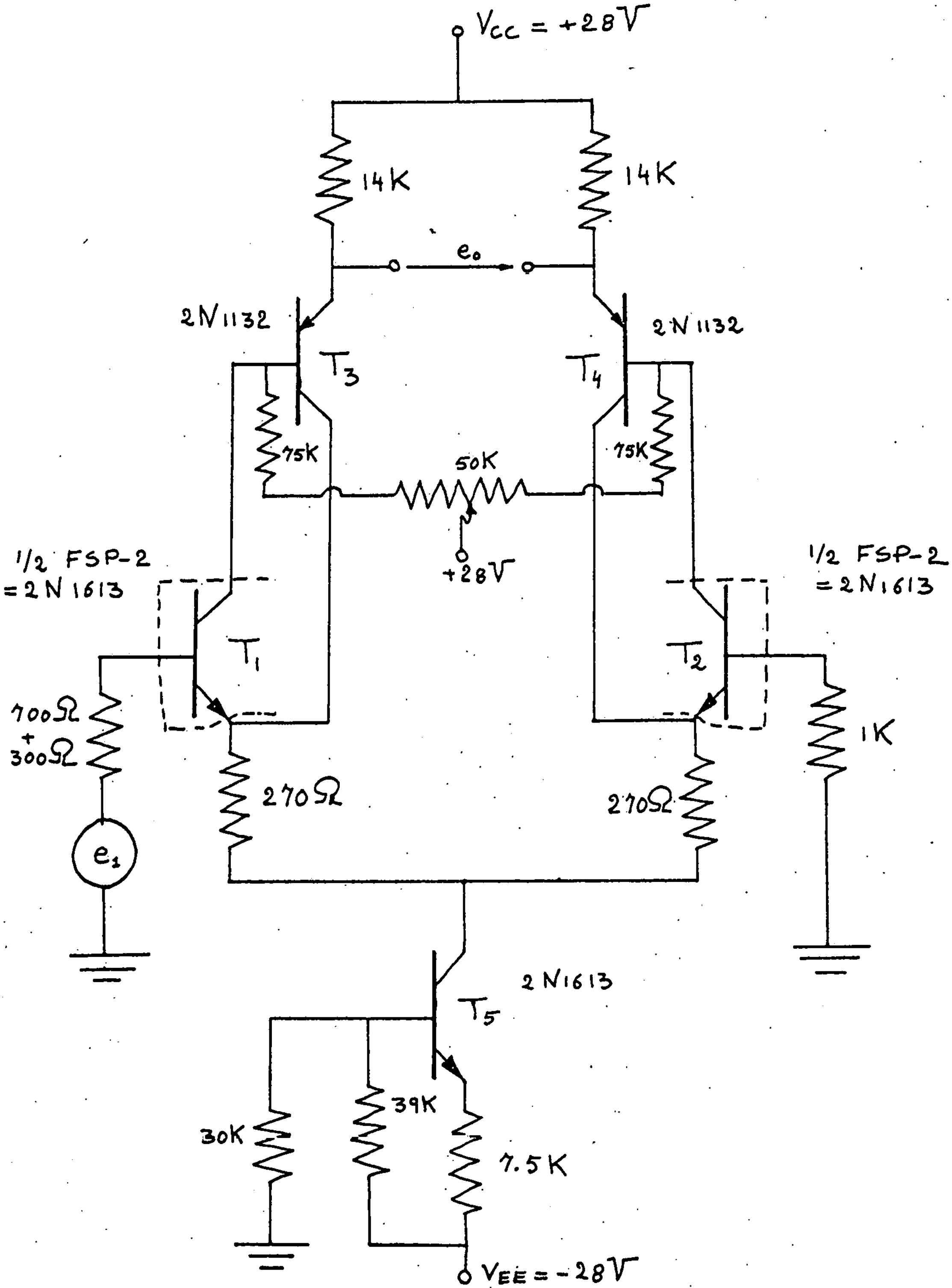
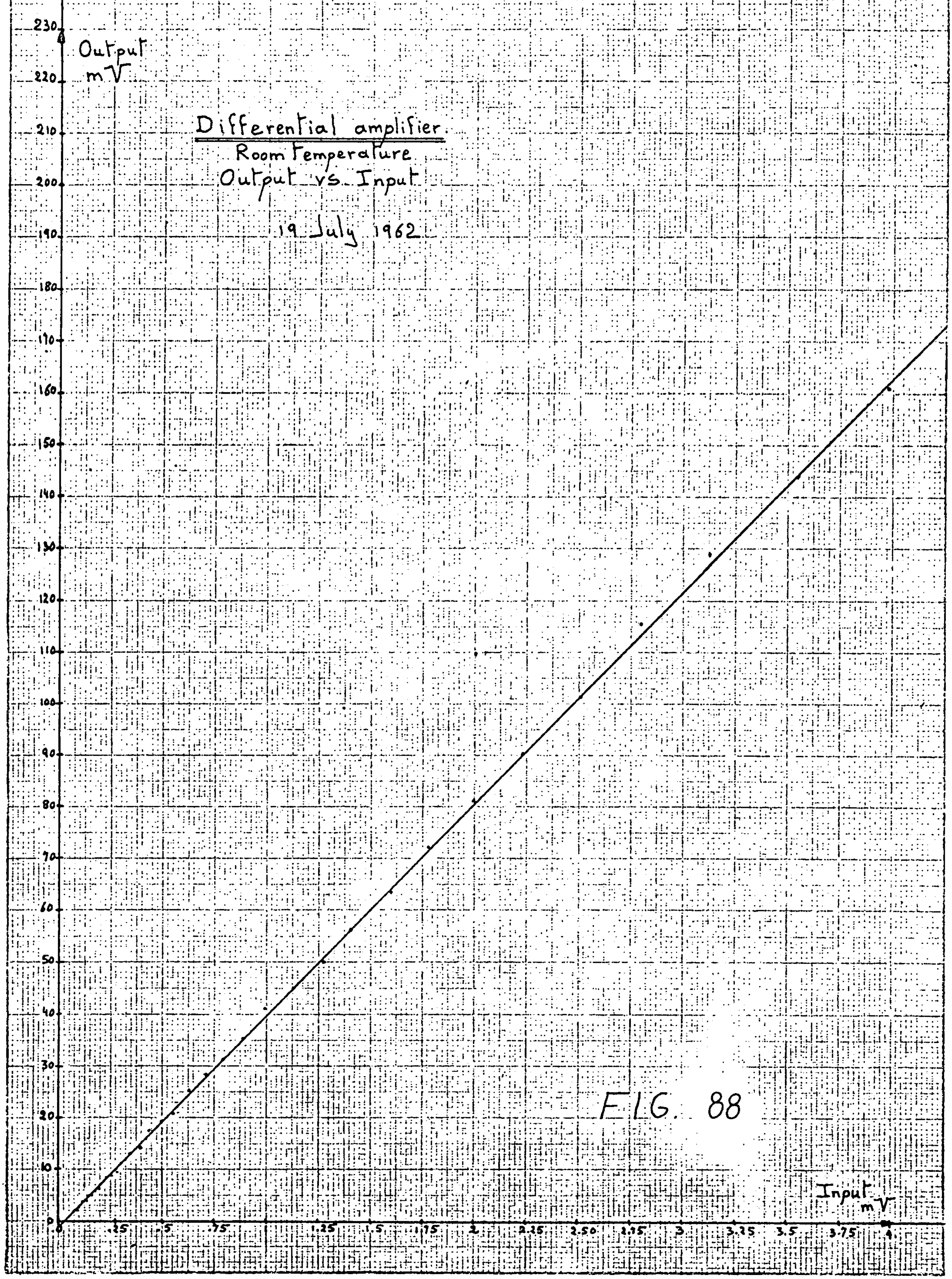

Vcc = +28V
14K
14K
e₀
2N1132
2N1132
T3
T4
75K
75K
50K
+28V
1/2 FSP-2
=2N1613
1/2 FSP-2
=2N1613
T1
T2
700Ω
+
300Ω
1K
270Ω
270Ω
e1
2N1613
T5
30K
39K
7.5K
VEE = -28V

FIG. 87

TABLE 10.1.- Differential Amplifier. Room temperature - Output vs. Input for two different days.

Attenuator db	Input mV	Output July 19 mV	Output July 23 mV	Diff. in Outputs mV	Attenuator db	Input mV	Output July 19 mV	Output July 23 mV	Diff. in Outputs mV
∞	0	≈ 0	≈ 0	0	93	0.627	25	24	1
110	0.0885	1	2	-1	92	0.703	28	27	1
109	0.0993	2	2	0	91	0.789	31	32	-1
108	0.112	3	2	1	90	0.885	35	35	0
107	0.125	3	3	0	89	0.993	41	39	2
106	0.140	4	3	1	88	1.115	45	45	0
105	0.157	4	4	0	87	1.250	50	50	0
104	0.176	5	4	1	86	1.403	56	56	0
103	0.198	6	5	1	85	1.575	63	63	0
102	0.222	8	6	2	84	1.767	72	71	1
101	0.249	8	7	1	83	1.982	81	80	1
100	0.280	9	9	0	82	2.224	90	89	1
99	0.314	11	11	0	81	2.490	101	101	0
98	0.352	13	13	0	80	2.800	115	113	2
97	0.396	14	16	-2	79	3.140	129	129	0
96	0.444	17	16	1	78	3.525	145	146	-1
95	0.498	19	19	0	77	3.955	162	163	-1
94	0.559	20	22	-2	76	4.440	182	182	0



Output mV
Differential amplifier
Room Temperature
Output vs Input
19 July 1962
230
220
210
200
190
180
170
160
150
140
130
120
110
100
90
80
70
60
50
40
30
20
10
0
.25
.5
.75
1
1.25
1.5
1.75
2
2.25
2.50
2.75
3
3.25
3.5
3.75
4
Input mV

FIG. 88

TABLE 10.2.- August 2, 1962. Differential Amplifier Output vs. Input.

INPUT		OUTPUT in mV							
Attenua-tor db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
∞	0	-113	-82	-56	-34	-20	-10	0	+22
110	0.0885	-109	-80	-55	-33	-17	- 8	+1	+23
109	0.0993	-109	-80	-55	-33	-17	- 8	1	23
108	0.112	-109	-80	-55	-32	-17	- 7	1	25
107	0.125	-109	-79	-54	-32	-16	- 7	2	25
106	0.140	-109	-78	-54	-31	-16	- 6	2	26
105	0.157	-108	-77	-53	-30	-15	- 5	3	27
104	0.176	-107	-77	-52	-30	-14	- 5	4	28
103	0.198	-105	-76	-51	-29	-13	- 4	6	29
102	0.222	-105	-75	-51	-28	-12	- 3	7	30
101	0.249	-103	-73	-50	-26	-11	- 1	9	31
100	0.280	-101	-72	-48	-25	-10	+ 1	10	32
99	0.314	-100	-70	-46	-23	- 7	+ 2	11	34
98	0.352	- 99	-69	-45	-22	- 6	3	12	35
97	0.396	- 97	-66	-43	-20	- 4	5	14	37
96	0.444	- 95	-64	-41	-18	- 2	7	16	39
95	0.498	- 92	-62	-39	-16	0	9	18	41
94	0.559	- 90	-60	-36	-14	+ 2	11	20	43
93	0.627	- 87	-55	-34	-11	5	14	24	46
92	0.703	- 84	-53	-31	- 7	7	17	27	48
91	0.789	- 80	-50	-27	- 4	11	21	31	52
90	0.885	- 75	-44	-23	0	15	25	35	57
89	0.993	- 71	-40	-17	+ 5	20	31	39	61
88	1.115	- 65	-35	-12	+10	25	36	44	66
87	1.250	- 60	-30	- 6	15	31	41	50	72
86	1.403	- 53	-23	0	22	37	47	56	78
85	1.575	- 46	-16	+ 7	28	44	54	63	85
84	1.767	- 38	- 8	+15	36	52	62	71	92

TABLE 10.2 Contd.- August 2, 1962. Differential Amplifier Output vs. Input.

INPUT		OUTPUT in mV							
Attenua-tor db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
83	1.982	- 28	+ 2	24	45	61	71	81	102
82	2.224	- 19	13	34	55	71	80	91	113
81	2.490	- 8	23	45	66	82	91	102	124
80	2.800	+ 5	36	58	79	95	104	114	136
79	3.140	+ 22	51	72	92	110	119	129	150
78	3.525	38	67	87	108	126	135	145	166
77	3.955	55	85	106	125	144	152	162	184
76	4.440	75	105	126	145	164	173	183	204
75	4.980	98	128	148	168	186	196	206	226
74	5.590	123	152	173	193	211	221	231	251
73	6.270	152	181	202	222	239	249	259	279
72	7.030	185	217	237	254	271	280	290	311
71	7.890	219	253	272	289	305	315	325	346
70	8.850	259	293	312	328	345	354	366	385
69	9.930	307	340	360	374	392	402	412	432
68	11.150	358	392	410	425	443	452	463	482
67	12.500	416	449	467	481	500	508	519	538
66	14.030	480	513	530	544	563	571	582	601
65	15.75	551	584	601	615	634	642	652	671
64	17.67	631	664	681	694	712	721	730	750
63	19.82	721	754	770	783	801	810	819	839
62	22.24	822	855	875	883	900	910	918	939
61	24.90	935	967	988	994	1012	1020	1028	1049
60	28.00	1062	1093	1113	1125	1136	1148	1153	1176
59	31.40	1214	1246	1263	1273	1288	1296	1306	1323
58	35.25	1374	1407	1423	1431	1447	1455	1464	1482
57	39.55	1554	1587	1601	1611	1625	1632	1641	1659

TABLE 10.2 Contd.- August 2, 1962. Differential Amplifier Output vs. Input.

INPUT		OUTPUT in mV							
Attenua-tor db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
56	44.40	1758	1788	1801	1809	1823	1831	1839	1857
55	49.80	1985	2014	2025	2033	2047	2054	2061	2079
54	55.90	2238	2267	2276	2283	2297	2304	2311	2329
53	62.70	2521	2551	2558	2564	2577	2586	2590	2607

TABLE 10.3.- August 3, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenuator db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
∞	0	-112	-81	-55	-35	-18	-6	+4	+20
110	0.0885	-112	-79	-52	-32	-15	-5	+8	+25
109	0.0993	-112	-79	-51	-32	-15	-4	8	+25
108	0.112	-111	-78	-51	-32	-14	-4	8	25
107	0.125	-111	-77	-50	-31	-14	-3	9	26
106	0.140	-110	-76	-50	-30	-13	-1	9	26
105	0.157	-109	-76	-49	-30	-12	0	10	27
104	0.176	-108	-75	-48	-29	-11	0	11	27
103	0.198	-107	-74	-47	-28	-11	+1	12	28
102	0.222	-106	-73	-46	-27	-10	+2	13	29
101	0.249	-104	-72	-45	-26	- 9	3	13	30
100	0.280	-102	-70	-44	-25	- 7	4	15	32
99	0.314	-101	-69	-42	-23	- 6	5	17	34
98	0.352	- 99	-67	-41	-21	- 4	7	19	35
97	0.396	- 97	-65	-39	-20	- 2	8	21	37
96	0.444	- 95	-63	-37	-18	0	10	23	39
95	0.498	- 92	-60	-35	-16	+ 3	13	25	41
94	0.559	- 89	-58	-32	-13	+ 6	15	27	43
93	0.627	- 86	-55	-29	-10	8	18	30	46
92	0.703	- 82	-53	-26	- 7	11	21	33	49
91	0.789	- 78	-49	-23	- 4	15	25	36	52
90	0.885	- 74	-44	-19	+ 1	19	29	40	57
89	0.993	- 69	-39	-14	+ 6	23	34	45	63
88	1.115	- 63	-34	- 9	11	28	39	50	68
87	1.250	- 57	-28	- 3	16	34	44	57	73
86	1.403	- 51	-22	+ 3	23	40	51	63	79
85	1.575	- 44	-15	+ 9	29	47	58	70	86
84	1.767	- 35	- 7	17	37	55	66	78	94

TABLE 10.3 Contd.- August 3, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenuator db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
83	1.982	- 26	+ 2	26	46	64	74	86	102
82	2.224	- 16	+12	36	56	74	84	96	112
81	2.490	- 5	23	48	67	85	95	108	123
80	2.800	+ 8	36	60	80	97	107	120	135
79	3.140	+ 23	51	75	95	114	122	135	150
78	3.525	39	67	91	111	130	139	150	167
77	3.955	57	85	109	129	147	156	168	185
76	4.440	78	106	129	149	167	176	188	205
75	4.980	100	128	152	172	189	198	211	227
74	5.590	126	153	177	197	214	225	236	251
73	6.270	154	182	205	225	242	253	264	279
72	7.030	186	213	237	256	274	284	294	310
71	7.890	222	249	272	291	309	319	330	347
70	8.850	262	289	312	331	348	359	370	387
69	9.930	310	337	360	379	395	406	417	432
68	11.150	361	388	411	429	445	455	467	482
67	12.500	418	445	467	485	503	513	523	538
66	14.030	482	508	530	548	566	576	585	601
65	15.75	553	579	601	619	637	646	656	671
64	17.67	633	659	681	698	715	725	734	750
63	19.82	722	748	770	788	804	813	823	838
62	22.24	823	849	869	887	903	912	922	938
61	24.90	935	961	981	998	1014	1023	1033	1049
60	28.00	1066	1086	1106	1130	1144	1147	1153	1177
59	31.40	1214	1239	1258	1277	1290	1299	1309	1323
58	35.25	1374	1399	1417	1435	1449	1458	1466	1481

TABLE 10.3 Contd.- August 3, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenua-tor db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
57	39.55	1554	1578	1596	1614	1627	1635	1643	1657
56	44.40	1755	1778	1796	1813	1826	1834	1841	1855
55	49.80	1980	2003	2019	2037	2049	2058	2064	2079
54	55.90	2232	2255	2270	2287	2299	2309	2314	2328
53	62.70	2515	2537	2552	2568	2579	2588	2593	2607

TABLE 10.4.- August 6, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenuator db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
∞	0	-121	-85	-56	-38	-20	-15	-5	+17
110	0.0885	-120	-84	-56	-35	-18	-11	-2	+21
109	0.0993	-120	-83	-55	-35	-18	-11	-1	21
108	0.112	-119	-83	-55	-35	-17	-10	-1	22
107	0.125	-118	-82	-54	-34	-17	-10	0	22
106	0.140	-117	-82	-53	-33	-16	-10	0	22
105	0.157	-116	-81	-53	-33	-16	- 9	+1	23
104	0.176	-115	-80	-52	-32	-15	- 7	+2	23
103	0.198	-114	-78	-51	-31	-15	- 6	2	24
102	0.222	-113	-77	-50	-30	-14	- 6	3	25
101	0.249	-112	-76	-49	-29	-13	- 4	4	26
100	0.280	-111	-75	-47	-27	-12	- 3	5	27
99	0.314	-109	-73	-45	-26	- 9	- 1	7	29
98	0.352	-106	-72	-43	-25	- 6	0	9	31
97	0.396	-105	-70	-42	-22	- 6	+ 1	11	33
96	0.444	-103	-68	-40	-20	- 4	+ 4	12	35
95	0.498	-101	-66	-37	-18	- 2	6	15	37
94	0.559	- 98	-63	-34	-15	0	8	17	39
93	0.627	- 95	-60	-32	-13	+ 3	12	20	42
92	0.703	- 92	-57	-28	- 9	+ 6	17	25	44
91	0.789	- 89	-54	-25	- 6	10	18	28	48
90	0.885	- 85	-49	-20	- 2	14	22	32	53
89	0.993	- 83	-44	-15	+ 3	19	26	36	58
88	1.115	- 78	-39	- 9	+ 8	25	31	41	62
87	1.250	- 72	-33	- 3	14	30	38	46	68
86	1.403	- 67	-27	+ 3	21	37	44	52	74
85	1.575	- 60	-20	+10	28	44	51	59	81
84	1.767	- 52	-11	18	36	52	59	68	89

TABLE 10.4 Contd.- August 6, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenua-tor db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
83	1.982	- 43	- 3	27	45	60	68	76	97
82	2.224	- 34	+ 8	37	55	70	78	86	107
81	2.490	- 22	+19	48	66	81	89	97	118
80	2.800	- 10	32	61	78	94	102	112	131
79	3.140	+ 5	47	75	94	109	115	127	147
78	3.525	+ 21	63	91	109	125	130	142	162
77	3.955	38	81	109	127	143	148	160	179
76	4.440	58	101	129	147	163	168	179	199
75	4.980	81	124	152	169	185	190	202	221
74	5.590	106	149	178	195	210	215	227	246
73	6.270	135	178	206	223	238	243	254	274
72	7.030	166	209	238	255	270	274	285	305
71	7.890	201	245	272	290	305	310	320	341
70	8.850	241	286	313	330	345	349	360	381
69	9.930	292	333	360	377	392	397	407	428
68	11.150	343	384	410	428	443	447	458	477
67	12.500	400	440	467	484	499	504	514	534
66	14.030	464	503	530	547	562	567	577	596
65	15.75	535	575	601	619	632	637	647	666
64	17.67	615	655	681	698	712	716	726	745
63	19.82	704	744	769	787	800	805	814	833
62	22.24	804	844	869	886	899	904	914	932
61	24.90	916	956	981	997	1010	1015	1024	1043
60	28.00	1048	1086	1112	1122	1140	1139	1149	1168
59	31.40	1196	1235	1259	1275	1289	1290	1299	1319
58	35.25	1356	1395	1418	1434	1447	1448	1458	1476

TABLE 10.4 Contd.- August 6, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenuator db	Input mV	20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
57	39.55	1535	1574	1597	1612	1625	1626	1635	1653
56	44.40	1736	1773	1797	1812	1824	1824	1833	1851
55	49.80	1961	1998	2021	2035	2047	2047	2056	2073
54	55.90	2214	2249	2272	2286	2298	2297	2306	2322
53	62.70	2498	2532	2553	2566	2579	2577	2585	2601

TABLE 10.5.- August 13, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenuator db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
∞	0	-113	-76	-55	-33	-17	-11	-2	+17
110	0.0885	-111	-73	-53	-30	-15	- 8	+1	+20
109	0.0993	-110	-73	-52	-30	-15	- 8	+1	+20
108	0.112	-109	-73	-52	-29	-14	- 7	2	21
107	0.125	-109	-72	-52	-29	-13	- 7	2	21
106	0.140	-108	-71	-52	-28	-13	- 6	3	22
105	0.157	-106	-71	-51	-28	-13	- 5	4	23
104	0.176	-105	-70	-50	-27	-12	- 5	5	24
103	0.198	-105	-70	-49	-26	-11	- 4	6	25
102	0.222	-103	-68	-48	-25	-11	- 3	7	27
101	0.249	-102	-67	-47	-24	-10	- 2	8	28
100	0.280	-101	-66	-46	-23	- 8	- 1	10	29
99	0.314	- 99	-64	-44	-22	- 5	0	11	29
98	0.352	- 97	-62	-43	-20	- 4	+ 3	13	31
97	0.396	- 95	-61	-41	-18	- 3	+ 4	14	33
96	0.444	- 93	-59	-39	-16	0	6	17	34
95	0.498	- 91	-56	-37	-14	+ 2	8	19	37
94	0.559	- 88	-54	-34	-12	+ 5	11	22	39
93	0.627	- 85	-51	-31	- 9	7	14	23	41
92	0.703	- 82	-48	-26	- 6	11	18	25	44
91	0.789	- 78	-44	-23	- 2	14	21	29	48
90	0.885	- 74	-40	-19	+ 2	18	25	34	54
89	0.993	- 68	-35	-14	+ 7	20	30	39	60
88	1.115	- 63	-30	- 7	11	25	35	44	64
87	1.250	- 57	-25	- 4	17	31	40	50	69
86	1.403	- 51	-18	+ 4	23	37	46	56	76
85	1.575	- 44	-11	+10	30	44	53	62	81
84	1.767	- 36	- 3	20	38	52	62	70	90

TABLE 10.5 Contd.- August 13, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenua-tor db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
83	1.982	- 27	+ 6	29	47	62	70	79	99
82	2.224	- 17	+15	39	57	70	80	89	109
81	2.490	- 5	26	47	68	84	91	99	118
80	2.800	+ 7	39	61	81	95	103	112	132
79	3.140	23	55	75	96	109	118	127	146
78	3.525	39	71	91	112	125	134	142	163
77	3.955	57	88	108	129	144	152	161	182
76	4.440	77	109	128	149	164	172	182	202
75	4.980	99	131	151	172	185	194	204	224
74	5.590	125	156	176	196	210	219	228	248
73	6.270	153	184	204	225	237	247	256	275
72	7.030	185	216	236	256	269	278	287	307
71	7.890	221	252	273	291	303	312	322	342
70	8.850	263	293	315	332	344	351	361	382
69	9.930	309	339	362	378	391	398	409	426
68	11.150	360	390	411	428	441	447	458	476
67	12.500	417	447	468	485	497	505	514	532
66	14.030	480	510	532	547	559	567	577	594
65	15.75	552	581	602	618	630	637	647	666
64	17.67	631	661	682	697	709	715	727	744
63	19.82	721	750	770	786	797	804	815	833
62	22.24	821	849	869	886	896	904	914	933
61	24.90	933	961	979	997	1008	1013	1024	1042
60	28.00	1064	1091	1110	1127	1137	1143	1153	1169
59	31.40	1212	1239	1258	1273	1283	1288	1299	1315
58	35.25	1372	1399	1417	1431	1441	1446	1456	1473

TABLE 10.5 Contd.- August 13, 1962. Differential Amplifier. Output vs. Input.

INPUT		OUTPUT in mV							
Attenua-tor db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
57	39.55	1551	1577	1595	1609	1619	1623	1633	1650
56	44.40	1751	1777	1795	1807	1817	1821	1830	1847
55	49.80	1976	2002	2018	2031	2040	2043	2052	2069
54	55.90	2228	2253	2269	2281	2289	2292	2301	2319
53	62.70	2511	2535	2549	2561	2569	2571	2580	2599

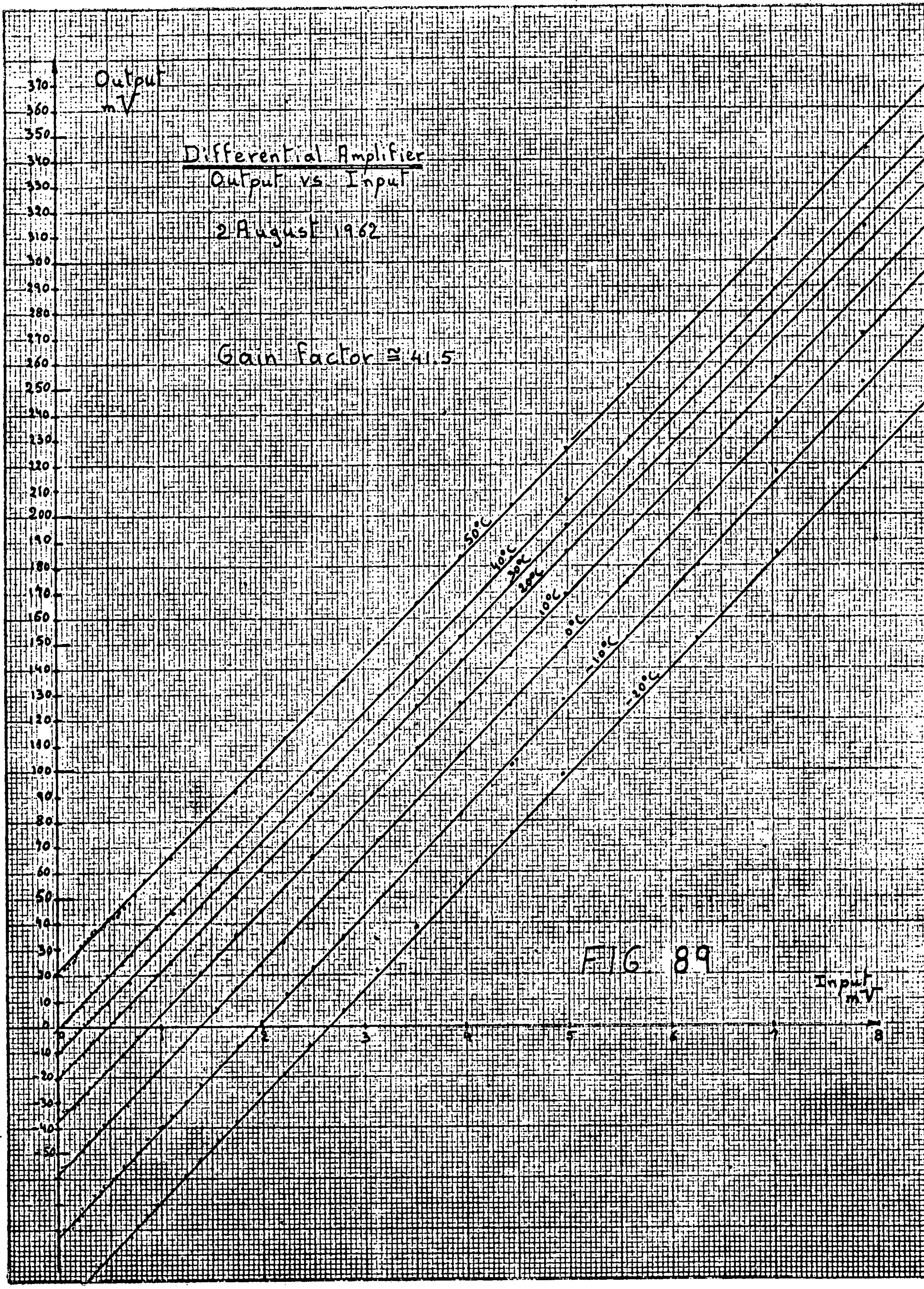

Output mV
Differential Amplifier
Output vs Input
2 August 1962
Gain Factor ≅ 41.5
50°C
40°C
30°C
20°C
10°C
0°C
-10°C
-20°C
Input mV

FIG. 89

TABLE 10.6.- Differential Amplifier. Table with expressions giving the regression lines for the several groups of data obtained.

Temp.	Group 1 (2 August)	Group 2 (3 August)	Group 3 (6 August)	Group 4 (13 August)
-20°C	Y = 42.0X-112.5 RMS = 47.5 μV	Y = 42.0X-111.2 RMS = 85.0 μV	Y = 41.9X-124.3 RMS = 65.0 μV	Y = 41.9X-110.5 RMS = 65.8 μV
-10°C	Y = 42.0X-81.8 RMS = 68.5 μV	Y = 41.9X-81.0 RMS = 51.3 μV	Y = 41.8X-85.5 RMS = 58.5 μV	Y = 41.7X-76.8 RMS = 42.8 μV
0°C	Y = 41.8X-58.8 RMS = 63.7 μV	Y = 41.6X-55.5 RMS = 43.5 μV	Y = 41.7X-56.9 RMS = 58.2 μV	Y = 41.7X-55.7 RMS = 60.5 μV
10°C	Y = 41.5X-37.4 RMS = 49.0 μV	Y = 41.6X-35.9 RMS = 45.8 μV	Y = 41.6X-38.0 RMS = 52.0 μV	Y = 41.4X-34.5 RMS = 39.2 μV
20°C	Y = 41.5X-20.9 RMS = 46.8 μV	Y = 41.5X-18.0 RMS = 42.8 μV	Y = 41.5X-22.0 RMS = 46.0 μV	Y = 41.3X-19.8 RMS = 45.0 μV
30°C	Y = 41.5X-11.4 RMS = 37.5 μV	Y = 41.4X-7.6 RMS = 44.5 μV	Y = 41.4X-14.7 RMS = 42.8 μV	Y = 41.2X-11.8 RMS = 36.5 μV
40°C	Y = 41.4X-1.7 RMS = 54.3 μV	Y = 41.3X+4.3 RMS = 48.3 μV	Y = 41.4X-5.1 RMS = 45.8 μV	Y = 41.2X-2.3 RMS = 40.5 μV
50°C	Y = 41.3X+20.4 RMS = 38.2 μV	Y = 41.3X+20.8 RMS = 38.6 μV	Y = 41.3X+16.2 RMS = 44.7 μV	Y = 41.2X+16.9 RMS = 36.5 μV

TABLE 10.7.- Differential Amplifier using 4 precision resistors. Output vs. Input.

INPUT		OUTPUT in mV							
Attenua-tor db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
∞	0	- 60	- 36	- 30	- 17	- 9	- 7	- 3	- 2
110	0.0885	- 57	- 35	- 28	- 15	- 6	- 4	- 1	0
109	0.0993	- 56	- 34	- 26	- 14	- 6	- 3	- 1	+ 1
108	0.112	- 56	- 32	- 26	- 14	- 5	- 2	0	+ 1
107	0.125	- 55	- 32	- 25	- 14	- 4	- 2	0	2
106	0.140	- 54	- 31	- 24	- 13	- 4	- 1	+ 1	2
105	0.157	- 54	- 30	- 23	- 12	- 3	0	+ 2	3
104	0.176	- 53	- 29	- 22	- 11	- 2	0	3	4
103	0.198	- 51	- 28	- 20	- 10	0	+ 1	4	5
102	0.222	- 51	- 27	- 19	- 8	+ 1	+ 3	5	5
101	0.249	- 50	- 26	- 18	- 7	+ 2	4	7	8
100	0.280	- 49	- 25	- 16	- 6	3	7	9	8
99	0.314	- 46	- 22	- 14	- 4	4	8	11	10
98	0.352	- 44	- 20	- 13	- 1	6	9	12	11
97	0.396	- 41	- 18	- 10	0	8	12	14	13
96	0.444	- 40	- 14	- 8	+ 4	9	14	17	15
95	0.498	- 37	- 12	- 5	+ 7	12	17	19	17
94	0.559	- 34	- 10	- 2	9	14	19	22	20
93	0.627	- 31	- 7	+ 1	13	18	23	25	23
92	0.703	- 28	- 3	+ 5	16	22	26	29	27
91	0.789	- 23	+ 1	10	21	26	30	33	31
90	0.885	- 19	+ 6	15	25	32	34	37	35
89	0.993	- 13	11	20	30	38	40	42	39
88	1.115	- 8	17	26	37	43	46	48	45
87	1.250	0	23	33	42	49	53	55	51
86	1.403	+ 7	30	40	50	56	60	62	58
85	1.575	15	39	48	58	64	67	69	66

TABLE 10.7 Contd.- Differential Amplifier using 4 precision resistors. Output vs. Input.

INPUT		OUTPUT in mV							
Attenuator db	Input mV	-20°C	-10°C	0°C	10°C	20°C	30°C	40°C	50°C
84	1.767	24	47	57	67	73	76	78	74
83	1.982	34	58	67	77	83	86	88	86
82	2.224	46	70	80	88	94	97	100	97
81	2.490	58	85	93	101	107	111	112	109
80	2.800	73	99	107	115	122	125	126	124
79	3.140	89	114	123	132	138	141	142	140
78	3.525	107	133	141	150	156	159	160	158
77	3.955	127	153	162	170	176	179	180	178
76	4.440	150	175	185	192	199	202	202	200
75	4.980	176	201	210	217	225	227	227	225
74	5.590	204	229	239	247	253	255	255	253
73	6.270	236	262	270	279	284	287	288	285
72	7.030	272	298	306	314	320	322	323	320
71	7.890	312	338	346	354	360	362	362	359
70	8.850	357	384	393	402	407	408	408	405
69	9.930	411	438	445	454	460	460	460	457
68	11.150	469	495	503	511	517	517	517	513
67	12.500	535	559	566	577	581	581	581	577
66	14.030	607	631	638	647	653	652	651	647
65	15.75	688	712	718	727	733	732	730	727
64	17.67	778	802	808	817	822	821	819	815
63	19.82	879	904	909	919	923	921	919	915

TABLE 10.8.- Differential Amplifier. Influence of positive voltage supply changes. (Outputs in mV).

Input \ Positive voltage in Volts	26.00	26.50	27.00	27.50	28.00	28.50	29.00	29.50	30.00
0.198 mV	2	2	4	5	6	7	8	9	10
0.498 mV	15	17	17	18	17	19	20	22	24
0.993 mV	35	37	37	38	38	41	41	43	43
1.982 mV	76	76	79	78	79	82	81	84	85
4.980 mV	200	202	202	203	204	205	206	208	208

TABLE 10.9.- Differential Amplifier. Influence of negative voltage supply changes (Outputs in mV).

Input \ Positive voltage in Volts	-26.00	-26.50	-27.00	-27.50	-28.00	-28.50	-29.00	-29.50	-30.00
0.198 mV	-8	0	-2	2	6	10	12	18	21
0.498 mV	4	8	10	16	18	21	25	30	32
0.993 mV	27	27	31	41	38	43	45	48	56
1.982 mV	70	68	72	76	79	83	86	90	94
4.980 mV	190	192	198	200	203	208	212	217	215