ELECTRONIC ENGINEERING

VOL. 36 No. 440

OCTOBER 1964

Commentary

THE S.B.A.C. Flying Display and Exhibition took place at Farnborough in the early part of September. It was the twenty-fourth display to be held and the first since the society changed its name to the Society of British Aerospace Companies.

This Display and Exhibition is of course the British Aerospace Industry's main 'shop window' and since its inception in 1932 it has grown enormously in size and popularity. In that year it was held at Hendon and 34 aircraft were on show and 21 member companies displayed their products on trestle tables in two tents. Attendance was by invitation only and was less than 1 000. In 1933, '34 and '35 Displays were also held at Hendon while in 1936 and '37 they were held at Hatfield which provided a longer runway and better exhibition facilities. The Displays were resumed after the war and in 1947 and '48 were held at Radlett. In 1948 the Society decided to open the Display to the public on certain days and it became necessary to find an airfield with good road approaches, adequate accommodation and more advanced flying control facilities. The Royal Aircraft Establishment, Farnborough, was made available on a rental basis and the first 'Farnborough' took place in September that year. On that occasion 163 companies took part in the Exhibition and it was visited by 14 000 guests, including 800 from overseas and by 100 000 members of the public. By 1962 the numbers had risen to 348 exhibitors, 90 000 guests (10 000 from overseas) and 250 000 members of the public. There was no 'Farnborough' in 1963 but this year's exhibition was larger than ever and, while attendance figures are not yet known, they undoubtedly broke all previous records.

As the S.B.A.C. Exhibition has grown in size so has its importance to the electronic industry increased and at this year's event electronics played a major role. Some idea of the importance of the aerospace industry to electronics can be gained from the fact that while the output of the capital goods sector of the U.K. electronic industry in 1963 was some £200M about £57M of this went into aviation. Aviation also accounts for a sizeable total of electronic exports and the figures in this sector have risen from £13M in 1961 to £21·4M in 1963. It is rather sad and somewhat paradoxical to note that, in the same period, the total exports of the aerospace industry have fallen from £149M to £112M.

Apart from the direct financial gain the aerospace industry also affects the electronic industry in another and highly beneficial way; it provides the impetus for the development of techniques that lead to the maximum in reliability. Thus from this year's Exhibition it is apparent that solid state and microminiaturization techniques, particularly thin films, are rapidly becoming more widespread in use. So far as rockets and missiles are concerned miniaturization is an end in itself but for general aircraft use the savings in size and weight are only secondary advantages and the outstanding feature of these new techniques is their inherent reliability.

The success with which the electronic industry is meeting the stringent demands made upon it can be judged from the way aircrews now have faith in their 'black boxes' and from the ever increasing use of electronics in all classes of aerospace vehicles. Indeed most aerospace developments are embarked upon on the assumption that the electronic engineers will be able to devise means of dealing with the difficult tasks envisaged for them; often tasks which can be solved in no other way.

In view of the close collaboration that is now necessary between the aerospace and electronic industries it has been decided to strengthen the cordial relations which have always existed between the two and to this end the Electronic Engineering Association and the Society of British Aerospace Companies have formed a joint committee at Presidential level to deal with questions of over-riding importance concerning aerospace electronics. This Committee, called the British Aerospace Electronics Joint Policy Committee, will consider matters of major policy where the aerospace and electronics industries are initially concerned, and where necessary will negotiate with the Ministry of Aviation and other bodies at home and abroad.

However, it would be wrong to pretend that all is well in the aviation section of the electronic industry and one of the main causes of dissatisfaction is the disregard for the welfare of the U.K. electronic industry shown by the Government when they concluded the Anglo-French treaty for the Concord. While it is agreed that there must be a great deal of give and take in any joint venture of this sort the fact remains that by placing the major responsibility for the electronics, communications and navigation systems with the French they have put our industry at a grave disadvantage and made it difficult for the Concord project to enjoy the services of the people most experienced in these fields.

During the course of the exhibition a brighter note was sounded at Government level by the announcement by Mr. Amery, Minister of Aviation, that Britain is to have its own space programme based on a super version of the Black Knight rocket. At the moment insufficient details are known to allow detailed comment and perhaps one's first reaction is surprise that such a programme is now possible whereas only a very short time ago it was deemed to be economically impossible. Nevertheless, it is very welcome news to the aerospace and electronic industries and gives real meaning to the word 'Aerospace' in the new title of the S.B.A.C.

The Design of Transistor-Resistor Logic Circuits by Graphical Means

By C. W. M. Barrow*, B.Sc.(Eng), A.M.I.E.E.

A graphical solution is described for NOR *transistor-resistor logic circuits. The* NOR *circuit requires that the transistor is non-conducting when all the input potentials of the input leads are at earth, but is bottomed if the potential of one of the input leads falls to a potential more negative than a certain threshold value. The calculations include the effects of tolerances in resistors, power supplies, and transistor parameters. The results are used in the design of a multi-stage counter using* NOR *circuit elements.*

(Voir page 726 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 733)

In designing large data processing equipments it is convenient to make them up of a large number of simple logic circuits. Circuits such as coincidence gates have been designed using transistors, others with diode-resistor, or resistor combinations followed by transistor amplifiers where necessary. Fig. 1 shows a gating circuit formed by four resistors followed by a transistor inverting amplifier.

If the potential of any of the input terminals falls below $-V$, then the transistor will conduct, and the collector potential will rise towards that of earth. This method of using the circuit is known as NOR gating, since a negative stimulus on *A* or *B* or *C* will make the transistor conduct and the prefix *N* implies that the output is an inversion of the input.

The NOR circuit using transistors and resistors uses a minimum number of semiconductors and for this reason can be cheaper, and if the resistors are conservatively rated may be more reliable[1]. They are however slow in operation. For data processing equipments where reliability and cheapness are desirable, but where high speed is unessential, transistor-resistor logic (t.r.l.) may be the most suitable.

T.R.L. has the added advantage that individual circuits can be designed fairly rigorously. Interest has already been shown in design methods, and a number of papers have been written on the subject[2,3,4,5]. A computer solution of the NOR circuit is described in one[2] and a graphical solution of a t.r.l. coincidence circuit described in another[3]. However, no simple solution which takes into account tolerances on resistors, power supplies and transistors seems yet to have been published. This article presents a graphical solution for the design of NOR circuits.

Graphical methods are particularly suitable for solution of design problems since the effects of relaxing or narrowing the tolerances may easily be seen. If by suitable choice of parameters, the graphs can be simplified to straight lines only two calculations are necessary for each condition. This has been done in the method described.

NOR Circuit

In all but the simplest data processing systems it is necessary to have logic circuits in tandem. Fig. 2 shows a NOR circuit being fed from and feeding similar circuits. There are two conditions to be met, if such a circuit is to function satisfactorily, i.e. under the worst-case conditions:

(1) An input signal on any one of the input leads should bottom the transistor VT_1. This is referred to as the 'on' condition.

(2) With no input signal the transistor will be cut-off. It has to perform these two functions whether it is feeding one circuit or feeding a maximum of *n* inputs to similar circuits.

SYMBOLS

I_b = Transistor base current
I_{co} = Base leakage current with emitter disconnected
I_{cl} = Collector leakage current
I_{eo} = Emitter leakage
$I_{co(max)}$ = Maximum value of I_{co}
$I_{cl(max)}$ = Maximum value of I_{cl}
I_{on} = Nominal value of current in R_L when the transistor is bottomed
I_{off} = Nominal value of current in R_L when the transistor is non-conducting
$I_{on(max)}$ and $I_{off(max)}$ = Maximum value of the preceding two items
m = Number of inputs
n = Number of outputs
R = Nominal value of input resistor
R_B = Nominal value of bias resistor
R_d = Incremental resistance of the base-emitter diode
R_L = Nominal value of collector load resistor
V = Nominal value of input signal amplitude
V_b = Voltage between the base and emitter
V_c = Nominal collector clamp potential
$V_{c(min)}$ = Minimum collector potential
V_d = Delay voltage of the base-emitter diode
V_i = Maximum bottoming voltage across transistor from emitter to collector
V_n = Nominal negative supply potential
V_p = Nominal positive supply potential
V_x = Minimum voltage between the base and emitter when the transistor is non-conducting
x = Tolerance of voltage supplies
y = Tolerance of resistors
β = Minimum value of current gain in the common emitter configuration
σ = $x + y$
I_B = V_p/R_B
I = $V_{c(min)}/R$

$$A = \frac{V_n(1+\sigma)}{(V_n(1-x)-V_{c(min)})(1-y)}$$

$$C = 1-\sigma-\frac{V_x}{V_p(1+y)}$$

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'OFF' CONDITION

The worst case for the 'off' condition is obtained when the values of the circuit parameters are taken on the extremes of their tolerances in such a way that the positive bias on the base is as small as possible.

That is, R_B becomes $R_B(1+y)$,

R becomes $R(1-y)$,

V_p becomes $V_p(1-x)$,

and I_{co} becomes $I_{co(max)}$.

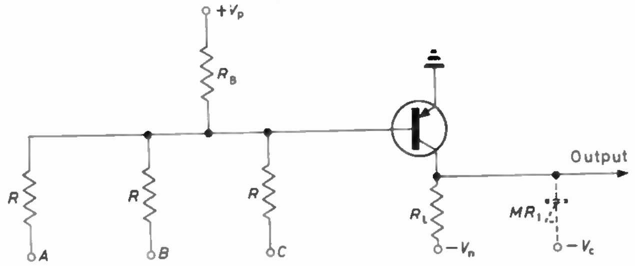


Fig. 1. TRL NOR circuit

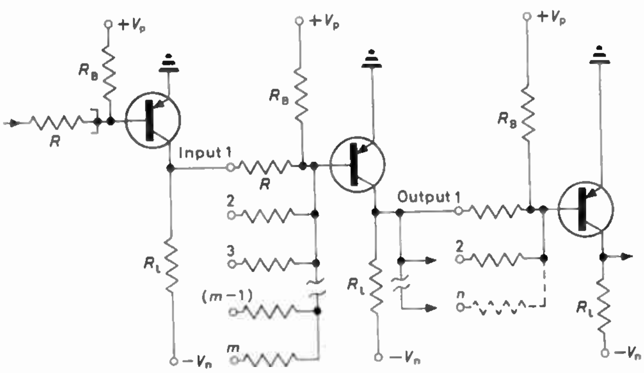


Fig. 2. NOR gating circuit in tandem

Using the nominal values, the 'off' condition is given by

$$\frac{V_b+V_i}{R/m}+I_{co}=\frac{V_p-V_b}{R_b}\quad\ldots\ldots\ldots\ldots(1)$$

Substituting the extreme values of R_B, R and I_{co} into equation (1) the worst conditions are given by:

$$\frac{V_b\,.\,m}{R(1-y)}+\frac{V_i\,.\,m}{R(1-y)}+I_{co(max)}=\frac{V_p(1-x)-V_b}{R_B(1+y)}\quad\ldots\ldots\ldots\ldots(2)$$

This condition can be simplified, as shown in Appendix (1), to give

$$I_B/I\geqslant(1/C)\left[\frac{m(V_x+V_i)}{V_c(1-y)}+\frac{I_{co(max)}}{I}\right]\ldots\ldots\ldots(3)$$

'ON' CONDITION

Using the nominal values of the circuit parameters, the 'on' condition is given by

$$\frac{V_c-V_b}{R}=\frac{V_p+V_b}{R_B}+\frac{V_b}{R/(m-1)}+I_b\ldots\ldots\ldots(4)$$

The worst case for the 'on' condition is obtained when the values of the circuit parameters are taken on the extremes of their tolerances in such a way that the base current is a minimum and the collector current a maximum. This condition occurs when:

R_B becomes $R_B(1-y)$

R_L becomes $R_L(1-y)$

V_p becomes $V_p(1+x)$

V_n becomes $V_n(1+x)$

V_c becomes $V_{c(min)}$.

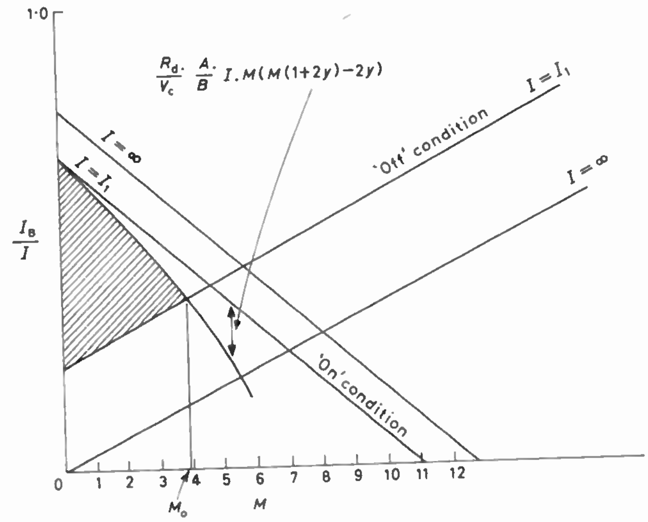
$(m-1)$ resistors R which are connected to the collectors of the bottomed transistors become $R(1-y)$.

The resistor R connected to the negative output from the preceding stage becomes $R(1+y)$.

Substituting these values into equation (4), together with the condition that $i_b\geqslant I_{on(max)}/\beta$ where β is the minimum end of life figure, the worst-case 'on' condition is obtained, i.e.

$$\frac{V_{c(min)}-V_b}{R(1+y)}-\frac{V_p(1+x)+V_b}{R_B(1-y)}-\frac{V_b}{R(1-y)/(m-1)}\geqslant I_{on(max)}/\beta$$

This conditon can be simplified in a similar manner to that



'On' Condition:

Intercept on M axis $=\dfrac{1-y-(A/\beta)\,I_{c1(max)}/(I_1(1+\sigma))}{V_d/V_{c(min)}\,(1+2y)+(A/\beta)(1+y)}$

Intercept on I_B/I *axis* $=\dfrac{1-y}{1+\sigma}-(A/\beta)\dfrac{I_{c1(max)}}{I_1(1+\sigma)}$

'Off' condition:

Intercept on I_B/I *axis* $=I_{co(max)}/CI_1$

Equation of 'off' condition

$$I_B/I=1/C\left[\frac{M(V_x+V_i)}{V_{c(min)}(1-y)}+\frac{I_{co\,max)}}{I_1}\right]$$

Fig. 3. Graphical design of NOR circuit
$n=m=M$

shown in Appendix (1) and using the value of $I_{on(max)}$ calculated in Appendix (2) it becomes

$$I_B/I\leqslant\frac{1}{1+\sigma}\left\{1-y-\frac{V_d}{V_{c(min)}}[m(1+2y)-2y]-(An/\beta)(1+y)-\frac{R_d}{V_{c(min)}}A/\beta\,In\,(m(1+2y)-2y)\right\}-A/\beta\,\frac{I_{c1(max)}}{I(1+\sigma)}\ldots\ldots(5)$$

For a first approximation the effect of the resistance of the base-emitter diode R_d may be ignored by leaving out the term $R_d/V_{c(min)}\;A/\beta\;In\;(m(1+2y)-2y)$, and by using the knowledge that $m(1+2y)\gg 2y$. This gives the more approximate relationship:

$$I_B/I<\frac{1}{1+\sigma}\left[1-y-\frac{V_d}{V_{c(min)}}m(1+2y)-(A/\beta)\,n(1+y)\right]-A/\beta\,\frac{I_{c1(max)}}{I(1+\sigma)}\ldots\ldots\ldots\ldots\ldots(6)$$

Graphical Solution of the NOR Circuit

A graphical solution of equations (3) and (6) is shown in Fig. 3 for the particular case of $n = m = M$. The axes have been chosen so that both equations are represented by straight lines. The 'on' condition limit line may be obtained by calculating the values of the intercepts on the I_B/I and M axis. The 'off' conditions may be calculated from the intercept on the I_B/I axis and from calculating a second value of I_B/I for a finite value of M. Two solutions are shown one for $I = I_1$ and one for $I = \infty$. The result of including the term $R_d/V_{c(min)}\ A/\beta\ In\ m(1+2y)$ is also shown. The shaded area shows the range of values of I_B/I for which the circuit would function for particular values of M. As M is increased the range gets more restricted until there is only one possible value of I_B/I for the maximum value of $M = M_o$.

In the solution of the design of a NOR circuit using this method, it is necessary to choose a suitable value of I. Increasing it has some effect in increasing M. It can be seen, however, that in Fig. 3 the maximum value of M for $I = \infty$ is 8. There are some advantages in increasing the current I, but the limitations of dissipation and transistor collector current put an upper limit on the choice.

A value of $V_{c(min)}$ to obtain a maximum value of M_o may be chosen by the following calculation. If the right-hand sides of equations (3) and (6) are equated and the value of dM/dV_c found, this may be equated to zero and the resulting quadratic in $V_{c(min)}$ solved. One of the roots will give the value of $V_{c(min)}$ which corresponds to the maximum value of M_o if all other circuit parameters are held constant. This equation is given below.

$$V_{c(min)} = -\frac{V_n(1-x)PS \pm \sqrt{[V_n(1-x)PS\,[V_n(1-x)PS - \{Q-PV_n(1-x)\}\ \{R-S\}]]}}{P(R-S)} \qquad \text{(7)}$$

where $P = \dfrac{1-y}{1+\sigma} - (I_{co(max)}/CI)$

$$Q = \frac{V_n\, I_{c1(max)}}{I(1-y)\beta}$$

$$R = \frac{V_n(1+2y)}{\beta}$$

$$S = \frac{V_d(1+2y)}{(1+\sigma)} + (1/C)\frac{V_x+V_i}{(1-y)}$$

Fig. 4. Graphical design of NOR circuit
$n \neq m$ but where n is known

'On' condition:

Intercept on I_B/I axis =

$$\frac{V_{c(min)}}{(1+\sigma)(1+2y)}\left[\frac{1-y-(An/\beta)(1+y)-(A/\beta)I_{c1(max)}/I_1}{V_d+R_d(A/\beta)In}\right]$$

Intercept on I_B/I axis =

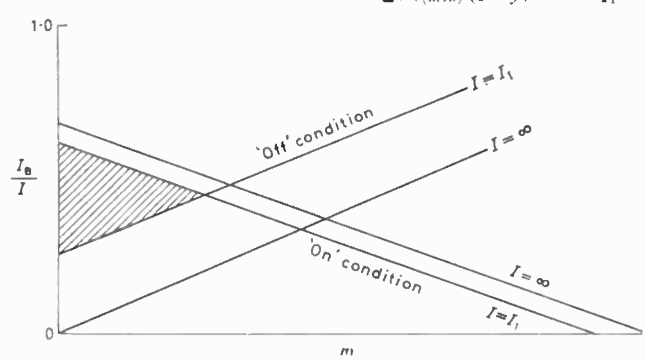
$$\frac{1-y}{1+\sigma} - (A/\beta)n\frac{(1+y)}{1+\sigma} - (A/\beta)\frac{I_{c1(max)}}{I_1(1+\sigma)}$$

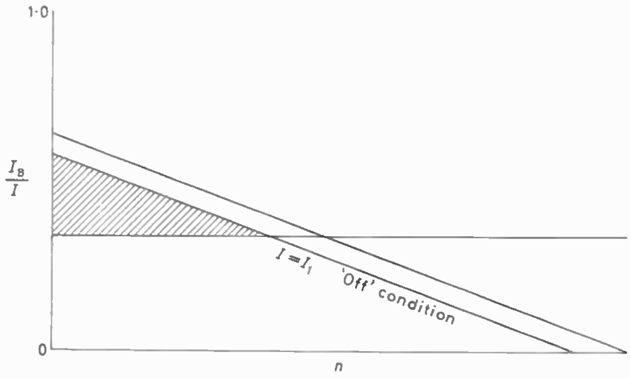
'Off' condition:

Intercept on I_B/I axis = $I_{co(max)}/CI_1$

Equation of 'off' condition

$$I_B/I = 1/C\left[\frac{m(V_x+V_i)}{V_{c(min)}(1-y)} + \frac{I_{co(max)}}{I_1}\right]$$





'On' condition:

Intercept on n axis =

$$\frac{\beta}{A(1+\sigma)}\left[\frac{1-y-(V_d/V_{c(min)})\,m\,(1+2y)-(A/\beta)I_{c1\,max}/I_1}{1+y+(R_d/V_{c(min)})\,Im\,(1+2y)}\right]$$

Intercept on I_B/I axis =

$$\frac{1}{1+\sigma}\left[1-y-\frac{V_{c(min)}}{V_d}m(1+2y)-(A/\beta)\frac{I_{c1(max)}}{I_1(1+\sigma)}\right]$$

'Off' condition:

Intercept on I_B/I axis = $1/C\left[\dfrac{m(V_x+V_i)}{V_{c(min)}(1-y)} + \dfrac{I_{co(max)}}{I_1}\right] = a$

Equation of 'Off' condition $I_B/I = a$

Fig. 5. Graphical design of NOR circuit
$n \neq m$ but where m is known

The intercepts for the 'on' condition may be calculated and drawn and the intercept and a second point may be calculated for the 'off' condition from equation (3).

The value of the correction term

$$(R_d/V_{c(min)})\,(A/\beta)\,I_1M^2\,(1+2y)$$

may now be calculated for the range of values below the intersection of the on and off limit conditions. It can then be subtracted from the vertical ordinates at the corresponding value of M.

It may be that due to consideration of the logical design, the number of inputs m or the number of outputs n is known, and a design is required where n is not necessarily equal to m. The calculation takes a similar course to that described previously for $n = m = M$.

Firstly the value of $V_{c(min)}$ for the maximum number of inputs or outputs must be calculated. This can be done using the same technique as before.

TABLE 1

x	=	0·05
y	=	0·03 (Total drift expected from a grade 1 resistor of 1 per cent initial selection tolerance)
σ	=	0·05 + 0·03 = 0·08
V_n	=	50V
V_D	=	50V
V_c	=	8V unless otherwise stated
V_i	=	0·1V
V_x	=	0·25V
β	=	23 (Estimated end-of-life figure)
I	=	0·34mA at 50°C
$I_{1(max)}$	=	0·5mA at V_b = 0·250V positive and at 50°C

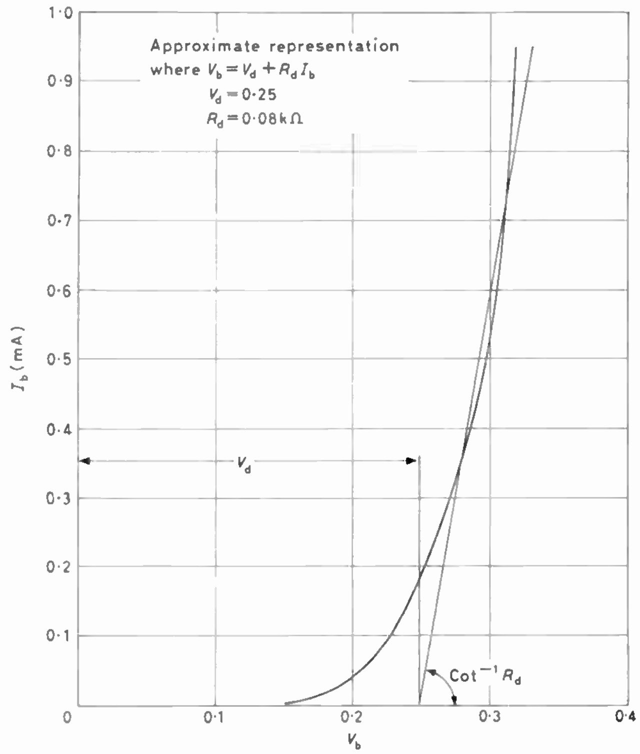


Fig. 6. Base/emitter diode characteristic of a high frequency alloy germanium transistor (see Table 1)

Approximate representation

where $V = V_d + R_d I_b$

$V_d = 0{\cdot}25$

$R_d = 0{\cdot}08k\Omega$

If n is known then the maximum value of m will be obtained for the following value of $V_{c(min)}$.

$$V_{c(min)} = V_n(1-x)\left[1-\sqrt{\left(\frac{((1-x)/(1-y))\,1/\beta\,[n(1+y)+(I_{c1(max)}/CI)]}{(1-y)/(1+\sigma)-(I_{co(max)}/CI)}\right)}\right] \qquad (8)$$

If m is known then the maximum value of n will be obtained for the following value of $V_{c(min)}$.

$$V_{c(min)} = \sqrt{\left[\frac{V_n\,.\,m\,[((1+2y)/(1+\sigma))\,V_d + (V_x+V_i)/C(1-y)]}{[(1-y)/(1+\sigma)-(I_{co(max)}/I)]}\right]} \qquad (9)$$

Fig. 4 shows the graphical solution of the problem of obtaining a value of m when n is known and Fig. 5 gives the value of n if m is known.

Example

In order to demonstrate the use of these graphical methods the solution of a design problem is included using a transistor of the OC44 type. This transistor has fairly low leakage currents between emitter and base and emitter and collector which greatly assists the design of a NOR circuit. The specification of the relevant characteristics of of the transistor and the circuit are listed in Table 1.

In order to obtain values of V_d and R_d a special test had to be made on a transistor which had a low gain ($\beta = 35$ at $I_c = 20$mA). The collector current was adjusted at each setting of base current so that the transistor was just bottomed. This was considered to simulate as near as possible the conditions for which the calculations are made. Fig. 6 shows the results of this test.

The Design of a Counter using NOR-Circuit Elements

Circuit Description

A counter having 16 positions was required. It was decided that a realizable design could most likely be achieved by making two counters each with four positions and feeding the carry from one into the other. The 16 outputs would then be obtained by combining the outputs of one counter with those of the other. Fig. 7 shows one ring of this counter, the second being identical to it. The counter has $m + 1$ stages. An output circuit showing the coincidence gate formed by MR_2, MR_3, and R_i and emitter-follower amplifier transistor VT_A is typical of $(m+1)^2$ possible outputs. Each counter element has m inputs connected via resistors to the collectors of each of the other elements. The resistor from the previous stage is shunted by a capacitor.

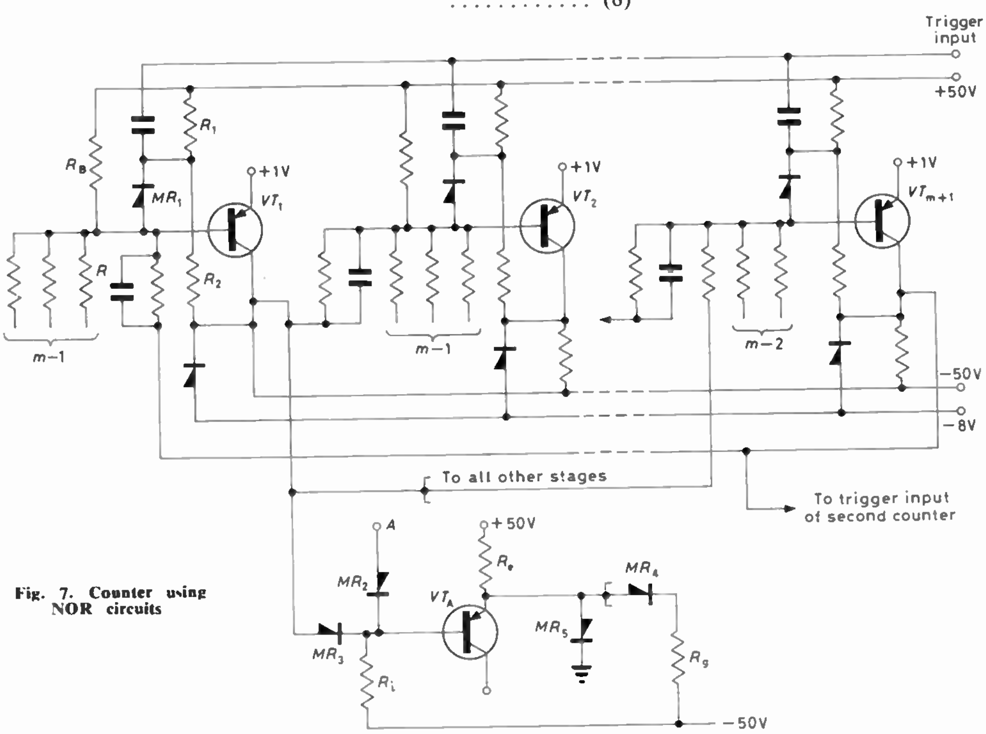


Fig. 7. Counter using NOR circuits

The operation of the circuit is as follows. Normally all transistors are bottomed except one, VT_1 say, which is cut-off. The current in the collector load resistor of this transistor is therefore holding the rest of the transistors on. The potential of the anode of MR_1 is more negative than that of the rest of the trigger input diodes. When the negative I/P pulse appears on the trigger input lead, VT_1 is brought on and the collector potential rises. This cuts the substantial transistor VT_2 off before it cuts-off the other transistors since the capacitor C increases the charge injected into the base region of VT_2. VT_2 cuts-off and the current in the collector resistor of VT_2 keeps all the rest of the transistors conducting. Thus the count has progressed by one unit and VT_2 is cut-off instead of VT_1.

The design calculation takes the following course. The current supplied to the output amplifiers, from each counter stage is first established. The number of inputs to each stage is known, i.e. 3, and the number of outputs is $3 + (I_o + I_1)/I$, where I_o is the current supplied to the output amplifiers, and I_1 is the current taken by the input bias resistors R_1 and R_2 and I is the nominal input current. The graphical solution for the element can now be made (Fig. 8) for 3 values of I. A choice of one if these values is taken and use is made of the base-emitter characteristic shown in Fig. 6 to get a more exact solution.

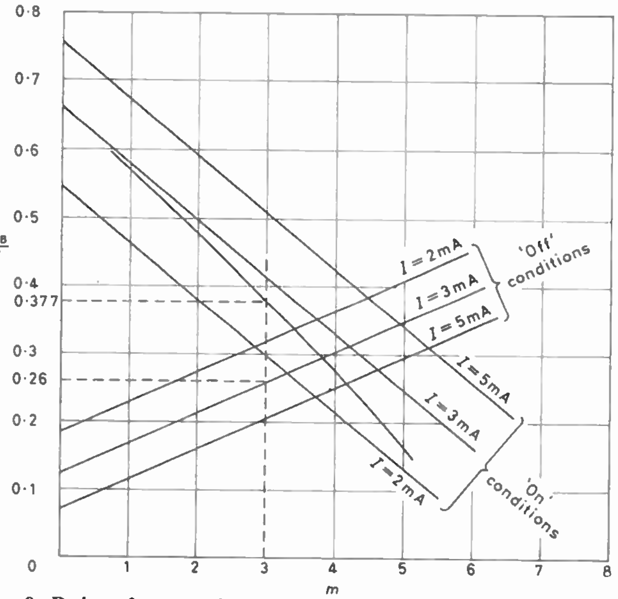


Fig. 8. Design of a stage in the counter shown in Fig. 7, using a high frequency alloy germanium transistor (see Table I)

The Calculation

The supply voltages and tolerances are shown in Table 1. Maximum gate current through R_g

$$=(V_n/R_g)\,(1+\sigma) = (50/18)\,(1{\cdot}08)\text{mA, since } R_g = 18\text{k}\Omega.$$

Maximum gate current through 10 gates =

$$(50/18) \times 1{\cdot}08 \times 10\text{mA}$$

Minimum current through $R_e = \dfrac{V_n(1-x) - 0{\cdot}4}{R_e(1+y)}$ mA

assuming a 0·4V drop across MR_5.

This can be equal to the maximum gate current for 10 gates.

Thus $\dfrac{47{\cdot}5 - 0{\cdot}4}{R_e x\, 1{\cdot}03} = (50/18) \times 1{\cdot}08 \times 10,$

and $R_e = 1{\cdot}55\text{k}\Omega$.

R_e is chosen as 1·5kΩ nominal.

Maximum current through $R_e = \dfrac{(V_p + V_c)}{R_e}\,(1+\sigma)$

$$= (50 + 8/1{\cdot}5) \times 1{\cdot}08$$
$$= 41{\cdot}8\text{mA}.$$

The base current must equal $41{\cdot}8/23 = 1{\cdot}82$mA assuming a minimum β of 23.

Minimum current through $R_1 = 1{\cdot}82$mA.

Maximum value of $R_1 = \dfrac{V_n(1-x) - V_c}{1{\cdot}82}\,(1-x) = 21{\cdot}2\text{k}\Omega$

The nominal value of R_1 was chosen to be 18kΩ. Maximum value of current through R_1, I_o, when the counter element feeding the amplifier is cut off

$$= \frac{V_n(1+x) - V_c(1-x)}{(R_1/4)\,(1-y)}$$
$$= \frac{50(1{\cdot}05) - 8(1{\cdot}05)4}{18\,(1-0{\cdot}02)}$$
$$= 10{\cdot}5\text{mA}.$$

The current taken by R_1 and R_2. $I_1 = \dfrac{V_n(1+x) + V_c(1-x)}{(R_1 + R_2)\,(1-y)}$

$$= \frac{50(1{\cdot}05) + 8\,(0{\cdot}95)}{(39+9{\cdot}1)\,(1-0{\cdot}03)} = 1{\cdot}3\text{mA and } n = m + \frac{I_o + I_1}{I}.$$

'On' conditions are given by the equation

$$I_B/I \leqslant \frac{1}{1+\sigma}\left[(1-y) - \frac{V_d}{V_{c(min)}}\,m(1+2y) - (A/\beta)\,\frac{(m+I_o+I_1)}{I}\,(1+y)\right] - (A/\beta)\,\frac{I_{c1}}{I(1+\sigma)}$$

found by substituting for n and $V_{c(min)} = V_c\,(1-x)$ in equation (6). This simplifies to:

$$I_B/I \leqslant$$
$$\frac{1}{1+\sigma}\,[1 - y - m\,\{(V_d/V_{c(min)})\,(1+2y) + (A/\beta)\,(1+y)\}]$$
$$-(A/\beta I)\left[\frac{(I_o + I_1)\,(1+y) + I_{c1(max)}}{1+\sigma}\right] \quad \ldots (10)$$

Where:

$$\frac{1-y}{1+\sigma} = \frac{1-0{\cdot}03}{1+0{\cdot}08} = 0{\cdot}896$$

$$A/\beta = \frac{V_n(1+2\sigma)}{(V_n - V_c)\beta} = \frac{58}{(50-8)\,23} = 0{\cdot}061$$

$$V_d/V_{c(min)}\,(1+2y) = \frac{0{\cdot}25 \times 1{\cdot}06}{9(1-0{\cdot}05)} = 0{\cdot}028$$

$$I_o + I_1 = 10{\cdot}5 + 1{\cdot}3 = 11{\cdot}8\text{mA}.$$

$$A/\beta \left[\frac{(I_o + I_1)\,(1+y) + I_{c1(max)}}{1+\sigma}\right] = 0{\cdot}061 \times \frac{11{\cdot}8 \times 1{\cdot}03 + 0{\cdot}5}{1{\cdot}08} = 0{\cdot}704$$

Substituting the values of these terms into equation (10), the following relationship is obtained:

$$I_B/I < 0{\cdot}896 - m\,(0{\cdot}082) - (0{\cdot}704/I)$$

'Off' conditions are given by equation (3),

$$I_B/I \geqslant (1/C)\left[\frac{m(V_x + V_i)}{V_c\,(1-y)} + \frac{I_{co(max)}}{I}\right]$$

where $C = 1 - \sigma - (V_x/V_p(1+y))$. Substituting in the values of the circuit parameters the 'off'-condition limit line is obtained,

$$I_B/I \geqslant m(0{\cdot}045) + 0{\cdot}372/I.$$

The limit lines for the 'on' and 'off' conditions are shown in Fig. 8, for $I = 2$, 3 and 5mA. The term $(R_d/V_{o(min)})\,(A/\beta)\,.\,n\,.\,I\,(m(1+2y) - 2y)$ (see equation (5)) is calculated for $1 = 3$mA and added to the limit line for 3mA. n is taken to be

$$m + \frac{I_o + I_1}{I} = m + 11{\cdot}8/3 = m + 3{\cdot}9$$

The value of R_d is obtained from Fig. 6.

$$R_d/V_c\;A\,.\,n\,.\,I\,\frac{(m(1+2y) - 2y)}{1+\sigma}$$

$$= 0{\cdot}08/9 \times \frac{0{\cdot}061 \times 3}{1{\cdot}08} \times 1{\cdot}01 \times m \times n,$$

ignoring $R_d/V_c(A/\beta) \cdot n \cdot I \cdot 2y$. This becomes

$$R_d/V_c(A/\beta) \cdot n \cdot I\, m(1+2y) = \cdot 00121\, m^2 + \cdot 0047\, m,$$

as $n = m + 3{\cdot}9$.

This correction is then added to the 'on'-condition limit line for $I = 3$mA.

The limits of I_B/I are found to be 0·26 to 0·377 for $I = 3$mA and $m = 3$.

Therefore I_B can have any value between $0{\cdot}26 \times 3$ and $0{\cdot}377 \times 3$mA, i.e. from 0·795mA to 1·13mA.

R_B can therefore have any value between $50/1{\cdot}13 = 43{\cdot}8$kΩ and $50/0{\cdot}765 = 62{\cdot}8$kΩ.

Actually 62kΩ was chosen.

The load resistor may now be calculated (see Appendix (2)).

$$R_L = \frac{(V_n - V_c)(1-\sigma)}{nI(1+y-x) + I_{cl(max)}}$$

$$n = m + \frac{I_o + I_1}{I} = m + 11{\cdot}8/3 = 3 + 3{\cdot}93 = 6{\cdot}93$$

$$R_L = \frac{(51-9)(0{\cdot}92)}{6{\cdot}93 \times 3(1-0{\cdot}01) + 0{\cdot}5}$$

$$= \frac{42 \times 0{\cdot}92}{21{\cdot}5} = 1{\cdot}8\text{k}\Omega$$

Conclusions

This article shows that general formulae may be found to assist the design of transistor-resistor circuits under worst case conditions. It may be argued that the worst case condition is too pessimistic a basis for design. This basis, however, is the only one which has any degree of certainty. Unless the probability distribution of resistors, power supplies and transistor characteristics are known in advance, the probability of the circuit failing owing to component values changing cannot be calculated.

The value of a graphical method for the solution of the equations lies in the fact that the effect of relaxing or tightening one or more of the conditions may be readily seen. It also makes it easy to obtain the more exact solution involving the nmI term of equation (5), which enters the calculation if R_d is taken into account.

In these calculations all the resistors are assumed to have the same tolerances. Differing values for the tolerances of R_B, R_L and input resistors R may be catered for by slight adjustments to the equations.

This method may, of course, be used for designing the d.c. conditions for a bistable trigger circuit, which is the special case of $m = 1$. Transient conditions were not dealt with but certain considerations arise. In general, a range of values of I_B/I will be found as solutions. The bottom limit will give the largest I_b and will improve the rate at which the transistor comes into conduction, but will slow up the 'turn off' time due to the excess holes in the base region. The lower limit will give the least value of I_B/I which will give a slower 'turn on' time but faster 'turn off' time. This effect may be significant when compared with the large variation of I_b caused by different numbers of signals coinciding at the input. The effect of the different number of coincident input signals on the value of I_b can be reduced if the rectifier MR_1, Fig. 1, is included. This minimizes and stabilizes the value of I taken by each output.

Acknowledgments

Acknowledgment is made to the Directors of the A.E.I. (Woolwich) Ltd, for permission to publish this article. Thanks are also due to Dr. J. E. Flood and Mr. G. H. Parks for encouragement and help in its preparation.

Appendix (1)

Equation (2) is:

$$\frac{V_b \cdot m}{R(1-y)} + \frac{V_1 \cdot m}{R(1-y)} + I_{co(max)} = \frac{V_p(1-x) - V_b}{R_B(1+y)}$$

with rearrangement, this becomes,

$$\frac{1}{m/(R(1-y)) + 1/(R_B(1+y))}\left[\frac{V_p(1-x)}{R_B(1+y)} - \frac{V_1 \cdot m}{R(1-y)} - I_{co(max)}\right] = V_b$$

In the worst case $V_b \geqslant V_x$.

Using this inequality together with the following relationships,

$$V_p/R_B = I_B,\ R = V_{c(min)}/I,\ R_B = V_p/I_B$$

Then

$$\frac{V_p(1-x)}{R_B(1+y)} - \frac{V_1 \cdot m}{R_B(1+y)} - I_{co(max)} \geqslant V_x \frac{Im}{V_{c(min)}(1-y)} + \frac{I_B}{V_p(1+y)}$$

Making the approximation

$$1 - x/(1+y) \simeq 1 - \sigma$$

and dividing by I,

$$I_B/I \geqslant \frac{1}{1-\sigma-(V_x/V_p)(1+y)}\left[\frac{m(V_x+V_1)}{V_{c(min)}(1-y)} + \frac{I_{co(max)}}{I}\right]$$

is obtained.

Appendix (2)

To calculate $I_{ON(max)}$:

When the maximum number of succeeding stages n are connected to the junction of R_L and the collector, the total current flowing through these resistors will be nominally nV_c/R and $nV_{c(min)}/(R(1-y))$ in the extreme. The current in R_L will be a maximum of

$$\frac{nV_{c(min)}}{R(1-y)} + I_{ol(max)}$$

where $I_{ol(max)}$ is the leakage current flowing from the collector of the transistor.

When R_L is taking the maximum current, V_n can rise positively to $V_n(1-x)$ and R_L may become $R_L(1+y)$. The collector voltage under this condition must not rise more than $-V_{o(min)}$. The following relationship may therefore be derived:

$$\frac{V_n(1-x) - V_{c(min)}}{R_L(1+y)} = \frac{nV_{o(min)}}{R(1-y)} + I_{cl(max)} \quad \ldots\ldots \text{(A1)}$$

$$= nI(1+y-x) + I_{cl(max)}$$

The maximum collector current in the 'on' state is:

$$\frac{V_n(1+x)}{R_L(1-y)} = I_{ON(max)} \quad \ldots\ldots\ldots\ldots \text{(A2)}$$

This assumes that no subsequent stages are connected. Eliminating the value of R_L from equations (A1) and (A2) and simplifying,

$$I_{ON(max)} = AnI(1+y-x) + I_{cl(max)}$$

$$\text{where } A = \frac{V_n(1+\sigma)}{(V_n(1-x) - V_{c(min)})(1-y)}$$

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Electronic Assessment of Pasture Growth

By F. J. Hyde*, D.Sc., M.I.E.E., F.Inst.P., and J. T. Lawrence*

A description is given of the design and construction of an instrument for the non-destructive measurement of the growth of grass. The capacitance of a probe is changed when it is placed on the grass and this causes a change in the frequency of the transistor oscillator to which it is connected. The difference between this frequency and that of an identical fixed frequency oscillator at 3Mc/s is converted to a deflexion on a d.c. microammeter which is approximately proportional to the capacitance change. It has been shown that the meter deflexion is correlated with the total water content of the grass under the probe. Laboratory measurements have been made on growing grass to throw light on the mechanism of the capacitance change.

(Voir page 727 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 734)

THE standard method of investigating the rate of growth of grass involves the periodic cutting of part of the plot. Such destructive testing is unsatisfactory. A non-destructive method was described by Campbell *et al*[1] in 1962. These workers discovered that a capacitance change could be produced in a suitably designed probe placed over a patch of grass; this was correlated to a first order with the fresh, dry and organic weight of the grass under the probe. In their instrument the capacitance change was used to alter the frequency of a valve oscillator attached to the probe head. This frequency was compared with that of a transistor oscillator held by the operator.

An instrument of improved performance and design has now been produced and is described in this article. The results of laboratory measurements on growing grass are also presented. It is shown that the change in capacitance produced by the grass is not solely a dielectric effect in the useful frequency range. It is also due to the conductance of the grass, which effectively emphasizes the capacitances between the electrodes and the grass adjacent to them.

Design of the Probe

The main structural requirements are that the electrodes of the probe should be longer than the highest grass to be measured and that the area covered should be sufficiently large to avoid positional errors over a non-uniform sward. To ensure good sampling over the measuring area, and to facilitate the placing of the probe without unduly disturbing the grass, it is convenient to use an electrode assembly comprised of rods.

The actual form of construction used is shown in Fig. 1. Sixteen rods are arranged uniformly in four rows of four, with 5in spacing between rows. The rods are of ¼in diameter brass and are 14½in long; they are coated with insulating varnish. Blocks of Tufnol 1in in diameter are fixed to the bottom of the rods to keep their ends approximately ½in above the ground. The rods pass through a Tufnol top plate of dimensions 19in by 19in by ½in. Alternate rods are connected together above this plate to form the electrode system. The value of the probe capacitance with air as the dielectric is 73pF. The plate is surmounted by a cover consisting of a shallow aluminium box which carries a switch and houses a variable standardizing capacitor. This can be switched in place of the probe for setting the zero of the instrument

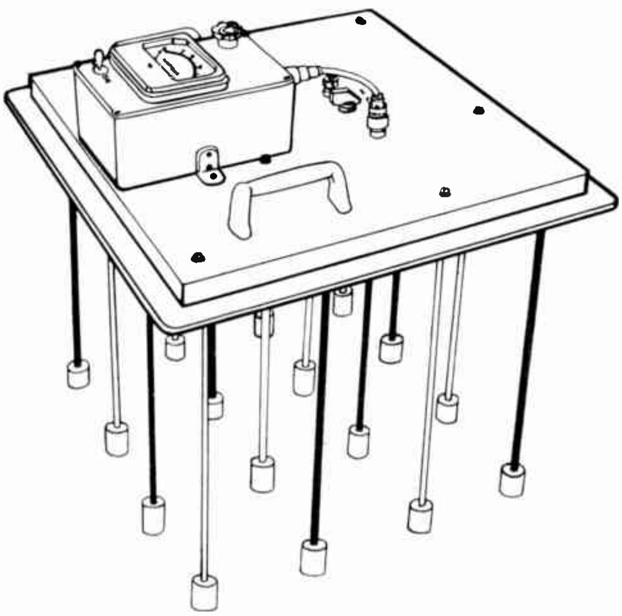

Fig. 1. Instrument assembly

when in use in the field, and for calibration purposes.

The electronic unit, incorporating the output micro-ammeter is fixed on top of the cover and is connected to the probe by a coaxial cable.

The Electronic Unit

Basically the unit consists of a 3Mc/s reference oscillator, a variable frequency oscillator, for which the probe forms part of the tuning capacitance, a beat frequency detector, a clipping amplifier and 'diode-pump' frequency meter with a d.c. output indicator.

The circuit is built on Veroboard panels with the tuned circuits contained in a small die-cast box. The whole unit

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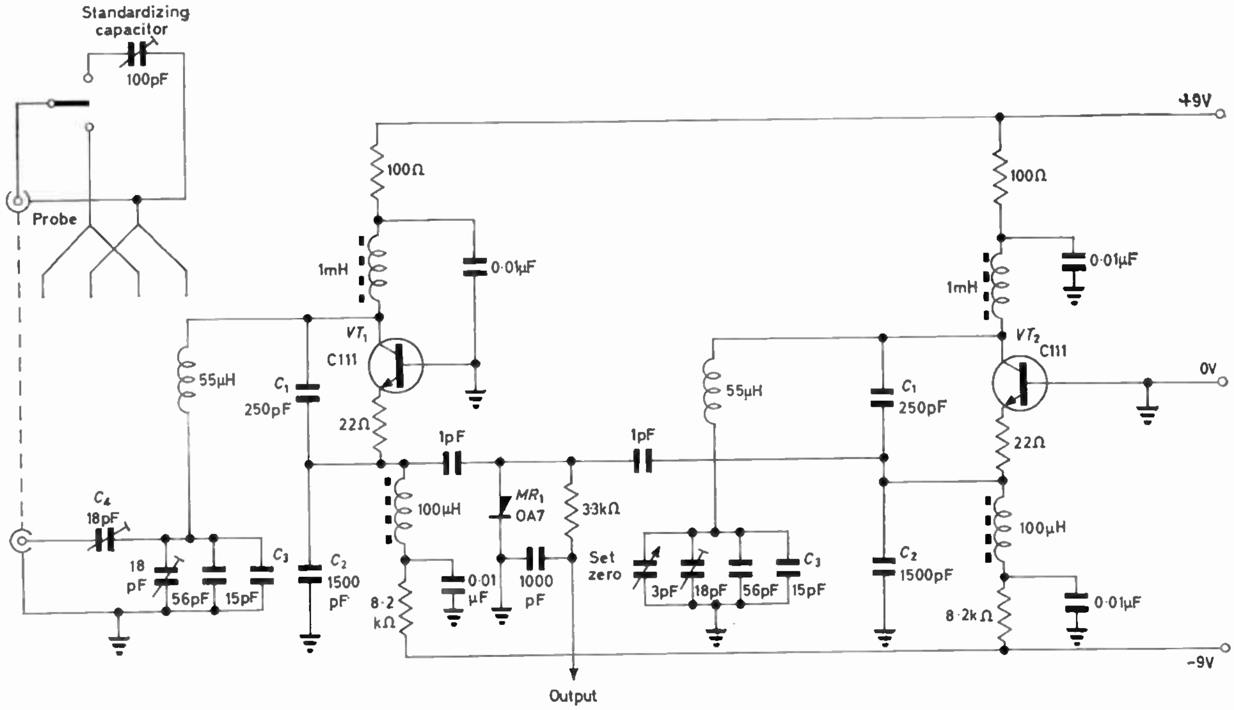

Fig. 2. Oscillator circuit

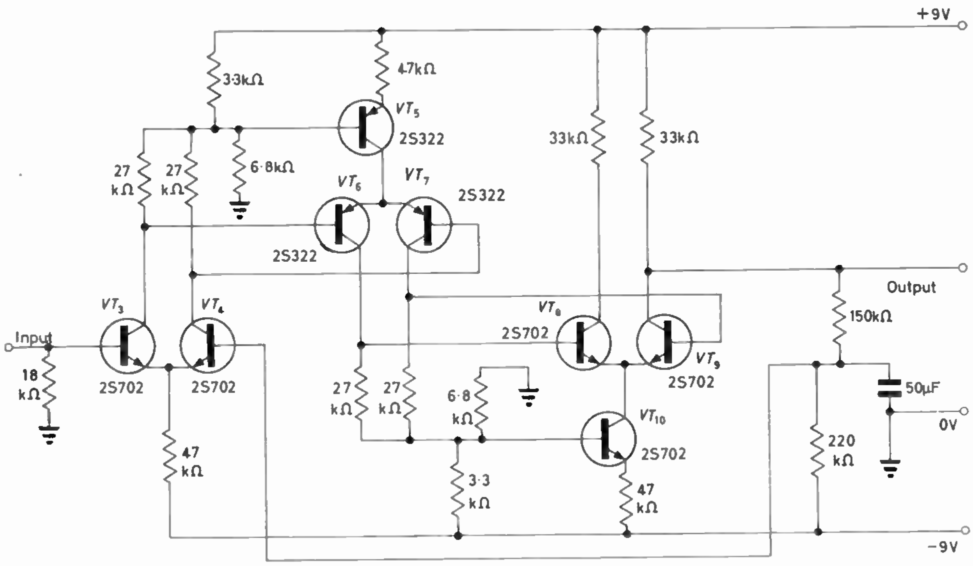

Fig. 3. Clipping amplifier

and batteries are housed in a larger die-cast box of dimensions 7¼in by 4½in by 3in. The 50μA, 4in moving-coil output meter is incorporated in the lid together with a zero-set control and on/off switch. As the output of the unit is a direct current, which depends on the difference frequency of two oscillators, it is important that they should have good differential frequency stability. As a first step in this direction good stability of the individual oscillators was aimed for. Silicon npn transistors type C111 were used for the oscillator circuits, which are shown in Fig. 2. Decoupled 8·2kΩ resistors are included in series with each emitter to provide constant current feed: this stabilizes the d.c. operating conditions against temperature changes. A.C. feedback resistance of 22Ω is included in each emitter circuit to reduce the changes in input impedance with temperature. The values of emitter current and collector voltage were chosen to give minimum frequency shift with changes in temperature; these values were approximately 1mA and +9V respectively.

Both oscillators are identical mechanically and electrically. A common-base series-tuned Clapp circuit is used. The capacitances of the feedback capacitors C_1 and C_2 swamp the transistor capacitances and therefore reduce the effect of any changes in the latter on the resonant conditions. Capacitors C_3 having a negative temperature coefficient of capacitance are incorporated in each tuned circuit to offset changes due to the net positive temperature coefficient of the effective inductance of the remainder of the circuit. Both circuits are mounted in the inner die-cast box to ensure that any temperature difference between them is small. For each oscillator it was found that the frequency stability was $-4 \times 10^{-6}/°C$ for changes in transistor temperature alone and $-10^{-5}/°C$ for changes in tuned circuit temperature alone. Although this could be improved it was considered adequate. The differential frequency stability of the pair of oscillators against ambient temperature change was found to be satisfactory. Although it varied somewhat with the mean temperature it was found to be of the order of $10^{-6}/°C$ in the temperature range 30 to 50°C. The probe of the variable frequency oscillator is connected to the remainder of the tuned circuit via a series capacitor C_4. This is set to provide a suitable change in frequency and associated output meter deflexion for the largest expected change in probe capacitance. The actual value of C_4 is about 8pF. As is shown in the Appendix, this type of coupling ensures that the change in oscillator frequency is approximately linearly related to change in probe capacitance.

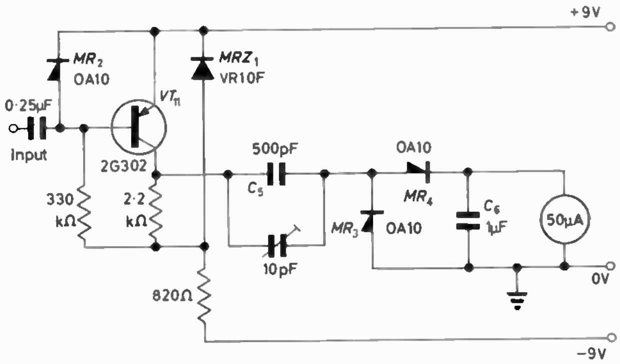

Fig. 4. 'Diode pump' frequency meter

The beat frequency detector incorporates a diode MR_1, which is loosely coupled to both oscillators. Coupling is effected at corresponding low impedance points of the oscillator circuit to prevent frequency pulling. After filtering, the output of the detector is amplified and clipped to give a square wave for input to the frequency meter. The circuit of the clipping amplifier is shown in Fig. 3. It consists essentially of three long-tailed pairs of transistors in cascade. The pairs are alternately npn and pnp and are directly coupled. A large feedback resistance is connected in the common emitter circuit of the first pair. Transistors VT_5 and VT_{10} ensure constant current feeds to the second and third long-tailed pairs at which the signal level is higher. The quiescent emitter current is chosen so that bottoming cannot arise, in view of the limited current available for each pair. Clipping of the signal waveform therefore only occurs because a transistor cuts off. This results in a very clean square wave output over the frequency range 10c/s to 10kc/s and from 3mV to 1V input. D.C. feedback from the output to the input, which is decoupled for signal frequencies, ensures that the operation of the circuit is almost unaffected by temperature changes. The overall current gain is of the order of 10^5.

Accuracy of counting depends on the constancy of the amplitude of the input waveform at the difference frequency, which is applied to the diode pump circuit. Transistor VT_{11} on the left of Fig. 4, to which the output of the clipper amplifier is fed, produces such a waveform. Its collector supply voltage is regulated by a 10V voltage reference diode, MRZ_1. The output from VT_{11} is fed to the counter circuit comprised of capacitors C_5 and C_6, diodes MR_3 and MR_4, and the output meter. During positive half cycles C_5 is charged through the transistor in series with the forward-conducting MR_4 and the parallel combination of C_6 and the output meter. The charging time-constant is short compared with the periods being counted, so that in effect a fixed amount of charge is passed to C_6 during each positive half cycle. During negative half cycles C_5 discharges through the forward conducting MR_3 and the transistor load. MR_4 prevents charge leaking to the left from C_6, which discharges through the meter. Since a fixed amount of charge passes to C_6 during each cycle it follows that the output current is proportional to the frequency.

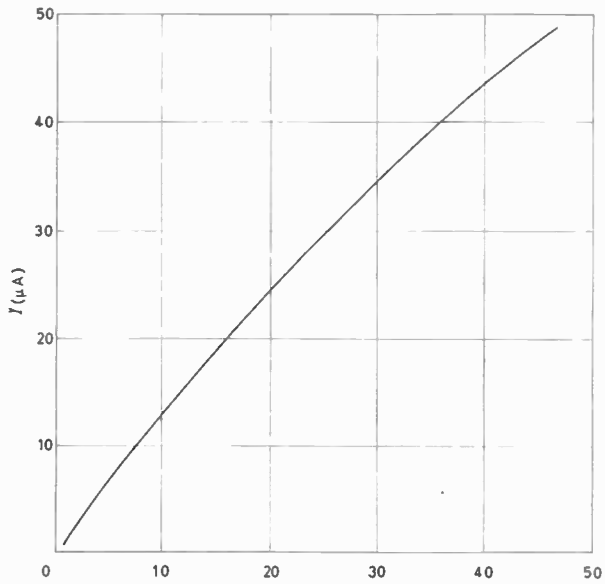

Fig. 5. Overall calibration of the instrument

A calibration of the circuit of Fig. 4 showed that the output was linearly related to input frequency over the frequency range 0 to 10kc/s. Full scale deflexion is obtained at a frequency of 2kc/s with $C_5 = 2\,500$pF and at 10kc/s with $C_5 = 500$pF. Capacitor C_4 in Fig. 2 was in fact adjusted so that frequencies up to 10kc/s could readily be generated. Although the choice of a smaller value of C_4 would have led to a better linearity between beat frequency and capacitance change in the probe (see Appendix) it was found that circuit noise caused spurious indications at very low beat frequencies. It was therefore decided to limit these spurious outputs to a very small part of the range near the origin by choosing a frequency scale with a maximum of 10kc/s. Because of the resulting non-linearity introduced between beat frequency change and probe capacitance the overall calibration of the instrument, shown in Fig. 5, is not quite linear. However, as will be seen from the discussion below, there is no particular merit in having an absolutely linear relation since the instrument must be calibrated in the field.

Laboratory Measurements

The order of capacitance change observed in the field was such that it seemed unlikely that it was solely due to the dielectric properties of the grass between the probe electrodes. Laboratory measurements were therefore made on growing grass using a simple parallel-plate type of capacitor. This is illustrated in Fig. 6. An insulated frame supported two thin brass electrodes E_1 and E_2 which were held in slots at a separation of 4¾in. Sheets of Tufnol 1/16in thick were made integral with the frame. These served to prevent direct contact between individual blades of grass and either of the electrodes. To keep capacitance to ground small the electrodes were terminated ½in above the base of the frame.

The interelectrode capacitance was measured with the frame standing on the bench with air as dielectric. A

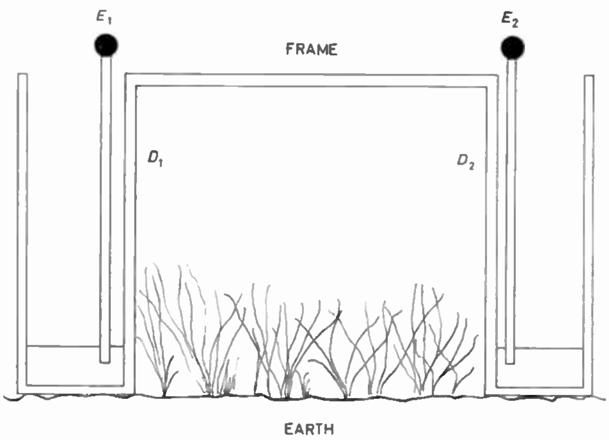

Fig. 6. Parallel-plate capacitor used for laboratory measurement

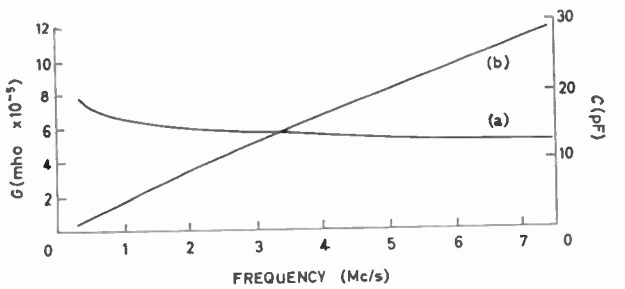

Fig. 7. Dependence of observed conductance and capacitance on frequency for a growing grass sample

Wayne Kerr transformer ratio arm bridge type B601 was used for the purpose. Neither of the electrodes was earthed and it was assumed that balanced measurements about earth were being made. The measured capacitance was almost constant over the frequency range 30kc/s to 7·5Mc/s. Its value was 9·75pF. The conductance was negligible.

A series of measurements was then made with the capacitor resting on flattened grass and it was found that there was a slight increase in the capacitance and a significant conductance was observed. The capacitance varied from 10·4pF at 30kc/s to 10·0pF at 7·5Mc/s. At 30kc/s the conductance was of the order of $2 \times 10^{-7}\Omega^{-1}$ and at 7·5Mc/s was approximately $14 \times 10^{-6}\Omega^{-1}$.

The capacitor was then located on growing grass about 5in high as shown in Fig. 6. Measurements of capacitance and conductance were made over the same frequency range. Typical variations with frequency of capacitance C and conductance G are shown as curves (a) and (b) respectively of Fig. 7. The corresponding variation of G/ω^2C is shown in Fig. 8 as the continuous curve (a). A logarithmic plot of this type is useful when evaluating the applicability of any equivalent circuit.

One possible equivalent circuit is shown in Fig. 9(a). Here C_1' is the capacitance between E_1 and the grass in contact or near-contact with D_1. C_2' is a similar capacitance on the other side. C_{g1} and R_{g1} represent the capacitance and effective shunt resistance of the grass

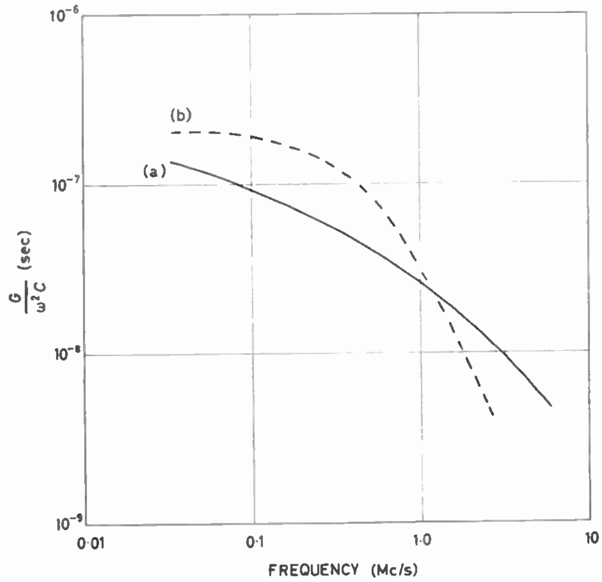

Fig. 8. Dependence of G/ω^2C on frequency
(a) Growing grass; (b) single relaxation time spectrum corresponding to Fig. 9(b)

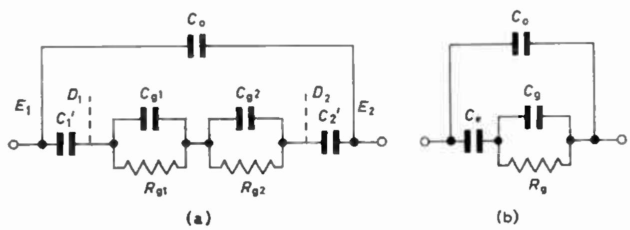

Fig. 9. S Simple equivalent circuits for growing grass
(a) Double-sided model; (b) reduced form if symmetry is assumed

between D_1 and the centre of the system; similarly C_{g2} and R_{g2} represent the grass on the other side. C_o is the total interelectrode capacitance between E_1 and E_2, which is associated with that part of the central region for which the dielectric is air. If symmetry is assumed so that $C_{g1}=C_{g2}=2C_g$, and $R_{g1}=R_{g2}=R_g/2$, and $C_1'=C_2'=2C_e$, then the equivalent circuit takes the simpler form shown in Fig. 9(b). Writing the terminal admittance as

$$Y = G + j\omega C \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

it follows that

$$G = \frac{\omega^2 C_e^2 R_g}{1 + \omega^2 (C_e + C_g)^2 R_g^2} \quad \ldots\ldots\ldots\ldots \quad (2)$$

and

$$C = C_o + \frac{C_e\{1 + \omega^2 (C_e + C_g) C_g R_g^2\}}{1 + \omega^2 (C_e + C_g)^2 R_g^2} \quad \ldots\ldots \quad (3)$$

In consequence one can write

$$G/\omega^2 C = \frac{C_e^2 R_g/(C_e + C_g)}{1 + \omega^2 (C_e + C_g)\left(C_g + \dfrac{C_o C_e}{C_o + C_e} \right) R_g^2} \quad \ldots \quad (4)$$

At low frequencies G/ω^2C is therefore almost constant and at high frequencies falls as ω^{-2}. This type of frequency-dependence, characterized by a single relaxation-time $R_g\{(C_e+C_g)[C_g+C_oC_e/(C_o+C_e)]\}^{\frac{1}{2}}$, is illustrated in Fig. 8 by the broken curve (b).

It is evident that the observed frequency-dependence of G/ω^2C cannot be represented by the simple model of Fig. 9(b), since the observed fall with frequency is less rapid than for the model. However, it is always possible to synthesize a frequency response such as that observed by considering a number of branches of the equivalent circuit, each characterized by a different relaxation time. One way in which such a model can be built up is by considering that each individual blade of grass in proximity to an electrode is represented by a three-element circuit such as that shown in Fig. 9(b). A large number of circuits of this type would effectively be in parallel on each side of the model.

As a result of laboratory measurements such as the above it has been shown that the observed capacitance change is frequency-dependent. Although a detailed equivalent circuit would be extremely complicated it appears that the grass acts as a lossy dielectric which is coupled to the electrodes via terminal capacitances which depend on the area of leaf in contact or near contact with the thin insulation on the electrodes. Support for this model derives from the fact that considerable changes in reading can be obtained for a given grass sample situated between the electrodes by deliberately adjusting the leaf area in contact with electrodes, without altering the total amount of grass involved.

Field Measurements

The instrument described has been used in the field. Before measurements are made the meter indication is adjusted to zero using the control shown in Fig. 2, with the instrument standing on bare ground. Calibration has been carried out for different sward types. The procedure is to obtain a meter reading and then correlate this with some characteristic of the volume of grass responsible for the reading. This involves cutting the patch of grass which was under the probes and evaluating it. It has been found[2] that there is a direct correlation between meter reading and the total water content of the grass (fresh weight—dry weight). Once such a calibration has been carried out for a particular sward type it is possible to use the instrument to compare the effect of different treatments on growing plots of the same type of grass.

Conclusions

An instrument has been made by means of which non-destructive investigation of the growth of grass can be made. Laboratory measurements have shown that the precise way in which growing grass creates a capacitance change in a capacitor probe is complicated. It appears to be in part due to the proximity of the grass to the electrodes and in part to dielectric effect in the grass itself. Although comparatively large capacitance changes are observed at low frequencies it seems that a choice of 1 to 3Mc/s for the standard frequency of operation has some merit, since it is then easy to produce useful difference frequencies of several kilocycles per second, which are approximately linearly related to the capacitance changes produced. It is possible that the correlation between capacitance change and water content might not be as good if a standard frequency considerably different from 3Mc/s was chosen.

Acknowledgments

The authors are indebted to Mr. M. Perry for preliminary work on the design of oscillators with good differential stability and to Mr. A. R. Owens for helpful suggestions concerning the clipping amplifier. Mr. E. Jones and Mr. G. Owen helped with the construction of the instrument, which was developed for Dr. M. B. Alcock of the Department of Agriculture, U.C.N.W. at his request.

Appendix

Relation between oscillator frequency and change in probe capacitance

A simplified version of the tuned circuit incorporating the probe is shown in Fig. 10. The probe capacitance with air as dielectric is represented by C_p and the change in probe capacitance caused by the grass by ΔC_p. The total capacitance C_T in series with L is given by

$$1/C_T = (1/C_1) + (1/C_2) + \cfrac{1}{C + \cfrac{C_4(C_p + \Delta C_p)}{C_4 + C_p + \Delta C_p}} \quad \ldots\ldots (5)$$

if the transistor impedance is neglected. Since C_1 and C_2 are in any case comparatively large they can be ignored so that

$$C_T \simeq C + \frac{C_4(C_p + \Delta C_p)}{C_4 + C_p + \Delta C_p} \quad \ldots\ldots\ldots\ldots (6)$$

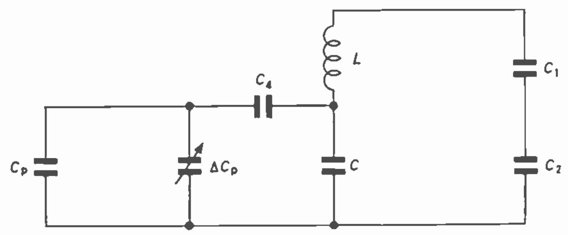

Fig. 10. Simplified equivalent circuit of the tuned circuit of the variable frequency oscillator

The oscillation frequency is given by

$$f \simeq \frac{1}{2\pi\sqrt{(LC_T)}} \quad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

Taking $f = f_o$ when $\Delta C_p = 0$, it follows that

$$(f/f_o)^2 = \frac{C(C_4+C_p) + C_4C_p}{C_4 + C_p} \cdot \frac{C_4 + C_p + \Delta C_p}{C(C_4+C_p+\Delta C_p) + C_4(C_p+\Delta C_p)} \quad \ldots\ldots\ldots\ldots (8)$$

Now $C_4 \ll C, C_p$, so that

$$(f/f_o)^2 \simeq \cfrac{C}{C + \cfrac{C_4(C_p + \Delta C_p)}{C_4 + C_p + \Delta C_p}} \quad \ldots\ldots\ldots (9)$$

After some manipulation this reduces to

$$f/f_o \simeq 1 - (C_4^2/2C)(\Delta C_p/C_p^2) \quad \ldots\ldots\ldots (10)$$

$$\text{and } \Delta f/f_o = \frac{f_o - f}{f_o} \propto \Delta C_p \quad \ldots\ldots\ldots\ldots (11)$$

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A Reliable D.C.-Amplifier for Thermocouple Vacuum Gauges

By U. Tallgren* and U. Kracht*

A simple circuit has been developed to amplify the signal from a thermocouple in a vacuum gauge. Output is sufficient for driving several series connnected 1mA instruments. The amplifier is based on a small transductor used as a second harmonic convertor followed by an a.c.-coupled amplifier. The output feeds a rectifier through a transformer coupled output stage. Complete galvanic insulation is obtained between input and output circuits. Linearity of the amplifier depends mainly on the symmetry of the transductor and can be better than ±3 per cent.

Stabilizing circuits for the heater current of two commonly used thermocouple vacuum gauge heads are given, and also circuits for switching a relay at a predetermined output level of the thermocouple amplifier.

Two types of circuits are described, one intended to work on mains supply, the other on a 48V standby battery.

The circuits in this article might also be of interest for other instrumentation problems, e.g. temperature control with thermocouples.

(Voir page 727 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 734)

THE thermocouple gauge is one of the most reliable and frequently used type of gauges in the fore vacuum range, e.g. from 1 torr down to 10^{-3} torr†. It consists essentially of a thermocouple with a heater element, both mounted in the vacuum system where the pressure is to be determined. As the pressure decreases the heat loss of the gauge system to the surrounding gas will diminish, hence the temperature of the thermocouple will increase. The thermocouple therefore gives its maximum output at high vacuum and very little at atmospheric pressure.

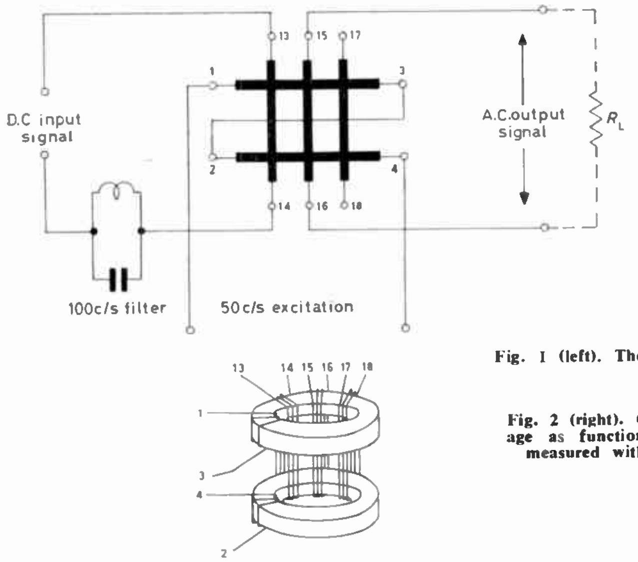


Fig. 1 (left). The d.c.-a.c. convertor

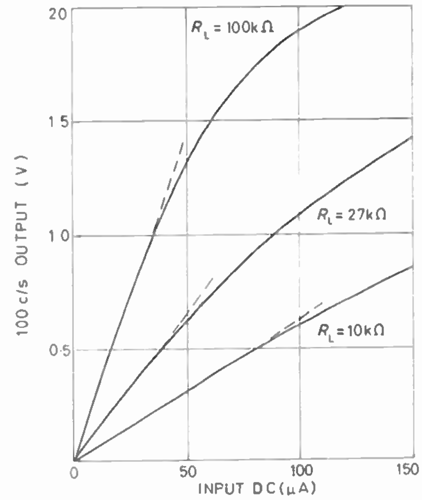


Fig. 2 (right). Convertor output voltage as function of input signal as measured with selective voltmeter

The associated circuit for a thermocouple gauge is simple. The output from the thermocouple usually feeds directly a sensitive moving-coil meter, and a variable heater supply permits one to adjust the heater power and counteract the influence of any variations in the supply voltage.

Frequently one needs to read control instruments at several locations, e.g. in the case of radiation hazards. Since the output power from a thermocouple gauge head is too small to drive several meters, it is necessary to use an amplifier between the gauge and the instruments. As automatic vacuum systems become more common, there is also need for a relay tripping circuit or 'pressure guard' that gives information when a certain pre-set pressure level is reached.

In 1960 the above-mentioned problems were considered in connexion with improving the security of the vacuum system for the linear accelerator of the CERN proton synchrotron (about 20m³ at 5.10^{-6} torr). Since no commercial units that could fulfil the requirements were available, the circuits to be described were developed. Later it became desirable to operate the whole Linac vacuum control system from the 48V standby battery, and the circuits were accordingly modified using transistors as amplifying elements instead of valves.

The aim has been to make a simple but reliable device with inherent 'fail-safe' conditions when used in automatic vacuum systems.

The D.C. to A.C. Convertor

A small transductor (Giesenhagen, Munich, Germany, type TUR 40 Y, BS 5 382, diameter 60mm, height 40mm, weight 200g) is used as a static convertor, changing the d.c. input from the thermocouple to a mainly 100c/s signal which can be further amplified without zero drift problems (Fig. 1). The advantages of this convertor are:

(a) High output voltage and convenient output frequency.

(b) No moving parts or active elements.

* *European Organization for Nuclear Research (CERN), Switzerland.*

† *1 torr = 1mm Hg.*

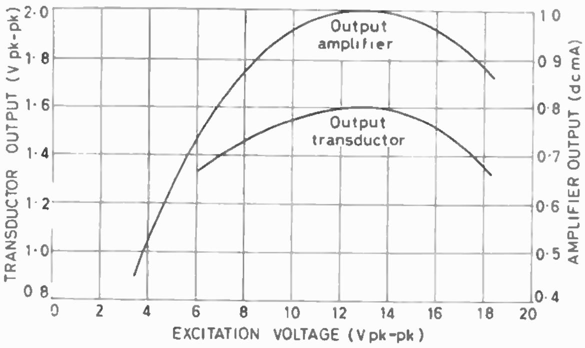


Fig. 3. Convertor and amplifier output voltage as function of excitation voltage for maximum d.c. input signal

(c) Negligible temperature dependence.
(d) Input and output insulated from each other and from earth.
(e) Linear relation over a wide range of d.c. input to a.c. output.
(f) Small size and convenient mounting.

The transductor has two cores which are excited with a 50c/s voltage. The excitation coils are so connected that the induced voltages in the control windings are cancelled. Due to imperfections in the symmetry of the system a small residual voltage will appear. The symmetry can be further disturbed if a static magnetic field is imposed on the two cores in different directions. This is done with the d.c. input signal, and as a result a 100c/s signal will appear in the control windings. Standard text-books on transductors (e.g. Geyger[1]) analyse this action in more detail. It should, however, be noted that the present use of the transductor TUR 40 Y is not foreseen by the manufacturer, but actual tests showed that it is well suited for this purpose.

Fig. 2 shows the 100c/s component appearing on the winding 15-16 of the TUR 40 Y for different resistive loading. Excitation voltage was 6·3V 50c/s. Note that one obtains rather high output voltages which are not difficult to amplify either with valves or transistors.

The output voltage from the convertor is somewhat dependent on the excitation voltage (Fig. 3) and the simplest means to stabilize this voltage was to use Zener diodes (Fig. 4). The excitation voltage is therefore now a square wave with about 12V peak-to-peak and with very little dependence on supply voltage variations. For the whole amplfier shown in Fig. 4, a variation of the mains supply of ± 10 per cent will cause a change in the output of less than ± 2 per cent.

Partly due to the excitation voltage not being sinusoidal, the a.c.-voltage from the convertor departs appreciably from a pure 100c/s signal (Fig. 5). That seems, however, not to influence the linearity of the amplifier. The linearity depends mainly on the selection of a transductor which has low residual voltage for zero input signal. The linearity over the range of interest can be better than ± 3 per cent, but greater deviations might be accepted in this particular application since the characteristic will not change with time.

Fig. 1 shows a 100c/s filter in series with the input signal. This is necessary when a low source impedance such as a thermocouple is used. Otherwise the low impedance of the thermocouple will short-circuit the signal output windings. If the thermocouple is disconnected from the amplifier a relatively high output voltage will appear from the amplifier since the input circuit is left open. In order to prevent this condition, a 1kΩ high stability resistor is connected across the input terminals.

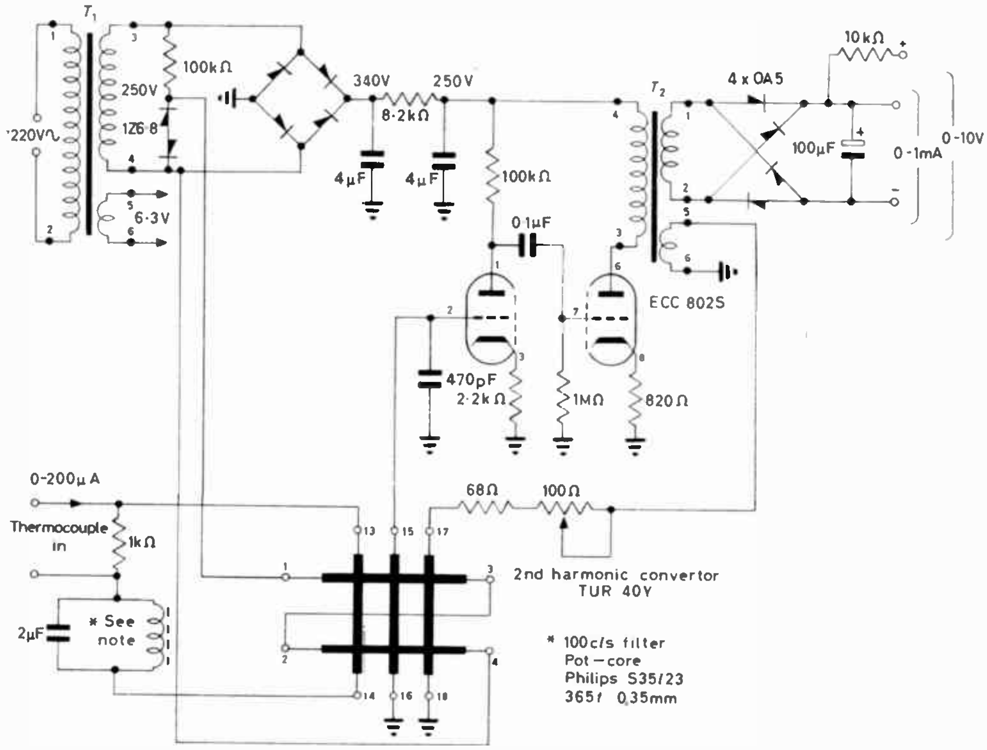
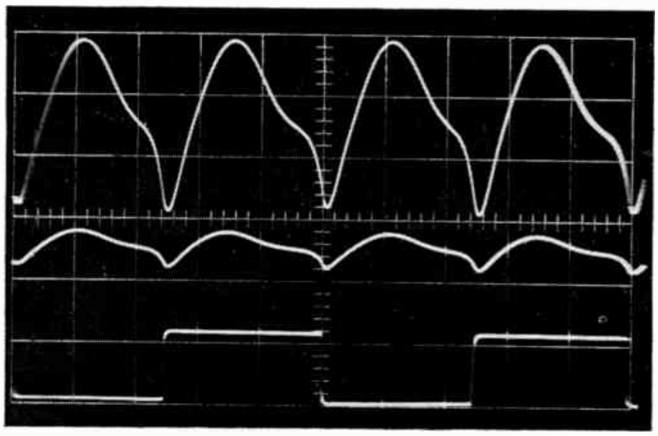


Fig. 4. Arrangement of mains operated thermocouple gauge amplifier

Fig. 5. Oscilloscope picture of convertor output signal
Upper trace: Maximum output from convertor
Centre trace: Low level output from convertor
Both 0·5V/cm
Lower trace: Excitation voltage 10V/cm
Time base: 4msec/cm



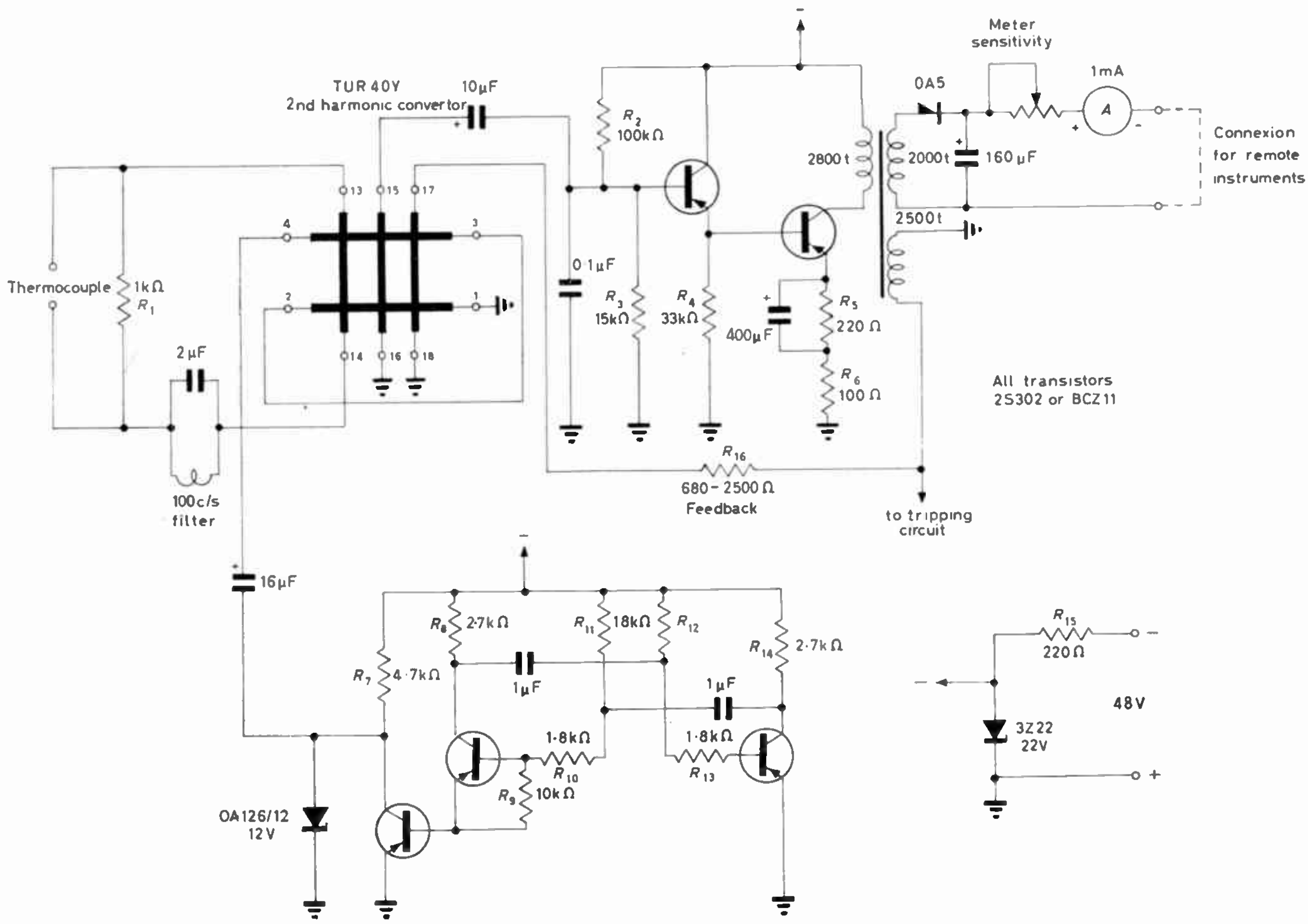


Fig. 6. Arrangement of battery operated thermocouple gauge amplifier

The Amplifier

Mains Operated Unit

The circuit diagram of the mains operated unit is shown in Fig. 4. A long life valve ECC802S is used as the amplifying element. As output transformer it was found possible to use the type chosen for the power supply. The normal 6·3V winding was useful for deriving a feedback signal which was impressed on the bias winding 17-18 of the transductor. A 100Ω potentiometer in the feedback chain gives control of the total gain. The feedback helps to reduce the residual output voltage from the amplifier at zero input.

It is normal to follow the a.c.-amplifier with a phase sensitive rectifier in order to restore the information of input signal polarity[1]. This is not required in this case, since the input polarity is always the same. The transformer coupled output stage of the amplifier therefore directly feeds a bridge of germanium diode rectifiers.

Battery Operated Unit

It seemed at the first investigation that for a transistorized version it would be both simpler and cheaper to use a balanced d.c.-amplifier[2] rather than a convertor and a.c.-amplifier. However, at the time this work was undertaken the only usable input transistor on the market was the OC200 (BCZ10). A practical test showed that due to insufficient gain and strong temperature dependence of the transistor, the overall performance of a balanced d.c.-amplifier was not completely satisfactory. With present day transistors, better results might be obtained, but the use of a convertor at the input still has the advantage of galvanic separation between input and output circuits. The adopted solution, based on the same convertor as for the mains operated unit therefore continues to be of interest. The circuit diagram of the battery operated version is given in Fig. 6.

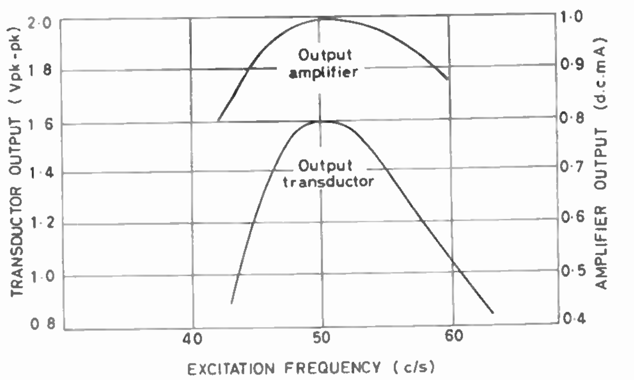


Fig. 7. Convertor and amplifier output voltages as functions of excitation frequency for maximum d.c. input signal

The oscillator needed to supply the excitation winding with 50c/s is a straightforward multivibrator whose frequency determining components are high-stability close-tolerance types to minimize frequency drift. Fig. 7 shows the influence of excitation frequency drift on convertor and amplifier output voltage. The shape of the curve depends mainly on the 100c/s input filter. The oscillator frequency changes less than 0·5c/s for supply voltages between 36 and 60V. A further transistor connected between the transductor and the multivibrator acts as a buffer between the multivibrator and the highly inductive excita-

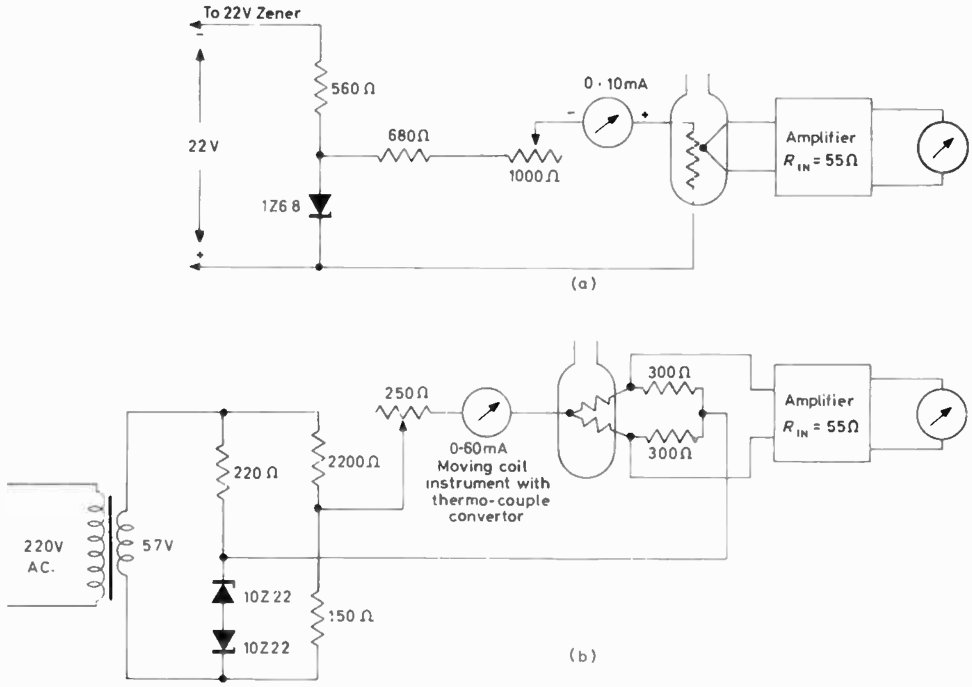


Fig. 8. Heater circuits for gauge heads

(*a*) *Best products gauge head (variation in heater current <0·3 per cent for supply voltage variations between 36 and 60V)*

(*b*) *Leybold gauge head (variation in heater current ±0·2 per cent for mains voltage variations between 190 and 250V)*

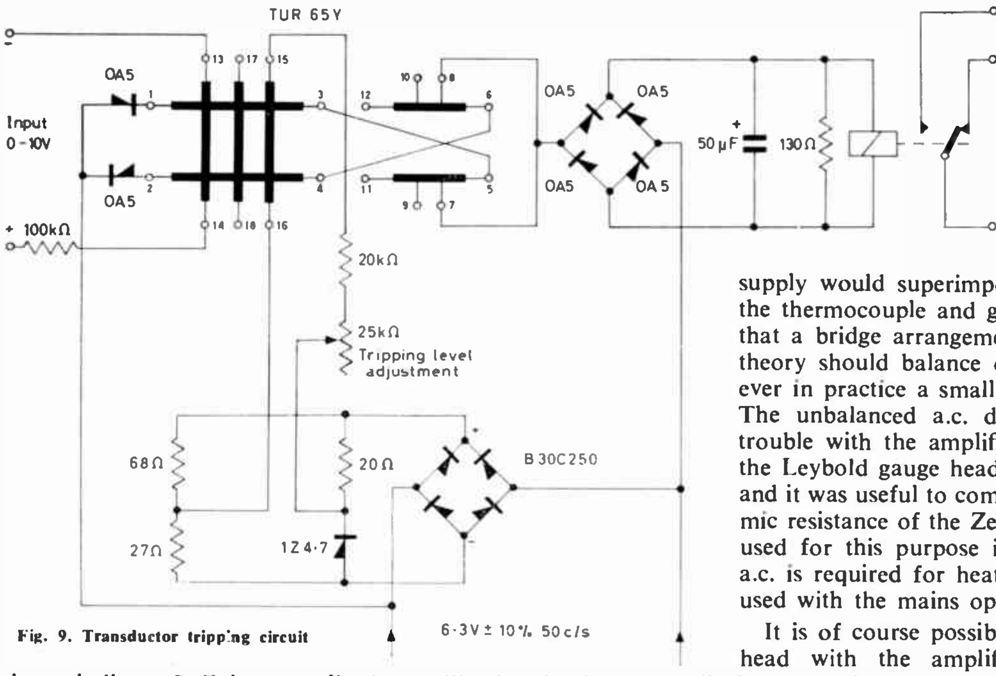


Fig. 9. Transductor tripping circuit

tion windings. Sufficient amplitude stabilization is simply obtained with a 12V Zener diode.

The a.c.-amplifier is straightforward using, as elsewhere in the battery operated device, silicon transistors to eliminate temperature dependence as far as possible. The use of silicon transistors increases the cost of the whole unit by less than 10 per cent. Internal feedback is provided by an un-bypassed part of the emitter resistor. This and additional overall feedback, through R_{16} from the output transformer secondary to the transductor, greatly improves the overall performance. A secondary winding feeds the output rectifier and another winding supplies the relay circuit and applies the feedback current. To reduce the dimensions of the transistorized version a special output transformer is used.

Disturbances from the 48V line are avoided by dropping the supply voltage to about 22V with a Zener diode. This voltage was chosen to allow for the greatest expected changes in the 48V supply and also for easy adaptation to other supply voltages (24 and 30V used elsewhere at CERN).

Heater Circuits for Thermocouple Gauge Heads

Two heater circuits for thermocouple gauge heads are given in Fig. 8. The first is for the gauge heads supplied by Best Products Ltd, Felixstowe, Suffolk. This gauge head has small dimensions (about 100mm length and maximum diameter 10mm) and low heater power requirements. Heater element and thermocouple are galvanically separated. The useful pressure range is, however, smaller than for the Leybold gauge head (Leybold, Cologne, Germany), Fig. 11, for which the second circuit is intended. The disadvantage with the Leybold head is that heater and thermocouple are not separated galvanically. It is therefore necessary to use a.c. for heating. A d.c. heater supply would superimpose itself on the signal from the thermocouple and give a false reading. It is true that a bridge arrangement is used, which at least in theory should balance out the heater current: however in practice a small unbalance is always present. The unbalanced a.c. does not seem to cause any trouble with the amplifier described. The heater for the Leybold gauge head takes an appreciable power, and it was useful to compensate for the internal dynamic resistance of the Zener diodes. The bridge circuit used for this purpose is described elsewhere[3]. Since a.c. is required for heating, this head has only been used with the mains operated amplifier.

It is of course possible to use any make of gauge head with the amplifier, but the heater circuit will have to be adopted to different heater power requirements.

Tripping Circuits

Transductor Tripping Circuit

For automatic vacuum systems it is usually required to have a relay signal when a pre-determined pressure is reached. Moving-coil relays have been used for this purpose, but they have the disadvantage of being very sensitive to mechanical vibrations.

A transductor circuit has been tried in connexion with the earlier described amplifier, with the advantage that a normal relay can be used. The circuit diagram is given in Fig. 9. The circuit behaves like a current controlled flip-flop. When the input control current exceeds a certain level, which can be controlled with a bias current on the transductor, the output from the transductor suddenly swings to its maximum value. There is, therefore, never any danger of relay chatter. Going back with the control current there is a hysteresis of about 5μA and then the output again abruptly falls to practically zero, well below the opening level for the relay. If desired the hysteresis can be made wider by using other taps on the feedback winding of the transductor.

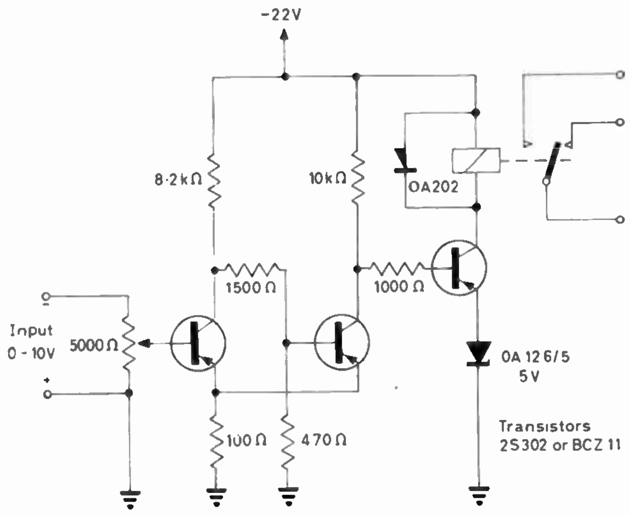


Fig. 10. Transistor tripping circuit

In order to have a quick response of the circuit, it is necessary to have a high resistance in series with the control winding. The tripping circuit is therefore connected to the gauge amplifier through a 100kΩ resistor. 10V output from the amplifier will therefore correspond to 100μA through the control winding.

The described tripping circuit has also been successfully used with Penning gauges. Several tripping circuits can be used in series if several trip-points are required. It is preferable to have a well filtered supply of the Penning gauge, and to take precautions for avoiding circulating a.c. in the control windings, as otherwise the hysteresis at the tripping point will be very broad.

Tripping Circuit for the Battery Operated Version

Since the tripping circuit for the mains version uses a rather large transductor which also requires a.c.-excitation it was thought interesting to use a circuit with transistors for the battery operated version. The circuit is given in Fig. 10.

The tripping circuit is a Schmitt-trigger followed by a relay drive stage. The circuit can be adjusted to trip the relay from about 0·65V (which corresponds to about 35 per cent of maximum amplifier output) across the input potentiometer. This corresponds to about $3{\cdot}5 \,.\, 10^{-1}$ torr when used with the transistorized amplifier. The relay should, however, not release at the same pressure as it operates in order to avoid relay chattering, rather, it should release at a higher pressure. This hysteresis is about 50mV and is fixed by the components used in the circuit.

Mechanical Design

In automatic vacuum systems there are usually required at least three vacuum gauges. In order to make a flexible system an amplifier plus tripping circuit has been constructed as a compact plug-in unit using the CERN standard plug-in chassis system[4]. For the transistorized version there is space for up to six units in one chassis, and for the mains operated version five, since one unit is taken up by a common power supply. Due to the plug-in system, a faulty unit can quickly be replaced by a spare. The mains operated version takes up a front panel height of four standard units (177mm) and the transistorized battery operated version, being more compact, three standard units (132·5mm).

Adjustments

The adjustment of the units is very simple, but somewhat different for the two versions. The heater currents of the gauges are set so they all give the same output in high vacuum ($< 10^{-3}$ torr), which is checked with a high-impedance millivoltmeter across the input of the amplifier. For the mains operated version, the feedback potentiometers in the amplifiers are then adjusted for full scale reading on the output instrument at high vacuum. For the battery operated version, the fixed feedback resistor R_{16} is selected for 2·0V across the tripping level potentiometer measured with a 20kΩ/V instrument, and afterwards the potentiometer in series with the output instrument is adjusted for full scale reading, also at high vacuum. Finally the trip level is set at the desired pressure with the help of a calibration curve (Fig. 11). Before the adjustments of the battery operated version the excitation oscillator frequency is adjusted to correct value (50c/s) by altering R_{11} and R_{12}.

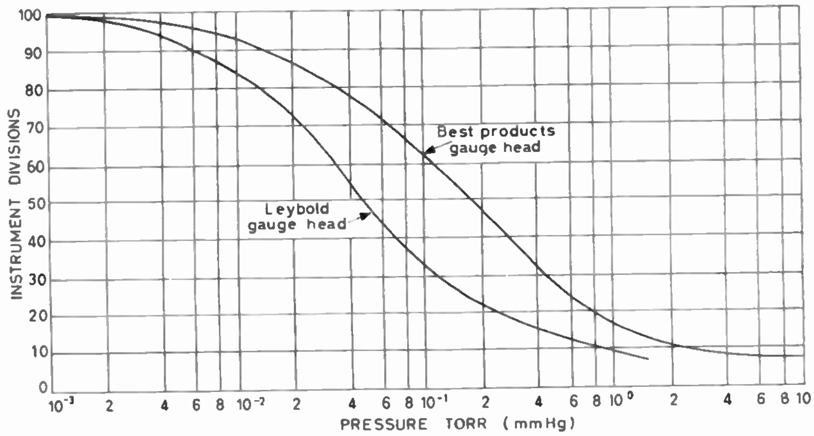


Fig. 11. Typical calibration curves for thermocouple gauge heads

Performance

More than twenty units have been made of each version and most of them have now made more than 15 000 hours of operation. So far no failures have been reported.

Acknowledgements

The authors wish to thank CERN for permission to publish this article and Mr. C. S. Taylor, head of the Linac Group, for his stimulating interest in this work. Gratitude is also expressed to Messrs. J. P. Buathier, A. Kubler, F. Malthouse and A. van der Schueren for their help.

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Some Aspects of SECAM Colour Television Receiver Design

By P. Cassagne* and G. Melchior*

A prototype commercial SECAM colour television receiver has been developed and in this article some of the more important design features are discussed. Particular attention is paid to the receiver controls and in ensuring that the receiver is simple to operate. It is concluded that a satisfactory SECAM receiver can be produced that compares favourably in cost with a comparable NTSC receiver.

(Voir page 727 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 734)

HAVING produced several types of experimental receivers (RS 08, RS 10, RS 11 and RS 41) designed to demonstrate the qualities of the SECAM system, the Compagnie Française de Television has developed a prototype domestic receiver; it is the RS 15. It will be the basis for the construction of some hundred units. In this way, the transfer to mass production as well as dealers' network education and servicing will be started and progressively developed.

Basic Principle

The SECAM system possesses certain novel characteristics as compared to the NTSC system, mainly sequential transmission of chromatic information and use of frequency modulation. For the receivers this results in special arrangements which will be described.

First a brief description of the coding principle will be given, and the main functions to be provided for when receiving colour pictures.

Colour television is based on the joint transmission of three signals, green, red and blue, from which a complete colour picture can be reconstituted.

As it is not desired that black-and-white programme be adversely affected by colour television (problems of compatibility), first the three primaries, green, red and blue are suitably mixed to form a normal black-and-white signal, designated luminance signal *Y*.

Two other signals, chrominance signals, are transmitted alongside the luminance signal *Y*, viz: *R-Y* and *B-Y*.

Black-and-white receivers are sensitive to the luminance signal only, whereas colour receivers are sensitive both to the luminance signal and to the two chrominance signals; from these three signals they can reconstitute—through the decoding unit—the three primaries, green, red, blue and display a colour picture on the screen of the three-colour tube.

The novelty of the SECAM process results from the fact that the *R-Y* and *B-Y* signals are not transmitted simultaneously, this being to ensure that they shall not unduly mix, so that colour rendering will remain faithful.

Since the chrominance signals are not simultaneous, ruggedness of the system as regards transmission vagaries, is secured by the use of frequency modulation (which would not be possible if *R-Y* and *B-Y* were transmitted simultaneously).

So the function of the decoding unit is:

(1) To make the *R-Y*, *B-Y* sequential signals simultaneous, since the primaries, green, red and blue, can be determined only if the luminance *Y* and the chrominance signals are simultaneously available.

(2) To demodulate the chrominance signals, the *Y* signal being demodulated by the usual process employed in black-and-white television.

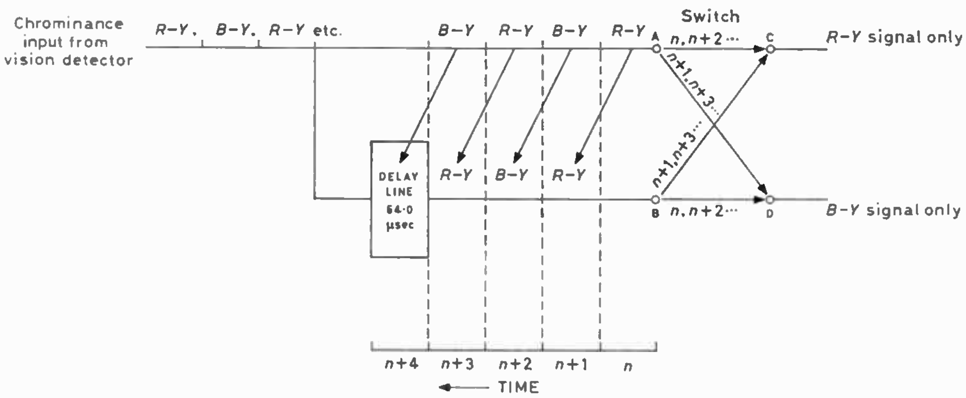

Fig. 1. Sequential to simultaneous transformation

To perform this dual function, the decoding unit contains essentially a ' memory ' and frequency discriminators.

Experience has shown that during the analysis of a television picture the eye is incapable of discerning if *R-Y* or *B-Y* vary from one line to the next. Thus the same *R-Y* or the same *B-Y* can be used for two successive lines.

Fig. 1 shows how, from the signals transmitted sequentially one line after another, the memory, consisting of a delay line, can reconstitute simultaneous signals.

On account of this repetition of the transmitted signals, colours are naturally spread in the vertical direction. From the large number of experiments made with a view to examining this effect, such an approximation is found to be perfectly valid. The resulting error is visible only in very special cases, and very infrequently in natural scenes.

The double switch which passes the direct and delayed chromatic signals towards the output *R-Y* and *B-Y* is operated at line frequency by a bistable flip-flop, not shown in the figure.

The second demodulation function is carried out on

* *Compagnie Française de Television.*

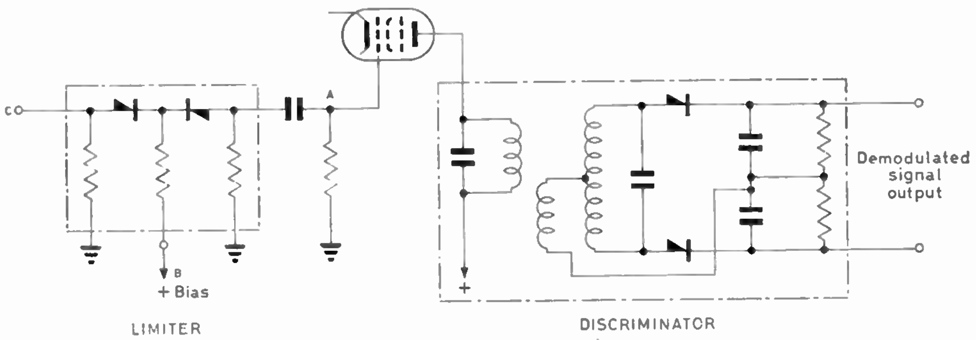

Fig. 2. Demodulation circuit of each chrominance channel

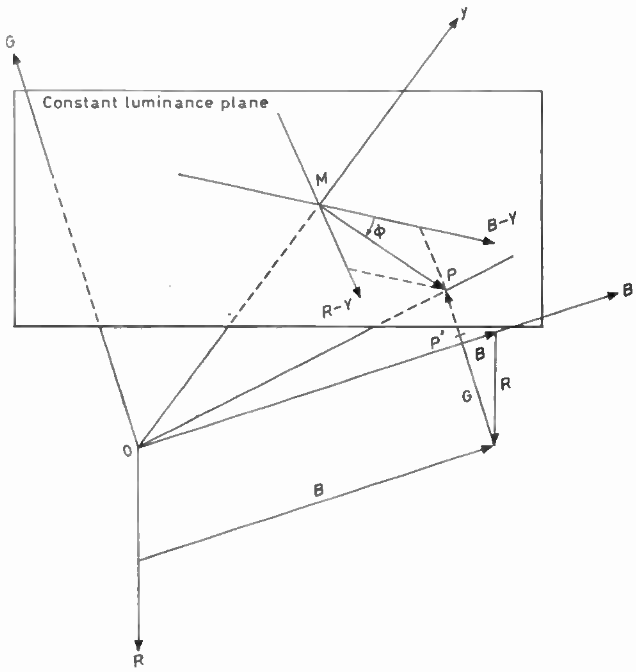

Fig. 3. Vectorial display of the colour space

each channel, R-Y and B-Y, by a limiter discriminator, the conventional circuit of which is shown in Fig. 2. As is usual for frequency modulation, this arrangement ensures a highly constant level for the demodulated output signal, and it is independent over wide limits of the level of the colour sub-carrier applied to point C. The sub-carrier level at point A is constant and can be modified by varying only the diode bias voltage (point B). This feature introduces an essential difference as compared to amplitude modulation systems and provides a stability and automaticity factor in reception of colour signals which deserves being examined in greater detail.

Receiver Controls

In colour television, picture reconstitution from three independent signals demands very accurate control of the transmission equivalent of the three channels which deliver each information. In the general case brightness, saturation and shade errors could result from incorrect adjustment of the gain of any one channel. Referring to Fig. 3, which shows the vector decomposition of the light P into its primary components, an error on any one of the primaries, G, for example, will shift the point P to be reproduced to P'. This produces a variation of Y, R-Y, B-Y, i.e. of luminance and of chromaticity. The use of the principle of constant luminance and of a single sub-carrier for chrominance, i.e. of the transmission primaries Y, R-Y and B-Y, very substantially simplifies the difficulty while not completely solving the problem; for in this case correct picture reproduction involves the assumption of adjustment of the gain of the luminance channel, as in black-and-white, but also the adjustment of the chrominance channel, in order to secure the suitable 'Y/chroma' ratio which defines saturation.

For, still from Fig. 3, if one transmits $Y = OM$, on the one hand, and the amplitude of the vector MP, on the other, as in the NTSC case, the error of gain on channel Y will cause a brightness error with corresponding variation of saturation, and an error on the gain of the chrominance channel will produce an error of saturation only. There will even be no saturation error if the luminance to chromaticity ratio OM/MP is kept correct. And shade, characterized by the angle ϕ of vector MC with a reference, in this case B-Y, will be unaffected by gain errors, provided the angle ϕ is retained and that the vector MP is not broken down to its two components R-Y and B-Y. The advantage of the 'constant luminance' principle is in this case the introduction of two degrees of freedom which are easily differentiated by observers well practiced in notions of brightness and of saturation. In addition, these two variables individually possess a wide range of acceptable adjustments for a normal viewer, faithfulness of reproduction demanding only high precision in the proportionality relation.

At least two parameters normally play a part in establishing correct receiver operating conditions; the distance between the reception point and the transmitter which defines the signal level available at the aerial terminals, on the one hand, and the general illumination of the room in which the receiver is situated and which will have to be taken into consideration when adjusting contrast.

The various conditions met with in practice regarding these two parameters demand a range of substantially different adjustment variations, often 50dB for the former, and a few decibels only for the latter. So different facilities will generally be required for controlliing the corresponding gains.

I.F. amplifier automatic gain control already universally introduced in black-and-white receivers, will also be used in colour reception in order to supply a constant detected video signal over the range of the usual aerial levels. The law of variation of the detected signal as a function of input level will be, for example, as shown in Fig. 4 which is characteristic of the circuit used in the RS 15.

From a certain level, about −70dB in this case, referred to 1mW, the detected level rises very slowly with the input signal, and since its variation does not exceed 2 to 3dB, the luminance gain adjustment does not need to be altered. As regards the amplification of the chromatic signals which modulate the colour sub-carrier situated in the higher frequencies of the luminance spectrum, a distinction has to be made as to the type of modulation. If amplitude modulation is used, the sub-carrier level available at the video detector and the gain of the following amplifiers will both play a part in defining the Y/chroma ratio of the reproduced picture.

Although a nominal transmission level is defined for the colour sub-carrier, a number of factors arising from transmission conditions between the production point and the transmitter, or from propagation conditions between the transmitter and the receiver (or even within the receiver) may affect that level, so making the sub-carrier amplitude at the receiver detector uncertain. So it will be necessary to provide a special, and generally independent, adjustment for the gain of the chrominance channels for use by the viewer to take these various conditions into account. This is the saturation adjustment.

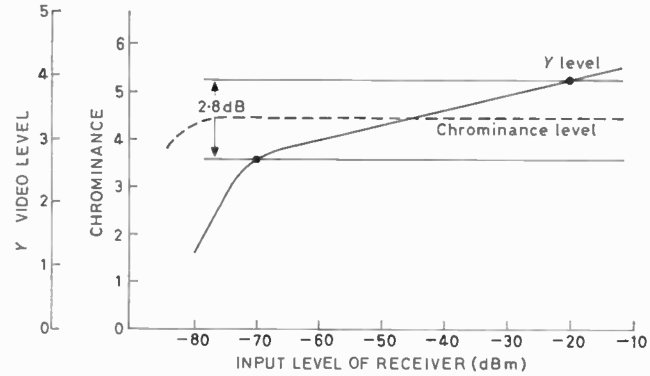

Fig. 4. Typical level characteristics of luminance and chrominance

−70dBm and for all transmission or reception conditions. For in this range of variation the chrominance signal is certain to retain a perfectly constant amplitude while the luminance signal varies, in the case of Fig. 3, by ±1·4dB. This variation of apparent saturation of the picture will generally not be detectable, it being considered that variations of ±2·5dB are normally acceptable[1].

And so the gain of the chrominance channels has to be established once for all, independently of the receiver input level and without having to provide a special saturation adjustment, or at least without its being available to the viewer.

Contrast Adjustment

Depending on viewing conditions and on the viewer's personal taste, contrast adjustment requires a level variation of some 8dB. This variation must of course be applied simultaneously to the luminance and chrominance components to ensure the proper scale alteration to the vectors of Fig. 3 without altering the established ratios. This is done automatically if the signals are transmitted in amplitude modulation, and the gain adjustment system will be able, as in black-and-white, to modify either the i.f. amplifier setting or that of the video amplifier ahead of the separation of the luminance and chrominance channels. In frequency modulation, colour sub-carrier variations are not normally transposed after demodulation, so that different arrangements will be required. Several solutions are applicable, but only two methods

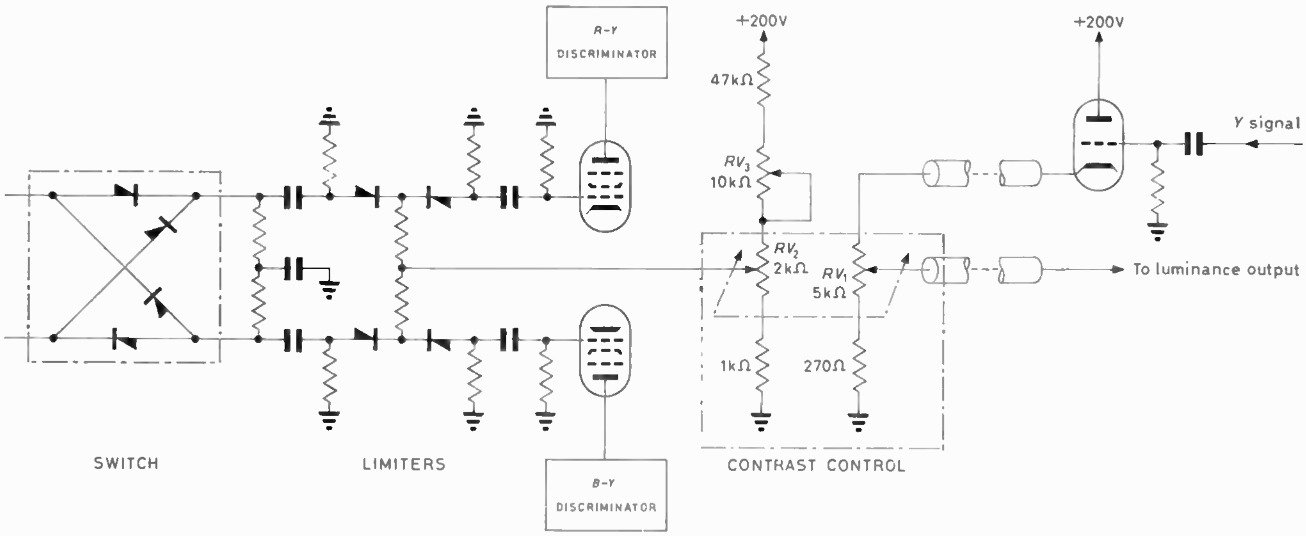

Fig. 5. Arrangement of contrast control

The use of frequency modulation for transmitting chrominance, on the other hand, provides considerable independence regarding these saturation variation factors. The receiver is capable of coping with large deviations (±6dB) in the level of the colour sub-carrier, through its amplitude limiters inserted ahead of the discriminators. The latter will deliver demodulated signals of constant amplitude, independent of the received level (dotted curve of Fig. 4) or of the variations mentioned above. So all that will be required is to set, once for all, the receiver nominal detection level, e.g. for reference white, 3·3V in the case of Fig. 4, corresponding to an input level of −45dBm, and to arrange the gains of the luminance channel and of the video circuits to supply the levels appropriate to the reproducer tube. Once established, this ratio is valid for all reception levels, from −20 to regularly used will be described.

The first, already described in a previous article[2] uses two potentiometers coupled together and acting, one on luminance and the other on the chrominance limiter bias. The method of applying this solution in the RS 15 is shown in Fig. 5 showing the limiters of the two channels, *R-Y* and *B-Y*, as well as the discriminator stages, on the one hand, and the luminance channel cathode-follower on the other, for providing remote control of the contrast control consisting of the two coupled potentiometers RV_1 and RV_2 situated on the front of the receiver. Potentiometer RV_3 is provided for works adjustment of the luminance-chrominance ratio by adjusting the mean limiter bias at the appropriate value to furnish the saturation corresponding to the gain of the luminance amplifier.

It should be noted that the sub-carrier can be extracted to feed the chrominance channel before or after the potentiometer RV_1, the only difference being a variation of the amount of limitation as a function of the contrast adjustment in the second case. It will need to be taken into account when setting the gain of the sub-carrier channel before limiting, so as not to lower is efficacy to too great an extent.

Fig. 6 shows the correlative curves of variation of the luminance and chrominance signals obtained from the circuit in question, against the position of contrast adjustment. The law of proportionality is maintained over a range of 8dB with a precision of about 10 per cent.

The second, and more direct, method is based on the use of the receiver i.f. amplifier gain control voltage. At least in the case of arrangements independent of picture

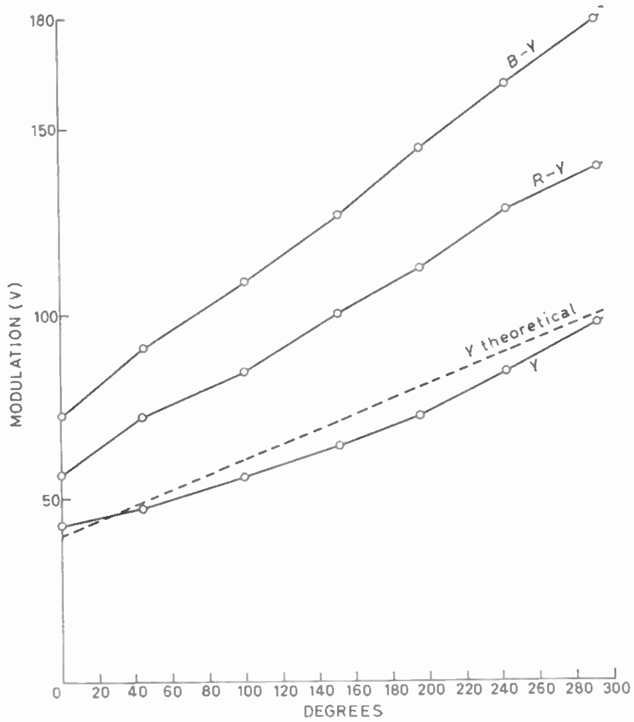

Fig. 6. Variation of luminance and chrominance signals

content, i.e. measuring the amplitude of the vision carrier during the line blanking interval, this voltage is proportional to the amplitude of the luminance signal, hence of picture contrast. In addition to its normal use as gain control for the i.f. amplifier, it is possible to deduce from it the value of the bias on the chrominance channel limiters, which sets the corresponding gain. Under these conditions contrast adjustment is provided by a single potentiometer which determines the d.c. voltage superimposed on the gain control voltage. The law of variation as a function of this voltage is very close to a proportionality law, both for the i.f. amplifier and for the limiters. So in order to 'line up' the luminance and chrominance gains all that is required is to apply this voltage in the ratio of the proportionality coefficients.

The other receiver controls are similar to those in black-and-white receivers and are designed in the conventional way. Brightness is adjusted by varying the reference voltage of the d.c. component restitution on the luminance output grid. Sound volume is adjusted by a potentiometer used to vary the level applied to the a.f. amplifier, and tuning is effected by varying the u.h.f. circuit capacitances. On this point, there is a substantial difference between the SECAM and NTSC receivers as regards mistuning. When the vision carrier is displaced along the Nyquist edge, the apparent depth of modulation of the higher frequencies transmitted on single sideband, hence the level of the colour sub-carrier available at the vision detector output varies. On amplitude modulation, the primary effect is a change in the luminance-chrominance ratio and can be interpreted as a maladjustment of the saturation control.

With the slope of the Nyquist edge of 23 to 34 per cent/Mc/s used in standards *I* and *L*, a mistuning of 500kc/s causes a variation of saturation of about 3dB, and this may even be intensified if the colour sub-carrier is situated in a non-uniform part of the i.f. amplifier response curve. This means making an adjustment by trial and error or aplying the more costly solutions of automatic tuning.

This interaction between the two adjustments completely disappears if the colour sub-carrier is frequency modulated, so that the effect of mistuning is merely a loss of

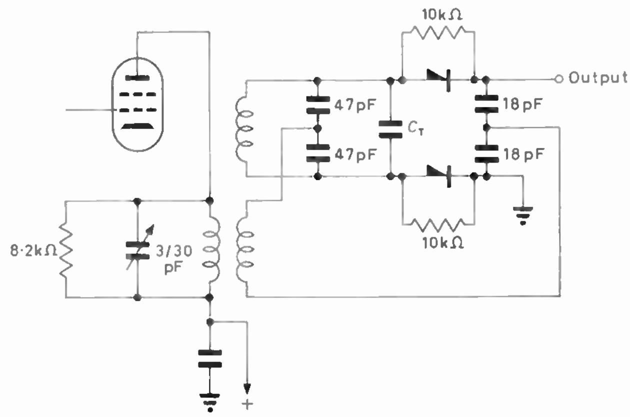

Fig. 7. Phase shift discriminator

resolution or, in the extreme, the presence of sound in the picture, as for black-and-white.

This completes the examination of the controls available to the viewer, and it will have been noted that no provision is made for shade adjustment in consequence of the use of the sequential principle. The two chromatic signals, which make use of the same channel in turn, are subjected to the same treatment throughout transmission up to the discriminator in the receiver. In this way their relative amplitude is maintained unaltered, and frequency modulation ensures that their absolute amplitude never changes. So shade distortion can be caused only by the demodulators and the following amplifiers, and the conditions required for fidelity will now be examined.

Colour Fidelity

Three parameters are mainly involved in the colorimetric quality of reproduction:

(1) The ratio of the video gains of the different channels.

(2) Linearity.

(3) D.C. component stability.

The output video stages which provide amplification and the matrix operation of the modulating signals on the picture tube raise very similar problems in the various systems. But the frequency demodulators in the SECAM system introduce differences with respect to synchronous detection; the results obtained will now be given.

(a) *Sensitivity*

A voltage gain of 25 can be obtained from the chrominance video amplifier. In view of the required levels, the voltage will have to be at least 5V at the discriminator output, so requiring a sensitivity of about 11V/Mc/s. This sets the sub-carrier level necessary at the limiter output, which is easily controlled by adjustment of the d.c. bias voltage.

(b) *Linearity*

In order to retain chromatic picture definition, the pass-band defined by the linear part of the discriminator must be not less than ±500kc/s. In the part of the spectrum corresponding to frequency deviation, ±230kc/s linearity can be achieved very satisfactorily.

(c) *D.C. component*

The d.c. component has to be faithfully demodulated and has to be equal to the voltage (generally zero) delivered by the undriven discriminator. In other words the d.c. voltage required, at a frequency of 4·43Mc/s and for normal sub-carrier level supplied by the limiters, has to be the same as when the sub-carrier is suppressed. That is the solution which was chosen for the RS 15 receiver[2] to avoid having to reconstitute the d.c. component.

(d) *Stability*

Subjective experience was made use of to ascertain the acceptable errors in respect of the chrominance d.c. components. This stability is conveniently determined by measuring on a test equipment the drift to be supplied to the sub-carrier frequency generated to produce a given amount of deterioration.

Measurements of this kind undertaken by the U.E.R. group concerned with colour television have led to the conclusion that stability of ±14kc/s was necessary[3].

Prior tests[2] had shown that average drifts of 600c/s per degree were frequently met with in circuits used up to that time. This first result ensured a satisfactory working range of 20°C only. In addition it was necessary to examine in greater detail circuit behaviour under humidity conditions and ageing.

Fuller examination has given a better definition of the above parameters and has provided much better results.

Stability of Discriminator Tuning

The frequency at which the output voltage is zero is determined by the tuning of the secondary circuit in a phase-shift discriminator, or by a given ratio of the mistuning between two circuits in the case of a stagger-tuned discriminator. So there is no *a priori* preferable type of discriminator, since the d.c. component is in both cases determined by the tuning frequency of oscillation circuits.

The tests finally led to the use of phase-shift discriminators (Fig. 7), on account of their practical advantages: smaller dimensions, less costly design, easier adjustments.

The main causes of instability that have to be considered are the ambient conditions (temperature, humidity) and time (ageing), since available components permit the design of stable windings and circuits with respect to the mechanical stresses normally applied to a domestic receiver.

Accelerated ageing obtained by placing the windings for some twelve hours in an oven held at 80°C is generally sufficient treatment to stabilize them, providing no drying out or slow polymerization varnish is used, and on this point it is preferable to restrict the varnish deposit to the points absolutely essential for securing the wires.

Humidity affects the tuning frequency, mainly on account of stray capacitance in the windings. The former used should be of non-hygroscopic material and, contrary to the technique regularly used in the construction of phase-shift discriminators, two-wire conductors are not to be used for the secondary windings; for, on account of the high sensitivity and of the wide bandwidth required, the values of the inductances are necessarily high, and the stray capacitance of a two-wire winding would on its own account for 50 per cent of the total secondary tuning capacitance. Perfect impregnation, necessarily costly, would then be essential to avoid unacceptable effects of humidity on the tuning frequency.

Increases of temperature expands the winding and varies the magnetic circuit permeability. Too high a stray capacitance between the secondary winding and the magnetic core has to be avoided, since capacitance variations with temperature are not properly reversible and are subject to hysteresis.

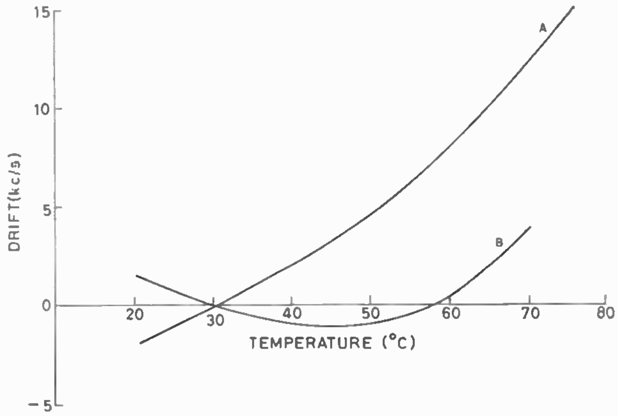

Fig. 8. Drift with temperature

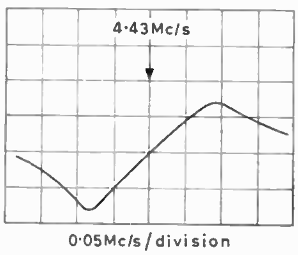

Fig. 9. Amplitude-frequency characteristic

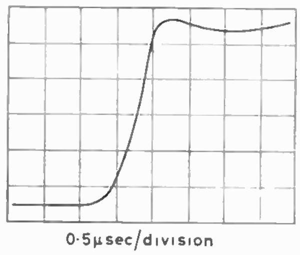

Fig. 10. Transient response

Unthreaded plastic formers are available, these being intended to take a magnetic core which cuts it own thread on its first introduction in the former. These formers are particularly useful on account of the air space left between the core and the winding.

With these precautions it is possible to obtain a discriminator with a positive temperature coefficient, reversible and constant with time; it is then easy, if required, to correct for drift due to temperature by the use of a negative capacitance coefficient.

Curve A of Fig. 8 shows drift against temperature measured in the absence of a correcting element. This curve is an average characteristic, probable dispersion about this average being about ±1kc/s at 60°C (assuming the discriminators are adjusted to ensure that the zero is correct at 30°C). So in spite of the absence of any correction, the required stability conditions are met.

Curve B shows the effect of correction consisting of a small negative temperature coefficient capacitor C_T connected in shunt across the secondary tuning capacitors. The probable dispersion widens out on account of capacitance dispersions, and is about $\pm 2{\cdot}5$kc/s at 60°C.

The temperature correction capacitor C_T (Fig. 7) is of 2·7pF, and the temperature coefficient is -2200×10^{-6}/°C.

This discriminator is simply and quickly adjusted: the single ferrite core, which tunes the secondary and also serves to effect the magnetic coupling of the windings, enables the zero to be very easily set to the frequency of 4·43Mc/s; the primary tuning capacitor is then adjusted to ensure best linearity of the characteristic; this second adjustment has no effect on the first.

Performance

Temperature stability has already been given (Fig. 8). A sensitivity of 12V/Mc/s is obtained by driving the discriminator by the pentode section of an ECF200, and with a level of 0·4V r.m.s. on the grid.

Fig. 9 shows the amplitude/frequency characteristic, and Fig. 10 the transient response for a sub-carrier frequency jump of 60kc/s.

Conclusion

The SECAM receiver RS 15 marks a new step in the evolution towards simplified construction for mass production.

Its cost, the main factor on the economic side, compares very favourably with that of NTSC receivers without this result being secured at the expense of quality or ease of operation. The simple solutions applied, the use of low-cost and perfectly satisfactory steel delay lines, should ensure for the future reliable and uniform mass production.

Acknowledgements

The authors wish to give their sincere thanks to M. Salun and M. Czekosky for their assistance in making measurements and obtaining the statistical results from which the curves of Fig. 4, 6 and 8 were obtained.

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A Photographic Positive-Negative Viewer

An entirely new method for rapid evaluation and interpretation of photographic negatives has been developed by The Marconi Co. Ltd. Based on the use of a high definition television system, the Marconi Photographic Positive-Negative Viewer enables an untrained observer to pick out details in aerial survey or reconnaissance photographs, without having to wait for positive prints.

Pictures are displayed on television monitors, and the same photograph can therefore be shown simultaneously to a large number of people. The equipment is particularly suitable as an aid to pilot briefing, where processing delays can be cut to a minimum.

The operator can instantly produce the effect of a wide variety of printing exposure times, and can magnify selected areas of the photograph up to twenty times. Controls are provided to correct for over or under exposure and other anomalies in the original photograph, while an 'edge enhancement' control can instantly highlight details which might take days to find using normal photographic techniques.

Designed to a stringent military specification, the equipment meets all the relevant specifications for arduous use on land or at sea.

The equipment will accept both negative and positive transparencies, either of which may then be shown in positive or negative form on the television screen.

A feature of this viewer is the facility which provides the effect of complex masking techniques in printing. This enables an operator to select any 20 per cent of the grey scale in the television picture information range black and white, and to effectively expand this to constitute 60 per cent of the total grey scale, the rest of the picture information being compressed to fill the remaining 40 per cent of the grey scale. This 'contrast expansion' technique makes it possible to examine a dark area of the photograph under conditions of under development, or a light area with the effect of a shorter printing exposure. The whole range of possible exposures can be swept through instantly, and the maximum detail extracted from any part of the photograph without going through the lengthy process of developing trial prints.

Another control on the equipment gives the effect of improving the resolution. This is achieved by an 'edge enhancement' process. At any edge, or point of transition between black and white, the blackness of the black, and the whiteness of the white are increased, giving a sharper appearance to the picture. This effect is particularly important at transitions between different degrees of grey, and edge enhancement can help a photographic interpreter to find detail in any part of a photograph.

Variations in lighting between different areas of the picture can be compensated for by a non-linear amplification of the signal, which is adjustable to cover any part of the picture which is under or over exposed, or affected by flaring. Five fixed values of 'gamma', between 0·4 and 0·7, are provided.

Film sizes up to 9½in square can be accommodated by the rapid wind-through mechanism, and a two-dimensional positioning system allows the original to be centred accurately on any part of the film, to permit closer examination by magnification at that particular point. The film is held accurately in the focal plane of the optical system by an electro-mechanically controlled pressure plate. Controls make certain that this plate is lifted before any movement of the film takes place, ensuring complete protection for the film at all times.

A flying spot scanner is used as the television scanning source. The cathode-ray tube source is mounted horizontally in the bottom of the cabinet, with a 45° mirror to direct the light vertically upwards through a lens system, and then through the film under examination. The light then passes through a Fresnel lens and on to the photocathode of a photomultiplier. The output from this photomultiplier is fed directly to the head amplifier of the television system and the signal processing circuitry. The final output of the system is a standard 625-line television waveform, which is fed along coaxial cable to the displays.

The size of the television picture raster is 9 inches square, and this raster is aligned down the electrical centre of the cathode ray tube to ensure that the beam deflexion and the resultant distortion is kept to a minimum. The faceplate of the 17in display tube is masked off to leave a 10in square in the centre, to ensure that the complete picture will appear, despite variations in the electrical centre of the tube.

The lens system incorporates three stages of optical magnification, up to a total of 10:1. In addition, there is an optical electronic magnification of 2:1 which can be provided by switching the amplitude of the flying spot scanning.

The resolution of the television system is equivalent to about 800 lines, and this high standard is maintained even in the corners of the picture.

Both the viewing unit, and the displays have been designed to operate in the most rugged environments. All doors and access panels are dust tight, and air is exhausted from the units by centrifugal blowers. Fresh air for cooling purposes is drawn in through replaceable air filters and interlocks ensure that the equipment is switched off in the event of one of the fans failing.

The Measurement of Permittivity and Conductivity at Temperatures up to 500°C

By R. H. A. Miles*, M.A., B.Sc., A.Inst.P.

Apparatus has been developed to give a semi-quantitative measure of the variations in permittivity and conductivity of materials over the temperature range 20° to 500°C. With some materials, such as the ferroelectric ceramics, large changes in these parameters occur, and it is necessary to employ an automatic system of range changing in order to record the results on a conventional recording meter to the approximate 10 per cent accuracy required. The total range of capacitance covered by the instrument is from 1pF to 1 000pF (at frequencies of 100kc/s and 1Mc/s), while the range of d.c. resistance recorded is 10Ω to 500MΩ. For samples in which the dielectric loss is appreciable the capacitance reading obtained is a function of both the capacitance and the loss resistance. Although it is not possible, in this case, to deduce the absolute values of the parameters, sufficient information is obtained to ascertain whether or not a more refined method of measurement is worthwhile.

(Voir page 727 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 734)

An instrument has been developed to meet the need for a comparatively simple and inexpensive method for rapidly surveying the dielectric behaviour of certain classes of new materials. The requirement was to investigate the variation of permittivity with temperature, and hence detect anomalies which might indicate ferroelectric or other phase transitions. The materials, typically mixed oxides, were often partially conducting due to oxygen deficiencies caused by imperfect preparation techniques. Measurement of permittivity, or in practice capacitance, requires an a.c. bridge technique to balance out the resistive component. It was known that automatic self-balancing bridges were available, but at high cost. An alternative approach, which forms the basis of the instrument described in this article, is to measure admittance at two different frequencies and measure the resistance under d.c. This allows most of the information to be obtained that an a.c. bridge would give, and in cases where anomalies were detected, then it was intended that the specimen should be remeasured with a manually balanced conventional bridge.

The apparatus automatically records capacitance and resistance with temperature, over the range 20°C to 500°C. Since wide variations of these parameters with temperature were possible, the apparatus has been designed to cover the ranges 1 to 1 000pF for pure capacitance and 10Ω to 500MΩ for d.c. resistance. Measurements are made at frequencies of 100kc/s and 1Mc/s, and for a pure capacitance give a direct reading, with an accuracy of better than ±10 per cent. Larger capacitances can also be measured but with reduced accuracy. For capacitances with resistive loss some information on the equivalent circuit can be obtained by comparison of the readings at the two frequencies, although it is not in general possible to separate resistance and capacitance since both may vary with frequency. In d.c. resistance measurements a non-linear function of the resistance is obtained and values are determined from a calibration graph, to an accuracy of ±20 per cent over most of the range.

It is clear that the variation of capacitance, namely $10^3:1$, and the resistance range of $10^6:1$ cannot be represented, even to the low accuracy required, within the normal width of a recording meter chart on a single range. The apparatus is therefore designed to change ranges automatically. Three ranges are employed in both capacitance and resistance measurements. In the former case these are 0 to 10pF, 10 to 100pF and 100 to 1 000pF and in the latter case 0 to 5kΩ, 3 to 500kΩ and 0·3 to 500MΩ.

The specimen under test is placed in a small, specially designed furnace. Measurements below room temperature are not possible with the present equipment because of the deleterious effect of condensation on the high impedance circuits employed.

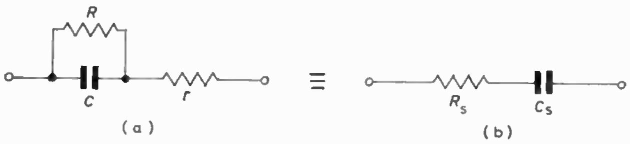


Fig. 1 (a) and (b). Equivalent circuits of specimen

The Basic Measuring Circuit

The equivalent circuit of a lossy dielectric between electrodes may be represented as shown in Fig. 1(a), where R is the parallel loss resistance, C the capacitance of the specimen and r the electrode loss. This may be reduced to the simple equivalent circuit (Fig. 1(b)) of a capacitance C_s in series with a resistance R_s where:

$$C_s = C + (1/\omega^2 CR^2) \quad \ldots\ldots\ldots\ldots\ (1)$$

and

$$R_s = r + \frac{R}{1 + R^2\omega^2C^2} \quad \ldots\ldots\ldots\ (2)$$

The basic circuit used for the measurement of the a.c. parameters of the materials is given in Fig. 2(a). The ratio of the output voltage (V_o) to the detector and the input voltage (V_i) to the specimen is given by the relationship:

$$V_o/V_i = \frac{C_s}{C_s + C_1 + j\omega C_1 C_s R_s}$$

If $C_1 \gg C_s$ (in fact $C_1 > 10C_s$ and thus to the 10 per cent accuracy of the equipment this assumption is justified), then:

$$V_o/V_i = (C_s/C_1)\frac{1}{1 + j\omega C_s R_s}$$

and

$$|V_o/V_i| = (C_s/C_1)\frac{1}{\sqrt{(1+\omega^2 C_s^2 R_s^2)}}$$

The meter reading of the detector, M, is proportional to V_o, and also the generator is designed to have a very low output impedance and a constant output voltage.

* *Formerly G.E.C. Limited, now with Cathodeon Crystals Limited.*

Hence:

$$M = K \frac{C_s}{\sqrt{(1+\omega^2 C_s^2 R_s^2)}}$$

where K is a constant.

By suitable adjustment of the meter sensitivity, V_i and C_1, the instrument can be made direct reading for pure capacitance ($R_s = 0$); the full scale readings obtained being 1 000pF, 100pF and 10pF respectively on the three ranges. Hence the capacitance reading C_r obtained for a lossy capacitance is given by:

$$C_r = \frac{C_s}{\sqrt{(1+\omega^2 C_s^2 R_s^2)}} \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

The loss angle δ of a capacitance and resistance is given by:

$$\tan \delta = \omega R_s C_s$$

Hence:

$$C_r = \frac{C_s}{\sqrt{(1 + \tan^2 \delta)}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

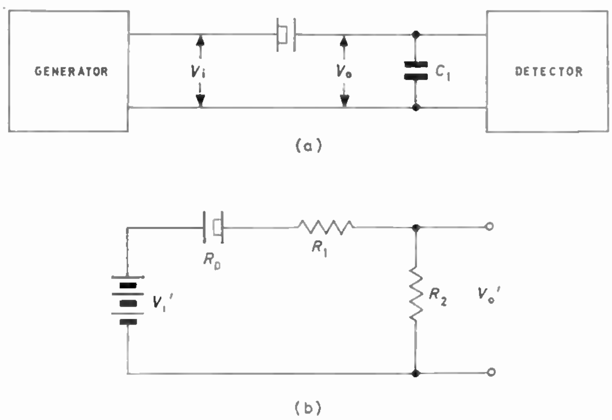


Fig. 2(a). Basic circuit for the measurement of a.c. characteristics

(b). Basic circuit for measurement of d.c. resistance

By substituting for R_s and C_s in equation (3) using equations (1) and (2), the readings obtained for various configurations of the circuit equivalent of the specimen can be obtained.

For the measurement of capacitances greater than 1000pF a series capacitance C' is inserted. In this case the reading is given by the expression:

$$C_r = \frac{C_s C'}{\sqrt{[(C_s + C')^2 + \omega^2 C_s^2 C'^2 R_s^2]}} \quad \ldots\ldots\ldots\ldots \quad (5)$$

For a pure capacitance equation (5) becomes:

$$C_r = \frac{C_s C'}{C_s + C'} \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

The basic equivalent circuit used in the measurement of resistance is shown in Fig. 2(b). The input voltage V_i' is constant and the d.c. resistance R_D is determined in terms of the output voltage V_o'.

It can be seen that

$$V_o'/V_i' = \frac{R_2}{R_1 + R_2 + R_D}$$

The meter reading M is again proportional to V_o', and V_i' is constant.

Hence:

$$M = \frac{K' R_2}{R_1 + R_2 + R_D} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

Circuit Details

A full circuit diagram of the apparatus is given in Fig. 3, and a photograph of the complete equipment is shown in Fig. 4.

In the measurement of capacitance a constant alternating voltage is applied to one electrode. This is obtained from the two valve generator (V_1 and V_2). The oscillator valve V_1 can operate either at 1Mc/s or at 100kc/s depending on the state of the frequency change relay D. This relay is normally operated automatically by the contacts in the recording meter, but for setting up can be actuated by the mechanically biased toggle switch S_1. The two oscillator frequencies are set up by adjusting C_1 and C_2. The output of V_1 is fed via C_9 to the buffer amplifier V_2. A step down transformer in the anode circuit of V_2 ensures that a low impedance is presented to the measuring network, and R_{14} forms the load for this valve. The adjustment of the capacitance calibration is carried out by varying the output of the generator; this output should be the same at 100kc/s and 1Mc/s. RV_1 varies the amplitude of oscillation of V_1 at both 100kc/s and 1Mc/s and thus adjusts the output at both frequencies. To equalize the outputs RV_2 is varied, this resistor producing a much greater change at 100kc/s than at 1Mc/s. C_{12} is used for the rough setting up of the drive to V_2 forming a capacitance attenuator in conjunction with C_9.

For the measurement of resistance, relay C is actuated by either the recording meter or by the spring loaded toggle switch S_2. This opens contact C_1 thereby applying a positive potential to the cathode of V_1, and causing it to cease oscillating. Contact C_2 is closed, applying a positive potential to the specimen. This potential is regulated by a Zener diode MR_2 to 5·6V. The resistance R_{11} forms the series resistance in the measuring circuit for the lowest resistance range, but is placed on the 'generator side' of the electrodes to ensure that accidental short-circuiting of the electrodes to chassis potential does not cause damage to the diode MR_1.

The apparatus was originally intended for the measurement of capacitance up to 1 000pF, but it was subsequently found that many ferroelectric ceramic specimens have higher capacitances. A capacitor of 1 000pF can be inserted in series with the specimen by opening S_3. Thus larger capacitances can be measured but the instrument is no longer direct reading. The capacitance is made up of C_{18} in parallel with a variable component C_{19} to enable the value to be accurately adjusted. The inductance L_6 across C_{18} provides a d.c. path when the apparatus is measuring resistance, but is arranged to have a high reactance compared with C_{18} and C_{19} at 100kc/s and 1Mc/s.

The basic measuring circuit has already been described above, but it may be seen from Fig. 3 that it is considerably complicated by the automatic range changing network. For the measuring of low capacitances the input capacitance of V_3 is about 100pF including strays. The screened lead used in connexions forms the major part of this capacitance, C_{24} providing an accurate adjustment. When the capacitance of the specimen under test rises to above 10pF contact A_2 closes and C_{22} and C_{23} are connected in parallel with C_{24}. These three capacitors are arranged to have a combined capacitance of about 1 000pF, adjustment being by C_{23}. For specimens having a capacitance greater than 100pF C_{20} and C_{21} in series are connected in parallel with C_{22}, C_{23} and C_{24}. A series combination is used to enable a combined capacitance of exactly 10 000pF to be obtained, with ordinary capacitors. Adjustment is made on this range by varying RV_1. Once

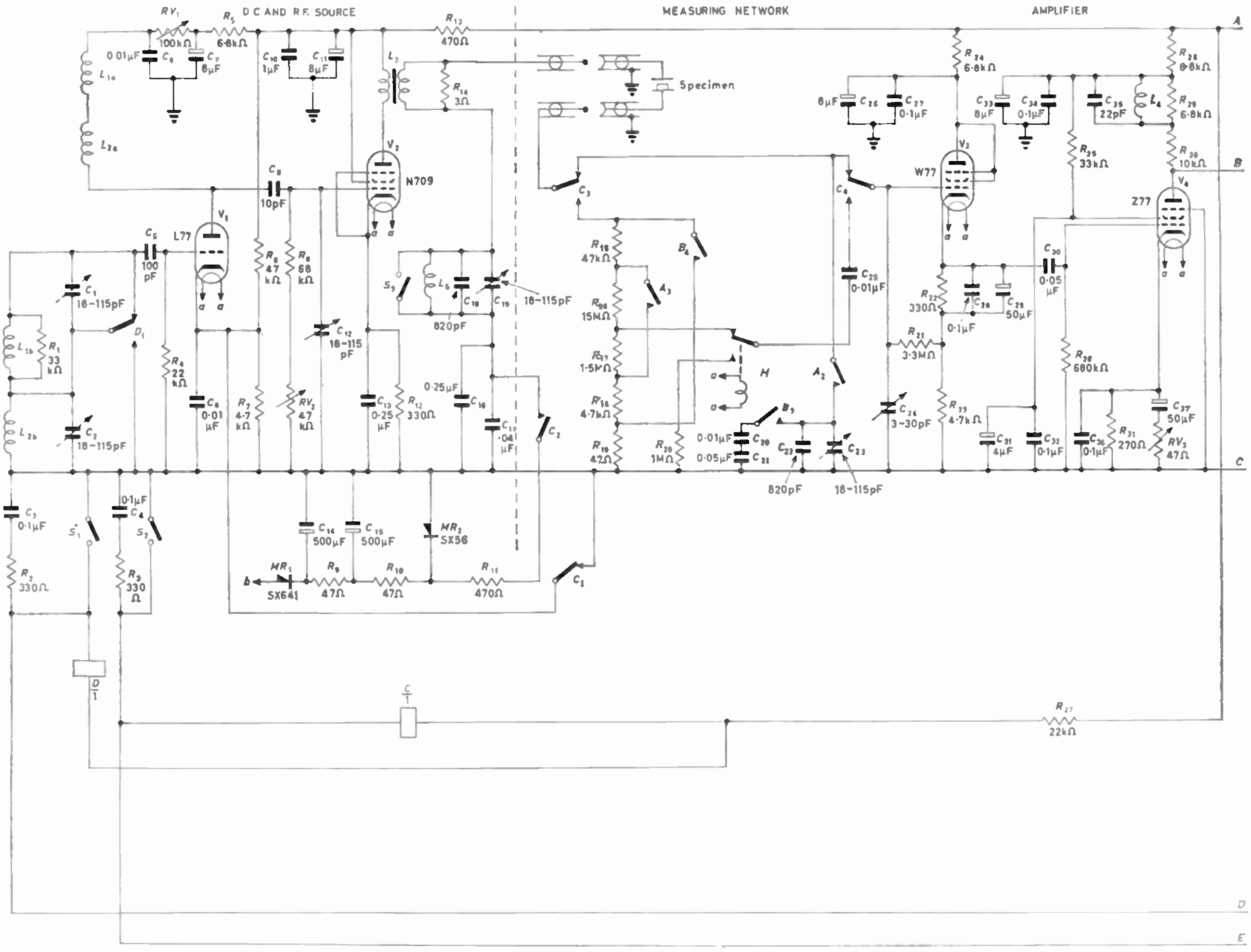


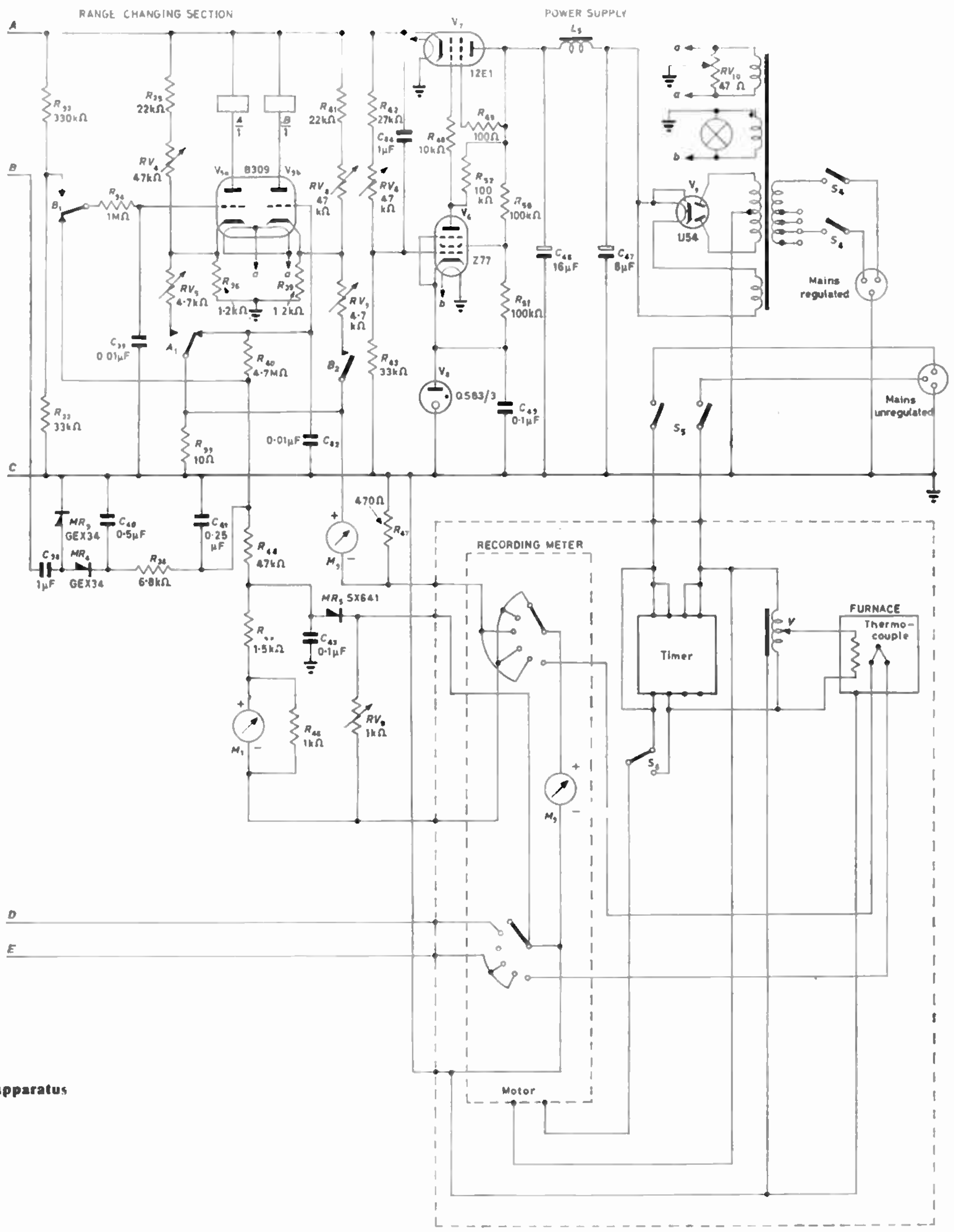
Fig. 3. Circuit diagram of

C_{23} and C_{24} have been set up the change in calibration on all ranges due to variations in valve characteristics, etc., can be corrected by RV_1.

For the measurement of d.c. resistance relay C is actuated. Contacts C_3 and C_4 remove the capacitive components from the grid of V_3 and introduce a resistive network and convertor unit H into the circuit. The convertor used has synchronous contacts operating at 50c/s. It is designed to have negligible pick-up and contact potentials, and may be used for converting direct voltages of as little as 1μV to alternating potentials. A changeover system of contacts is employed and at no time are both contacts simultaneously open-circuited. One contact is earthed via R_{30}, this resistance preventing excessive currents through the contacts when both are closed.

In order to reduce the number of relay contacts required, the switching circuit for changing ranges during resistance measurement is somewhat unorthodox. On the highest resistance range the two resistances in the circuit are the series combinations $(R_{11}+R_{15}+R_{16})$ and $(R_{17}+R_{18}+R_{19})$ respectively, corresponding to R_1 and R_2 in Fig. 2(b). When the resistance under measurement falls to below 300kΩ contact A_3 closes. This reduces the resistors in the measuring circuit to $(R_{11}+R_{15})$ and $(R_{18}+R_{19})$ respectively. The effect of R_{16} and R_{17} (in parallel) on the voltage fed to the convertor H is negligible since R_{16} is much less than the input impedance of the cathode-follower V_3. In a similar fashion when the resistance drops below 3kΩ contact B_4 closes and the effective measuring network then consists of the series resistance R_{11} and the parallel resistance R_{19}.

To present a sufficiently high impedance to the measuring circuit a cathode-follower V_3 is employed. This is of conventional design but must be capable of handling frequencies of 50c/s (for resistance measurements) and 100kc/s and 1Mc/s (for capacitance measurements). The output of this stage is fed via C_{30} to the amplifier valve V_4. This valve is arranged to have equal gain at 100kc/s and 1Mc/s, and this is achieved by suitable choice of L_4 and C_{35}. At 50c/s the gain may be adjusted by varying RV_3. This applies a variable amount of negative feedback



apparatus

to V_4 since C_{36} only provides adequate cathode by-passing at the higher frequencies.

The output of V_4 is fed via the capacitor C_{38} to the voltage doubling detector comprising MR_3 and MR_4. The potential obtained is smoothed by C_{40}, R_{38} and C_{41} and is fed via the dropping resistor R_{44} to the recording meter M_3 and the setting up meter M_1. MR_5 is used to limit the voltage reaching M_1 and M_3, should either a short-circuit occur across the specimen during a capacitance measurement, or any voltage (e.g. mains pick-up) be introduced into the high impedance electrode circuit. RV_9 is a shunt which is adjusted so that M_1 and M_3 give the same reading.

The output from the detector is also fed to the grid of the left-hand section of the range change valve V_{5a} via the relay contacts B_1 and the time delay circuit $R_{34}C_{39}$. V_{5a} is normally kept non-conducting by a positive cathode potential, which may be adjusted by RV_4. When the voltage from the detector exceeds a certain level, determined by RV_4, V_{5a} conducts and relay A closes, performing three functions. Firstly contact A_1 connects RV_5, in series with R_{37}, across R_{36}, and the cathode potential of V_{5a} is reduced by about 10:1. Hence A_1 will not open again unless the voltage on the grid of V_{5a} falls to just under $1/10^{th}$ of its original value. Secondly the opening of A_1 allows a voltage from the detector to reach the grid of V_{5a} after a suitable time delay produced by C_{42} charging through R_{40}. This allows the circuit to stabilize before a second range change can occur. Thirdly the closing of contacts A_2 and A_3 causes a change of range in the appropriate measuring circuit. This will of course produce a reduction of 10:1 in the voltage from the detector. RV_5 is adjusted to ensure that a drop of slightly greater than 10:1 is required to release relay A and therefore 'hunting' between ranges will not occur.

The flow of current through R_{34} gives a deflexion of the range indicating meter M_2 and also a deflexion of the 'range' trace on the recording meter M_3.

For the second range change V_{5b} operates in a similar manner to V_{5a}. A set of changeover contacts B_1 in the grid of V_{5a} ensure that after relay B is closed V_{5a} is kept conducting. Moreover should the range change in the reverse direction, V_{5a} will not immediately cease to conduct because of the time delay introduced by R_{34} and C_{39}. This will allow the measuring circuit to stabilize before the first range change circuit becomes operative once again.

The current passing through RV_7 when V_{5b} conducts causes a further increase in the current through R_{37} giving an abrupt discontinuity in the readings of M_1 and M_3 indicating the second change of range.

Since the calibration of the instrument depends not only on a constant output from the generator but also a constant gain in the amplifier circuit, a regulated power supply is essential. The h.t. is supplied from a conventional series stabilizer circuit comprising V_6, V_7 and V_8 and the associated components. The electronic part of the equipment is also fed from a constant voltage transformer ensuring adequate stabilization of the heater voltage. The furnace timing switch and recorder motor are fed from the second mains socket via S_5, since neither of these require regulation.

The rate of heating of the furnace can be adjusted by means of the Variac transformer. The time for which the furnace is heating is determined by the process timer. When the set time has elapsed the furnace is switched off by means of a self-holding relay within the timer. This relay is released by momentarily opening S_5, thus restarting the cycle. The recording meter, which as previously mentioned actuates the function changing relays within the equipment, may be operated either directly from the mains via S_5 or from the time switched supply depending on the position of S_6. In the former case, curves for both heating and cooling are automatically plotted, and in the latter case only the heating cycle is recorded.

It should be noted that the negative side of the detector is earthed direct to the meter. This prevents the current through the meter leads giving a deflexion on the meter when the function selection relays are operated.

Furnace and Electrode Design

A general view of the furnace and electrode system is given in Fig. 5. A small furnace is employed, non-inductively wound on a steel tube. This is designed to operate at voltages up to 240V and has a maximum temperature of 500°C.

Within the furnace fits a stainless steel container for the electrodes and specimen, and this also forms additional screening. For the measurement of small capacitances it is essential that the strays should be kept to a minimum. This is achieved by taking the leads into the furnace through two steel tubes and dividing the container by a steel plate. The lower electrode is mounted on this plate using mica insulation. This electrode is connected to the generator and thus the stray capacitance to the plate has negligible effect because of the low impedance of the circuit. The upper electrode is movable and is clamped under a Nimonic strip. Connexion to this electrode is by a flexible silver wire seen at the rear of the system. The thermocouple for temperature measurement is visible in front of the lower electrode. Because of the method of connexion of the meter, the thermocouple must not make any electrical connexion to the rest of the system.

Considerable difficulty has been encountered in the design of the furnace interior. The first setback was due to oxidation of the components including the electrodes which were originally of stainless steel. Hard soldering all the electrical connexions within the furnace and facing the electrodes with heavy silver plates gave improved results, but because of the high temperature employed

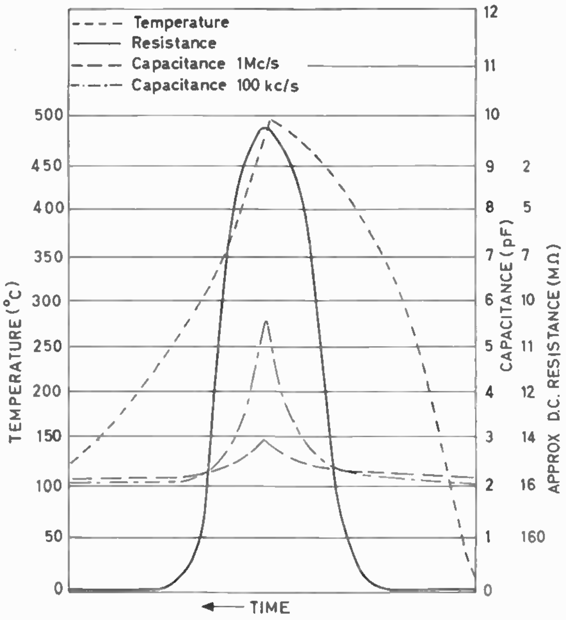


Fig. 6. Measurements on a lossy dielectric

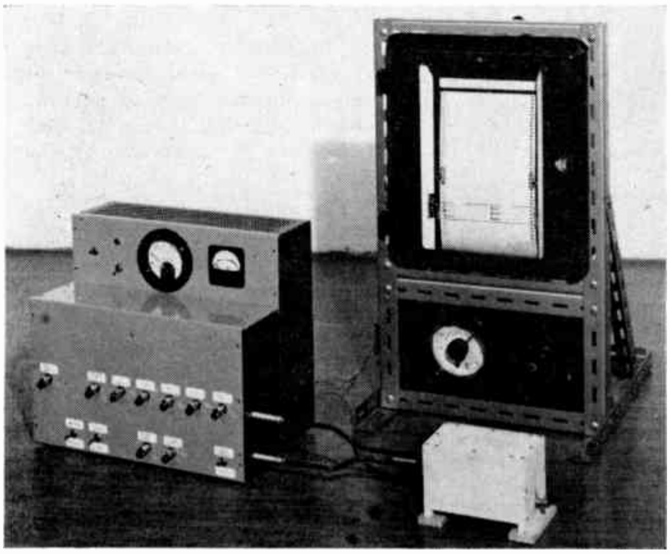


Fig. 4 (above). Complete measuring equipment

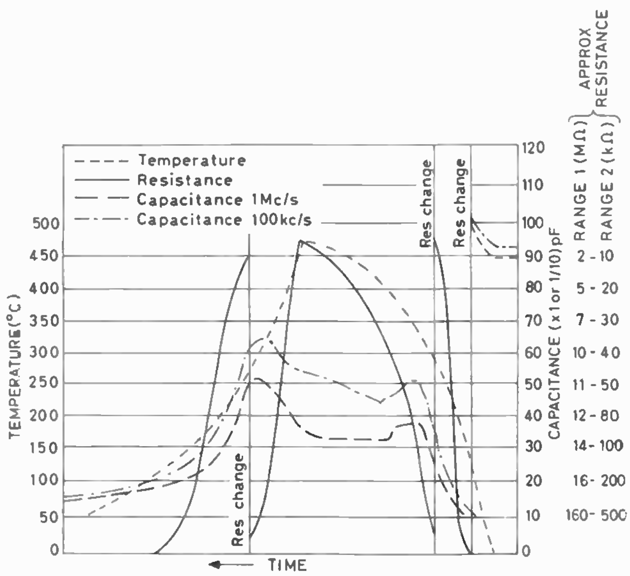
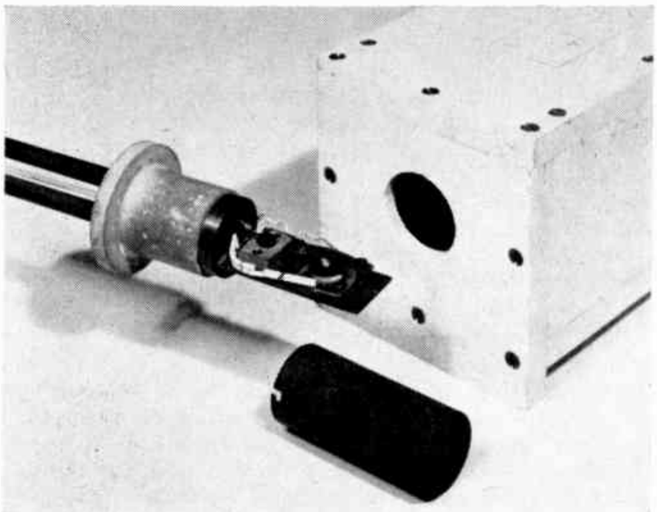


Fig. 7. Measurements on a ferroelectric ceramic

Fig. 5 (below). Furnace and electrodes



the silver tended to corrode and distort, and the hard solder deteriorated, possibly because of diffusion into the silver. Electrodes of 25 per cent iridium/platinum alloy are now employed and these give satisfactory results.

Some difficulty has also been occasioned by the condensation of water in the connecting tubes. This produces a voltaic e.m.f. causing incorrect readings of resistance and also losses causing errors on the capacitance ranges. Initially the water was found to come from the fish-spine beads used for insulation within the tubes. This has been eliminated by using small glass beads for insulation, and also by making a number of holes in the heated portion of the upper tube (see Fig. 5). These holes allow any water vapour from within the furnace to escape instead of condensing in the cooler portions of the tube.

Interpretation of Results

Two typical results plotted by the apparatus are reproduced in Figs. 6 and 7. The sets of curves in Fig. 6 show a material which merely becomes lossy at high temperature, and, in Fig. 7, a material which is ferroelectric and exhibits large changes of capacitance near the Curie temperature.

The three parameters plotted are:

(1) D.C. resistance.

(2) A function of resistance and capacitance at 100kc/s, (represented by equation (3)).

(3) A similar function of resistance and capacitance at 1Mc/s.

The range change traces have been omitted for clarity.

Most materials show a rise in d.c. conductivity with rising temperature. It is also found that when this occurs both $\tan\delta$ and the equivalent series capacitance C_s become functions of frequency, and a separation of the two 'capacitance' traces occurs. In the case of the curves plotted in Fig. 6 it can be seen that both the 'capacitance' curves rise as the loss increases. This denotes a fall in the parallel loss resistance. From the curves:

At a temperature of 500°C

D.C. resistance	250kΩ
'Capacitance' reading C_r at 100kc/s	5·6pF
'Capacitance' reading C_r at 1Mc/s	3·0pF

At a temperature of 20° (room temperature)

D.C. resistance	Effectively infinite
Capacitance reading C_r	2·2pF

(Approximately equal at both frequencies)

Let it be assumed that the d.c. resistance R_d is equal to the parallel loss resistance R, the contact resistance r is zero, and that the equivalent parallel capacitance C is independent of temperature. If the value of C obtained at room temperature and the value of R assumed at 500°C are substituted in equations (1) and (3) then the calculated values obtained for capacitance reading C_r are 6·7pF at 100kc/s and 2·3pF at 1Mc/s. It can be seen that these values, although not showing exact agreement with those obtained in practice are of the right order of magnitude, and the assumptions made represent a reasonable approximation.

Considering Fig. 7, it can be seen that the d.c. conductivity rises rapidly with temperature. The changes of range are marked by vertical lines for clarity. The rate of rise shows a discontinuity at about 360°C. The 'capacitance' readings initially rise with temperature but reach a maximum at about 320°C; a small discontinuity occurs at about 415°C after which a further general rise in apparent capacitance occurs. It should be noted, however, that if a ceramic becomes conducting the value of 'capacitance' is inflated by conduction effects between the grains in the material and the results have little meaning. Generally it is considered that the 'capacitance' of the material has little significance if $\tan\delta$ exceeds 0·5. It should be noted that due to hysteresis effects and different rates of heating and cooling, the values for 'capacitance' are different when the temperature is rising from those when it is falling.

As both capacitance and resistance are varying it is not possible to separate these two parameters but transition temperatures can easily be seen. Normally, inspection of the curves shows if any sample is worthy of subsequent investigation, but as shown above further information can be obtained by inserting values in equation (3).

Preparation of Specimens

Samples for measurement are prepared in the form of disks 1·2cm in diameter and about 0·5 to 2mm thick. With high permittivity materials, which are usually ceramics, it is essential that the electrodes make good electrical contact with the whole surface area so that series-capacitance effects are eliminated. This is achieved by coating the major surfaces with sputtered or evaporated gold. Materials in the form of powders or small crystals can be converted into suitable disks by grinding, die-pressing at 20 000 lb/in^2, and subsequent hydrostatic pressing at 70 000 lb/in^2. If the materials have a low permittivity the use of electrodes deposited directly on to the material is unnecessary.

Lithium Nickel Ferrite Core Stores

Lithium-nickel ferrite storage cores offer considerable advantages to the computer manufacturer in comparison with the cores previously available. In particular, cores made from the new ferrite material can operate over a wide temperature range without the need for current compensation.

These cores can readily operate over a temperature range of 60°C with constant drive-current amplitude so that neither temperature compensation of the drive current nor temperature stabilization of the matrix is required. The sense amplifiers of the core store can also be simplified because over this range with a constant drive current, the peaking and switching times remain substantially constant and no compensation of strobing time is required.

The lithium-nickel ferrite cores have 'usable' properties over the temperature range −20°C to +700°C. The magnetostrictive coefficient is very low, resulting in low stress sensitivity and negligible oscillation through magnetostrictive excitation.

At present two types of core made from this material are available from Mullard Ltd. The FX2763 (0·03in outside diameter) is intended for use in high-speed stores with cycle times of less than 2μsec whereas the FX2764 (0·05in outside diameter) is for use in slower stores.

Both cores are available assembled as single matrix planes or complete stacks.

True Third Harmonic Tuning of E.H.T. Transformers

By F. D. Bate*, B.Sc., A.M.I.E.E.

An analysis is given of third harmonic tuning as used in line flyback e.h.t. supplies on the assumption that the two frequencies present are f and 3f.

Various curves have been drawn which enable a design to be made for optimum third harmonic tuning.

(Voir page 727 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 734)

The use of third harmonic tuning in line flyback e.h.t. supplies as used in television receivers is well established. It arose because of the fact that it is impossible to construct a transformer having a coefficient of coupling exactly equal to unity. Thus there was always some leakage inductance between the e.h.t winding, and the rest of the transformer.

The presence of this leakage inductance also meant that certain self- and stray-capacitances of the e.h.t. winding could no longer be treated as being effectively in parallel with the transformer primary. The result of this is that the transformer part of the actual circuit of Fig. 1(a)

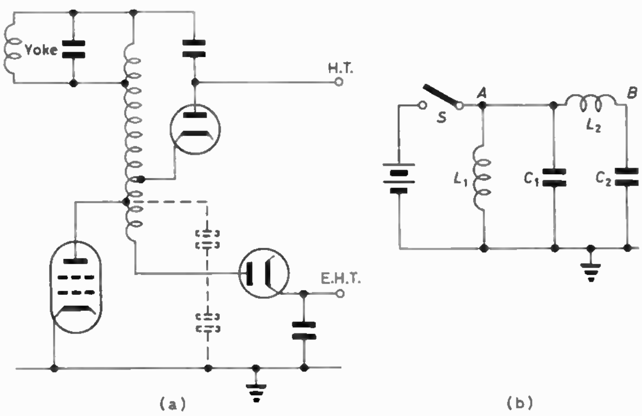


Fig. 1(a). Basic valve line scanning circuit

(b). Equivalent circuit to explain third harmonic tuning

has an equivalent circuit as shown by Fig. 1(b). A detailed analysis giving the equivalent circuit has already been covered in articles by Cherry[1] and Murakami[2].

In Fig. 1(b), C_1 is the external capacitance across the scanning coil, L_1 is the combined inductance of the scanning coil and the primary inductance of the transformer, L_2 is the leakage inductance between primary and secondary windings and C_2 a combination of the self-capacitance of the e.h.t. overwind and stray-capacitances, all components being referred to the primary of the transformer. More exact equivalent circuits usually have an extra capacitor across L_2 and an extra inductor across C_2, these however, only make the problem more difficult and to a first approximation, can be neglected.

The object of the investigation is to find out what are the currents and voltages of the equivalent circuit as given by Fig. 1(b), after the interruption of a current flowing in the inductance L_1. The analysis of the circuit usually given assumes that a linearly rising current flows in L_1 which is then open-circuited by the switch S. This leads to a ratio of the two frequencies present of approximately 2·8 : 1 if there is to be no current in L_2 at the end of the fly-back time. Present 110° tubes require this current to be S-shaped, a condition that will make the frequency ratio approach closer to 3.

In the case of transistor circuits the current source cannot be reduced to zero value in zero time, but requires a finite time of 1 to 2μsec for zero current to be reached. Both of these conditions make the actual problem more complex. It was thus felt that the essentials of the problem would not be lost by assuming that the frequency ratio is exactly 3 : 1, this assumption having the merit of relative simplicity. The waveforms of the current pulses applied to the circuit for the different assumptions, and the actual current in practical circuits, are shown in Fig. 2.

Determination of the Minimum Voltage Obtainable Using Third Harmonic Tuning

In the equivalent circuit of Fig. 1(b) point A is the anode of a valve, or collector of the transistor, and the voltage waveform here consists of two frequencies f_1 and $3f_1$ added together in phase as shown in Fig. 3(b). The voltage waveform which would be present if no third harmonic tuning were used, that is, with L_2 and C_2 omitted is shown in Fig. 3(a), and since the mean value of these

NOMENCLATURE USED

C_1	=	External capacitance across the reflected yoke inductance
C_2	=	Reflected self-capacitance of overwind
f	=	Frequency of oscillation in general
f_1	=	Fundamental frequency
f_2	=	Third harmonic frequency
I	=	Current at any instant through L_2C_2
L_1	=	Reflected yoke inductance
L_2	=	Reflected leakage inductance
n	=	$(\omega_2/\omega_1)^2$
P	=	$\sqrt{(L_2C_2/L_1C_1)}$
t	=	Time at any instant
T_f	=	Flyback time
v	=	Voltage at A (Fig. 1) at any instant
V	=	Voltage at A (Fig. 1) if the fundamental frequency only is present
V_1	=	Fundamental frequency voltage at point A
V_2	=	Third harmonic frequency voltage at point A
V_B	=	Voltage at point B (Fig. 1) = $V_{f1} + V_{f2}$
V_{f1}	=	Fundamental frequency voltage at point B
V_{f2}	=	Third harmonic frequency voltage at point B
ω	=	$2\pi f$
ω_1	=	$2\pi f_1$
ω_2	=	$2\pi f_2$
V_{max} and V_{min}	=	Maximum and minimum voltage points during flyback (Fig. 3).

* *Thorn-AEI Radio Valves & Tubes Ltd*

waveforms must be equal for the same flyback time then,

Mean value under V (Fig. 3(a)) = Mean value under V_1 and V_2 (Fig. 3(b))

$$2V/\pi = (2/\pi)V_1 + (2/3\pi)V_2$$

or:

$$V = V_1 + (V_2/3) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

the equation of the voltage waveform resultant in Fig. 3(b), i.e. point A in Fig. 1(b) is given by:

$$v = V_1 \sin \omega t + V_2 \sin 3\,\omega t \quad \ldots\ldots\ldots\ldots \quad (2)$$

and using equation (1) with $\sin 3\omega t = 3 \sin \omega t - 4 \sin^3 \omega t$ gives

$$v = (9V - 8V_1)\sin \omega t - 12(V - V_1)\sin^3 \omega t \quad \ldots\ldots \quad (3)$$

Assuming for the moment that V and V_1 are constant the positions of maximum or minimum can be found by differentiating equation (3) with respect to ωt. This gives

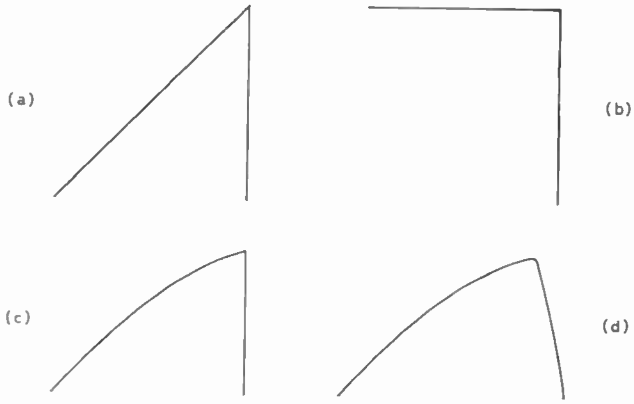


Fig. 2(a). Current pulse normally applied to Fig. 1(b) which gives a frequency ratio approximately 2·8

(b). Current pulse applied to Fig. 1(b) which gives a frequency ratio of exactly 3

(c). Actual current pulse with 110° tubes. Valve output stage.

(d). Actual current pulse with 110° tubes. Transistor output stage

i.e. $V_2 = (1/9)V_1$

This means that provided $V_2 \lesseqgtr (1/9)V_1$ then only one maximum occurs during the flyback. When $V_2 > (1/9)V_1$ then $\sin^2 \omega t < 1$ and a minimum occurs at half-way through the scan with two maxima placed symmetrically on either side.

Taking the value of $\sin^2 \omega t$ given by equation (4) and substituting it in equation (3):

$$v = 1/9\,\frac{(9V - 8V_1)^{3/2}}{(V - V_1)^{1/2}} \quad \ldots\ldots\ldots\ldots \quad (5)$$

Differentiating v with respect to V_1 and equating to zero to obtain the maximum and minimum values of v then:

$$V_1 = (15/16)V \text{ and } V_2 = (3/16)V$$

(restricting to positive values for V_1 and V_2)

$$\text{or } V_2 = (1/5)V_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

Substituting these values of V_1 and V_2 in equation (5) gives for the minimum value of v

$$v = V(\sqrt{2}/\sqrt{3}) = 0{\cdot}815V$$

at halfway through the scan $t = T/2$

$$v = V_1 - V_2 = 0{\cdot}75V$$

These results show that the minimum voltage that can theoretically be obtained using third harmonic tuning is 81·5 per cent of the voltage with no third harmonic tuning, two positive peaks occur and the value of the minimum in between these peaks is 75 per cent of the voltage with no third harmonic tuning, see Fig. 3(c).

Determination of the Optimum Values of C_1, L_2 and C_2 When L_1 is Fixed

In determining the choice of components so that the minimum voltage occurs across L_1 and C_1 in the equivalent circuit three independent equations must be obtained. L_1 is the reflected yoke inductance and this is determined by parameters outside the circuit, hence C_1, L_2 and C_2 remain to be found for optimum third harmonic tuning.

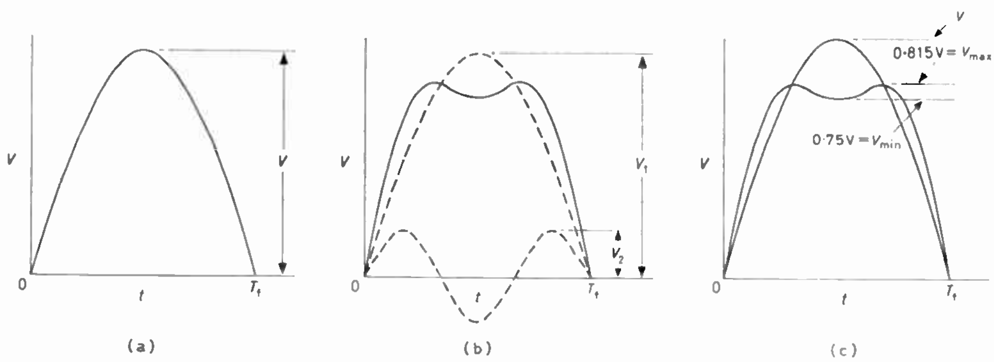


Fig. 3(a). No third harmonic tuning

(b). Third harmonic tuning

(c). Relative magnitudes of the voltage present with and without third harmonic tuning

either

$$\cos \omega t = 0$$

or

$$\sin^2 \omega t = \frac{9V - 8V_1}{36(V - V_1)} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

When $\cos \omega t = 0$ then a maximum or minimum occurs half-way during the flyback, and if a maximum occurs, then clearly only one maximum is present during the flyback.

$$\text{If } \sin^2 \omega t = 1 = \frac{9V - 8V_1}{36(V - V_1)}$$

then $V_1 = (27/28)V \qquad V_2 = (3/28)V$

Referring to the equivalent circuit, Fig. 1(b) it can easily be shown that the impedance looking across L_1 is infinite at the frequencies given by the solution of

$$\omega^4 L_1L_2C_1C_2 - \omega^2 (L_1C_1 + L_1C_2 + L_2C_2) + 1 = 0$$

i.e.

$$\omega^2 = \frac{(L_1C_1 + L_1C_2 + L_2C_2) \pm \sqrt{[(L_1C_1 + L_1C_2 + L_2C_2)^2 - 4L_1L_2C_1C_2]}}{2L_1L_2C_1C_2}$$

writing the two frequencies as ω_1 and ω_2, and $\omega_2 = 3\omega_1$ gives:

$$L_1C_1 + L_1C_2 + L_2C_2 = 10/3\,\sqrt{(L_1C_1L_2C_2)} \quad \ldots\ldots\ldots \quad (7)$$

$$\text{This gives } \omega_1^2 = \frac{1}{3\sqrt{(L_1 C_1 L_2 C_2)}} \qquad \omega_2^2 = \frac{3}{\sqrt{(L_1 C_1 L_2 C_2)}} \quad \Bigg\} \ldots\ldots (8)$$

Thus the time of flyback is given by

$$T_f = \pi\sqrt{3}(L_1 C_1 L_2 C_2)^{1/4} \quad \ldots\ldots\ldots\ldots (9)$$

Equations (7) and (9) give two independent equations between the quantities, the third equation being obtained when $V_2 = 1/5\ V_1$ equation (6). Returning to the equiva-

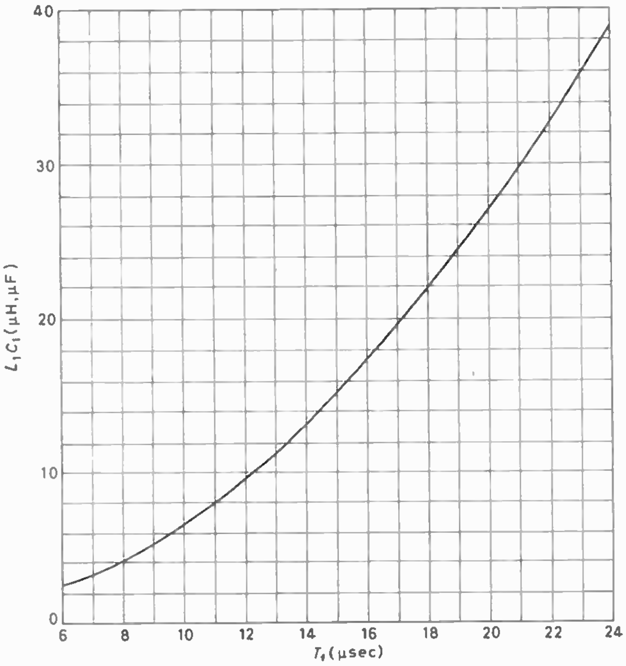


Fig. 4. Graph of L_1C_1 against T_f

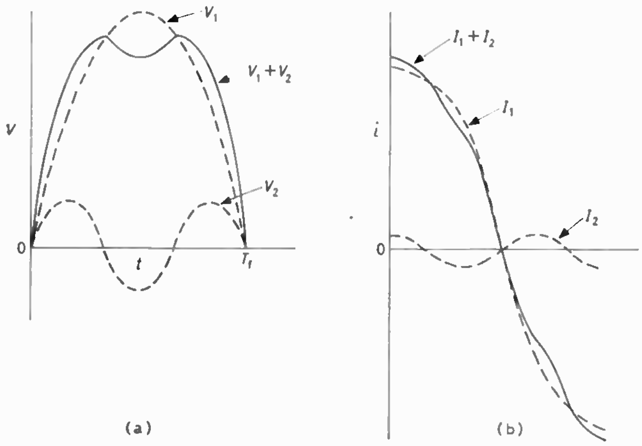


Fig. 5(a). Voltage at point A (see Fig. 1) optimum value
(b). Current through L_1 (see Fig. 1) optimum value

lent circuit once again and assuming in the $L_2\ C_2$ arm that the currents have the same magnitude at each frequency (they are in fact produced 180° out of phase in this arm) then from Fig. 5,

$$V_1 = I\,(j\,\omega_1 L_2 + (1/j\,\omega_1 C_2)) = (15/16)\,V \quad \ldots\ldots (10a)$$

$$V_2 = -I\,(j\,\omega_2 L_2 + (1/j\,\omega_2 C_2)) = (3/16)V \quad \ldots\ldots (10b)$$

dividing equation (10a) by equation (10b) gives

$$\frac{1 - \omega_1^2 L_2 C_2}{1 - \omega_2^2 L_2 C_2} = -5\,(\omega_1/\omega_2) \quad \ldots (11)$$

Using equation (8) and $(\omega_1/\omega_2) = (1/3)$ one arrives at the equation $4L_2C_2 = L_1C_1$ which is the third equation.

Thus the three equations necessary for the choice of components are

$$L_1C_1 + L_1C_2 + L_2C_2 = (10/3)\sqrt{(L_1C_1L_2C_2)} \quad \ldots\ldots (7)$$

$$T_f = \pi\sqrt{3}\,(L_1L_2C_1C_2)^{1/4} \quad \ldots\ldots\ldots\ldots (9)$$

$$4\,L_2C_2 = L_1C_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots (12)$$

These may be re-written in the form.

$$L_1L_2C_1C_2 = T_f^4/9\pi^4 \quad \ldots\ldots\ldots\ldots (9a)$$

$$L_1C_1 + L_1C_2 + L_2C_2 = 10T_f^2/9\pi^2 \quad \ldots\ldots\ldots (7a)$$

$$4L_2C_2 = L_1C_1 \quad \ldots\ldots\ldots\ldots\ldots (12)$$

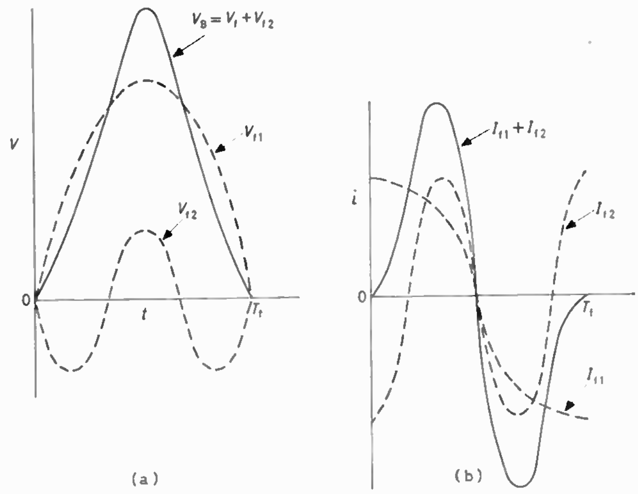


Fig. 6(a). Voltage at B (see Fig. 1)
(b). Current through L_2 (see Fig. 1)

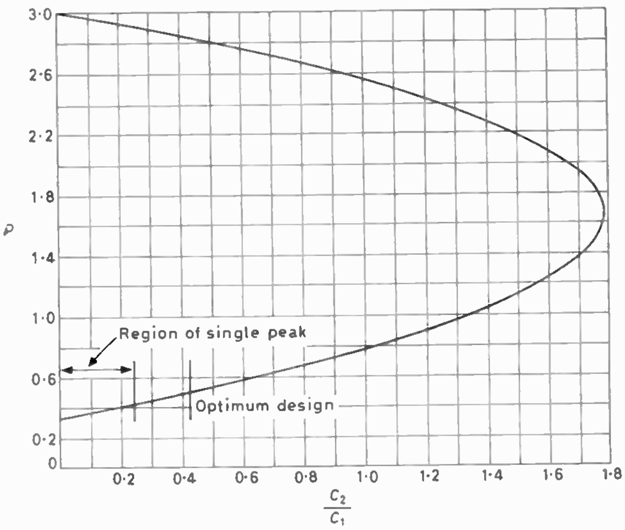


Fig. 7. Graph of $P \equiv (L_2C_2/L_1C_1)$ against C_2/C_1

Combining equations (9a) and (12) gives:

$$\left.\begin{aligned} L_2C_1 &= (3/\pi^2)T_f^2 \\ \text{hence also } C_2 &= (5/12)\,C_1 \\ L_2 &= (3/5)\,L_1 \end{aligned}\right\} \ldots (13)$$

Equations (13) enable all the components to be chosen for optimum performance.

Calculation of the Voltage at the Point *B* (Fig. 1)

It is of interest to calculate the voltage present at point *B* in Fig. 1(b). If V_{f1} and V_{f2} are the two voltages at point *B* then

$$V_{f1}/V_{f2} = \frac{I/\omega_1 C_2}{I/\omega_2 C_2} = \omega_2/\omega_1 = 3 \quad \ldots\ldots\ldots (14)$$

this means that at the point B the voltage of the fundamental frequency f_1 is three times the value of the voltage at the third harmonic frequency f_2 for all possible values of $L_1C_1L_2C_2$ which satisfy equations (7) and (9). Returning to equations (10):

$$V_1/V_2 = -\left(\frac{1-\omega_1^2L_2C_2}{1-\omega_2^2L_2C_2}\right)\omega_2/\omega_1 \quad \ldots\ldots\ldots (15)$$

and using equations (8), (10) and (1) it can be shown that

$$V_B = 1{\cdot}5V \quad \ldots\ldots\ldots\ldots\ldots\ldots (16)$$

This equation indicates that the final voltage peak at B is 1·5 times the voltage at A, in other words provided the transformer is third harmonically tuned the effect of the leakage inductance and self-capacitance is to increase the voltage by 50 per cent. Thus, finally, the true step-up of the circuit relative to the non-third harmonic case under optimum conditions is given by $(1{\cdot}5/0{\cdot}815) = 1{\cdot}84$ or a voltage gain of 84 per cent over the case with no third harmonic tuning. This compares with a value in practice of about 55 per cent step-up. The voltage at A in the circuit being 85 per cent in practice compared with 81·5 per cent calculated, when compared with no third harmonic tuning.

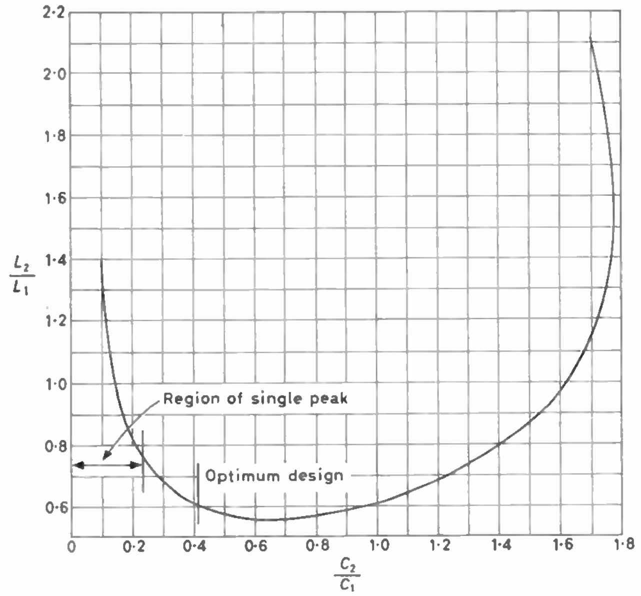


Fig. 8. Graph of C_2/C_1 against L_2/L_1

A graph is plotted of L_1C_1 as a function T_f, the flyback time, Fig. 4, for the case where optimum components are used, equation (13).

Example (Using a Transistor Circuit)

Suppose there is a value for L_1 from external considerations of 100μH and a flyback time of 20μsec, then C_1 from the graph $= 0{\cdot}27\mu$F, $C_2 = 0{\cdot}11\mu$F and $L_2 = 60\mu$H from equation (13).

Third Harmonic Tuning in General

In general the equivalent circuit of Fig. 1 can be third harmonically tuned over a wide range of values of $L_1C_1C_2$ and L_2, the limitation on these values being given by equation (7),

i.e $L_1C_1 + L_1C_2 + L_2C_2 = (10/3)\sqrt{(L_1C_1L_2C_2)}$ …… (7)

dividing this by L_1C_1 and writing $P \equiv \sqrt{(L_2C_2/L_1C_1)}$

gives $\quad 1 + (C_2/C_1) + P^2 = (10/3)P$ ………. (17)

Using equation (17) graphs are plotted of P against C_2/C_1, Fig. 7, and C_2/C_1 against L_2/L_1, Fig. 8. These graphs show that $C_2/C_1 \leqslant (16/9)$, that is the reflected self-capacitance of the coil must always be less than $(16/9)\,C_1$, and also that $L_2/L_1 \geqslant (9/16)$, that is the reflected leakage inductance cannot be less than $(9/16)\,L_1$. The points on the curve for optimum design are also shown, as also the region where only one peak is present on the curve, this being the region where $C_2/C_1 \leqslant (12/49)$ and $L_2/L_1 \geqslant (3/4)$.

Method of Designing a Transformer Having the Minimum Voltage Present at Point A (Fig. 1(b))

Although equation (13) gives component values for optimum performance, a transformer may be designed which is third harmonically tuned, but where the peak voltage is not a minimum, the following method enables one to calculate the change in components necessary for optimum performance. Now, given that peak voltage at A in Fig. 1(b) where $V_1/V_2 \leqslant 9$ is given by equation (5) and V_{min} which occurs half-way through the flyback is given by:

$$V_{min} = V_1 - V_2 \quad \ldots\ldots\ldots\ldots\ldots (18)$$

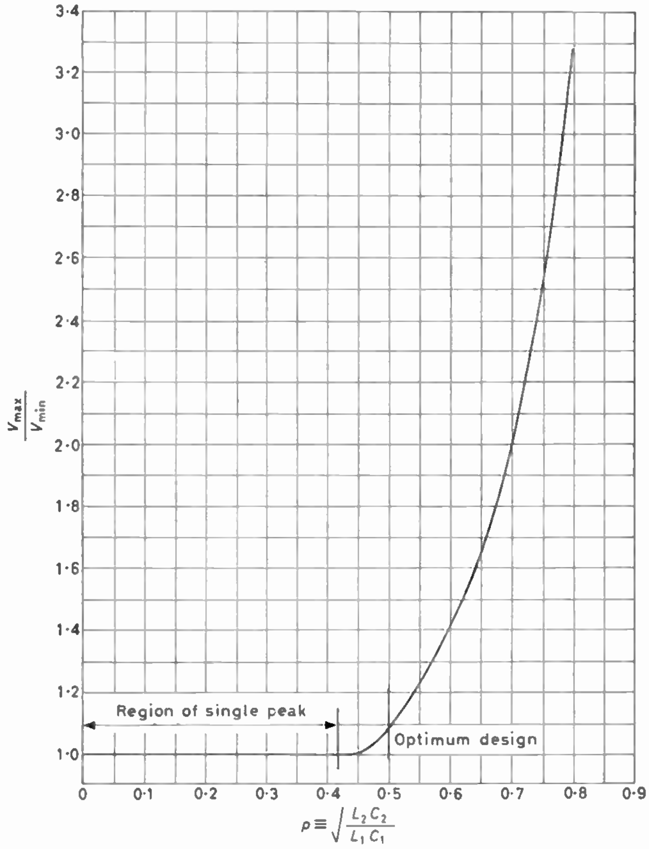


Fig. 9. Graph of $P \equiv (L_2C_2/L_1C_1)$ against V_{max}/V_{min}

combining equations (5), (18) and (1) gives

$$V_{max}/V_{min} = 1/(3\sqrt{3})\,\frac{(\alpha+3)^{3/2}}{(\alpha-1)}$$

where $\alpha \equiv V_1/V_2 = \dfrac{P-3}{1-3P}$ from equations (8) and (15).

Hence

$$V_{max}/V_{min} = 4\sqrt{2}/(3\sqrt{3})\,\frac{P}{1-P}\,\frac{P}{3P-1} \quad \ldots\ldots (19)$$

this equation is valid provided $\alpha \leqslant 9$ i.e. $P \geqslant 3/7$. When $\alpha = 9$ i.e. $P = 3/7$ $V_{max}/V_{min} = 1$.

A graph is plotted of V_{max}/V_{min} against P, Fig. 9. Before the graphs can be used it is necessary to take equation (9a) which determines the flyback time.

$$L_1L_2C_1C_2 = T_f^4/9\pi^4 \quad \text{.............. (9a)}$$

which when combined with $P = \sqrt{(L_2C_2/L_1C_1)}$ gives:

$$P L_1C_1 = T_f^2/3\pi^2 \quad \text{............. (27)}$$

Example Using the Graphs and Equation (27)

Suppose there is a transformer third harmonically tuned and where V_{max}/V_{min} as measured on an oscilloscope is 1·5 and it is desired to redesign this for optimum performance i.e. $V_{max}/V_{min} = 1{\cdot}088$, the flyback time T_f and L_1 to remain the same. Then from Fig. 9 for $V_{max}/V_{min} = 1{\cdot}5$, $P = 0{\cdot}62$ whereas for optimum design P should be 0·5, thus, using equation (27), since T_f and L_1 are fixed the new value of C_1 should be $0{\cdot}62/0{\cdot}5\ C_1 = 1{\cdot}24\ C_1$ i.e. C_1 should be increased by 24 per cent. Referring to Fig. 7 when $P = 0{\cdot}62$, $C_2/C_1 = 0{\cdot}7$ and when $P = 0{\cdot}5$, $C_2/C_1 = 0{\cdot}42$ thus the new value of C_2 is given by $0{\cdot}42/0{\cdot}7\ 1{\cdot}24$, $C_2 = 0{\cdot}75\ C_2$ i.e. C_2 must be reduced by 25 per cent. Finally from Fig. 8 L_2/L_1 must be changed from 0·56 when $C_2/C_1 = 0{\cdot}7$, to 0·6 for optimum design when $C_2/C_1 = 0{\cdot}42$, and since L_1 is fixed the new value of L_2 is $0{\cdot}60/0{\cdot}50\ L_2 = 1{\cdot}07$, i.e. L_2 must be increased by 7 per cent.

A similar procedure is used when T_f and L_1 have also to be changed.

Conclusion

It has been shown theoretically that the minimum voltage obtainable at the valve anode or transistor collector using third harmonic tuning is 81·5 per cent of the voltage with only the fundamental frequency present, and that the voltage step up of the transformer is increased by 84 per cent over the case of no third harmonic tuning, assuming a four component equivalent circuit. These compare with measured values of 85 per cent and 55 per cent respectively. Various curves have been drawn which enable a design to be made for optimum third harmonic tuning.

Acknowledgments

The author wishes to thank the Management of the Thorn-AEI Applications Laboratory for permission to publish this article.

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Non-Linear CR Circuits Using Silicon Carbide Varistors

By W. G. P. Lamb*, Ph.D.

It is shown in this article that the transient behaviour of non-linear series resistance-capacitance circuits may be demonstrated or used for function generation by the use of silicon carbide non-linear resistors in conjunction with operational amplifiers.

(Voir page 727 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 734)

SEVERAL workers on analogue computing techniques have made use of the non-linear properties of silicon carbide in the generation of mathematical functions. For example, Brown and Walker[1] have developed a multiplier based on the quarter squares principle using square law characteristics obtained with the aid of silicon carbide resistors and Kovach and Comley[2] have, in addition, used them to generate a number of other functions.

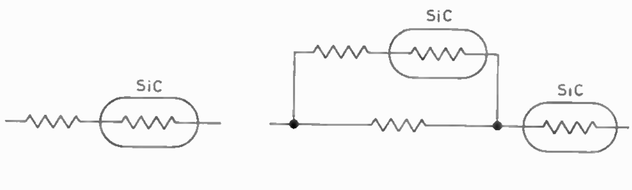
Silicon carbide resistors obey a relationship of the type $Ri = v^\alpha$ where i and v are the current and potential respectively and α has a value changing gradually from 1 to 4 as v, and consequently i, are increased. By the use of a suitably chosen series resistor it is possible to keep α sensibly constant over a useful working range for values of α lying between 1 and about 2·5. Moreover it is possible to extend the useful range by placing further non-linear resistors in parallel with the series resistor. These arrangements are shown in Figs. 1(a) and 1(b).

The present article describes how such non-linear networks may be used in conjunction with operational amplifiers to demonstrate the transient behaviour of simple series circuits comprising resistances and capacitances which are forced to exhibit non-linear characteristics.

Such a method offers an extension of the ordinary exponential decay function achieved with a conventional linear CR circuit, and yields a range of functions which retain some similarity with exponentials but have a number of different features about them.

Fig. 1(a). Compensating series resistor

(b). Compensating circuit employing additional SiC resistor



(a) (b)

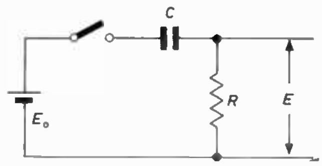


Fig. 2. The basic circuit

Non-linear Capacitance-Resistance Circuits

The present work was restricted to simple series circuits incorporating one linear and one non-linear component, the latter being either the resistor or the capacitor, and the type of non-linearity being of the form $Ri = v^\alpha$ in the case of the resistor, and $q = Cv^\alpha$ for the capacitor. The cases of $\alpha = \frac{1}{2}$ or 2 serve to illustrate, although it is possible with careful choice of series resistor to adjust α to have any value between about 1/2·5 and 2·5.

The transient behaviour of some of these circuits has been described in text-boks such as that by Hughes[3] but will be reviewed here for completeness.

The Case of a Linear Capacitance and a Non-linear Resistance

Let the resistance R be such that the relation $Ri = v^\alpha$ holds between the current and the potential across it.

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Then $E_o = q/C + (Ri)^{1/\alpha}$ at all times .... (1)

When $\alpha = 2$, this gives, by simple integration, the equation:

$$E/E_o = \frac{1}{(E_o t/RC) + 1} \quad \ldots\ldots\ldots\ldots \quad (2)$$

and

$$\frac{d(E/E_o)}{dt_{t=o}} = -E_o/RC \quad \ldots\ldots\ldots\ldots \quad (3)$$

Where t refers to time.

When $\alpha = \frac{1}{2}$, $E/E_o = (1-(t/2RCE_o^{\frac{1}{2}}))^2$ .... (4)

and

$$\frac{d(E/E_o)}{dt_{t=o}} = \frac{-1}{RCE_o^{\frac{1}{2}}} \quad \ldots\ldots\ldots\ldots \quad (5)$$

Equation (2), relevant to $\alpha = 2$, gives a curve which is initially steeper than the exponential case (for which $\alpha = 1$ of course), provided E_o is greater than unity but like the exponential it approaches zero asymtotically at time infinity. When $\alpha = \frac{1}{2}$ the decay curve as given by equation (4) is less steep initially than the exponential, provided E_o is greater than unity, but reaches a zero value at time $2CRE_o^{\frac{1}{2}}$ and has a zero gradient at that time.

the transients become more removed from exponentials.

Realization of the Non-linear Circuits

The case of $\alpha = 2$ for the non-linear resistor circuit needs little comment since this can be constructed by simply placing a non-linear silicon carbide unit in series with a conventional capacitor.

When $\alpha = \frac{1}{2}$, however, an operational amplifier circuit may be used as shown in Fig. 3.

The circuit is of the 'bootstrap' type and is arranged so that if v is the potential at the point A on one side of the linear resistor r then the other side B is held at $v - v^{\frac{1}{2}}$. The current through r is given by $v^{\frac{1}{2}}/r$ so that if r is chosen to be equal to R one has a resistor which obeys the relation $Ri = v^{\frac{1}{2}}$ as required.

Care must be taken to ensure that the loading of the first amplifier which produces $v^{\frac{1}{2}}$ by action of the non-linear silicon carbide network in the feedback loop, is kept small. This may be accomplished by the use of a suitably biased cathode-follower stage preceding the input resistor

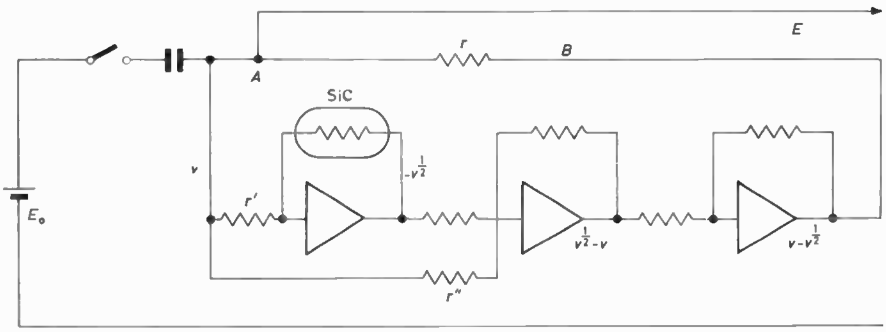

Fig. 3. Circuit for non-linear resistor $\alpha = \frac{1}{2}$

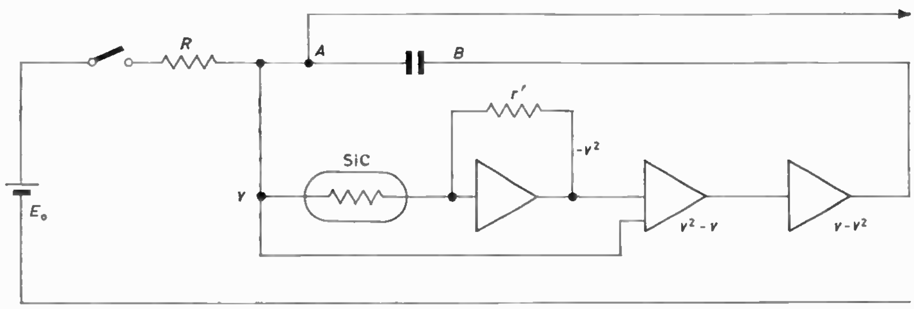

Fig. 4. Circuit for non-linear capacitor, $\alpha = 2$

THE CASE OF A LINEAR RESISTANCE AND A NON-LINEAR CAPACITANCE

Let the capacitance C be such that its charge-potential relation is given by $q = Cv^{\alpha}$

then $E_o = iR + (q/C)^{1/\alpha}$ .... (6)

When $\alpha = 2$, the solution is

$$\ln (E/E_o) - (E/E_o) + 1 + (t/2RCE_o) = 0 \quad \ldots\ldots \quad (7)$$

and

$$\frac{d(E/E_o)}{dt_{t=o}} = \frac{-1}{2RC((E_o/E)-1)E_o} = -\infty \quad \ldots\ldots \quad (8)$$

and when $\alpha = \frac{1}{2}$,

$$\ln \sqrt{\left[\frac{1-(1-E/E_o)^{\frac{1}{2}}}{1+(1-E/E_o)^{\frac{1}{2}}}\right]} + \sqrt{E_o t/RC} = 0 \quad \ldots\ldots \quad (9)$$

and

$$\frac{d(E/E_o)}{dt_{t=o}} = \frac{-2E(E_o - E)^{\frac{1}{2}}}{RCE_o} = 0 \quad \ldots\ldots\ldots \quad (10)$$

Equation (7) gives a decay which is very steep initially but reaches zero at time infinity whereas (9) starts with a zero slope, passes through a point of inflection and then eventually reaches zero at time infinity.

Other values of α allow curves which approach the exponential decay to be produced if α lies between $\frac{1}{2}$ and 1 or 1 and 2. However, if α is less than $\frac{1}{2}$ or more than 2

to the amplifier. The slight deviation from unity gain in the cathode-follower may be counteracted by a corresponding reduction of the resistor r', and a similar adjustment may be made to the value of r''.

To obtain the effect of a non-linear capacitor the positions of r and C are merely reversed as shown in Fig. 4.

Again, when A is at potential v to earth, B is at $v - v^2$. The capacitor is thus charged to v^2 and so carries a charge given by $q = Cv^2$. Equation (6) is therefore obtained. Actually, $E_o - iR$, that is $E_o - E$, is produced instead of E.

Interchange of the positions of the silicon carbide network and r' gives the case of $\alpha = \frac{1}{2}$.

As well as demonstrating some of the peculiar features of non-linear circuits, it would seem that the functions generated might find some application as forcing functions in analogue methods, particularly for solving vibration and deflexion problems in the field of atmospheric acoustics and blast wave propagation.

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An Electromagnetic Brake Activated by Eyebrow Muscles

By L. Vodovnik*, Dr. Ing.

After a discussion of the processes which are initiated when the driver decides to brake, the possibilities of shortening a part of the reaction time are considered. An experimental system with an electromagnetic brake activated by the action potentials of the eyebrow muscles is described. Some possible improvements and further applications (e.g. for amputees) are proposed.

(Voir page 727 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 734)

IT is a known fact that a number of road accidents are caused owing to the long reaction times of drivers. In order to see if there is a possibility to shorten that time it is necessary to consider the processes from the moment the driver perceives a dangerous situation to the moment he puts on the brake. A picture of the traffic situation is formed on the retina of the driver's eyes. Through the optical nerve the picture is transferred to the optical centre in the cortex. After some delay due to fear and other psychological parameters which vary individually, the driver's brain 'computes' the situation, and the output of such a computation is the command for a reaction which originates in the motor-centre of the brain. The order passes, encoded in nerve pulses, via the spinal cord and efferent nerve fibres to different muscle groups of the effectors: arms and legs. The resulting effects of such commands are normally turning the car to the left or to the right, speeding up, braking, etc. In very dangerous situations and at not too high speeds, the command from the brain is mostly braking. After such an order is given, it is desirable that the braking should be carried out as fast as possible. The time from the moment the driver decides to brake until the activation of the brake may be divided into the following periods:

(1) Transmission-time of the nerve pulses from the brain to the leg-muscles. At pulse-velocities of about 100m/sec, this transmission time may be estimated to be 0·15sec.

(2) Lifting-time of the foot from the accelerator pedal.

(3) Braking time needed to press on the brake pedal until the wheels lock.

The total delay time of the three periods T_1 has been found to be in the range of 0·4 to 0·5sec. This lag could be eliminated if there was a possibility to pick up the signal directly from the motor-centres in the brain and feed it (amplified) to an electromagnetic brake.

A compromise between a complete elimination of the time lag and a practical realizability of detecting signals near the brain centres, is given by the following solution. The foot-braking is essentially a conditioned reflex which has to be learned. Now, in principle, any other muscle group could be, after a reasonable learning period, conditioned to activity when the braking command comes from the brain. If this 'auxiliary' muscle group lies nearer to the brain, the first of the above mentioned periods may be shortened. The next point to consider in the selection of an appropriate muscle is its mass. Small muscles contract much faster than large ones. If the activity of such a muscle is related only to the brake, and not to the accelerator, the second period is eliminated and the third one much shortened. Assume for the present the 'auxiliary brake path' may take only two states: yes or no. If now an apparatus is built which enables the driver to use the auxiliary brake path in parallel with the normal one, two braking possibilities arise:

(1) Normal, analogue foot-braking with a delay T_1.

(2) Auxiliary, digital braking with a delay $T_2 < T_1$.

The second possibility is made use of in extremely critical situations where only strong and fast braking may prevent an accident. In this case the nerve pulses from the brain take the normal route via the spinal cord to the foot, but simultaneously they are sent to the small muscle which gets activated in this manner. As the usefulness of the auxiliary brake is limited to the period of T_1, the digital braking is limited to this period, after which the foot puts on the brake and normal braking may follow.

Description of an Experimental System

For the auxiliary muscle group the muscles near the eye (muscles orbiculary oculi and venter frontalis) were chosen. On the bars of a pair of spectacles, steel springs were fastened. At the other end of the springs, silver electrodes were mounted. When the driver puts on the spectacles the electrodes are softly pressed above his eyebrows. The wires from the electrodes are led to a transistor-amplifier built with Siemens silicon transistors type BCY17. The amplifier is a conventional differential amplifier with two cascaded input emitter-followers, three amplifying stages, and an output emitter-follower. The overall voltage gain is 10 000, input resistance 1MΩ, frequency range 1c/s to 10kc/s. The output of the amplifier is via a double-phase diode-rectifier connected to a monostable multivibrator with a quasistable time of about 0·5sec. In the collector circuit of the multivibrator there is a fast relay which activates a larger one. Near the brake pedal of a Fiat-750, a strong electromagnet was mounted which—when under voltage—pulls the pedal of the brake. The power consumption of this magnet is about 500W

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(12V 40A). The electromagnet is connected to the accumulator via the contacts of the larger relay.

The driver is trained to a short intensive blinking when a situation arises in which fast and strong braking is imperative. Simultaneously, the normal braking process is initiated as well. The action-potentials due to the short blinking of the eyebrow muscles (about 1mV) are amplified, rectified and trigger the monostable multivibrator, which activates the electromagnet for half a second. After this period normal or no braking is possible.

Discussion of the System and Possible Further Improvements

In the first experiments, special electrode jelly was used to render better contact between the electrodes and the skin. Later experiences showed that this was not necessary, and satisfactory contact during some hours was achieved without any special precautions. The spectacles with electrodes had no annoying effect on the driver. The threshold of the multivibrator was adjusted high enough that normal, unconscious blinking and even that due to sudden light at night, did not activate the brake. Measurements had shown that a learned blinking needs a reaction time (after a light-stimulus) of less than 0·1sec. After the same light-stimulus the learned foot-braking occurs with a delay of about 0·45sec. So, theoretically, a reduction in delay time of 0·35sec could be achieved. At a speed of 50km/h, an equivalent gain in length is about 5m. Such a distance may, in a large percentage of critical situations, prevent an accident. With the experimental system the theoretical limit has not yet been achieved.

The main reasons are time lags in the relay and the electromagnet. In future experiments the magnet should not be applied to the lever of the brake but directly to the wheels. Such a construction would reduce the time lag of the magnet to some tens of milliseconds. The relays should be replaced by transistor switches, and so this source of delays would be eliminated as well. In principle, the electrodes could be substituted by microswitches or strain gauges. The advantages of these elements have not yet been studied thoroughly enough, but it seems feasible that a strain-gauge activated brake could perform the same analogue response as the foot-brake. Its applicability could therefore be extended to amputees without the right leg.

The electromagnetic brake is in its early stage of development. But in general it is supposed that the main advantages of the electromagnetic brake will be proved in city-traffic, in crowded streets at speeds below or about 50km/h. Statistical data show that about 50 per cent of mortal accidents happen at speeds below 60km/h, and it is in this range that the proposed brake is intended to be used.

Acknowledgments

The author is indebted to his friend, Ing. J. Vida who, unselfishly and with great patience, 'risked' his car and himself for the first experiments. Thanks are due to the firm Elektromedicina in Ljubljana for its financial support, and to Ing. M. Sisakovic for the wiring and measurements of the electronic circuits.

A Three-Dimensional Radar Display

Many attempts have been made in recent years to deal with the problem of presenting to an air traffic controller a three-dimensional celestial view of a volume of airspace. A method devised by E.M.I. Electronics Ltd is claimed to suffer from none of their drawbacks and, in addition, offers easy and rapid control over the aspect from which the volume of airspace can be viewed by a seated observer.

It is assumed that the 'raw' radar information has been processed and that the spatial position information of all aircraft is stored in an electronic memory which is continuously being undated as fresh data become available. Such a memory forms an integral part of the computer which controls a beam positioning type of radar, but any other air traffic control system in which radar information is stored and not displayed simultaneously with being received can use this method of three-dimensional display.

The actual display utilizes a conventional, flat cathode-ray tube which may have a colour screen to enable air-aircraft characteristics and other features to be suitably distinguished, on the face of which appears a synthetic picture in the form of the faint outline of a perspective view of a transparent cube, representing the airspace under consideration. Within this cube the aircraft appear as luminous spots in their correct relative positions.

The principle of variable parallax is used to convey to an observer the impression of 'depth' as well as relative positions of aircraft, so the aspects of the cube is under control of the viewer who can rotate and tilt the cube as required. The observer can also introduce a number of graticules in any of the three orthogonal planes, which will enable him to make a quantitative assessment of a particular air traffic situation. The outlines of airways and geographical features can be inserted if and when required.

Different colours may be used to distinguish the various kinds of information, although the additional benefit which may be derived from the use of colour may not warrant the increased complexity and expense of the dislay unit.

The picture on the face of the display unit is generated in two distinctly separate operations. The beam first traces out the outlines of the cube, airways, geographical features and any graticules that may be required, as they would have to appear in the projected view after appropriate co-ordinate transformation, and subsequently paints in the positions of aircraft echoes. The latter are obtained by interrogating sequentially the main computer store containing the aircraft locations and, after converting the digitally stored information into analogue data, feeding the aircraft positions to the co-ordinate transforming circuits.

A picture regeneration rate of 20 per second can readily be achieved and the presentation can either use a 21in diameter. direct viewing cathode-ray tube, or a projection tube in conjunction with a 6ft square screen.

Since the pattern is generated synthetically, independent expansion or contraction of any of the three scales as well as a shift of the centre of the picture is a simple matter.

The latter facility enables a controller to single out temporarily any small sector within the airspace he is controlling. Similarly, the controller would not display all aircraft within his zone, but only select those that are of immediate concern to him.

In a semi- or fully-automatic air traffic control system the computer would make the selection and, if necessary, override the decision of the controller. The spots in the display representing aircraft can be labelled with basic 'tags'. A 'tail', which can be painted in electronically by storing the aircraft's previous positions, will indicate to an observer directly direction and speed, and the size of the spot can be made representative of the size of aircraft. Intermittent flashing can be used as a means of keeping visual track of one or two specific aircraft and, if colour is available, this facility can be used to distinguish between civil, military and other categories of aircraft. A complete set of detailed alphanumeric data pertaining to one aircraft can be displayed in each of the four corners of the screen, with electrically generated arrow lines pointing to the aircraft concerned.

Microelectronics in Equipment Design

(Part 2)

By S. S. Forte*, Ph.D., B.Sc., A.M.I.E.E., M.I.E.E.E.

(Voir page 650 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 657)

IN the previous issue the general philosophy of micro-miniaturization was discussed, and a practical example of the use of solid circuits cited in the form of details of an experimental digital computer. It is now proposed to describe some of the problems encountered in the design of an experimental 75Mc/s airborne marker receiver, the building of which was carried out as an exercise in the application of microelectronic principles to linear (as distinct from digital) circuits. Perhaps at the onset it may be useful for those readers who are not particularly familiar with aviation electronics if a short description of the functions of an airborne marker receiver is given.

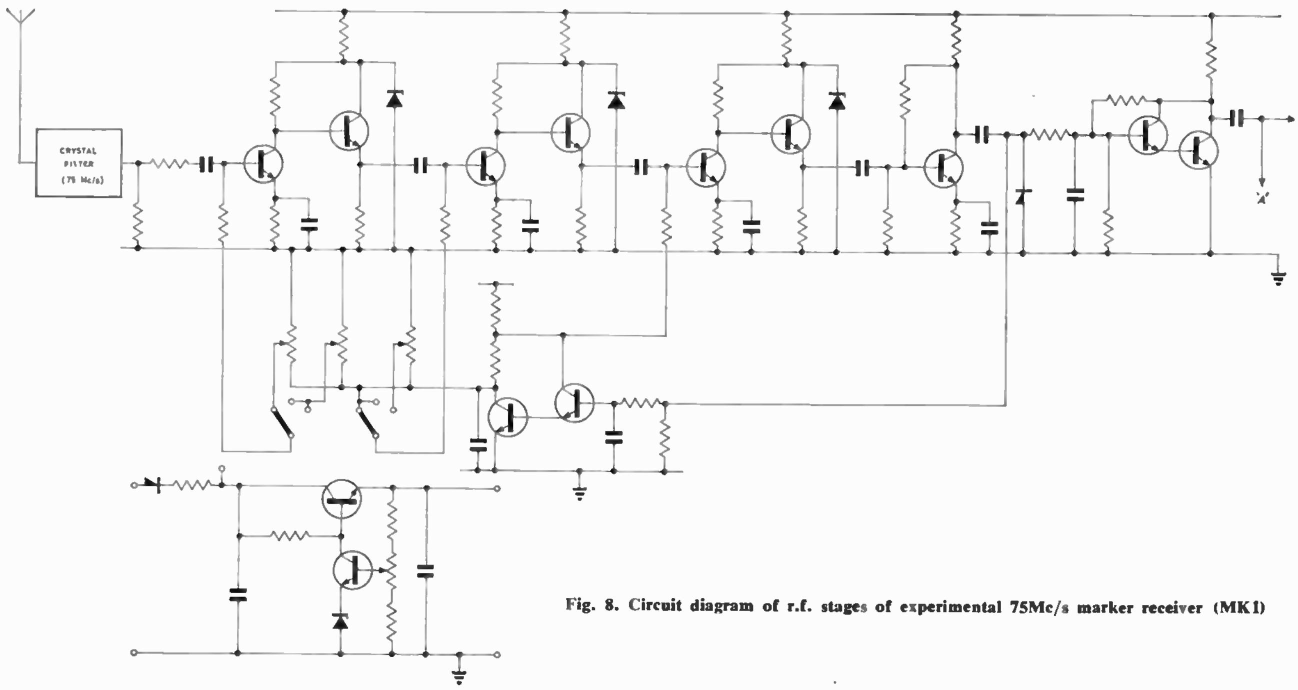


Fig. 8. Circuit diagram of r.f. stages of experimental 75Mc/s marker receiver (MK1)

A marker receiver forms part of the internationally agreed instrument landing system (i.l.s.) used on the majority of aircraft in airline service. In brief, the system provides the pilot with elevation, azimuth and distance information during the approach phase to the runway, the marker receiver's function being to provide the distance information. This is achieved by receiving signals from radio beacons sited along the runway axis at pre-determined distances from the landing threshold.

The detailed specifications for this equipment are laid down by the Radio Technical Commission for Aeronautics (RTCA) so that the exact function of this receiver is well known.

There may be up to three marker beacons, all radiating a fan beam across the approach line at 75Mc/s with horizontal polarization. The inner marker is sited 250ft from the landing threshold and is amplitude modulated at 3kc/s to a depth of 95 per cent. The modulation is keyed at a rate of six dots per second continuously. The middle marker is 3 500ft away from the landing threshold and is amplitude modulated with 1 300c/s. This beacon is also keyed, the keying consisting of a series of continuous dots and dashes, the former at the rate of six per second and the latter at the rate of two per second.

The outer marker is situated at a distance of 3·9 nautical miles and is amplitude modulated with 400c/s, keyed at a rate of two dashes per second continuously.

The receiver in the aircraft must light a different coloured lamp as the aircraft passes over each beacon. The receiver is also used to indicate the position of 'en route' markers. These are generally not keyed, and are modulated at 3kc/s. Since the aircraft would be at its cruising height, the sensitivity of the receiver has to be greater than is necessary for its i.l.s. mode of operation. This places a requirement on the equipment to have three pre-set sensitivities. The 'High' mode requires an r.f. input level of 100 to 200μV to give lamp-operate threshold. The 'Medium' mode requires a lamp-operate threshold between the level selected for the 'High' mode and 5mV. Similarly the 'Low' mode requires the threshold to be between the 'Medium' sensitivity level and 200mV.

The automatic gain control must be such that the audio frequency variation is less than 3 : 1 when the r.f. input is increased from threshold to 200mV.

The audio output must be 100mV into 100Ω or 500Ω, and the receiver has to be operated from the aircraft 28V d.c. supply. The equipment is to be capable of operation at an altitude of 60 000ft and in a temperature range of −25°C

* *The Marconi Co. Ltd.*

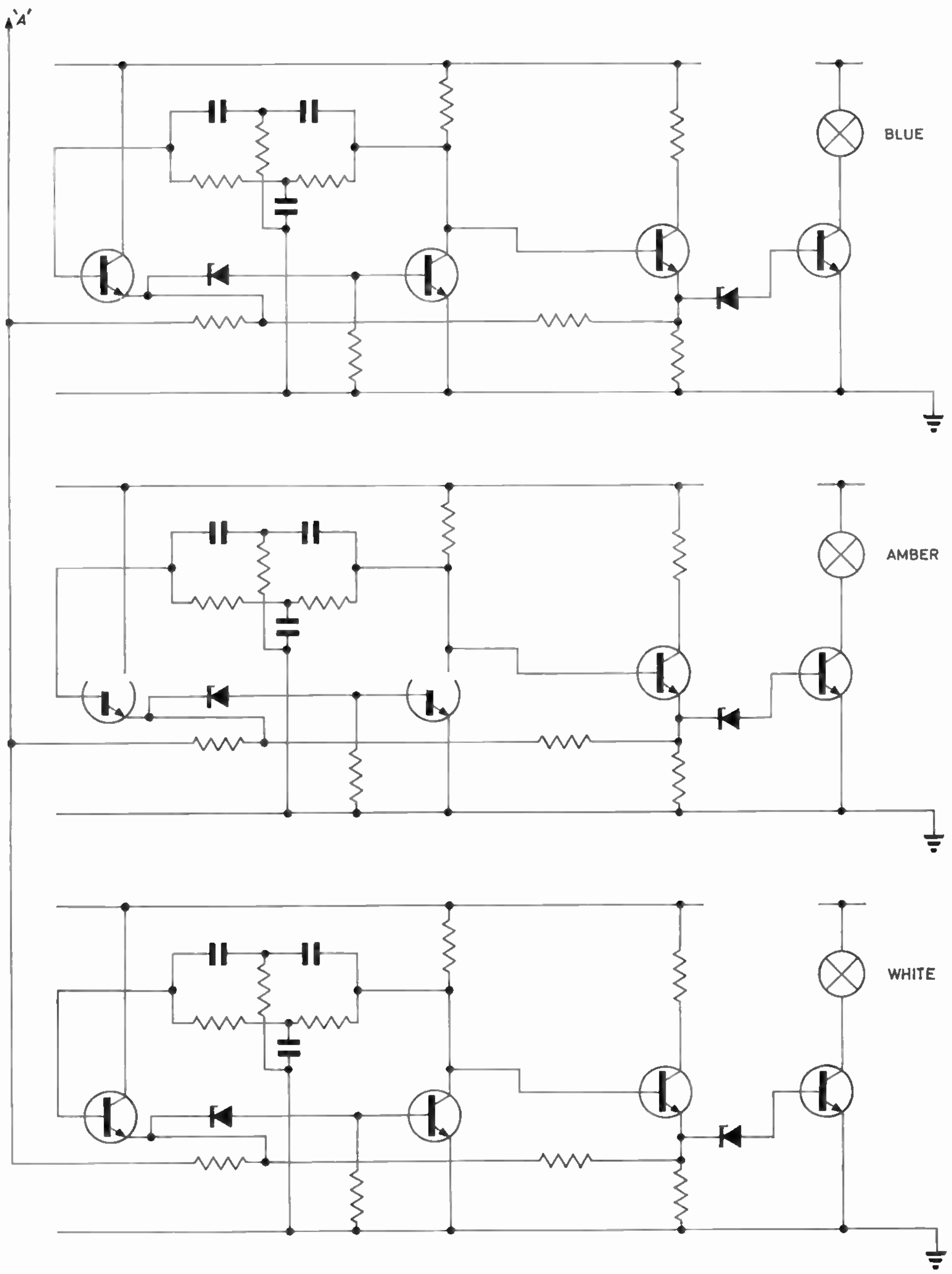


Fig. 9. Circuit diagram of a.f. stages experimental 75Mc/s marker receiver (MK1)

to +80C° after one minute warm up in free air. It must moreover survive without deterioration from −55°C (power off) to +100°C (power on).

These then are the specifications which the new design of marker receiver would have to meet at the outset. Because of the design objective of maximum reliability, and the constraints imposed by the use of microelectronic techniques, the new receiver differs widely from conventional transistorized receivers now entering airline service.

Inevitably in the origination of a new design using new techniques, where experience is necessarily limited, a second look at the problem will reveal a better approach. This was so in the equipment under discussion, for after completion of the original project a second attack was made which resulted in a considerable improvement upon the original. Both phases (hereinafter designated Mark I and Mark II) are described below as illustrating the various stages of experimental work involved.

The Mark I Receiver

The first decision which had to be made was as to the type of receiver—straight or superheterodyne— to employ. In this, the availability of silicon planar transistors with f_T of the order of 1 000Mc/s decided the issue in favour of a straight receiver, as these components offer worthwhile stage gains at the operating frequency of 75Mc/s, and so enable a straight receiver to be used, with the required selectivity derived from a crystal filter. Such a design confers a particular advantage in that it eliminates tuned circuits, which are incompatible with the microelectronic techniques it was proposed to use, and moreover eliminates all variables from the r.f. section, thereby obviating the risk of maladjustment during the operational life of the receiver.

In the equipment under discussion, after demodulation of the carrier the audio is analysed in a conventional manner using twin-T frequency selective *RC* networks in a feedback amplifier to give filter peaks at 400, 1 300 and 3 000c/s respectively. The a.c. output then drives directly a class-C lamp-drive stage.

In the interest of standardization, with its attendant advantages, the number of microcircuit types was kept to a minimum, the three r.f. amplifiers being identical, as also were the three audio filter amplifiers. The a.g.c. amplifier, output stages and stabilizer stage remained as 'one-off' circuits.

Fig. 8 serves to illustrate the more detailed circuit description which follows.

The majority of the gain at 75Mc/s is derived from three identical amplifiers whose response is peaked at approximately the correct frequency by a suitable choice of emitter decoupling capacitors and interstage coupling capacitors. Each stage consists basically of a grounded emitter amplifier coupled to the next module by an emitter-follower for impedance matching purposes. The supply is stabilized and decoupled by a Zener diode incorporated in each stage, and the resulting gain is 20dB per stage. The final r.f. amplifier is a grounded emitter amplifier fed from the +14V rail to give greater signal handling capacity. This feeds a shunt detector diode. The detector output is then decoupled to the class-A audio amplifier; this amplifier thus has its working current increased as the audio signal increases, and will thus take maximum standing current only at full output signal. The d.c. is also fed to an a.g.c. amplifier using a Darlington pair to give a high current gain and to permit use of a high input resistance, thereby resulting in negligible loading of the detector and reducing the capacitor controlling the a.g.c. time-constant to an acceptable value. The delay voltage for the a.g.c. is provided by the sum of the base emitter voltages of the two transistors in the amplifier. The output is taken from the collector to the base voltage rail of the three broadband stages via three gain setting potentiometers to provide control in the usual manner. The whole of the circuit just described is mounted on a double-sided printed circuit board using plated-through hole techniques.

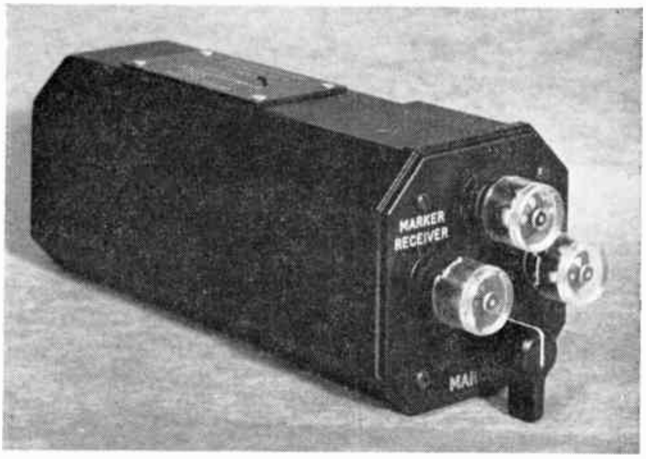
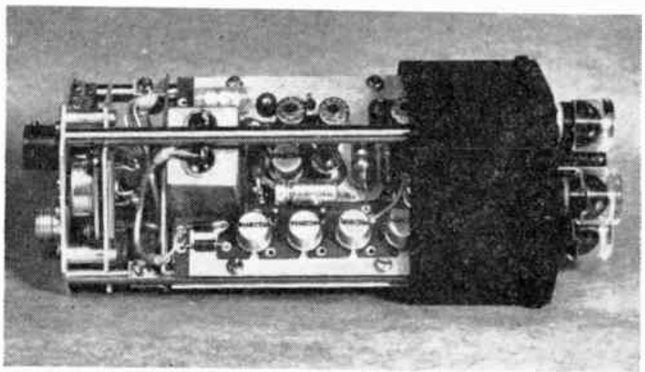


Fig. 10 (above). Completely assembled 75Mc/s marker receiver (MK1)

Fig. 11 (below). Completely assembled 75Mc/s marker receiver (MK1) with cover removed



At this stage, the audio frequency output may be applied either to head telephones or to three audio frequency filters resulting in the conventional three lamp visual display. The audio circuit which is mounted on a second printed circuit board is shown in Fig. 9. Three filter amplifiers accept the audio signal and are identical except for the rejection frequency of the twin-T filters. These are of normal design using conventional close toleranced components, such components not being possible in the microelectronic form with the present state of the art.

Each amplifier is d.c. coupled with an emitter-follower output stage, and another emitter-follower terminating the *RC* networks. Overall feedback is then applied to give unity gain at the filter rejection frequency. At frequencies remote from the reject point of the filter, the gain of the grounded emitter stage itself will be low, thus causing the overall gain to fall below unity and thereby resulting in maximum gain at the null point of the twin-T. Signal is taken from the output emitter-follower and is d.c. coupled by a Zener diode to a lamp-drive stage so that, in the absence of signals, or in the presence of small signals at the output of the filter amplifier, this stage is non-conducting; but when an audio signal, large enough to exceed the Zener voltage, appears at the output, current pulses will flow in the base, causing output pulses to flow in the appropriate lamp, thus giving the required indication.

Fig. 10 shows the completely assembled marker receiver, which, as is seen, has been mounted entirely in an instrument case fitting behind the lamps in the instrument panel, thereby achieving a 4 to 1 reduction in both volume and weight compared with the present equipment.

Fig. 11 shows the receiver with the cover removed, and the r.f. board with its microcircuits in T05 encapsulation can be seen.

The performance of the prototype receiver conforms to the specification, and preliminary calculations have shown a predicted mean time between failures (m.t.b.f.) of 34 000 hours. This compares with 10 000 hours m.t.b.f. for the latest conventional transistorized equipment now entering airline service.

The Mark II Receiver

The Mark I receiver was very satisfactory as far as it went, but it seen became clear that while the original concept of a straight receiver and crystal filter was technically sound, further improvement in terms of reliability was possible, and accordingly the Mark II marker receiver took shape. In this the detailed circuit design has been changed almost entirely in order to simplify the microcircuit construction by reducing the number of components per module; at the same time circuits were designed which would accept wider toleranced elements. The modified circuits are illustrated in Figs. 12 and 13.

The r.f. amplifier module was the first one to be

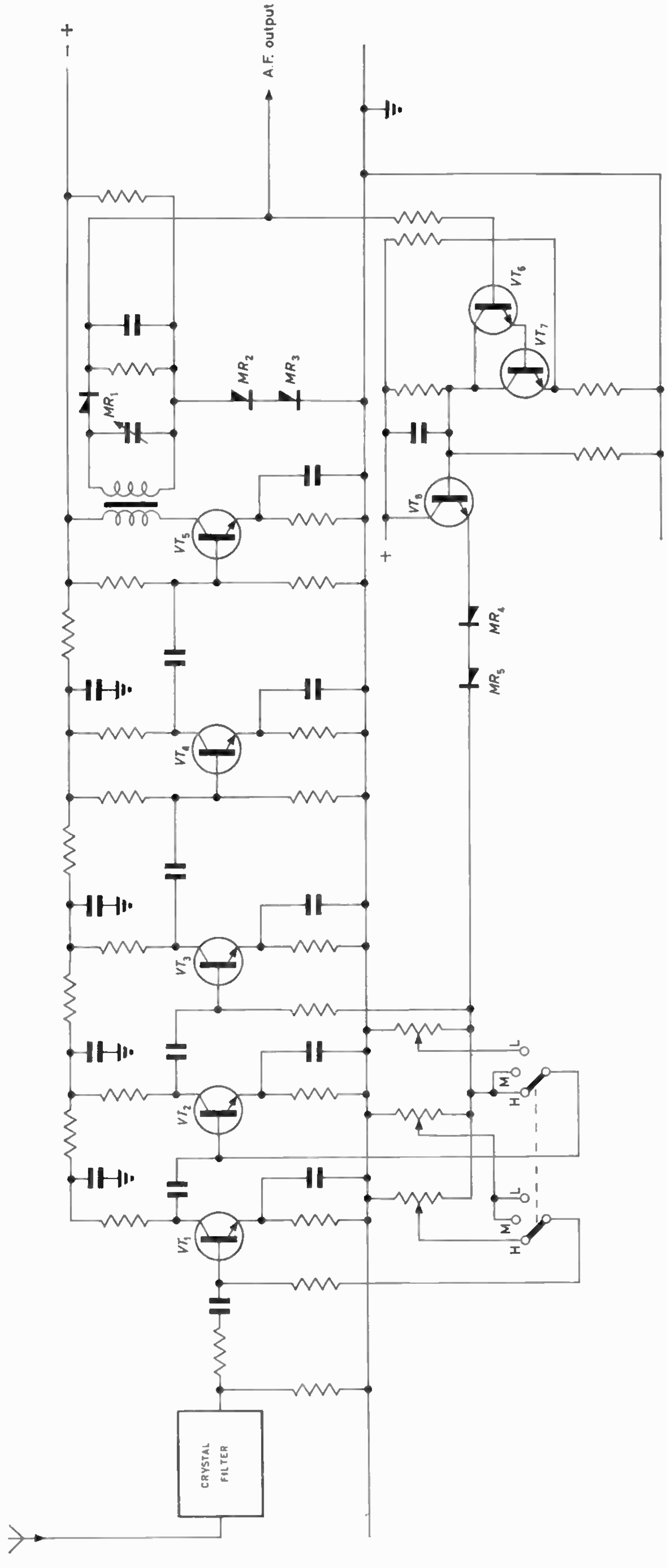


Fig. 12. Modified r.f. circuits used in MKII receiver

modified. This was changed to a single transistor amplifier VT_1, since the earlier configuration of two transistors, a Zener diode, two capacitors and five resistors was found somewhat to overload the mounting capacity of an eight-pin T05 header. The single transistor circuit was therefore adopted, without the Zener stabilization, since the d.c. stability of the supply rail is of secondary importance when compared with the r.f. decoupling. The loss of gain due to removal of the impedance matching emitter-follower is approximately 7dB per stage, requiring an adidtional r.f. stage to maintain the gain at the original level.

An additional advantage resulting from the reduction of the number of components per header is that more points of the internal structure may be brought out to pins, making modifications and monitoring simpler. To illustrate this point, the five modules containing VT_1 to VT_5 are identical; however, VT_4 and VT_5 have an external base-to-supply-rail resistor connected to them, giving a fixed bias, and VT_5 has a tuned circuit load instead of a resistor. This gives three circuit configurations with the same module. Furthermore, by making the base and emitter capacitors externally accessible, these may be replaced by any desired external capacitor in order to modify the frequency response of the module, thereby approaching the desired objective of a small number of more or less universal building-blocks.

A d.c. amplifier using a Darlington pair is coupled to the detector load as before, the load itself sitting on a pedestal of two diodes biased to conduction so as to compensate for the two V_{be}'s of the pair of transistors. Any temperature effects caused by changes in V_{be} with temperature will be compensated since VT_6, VT_7 and the diodes MR_2 and MR_3 are mounted in the same header; the a.g.c. voltage is introduced at the emitter of VT_7 and is thus independent of temperature. The output of the d.c. amplifier is taken to the a.g.c. rail via an emitter-follower VT_8 to avoid loading VT_7, and is then applied to the r.f. modules as before.

It was found in the course of the design that the use of two diodes and a Darlington pair in a single header results in another very useful standard module, one great advantage being that high input resistances are obtainable, even at 100°C, which provides isolation between circuit elements, thus simplifying design. Five such modules are in fact used in the final receiver.

The a.f. output is taken from the detector to an a.c. coupled amplifier, another departure from the sliding bias arrangement used in the Mark I design. A class-B complementary symmetry stage is used (VT_{13}, VT_{14}) which, by

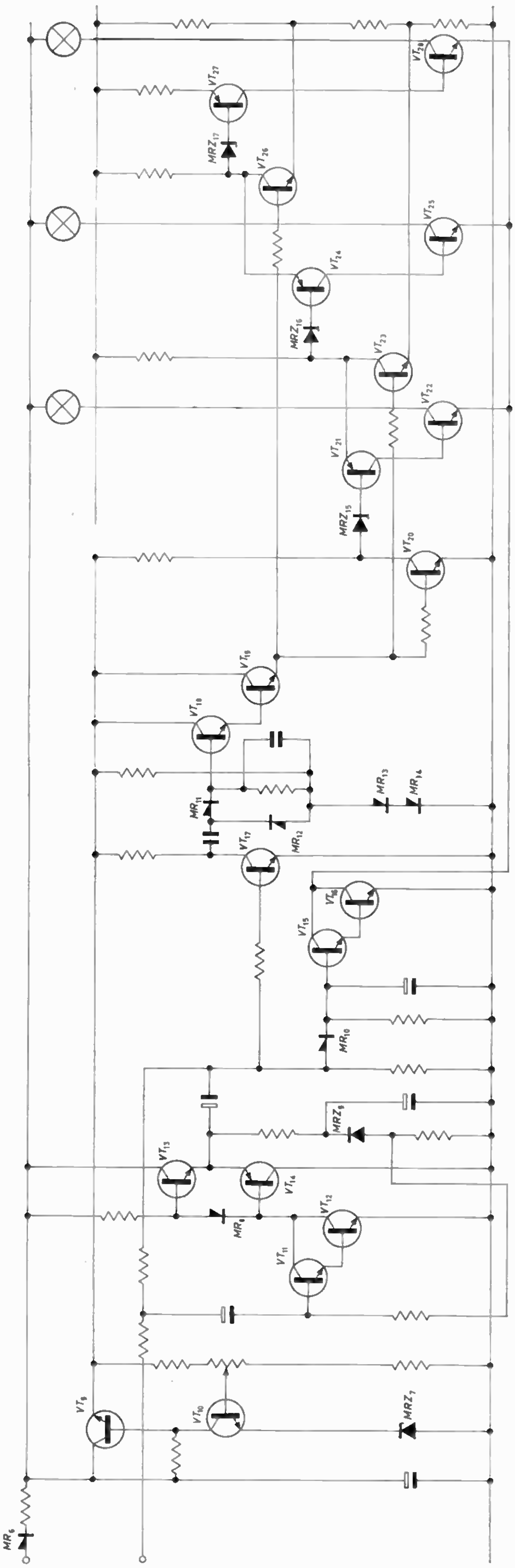


Fig. 13. Modified a.f. circuits used in MKII receiver

action of the back-to-back emitter-follower, gives low output impedance and low distortion once the input knee voltage has been exceeded. The standing current at no signal is almost entirely that of the driver stage VT_{11}, VT_{12}. Crossover distortion, due to the 'dead' voltage of the sum of the two base-emitter voltages of VT_{13} and VT_{14}, is minimized by placing a diode in conduction from base to base as part of the driver load. Overall feedback is also applied, reducing the overall a.f. voltage gain to 5. The resultant distortion at both the 100mW and 1mW levels is negligible. The d.c. stability of the output stage is controlled by a Zener diode MRZ_9 which maintains the output at a fixed working point, by means of d.c. feedback to VT_{11}, independently of temperature and device tolerances.

The most marked departure from the Mark I receiver design takes place in the a.f. analysing section. In the first model the selectivity was derived from twin-T networks using conventional components. This had two major drawbacks—firstly the component density in the feedback amplifier module was rather high for ease of construction, and secondly the circuit made use of six close-tolerance components per channel giving a total of eighteen bulky conventional components of close tolerance and doubtful reliability. The solution adopted in the second model employs a d.c. logic system. In this case any frequency within a specified band, rather than a spot frequency, will operate the indicators. This still meets the specifications laid down by the RTCA. The overall system is made up of a limiter VT_{17} driven from the output stage, which in turn drives a diode pump MR_{11}, MR_{12} giving a d.c. output which increases with frequency. This circuit, as in the case of the detector, stands on a pedestal of two diodes and is read out by a Darlington pair VT_{18}, VT_{19}, to give a low source resistance output with the voltage independent of temperature. The d.c. voltage from the emitter-follower is applied to the bases of three transistors VT_{20}, VT_{23}, VT_{26} (in three identical modules), whose emitters are taken to three fixed reference voltages. The emitter of VT_{20} is in fact taken to earth, the reference voltage being set by the input knee voltage. Thus when V_{in} from the diode pump exceeds this voltage, VT_{20} will conduct, causing its collector to fall to the saturation voltage; in turn, current will flow through the Zener diode MRZ_{15} causing VT_{21} and VT_{22} to conduct. An indicator LP_3 forming the collector load of VT_{22} will thus be turned on.

The emitter of VT_{23} is taken to a potential V', positive with respect to earth. When V_{in} exceeds this level, the transistor will conduct; in so doing the collector voltage will fall to a value equal to V' plus the saturation voltage. At this point VT_{24} and VT_{25} will conduct, turning on the indicator LP_2 which forms the collector load of VT_{25}, as before. However, in this case, the base current in VT_{21} will cease, since the voltage between the collector of VT_{20} and VT_{23} is insufficient to break down the Zener diode MRZ_{15}. The current in VT_{21} and VT_{22} is thus cut off, thereby extinguishing LP_3.

Similarly, the emitter of VT_{26} is taken to a higher potential V'' and when V_{in} exceeds this level the indicator LP_1 will be turned on as LP_2 is extinguished.

All that is required therefore is to position the points of change-over between the desired audio tones on the voltage/frequency characteristic. The tolerance on this is much less critical than would be the case when a twin-T network was used for each frequency.

The number of close toleranced conventional components in the audio circuits falls from eighteen in the first model to only five in the new design. One capacitor and one resistor are used in the diode pump, and the remaining three components are the resistors which form the potential divider for the determination of the voltages V' and V'' and of the corresponding cross-over frequencies.

An additional circuit was added as a safeguard against a false lamp indication caused by low signal level. This is the Darlington pair VT_{15}, VT_{16} which is turned on only when sufficient a.f. (detected by MR_{10}) is present for the correct operation of the limiter VT_{17}. The emitters of the lamp-drive transistors VT_{22}, VT_{25}, VT_{28} are returned through the collector of VT_{16} which saturates when the signal is of adequate amplitude, thus enabling the lamp-drive modules to operate.

The result of this second design is that the number of close tolerance components in the new receiver has been reduced by 70 per cent. Some external, conventional components have been left or indeed added in the interest of standardization of the modules. Thus the effect on reliability when compared with the Mark I microminiaturized receiver is only expected to be of second order, except for the fact that the new modules are less complex and therefore likely to be more reliable. There will, however, be a major reduction in cost because of the overall simplification of the design and of the standardization of modules.

The Mark II marker receiver is now in the latter stages of development. The new system has been thoroughly evaluated in breadboard form, and has successfully undergone the required environmental tests. The solid circuit modules have successfully passed individual functional tests and the new model of the receiver is being assembled and tested.

The two applications described above illustrate that microelectronics can be successfully applied to linear or analogue equipment as well as to the more commonly accepted digital systems. The techniques at present available in the solid circuit art do, it is true, impose severe constraints on the circuit design, particularly in the case of analogue applications. By adaptation of the design to suit the techniques available, and by a very close collaboration between circuit designers and semiconductor manufacturers, these constraints need not be intolerable as, it is hoped, this article has demonstrated.

There is not much doubt that one of the outstanding problems existing at the moment in the application of microelectronics to equipment design is the problem of interconnexions. Several solutions have been suggested—double sided printed wiring boards with plated-through holes, multi-layered printed wiring boards, preformed welded wiring, etc.—but none meets all the desired criteria. No ideal solution can be said to exist at this time, and until such a solution emerges, the ultimate capabilities in the reliability and in the packaging of equipments using microelectronic techniques will not have been attained.

As the technology improves, so new materials will become available, and so tolerances will be held more tightly to make possible the use of improved passive 'components', thereby reducing some of the current limitations. Moreover, yields of acceptable components will undoubtedly rise markedly so that solid circuit costs will fall, perhaps not quite as spectacularly as transistors have done in the past few years, but nevertheless quite appreciably.

The age of microelectronics is undoubtedly dawning—indeed it is with us now! The day is not far removed when major portions of electronic systems and equipment, spreading progressively from the space and missile applications, via the services requirements to the general run of radio and communication equipment, will be produced using microelectronic techniques with resultant spectacular improvements in reliability and easing of the maintenance problem.

Acknowledgments

The author would like to record his gratitude to his colleagues in the various Divisions of the Company for their assistance and co-operation, and to the Director of Engineering & Research, The Marconi Company, for permission to publish this article.

A Radio Frequency Anechoic Chamber

A radio frequency anechoic chamber has been developed by The Wayne Kerr Laboratories Ltd under contract to the Ministry of Aviation in relation to work being carried out on the problem of the radio frequency radiation hazard to health. In its initial form it is just over 20ft in length and about 9ft by 9ft in width and height. Transverse baffles situated along the walls, floors and ceiling reduce the height and width at places along the length of the chamber to a few feet less than the quoted values. The anechoic material consists of blocks of rigid foam absorber which are mounted in a semi-permanent fashion to form the internal surface of the chamber. The chamber can accommodate radiation frequencies from 250 to 50 000Mc/s.

The term semi-permanent is used as the blocks are secured in such a manner that the internal aspect of the room may be altered very quickly. Also the chamber is readily available for extension in length by the addition of extra blocks. Since the absorption properties of the anechoic material depend to a certain degree upon the angle of incidence of the radiation, transverse baffles are used so as to present a surface giving nearer normal angle of incidence. Even so, a chamber having transverse baffles structures may only be necessary, for instance, in the case of measurement of scattering cross section or for aerial pattern measurement where even the smallest undesired signals should be reduced to a minimum. A simpler smooth walled room would suffice for many other purposes.

The Wayne Kerr chamber is designed primarily to contain radiating systems producing accurately known fields of intensities up to 20mW/cm^3 (about 280V/m). The systems being developed in the first instance are for S and X bands and cover the frequency ranges 2 500 to 4 000Mc/s and 8 500 to 10 400Mc/s. Briefly, these consist of high gain pyramidal horn radiators with waveguide feeds monitored for power and frequency fed by M type backward wave oscillators (bwo's). The length of the chamber has been chosen with respect to the minimum distance requirements relating to far field or Fraunhofer conditions. Eventually the intention is to expand the 'standard field' work over a frequency range extending from 400 to 40 000Mc/s. Generally speaking, the power outputs of available high power c.w. oscillators decrease with increase in frequency and resource may therefore have to be made to alternative techniques of producing the required high intensity fields. Such a one would be to use a focusing technique whereby the Fraunhofer field pattern is produced in the near field region. Again at the lower frequencies, where high level power sources are available, aerial gains are inherently lower and hence the fields may have to be supported by TEM wave guiding structures rather than by radiating them. It is for these alternative techniques that it may become necessary to re-align the anechoic material of the chamber.

For the weapon electromagnetic radiation hazard problem, irradiation of weapons or their component parts can be carried out under controlled laboratory conditions. In addition to checking on the radio frequency vulnerability of fusing circuits, the various hazard detecting systems that have been developed to monitor fuse circuits for induced radio frequency currents can themselves be quickly assessed and if need be corrected so as to give reliable and accurate performance in their eventual fields of operation.

The Input Resistance of Transistor Feedback Amplifiers

By E. de Boer*, Ph.D.

This article describes a general theory of negative feedback, which is concentrated upon the impedance levels in the circuit rather than on the amplification. Simple as well as complex feedback systems can be treated and useful insight can be gained almost by inspection. The rules for applying the theory are simple and logical. When feedback is taken from or fed to a point in shunt, the point is to be considered as having infinite internal resistance. When, alternatively, feedback necessitates the opening of a loop, the loop should be considered as having zero internal resistance. Any inherent internal resistances of such points are to be absorbed into the sourceor the load. A number of illustrative examples, including the design of a line amplifier designed to work between prescribed impedance levels, serves to show the salient points of the theory.

(Voir page 727 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 734)

IN a previous article[1] it was shown how the internal resistance of a feedback amplifier can most easily be computed with the help of two unconventional theorems. For complicated systems it is often possible to simplify the circuit to such an extent that the theorems can be applied directly and that, consequently, the internal resistance can be obtained almost by inspection.

For valve amplifiers this technique is usually sufficient, since only the quantities gain and internal impedance are required. Transistor amplifiers are different, however. Especially in interstage design, it is required to know the input resistance of a succeeding stage, and this quantity too is liable to change with the application of negative feedback. In this article the method referred to above will be extended to the input impedance of feedback amplifiers.

To recall the trend of the preceding article, negative voltage feedback is to be treated by regarding the amplifier as a current source, and current feedback with an amplifier acting as a voltage source. The current source acquires a shunt resistor, and the voltage source a series resistor, the values of which depend on the amplifier and the feedback circuit. The load impedance has nothing to do with this, provided that any inherent output impedance of the original amplifier (without feedback) be absorbed in it. The feedback factor depends only upon the ratio of the internal impedance (as a resultant of the feedback) and the load impedance (including the inherent output impedance).

To simplify matters all dependence on frequency will be neglected, hence the concept of resistance will be used throughout. As in the formerly treated case attention will be focused on amplifiers behaving either as a short-circuit or as an open-circuit, this time with respect to the driving source. The former type will be called current-accepting, it is assumed to require only current for driving, and it develops no voltage across the input. It is an idealized form of the common-base transistor amplifier. The second type, like a valve, accepts only an input voltage, and no input current is flowing. Which of these two is chosen depends entirely and solely upon the form of the feedback network. Two types of negative feedback to the input are possible: shunt feedback, where a feedback current is applied in parallel to the input, and series feedback, where a feedback voltage is placed in series with the input. It is easily seen that series feedback tends to increase, and shunt feedback to decrease, input resistance.

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Series Feedback

Consider an amplifier with zero input resistance. It is of the current-accepting type, and is driven by a current i. The amplifier operates on this current, and developes at some point a voltage or current that is the required output.

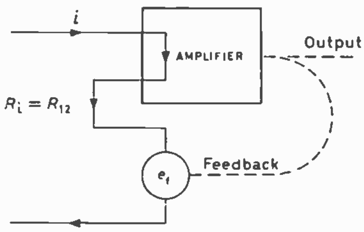


Fig. 1. Amplifier with series feedback to the input (idealized)

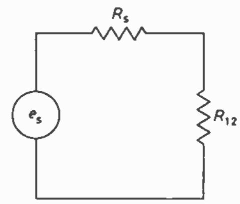


Fig. 2. Source circuit

From this point backward, negative feedback is applied, e.g. of the pure series form. With respect to the input, the feedback really amounts to generating some feedback voltage e_f that is to be subtracted from the input voltage; see Fig. 1. This feedback voltage is proportional to i, at least for a linear amplifier:

$$e_f = i \,.\, R_{12} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

The factor R_{12} is a transfer conductance which describes the feedback situation. Any other effects of the feedback network will be neglected.

The total input voltage, developing when the current i flows, is just e_f, since the actual amplifier requires no input voltage. As far as the driving source is concerned, the input impedance is then R_{12} with feedback, and zero without it. This constitutes the first important result. In order to obtain an expression for the feedback factor, i.e. the reduction of amplification due to the feedback, one has to go one step further. If the amplifier was driven by a current source, there would be no reduction of amplification by the feedback at all. Without feedback a voltage-source drive is not permitted. Of necessity one must consider the amplifier to be driven by a voltage source (e_s) with an internal resistance R_s (Fig. 2). Without feedback the current i is:

$$i = e_s / R_s$$

With feedback the current changes to i':

$$i' = e_s / (R_s + R_{12})$$

The feedback factor i/i' is thus found to be:

$$f = i/i' = 1 + (R_{12}/R_s) \quad \ldots\ldots\ldots\ldots \quad (2)$$

That this is also the factor by which frequency dependence and non-linear distortion are reduced, is well known. The result expressed by equation (2) shows how the feedback factor depends exclusively on the ratio of input resistance to source resistance. Large amounts of feedback are only realizable when these two resistances are highly dissimilar. Power matching, as for example required to minimize noise, is not compatible with large single-loop negative feedback on either side of an interstage network.

These results can be formulated in the following theorems.

Theorem A. A pure current-accepting amplifier, which is provided with series-type negative feedback described

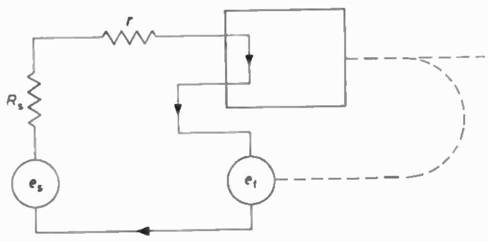


Fig. 3. Realistic representation of feedback amplifier

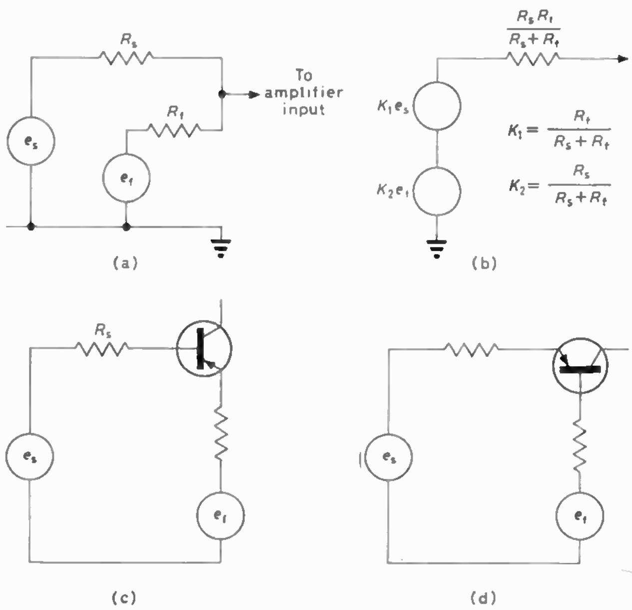


Fig. 4. Various forms of the feedback circuit

by a transfer resistance R_{12}, acquires an input impedance equal to R_{12} by the negative feedback.

Theorem B. If this amplifier is driven by a voltage or current source with an internal resistance R_s, the feedback factor is unity plus the ratio of R_{12} and R_s.

A real amplifier will have a non-zero input impedance, and hence will not behave as required in theorem A. As Fig. 3 shows, the input resistance r (without feedback) appears as in series with the source. It can therefore most conveniently be absorbed into the source's internal resistance. Formula (2) then becomes:

$$f = 1 + \frac{R_{12}}{R_s + r} \quad \ldots\ldots\ldots\ldots\ldots \text{(2a)}$$

This shows that the inherent input resistance r of the amplifier should be small, (preferably smaller than R_s), in order to attain a large feedback factor.

A second amendment concerns the feedback network itself. It is usually not possible to design this so that a pure voltage source is created that acts in series with the input. Again the real situation can be described by including a resistor R_f, as Fig. 4(a) shows. By using Thevenin's theorem this situation can be represented as in Fig. 4(b). A very common method of applying series feedback is illustrated by Fig. 4(c). It has the disadvantage that the input resistance is rather high. Much more in line with the theorem's requirements is the little-used configuration depicted in Fig. 4(d).

Shunt Feedback

The second type of feedback to the input is shunt feedback: the feedback current is applied in parallel with the input. By way of the duality principle the theorems that apply in this case can be written down immediately.

Consider an amplifier with infinite input resistance, i.e. an amplifier of the voltage-accepting type. The feedback is applied in the form of a current i_f:

$$i_f = e \,.\, G_{12} \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(3)}$$

Here e is the input voltage and G_{12} the transfer conductance of the complete feedback loop. See Fig. 5. Since the input of the amplifier originally had zero conductance, all of the resulting input conductance results from the

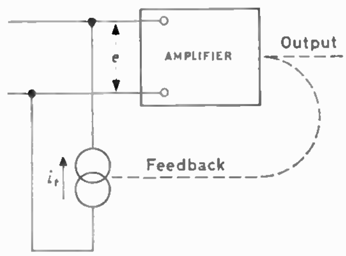


Fig. 5. Amplifier with shunt feedback (idealized)

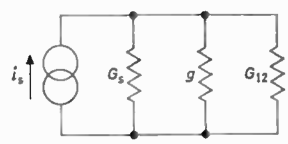


Fig. 6. Equivalent input circuit

feedback current i_f. The input conductance of the feedback amplifier will thus be equal to G_{12}. Hence:

Theorem C. A pure voltage-accepting amplifier, which is equipped with a shunt-type negative feedback described by a transfer conductance G_{12}, acquires an input conductance equal to G_{12} as a result of the feedback.

To derive a form for the feedback factor it is necessary to consider the character of the source somewhat further. Assuming the driving source to be described by a current source with an internal conductance G_s the following form can be derived for the feedback factor:

$$f = e/e' = 1 + (G_{12}/G_s) \quad \ldots\ldots\ldots\ldots \text{(4)}$$

This result is of the same form as equation (2), so that the same general remarks apply. In this way the following theorem is found:

Theorem D. If the pure voltage-accepting amplifier, equipped with shunt feedback, is driven by a current source with an internal conductance G_s, the feedback factor is unity plus the ratio of the effective input conductance G_{12} and the source conductance G_s.

The inherent input conductance of a real amplifier, as well as the conductance contributed by the feedback network, can be absorbed into the source conductance G_s. If the latter is changed into $G_s + g$, (Fig. 6) a more general expression is obtained:

$$f = 1 + \frac{G_{12}}{G_s + g} \quad \ldots\ldots\ldots\ldots\ldots \text{(4a)}$$

Again it is found that a small value of the intrinsic conductance g is needed to attain a large feedback factor.

Hence, in order to apply shunt feedback, the original amplifier should have as large an input resistance as possible.

These four theorems serve to treat any real problem, provided the feedback network is of a suitably simple form. More complicated feedback systems, in which the various feedback loops completely encircle one another, can be treated equally well after some simple manipulations. Inversely, one can easily design an amplifier that is required to have a specific input resistance. If this is smaller than the intrinsic input resistance, one should apply shunt feedback; the amplifier is considered to be of the voltage-accepting type, and its actual input conductance is for the moment absorbed into the source conductance. The situation is then suited to the application of Theorems C and D. Conversely, if the required input resistance is larger than the inherent input resistance, the dual description, leading to the use of Theorems A and B, is indicated.

If power matching is required, a feedback factor of only two is possible. In this case an internal feedback loop should be used to obtain sufficient reduction of noise and distortion.

able—voltage or current—to be taken as the origin of the feedback and to the manner—in shunt or in series—the feedback is applied to the input. Take case (b) as an example (Fig. 7). The amplifier is of the current-accepting type, as far as the input is concerned. Towards the load it behaves as a current source as is required by the fact that voltage feedback is applied. Any inherent input or output conductance of the amplifier has been absorbed into the source resistance R_s and the load resistance R_L, respectively. The amplifier itself is characterized by its current gain h_{21}, the amount of short-circuit output current at one unit of input current. The output voltage e_L is by way of the feedback network converted into a feedback voltage e_f, that acts in series with the input. The proportionality factor is h_{12}:

$$e_f = h_{12} e_L$$

Substituting:

$$\begin{cases} i_L = h_{21} i \\ e_L = i_L \cdot R_L \end{cases}$$

gives:

$$e_f = h_{12} h_{21} R_L i$$

The total input voltage to the amplifier is equal to e_f, so

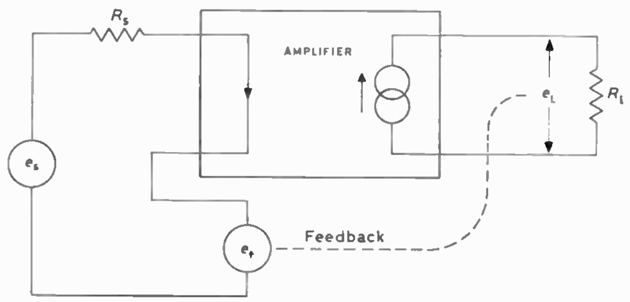


Fig. 7. Idealized amplifier with overall feedback, case (b)

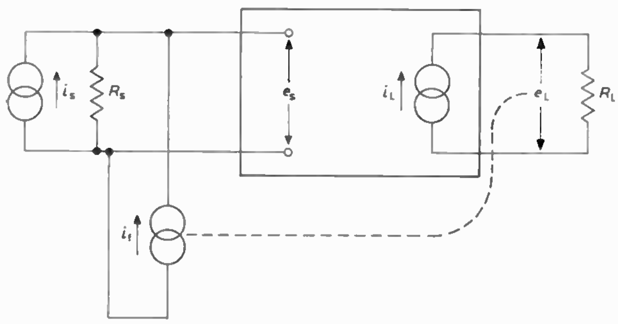


Fig. 8. Idealized amplifier with overall feedback, case (a)

Mutual Influence of Input and Output Resistance

The results, obtained so far, are entirely analogous to those derived for the internal resistance of a feedback amplifier[1]. In that case also the internal, or output, resistance depends entirely on the feedback configuration, and the feedback factor is solely dependent upon the ratio of the load impedance and the effective internal resistance. This correspondence is, of course, no coincidence. The two types of treatment express in various forms the same fundamental properties. This makes it feasible to try to go on and attack a still more general situation. In many cases a feedback path to the input originates from the actual output, and hence the feedback depends upon the load resistance. The same can be said from the point of view of the output.

Here one often resorts to the most general techniques, writing down and solving the circuit equations or using flow graphs, in order to get any answer at all. Clear insight into the situation is almost entirely lost. In many situations, however, the methods described here can be used to advantage. To demonstrate this the case will be studied of a single feedback path from output to input. There are four possibilities:

(a) Voltage shunt feedback
(b) Voltage series feedback
(c) Current series feedback
(d) Current shunt feedback

of which (a) and (c), as well as (b) and (d) form pairs of duals. These designations correspond to the output vari-

that the effective input resistance is:

$$R_i = h_{12} h_{21} R_L \quad \ldots\ldots (5)$$

Similarly, if the amplifier is driven by a source with a resistance R_s, its effective output resistance with respect to the load is:

$$R_o = R_s / h_{12} h_{21} \quad \ldots\ldots (6)$$

The feedback factor f can be found from the input circuit as well as from the output circuit:

$$f = 1 + h_{12} h_{21} \,.\, (R_L / R_s) \quad \ldots\ldots (7)$$

It is seen that the input resistance is simply proportional to the load resistance, a fact which is not unexpected for an amplifier stripped of all its inherent resistance. By absorbing all inherent resistances into the source and load resistances, the application of this principle is simple and straightforward.

Case (b), and, by duality, case (d) being disposed of, it remains to treat the cases (a) and (c). As Fig. 8 shows, case (a) implies a voltage-accepting amplifier that acts as a current source to the load. All input current flowing originates from the feedback, and loading of the output results in a current drop only via the feedback. The amplifier is described by a transductance, g_{21}:

$$i_L = g_{21} e_s$$

The output voltage is:

$$e_L = i_L \,.\, R_L$$

where R_L is the load resistance. The feedback causes a feedback current i_f to flow to the input circuit which is proportional to e_L:

$$i_f = g_{12} e_L \quad \ldots\ldots (8)$$

The input conductance is, therefore:

$$G_i = g_{12}g_{21}R_L$$

Here it is the input conductance which is proportional to the load resistance. Notice, however, that other factors of proportionality appear, to represent the gain of the amplifier and the reduction of the feedback network. In a similar way one derives that the output conductance is proportional to the source resistance R_s:

$$G_o = g_{12}g_{21}R_s \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

Finally, the feedback factor depends entirely upon the product of resistances in the input and output circuits:

$$f = 1 + g_{12}g_{21}R_sR_L \quad \ldots\ldots\ldots\ldots \quad (10)$$

Applications

The description of a few applications will serve to clarify the theory. To simplify matters only one type of transistor will be used, a unit with the following basic parameters at the working point:

current gain $\beta = 100$

emitter resistance $R_{eb} = 50\Omega$.

In common-emitter connexion the input resistance will be 5kΩ. All other factors affecting transistor performance will be neglected.

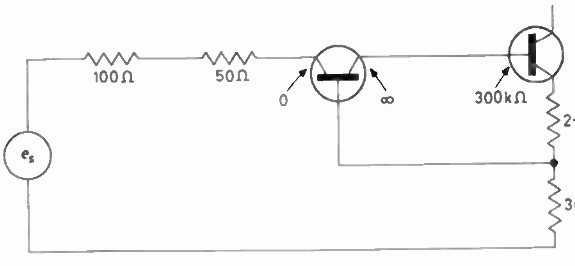


Fig. 9. Example of the application of the theory

Is it possible to design an amplifier with ten times amplification, equipped with negative feedback of 200 times, using two of these transistors? If so, what is the output resistance going to be? The solution proceeds as follows. Suppose the source has an output resistance of 100Ω. The feedback can create an input resistance either much larger or much smaller than the source resistance. The latter solution not being feasible, it will be necessary to create an input resistance of at least 200 × 100Ω = 20kΩ by series-type negative feedback. It is best to start from a low input resistance, as is exhibited by the common-base configuration. For series feedback the amplifier's input must be treated as a current acceptor, having zero input resistance. The total source resistance then becomes 150Ω, hence an effective input of 30kΩ has to be realized by the feedback.

An input current of 1μA should give rise to a feedback voltage of 30mV. Since the latter voltage is nearly equal to the input voltage, the output voltage at this current should be ten times larger: 300mV. A collector resistance of 300kΩ is sufficient for this purpose. The stage can now be terminated by an emitter-follower which at the same instance realizes the required resistance of 300kΩ at its input. Fig 9 gives the resulting design, drawn for a.c. only.

The output resistance without feedback is 3kΩ, since the emitter-follower, its base being connected to a point of infinite impedance, cannot operate as such. With feedback applied, 1·5mV applied at the output gives rise to an input current of 1μA, and to an output current of 100μA. Hence the effective output resistance is 15Ω.

With the common-emitter instead of the common-base configuration in the first stage, extra difficulties would have been encountered. An emitter resistance of only 100Ω necessary for the feedback loop, entails an increase of the input resistance to 15kΩ. Adding this to the source resistance leaves the necessity of realizing an effective input resistance of 3MΩ. For an input current of 1μA a feedback voltage of 3V should be developed across the 100Ω emitter resistor, which requires a current of 30mA. A total current amplification of 30 000 would be needed, which is more than two transistors can provide.

One might try to alleviate the problem by adding an auxiliary feedback loop. This should not add to the total feedback factor, but serve the sole purpose of modifying the resistance levels. Consider the situation discussed above. Shunt feedback, in the form shown by Fig. 10(a),

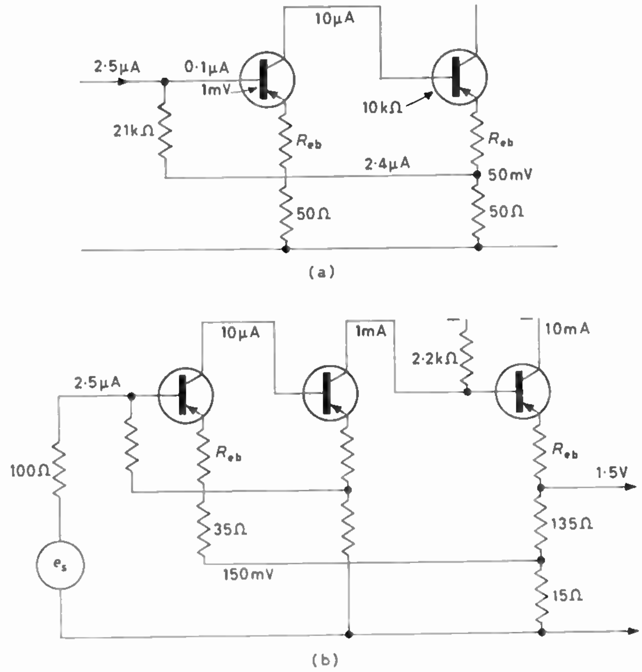


Fig. 10(a). Internal feedback loop

(b). Complete circuit, alternative design

reduces the input resistance to 200Ω. Overall series feedback has to enlarge the input resistance to 60kΩ for a feedback factor of 200. At 2·5μA input current the feedback voltage should be 150mV. The base current of the first transistor is 0·1μA and the feedback current to the emitter is 3mA. Again a current amplification of 30 000 is needed, which necessitates the use of an extra stage of amplification, e.g. an emitter-follower. Fig. 10(b) shows the resulting circuit.

A second example of the application of the theory concerns an amplifier required to work between two characteristic impedances of 1kΩ each. At either end the amplifier should be matched to 1kΩ, and the internal resistance should be virtually independent of the load impedance at the other end. The question arises: can an amplifier be designed such that the internal resistances are exclusively realized by negative feedback?

The theory has shown that with matched impedances a feedback factor of only two can be attained. Any larger feedback factor must be realized by an internal feedback loop in the amplifier. Furthermore, an ideal amplifier acquires an input resistance that is proportional (or inversely proportional) to the load resistance, when feedback is applied from input to output. This dependence can be decreased by shunting the load with a small resistance (R), as shown by Fig. 11(a). The square represents an ideal amplifier, in this case with zero input resistance and zero output conductance. The arrow indicates a feedback loop (series-type voltage feedback) which causes the input resistance to be proportional to the parallel value of R and R_L. When $R \ll R_L$, this is practically independent of

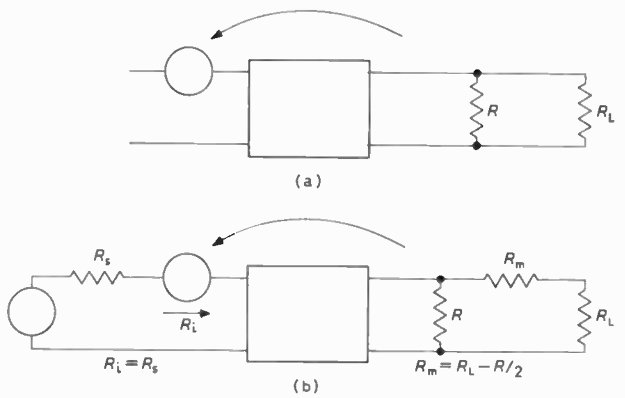


Fig. 11. Line amplifier, equivalent circuit

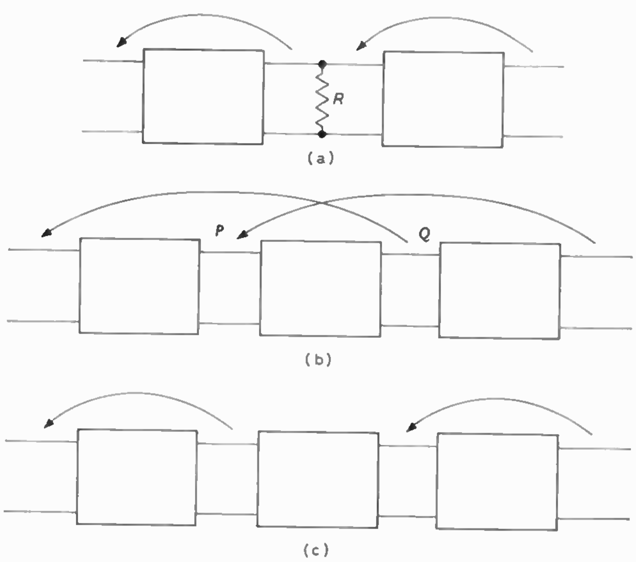


Fig. 12. Various possibilities of making input and output circuit independent of one another

R_L. Now one cannot obtain matching on both sides simultaneously. If the input end is matched, the feedback loop creates an output resistance again equal to R. The total output resistance is $R/2$ which is much smaller than R_L. To restore matching it is necessary to introduce extra resistance (Fig. 11(b)). Alternatively, one may connect a shunt resistor at the input and realize the matching at the output by feedback. The latter solution has the advantage of allowing a larger output power, since no power is lost in any actual series resistor. The system may, however, be more noisy.

It appears that the question posed above has to be answered in the negative, at least when one feedback loop is utilized. Consider next the effects of two consecutive feedback loops as given by Fig. 12(a). The same reasoning shows that R must be much smaller than the transformed value of either the source or the load resistance. In any event, simultaneous realization of the proper values of input and output resistance by feedback is impossible. The same applies to the situation of Fig. 12(b), as can be seen as follows. The voltage or current at Q is proportional to that at P, thus the situation does not principally differ from the one of Fig. 12(a), and the same reasoning will hold again. Only a situation like the one depicted by Fig. 12(c) meets all requirements. An isolating stage, either a unilateral amplifier or a resistive attenuator provides the necessary independence. From a practical point of view this solution is a quite complicated one, compared to the one in Fig. 11.

Conclusion

The rules for applying the theory are simple and logical. When feedback is taken from or fed to a point by a shunt connexion, that point should be considered as having an infinite internal resistance. For instance, voltage feedback demands the output circuit to be drawn as a current source, and shunt feedback causes the input circuit to be considered as a voltage acceptor. Alternatively, when application of feedback necessitates the opening of a loop, this loop should be considered to have zero internal resistance, at least on the side of the amplifier. To be specific: current feedback necessitates a current-sensing element to be inserted in the output circuit; consequently, the amplifier is to be regarded as a voltage source. Similarly, series feedback entails a voltage source to be inserted in the input circuit, and the amplifier's actual input should act as a pure current acceptor with zero resistance.

In all cases the inherent input and output resistances are to be absorbed into the resistances of the driving source and the load, respectively. This process is to be executed in accordance with the character of the amplifier as specified above: a current-acceptor input demands a closed current loop including the driver, and all inherent resistances are to be taken in series, etc. etc.

The result of applying, the theory is that the equivalent circuit of the amplifier with feedback contains at least one resistance, although the original amplifier did not contain any resistance. This effective internal resistance portrays a new dimension the circuit has acquired as a result of the negative feedback. The dynamics of the circuit, and especially the exchange of energy between the amplifier and its surroundings can now most proficiently be described.

The next step is to compute the amount of feedback applied, the feedback factor. This is important for two reasons: it is the factor with which the amplification as well as the distortion and noise are decreased as a result of the feedback, and it also governs the stability of the circuit. Why the treatment proposed in this article is to be preferred over others, is evident from the simplicity of the resulting relations. It makes it possible to understand the operation of the system completely without any difficulty. Even complicated feedback systems with concentric or interleaving feedback loops can be treated and the operation will be understood easily.

In textbooks negative feedback is usually dealt with in a completely different way. If one looks at it closer, these cover just the cases not mentioned above, e.g., current feedback to a current-source amplifier or series feedback to a voltage acceptor. The equivalent circuits of the amplifier with feedback then do not differ in form from those of the original amplifier. Simple expressions for the amplification are found easily, but only when no loading is present or when a pure current (or voltage) source acts as driver. The formulae become quite complicated when load and driver resistances are taken into account.

REFERENCE

1. De Boer, E. Internal Resistance of Feedback Amplifiers. *Electronic Engng. 34*, 600 (1962).

Short News Items

The Royal Norwegian Navy has placed a contract with Decca Radar Ltd for true motion radar equipment which is to be fitted in a completely new class of motor gunboat.

The decision to fit this equipment in their latest m.t.b.'s has been made by the Royal Norwegian Navy in the light of operational experience of similar radar. The British-developed technique of true motion display has proved a key factor in providing rapid, accurate and easily interpreted tactical information for the operation of high speed craft in land-locked waters.

The Ministry of Defence are using a digital data logging system supplied by Digital Measurements Ltd for the study of patterns of air turbulence.

The outputs from a cross-wind line of anemometers on masts are fed to a 20-channel data logging equipment. This equipment enables readings to be taken at intervals ranging from about ½sec to a few seconds, according to the particular experimental requirements. The digital output is on punched paper tape which is subsequently analysed by a Mercury computer.

In the computer the wind components are subjected to calculations which highlight the eddy structure of the fluctuations. The patterns of wind arrows which can be drawn from the analysed data illustrate the shapes and sizes of the eddies. The punched data is also used to calculate the statistics of wind fluctuations.

Erie Resistor Ltd has formed a European subsidiary company to promote sales of the company's products in Benelux, Western Germany and neighbouring countries. The name of the company is Erie Continental S.A., and it is located at 140 Avenue Eugene Plasky, Brussels 4, Belgium. The new company will work in close liaison with Erie's London Export Department.

'Apparatus and Instrumentation of High Frequency Pulse Measurements' is the title of an E.R.A. Report first issued to members in 1960 but now generally available.

The report describes apparatus developed for generating pulses of oscillation (frequency in the 200Mc/s range) of adjustable duration and of some kilovolts in amplitude; it describes the measures adopted for obtaining calibrated oscillograms of the pulse profiles, and explains how controlled mid-gap irradiation is applied.

The pulse is initiated by the discharge of a bank of capacitors through a gas relay in series with the oscillator and is ended by the application of a negative bias to the oscillator grids. A high-Q coaxial transmission line is coupled at one end to the oscillator; the other end carries the electrodes of the test gap, a high-frequency electrometer of novel design and the launching disk of a piston attenuator. This is interposed between the test gap and the oscilloscope, and with the high-frequency voltmeter provides the means of calibrating the system.

Copies of the report priced at 12s. 6d., plus 6d. postage, are obtainable from the Electrical Research Association, Cleeve Road, Leatherhead.

The first E.L.D.O. three-stage satellite launcher is now undergoing static test.

In a specially converted test tower at the Hatfield factory of Hawker Siddeley Dynamics, the 104ft tall Europa I has been set up for vibration tests that will yield information on structural whipping and bending stresses and will determine the position of rate gyros in the autopilot loop.

The first Europa rocket will never leave the ground. It has been built solely as a static test round, and although externally identical to the flight standard vehicles, with complete fuel tanks, substitute materials suitable for ground testing but not for flight have been used in its manufacture. Heavy steel pipes represent the mass and weight of the engine thrust chambers, and these will be gimballed during the tests.

Europa I components have been arriving at Hatfield since last December, and individual stages were mated together on the ground before the intricate task of assembly in the tower was undertaken.

During the vibration tests the fuel tanks of the three stages which make up Europa I will be ballasted, either with water or with chemicals which have specific gravities similar to the fuels involved, to make up the launch weight of the vehicle.

Four steel ropes, attached to the launcher pins on the propulsion bay of the Blue Streak first stage, will suspend the entire vehicle from the top of the tower. The vehicle is strain gauged with more than 200 readings being continuously fed to a battery of recorders for eventual computer analysis.

The West German Navy has ordered a fully transistorized marine radar simulator from Redifon Ltd for training naval officers in the interpretation of marine radar displays, 'blind' pilotage and navigation.

The equipment can simulate the operation of two radar-equipped ships and eight target ships. The course and speed of the eight target ships can each be controlled by the instructors to present the trainee with operational problems in collision avoidance, station keeping, tactical fleet exercises and manoeuvres.

The instructor can select any radar indication problem normally met with during marine operations, including receiver noise, sea clutter, tide and yaw conditions.

A coastline generator is also provided to present a realistic radar picture of any selected coastal or harbour area.

A similar installation is being provided for the Japanese Marine University at Kobe. This version provides one radar-equipped ship and four target ships and will be used for training marine officers in 'blind' pilotage and radar navigation.

The British Standards Institution has prepared a supplement to its Code of Practice on 'The reception of sound and television broadcasting' (CP 327.201) to give guidance on some of the problems likely to be encountered in the reception of the new u.h.f. service.

The supplement considers in particular the basic requirements for good reception, types of aerial system (including communal serials) and the siting of aerials.

Since BBC 2 television programmes are already under way—making the need for guidance more urgent—the BSI has issued the supplement without taking the usual step of inviting comments from industry generally.

Copies of Supplement No. 1, 'The reception of u.h.f. television broadcasting' (to be ordered as PD 5282), may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1, price 2s. 6d. each (postage will be charged extra to non-subscribers).

The English Electric Valve Co. Ltd has purchased from Associated Electrical Industries Ltd its Carholme Road factory at Lincoln, together with the valve business being carried on there.

The transfer of the factory will enable the English Electric Valve Company to meet the rapid expansion of its business which is out-growing the facilities now provided by its factory at Chelmsford, Essex. It is intended to increase still further the production of specialized electronic valves and tubes used in television transmission and telecommunications and eventually to build up the labour force at the Lincoln factory to some 1 000 people.

The English Electric Valve Company will lease to Associated Electrical Industries a part of the factory, and A.E.I. are able to continue the development and manufacture of semiconductors in the present premises.

Associated Electrical Industries, one of the major manufacturers of semiconductor devices in the United Kingdom, will extend its operations in this field and semiconductor production will continue at the Lincoln factory and at the A.E.I. Rugby works as at present; output is to be stepped up.

The Paul Instrument Fund Committee has made grants as follows:

£1100, as a supplementary grant, to Dr. P. B. Hirsch, F.R.S., lecturer in physics in the University of Cambridge, for continuation of his work on an electron microscope with velocity analyser.

£5400 to Professor D. J. E. Ingram, professor of physics, University of Keele, for the design and construction of a millimetre wavelength electron resonance spectrometer to work at zero or near zero magnetic field strengths and with as wide a frequency coverage as possible.

£2200 per annum for two years to Dr. H. Motz, Donald Pollock reader in engineering, University of Oxford, for the construction of a pulsed millimetre wave generator.

£3400, as a supplementary grant, to Dr. E. E. Schneider, reader in solid state physics, University of Newcastle upon Tyne, for continuation of his work on the application of superconducting cavities to magnetic resonance spectrometers.

The Paul Instrument Fund Committee, composed of representatives of the Royal Society, the Institute of Physics and the Physical Society and the Institution of Electrical Engineers, was set up in 1945 'to receive applications from British subjects who are research workers in Great Britain for grants for the design, construction and maintenance of novel, unusual or much improved types of physical instruments and apparatus for investigations in pure or applied physical science'.

A co-ordinated programme of research into superconductors and the development of their applications in superconducting magnets has been initiated by the D.S.I.R. Success in this field could assist physical research where powerful magnets are needed, help communications systems, lead to increased efficiency in electric power generation and in due time result in improvements in power transmission lines.

The first part of the programme, costing about £200 000, of which industry will be required to contribute at least £40 000, involves Government research stations, independent research organizations and industry, and also research at universities.

Research contracts placed with manufacturing organizations will be on a cost-sharing basis and the scientific results will in general be published. Arrangements will be made to secure an appropriate financial return to the department on patentable results.

The three broad lines of enquiry are:

(1) Investigation of superconducting materials and their behaviour.

(2) The problems of constructing solenoid magnets from them.

(3) The investigation of technical applications of superconductors other than in magnets.

Ariel I (originally called U.K.1) is to be monitored again at United States telemetry stations and at the three British-run stations at Winkfield, Singapore and at the South Atlantic station shortly for a period of two months. This will be done at the request of the University College, London, group responsible for one of the experiments fitted into Ariel I.

The satellite, containing British-designed experiments, was successfully launched from Cape Canaveral by an American Delta rocket on 26 April 1962 and for a long time valuable information was received from it, but the monitoring of the transmissions was discontinued at the end of June 1964.

The resumption of monitoring of Ariel I has been prompted by the launch recently of the new American satellite Explorer 20 (nicknamed TOPSI). The experiments in Explorer 20 are predominantly American and are designed to study the properties and behaviour of the ionosphere; but they include one experiment designed by the University College, London, group under Drs. Boyd and Willmore. This British experiment supports the main purpose of the American experiments in the new satellite. The reason for the request for resumption of monitoring of the experiment still working in Ariel I is that it also makes ionospheric measurements which may be correlated with those to be made now by Explorer 20.

For a short period, therefore, British experiments in three satellites—Ariel I, Ariel II (launched as U.K.2 on 27 March 1964) and Explorer 20—will be under observation simultaneously.

EMI Electronics Ltd and Ing. C. Olivetti & Co. S.p.A., of Ivrea, Italy, have recently concluded an agreement under which EMI will assume responsibility for the marketing and maintenance of the Olivetti continuous path machine tool control system in the United Kingdom, EFTA (except Switzerland), British Commonwealth countries (except Canada) and a number of other areas.

EMI will also provide a computer programming bureau to give customers a speedy tape preparation service for the recorded magnetic tapes which are employed with the Olivetti system of numerically-controlled machine tools.

Ferranti Ltd has moved its London office from Kern House, 36 Kingsway, London, W.C.2, to Millbank Tower, Millbank, London, S.W.1. The telephone number of the new London office is VICtoria 6611.

Exports of valves, tubes and semiconductor devices during the second quarter of 1964 totalled £2 626 723, according to figures issued by the BVA and VASCA, based on the Customs and Excise returns.

While the overall total for the quarter is slightly down, most of the valve categories show an increase over the first quarter of the year.

Digital Measurements Ltd has recently supplied data-logging equipment to be used for the monitoring of the air conditioning system installed at the Shell Centre in London. The equipment consists of temperature transducers (platinum resistance air temperature thermometers) with their associated bridge circuits, and a multi-channel data logger producing a printed output from an automatic electric typewriter.

The data logger is in a centralized control room and the transducers are at widely spaced locations, necessitating very long cable runs. The equipment monitors temperature over a wide range and the accuracy between 50°F and 90°F is ±0·5°F, the resolution being 0·1°F. The output, which is scaled to read directly in degrees, is displayed on a digital voltmeter in the data logger, and is also printed out by the typewriter. The output of any transducer can be indicated on demand and, in addition, a timing circuit commands the logger to produce a complete statement of all temperatures every half hour.

Apart from the typewriter, the standard rack-mounting units are contained in a cabinet 20in wide by 18in deep by 35in high.

In Spain this autumn 29 members of the Scientific Instrument Manufacturers' Association will be taking part in exhibitions being held first in Madrid and then in Barcelona.

The organization of these exhibitions by SIMA in conjunction with the Board of Trade is part of a campaign to promote the sales of British instruments in Spain.

At the moment fixed buying patterns in the Spanish market for instruments have not yet been firmly established. Important suppliers are West Germany, France, U.S.A. and U.K., but the totals fluctuate widely from year to year and no clear trend has yet been established. It is thus of prime importance that the British instruments industry should establish itself in this market which is both growing and flexible.

In 1962, exports of instruments to Spain totalled £650 000 and in 1963 this rose to £850 000.

Over the four-year period of the Development Plan, instrument expenditure is expected to increase by 100 per cent and by the end of this year it is hoped that British instrument sales to Spain will be more than £1M.

The liberalization of trade in recent years, together with the progress anticipated under the Development Plan, is bringing with it increases in our exports to Spain. The aim of the SIMA exhibitions is to promote this.

Redifon Ltd has recently delivered a number of automatic 80W m.f. beacon transmitters for operation in remote areas of Iceland. These beacons will be used as aircraft navigational aids.

The G.142 dual transmitter covers a frequency range of 200 to 420kc/s and complies fully with ICAO requirements. It can be used as an airfield beacon, a route marker, or an approach or 'holding' aid in the vicinity of an airfield, and can be employed as a locator beacon in an instrument landing system. The power of the beacon can be controlled in steps down to one-sixteenth of full power.

The equipment, supplied in weatherproof kiosks, can be left unattended in remote sites for long periods and provides continuous operation under widely varying conditions. It incorporates an automatic changeover device and monitoring unit, a dual automatic capacitor, and two transmitters each with its own power unit. Reserve transmitter and keying unit are automatically switched into service if the transmitter in use does not radiate signals at adequate power with correct modulation, or if the keying unit fails.

B.O.A.C. has placed a contract valued at £½M for the Epsylon flight data acquisition system, and flight trials in the 707's and VC.10's are scheduled for the end of the year.

Epsylon Industries Ltd, a member of the Stone-Platt group, has based its development of the EFDAS system on its long experience of analogue and digital data acquisition schemes for flight trials and research projects.

The purpose of the integrated system ordered by B.O.A.C. and designed to the airline's own specifications is to meet the Ministry of Aviation's mandatory requirement for accident data information (the so-called crash recorder) as well as to provide operational data which will assist the airline in its search for greater efficiency and economy.

The equipment consists of airborne monitoring and recording units and ground automatic data processing equipment, both of which use the very latest techniques.

EFDAS is designed on the modular principle so as to meet the requirements of small, medium and large airlines and can be introduced in easy 'add on' stages if desired. The minimal system meets the mandatory requirement with eight data channels to spare and can be extended up to a 48- or even 96-channel capacity. The overall system accuracy is within ±1 per cent and the airborne recording capacity is up to 200 hours for mandatory information and 125 hours for operational data. The operational data can be processed on the ground at 100 times recording speed and selected information, corrected for any non-linearity of aircraft instrumentation or transducers, fed out in computer language, typescript or in graphical form.

The first Noise and Vibration Reduction Exhibition—NAVREX—is to be held at Earls Court, London, from 15 to 19 June 1965.

The exhibition, which will be sponsored by Trade & Technical Press Ltd and organized by Iliffe Exhibitions Ltd, will be the first to deal objectively and exclusively with the increasing problems associated with the reduction and control of noise and vibration.

In conjunction with NAVREX a conference will be held within Earls Court, at which technical papers will be presented.

The Marconi Co. Ltd has won an order to develop and supply the Doppler navigation sensor for the prototype Anglo-French supersonic airliner, the Concord.

The Marconi Company has already sold a considerable quantity of Doppler navigation equipment, types AD2300 D and E, to the French Air Force for the Mirage IV supersonic bomber and the Mirage IIIE strike fighter.

The equipment for the Concord will be a development of the recently introduced Doppler navigator, type AD560. This equipment, which forms part of the Marconi 'Sixty Series' of airline standard navigation and communication equipment, will be modified to operate in the high temperature environment resulting from the kinetic heating of the aircraft skin at high speeds.

The Radar Division of Cossor Electronics Ltd has received an order valued in the region of £200 000 for CRD.100 transistorized display systems, and extensive back-up equipment, for the Ministry of Aviation Air Traffic Control Training School and Evaluation Unit at Hurn Airport.

A total of twenty-two operational positions and two monitoring positions will be fitted with the viewing units and three spare units will also be included. The equipment provides a comprehensive system capable of accepting unprocessed radar data from a radar simulator, a radar film recorder, video maps and external radar heads, and after processing will distribute the information to any or all of twenty-four display positions.

The system is also capable of accepting and distributing the output of the radar data processing and storage system and of the alpha-numeric symbol generator.

Installation of the system will be carried out by Cossor's planning and Installation Unit and is being planned so that training may continue while the work is in progress.

The Home Office has placed an order with the Marconi International Marine Co. Ltd for 170 'Viking IV' power megaphones, to be supplied with extension leads and remote microphones.

These will be used for training Civil Defence Corps personnel.

Elliott-Automation Ltd has been selected to lead a three-nation team for the development of the automatic flight control system for the Anglo-French Concord supersonic airliner.

Elliott's, as prime contractor, will have working with them the French company S.F.E.N.A. and The Bendix Corporation of America, which is to make available its full technical support for the development of the project.

The automatic flight control system includes the automatic pilot, including automatic landing, the electrical trim system, the auto-stabilization and auto-throttle systems.

The decision to place this work with this team of companies was taken by Sud Aviation in agreement with The British Aircraft Corporation and with the official approval of the British and French Governments.

S. G. Brown Ltd, a Hawker-Siddeley company, has finalized an agreement with the Wayne-George Corporation, U.S.A., for the sole British distribution rights of the Corporation's complete range of optical shaft encoders and gyro test tables.

These encoders will be additional to those already designed and manufactured at the Chiswick factory of S. G. Brown. Typical applications include radar aerials, telescopes, servo systems, stable platforms, guidance systems, machine tools and industrial control positioning systems where accuracy, reliability and power consumption are important considerations.

BOOK REVIEWS

Control System Design

By C. J. Savant. 457 pp. 2nd Edition. Mc-Graw-Hill Publishing Co. Ltd. 1964. Price 89s.

CONTROL system technology is no longer an upstart science crowded with enthusiasts and new ideas. Some degree of middle-age is becoming apparent and, with it, an accompanying discretion. This is demonstrated most clearly by the appearance of this second edition of a notable textbook. It would appear that the first edition was a commercial success and, not surprisingly, since the author's style and original choice of material was extremely well judged.

This new edition has been meticulously revised and a large number of pages have at least subtle changes of word or style, bringing the text into line with contemporary usage. Here and there considerable modifications from the first edition are noticeable; in particular, much attention has been paid to improving the already useful section on network synthesis as applied to system equalization. However, the broad topics are almost as before, commencing with a very general chapter introducing feedback control systems. Chapters 2 and 3 deal with setting up the differential equations for systems under study, and steady state errors of systems. Chapter 4 is a nice presentation of the now well-accepted root-locus method, and in the following chapter the frequency analysis technique is clearly expounded, bringing in all the well-known graphical methods including Bode plots and the Nichols chart.

Chapter 6 is entitled 'The design of feedback control systems' and is an excellent, concise treatment on network synthesis methods suitable for the design of compensating networks to improve system performance for both d.c. and a.c. systems. Choice of primary components such as servomotors is not dealt with here despite the heading. Chapter 7 is a comprehensive catalogue of mathematical models of system components, including servomotors, gyroscopes, accelerometers, amplifiers, and modulation equipment, although nothing is given on fluid power system components. The final chapter sets out the two most widely used design tools for non-linear systems, namely, the describing function method and the phase plane analysis.

Several appendices provide the theoretical and mathematical background required for the book including Laplace Transform theory, an introduction to determinants, roots of equations, and derivations of the Nyquist stability criterion. Appendix VI gives design procedures for bridge-T and parallel-T networks, and appendix VII provides a useful test for the stability of linear systems.

Throughout the book use is made of well chosen examples to illustrate the argument. There are fewer references in the new edition, but since these are of other basic textbooks, they are quite adequate for the purpose.

The book has been enhanced by the careful revision it has been given, and it remains one of the clearest and best presented textbooks in the field.

K. C. GARNER.

Theory of Superconductivity

By J. M. Blatt. 486 pp. Med. 8vo. Academic Press. 1964. Price 89s.

SINCE its discovery by Onnes in 1911, the phenomenon of superconductivity has presented many problems to the theoretical physicist. Professor Blatt's book explains, in considerable detail, how theories have changed and developed over the years, and brought us to a position where the theory of soft superconductors is reasonably well understood.

Because of the complexity of the problem, a great deal of the book is concerned more with the mathematical techniques rather than the physical basis of superconductivity. A high standard of quantum mechanics and solid state physics is assumed in the text, the more specialized material being developed in a 57 page Appendix. A comprehensive bibliography allows ready access to papers on many aspects of superconductivity.

The aim of this book is to put the present theories into historical perspective but there is so much detail that the book will appeal primarily to specialists in superconductivity or those wishing to explore the field in considerable depth. In view of this it would have been instructive if there were a detailed discussion of the direction in which theories may move in the future, to include, for example, the problem of the correlation of superconducting properties with the detailed Fermi surface metals and the theory of superconductors that do not show the simple isotope effect ($T_c \propto M^{-\frac{1}{2}}$).

At present, much of the experimental work on superconductivity is concerned with HARD superconductors which allow flux penetration at field strengths below the thermodynamical critical field, but remain superconducting to much higher fields. The interest in this work is stimulated by the fact that some of the hard superconductors (e.g. NbZr and Nb_3Sn) can support high current densities and are therefore important both practically and theoretically. The rapid progress made in this topic since the book was written underlines the difficulty of keeping pace with an expanding subject. However, an excellent discusion is given of the theoretical position as seen in early 1963, and of the developments that resulted in a very satisfactory theory of soft superconductors.

P. TOWNSEND.

Optimization Theory and the Design of Feedback Control Systems

By C. W. Merriam III. 391 pp. Med 8vo. McGraw-Hill. 1964. Price £5 12s.

ONE of the recent developments in the field of control engineering theory has been the application of optimization techniques to the design of control systems. The real usefulness of these methods has arisen in the design of complex control systems, for example, in guided missiles, in aircraft landing systems, or in complicated manufacturing processes.

When conventional methods (e.g. root locus and frequency response) are applied, the experience and intuition of the designer must be employed in the variation of system gains in an iterative process by which it is hoped a satisfactory design will be reached. Clearly this process need not always converge. These procedures also become extremely complicated if it is desirable to vary the system gains automatically during the process.

In the optimization method, an error index is defined, which is the weighted sum of the errors in the responses of importance in the system. It is in the choice of weighting that the designer's skill is employed. The system is then solved to give a minimum value of the error index. The book is mainly concerned with the mathematical techniques involved in this solution.

The first three chapters are introductory. In the first chapter the chronological developments in optimization theory applied to control system design are reviewed. Chapter 2 and 3 deal with the simpler cases of parameter and impulse response optimization, and serve as an introduction to the application of optimization theory to this type of problem.

Chapters 4, 5 and 6 deal with the mathematical background required for control system optimization. Examples are given of various mathematical forms of weighting function.

Chapters 7 and 8 introduce the control-equation synthesis problem and deal with linear control systems. The last three chapters deal with nonlinear control systems.

There are eight appendices. Appendix E, which gives the application of optimization theory to the non-trivial case of an aircraft landing system, is particularly useful.

The text is well written and the subject logically presented, however, a reasonably high degree of mathematical proficiency is necessary to derive full benefit from this book.

G. D. BERGMAN.

Precis d'Electromagnetisme Theorique

By P. Poincelot. 456 pp. Demy 8vo. Dunod. 1963. Price 76NF.

THIS book represents a treatise of electromagnetic theory starting from Maxwell's equations. It is highly mathematical and an engineer wishing to find a quick solution to a practical problem in, say, waveguides would be disappointed. What he would find is a treatment which, although sometimes restricted to ideal cases (e.g. infinite conductivity) would give him a deep understanding of the mechanisms involved, provided he has a good mathematical background. It should be easy to read by those with little knowledge of French due to the mathematical nature of the content. Written text, where it occurs, is clear and concise.

Rationalised m.k.s. units are used throughout, the fourth fundamental unit chosen being the ampere. A short chapter is devoted to deriving relationships between the m.k.s. and c.g.s. systems.

Three more chapters deal with general electromagnetic theory based on Maxwell's equations. Of the remaining 75 per cent of the book, about one-third is devoted to electrostatics and two-thirds to electromagnetism.

Both these sections start with a rapid but thorough treatise after which a number of specific cases is treated under electrostatics; the dipole, charged planes, spherical surfaces, conductors, dielectrics and forces in the electric field represent the main headings. Some 19 particular problems are then considered and it is significant that the author has eliminated those problems, which are no more than an exercise in mathematics, to retain those which are significant to a study of electric field phenomena.

The chapters on electromagnetism follow a similar pattern and lead naturally to cavity resonators, the coaxial line. and waveguides. There is a good section on skin effect and several problems relating largely to electromagnetic radiation are treated, including a thorough treatment of the dipole.

G. ASHDOWN.

Laser Abstracts—Vol. 1

By A. K. Kamal. 177 pp. Med. 8vo. Plenum Press. 1964. Price $12.50

This book contains some 750 abstracts of published laser papers from the earliest researches until 1963.

Entries are indexed by category as well as contributing authors.

An Introduction to Counting Techniques and Transistor Circuit Logic

By K. J. Dean. 223 pp. Crown 8vo. Chapman and Hall. 1964. Price 25s.

This book explains the design and use of circuit modules for transistorized counting, static switching system and simple logical machines, and also discusses both transistor-resistor logic and diode logic.

Dictionary of Plastics

By J. A. Wordingham and P. Reboul. 211 pp. Crown 8vo. George Newnes. 1964. Price 30s.

This dictionary has been based on information on plastic materials and processes and gives definitions of a wide range of terms in present day use.

Included also are appendices giving trade names, a list of British Standards on plastic materials and a bibliography covering the more important industrial processes.

Quantum Electronics III

Edited by P. Grivet and N. Bloembergen. 1923 pp. Med. 8vo. Columbia University Press. 1964. Price £12 12s. (two volumes)

These volumes contain the papers, and the subsequent discussions, which were presented at the Third International Conference on Quantum Electronics held at Paris in February 1963, and includes the latest applications and theoretical considerations of the laser and maser.

Proceedings of the Symposium on Electron and Vacuum Physics, Hungary 1962

Editor E. Winter. 506 pp. Crown 4to. Akademiai Kiado. 1963. Price $11.00

This volume contains the papers read at the Symposium on Electron and Vacuum Physics held at Balatonfoldvar in 1962. They deal with the problems of producing and measuring high and ultrahigh vacuum, of preparing various cathode types and substances, of current passing through vacuum and gases, of plasma physics, of electric discharges in gases, etc.

Mechanical and Electrical Vibrations

By J. R. Barker. 221 pp. Crown 8vo. Methuen & Co. 1964. Price 21s.

This is a suitable textbook for students taking a degree or diploma in engineering or physics, and is still useful at the post-graduate level. It is an attempt to link up mechanics and electricity in a way which will make condensed and profitable reading for students.

Comparatively elementary mathematics is used as far as possible, but the reader is prepared for more advanced mathematical techniques such as the Laplace Transforms or matrices. The development of the electro-mechanical analogies and their application over a wide field is an important theme of the book.

Schaltungen und Elemente der digitalen Technik (Circuits and Components in Digital Techniques)

By Bartels and Oklobdzija. 150 pp. Med. 8vo. Verlag für Radio-Foto-Kinotechnik GmbH. Berlin. 1964. Price DM. 21

The authors, both with Standard-Elektrik-Lorenz AG, present basic circuits and components used in digital techniques in a manner suitable for direct application by engineers of different attainments. They have drawn on their own wide experience to prepare a kind of digest. Part I deals with circuitry, starting with simple relay applications progressing via code processing to analogue-digital conversion. Part II gives the rules for the application of some specific relays, including those of the dry reed type, and of electronic elements including logic.

Electronic Universal Vade-Mecum, Vols. 1 and 2

Edited by P. Mikolajczyk and B. Paszkowski. 1 449 pp. Demy 4to. Pergamon Press. 1964

This Vade-Mecum comprises a compilation of data on some 25 000 valves and semiconductors and is printed in seven languages—English, French, Spanish, German, Polish, Russian and Italian.

Volume 1 covers information on radio-receiving type valves manufactured over the world and which are classified into 442 groups comprising valves of identical or similar characteristics.

Volume 2 contains information on transmitting valves, magnetrons, nuclear radiation detectors and semiconductors classified into 580 groups.

Two indexes are included for easy location of a given type of device and its equivalents.

Electromagnetism for Engineers

By P. Hammond. 209 pp. Crown 8vo. Pergamon Press. 1964. Price 17s. 6d.

This is an elementary book in that no knowledge of mathematics is required of the reader other than simple differentiation and integration. It is intended as a text book for first- and second-year students in universities and colleges of technology.

Electron Tube Handbook, 1964

800 pp. Demy 8vo. Brown-Boveri & Co. Ltd. 1964

This new tube handbook replacing the previous 1961 edition includes many supplementary items. Eleven chapters are presented in German, English and French. The first chapter sets out the symbols used. Chapter 2 'Definitions and Useful Information' has been greatly increased and gives all the necessary information for determining application, choice and operation of transmitting and rectifier tubes and thyratrons. Two hundred and twenty pages are devoted to this chapter which is supplemented by tables for the recognition and prevention of faults in electron tubes. Chapter 3 tabulates formulae, tables and wiring diagrams for easy reference.

The next seven sections present detailed information on the high voltage rectifier tubes, thyratrons, force-cooled transmitting tubes for communication purposes and industry as well as radiation cooled transmitting tubes.

LETTERS TO THE EDITOR

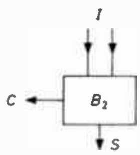
(We do not hold ourselves responsible for the opinions of our correspondents)

A Decimal to Binary Decoder

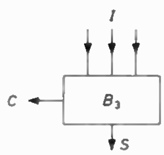
DEAR SIR,—While scanning through Volume 35 of *Electronic Engineering* recently in search of a particular reference, I came upon Mr. Nairn's article[1] describing a decimal to binary decoder. My interest was at once aroused because only a month back I was presented with a very similar problem. It was sufficiently different from Mr. Nairn's, however, to make it worthwhile bringing it to your notice.

Whereas Mr. Nairn began with four outputs carrying the units in binary, six outputs carrying the tens in binary and eight ouputs carrying the hundreds in binary, and, very logically, added these three binary numbers to give the net 3-digit decimal number in binary form, I was presented with three identical ten-position switches ("Multiswitches", made by Messrs. Contraves of Zurich, and marketed in this country by Messrs. Kynmore Engineering Co. Ltd.), each giving the numbers 0 to 9 in binary. Each of these switches has four outputs therefore, and is of the form of Fig. 2 in Mr. Nairn's article.

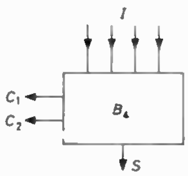
Fig. 1 gives my solution to this problem. The boxes *B* have truth tables as follows:



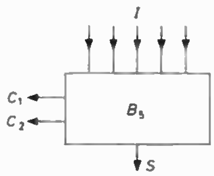
I	S	C
0	0	0
1	1	0
2	0	1



I	S	C
0	0	0
1	1	0
2	0	1
3	1	1

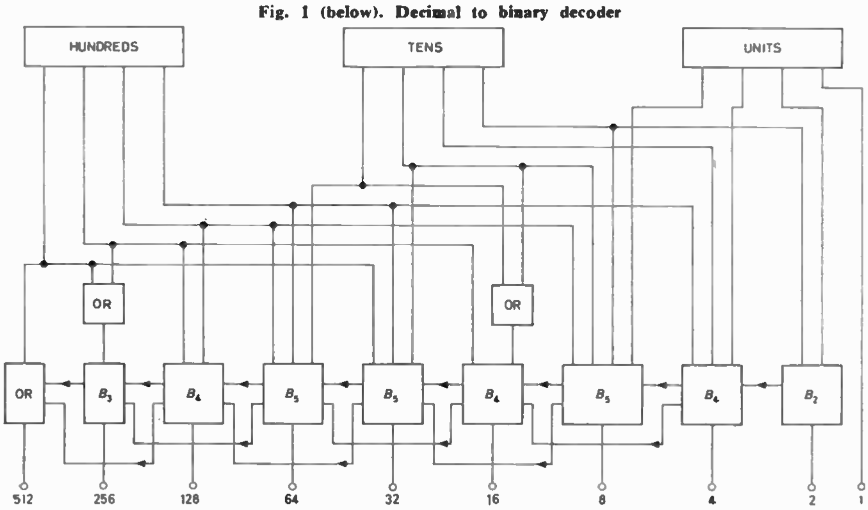


I	S	C_1	C_2
0	0	0	0
1	1	0	0
2	0	1	0
3	1	1	0
4	0	0	1



I	S	C_1	C_2
0	0	0	0
1	1	0	0
2	0	1	0
3	1	1	0
4	0	0	1
5	1	0	1

Fig. 1 (below). Decimal to binary decoder



and etc. In general there are as many outputs as there are binary digits in the number of inputs.

Fig. 2 shows the reduction of B_7 to a system of half-adders and OR gates. The pattern for lower and higher order boxes can easily be deduced from this. OR gates are needed in general to separate the paired 'carry' outputs. Diodes are quite satisfactory for this purpose, though I notice Mr. Nairn preferred to use instead an extra relay contact.

It is interesting to see why the solution of my problem is distinctly more complex than that of Mr. Nairn's. In the latter case, each decimal digit can only supply *one* input to any one of the adders. In the former case, however, the 'tens' and the 'hundreds' switches supply *two* inputs to certain of the boxes. A slight simplification has been effected by using OR gates (diodes) to separate those lines which cannot carry voltages simultaneously. Thus the 'tens' switch cannot provide simultaneously an output on its first and third terminals, neither can the 'hundreds' switch provide simultaneously outputs on its first and second terminals, nor can there be more than one input to the '512' output at any time.

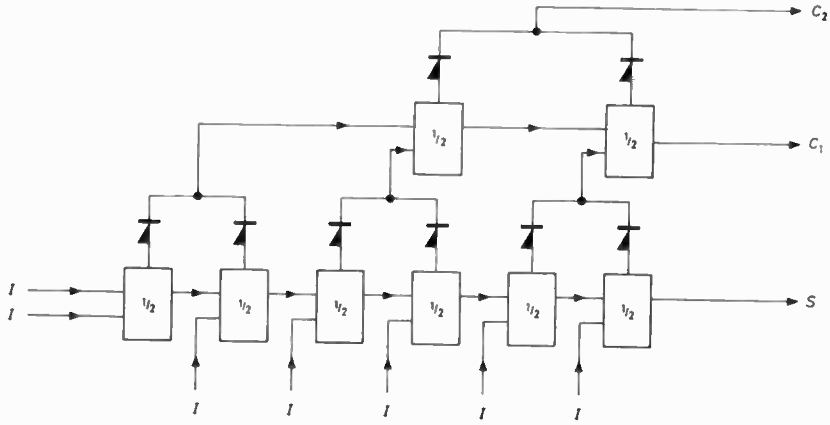


Fig. 2. Reduction to half-adders and OR gates

I am much indebted to Mr. Nairn for suggesting the use of a double-coil Post Office relay as a half-adder, and I was intigued by the ingenious way he succeeded in forming a treble half-adder out of only two such relays. There does seem a real case in problems of this kind for developing a really simple and compact half-adder.

In conclusion, I would point out that since there is a one-one correspondence between the configuration of output voltages and the setting of the decimal switches, it must be possible to construct a decoder entirely of ten-position wafers grouped into three sets (units, tens and hundreds), using diodes where necessary to eliminate unwanted loops. The problem is pretty formidable and probably of academic interest only, but to anyone attempting it, I would suggest the simpler problem of a 3-digit ternary to 5-digit binary conversion as a first step.

Yours faithfully,

G. HOFFMANN DE VISME,
The Manchester College of Science and Technology.

REFERENCE

1. NAIRN, D. A Simple Decimal to Binary Decoder Using Relays. *Electronic Engng. 35*, 232 (1963).

THE AIRCRAFT EXHIBITION

A description, compiled from information supplied by the manufacturers, of a few of the electronic exhibits shown at the recent exhibition of the Society of British Aerospace Companies.

(Voir page 721 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 728)

AMPLIVOX LTD

Beresford Avenue, Wembley, Middlesex

MINIATURE TRANSCEIVER

(Illustrated below)

The Televox is battery operated, completely portable and fits easily in the pocket. This will be one of the smallest v.h.f. 'walkie talkie' sets to be made in this country and will find many applications where communications between dispersed personnel is required.

In the design of this instrument, special attention has been paid to the differing requirements of various industries and several interesting features will be incorporated:

(1) Each transceiver will include circuits for up to four crystal controlled channels. This provides communication between operators on the same frequency as well as the facility to allocate different channels to other personnel in the same area.

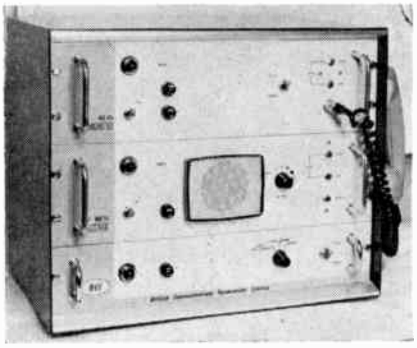
(2) The Televox will be supplied with a detachable battery compartment. An alternative compartment (slightly larger) will also be available when exceptionally long periods of use are envisaged. These battery units fit quite simply on the base of the radio section. Alternatively the battery unit can be replaced by a plug/adaptor connecting to a separate power supply.

(3) The Televox will incorporate an internal speaker/microphone and will, therefore, be completely self-contained.
By simple switching it can be used with an external hand microphone/speaker unit such as the Amplivox Mini-Mike (as illustrated), or under very noisy conditions, in conjunction with a noise excluding headset, such as the Amplivox Ampligard series.

(4) A telescopic aerial is incorporated, and provision is made for connexion to an external aerial for increased range.

Accessories include a high quality leather case, crystals, and a range of microphones, listening units and headsets.

EE 74 751 *for further details*



BRITISH COMMUNICATIONS CORPORATION LTD

South Way Exhibition Grounds, Wembley, Middlesex

U.H.F. RADIOTELEPHONES

(Illustrated above)

With the ever increasing call for new services already necessitating channel sharing in the overcrowded v.h.f. bands, BCC are anticipating the use of the u.h.f. band by introducing new mobile and fixed station radiotelephone equipments which were shown for the first time.

The BCC 55 is a six channel 10W p.m. mobile equipment. Its associated fixed station (illustrated), the (BCC415/115) transmitter receiver combination, is a 15W p.m. station. Both equipments are designed for operation in the 440 to 470Mc/s band.

EE 74 752 *for further details*

V.H.F. RADIOTELEPHONES

(Illustrated above right)

This range of lightweight portables covers all the v.h.f. bands and both a.m. and p.m. versions are available. Each set provides a choice of six channels and has a loudspeaker included which can also serve as a microphone. A dry battery pack is available in addition to an already extensive list of accessories. This contains 9 standard U7 dry cells and is an alternative to using a rechargeable nickel cadmium battery. The dry battery pack, being exactly the same dimensions as the rechargeable battery is a direct replacement.

The frequency range covered is 41 to 174Mc/s with either 25 or 50kc/s channel spacing. The transmitter output power is between 120 and 500mW depending on frequency and type of modulation: the frequency stability is better than ±2kc/s.

The maximum audio output of the receiver is 50mW and the selectivity such as to provide a response better than 3dB within ±7·5kc/s and better than −60dB beyond ±25 kc/s (with 25kc/s channel spacing).

EE 74 753 *for further details*

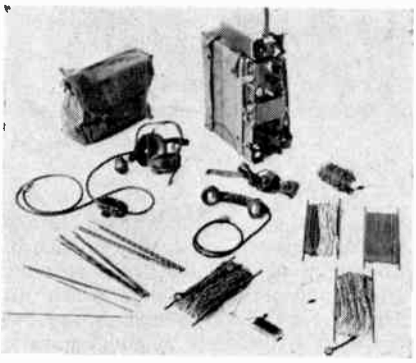
H.F. PACKSET

(Illustrated on page 714)

The BCC30 HP is an entirely transistorized h.f. packset combining a high power (20W) with a low all-up weight of only 35 lb. The low weight of the BCC30 makes it ideal for use by paratroops to provide the long range contacts which cannot be satisfied by v.h.f. sets. The main features of the set apart from its low all-up weight for a complete 20W portable station, are its comprehensive facilities. C.W. as well as a.m. and p.m./r.t. are provided with 18 crystal controlled channels within the 2 to 8 (or 2·5 to 10) Mc/s band. Free tuning is also provided to cover the entire band with a selectivity allowing 600 speech channels. A full range of lightweight

ELECTRONIC ENGINEERING

will occupy Stand 42 at the R.E.C.M.F. British Electronic Component and Instrument Exhibition to be held at Ostermans Marmorhallar, Stockholm, from 13 to 16 October 1964.



supporting accessories includes a hand generator power supply, a.c. and d.c. operated battery chargers and vehicle installation kits. The small size of the BCC30 and its very low power consumption make it an attractive multi-role set for use as a packset, fixed station or mobile unit.

EE 74 754 *for further details*

A. C. COSSOR LTD

The Pinnacles, Elizabeth Way, Harlow, Essex

Airborne Transponder Units

(Illustrated below)

In addition to a complete secondary surveillance radar ground station, two airborne transponders were shown. These were the military type 1500 and the civil type 1600.

The 1500 miniaturized i.f.f. transponder, developed under MOA contract, has been designed for continuous unblown operation over the extremely wide temperature range −55°C to +140°C and the high m.t.b.f. factor sets new standards in reliability for aircraft equipment.

The 1600 is built to the standards of ICAO Annex 10 and ARINC Characteristic 532D and is designed to conform to the type approval requirements of the Air Registration Board and the United States FAA, with 12 bit encoding facilities on modes A, B, C and D. The equipment is fully transistorized with the exception of the final stage of the transmitter which provides power output in excess of 1kW. The unit is designed for continuous operation within the temperature range −40°C to +55°C with convection cooling and for unpressurized operation up to 40 000ft. There is a built-in self test facility, operated from the control panel in the crew compartment.

EE 74 755 *for further details*

Miniature Radiotelephone

(Illustrated below)

The 'Companion' range consists of a series of personal radiotelephone equipments designed to be carried or worn by the operator. Included in the range are the CC2/8 v.h.f. f.m. pocket set, the CC3 a.m. packset and the new CC4 f.m. six channel light weight packset.

Weighing only 1 lb (approx.) and measuring $5\frac{3}{4}$in by $3\frac{1}{4}$in by $1\frac{1}{4}$in the CC2/8 has a frequency range from 71·5Mc/s to 174Mc/s at a transmitter power of 500mW. The receiver sensitivity is 0·6µV for 20dB quieting and the spurious response level is better than −60dB. With an audio power output of 200mW the equipment can be used in areas of high noise level. By using the basic transmitter-receiver with various accessories, different configurations are possible to meet individual requirements and power output may be adjusted to suit operational requirements in terms of battery life.

EE 74 756 *for further details*

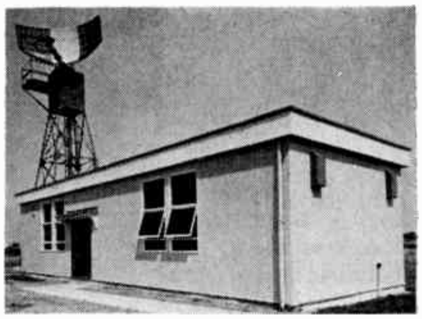
DECCA RADAR LTD

Albert Embankment, London, S.E.1

A.T.C. Surveillance Radar

(Illustrated above right)

The AR1 air traffic control surveillance radar is a 10cm high definition, high data rate equipment which is already in quantity production for the Royal Air Force and for civil aviation authorities. The AR1, which is capable of fulfilling all a.t.c. roles within 75 miles of a terminal, has been designed specially to meet operational requirements at airports where the very high cost of installing major surveillance radars, backed up by close range equipments, cannot be justified. By placing design emphasis from the outset on versatility, low installation cost and low day-to-day operating cost, Decca have produced in the AR1 a compact surveillance radar capable of meeting the growing civil demand for a practical radar system for use wherever high performance air liners are operating. A stable transistorized m.t.i. system with no blind speeds below 560 knots, and a variable polarization



system of very advanced design, enable the radar to exceed substantially ICAO requirements in all close range roles down to p.p.i. approach, at the same time providing a surveillance potential to about 100 miles at 40 000ft, depending on aircraft size.

The AR1 radar has been fully evaluated both technically and operationally by the Royal Air Force.

EE 74 757 *for further details*

Display Units

(Illustrated below)

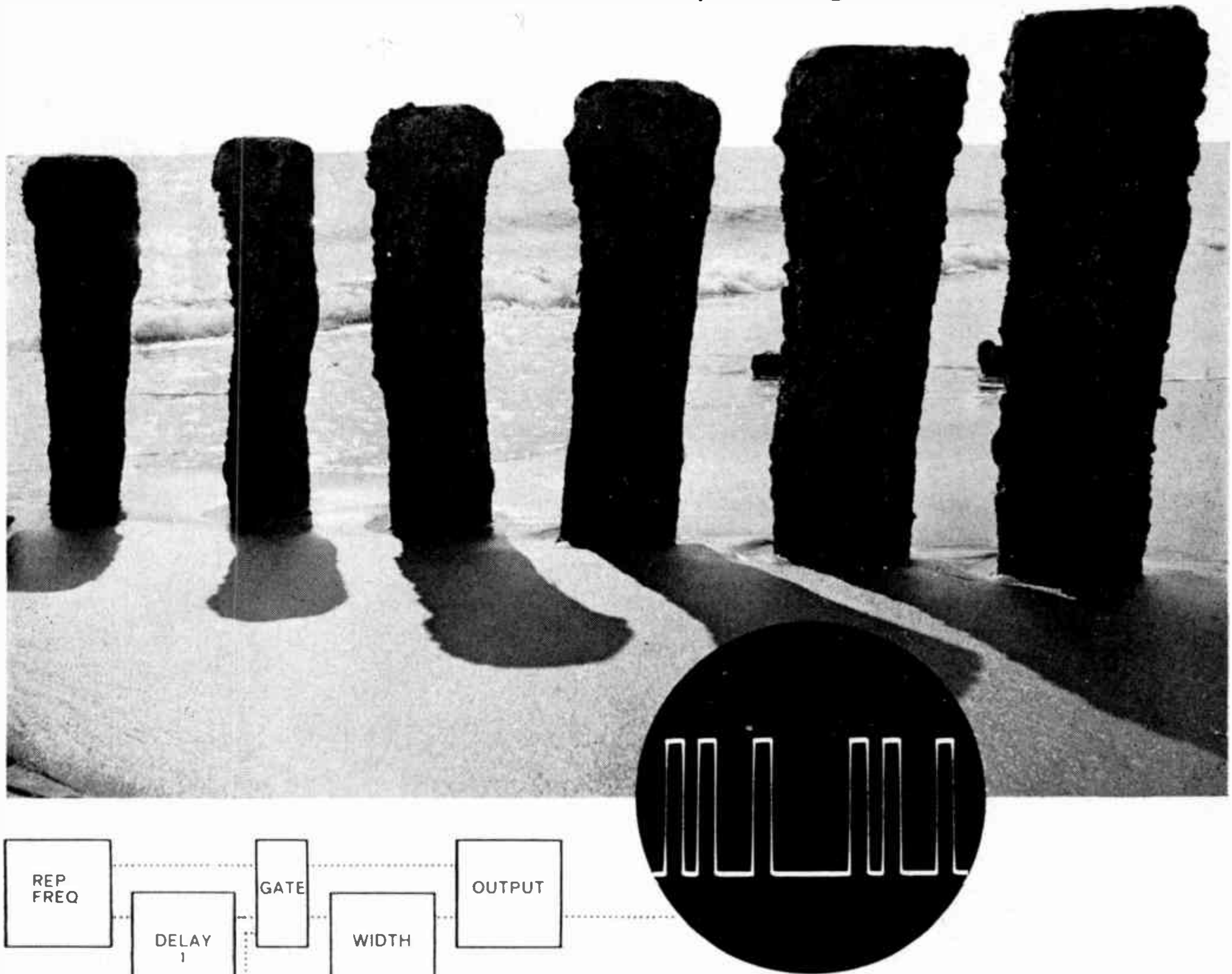
A live demonstration was given of an integrated air traffic control system in which Decca Navigator data, transmitted via an air/ground data link, was presented simultaneously with surveillance radar data on Decca series 5 autonomous display units. At Farnborough the display unit operated from the station surveillance radar while an air/ground Decca Navigator link was located immediately outside the main exhibition tent.

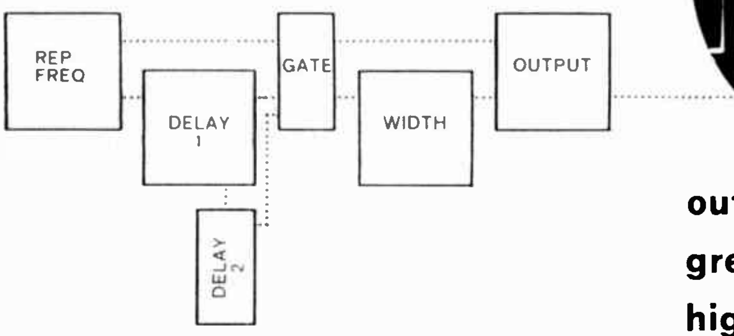
The radar display unit employed for this new air traffic control presentation is a type 12C display from the Decca series 5 range of autonomous transistor display units. The facilities available comprise a normal p.p.i. presentation with off-centring up to eight radii, three characteristically shaped markers whose position is determined by data relayed from airborne Decca Navigator receivers, and one marker symbol under local control at the display position.

The use of a standard display unit for this comparatively complex presentation is possible because of the versatility of the series 5 design. Operational requirements ranging from simple p.p.i. or r.h.i. presentations to those demanding use of electronic marking symbols, range and

EE 74 133 *for further details*

a new approach to pulse generation



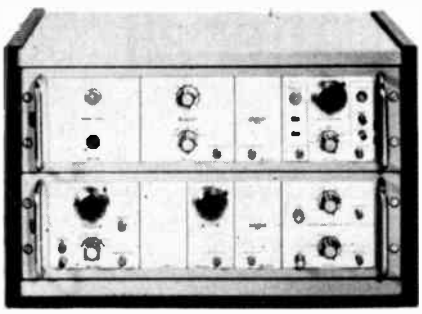


outstanding quality

great flexibility

high accuracy

Modular Pulse Generator type PM 5720-40



Fully transistorised
Pulse frequency 10 c/s – 10 Mc/s
Delay/pulse width 10 ns – 1 s
Output 5 V across 50 Ω
Max. attenuation 1000x
Rise/fall time < 10 ns

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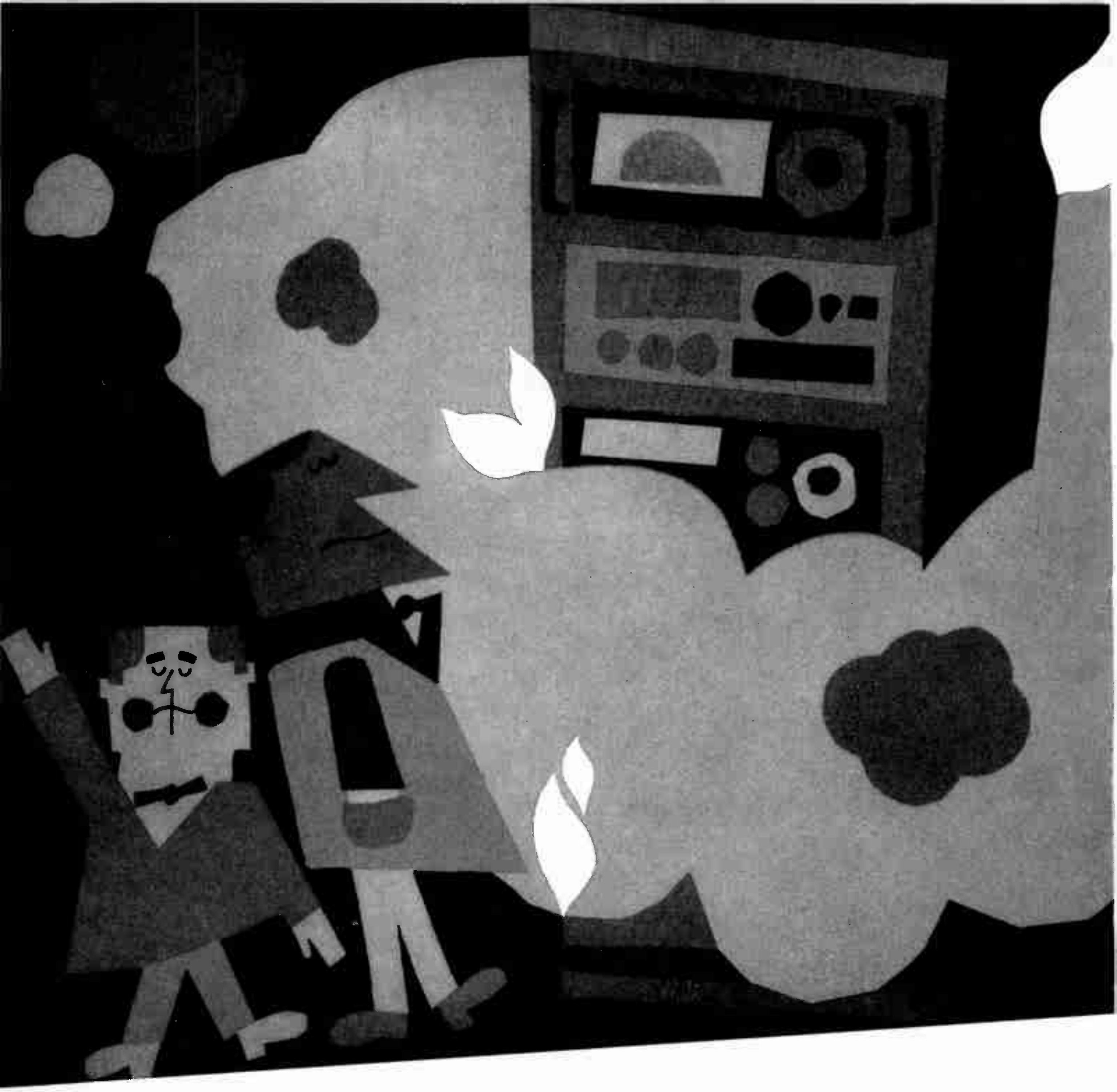


bearing measurement lines, direct v.d.f. presentation on the cathode-ray tube, interconsole marking, and similar data handling aids can all be met from this versatile range, comprising some 100 variations. Another recent example of the use of this equipment is in the Eurocontrol simulator project for which Decca, Telefunken of Berlin and CSF of Paris are jointly responsible. The simulator installation at Bretigny will employ a special console version of these displays as the primary display system for both 'pilots' and controllers.

EE 74 758 *for further details*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

Automatic V.H.F., D.F. Equipment

(Illustrated below)

This high accuracy, low cost system consists of three units: the display unit, a rack assembly and the aerial unit.

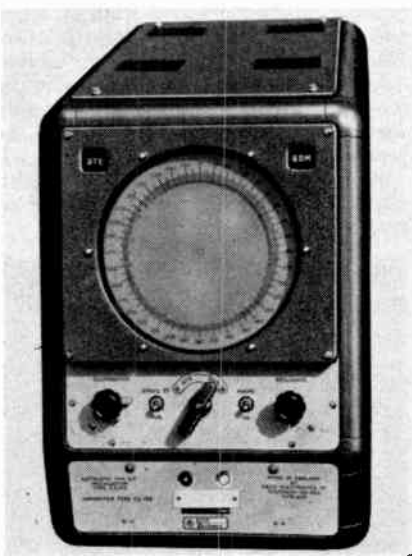
The indicator unit houses a 6in c.r.t., on which the bearings are displayed, and the operational controls. Edgewise illumination of the scale permits operation in darkened conditions.

The rack assembly houses the main equipment and is normally situated in a suitable hut at the base of the aerial tower. It comprises a receiver, two units for deriving display pulses from the aerial bearing information, a remote control termination unit and two power units.

The Adcock aerial with sense element is mounted on a tubular steel tower at a height of 28ft. It is continuously rotated by a synchronous motor at a speed of 250 rev/min. The housing for the motor also contains an inductive coupling unit for the aerial down lead and a magslip transmitter coupled to the aerial shaft. The tower is hinged at the base for simple lowering for maintenance.

The method of operation is as follows.

An Adcock aerial is rotated continuously at a speed of 250 rev/min and its



output is combined with a small portion of an omnisignal from a dipole mounted centrally between the Adcock elements.

As the aerial polar diagrams approximate a 'figure of eight' in which the nulls are not in exact diametric opposition the spacing of successive nulls during reception is equivalent. alternately, to slightly more or slightly less than 180° of aerial rotation.

These nulls appear in the output of the receiver second detector but are, at this stage, unsuitable for sensing or display because of variations in sharpness due to received signal strengths and other factors. They are, therefore, applied to a pulse generator circuit in which corresponding narrow pulses of constant width are formed. This circuit consists essentially of a single stroke time-base which is triggered at a predetermined signal voltage before the null and runs at half speed until the signal attains the same pre-determined level after the null. At this point the run-down is continued at full speed until the time-base bottom limit is reached and flyback is initiated.

A narrow pulse is formed from the flyback, with a constant delay after the mid-point in time between the instants at which the signal reaches the same voltage before and after the null. The timing of the pulse so generated is, therefore, unaffected by null sharpness within the working range of the equipment, provided that the signal voltage about the null is symmetrical. The effect of the slight asymmetry required for sensing is minimized by confining the operation of the circuit to the immediate null region. Broad nulls, which are not within the working range of the equipment, are excluded by a width gate.

The narrow pulses leaving the pulse generator unit are applied to a resolver transmitter coupled to the aerial shaft and thereby split into sine and cosine components relative to aerial position. These components are applied to the scan coils of a cathode-ray tube so that deflexions result in a substantially radial direction dependent upon the angular position of the aerial at the instant a pulse occurs.

The scan coils are suitably orientated relative to the bearing scale around the c.r.t. screen to that aerial position at that instant and hence the direction of the signal source is indicated.

To obtain sensed bearings, the generated narrow pulses are applied separately to a circuit controlling the c.r.t. brightness, which functions to eliminate pulses that are preceded by intervals of less than 180° of aerial rotation.

EE 74 759 *for further details*

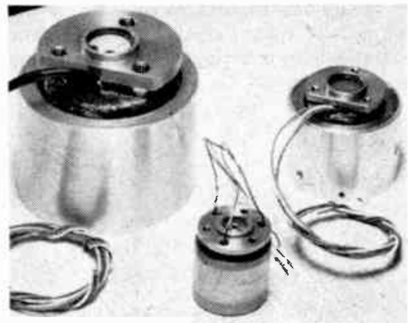
FERRANTI LTD

Hollinwood, Lancashire

External Rotor Motors

(Illustrated above right)

A new range of 400c/s d.c. external



rotor motors was shown which is broadly in line with Ministry of Aviation Spec. EL. 1892. The motors were designed originally as gyroscope spin motors, but have found applications in several other fields such as small blowers, combined blower/power sources, drum drives, powered pulleys and powered swash plates. They are very efficient enabling a high output/volume ratio to be obtained, which is valuable where space, weight, temperature or limited available power are problems.

EE 74 760 *for further details*

Moving Map Display

The moving map display presents to both pilot and navigator an optically-projected image of a map over which the aircraft is flying. Aircraft present position and track are indicated on the ground-glass screen of the unit as a central circle and radial line. The map image, which is projected optically on to the screen from a colour film strip of the appropriate topographical charts, moves behind these indications. The reduction used for micro-filming the charts is 20 diameters, the optical system of the display being so designed, that when the film is projected, it is magnified 20 times.

The film strip is driven along its length for change of Eastings and across its width for change of Northings using servo motors on each axis which take their inputs from the Doppler or inertial navigation system via the navigation computer and an electromechanical analogue computer. The displays provide both the pilot and navigator with an easily interpreted pictorial representation of the situation during the progress of a sortie. A present position marker constantly shows where the aircraft is, and the radial track line shows what points the aircraft will fly over. In good weather, the pilot can use his display to map read his route, relating the image on his display to the configuration of the terrain he sees. He can make visual fixes and also make a check on the correct functioning of the navigation system.

The navigator uses his moving map as a situation display presenting navigation information and as a monitor of the navigation system. He can also compare the output from the 'sideways looking' radar with his display in order to take a radar fix. In this mode, the display helps him to anticipate the appearance of radar fixpoints, and to recognize them as they appear.

EE 74 761 *for further details*

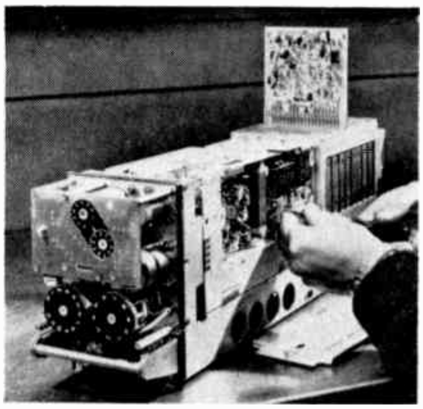
THE MARCONI CO. LTD

Chelmsford, Essex

Distance Measuring Equipment

(Illustrated below)

The AD70 airborne d.m.e. interrogator continuously and automatically measures the slant range from an aircraft to a selected ground beacon. This distance is displayed directly on a meter on the pilot's instrument panel. The ground transponders used with this type of equipment may be either VORTAC or TACAN installations, both of which fulfil other navigational roles in addition to the d.m.e. function.



The distance information, when combined with a single v.o.r. or radio-compass bearing, allows the aircraft position to be accurately fixed. In the majority of ground installations, a v.o.r. type of beacon is co-sited with the d.m.e. transponder, as in the VORTAC equipment.

A d.m.e. beacon sited at the end of a runway can also provide a valuable additional landing aid, providing a continuous reading of 'distance to go' to the touchdown point.

The AD70 distance measuring equipment comprises 17 separate modular assemblies, mounted on a ½ ATR aluminium main frame, with a total weight of 31 lb. Transistors are used in this equipment, except where high power levels or other design considerations preclude their use. Plug-in printed circuit modules are also used extensively, with extension boards to simplify routine operational testing of the equipment. One hundred and twenty-six crystal controlled frequency channels are provided, although this number can be increased to 252 should the d.m.e. system be expanded at any time in the future.

Frequencies are set by decade switches on a separate controller, and once the selected frequency has been set in, the action of the equipment is entirely automatic.

A continuous succession of interrogating pulse pairs is sent out by the transmitter and each one causes the ground transponder to transmit reply pulses after a fixed time interval. The distance of the aircraft from the ground beacon can then be measured from the time interval between each interrogation pulse pair and its reply. The equipment determines this time interval by means of a time sensitive receiving gate which will only pass signals received from the ground transponder during a discrete time interval after the transmission of a pair of interrogating pulses. The separation between this time interval and the transmission is varied automatically from zero to the time corresponding to 200 miles. When a response falls within this gate the equipment 'locks on' to this signal and then changes the position of the gate at the rate necessary to track the incoming signals as the aircraft moves towards or away from the ground beacon. The distance corresponding to the instantaneous measured time intervals is displayed simultaneously on the meter in the pilot's instrument panel.

An important feature of the AD70 is the velocity memory circuit which ensures that any momentary loss of signal does not cause a break in the distance information. The velocity memory supplies a simulated signal to the tracking circuits for as long as a true signal has been received, up to a maximum of ten seconds. If the signal has not reappeared by this time, the equipment will revert to the standby mode, and start to search for a signal throughout the full distance range.

EE 74 762 *for further details*

Solid State Airborne H.F. Equipment

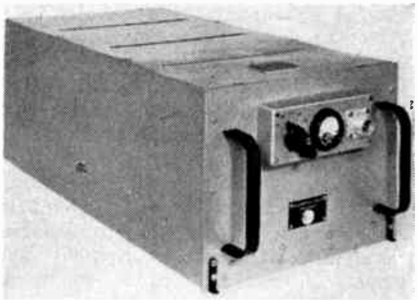
(Illustrated above right)

The h.f. communications system, type AD460, follows the concept of engineering for reliability associated with all the Marconi Sixty Series equipment. The latest design techniques have been employed throughout, and at all stages of both the design and the manufacture of each unit, stringent tests were applied to ensure absolute reliability.

The AD460 is tunable throughout the frequency band 2 to 29·999Mc/s by decade switches on a separate controller unit. Channel spacing of 1/4kc/s will provide for all possible future needs for extra channels in the densely crowded h.f. band.

Semiconductor diodes are used for all switching functions and solid state tuning is employed throughout, except in the final power output stage. Operating frequencies are generated by the frequency synthesizer method employing decade oscillators individually phase-locked to the stable-frequency reference oscillator. The use of silicon transistors throughout, enables the equipment to operate reliably over a temperature range from −55°C to +55°C.

Modes of operation include—double sideband providing r.t. and c.w.—single sideband with suppressed-carrier r.t.—



and single sideband (floating carrier and a.f.c.) providing r.t. or two-tone f.s.k. for data links. Provision has been made for the operation of SELCAL in accordance with ARINC Characteristic 531 and separate input and output connexions are available for use with data-link equipment. Separate data i.f. and discriminator stages are also included for data links. Narrow or broadband c.w. transmission is provided for and a separate beat frequency oscillator is available in this mode of operation.

The AD460 has an output power of ¼kW although a 1kW version, type AD470, is also available. A comprehensive range of test facilities is incorporated to permit a rapid checking in all modes of operation.

EE 74 763 *for further details*

MARCONI INSTRUMENTS LTD

St. Albans, Hertfordshire

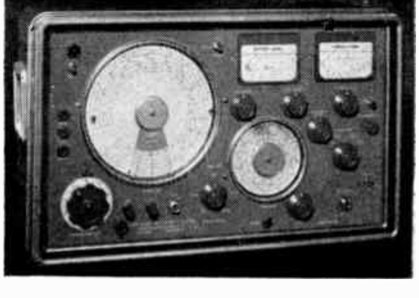
A.M. Signal Generator

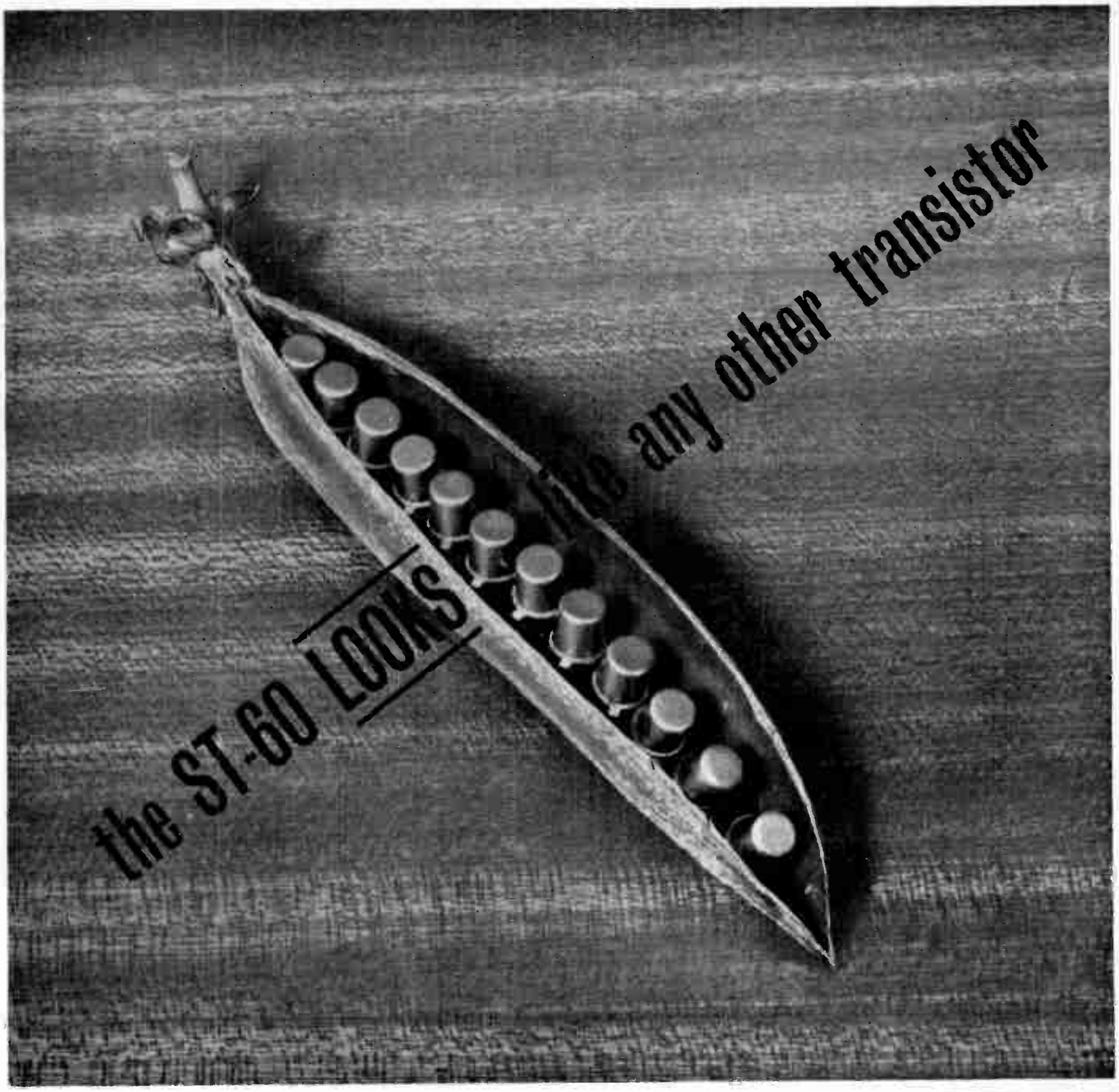
(Illustrated below)

The a.m. signal generator type TF 801D/SM1 is generally similar to the standard model TF 801D/1 (10 to 470Mc/s), but special features are incorporated to adapt it to testing certain types of air navigational aids, and the frequency coverage is altered to provide better discrimination in the appropriate bands. The output is 0·1μV to 1V, e.m.f., continuously variable, and frequency drift is less than $\pm 5 \times 10^{-5}$/ten minutes.

In order to cater for v.o.r. requirements, the external modulation circuits produce negligible phase shift at 30c/s, and the demodulated waveform is available at the front panel from an internal detector for comparison purposes, while the v.h.f. carrier band is approximately centred in the fourth range of the generator (70Mc/s to 150Mc/s).

A crystal-controlled oscillator is also provided internally, for use when an accurate 75Mc/s signal is required, e.g.




the ST-60 LOOKS like any other transistor

tor marker beacon receivers. This is normally used with the third range of the generator (48Mc/s to 110Mc/s), the variable oscillator being switched off, but with the amplifier still functioning. A high level signal capable of deep modulation is obtained. The fourth range amplifier can also be used for the same purpose when this is more convenient.

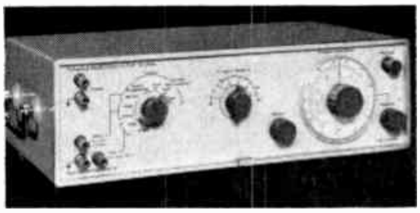
EE 74 764 *for further details*

Tunable Rejection Filter

(Illustrated below)

In addition to its use as an accessory for the Marconi wave analyser, type TF 2330, to permit the measurement of relative amplitudes differing by more than 75dB, this instrument can be employed in other applications where it is necessary to suppress or eliminate a single frequency.

The TF 2334 provides over 80dB rejection of fundamentals in the range 20c/s to 20kc/s, with minimum attenuation of harmonics. It is basically a 'twin-T' filter of high impedance, designed for use with sources of less than 1000Ω. High impedance or 600Ω input modes are available. When the latter is used, switched compensation for the



attenuation of harmonics up to the fourth is available, so that their values can be read directly from the wave analyser.

No active components are used and a power supply, therefore, is not required. The instrument will be available in bench or rack-mounting form.

EE 74 765 *for further details*

PLESSEY-UK LTD

Ilford, Essex

Flight Data Recording System

This equipment, which has been specified by BEA for fitting to its entire fleet and by Aer Lingus for inclusion in the equipment of all its BAC one-eleven jets on order, fully meets the Ministry of Aviation's mandatory requirements for the provision of a record which will survive shock, fire and chemical attack.

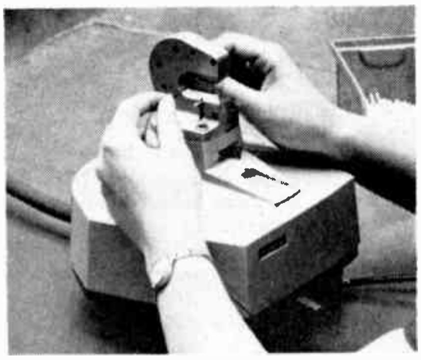
In addition to the six parameters specified by the MCA—indicated air speed, indicated altitude, magnetic heading, vertical acceleration, pitch attitude, time—the design caters for special investigation work by providing spare channels for extra sampling circuits. Digital recording of all signals guarantees the preservation of accuracy throughout the entire sequence of operations, from the recording in flight to the data processing and display on the ground, and also permits built-in self-checking facilities.

The system comprises three main units—a cockpit control unit, a processing unit and the recorder—to which are connected a number of transducers. In the recorder—which was specially developed by S. Davall & Sons Ltd—the use of magnetic recording on stainless steel wire results in far greater resistance to crash hazards than is offered by any other medium.

The recorder unit itself, which will be installed in the tail of the aircraft, is so robustly constructed that automatic ejection has been considered unnecessary. For example, it is built to withstand loads up to 1 ton, acceleration up to 100*g*, temperatures up to 800°C for 15min and total immersion in sea water for up to 30 hours.

Each recorder has a 200-hour running capacity without attention and is switched on and off automatically at the beginning and end of each flight by the air speed sensor unit.

EE 74 766 *for further details*



Crimping Tools

(Illustrated above)

Power operated crimping tools have now been introduced. These are compressed-air driven versions of the hand UNIcrimp—a universal, one-handed, automatic tool capable of making perfect square-form crimps on any size of wire or contact up to 12AWG (40/·0076in).

The power version is now offered in two forms: a portable model similar to the hand UNIcrimp, and a new bench model with a heavy base for workshop use. The power models will crimp pre-insulated tags as well as all connector contacts. In any of its three forms the UNIcrimp is a foolproof, fully-proven production machine for use by unskilled operators.

EE 74 767 *for further details*

RANK BUSH MURPHY ELECTRONICS

Welwyn Garden City, Hertfordshire

Radio Beacon

(Illustrated above right)

A radio beacon designed to aid air

support of front-line ground troops was shown for the first time. The equipment is the latest addition to the Rank Bush Murphy Electronics' internationally famous Eureka/Rebecca Navigational System.

Named the Eureka MR343, it is a transistorized low power beacon designed for parachuting with airborne assault forces to provide a distance and homing signal for support aircraft and for use as a terminal aid for forward airfields.

A striking reduction in size, weight and power requirement has been achieved and the equipment is capable of being operated by non-technical Service personnel. It has a built-in facility for checking receiver and transmitter performance.

EE 74 768 *for further details*

REDIFON LTD

Broomhill Road, London, S.W.18

Transportable Radio Beacon

The TG.142R m.f. beacon is specially designed for speedy erection in the field. The complete station comprises eight transportable items and can be set up by two men, using the 20ft derrick boom provided with the equipment to erect a 60ft vertical mast aerial.

The equipment consists of transmitter, an aerial mast in six 10ft sections with a capacitance top, twelve 110ft radial earth wires, an aerial coupling unit, a petrol driven generator, two spares containers, a tent and a derrick boom.

The beacon operates in the 200 to 420kc/s frequency range on c.w. or m.c.w. service with local or remote control. A special keying disk is used to select signal characteristics, which, including the callsign, can be easily altered to suit circumstances.

EE 74 769 *for further details*

S. SMITH & SONS (ENGLAND) LTD

Kelvin House, Wembley Park Drive, Wembley, Middlesex

Flight Control Systems

The Series 6 system is the latest in Smiths range of flight control equipment and is already specified for Fokker's Turbofan Transport—the F28 Fellowship.

The series 6 range comprises the SEP.6 Autopilot, SFS.6 Flight System, SCS.6 Compass System and the STS.6 Automatic Throttle Control System. Designed on the 'building-brick' principle, the system can be extended from a simple autostabilizer or heading reference to the complete automatic and instrumental control necessary for operation in reduced weather minima.

Series 6 equipment conforms to ARINC recommendations and will meet the flight control requirements of aircraft ranging in size from the executive to the large public transport jet airliners.

Safety and reliability have been the dominant factors in the development of the Series 5 flight control system which satisfies civil requirements for all-weather operation and automatic landing.

In addition to autolanding, the system provides the widest range of autopilot control and flight director modes which include automatic turn and lock on to a pre-selected heading: automatic acquisition of pre-selected height; height lock; rate of descent lock; i.l.s. and v.o.r. coupling with automatic 'over-station' sensor; automatic or manual acquisition of i.l.s. glidepath and descent along the glidepath.

Fitted initially at duplex level to the Hawker Siddeley Trident, the Series 5 flight control system gives the aircraft an automatic flare-out capability. In due course, the system will be extended to triplex level for full autolandling. The Trident is the first aircraft in the world to be equipped in this way. The new Short Belfast autolanding freighter, on order for Britain's Royal Air Force is fitted with Series 5 equipment at triplex level.

EE 74 770 *for further details*

THE SOLARTRON ELECTRONIC GROUP LTD

Farnborough, Hampshire

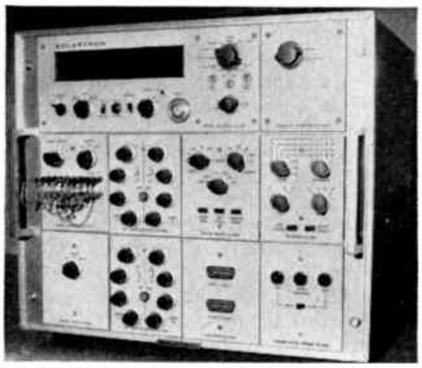
Digital Data Logger

(Illustrated above right)

The 'compact logger' is a small 20 channel digital data logging system having a maximum scanning speed of 3 points per second. It incorporates the integrating digital voltmeter LM 1420 which has an accuracy of 0·05 per cent at a sensitivity of $10\mu V$ per digit. The voltmeter uses a voltage-frequency convertor which permits a fully isolated input circuit. As a result, a common mode rejection of 150dB is achieved and with the input filter in use, a series mode rejection of 60dB is obtained.

The compact logger is modular in construction and may be supplied in a two or three-tier case depending on the facilities required.

The plug-in modules available include a command range unit for the digital voltmeter. 20 channel scanner, digital clock, punch encoder, printer drive, typewriter drive, off-limit detector, linearizer and thermocouple reference. These units may be added as required, being replaced by blank panels when not fitted. Each unit has a specific position in the cabinet.



The illustration shows the three tier version with a full complement of subunits fitted.

EE 74 771 *for further details*

High Resolution Video Map

The Solartron high-resolution video map may be integrated with any standard radar system. It is claimed to be the most accurate map currently available. A micro-spot cathode-ray tube with a spot size of $1{\cdot}5 \times 10^{-3}$in and an 'A' phosphor is used for scanning. The tube is clamped at the neck by a special collar which provides correct reference to the plane of the tube face and simplifies focusing.

Deliberate distortion in the deflexion amplifiers compensates for non-linearity due to the flat faced tube thus simplifying map production. A very high standard of draughtsmanship and photography are used to ensure the required accuracy.

The overall specification is as follows:

Maximum range: 40 to 400 nautical miles as required. Resolution: 1 part in 1 000 (making an expansion of over twenty times perfectly usable). Accuracy: On a map of 100 miles radius a point at 10 miles range is displayed with a positional accuracy of better than 200 yards.

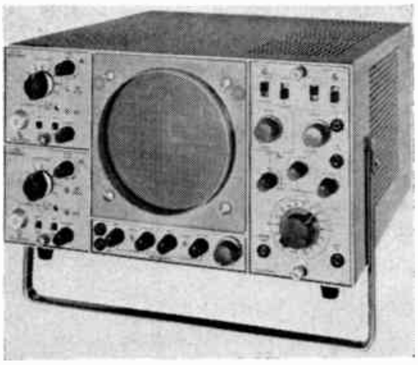
EE 74 772 *for further details*

Double Beam Oscilloscope

(Illustrated above right)

The new Solartron double beam oscilliscope system CD 1 400 has a 5in double gun cathode-ray tube which gives a large, bright display. Both X and Y systems plug into the instrument and drive the tube plates directly so that no built-in plate-drive amplifiers are required in the main frame. Four plug-in units are currently available and more are being developed. The wide-band amplifier CX 1441 has a −3dB bandwidth of d.c. to 15Mc/s with a sensitivity range from 100mV/cm to 50V/cm. Additional gain ×10 is available on all ranges at reduced bandwidth.

The high-gain differential amplifier CX 1442 has a −3dB bandwidth of d.c. to 75kc/s with a sensitivity range from 1mV/cm to 5V/cm. Additional gain ×10 is available, a.c. coupled only, at re-



duced bandwidth. Common mode rejection is approximately 60dB.

The time-base unit CX 1443 offers a velocity range from $0{\cdot}5\mu$sec/cm to 200msec/cm with expansion of ×5 approximately, available if required. Comprehensive triggering and h.f. sync facilities are provided from Y_1, Y_2 or external signals with choice of polarity.

The sweep-delay time-base unit CX 1444 is similar to the CX 1443 but incorporates a sweep delay time-base with maximum delay time of 100msec. The measurement of amplitude and time is accurate to ±5 per cent on most ranges.

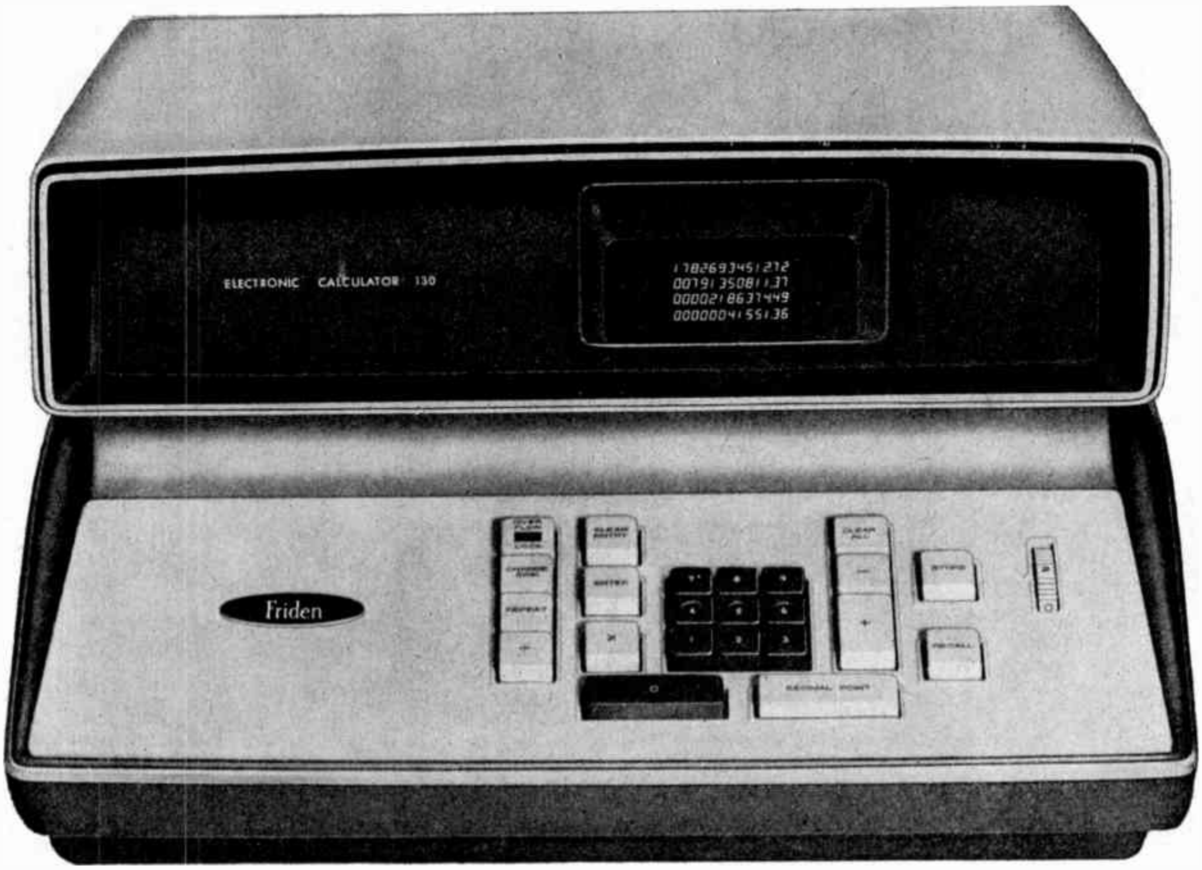
The illustration shows the CD 1400 fitted with two wideband amplifiers and a time-base unit.

EE 74 773 *for further details*

A.T.C. Simulator

(Illustrated below)

A four-target air traffic control simulator was demonstrated using an airspace surveillance p.p.i. radar display. This simulator is one of the SY2020 series of equipments now in production for current orders. Rationalization of module design and the advantages of batch production have enabled costs to be reduced significantly, while certain engineering improvements have increased accuracy and reliability. An example of this is the method employed for integrating target positions. The new highly-accurate electronic integrators allow, for the first time, a target response to be located in s.r.e. at a range of up to 75 nautical miles and then to be brought in for close approach on p.a.r.



Most advanced *electronic* calculator in existence?

Probably

Most beautifully designed?

Certainly

The new Friden* 130 Electronic Calculator is mid-way between a mechanical desk calculator and an electronic computor. It has no moving parts — all calculations are performed by solid-state electronic components, and entries and answers are displayed on a cathode ray tube like a miniature T.V. screen. It works in milliseconds and it works silently.

Operation of the 130 is through a very simple keyboard with clearly marked controls. Results appear on four registers making it possible to work on multistop problems in logical order without manually re-entering amounts. All entries and answers are automatically aligned around a selected decimal point. In classic simplicity and elegance of design, it is unmatched among today's office machines. On display at the Business Efficiency Exhibition Stand 71/73 at Olympia, London, 6th – 14th October, 1964.

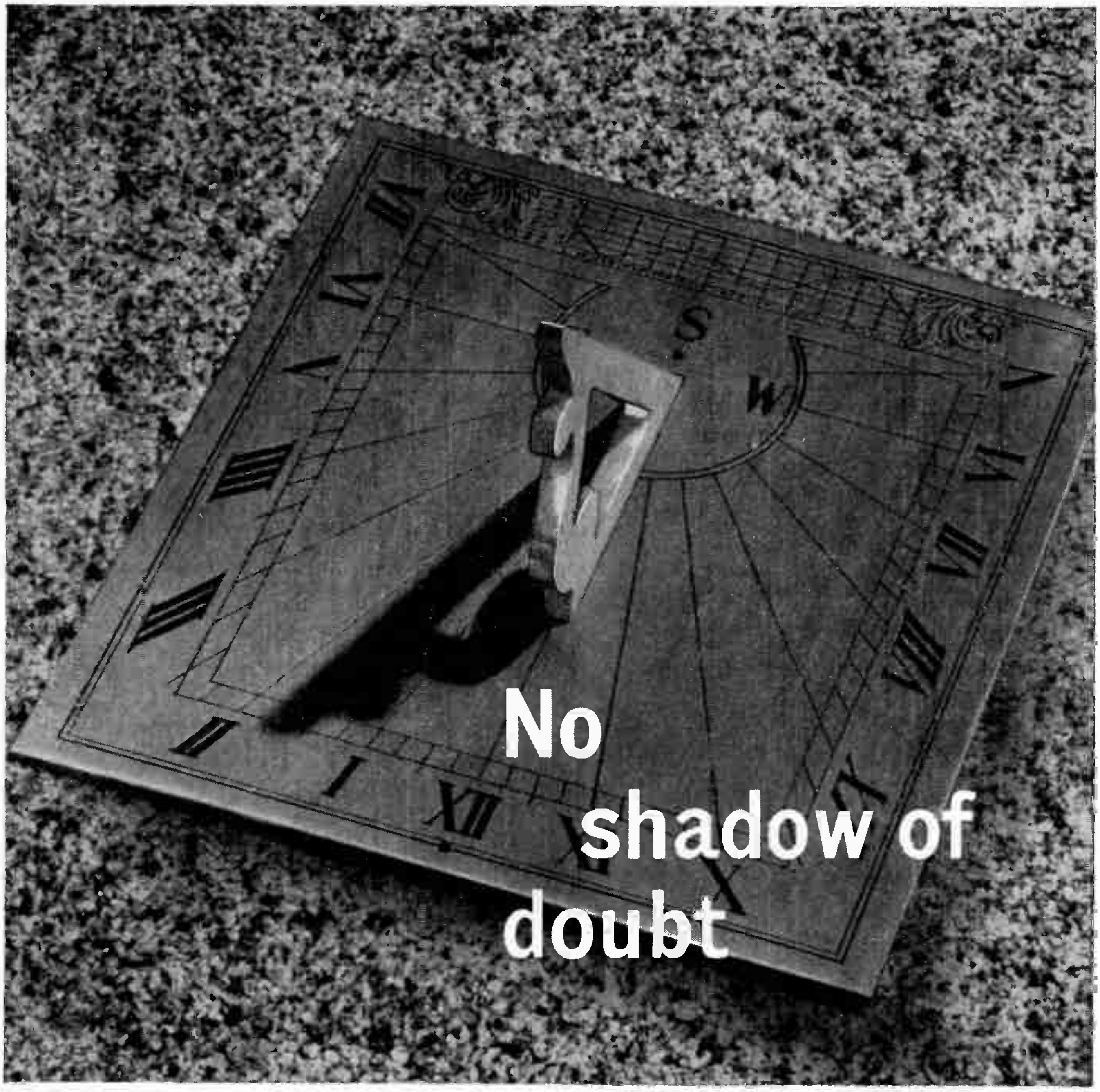
Friden *Limited*

Friden House, 101 Blackfriars Road, London, SE1.
WATerloo 1301.

Branches at: BELFAST, BIRMINGHAM, BRISTOL, CARDIFF, COVENTRY, EDINBURGH, GLASGOW, HANLEY, HITCHIN, LEEDS, LIVERPOOL, MANCHESTER, NEWCASTLE, NOTTINGHAM, PLYMOUTH, SHEFFIELD, SOUTHAMPTON.

* A Trademark of Friden Inc.

EE 74 138 *for further details*

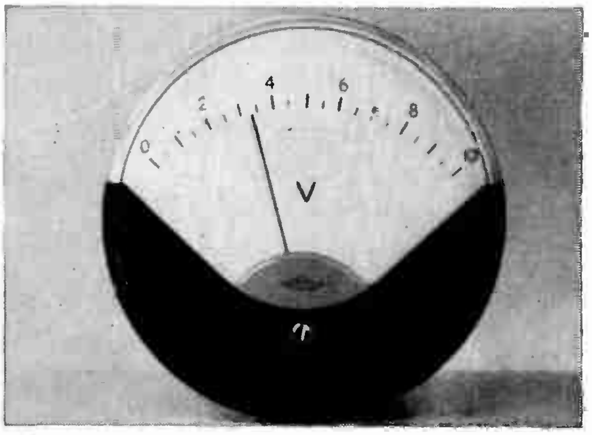


The Pullin Clear View range gives readability beyond the shadow of a doubt; because there can't be shadows. The case is translucent, the dial designed to be seen . . . not blurred.

one of the range

The Series 38 CY (3.8" scale) *Approximate viewing distance 4 ft. 6 ins. with moving iron or internal magnet moving coil movements.*

See for yourself, write for details of the Clear View range.



PULLIN FOR PRECISION

MEASURING INSTRUMENTS (PULLIN) LTD.

Head Office: *Electrin Works, Winchester Street, London, W.3. Telephone: ACOrn 8801/5*

London Showrooms: Electrin House, 93-97 New Cavendish Street, London, W.1. Telephone: LAN 4551-6 with sufficient accuracy over the entire range to give utmost realism.

The illustration shows procedure training on the p.a.r. equipment associated with the Royal Air Force Shawbury Simulator. The aircraft is at eight and a half miles of final approach, a little right of the extended centre-line, and not yet in the glidepath. The target Control Unit can be seen behind the instructor's arm.

EE 74 774 *for further details*

STANDARD TELEPHONES & CABLES LTD

Connaught House, 63 Aldwych, London, W.C.2

SECONDARY SURVEILLANCE RADAR

The SGR.1 is the ground interrogator element of a secondary surveillance radar system. It is designed to operate either independently or in conjunction with a primary radar to provide a correlated display for air traffic control.

The system provides for two basic configurations depending upon whether the aerial is mounted coaxially on a primary radar aerial or as a separate system with its own turning gear. In the latter case, a turning motor, a motor generating set and a servo control cabinet are provided.

The interrogator equipment is housed in a single cabinet containing a transmitter and a receiver together with the associated power supply equipment, control and monitoring units.

The transmitter employs a single grid pulsed travelling wave tube of high gain to amplify all mode pulses which are then fed to a high power coaxial semiconductor switch. This switch, which has a very short operating time, feeds the individual pulses, via preset r.f. attenuators, to either the interrogate or to the control elements of the aerial as required.

This method of deriving the radiated interrogate and control pulses from one transmitter via passive networks ensures long-term pulse amplitude stability and obviates the need for balancing separate transmitters. In the interest of long-term stability and reliability the equipment employs solid state devices throughout with the exception of the travelling wave tube.

The aerial system consists of two aerials mounted as a single assembly.

The equipment can be supplied either as a single or dual system. In the latter case changeover facilities are provided to interconnect the aerial and the interrogator cabinets, for local control and supervision, remote control and supervision, and aerial drive and synchronization.

EE 74 775 *for further details*

MICRO-MINIATURE RADIO ALTIMETER

(Illustrated above right)

To improve further the reliability of the low-level frequency-modulated radio altimeter for use in automatic landing

systems, STC has produced a laboratory model employing micro-miniaturization techniques. Although this development was conducted primarily in an attempt to improve reliability, the saving in size, weight and power consumption is significant.

The integrity of the existing range of STC radio altimeters is well established. They have been used to provide height information for all-weather landing systems from the inception of the British Ministry of Aviation's Blind Landing Experimental Unit to the recent installations in the Trident, VC.10, Belfast and numerous other civil and military aircraft. Throughout this period the Company has constantly striven to improve the design techniques, the manufacturing methods and the conditions under which the equipments are produced. This work has culminated in the miniaturized radio altimeter which was demonstrated at the Air Show.

This altimeter consists of a 4 300Mc/s f.m. solid state transmitter, a video amplifier and associated circuits and the frequency counter circuits. The equipment presents an ideal subject for the application of micro-electronics.

The model exhibited is housed in a short 1/4 ATR pigmy case. This is not necessarily the final form of packaging but it serves as a demonstration of what can be achieved by the use of modern techniques.

The photograph shows the micro-miniature altimeter by the side of a conventionally constructed altimeter.

EE 74 776 *for further details*

VENNER ELECTRONICS LTD

Kingston By-Pass, New Malden, Surrey

WIDE RANGE OSCILLATOR

(Illustrated below)

The TSA 625 wide range oscillator

covers 10c/s to 1Mc/s in five switched ranges. Features of the unit include push-button range selection and battery test, attenuator settings covering 2·5mV full-scale to 2·5V and output impedance better than 600Ω on all ranges.

Both sine waves and square waves are provided.

EE 74 777 *for further details*

WESTLAND AIRCRAFT LTD

Saunders-Roe Division, Strain Gauge Department, East Cowes, Isle of Wight

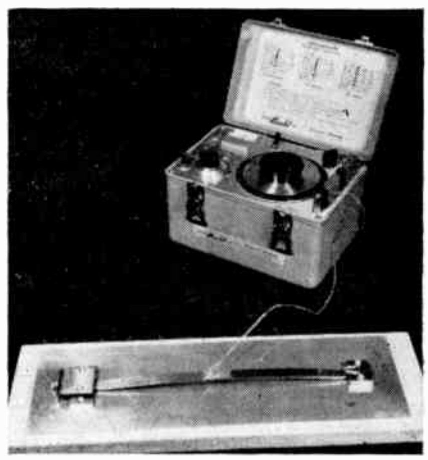
STRAIN GAUGED EXTENSOMETER

(Illustrated below)

The Strain Gauge Department of Saunders-Roe Division of Westland Aircraft Ltd has developed a simple form of strain gauged extensometer for surface mounting on large structures. It consists of a strip of steel, or other suitable material, having ball ends which is sprung into position between end fittings attached to the structure. Any deflexion over the extensometer length causes a change in the amount of bending in the strip which is measured by strain gauges attached at its mid point. The strain gauge output is calibrated directly in terms of change in the effective length of the strip. The nominal length of the extensometer is twelve inches but it may be made with other lengths to suit particular applications.

The device has advantages over rigidly attached extensometers in that it is only held in contact with the surface of the structure at the ball ends by the end load on the strip. This minimizes the effects of end rotation, local distortion and shear, such that the device will measure only the actual linear movement between the end contact points.

The end fittings which locate the strip can be attached by adhesives, screws in tapped holes, or by masonry plugs in the case of concrete or brick structures. Readings can be taken on any strain measurement equipment or galvanometer but for field use a portable, battery operated, strain indicator of the type supplied by Saunders-Roe, is most convenient.



EE 74 778 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRICAL ENGINEERS

Unless otherwise stated, all meetings will be held at Savoy Place, commencing at 5.30 p.m.

Electronics Division

Date: 5 October Time: 2.30 p.m.
Discussion: Transistorized Biological Amplifiers (Joint meeting with I.E.R.E. Medical and Biological Electronics Groups)
Date: 12 October
Lecture: Application of Electronic Vector Scanning Techniques applied to Height-Finding Landing Systems
By: D. E. N. Davies
Lecture: Frequency Scanning Aerial Systems
By: Miss E. Killick
Date: 21 October
Chairman's Address: Through Communications to Electronics
By: G. G. Gouriet
Date: 26 October Time: 2.30 and 5.30 p.m.
Colloquium: Design Automation (Joint meeting with I.E.R.E. Computer Group)
Date: 28 October Time: 6 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1
Lecture: Magnetic Circuit Design Applied to Electro-Mechanical Transducers. (Joint meeting with I.E.R.E. Electro-Acoustics Group)
Date: 29 October
Discussion: Hollow Beams
Opened by: A. H. W. Beck

Science and General Division

Date: 13 October
Lecture and Control Revolution
By: M. W. Humphrey Davies
Date: 20 October
Lecture: Anomalous Iron Losses in Cold-Reduced Grain-Orientated Transformer Steel
By F. Brailsford and R. Fogg
Date: 26 October
Lecture: Millimetre Wave Examination of Gas Discharges
By: A. L. Cullen.

Power Division

Date: 28 October
Lecture: Development in Power Engineering
By C. H. Flurscheim

THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

All meeting at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, W.C.1, unless otherwise stated.

Joint I.E.R.E.-I.E.E. Medical Electronics Group

Date: 5 October Time: 5.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2
Discussion: Transistorized Biological Amplifiers
Date: 7 October Time: 6 p.m.
Lecture: The Contribution of the Engineer to the Economy
By: A. C. Copisarow
(Tickets will be necessary and those wishing to attend should apply to the I.E.R.E. at 9 Bedford Square, London, W.C.1)

Radar Group

Date: 21 October Time: 6 p.m.
Lecture: The Effect of Tolerances on Automatic Landing
By: M. G. Henley

Joint I.E.R.E.-I.E.E. Computer Groups

Date: 26 October Time: 5.30 p.m.
Colloquium: Design Automation
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2

Joint I.E.R.E.-I.E.E. Electro-Acoustic Groups

Date: 28 October Time: 6 p.m.
Symposium of Short Papers on ' Magnetic Circuit Design Applied to Electromechanical Transducers '
By: A. E. Falkus, S. Kelly, A. K. Hathway and F. Knight

South Western Section

Date: 7 October Time: 7 p.m.
Held at: Bristol College of Commerce, Unity Street, Bristol
Annual General Meeting of the Section followed by a programme of recent technical films

South Wales Section

Date: 14 October Time: 6.30 p.m.
Held at: Welsh College of Advanced Technology, Cardiff.
Lecture: The N.T.S.C. Colour Television System
By: P. Lowry

Scottish Section

Date: 7 October Time: 7 p.m.
Held at: The Department of Natural Philosophy, The University, Edinburgh
Lecture: Electronic Control Systems in Coal Mining
By: G. M. Rendall

Southern Section

Date: 13 October Time: 6.30 p.m.
Held at: Lanchester Theatre, University of Southampton, Southampton
Lecture: Compatible Stereophonic Broadcasting
By: A. R. V. Roberts
Date: 29 October Time: 7 p.m.
Held at: Farnborough Technical College, Farnborough
Lecture: High Frequency Induction and Dielectric Heating
By: W. D. Wilkinson

East Midlands Section

Date: 14 October Time: 6.30 p.m
Held at: University of Leicester, Leicester
Lecture: Transistor Amplifiers
By: P. J. Baxandall

Merseyside Section

Date: 21 October Time: 7.30 p.m.
Held at: Walker Art Gallery, Liverpool
Lecture: Cold Cathode Tubes
By: D. Reaney

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 7 October Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2
Lecture: Character Recognition (Postal Mechanisation)
By: A. W. M. Coombs
Date: 20 October Time: 5 p.m.
Held at: The Conference Room, Fleet Building, 70 Farringdon Street, E.C.4
Lecture: Maintenance in the London Director Network
By: T. J. Rees

THE RADAR & ELECTRONICS ASSOCIATION

Date: 8 October Time: 7 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2
Lecture: Telecommunications from the New Post Office Tower
By: W. L. Newman

THE TELEVISION SOCIETY

Date: 1 October Time: 7 p.m.
Held at: The Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3
Lecture: Microphones and Sound Effect for Television
By: D. J. Basinger
Date: 16 October Time: 7 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2
Lecture: Transcoding from SECAM to N.T.S.C.
By: H. Steele

PUBLICATIONS RECEIVED

HIGH FREQUENCY CRYSTAL FILTERS is the title of a new STC Publication which describes 22 types of filter specially designed for use in mobile radio equipment. The characteristics of each device are given in tabular and graphic form. Filters are listed for 12·5, 20, 25 and 50kc/s channel spacing. Copies of this publication are available from Standard Telephones and Cables Ltd, Components Group, Footscray, Sidcup, Kent.

VARIAC CATALOGUE describes the variable transformers now produced by Claude Lyons Ltd, Valley Works, Hoddesdon, Herts, and contains their new price list. Copies of the catalogue and price list are obtainable direct from the company at the above address.

THE ' ETCHINEERING ' PROCESS of photo-etching metals is described in a booklet recently published by Microponent Development Ltd, 17 Newhall Hill, Birmingham 1. The procedures are described, many examples of different types of flat metal parts produced by it are shown and there is a list of normal tolerances attainable on a wide range of metals. The issue coincides with the firm moving to a new address in much larger premises.

PRECISION ELECTRONIC INSTRUMENTS is the title of the new catalogue recently published by the Belix Company Ltd, Victoria Road, Surbiton, Surrey, describing their range of electronic units now available.

BEAUDOUIN X-RAY CRYSTALLOGRAPHY EQUIPMENT is described in a catalogue being distributed in this country by Scientific Techniques Ltd, 10 Dover Street, London, W.1. Notable in the Beaudouin range are the cameras, the curved crystal monochromators and a variety of units for film preparation, and diagram analysis.

TUBE AND SEMICONDUCTOR INFORMATION is contained in a number of pamphlets recently published by Telefunken of Western Germany. They are as follows: No. 64 05 102—Data and Characteristics of Transistor AC 160; No. 64 04 103—Transistorized a.f. Amplifiers for Outputs of 4W, 5W, 8W and 20W; No. 64 03 107—15W Transistor Amplifier for High Quality Reproduction; No. 64 04 108—A Stereo Adaptor incorporating Automatic Indication; No. 63 12 109—Design of Class B Push-Pull Transistor Amplifier. Further information and copies of these publications may be obtained from the Press Office, Telefunken AG, 1 Berlin-Charlottenburg, Ernst-Reuter-Platz, Germany.

MULLARD EDUCATIONAL SERVICE have recently issued four new publications which are as follows: One is a list of elements which gives their atomic number and weight, electron and shell dispositions, periodic group and neutron complement. Unstable elements are indicated and atomic weights are shown to four significant figures where these are available. If only a few lists are required they are issued free of charge: a nominal charge of 2s. 6d. per 25 is made for larger quantities.

The other publications are three new pamphlets in the series ' Educational Electronic Experiments '.

The first, number 8 in the series, describes the construction of a low-voltage electrometer which is particularly suitable for pH measurements.

Number 9 describes the construction of a decade scaler which has a maximum count rate of 400 cycles per second and will accept and count most standard waveforms. Power supplies are derived from a 12V d.c. source which gives the added advantage of portability.

A method of making Hall effect measurements is described in pamphlet number 10.

The pamphlets are available free of charge to schools and other educational establishments direct from the Public Relations Department, Mullard House, Torrington Place, London, W.C.1.

LE SALON AÉRONAUTIQUE

Une description, basée sur des renseignements fournis par les fabricants, de certains des appareils électroniques exposes au dernier salon de la SBAC.

Traduction des pages 713 à 719

AMPLIVOX LTD

Beresford Avenue, Wembley, Middlesex

ÉMETTEUR-RÉCEPTEUR MINIATURE

(Illustration à la page 713)

L'émetteur-récepteur Televox fonctionnant sur batterie est entièrement portatif et se loge facilement dans la poche. C'est l'un des plus petits combinés portatifs fabriqués en Angleterre et qui trouvera de nombreuses applications pour les communications entre personnes se trouvant à une certaine distance les unes des autres.

Dans la réalisation de ce combiné, on a tenu compte de la diversité des conditions d'emploi dans différentes industries et l'appareil comporte donc plusieurs caractéristiques tout à fait particulières:

(1) Chacun des émetteurs-récepteurs comprend des circuits pour un maximum de quatre voies à pilotage piézoélectrique. Ces circuits assurent les communications entre opérateurs à la même fréquence, ainsi que la possibilité d'accorder différentes voies à d'autres personnes dans la même zone.

(2) Le Televox sera fourni avec un compartiment à batterie amovible. Un compartiment légèrement plus grand sera également prévu en variante pour des périodes d'emploi d'une très longue durée. Ces éléments à batterie se logent très facilement à la base de la section à radio. L'élément à batterie peut être remplacé par un adaptateur ou une fiche reliés à un bloc d'alimentation indépendant.

(3) Le Televox comportera un haut-parleur/microphone et sera donc entièrement autonome.

Par simple commutation il pourra être utilisé avec un microphone à main/haut-parleur extérieur tel que le Mini-Mike Amplivox (voir notre gravure), ou dans des conditions de bruit environnant très élevé, en liaison avec un casque excluant le bruit, comme ceux de la série Amplivox Ampligard.

(4) Le Televox comprend une antenne téléscopique et il peut être relié à une antenne extérieure pour pouvoir obtenir une gamme plus étendue.

Les accessoires comprennent un étui en cuir de haute qualité, des cristaux, une série de microphones, des éléments d'écoute et des casques.

EE 74 751 *pour plus amples renseignements*

BRITISH COMMUNICATIONS CORPORATION LTD

South Way Exhibition Grounds, Wembley, Middlesex

RADIOTÉLÉPHONES U.H.F.

(Illustration à la page 713)

En raison de la demande croissante de nouveaux services nécessitant déjà la répartition des voies dans des bandes très haute fréquence encombrées, la BCC prévoit l'emploi de la bande UHF et elle a réalisé à cet effet un nouveau matériel radiotéléphonique à stations fixe et mobile qu'on a pu voir pour la première fois.

Le BCC 55 est un matériel mobile à six voies de 10 W p.m. L'équipment à station fixe (voir notre gravure), c'est à dire le combiné récepteur—émetteur BCC 415/115, est de 15 W p.m. Les deux appareils sont prévus pour le fonctionnement dans la bande 440 à 470 MHz.

EE 74 752 *pour plus amples renseignements*

RADIOTÉLÉPHONES VHF

(Illustration à la page 713)

Cette gamme de radiotéléphones portatifs de poids léger couvre toutes les bandes très haute fréquence et elle comprend des modèles à modulation d'amplitude et à modulation de phase. Chaque appareil comporte un choix de six voies ainsi qu'un haut parleur pouvant servir également de microphone. Un bloc d'alimentation à batterie sèche peut être fourni également dans un choix très étendu d'accessoires.

La gamme de fréquences s'étend de 41 à 174 MHz avec espacement de voies de 25 ou 50 kHz. La puissance de sortie de l'émetteur varie entre 120 et 500 mW selon la fréquence et le genre de modulation. La stabilité de la fréquence est supérieure à ± 2 kHz.

La puissance BF de sortie maxima du récepteur est de 50 mW et la sélectivité est telle qu'elle assure un réponse supérieure à 3 dB entre $\pm 7{,}5$ kHz et supérieure à -60 dB au-dessus de ± 25 kHz (avec espacement de voies de 25 kHz).

EE 74 753 *pour plus amples renseignements*

EMETTEUR-RÉCEPTEUR PORTATIF H.F.

(Illustration à la page 714)

L'émetteur-récepteur portatif H.F. type BCC 30 HP est un appareil "tout transistors" alliant une puissance élevée (20 W) à un poids total réduit de 17 kg. Ce faible poids le rend idéal à l'emploi par les parachutistes car il leur permet d'effectuer des missions de longue durée qu'ils ne pourraient exécuter avec des appareils très haute fréquence. Il se caractérise en particuleir, en dehors de son poids global réduit pour une station portative complète de 20 W, par les possibilités multiples qu'il offre. Il prévoit, en effet, des ondes entretenues ainsi que la modulation d'amplitude et de phase en phonie. Il comprend 18 voies pilotées par cristal dans la bande de 2 à 8 (2,5 à 10) MHz. On peut également effectuer l'accord libre sur toute la bande avec une sélectivité permettant 600 voies de parole.

EE 74 754 *pour plus amples renseignements*

A. C. COSSOR LTD

The Pinnacles, Elizabeth Way, Harlow, Essex

RÉPONDEURS DE BORD

(Illustration à la page 714)

En plus d'une installation complète de radar de sol de surveillance, on a pu voir deux répondeurs de bord, soit un appareil militaire, type 1500, et un appareil civil, type 1600. Le répondeur miniaturisé 1500 d'identification "ami ou ennemi", mis au point sous contrat du ministère britannique de l'aviation, a été conçu pour l'utilisation continue dans une gamme de températures extrêmement étendue, à savoir de $-55°$ C à $+140°$ C. Son facteur élevé de maintien automatique du relèvement vrai lui assure une fiabilité exceptionnelle pour l'appareillage de bord.

Le type 1600 a été construit suivant les normes de l'ICAO, Annexe 10, et la caractéristique 532D de l'ARINC. Il est conforme aux conditions d'admissibilité du Air Registration Board et de la United States FAA, et il comporte des dispositifs d'encodage à 12 chiffres binaires sur les modes A, B, C et D. C'est un appareil entièrement transistorisé, à l'exception de l'étage final de l'émetteur dont la puissance de sortie dépasse 1kW. Il a été prévu pour le fonctionnement continu dans la gamme de températures allant de $-40°$ C à $+55°$ C, avec refroidissement par convection, ainsi que pour le fonctionnement non pressurisé jusqu'à une altitude de 12160m. Il comprend un dispositif incorporé de contrôle automatique actionné à partir du panneau de commande dans le compartiment de l'équipage.

EE 74 755 *pour plus amples renseignements*

RADIOTÉLÉPHONE MINIATURE

(Illustration à la page 714)

La gamme d'appareils "Companion" consiste en une série de radiotéléphones à usage personnel conçus pour être transportés ou portés par l'opérateur. Ils comprennent un récepteur-émetteur de poche VHF à modulation de fréquence CC2/8, un récepteur-émetteur à modulation d'amplitude CC3 et un nouveau radiotéléphone à modulation de fréquence CC4 à six voies et de poids léger.

Ne pesant que 500 grammes (environ) et mesurant 14,3 cm × 8,89 cm × 3,81 cm, le CC2/8 a une gamme de fréquence de 71,5 MHz à 174 MHz, pour une puissance d'émission de 500 mW. La sensibilité de réception est de 0,6 μV pour un accord silercieux de 20 dB et le niveau de réponse parasite est supérieur à −60 dB. Ayant une puissance BF de sortie de 200 mW, l'appareil peut être utilisé dans des zones de bruit ambiant élevé. En utilisant l'émetteur-récepteur de base avec différents accessoires, on peut réaliser divers montages répondant à des nécessités particulières et la puissance de sortie peut être réglée en fonction des exigences opérationnelles permises par la durée de la batterie.

EE 74 756 *pour plus amples renseignements*

DECCA RADAR LTD

Albert Embankment, London, S.E.1

RADAR DE SURVEILLANCE AÉRIENNE

(*Illustration à la page 714*)

Le radar de surveillance du trafic aérien AR1 est un appareil à définition de 10 cm de haut et à vitesse d'indication élevée. Il est déjà produit en série pour la Royal Air Force ainsi que pour les autorités de l'aviation civile. Pouvant effectuer toutes les opérations de contrôle du trafic aérien dans un rayon de 120 km d'une aérogare, l'AR1 a été spécialement conçu pour satisfaire aux exigences opérationnelles des aéroports qui ne pourraient envisager le coût d'installation très élevé des grands radars de surveillance, secondés par des appareils à faible portée. En s'attachant d'emblée à lui assurer une grande souplesse d'emploi, à réduire les frais d'installation et les frais d'entretien journaliers, la société Decca a pu ainsi réaliser un radar de surveillance compact pouvant répondre à la demande accrue d'un système de radar pratique à l'intention des grands avions de transport civil. Grâce à un dispositif d'élimination des échos fixes, stable et transistorisé, pour vitesses de vol à l'aveugle non inférieures à 560 noeuds, et grâce également à un dispositif de polarisation variable d'une conception hardie, le radar AR1 dépasse sensiblement les conditions requises par l'International Civil Aviation Organization pour les opérations à courte portée dont l'approche au PPI. Il fournit, en même temps, un potentiel de surveillance d'environ 160,9 Km à une altitude de 12192 m, selon les dimensions de l'avion.

EE 74 757 *pour plus amples renseignements*

INDICATEURS

(*Illustration à la page 714*)

On a pu voir une démonstration de ligne d'un système de contrôle de trafic aérien intégré au cours de laquelle des indications d'un radar de navigation Decca, émises au moyen d'une liaison de données air/sol, furent présentées, en même temps que des indications de radar de surveillance, sur les indicateurs autonomes Decca série 5. A Farnborough, l'indicateur fut utilisé à partir du radar de surveillance de la station tandis qu'une liaison de radar de navigation Decca air/sol était logée immédiatement à l'extérieur de la tente principale de l'exposition.

L'indicateur de radar utilisé pour cette nouvelle démonstration de contrôle de trafic aérien est un indicateur du type 12 C de la gamme des indicateurs autonomes à transistors Decca série 5. Cette installation comporte une présentation PPI normale avec décentrage jusqu'à 8 rayons, trois marqueurs de forme caractéristique dont la position est déterminée par les renseignements relayés à partir de récepteurs de bord de radar de navigation Decca, et un symbole marqueur sous contrôle local à la position d'indication.

L'utilisation d'un indicateur standard pour cette présentation relativement compliquée est rendu possible grâce à la souplesse d'emploi des appareils de la série 5. Des conditions opérationnelles allant de l'affichage simple au PPI ou par indicateur de hauteur relative à l'affichage exigeant l'emploi de symboles de marqueurs électroniques, de lignes de mesure de la portée et dugisement, de la présentation radiogoniométrique directe sur le tube cathodique du marquage entre supports de radar et d'autres accessoires analogues pour le traitement des données peuvent toutes être satisfaites grâce à cette installation, comprenant une centaine de variations. Un autre exemple récent de l'emploi de ce matériel est fourni par le projet de simulateur Eurocontrol, réalisé conjointement par les sociétés Decca, Telefunken de Berlin et CSF de Paris. L'installation de simulateur à Bretigny utilisera une version à support spécial de ces indicateurs comme système d'affichage primaire pour les "pilotes" aussi bien que pour les contrôleurs.

EE 74 758 *pour plus amples renseignements*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

EQUIPEMENT RADIOGONIOMÉTRIQUE AUTOMATIQUE VHF

(*Illustration à la page 715*)

Cet équipement de haute précision et à bas prix se compose de trois éléments, à savoir un indicateur, un élément de bâti et une antenne.

Le mode de fonctionnement est comme suit:

La rotation continue de l'antenne Adcock s'effectue à la vitesse de 250 tours/minute et son débit est combiné avec une petite partie d'un omnisignal d'un dipôle monté au centre de deux éléments Adcock.

Vu que les diagrammes polaires de l'antenne ont à peu près la forme d'un 8 dont les zéros ne sont pas diamétralement opposés de manière exacte, l'espacement entre zéros successifs au cours de la réception équivaut tantôt à un peu plus, tantôt à un peu moins de 180° de rotation de l'antenne.

Ces zéros apparaissent donc à la sortie du deuxième détecteur de réception mais ils sont encore, à ce stade, inappropriés à la perception sensible ou à l'affichage, en raison des variations d'acuité dues à la force des signaux reçus et à d'autres facteurs. Ils sont donc appliqués à un circuit générateur d'impulsions dans lequel sont formées des impulsions étroites correspondantes d'un largeur constante. Ce circuit se compose essentiellement d'une base de temps monocourse, déclenchée à une tension de signal prédéterminée avant le zéro et fonctionnant à mi-vitesse jusqu'à ce que le signal atteigne le même niveau prédéterminé après le zéro. A ce stade, la décharge continue à pleine vitesse jusqu'à ce que la limite inférieure de la base de temps soit atteinte et que la retour du spot soit commencé.

Une impulsion étroite se forme à partir du temps de retour, avec un retard constant après le point médian dans le temps entre les instants auxquels le signal atteint la même tension avant et après le zéro. Le minutage de l'impulsion ainsi produite n'est donc pas affecté par la netteté du zéro dans la gamme de travail de l'équipement, à condition que la tension de signal autour du zéro soit symétrique. L'effet d'une légère asymétrie nécessaire à la détection est minimisé en limitant le fonctionnement du circuit à la zone de zéro immédiate. Les grands zéros, qui ne sont pas inclus dans la gamme de travail de l'équipement, sont éliminés par une porte de largeur.

Les impulsions étroites produites par le générateur d'impulsions sont appliquées à un émetteur de résolution accouplé à l'arbre d'antenne et divisées en composantes sinus et cosinus par rapport à la position de l'antenne. Ces composantes sont appliquées aux bobines d'analyse d'un tube cathodique de sorte que les déviations s'effectuent dans une direction sensiblement radiale qui dépend de la position angulaire de l'antenne au moment où se produit l'impulsion.

Les bobines de balayage sont orientées de manière appropriée par rapport à l'échelle de relèvement autour de l'écran à tube cathodique, de sorte que la position de l'antenne à ce moment-là et, par conséquent, la direction de la source du signal, est clairement indiquée.

Pour obtenir des relèvements détectés, les impulsions étroites produites sont appliquées séparément à un circuit contrôlant la brillance du tube cathodique dont le rôle est alors d'éliminer les impulsions précédées par des intervalles inférieurs à 180° de rotation de l'antenne.

EE 74 759 *pour plus amples renseignements*

FERRANTI LTD

Hollinwood, Lancashire

MOTEURS À ROTOR EXTÉRIEUR

(*Illustration à la page 715*)

In a pu voir une nouvelle gamme de moteurs à rotor extérieur généralement conformes à la Spécification EL. 1892 du ministère de l'aviation. Ces moteurs, initialement conçus comme moteurs de rotation de gyroscopes, ont trouvé des applications dans de nombreux autres domaines. Ils peuvent ainsi être utilisés comme petits ventilateurs, sources de puissance et de ventilation combinées, entraîneurs de tambours, poulies à moteur, plateaux isolants actionnés par moteur, etc. Ils sont d'une grande efficacité et permettent d'obtenir un rapport élevé débit/volume, ce qui est un facteur d'importance dans des considérations d'espace, de poids, de température ou de puissance limitée.

EE 74 760 *pour plus amples renseignements*

INDICATEUR À CARTE MOBILE

L'indicateur à carte mobile offre tant au pilote qu'au navigateur une image optique de la carte survolée par l'avion. La position momentanée et la piste suivie par l'avion sont indiquées sur l'ecran en verre dépoli de l'appareil sous forme de cercle central et de ligne radiale. L'image de la carte, projetée optiquement sur l'écran à partir de la bande de film en couleurs des cartes topographiques appropriées, se déplace derrière ces indications. La réduction utilisée pour le microtournage du film des cartes est de 20 diamètres, le système optique de l'indicateur étant conçu de telle manière que le film est agrandi 20 fois durant la projection.

La bande de film est entraînée le long de sa longueur pour les changements de cap vers l'est et le long de sa largeur pour les changements vers le nord. Ces changements s'effectuent à l'aide de moteurs asservis sur chaque axe. Ces moteurs reçoivent leurs entrées du système de navigation. Doppler ou à inertie par le truchement de la calculatrice de navigation et d'une calculatrice analogique électromécanique. Les indicateurs fournissent au pilote et au navigateur une image facilement interprétée des conditions de vol au cours d'une sortie. Un marqueur de la position momentanée montre constamment le point précis où se trouve l'avion et la ligne de piste radiale montre le point qui sera survolé par l'avion. Par beau temps, le pilote peut utiliser son indicateur pour lire sur la carte le cours suivi en reliant l'image sur son indicateur à la configuration du terrain survolé. Il peut faire le point visuel et vérifier le fonctionnement correct du système de navigation.

Le navigateur utilise sa carte mobile pour obtenir des renseignements de navigation de même que comme contrôleur du système de navigation. Il peut également comparer les indications de son radar à vision latérale avec son indicateur afin de faire le point au radar. De cette manière, l'indicateur lui permet de prévoir l'apparition de points fixes de radar et de les identifier au fur et à mesure de leur apparition.

EE 74 761 *pour plus amples renseignements*

THE MARCONI CO. LTD

Chelmsford, Essex

ÉQUIPEMENT DE MESURE DE LA DISTANCE

(*Illustration à la page 716*)

Le relais combiné de l'equipement de bord de mesure de la distance AD70 mesure automatiquement et de manière continue la gamme d'inclinaison d'un avion par rapport à une balise de sol déterminée. Cette distance est affichée directement sur un enregistreur se trouvant sur le panneau d'instruments du pilote. Les répondeurs de sol utilisés avec ce type d'equipement peuvent être des installations VORTAC ou TACAN, toutes deux remplissant également d'autres fonctions de navigation en plus de la mesure de distance.

L'indication de la distance, lorsqu'elle est combinée avec un seul relèvement de radio-compas ou de balise omnidirectionnelle très haute fréquence, permet de déterminer avec précision la position de l'avion. Dans la plupart des installations de sol, une balise du type omnidirectionnel très haute fréquence est utilisée en même temps que le répondeur de l'équipement de mesure de distance, comme c'est le cas pour l'équipement VORTAC.

Une balise d'équipement de mesure de distance placée à l'extrémité d'une piste d'atterrissage peut également constituer un appareil auxiliaire précieux pour les manoeuvres d'atterrissage, car elle fournit une indication continue de la distance à parcourir jusqu'au point d'atterrissage.

L'équipement de mesure de la distance AD70 comprend 17 assemblages modulaires séparés, montés sur un bâti principal en aluminium selon les normes du Air Traffic Registration, pesant au total une quinzaine de kilos. Ces appareils sont transistorisés, sauf dans de cas où des niveaux de puissance élevés ou d'autres considérations de réalisation empêchent leur emploi. L'appareil est muni de plusieurs modules à circuit imprimé à fiches, ainsi que de plaquettes supplémentaires permettant de simplifier le contrôle opérationnel courant de l'équipement. Il comporte 126 voies de fréquence pilotées par quartz, mais ce nombre peut être porté à 252 au cas où le système de mesure de distance serait agrandi par la suite. Les fréquences sont fixées par commutateur de décades sur un contrôleur à part et, une fois la fréquence fixée, le fonctionnement de l'appareil devient entièrement automatique.

Une succession ininterrompue d'impulsions d'interrogation en paires est émise par l'émetteur et chacune de ces impulsions provoque l'émission d'impulsions de réponse du répondeur de sol après un intervalle de temps déterminé. On peut ensuite mesurer la distance séparant l'avion de la balise de sol par l'intervalle de temps entre chaque paire d'impulsions d'interrogation et la réponse reçue. L'appareil détermine cet intervalle de temps au moyen d'un circuit de porte récepteur, sensible à la durée, qui n'acceptera les signaux reçus du répondeur de sol qu'après un intervalle de temps distinct suivant l'émission d'une paire d'impulsions d'interrogation. L'espace de temps entre cet intervalle et l'émission varie automatiquement de zéro à la durée correspondant à 322 km. Lorsqu'une réponse parvient à ce circuit de porte, l'appareil "verrouille" ce signal puis change la position du circuit de porte à la vitesse nécessaire à la mise en place des signaux reçus à mesure que l'avion s'approche ou s'éloigne de la balise de sol. La distance correspondant aux intervalles de temps mesurés instantanément est affichée simultanément sur l'enregistreur du panneau d'instruments du pilote.

EE 74 762 *pour plus amples renseignements*

EQUIPEMENT HF DE BORD CONSTITUÉ DE CORPS SOLIDES

(*Illustration à la page 716*)

Le système de communications HF, type AD460, présente les mêmes qualités de fiabilité que tous les appareils Marconi de la série 60. Les techniques de réalisation les plus modernes ont été employées, et à toutes les étapes de la mise au point et de la construction de chaque appareil, des essais rigoureux ont été appliqués pour leur assurer une fiabilité absolue.

Le AD460 est accordable sur toute la bande de fréquence de 2 à 29,999 MHz par commutateurs de décades sur un contrôleur à part. L'espacement des voies de $\frac{1}{4}$ kHz permettra de répondre à tous les besoins futurs éventuels de voies supplémentaires dans la bande encombrée des hautes fréquences.

Des diodes semiconductrices sont utilisées pour toutes les opérations de commutation et l'accord se fait exclusivement par éléments constitués de corps solides, sauf à l'étage final de sortie de puissance. Les fréquences de fonctionnement sont produites par la méthode de synthétisation de la fréquence utilisant des oscillateurs à décades, individuellement verrouillés en phase à l'oscillateur de référence à fréquence stable. L'emploi exclusif de transistors au silicium permet le fonctionnement sûr de l'appareil dans une gamme de températures allant de $-55°$ C à $+55°$ C.

Le fonctionnement s'effectue soit par deux bandes latérales pour radiotéléphonie et ondes entretenues, soit par bande latérale unique pour radiotéléphonie sans ondes entretenues, soit enfin par bande latérale unique (porteuse flottante et commande automatique de fréquence) pour radiotéléphonie ou modulation par déplacement de fréquence pour liaison de renseignements. On a également prévu le fonctionnement par système SELCAL (appel sélectif), conformément à la Caractéristique 531 de l'ARINC, et des connexions d'entrée et de sortie séparées sont prévues pour l'emploi avec le matériel de liaison de renseignements. Des étages séparés de renseignements moyenne fréquence et des discriminateurs sont également inclus pour les liaisons de renseignements. On peut, en outre, effectuer l'émission par ondes entretenues à large bande ou à bande étroite et un oscillateur à battements séparé est prévu pour ce mode de fonctionnement.

EE 74 763 *pour plus amples renseignements*

MARCONI INSTRUMENTS LTD

St. Albans, Hertfordshire

GÉNÉRATEUR DE SIGNAUX À MODULATION D'AMPLITUDE

(*Illustration à la page 716*)

Le générateur de signaux à modulation d'amplitude, type TF 801D/SM1, est à peu près semblable au modèle normal type TF 801D/1 (10 à 470 MHz), mais il comporte des dispositifs spéciaux permettant de l'adapter au contrôle de certains types d'aides à la navigation aérienne. A cet effet, on peut modifier la gamme de fréquences afin d'assurer une meilleure discrimination dans les bandes appropriées. La puissance de sortie s'étend de 0,1 mV à 1 V, f.é.m., à variation continue, et la dérive de fréquence est inférieure à $\pm 5 \times 10 - 5/10$ minutes.

Afin de pouvoir utiliser l'appareil comme radiophare VHF omnidirectionnel, les circuits extérieurs de modulation produisent un déphasage négligeable à 30 Hz, et la forme d'onde démodulée est fournie sur le panneau frontal à partir d'un détecteur intérieur pour la vérification, tandis que la bande des ondes porteuses très haute fréquence est centrée environ à la quatrième gamme du générateur (70 MHz à 150 MHz).

Un oscillateur piloté au quartz est également prévu à l'intérieur. Il est utilisé lorsqu'un signal précis de 75 MHz est exigé; par exemple, pour les récepteurs de balises marqueuses. On l'emploi normalement avec la troisième gamme du générateur (48 MHz à 110 MHz), l'oscillateur variable étant débranché, mais l'amplificateur demeurant en fonctionnement. On obtient alors un signal d'un niveau élevé susceptible d'une modulation profonde.

EE 74 764 *pour plus amples renseignements*

FILTRE DE REJET ACCORDABLE

(*Illustration à la page 717*)

En plus de son emploi comme accessoire pour l'analyseur d'ondes Marconi type TF 2330, afin de permettre la mesure des amplitudes relatives variant de plus de 75 dB, cet instrument peut être employé pour d'autres applications exigeant la suppression ou l'élimination d'une seule fréquence.

Le filtre TF 2334 assure le rejet sur 80 dB des fondamentales dans la gamme de 20 Hz à 20 kHz, avec une atténuation minima des harmoniques. C'est essentiellement un filtre à "T jumelés" d'une impédance élevée, conçu pour l'emploi avec des sources inférieures à 1 000 Ω. Des modes d'entrée de 600 Ω ou à impédance élevée sont prévus. Lorsque les modes de 600 Ω sont utilisés, on peut compenser par commutation l'atténuation des harmoniques jusqu'au quatrième, de sorte que leurs valeurs peuvant être lues directement sur l'analyseur d'ondes.

EE 74 765 *pour plus amples renseignements*

PLESSEY-UK LTD

Ilford, Essex

ENREGISTREUR DE RENSEIGNEMENTS DE VOL

Cet enregistreur qui sera installé sur tous les appareils de la flotte aérienne de la British European Airways, ainsi que sur tous les avions à réaction "BAC one-eleven" actuellement en cours de commande pour Aer Lingus, répond pleinement aux conditions exigées par le Ministère britannique de l'aviation pour pouvoir assurer un enregistrement susceptible de résister aux chocs, au feu et aux attaques chimiques.

En plus des six paramètres exigés par le ministère de l'aviation civile, à savoir: la vitesse propre nominale, l'altitude nominale, le cap magnétique, l'accélération verticale, l'angle d'inclinaison, le temps, le nouvel enregistreur a été également étudié pour les travaux de recherche spéciaux, car il est muni de canaux de réserve pour des circuits d'échantillonnage supplémentaires. L'enregistrement numérique de tous les signaux garantit le maintient du plus haut degré de précision dans toute la séquence des opérations, c'est à dire depuis l'enregistrement en cours de vol jusqu'au traitement des données et à l'affichage au sol. On peut aussi lui ajouter des dispositifs de contrôle automatique incorporés.

Le système comprend trois éléments principaux: un élément de contrôle placé dans le poste de pilotage, un élément de traitement des données et un enregistreur, auxquels sont reliés un certain nombre de transducteurs. L'enregistreur—spécialement conçu par la société S. Davall and Sons Ltd—utilise l'enregistrement magnétique sur fil d'acier inoxydable, ce qui garantit une résistance aux risques d'écrasement supérieure à celle que pourrait assurer n'importe quel autre moyen.

L'enregistreur, qui sera installé dans la queue de l'avion, est de construction si robuste que l'éjection automatique a été considérée comme superflue. Ainsi, il a été construit pour pouvoir résister à des charges maxima d'une tonne, à une accélération pouvant atteindre 100 g à des températures maxima de 800°C pendant 15 minutes et à une immersion totale dans l'eau de mer pendant une trentaine d'heures au maximum.

EE 74 766 *pour plus amples renseignements*

OUTILS DE SERTISSAGE

(*Illustration à la page 717*)

Des outils de sertissage servo-commandés viennent d'être réalisés. Il s'agit de versions à l'air comprimé du "UNIcrimp" manuel: un outil universel automatique et actionné par une seule main, capable d'effectuer un sertissage carré parfait sur n'importe quel format de fil ou de contact jusqu'à une épaisseur de jauge de 12 AWG.

Le modèle mécanique se présente sous deux formes différentes: un modèle portatif semblable au UNIcrimp manuel et un nouveau modèle de bande DC avec une base robuste pour les travaux d'atelier. Le modèle mécanique peut sertir des cosses pré-isolées ainsi que tous les contacts de connecteur. Dans n'importe laquelle de ces trois formes, le UNIcrimp constitue un outil de production éprouvé et à l'abri des fausses manoeuvres pouvant être utilisé par des ouvriers non spécialisés.

EE 74 767 *pour plus amples renseignements*

RANK BUSH MURPHY ELECTRONICS

Welwyo Garden City, Hertfordshire

RADIOPHARE

(*Illustration à la page 717*)

On a pu voir pour la première fois au Salon de l'Aéronautique de Farnborough, en Septembre, un radiophare prévu comme auxiliaire d'appui aérien des troupes de première ligne. Le nouveau radiophare constitue le "dernier-né" de la série des systèmes de navigation Eureka/Rebecca de renom mondial, réalisé par la société Rank Bush Murphy Electronics.

Il s'agit d'un radiophare transistorisé à faible puissance, portant le nom d'Eureka MR 343, conçu pour le parachutage avec des troupes d'assaut aéroportées et pouvant fournir un signal de distance et de radioguidage pour les avions d'appui et constituant, en outre, un auxiliaire terminal pour les aérodromes avancés.

L'encombrement, le poids et la puissance de l'appareil ont été considérablement réduits il peut être utilisé par des opérateurs militaires non spécialisés. Il comporte un dispositif incorporé pour contrôler le comportement du récepteur et de l'émetteur.

EE 74 768 *pour plus amples renseignements*

REDIFON LTD

Broomhill Road, London, S.W.18

RADIOPHARE TRANSPORTABLE

Le radiophare m.f. TG. 142 R a été spécialement conçu pour le montage rapide en campagne. La station complète comprend huit éléments transportables et peut être montée par deux hommes à l'aide de la flèche de chevalement de 6 m fournie avec l'installation pour l'érection d'une antenne à mât vertical de 18,3 m.

L'installation comprend un émetteur, un mat d'antenne en six sections de 3 m avec un sommet de capacité, douze câbles radiaux de mise à la masse de 33,5 m, un élément de couplage à l'antenne, un générateur à essence, deux coffrets à accessoires, une tente et une flèche de chevalement.

Le radiophare fonctionne dans la gamme de fréquence de 200 à 420 kHz sur ondes entretenues ou ondes entretenues modulées, avec télécommande ou commande locale. Un disque de manipulation spécial est utilisé pour le choix des caractéristiques de signal qui peuvant être facilement modifiées selon les circonstances. Cette modification s'applique également à l'indicatif.

EE 74 769 *pour plus amples renseignements*

S. SMITH & SONS (ENGLAND) LTD

Kelvin House, Wembley Park Drive, Wembley, Middlesex

Systèmes de contrôle de vol

Le systèm série 6 est le plus récent des appareils de contrôle de vol Smith et il a déjà été commandé pour les avions Fokker Turbofan Transport, du type F 28 Fellowship.

Les éléments de la série 6 comprennent le pilote automatique SEP. 6, le système de vol SFS. 6, le système de compas SCS. 6 et le système de contrôle automatique du régulateur STS. 6. Conçu suivant le principe dit des "éléments de construction", le système peut être étendu du simple stabilisateur automatique ou référence de cap au contrôle automatique complet des instruments, nécessaire au fonctionnement dans des conditions climatériques minima.

Le matériel de la série 6 est conforme aux recommandations de l'ARINC et répond aux conditions de contrôle de vol d'avions allant du plus petit modèle aux grands avions de transport à réaction.

La sécurité et la fiabilité ont été les considérations dominantes dans la mise au point du système de contrôle de vol série 5 qui répond aux conditions civiles de fonctionnement par tous les temps et d'atterrisage automatique.

Le système permet non seulement l'atterrisage automatique, mais il assure également une gamme étendue d'opérations de commande de pilotage automatiques et de direction de vol comprenant, entre autres, la mise automatique sur cap prédéterminé, l'acquisition automatique d'une hauteur prédéterminée, le blocage de hauteur, le blocage de la vitesse de descente, l'accouplement du système d'atterrisage aux instruments et du radiophare VHF omnidirectionnel à l'élément sensible automatique de survol d'une station, l'acquisition automatique ou manuelle d'une trajectoire de plané par atterrisage aux instruments et la descente le long de la trajectoire de plané.

EE 74 770 *pour plus amples renseignements*

THE SOLARTRON ELECTRONIC GROUP LTD

Farnborough, Hampshire

Appareil de concentration de données numériques

(Illustration à la page 718)

Le système de concentration de données numériques Solartron est un petit appareil à 20 voies dont la vitesse d'analyse maxima est de 3 points par seconde. Il comprend le voltmètre numérique intégrateur LM 1420 dont la précision est de 0,05% à une sensibilité de 10 μV par chiffre. Le voltmètre utilise un convertisseur tension/fréquence qui assure un circuit d'entrée entièrement isolé. On obtient ainsi un mode de rejet commun de 150 dB et, lorsque le filtre d'entrée est utilisé, on obtient un mode de rejet en série de 60 dB.

Les modules à fiches comprennent un élément de commande de gamme pour l'antenne de 20 voies à voltmètre numérique, une minuterie numérique, un encodeur à poinçonneuse, un entraînement d'imprimeur, un entraînement de machine à écrire, un détecteur de limite, une référence de linéariseur et de thermocouple. Ces éléments peuvent être ajoutés suivant les besoins et ils sont remplacés par des panneaux blancs en cas de non nécessité. Chaque élément à une place particulière dans le coffret.

Notre illustration montre le modèle à trois compartiments avec tous ses accessoires.

EE 74 771 *pour plus amples renseignements*

Carte vidéo à haute résolution

La carte vidéo à haute résolution Solartron peut être intégrée à n'importe quel système de radar classique. Elle constitue la carte la plus précise actuellement disponible. Un tube cathodique à micro-spot dont la dimension du point lumineux est de 3,8 cm × $25{,}4^{-3}$ cm et une substance luminescente "A" sont utilisés pour le balayage. Le tube est serré au cou par un collier special qui assure une référence correcte par rapport au plan de la face du tube et simplifie la concentration.

La distorsion délibérée dans les amplificateurs de déviation compense la nonlinéarité due au tube à face plate, simplifiant ainsi la production de la carte. Un niveau très élevé de dessin et de photographie garantit la précision voulue.

Les spécifications générales sont comme suit:—

Gamme maxima: 40 à 400 milles marins, selon les nécessités. Résolution: 1 partie dans 1 000 (donnant une expansion de plus de vingt fois, parfaitement utilisable). Précision: sur un carte de 100 milles marins de rayon une point à une distance de 10 milles est indiqué avec une précision de position supérieure à 90 m.

EE 74 772 *pour plus amples renseignements*

Oscilloscope à double faisceau

(Illustration à la page 718)

Le nouvel oscilloscope à double faisceau Solartron type CD 1400 comporte un tube cathodique de 12,7 cm à double canon donnant une image large et brillante. Tant le système X que le système Y peuvant être enfichés dans l'instrument et actionnent les plaques de tube directement de sorte que des amplificateurs d'entraînement de plaque incorporés ne sont nullement nécessaires dans le châssis principal. Quatre éléments à fiches sont disponibles actuellement et d'autres sont en cours de mise au point. L'amplificateur à large bande CX 1441 a une largeur de bande de −3 dB du courant continu à 15 MHz et sa gamme de sensibilité s'étend de 100 mV/cm à 50 v/cm. Un gain supplémentaire de ×10 peut être fourni sur toutes les gammes sur une largeur de bande réduite.

L'amplificateur différentiel à gain élevé CX 1442 a une largeur de bande de −3 dB du courant continu à 75 kHz avec une gamme de sensibilité de 1 mV/cm à 5 V/cm. Un gain supplémentaire de ×10 est prévu, accouplé au courant alternatif seulement, et sur bande réduite. Le rejet de mode commun est d'environ 60 dB.

La base de temps CX 1443 offre une gamme de vitesses de 0,5 μsec/cm à 200 msec/cm avec extension jusqu'à ×5 environ, prévu sur commande. Le déclenchement général et la synchronisation haute fréquence s'effectuent à partir de Y 1, Y 2 ou de signaux extérieurs, avec choix de la polarité.

La base de temps à balayage différé CX 1444 est semblable à l'élément CX 1443 mais elle comprend une base de temps à balayage différé avec temps de retard maximum de 100 msec. La précision du temps et de l'amplitude est de ±5% sur la plupart des gammes.

Notre gravure montre le CD 1400 muni de deux amplificateurs à large bande et d'une base de temps.

EE 74 773 *pour plus amples renseignements*

Simulateur de contrôle de trafic aérien

(Illustration à la page 718)

On a pu voir un simulateur de contrôle de trafic aérien à quatre cibles utilisé en liaison avec un indicateur de radar de surveillance air/espace. Ce simulateur fait partie des appareils de la série SY 2020 en cours de production pour des commandes courantes. La rationalisation de la construction modulaire et les avantages de la production par lots ont permis de réduire considérablement le coût de ces appareils, cependant qu'un certain nombre de perfectionnements techniques ont augmenté leur précision et leur sécurité de fonctionnement. Ceci s'applique en particulier à la méthode employée pour l'intégration des positions de cibles. Les nouveaux intégrateurs électroniques de haute précision permettent, pour la première fois, d'obtenir une réponse de cible dans des radars de surveillance à une distance maxima de 75 milles marins, puis de les transmettre à un radar d'approche de précision avec un degré de fidelité suffisant pour produire une image aussi exacte que possible dans toute la gamme.

EE 74 774 *pour plus amples renseignements*

STANDARD TELEPHONES & CABLES LTD

Connaught House, 63 Aldwych, London, W.C.2

RADAR DE SURVEILLANCE SECONDAIRE

Le SGR. 1 est l'élément interrogateur de sol d'un radar de surveillance secondaire. Il peut être utilisé soit indépendamment, soit en liaison avec un radar primaire pour fournir une image corrélative pour le contrôle du trafic aérien.

Le système prévoit deux configurations de base, selon que l'antenne est montée coaxialement sur une antenne de radar primaire ou comme système à part avec son propre mécanisme rotatif. Dans ce dernier cas, il est fourni avec un moteur rotatif, un groupe électrogène de moteurs et un coffret de servo-commande.

L'interrogateur est logé dans un seul coffret contenant un émetteur et un récepteur, ainsi que le bloc d'alimentation et les éléments de contrôle.

L'émetteur emploi un tube à propagation d'ondes par impulsions à réseau unique d'un gain élevé servant à amplifier toutes les impulsions de mode qui sont ensuite injectées à un commutateur semi-conducteur coaxial de grande puissance. Ce commutateur, à temps de fonctionnement très reduit, transmet les impulsions individuelles, au moyen d'atténuateurs haute fréquence préréglés, aux éléments d'interrogation ou de contrôle de l'antenne, selon les besoins.

Ce moyen d'obtenir les impulsions de contrôle ou d'interrogation de l'un des émetteurs à travers des circuits passifs, assure aux impulsions une stabilité d'amplitude à long terme et obvie à la nécessité d'équilibrer des émetteurs séparés. Afin précisément de conserver cette stabilité à long terme et cette fiabilité, l'appareil est entièrement muni de composants constitués de corps solides à l'exception du tube à ondes progressives.

Le dispositif analyseur se compose de deux antennes montées sous forme d'un seul assemblage.

EE 74 775 *pour plus amples renseignements*

ALTIMÈTRE RADIOÉLECTRIQUE MICROMINIATURE

(*Illustration à la page 719*)

Afin d'accroître davantage la sécurité du fonctionnement de l'altimètre radioélectrique modulé en fréquence à bas niveau pour les systèmes d'atterrisage automatiques, la société Standard Telephones and Cables Ltd a réalisé un modèle de laboratoire à l'aide des méthodes de microminiaturisation. Bien que le but original de cette réalisation fut d'améliorer la fiabilité, l'économie de dimensions, de poids et de consommation électrique qui a pu être obtenue par cette technique est fort importante.

La qualité de la gamme actuelle d'altimètres radioélectriques STC est chose connue. Ils ont été utilisés avec succès pour fournir des renseignements d'altitude pour les systèmes d'atterrisage par tous temps depuis la création de l'élément d'atterrisage aveugle du Ministère britannique de l'Information jusqu'au plus récentes installations dans les appareils Trident, VC10, Belfast, ainsi que dans de nombreux autres avions civils et militairse. Durant cette période, la société Standard Telephones and Cables Ltd s'est toujours efforcée d'améliorer les techniques de mise au point, les méthodes de construction et les conditions dans lesquelles ses appareils sont produits. Ces travaux ont abouti à la réalisation de l'altimètre radioélectrique miniature qu'on a pu voir à l'exposition aéronautique.

L'altimètre se compose d'un émetteur constitué de corps solides à modulation de fréquence et d'une puissance de 4 300 MHz, d'un vidéo-amplificateur et de ses circuits, ainsi que de circuits de comptage de fréquence. Cet appareil constitue la matière idéale pour l'application de l'électronique microminiaturisée.

Le modèle exposé était logé dans un coffret "pygmée" format 1/4 ATR. Ce n'est pas nécessairement la forme la plus compacte qu'on puisse lui donner mais elle montre ce qui peut être réalisé par l'emploi des méthodes modernes.

Notre illustration montre l'altimètre microminiature aux côtés d'un altimètre de construction classique.

EE 74 776 *pour plus amples renseignements*

VENNER ELECTRONICS LTD

Kingston By-Pass, New Malden, Surrey

OSCILLATEUR À LARGE BANDE

(*Illustration à la page 719*)

L'oscillateur à large bande TSA 625 couvre une puissance de 10 Hz à 1 MHz en cinq gammes à commutation. Il offre la sélection de gamme par boutons-poussoirs et le contrôle de batterie, des réglages d'atténuateur allant de 2,5 mV sur la totalité de l'échelle à 2,5 V et une impédance de sortie supérieure à 600 Ω sur toutes les gammes. Il fournit des ondes sinusoïdales et des ondes carrées.

EE 74 777 *pour plus amples renseignements*

WESTLAND AIRCRAFT LTD

Saunders-Roe Division, Strain Gauge Department, East Cowes, Isle of Wight

EXTENSOMÈTRE

(*Illustration à la page 719*)

Le service d'extensométrie de la Division Saunders-Roe de la société Westland Aircraft Limited vient de mettre au point un extensomètre de forme simple pouvant être monté à la surface de grandes structures. Il se compose d'une bande de métal ou d'un autre matériau approprié à extrémités sphériques, montée sur ressorts entre des pièces d'extrémité fixées à la structure. Toute déviation sur la longueur de l'extensomètre entraîne un changement dans la flexion de la bande. Ce changement est mesuré par des jauges de contrainte fixées à son point médian. Les données de l'extensomètre sont étalonnées directement en fonction du changement de longueur effectif de la bande. La longueur nominale de l'extensomètre est de 30 cm mais on peut l'adapter à des applications particulières.

Les pièces d'extrémité qui tiennent la bande peuvent être fixées par matières adhésives, par des vis dans des ouvertures taraudées, ou par des tampons dans le cas de structures en briques ou en béton. Les indications peuvent être données sur n'importe quel équipement extensométrique ou galvanomètre. Cependant, pour le service de campagne, un extensomètre portatif, fonctionnant sur batterie, du type fourni par Saunders-Roe, est des plus indiqués.

EE 74 778 *pour plus amples renseignements*

Résumés des Principaux Articles

La réalisation des circuits logiques à résistances à transistors par la méthode graphique par C. W. M. Barrow

Résumé de l'article aux pages 660 à 665

Cet article traite d'une méthode graphique de réalisation de circuits logiques à résistances à transistors NI. Le circuit NI exige un transistor non-conducteur lorsque tous les potentiels d'entrée des cables d'entrée sont à la masse. En contre-partie, le transistor doit être saturé si le potentiel de l'un des conducteurs d'entrée tombe à un potentiel plus négatif qu'une certaine valeur de seuil. Les calculs effectués englobent les effets de tolérances dans les résistances, les blocs d'alimentation et les paramètres de transistors. Les résultats ont été utilisés pour l'étude d'un compteur à plusieurs étages utilisant des éléments de circuit NI.

Évaluation électronique de la croissance des pâturages par F. J. Hyde et J. T. Lawrence

Résumé de l'article aux pages 666 à 670

Les auteurs décrivent la conception et la constructi●n d'un instrument pour la mesure non-destructive de la croissance de l'herbe. La capacité de la sonde employée à cet effet change lorsqu'elle est placée sur l'herbe et provoque à son tour un changement dans la fréquence de l'oscillateur à transistors auquel elle est reliée. La différence entre cette fréquence et celle d'un oscillateur à fréquence fixe identique à 3MHz est convertie en une déviation sur un microampèremètre à courant continu, ceete déviation étant approximativement proportionnelle au changement de capacité. On a constaté que la dèviation du microampèremètre est en rapport avec le contenu d'eau total de l'herbe sous la sonde. On a mesuré en laboratoire la croissance de l'herbe afin de mieux s'expliquer le processus du changement de capacité.

Un amplificateur à courant continu d'une grande fiabilité pour vacuomètres à thermocouples par U. Tallgren et U. Kracht

Résumé de l'article aux pages 671 à 675

Un circuit simple a été réalisé qui permet d'amplifier le signal d'un thermocouple dans un vacuomètre. La sortie est suffisante pour entraîner plusieurs instruments de 1mA reliés en série. L'amplificateur est basé sur un petit transducteur utilisé comme deuxième convertisseur d'harmoniques suivi d'un amplificateur accouplé au courant alternatif. La sortie alimente un redresseur à travers un étage de sortie accouplé à un transformateur. L'isolement galvanique complet est obtenu entre les circuits d'entrée et de sortie. La linéarité de l'amplificateur dépend principalement de la symétrie du transducteur et elle peut être supérieure à ±30%.

Les circuits stabilisateurs du courant de chauffage de deux têtes de vacuomètre à thermocouple d'usage courant sont indiqués, ainsi que les circuits pour commuter un relais à un niveau de sortie prédéterminé de l'amplificateur à thermocouples.

Deux types de circuits sont décrits, dont l'un est destiné à l'utilisation sur alimentation secteur et l'autre sur batterie de secours de 48V.

Les circuits dont il est question dans cet article pourraient également être d'intérêt pour d'autres problèmes d'instrumentation tels que le contrôle de la température à l'aide de thermocouples.

Projet de base d'un récepteur de télévision en couleurs SECAM par P. Cassagne et G. Melchior

Résumé de l'article aux pages 676 à 681

Un prototype commercial de récepteur de télévision en couleurs SECAM vient d'être mis au point. Cet article traite de certaines de ses caractéristiques les plus importantes. Un soin tout particulier a été apporté aux commandes de réception et à la simplicité de fonctionnement du récepteur. Ce prototype prouve qu'on peut produire un récepteur SECAM donnant pleine satisfaction et se comparant favorablement, au point de vue prix, à un récepteur NTSC d'un type équivalent.

Mesure de la permittivité et de la conductivité à des températures maxima de 500°C par R. H. A. Miles

Résumé de l'article aux pages 682 à 687

Un appareil a été mis au point qui effectue la mesure semi-quantitative des variations de permittivité et de conductivité de matériaux dans la gamme de températures de 20° à 500°C. Pour certains matériaux, tels que la céramique ferroélectrique, il se produit de grandes variations dans ces paramètres et il est nécessaire d'employer un système automatique de changement de gamme pour pouvoir enregistrer les résultats sur un enregistreur classique à la précision approximative voulue de 10%. La gamme totale de capacitance couverte par l'appareil s'étend de 1,0pF à 1000pF (à des fréquences de 100 kHz et de 1 MHz), tandis que la gamme de résistance de courant continu enregistrée va de 10Ω à 500 MΩ. Pour les pièces mesurées dont la perte diélectrique est appréciable, la lecture de capacitance obtenue est en fonction de la capacitance et de la résistance de perte. Bien qu'il ne soit pas possible, en pareil cas, de déduire les valeurs absolues des paramètres, des indications suffisantes sont obtenues pour établir s'il y a lieu de recourir à une méthode de mesure plus précise.

Accord réel des troisièmes harmoniques par F. D. Bate

Résumé de l'article aux pages 688 à 691

Cet article constitue une analyse de l'accord des troisièmes harmoniques, tel qu'il s'effectue dans les alimentations très haute tension à impulsions de retour en ligne, en assumant que les deux fréquences présentes soient f et 3f.

A l'aide d'un tracé de différentes courbes on a pu aboutir au principe d'un accord optimum de troisièmes harmoniques.

Circuits RC non-linéaires à varistors au carbure de silicium par W. G. P. Lamb

Résumé de l'article aux pages 692 à 693

Cet article indique que le comportement transitoire des séries non-linéaires de circuits à résistance-capacité peut être démontré ou utilisé pour la production de fonctions en utilisant des résistances non-linéaires en liaison avec des amplificateurs opérationnels.

Un frein électromagnétique actionné par les muscles des sourcils par L. Vodovnik

Résumé de l'article aux pages 694 à 695

Après avoir examiné les processus mis en action lorsqu'un conducteur décide de freiner, l'auteur considère les possibilités de pouvoir réduire la durée de réaction. Il décrit un système expérimental à frein électromagnétique mis en oeuvre par le potentiel d'action des muscles des sourcils. Il propose des améliorations éventuelles ainsi que d'autres applications possibles (par exemple, par les amputés).

La résistance d'entrée d'amplificateurs à réaction à transistors par E. de Boer

Résumé de l'article aux pages 702 à 706

Cet article décrit une théorie générale de la contre-réaction s'appuyant sur les niveaux d'impédance du circuit, plutôt que sur l'amplification. Des systèmes à réaction complexes ou simples peuvent être étudiés et on peut obtenir presque par contrôle une meilleure conception de leur fonctionnement. Les règles d'application de la théorie sont simples et logiques. Lorsque la réaction est prise d'un point en dérivation ou transmise à ce point, ce dernier peut être considéré comme ayant une résistance interne infinie. Lorsque au contraire, la réaction exige l'ouverture d'un circuit en boucle, ce dernier doit être considéré comme ayant une résistance intérieure nulle. Toute résistance interne propre de ces points doit être absorbée par la source ou la charge. Un certain nombre d'exemples, comprenant, entre autres, la réalisation d'un amplificateur de ligne destiné au fonctionnement entre des niveaux d'impédance déterminés, mettent en lumière les points marquants de la théorie.

DIE LUFTFAHRTSCHAU

Beschreibung einiger auf der Luftfahrtschau des Vereins Britischer Flugzeugkonstrukteure (SBAC) gezeigten elektronischen Geräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 713 bis 719

AMPLIVOX LTD

Beresford Avenue, Wembley, Middlesex

MINIATUR-SENDER-EMPFÄNGER

(Abbildung Seite 713)

Der "Televox" ist batteriegespeist, volltragbar und ohne Schwierigkeiten in die Tasche zu stecken. Es handelt sich hier um eins der kleinsten tragbaren, in England gefertigten UKW-Sprechfunkgeräte und wird dort viele Anwendungsmöglichkeiten haben, wo Nachrichtenverkehr zwischen verteiltem Personal erfordlich ist.

In der Konstruktion wurde den unterschiedlichen Anforderungen verschiedener Industrien besondere Aufmerksamkeit gewidmet, und mehrere einzigartige Eigenschaften sind vorgesehen:

(1) In jedem Sender-Empfänger gibt es Schaltungen für bis zu vier kristallgesteuerte Kanäle. Verkehr kann daher auf derselben Frequenz abgewickelt werden; andererseits kann man anderen Personen im selben Gebiet verschiedene Kanäle zuteilen.

(2) Der Televox wird mit einem abnehmbaren Batterieteil geliefert werden und wahlweise auch ein etwas grösserer Batterieteil für ungewöhnlich lange Betriebsdauer zur Auslieferung gelangen. Die Batterieteile sind für einfache Montage an die Grundplatte des Radioteils ausgelegt. Mittels eines Steckerzwischenstückes lässt sich statt des Batterieteiles eine getrennte Stromversorgung anschliessen.

(3) Der Televox wird ein eingebautes Lautsprecher-Mikrofon haben und ist daher ein völlig geschlossenes Gerät. Durch einfaches Umschalten ist es möglich, auch einen Handmikrofon-Lautsprecher, z.B. das abgebildete Amplivox-Mini-Mike, oder in sehr geräuschvoller Umgebung einen geräuschunterdrückenden Kopfhörer, z.B. der Amplivox-Ampligard-Serie, zu benutzen.

(4) Eine Teleskopantenne ist eingebaut, für grössere Reichweite kann man jedoch eine Aussenantenne anschliessen. Eine Qualitätsledertasche, Kristalle, sowie eine Auswahl von Mikrofonen, Abhörgeräte und Kopfhörer gehören zum Zubehörsortiment.

EE 74 751 *für weitere Einzelheiten*

BRITISH COMMUNICATIONS CORPORATION LTD

South Way Exhibition Grounds, Wembley, Middlesex

UHF-FUNKFERNSPRECHER

(Abbildung Seite 713)

Der laufend wachsende Bedarf an neuen Diensten hat bereits zur Zuteilung von Gemeinschaftswellen in den überbelegten UKW-Bändern geführt. BCC hat daher der Benutzung des UHF-Bandes durch Einführung neuer beweglicher und ortsfester Funkfernsprechanlagen vorgegriffen, die erstmalig ausgestellt wurden.

Der BCC 55 ist eine 6-Kanal-10 W-PM - Sender - Empfängerkombination. Beide Ausrüstungen sind für Betrieb im Band 440 ... 470 MHz ausgelegt.

EE 74 752 *für weitere Einzelheiten*

UKW-FUNKFERNSPRECHER

(Abbildung Seite 713)

Diese leichten, tragbaren Geräte überstreichen alle UKW-Bänder und sind in Ausführungen für AM und PM lieferbar. In jedem Gerät lassen sich sechs Kanäle einstellen, und der eingebaute Lautsprecher kann auch als Mikrofon benutzt werden. Zusätzlich zu der bereits umfangreichen Zubehörliste ist auch eine Trockenbatterieversorgung lieferbar, die neun Standard-Trockenelemente U7 enthält und eine Alternative zum aufladbaren Nickel-Kadmium-Akkumulator ist. Die Trockenbatterien haben dieselben Abmessungen wie der aufladbare Akkumulator und können daher direkt ausgewechselt werden.

Der Frequenzbereich ist 41 bis 174 MHz und wird entweder mit 25 kHz oder 50 kHz Kanalabstand überdeckt. Die Ausgangsleistung des Senders hängt von der Frequenz und Modulationsart ab und liegt zwischen 120 und 500 mW; die Frequenzkonstanz ist besser als ± 2 kHz.

Die maximale Tonfrequenzleistung des Empfängers ist 50 mW; die Empfindlichkeit ergibt eine Kurve, die innerhalb $\pm 7{,}5$ kHz besser als 3 dB, über ± 25 kHz hinaus und bei 25-kHz-Kanalabstand besser als -60 dB ist.

EE 74 753 *für weitere Einzelheiten*

HF-TRAGGERÄT

(Abbildung Seite 714)

Typ BCC 30HP ist ein volltransistorisiertes HF-Traggerät, das hohe Ausgangsleistung (20 W) mit einem niedrigen Gesamtgewicht von rund 16 kg vereint. Durch sein niedriges Gewicht eignet sich der BCC 30 besonders für Fallschirmjäger zur Herstellung von Verbindungen über Entfernungen, die UKW-Geräte nicht mehr überbrücken können. Ausser dem niedrigen Gesamtgewicht für eine komplette 20-W-Tragstation zeigen sich als Hauptmerkmale seine umfangreichen Einrichtungen. Sowohl CW-Betrieb wie AM- und PM-Funkfernsprechen sind innerhalb des Frequenzbandes 2 ... 8 (oder 2, 5 ... 10) MHz in 18 quarzgesteuerten Kanälen vorgesehen. Ausserdem kann das gesamte Band mit einer Selektivität durchgestimmt werden, die 600 Sprechkanäle ermöglicht. Ein breites Sortiment von leichten Zusatzgeräten wie z.B. Handgenerator-Stromversorgungen, Batterieladegeräte für Gleich- und Wechselstrom, sowie ein Montage-Bausatz für Wageninstallation sind lieferbar. Durch kleine Abmessungen und sehr niedrigen Energiebedarf ist der BCC 30 vor allem für Mehrzweckeinsatz als Traggerät, ortsfeste, oder bewegliche Ausrüstung geeignet.

EE 74 754 *für weitere Einzelheiten*

A. C. COSSOR LTD

The Pinnacles, Elizabeth Way, Harlow, Essex

BORDTRANSPONDERGERÄTE

(Abbildung Seite 714)

Ausser einer kompletten Bodenstation für Sekundär-Überwachungsradar wurden zwei Bordtransponder ausgestellt, und zwar die Militärausführung 1500 und das Zivilmodell 1600.

Modell 1500 ist ein miniaturisierter Kennungstransponder, der im Auftrag des britischen Luftfahrtministeriums für Dauerbetrieb ohne Luftkühlung in dem äusserst breiten Temperaturbereich von $-55°C \ldots +140°C$ entwickelt wurde und damit einen neuen hohen Massstab für Zuverlässigkeit von Bordausrüstungen darstellt.

Die Eigenschaften des 1600 entsprechen den Vorschriften des ICAO-Anhanges 10 sowie des ARINC-Pflichtenblattes 532D und genügen den Zulassungsbestimmungen der britischen und amerikanischen Luftfahrtbehörden mit 12-Bit-Verschlüsselung in Betriebsarten A, B, C und D. Mit Ausnahme der Senderendstufe, die die 1 kW übersteigende Ausgangsleistung abgibt, ist das Gerät volltransistorisiert. Die Ausrüstung ist für Dauerbetrieb im Temperaturbereich $-40°C \ldots +55°C$ mit Konvektionskühlung und für Betrieb ohne Druckkabine bis zu 12 200 m geeignet. Eine vom Bediengerät im Besatzungsabteil ausgelöste automatische Prüfeinrichtung ist eingebaut.

EE 74 755 *für weitere Einzelheiten*

MINIATUR-FUNKFERNSPRECHER

(Abbildung Seite 714)

Das Companion-Programm besteht aus einer Serie vom Benutzer getragener Funksprechgeräte, und zwar u.a. aus dem UKW-FM-Taschengerät CC2/8, dem AM-Traggerät CC3 und dem neuen leichten 6-Kanal-FM-Traggerät CC4.

Der bei Abmessungen von 143 × 89 × 38 mm nur ca. 450 g wiegende CC2/8 hat einen Frequenzbereich von 71,5 ... 174 MHz und eine Senderleistung von 500 mW. Die Empfängerempfindlichkeit ist 0,6 μV für 20 dB Geräuschdämpfung und der Störsignalpegel besser als −60 dB. Mit einer Tonfrequenzleistung von 200 mV lässt sich das Gerät bei hohem Lärmspiegel einsetzen. Wird das Grundgerät mit verschiedenen Zusätzen zusammen eingesetzt, lassen sich den einzelnen Anforderungen entsprechend unterschiedliche Anordnungen zusammenstellen; die Ausgangsleistung kann den Betriebsanforderungen—besonders hinsichtlich der Lebensdauer der Batterie—angepasst werden.

EE 74 756 *für weitere Einzelheiten*

DECCA RADAR LTD

Albert Embankment, London, S.E.1

FS-ÜBERWACHUNGSRADAR

(Abbildung Seite 714)

Das Flugsicherungs-Überwachungsradar AR1 ist eine detailreiche 10-cm-Anlage für hohe Datenraten, die bereits für die Kgl. Luftwaffe und Behörden der Zivilluftfahrt in Serienfertigung ist. Das AR1 kann alle Flugsicherungsaufgaben innerhalb 120 km vom Abfertigungsgebäude erfüllen und wurde hauptsächlich zur Befriedigung der Betriebsanforderungen derjenigen Flughäfen entwickelt, die den für Installation von Grossanlagen mit unterstützenden Nahbereichausrüstungen erforderlichen Aufwand nicht rechtfertigen können. Decca hat von Anfang an in der Konstruktion Vielseitigkeit, niedrige Installationskosten sowie geringe laufende Betriebskosten betont und im Typ AR1 somit ein kompaktes Überwachungsradar geschaffen, das den wachsenden Bedarf der Zivilluftfahrt an einem praktischen Radarsystem für alle Plätze, die Hochleistungs-Verkehrsmaschinen abfertigen, befriedigt. Mit seinem stabilen transistorisierten MTI-System mit keinen Blindgeschwindigkeiten unter 1038 km/h und einem regelbaren Polarisationssystem sehr fortschrittlicher Konstruktion kann das Radar die ICAO-Anforderungen für alle Nahaufgaben bis zum Rundsichtanflug herunter wesentlich überfüllen und gleichzeitig ein Überwachungspotential bis zu ungefähr 160 km bei 12 200 m Höhe—je nach Flugzeuggrösse—geben.

Das Radar AR1 wurde von der Kgl. Luftwaffe sowohl technisch wie auch betriebsmässig voll erprobt.

EE 74 757 *für weitere Einzelheiten*

SICHTGERÄTE

(Abbildung Seite 714)

Betriebsmässig vorgeführt wurde ein integriertes Flugsicherungssystem, in dem die über eine Bord-Bodenverbindung übermittelten Decca-Navigator-Daten gleichzeitig mit den Überwachungsradardaten auf autonomen Sichtgeräten der Decca-Serie 5 dargestellt wurden. In Farnborough wurde das Sichtgerät vom Überwachungsradar der FS-Stelle gesteuert, während die Bord-Bodenverbindung für den Decca-Navigator unmittelbar neben dem Hauptausstellungszelt installiert war.

Für diese neue Flugsicherungsdarstellung wurde das Sichtgerät 12C der autonomen Transistor-Sichtgeräte-Serie 5 der Decca Ltd eingesetzt. Die in diesem Gerät zur Verfügung stehenden Einrichtungen umfassen eine normale Rundsichtdarstellung mit Dezentrierung auf bis zu acht Radien, drei kennzeichnend geformte Marken, deren Stellung durch die von den Bord-Decca-Navigator-empfängern übertragenen Daten bestimmt wird, und ein Markensymbol unter örtlicher Kontrolle am Sichtgerätstandort.

Einsatz eines Standard-Sichtgerätes für diese verhältnismässig komplexe Darstellung wird durch die Vielseitigkeit der Serie-5-Modelle ermöglicht. Betriebliche Anforderungen können sich von der einfachen Panorama- oder Abstand-Höhenanzeige bis zu solchen erstrecken, die elektronische Markensymbole, Entfernungs- und Peilwinkelmesslinien, direkte VHF-Peilanzeige auf Elektronenstrahlröhren, Markierungsübertragung zwischen Konsolen und ähnliche Datenaufbereitungshilfen erfordern. Die Vielseitigkeit der Serie 5 ermöglicht 100 Abwandlungen, die allen gestellten Anforderungen genügen können. Ein weiteres Anwendungsbeispiel dieser Geräte aus jüngster Zeit ist das Eurocontrol-Simulatorprojekt, für das Decca, Telefunken in Berlin und CSF in Paris gemeinsam verantwortlich sind. Für das Primär-Darstellungssystem der Simulatoranlage in Bretigny wird eine Konsole in Sonderausführung mit diesen Sichtgeräten erstellt.

EE 74 758 *für weitere Einzelheiten*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

AUTOMATISCHES UKW-PEILGERÄT

(Abbildung Seite 715)

Dieses äusserst genaue, preisgünstige System besteht aus drei Einheiten: dem Sichtgerät, einem Gestell und einer Antenne.

Das Anzeigegerät enthält eine 14-cm-Oszillografenröhre, auf der die Peilergebnisse dargestellt werden, und die Bedienelemente. Randbeleuchtung der Skala erlaubt Beobachtung bei Verdunkelung.

Die Hauptausrüstung ist im Gestell untergebracht, das üblicherweise in einer geeigneten Hütte am Fuss des Antennenturmes untergebracht wird. Sie besteht aus einem Empfänger, zwei Geräten, in denen die Antennenpeilwinkelinformation in Anzeigeimpulse umgewandelt wird, einem Abschlussgerät für die Fernsteuerung und zwei Stromversorgungen.

Die Adcock-Antenne mit Seitenbestimmungselement ist in 8,35 m Höhe an einen Stahlrohrturm montiert. Sie wird kontinuierlich durch einen Synchronmotor mit 250 UPM gedreht. Das Gehäuse für den Motor enthält auch eine induktive Ankopplung für die Zuleitung und einen mit der Antennenwelle gekuppelten Drehfeldgeber. Der Antennenturm ist am Fuss mit Scharnieren versehen, so dass er sich für Wartung einfach umkippen lässt.

Die Betriebsweise ist wie folgt:

Eine Adcock-Antenne wird kontinuierlich mit 250 UPM rotiert und ihr Ausgang mit einem kleinen Anteil eines Allrichtungssignals von einem mittig zwischen die Adcock-Elemente montierten Dipol kombiniert.

Da das Antennenpolardiagramm näherungsweise die Form einer Acht darstellt, in der die Nullpunkte sich nicht genau gegenüber liegen, entsprechen aufeinanderfolgende Nullpunkte während des Empfangs abwechselnd etwas mehr oder etwas weniger als 180° der Antennendrehung.

Diese Nullpunkte erscheinen im Ausgang des zweiten Detektors des Empfängers, sind aber in dieser Stufe für Seitenkennung oder Anzeige ungeeignet, da ihre Schärfe auf Grund der empfangenen Signalstärken und anderer Faktoren schwankt. Sie werden daher an eine Impulsgeberschaltung gelegt, in der entsprechende schmale Impulse konstanter Breite erzeugt werden. Diese besteht im wesentlichen aus einer Einmal-Kippschaltung, die durch eine vorgewählte Signalspannung vor dem Nullpunkt getriggert wird und mit halber Geschwindigkeit läuft, bis das Signal nach dem Nullpunkt denselben vorgewählten Pegel erreicht hat. An diesem Punkt wird der Ablauf mit voller Geschwindigkeit fortgesetzt, bis die untere Grenze der Kippschwingung erreicht und der Rücklauf eingeleitet wird.

Durch den Rücklauf wird ein schmaler Impuls mit konstanter Verzögerung nach dem Zeitmittelpunkt zwischen den Zeitpunkten erzeugt, in denen das Signal vor und nach dem Nullpunkt dieselbe Spannung erreicht. Der Zeitpunkt des auf diese Weise erzeugten Impulses wird innerhalb des Betriebsbereiches der Ausrüstung daher nicht durch die Nullpunktschärfe beeinflusst, wenn die Signalspannung um den Nullpunkt herum symmetrisch ist. Der Einfluss der für die Seitenbestimmung erforderlichen leichten Unsymmetrie wird durch Begrenzung des Arbeitens der Schaltung auf das unmittelbare Nullpunktgebiet so klein wie möglich gehalten. Breite Nullpunkte ausserhalb des Arbeitsbereiches der Ausrüstung werden durch eine Breiten-Torschaltung ausgeschlossen.

Die vom Impulsgeber gebildeten schmalen Impulse werden an einen Funktionsdrehgeber gelegt, der mit der Antennenwelle gekuppelt ist und dadurch auf die Antennenstellung bezogen in Sinus- und Kosinuskomponente unterteilt. Diese Komponenten werden an die Abtastspulen einer Elektronenstrahlröhre gelegt, so dass die Ablenkungen eine im wesentlichen radiale Richtung ergeben, die von der Winkelstellung der Antenne zu dem Zeitpunkt, in dem ein Impuls auftritt, abhängt.

Die Abtastspulen werden in geeigneter Weise in Bezug auf die Peilskala um den Bildschirm herum ausgerichtet, so dass die Antennenstellung in diesem Augenblick und damit die Richtung der Signalquelle angezeigt wird.

Zur Bestimmung der Peilseite werden die erzeugten schmalen Impulse getrennt an eine Schaltung gelegt, die die Helligkeit der Elektronenstrahlröhre steuert; sie blendet Impulse aus, denen Intervalle von weniger als 180° Antennendrehung vorausgehen.

EE 74 759 *für weitere Einzelheiten*

FERRANTI LTD

Hollinwood, Lancashire

AUSSENLÄUFERMOTOREN

(Abbildung Seite 715)

Eine ausgestellte neue Serie von Aussenläufermotoren für 400 Hz Wechselstrom entspricht im allgemeinen dem Pflichtenblatt EL.1892 des Luftfahrtministeriums. Die Motoren wurden ursprünglich als Gyroskopmotoren entwickelt, haben inzwischen aber auf anderen Gebieten Verwendung gefunden, u.a. für kleine Gebläse, kombinierte Gebläse-Kraftquellen, Trommeltriebe, Kraftriemenscheiben und Krafttaumelscheiben. Ihr hoher Wirkungsgrad ermöglicht ein hohes Leistung-Volumenverhältnis, das überall dort von Wert ist, wo Raum, Gewicht, Temperatur oder begrenzte verfügbare Energie eine Rolle spielen.

EE 74 760 *für weitere Einzelheiten*

BEWEGLICHE KARTENDARSTELLUNG

Die bewegliche Kartendarstellung gibt sowohl dem Flugzeugführer wie Navigator ein optisch projiziertes Kartenbild der überflogenen Gegend. Der Standort des Flugzeuges und Kurs über Grund werden auf dem Mattscheibenschirm des Gerätes als ein zentrischer Kreis und eine radiale Linie gezeigt. Das von einem Farbfilmstreifen der entsprechenden topografischen Karten optisch projizierte Kartenbild bewegt sich hinter diesen Anzeigen. Der benutzte Mikrofilm der Karten ist 20mal verkleinert und 20mal vergrössert, wenn der Film projiziert wird.

Für Änderungen in östlicher Richtung wird der Film längsweise, für Änderung in nördlicher Richtung quer getrieben, wobei der Eingang der Stellmotoren für jede Achse vom Doppler- oder Trägheits-Navigationssystem über den Navigationsrechner und einen elektromechanischen Analogrechner gespeist wird. Die Sichtgeräte geben sowohl Flugzeugführer wie Navigator eine leicht deutbare bildliche Darstellung der Lage während des Einsatzablaufes. Eine Standortmarke zeigt laufend, wo sich das Flugzeug befindet, und die radiale Kurslinie, welche Punkte die Maschine überfliegen wird. Bei gutem Wetter kann der Flugzeugführer die Darstellung nutzen, um die Strecke nach der Karte zu fliegen, wobei er das Kartenbild mit dem Terrainbild, das er selbst sieht, in Einklang bringt. Er kann Sichtstandortbestimmungen vornehmen und das fehlerfreie Funktionieren des Navigationssystems kontrollieren.

Der Navigator benutzt die bewegliche Karte als Lagesichtanzeige, die Navigationsinformation darstellt, sowie zur Überwachung des Navigationssystems. Er kann auch den Ausgang seines seitwärtssehenden Radars mit seiner Kartendarstellung vergleichen, um eine Radar-Standortbestimmung durchzuführen. In dieser Betriebsart hilft das Gerät das Erscheinen der Radar-Bestimmungspunkte vorauszusehen und sie zu erkennen, wenn sie erscheinen.

EE 74 761 *für weitere Einzelheiten*

THE MARCONI CO. LTD

Chelmsford, Essex

ENTFERNUNGSMESSGERÄT

(Abbildung Seite 716)

Das DME-Bordabfragegerät AD70 misst laufend und automatisch die Schrägentfernung vom Flugzeug zu einem gewählten Bodenfunkfeuer. Die Entfernung wird am Instrumentbrett des Flugzeugführers auf einem Messgerät direkt angezeigt. Der mit diesem Gerät verwendete Bodentransponder kann entweder zu einer VORTAC- oder TACAN-Anlage gehören, die beide ausser der DME-Funktion auch andere Navigationsaufgaben erfüllen.

Zusammen mit einem VOR- oder Funkkompass-Einzelpeilergebnis ermöglicht die Entfernungsinformation die genaue Bestimmung des Flugzeugstandortes. Die meisten Bodenanlagen haben ein VOR-Drehfunkfeuer neben dem DME-Transponder, wie z.B. in der VORTAC-Ausrüstung.

Ein am Ende der Landebahn aufgestelltes DME-Funkfeuer kann eine wertvolle zusätzliche Landehilfe bilden, die eine laufende Anzeige der Entfernung bis zum Aufsetzpunkt gibt.

Die Entfernungsmesserausrüstung AD70 besteht aus 17 modularen Einzelbausteinen, die in einen 1/2ATR-Aufbaurahmen aus Aluminium eingebaut sind und insgesamt 14 kg wiegen. Die Ausrüstung ist mit Ausnahme derjenigen Stufen, in denen hohe Leistungen oder Konstruktionsrücksichten es nicht zulassen, mit Transistoren bestückt. Einsteck-Druckschaltungsplatten finden weitgehend Verwendung, und die betriebsmässige Routineprüfung wird durch Verlängerungsplatten erleichtert. Es gibt 126 quarzgesteuerte Frequenzkanäle, die Zahl kann aber auf 252 vergrössert werden, sollte es durch Ausbau des DME-Systems in Zukunft erforderlich werden.

Frequenzen werden mittels Dekadenschalter in einem getrennten Bediengerät eingestellt, und sowie die gewählte Frequenz einmal eingestellt ist, arbeitet das Gerät völlig automatisch.

Der Sender strahlt eine laufende Folge von Abfrage-Impulspaaren aus, die jeweils im Bodentransponder nach einem Festzeitintervall Ausstrahlung von Antwortimpulsen bewirken. Die Entfernung des Flugzeuges vom Bodenfunkfeuer kann dann durch das Zeitintervall zwischen jedem Abfrage-Impulspaar und seiner Beantwortung gemessen werden. Das Gerät bestimmt das Zeitintervall mittels einer zeitempfindlichen Empfangstorschaltung, die nur die während eines diskreten Zeitintervalls nach Ausstrahlung eines Abfrage-Impulspaares vom Bodentransponder empfangenen Signale durchlässt. Der Abstand zwischen diesem Zeitintervall und der Ausstrahlung wird automatisch zwischen Null und der einer Entfernung von 322 km entsprechenden Zeit variiert. Wenn ein Signal in dieser Torzeit ankommt, so wird das Gerät von diesem Signal "mitgenommen" und die Einstellung des Tors so geändert, dass die Ausrüstung einfallenden Signalen mit der Flugzeugbewegung auf das Funkfeuer zu oder von ihm weg folgen kann. Die dem momentan gemessenen Zeitintervall entsprechende Entfernung wird gleichzeitig auf dem Anzeigegerät im Instrumentenbrett des Flugzeugführers dargestellt.

Eine bemerkenswerte Einrichtung des AD70 ist die Geschwindigkeits-Speicherschaltung, die gewährleistet, dass bei momentanem Signalverlust keine Unterbrechung in der Entfernungsinformation eintritt. Das Geschwindigkeitsgedächtnis speist so lange ein simuliertes Signal in die Folgeschaltungen, bis das wahre Signal wieder empfangen wird, und zwar bis zu maximal zehn Sekunden. Wenn in dieser Zeit das Signal nicht wieder ankommt, geht das Gerät wieder in die Bereitschaftsarbeitsweise zurück und beginnt die Suche nach einem Signal über den vollen Entfernungsbereich.

EE 74 762 *für weitere Einzelheiten*

HF-FESTKÖRPER-BORDAUSRÜSTUNG

(Abbildung Seite 716)

Dem HF-Nachrichtenverkehrssystem AD460 liegt dasselbe Konstruktionskonzept für Zuverlässigkeit zugrunde, mit dem man die anderen Geräte der Marconi-Serie Sechzig verbindet. Die Konstruktion folgt durchweg dem neusten Stand der Technik, und in allen Entwurfs- und Fertigungsstufen untergeht jedes Gerät die schärfsten Tests, um absolute Zuverlässigkeit zu gewährleisten.

Der AD460 kann mittels Dekadenschalter in einem getrennten Bediengerät über das Frequenzband 2 ... 29 999 MHz durchgestimmt werden. Ein Rastabstand von ½ kHz sollte allen zukünftigen Anforderungen nach Extra-Kanälen in dem stark überbelegten HF-Band genügen.

Für alle Schaltfunktionen werden Halbleiter-Dioden benutzt, und mit Ausnahme der Endstufe findet durchweg Festkörperabstimmung Verwendung. Die Betriebsfrequenzen werden nach dem Frequenzsyntheseverfahren erzeugt, in dem die einzelnen Dekadenoszillatoren mit dem frequenzkonstanten Bezugsoszillator phasenstarr laufen. Da das Gerät nur mit Silizium-Transistoren bestückt ist, kann es zuverlässig in einem Temperaturbereich von −55° C ... +55° C arbeiten.

Als Betriebsarten gibt es u.a.: Zweiseitenband für Sprechfunk und CW, Einseitenband mit Trägerunterdrückung für Sprechfunk und Einseitenband (mit Trägersteuerung und AFN) für Sprechfunk oder Zweiton-Frequenzumtastung für Datenübertragung. Selektivrufbetrieb (SELCAL) nach ARINC-Pflichtenblatt 531 ist möglich, und für Datenübertragung sind getrennte Ein- und Ausgangsanschlüsse vorhanden. Auch Schmal- und Breitband-CW-Sendebetrieb ist möglich, wofür ein getrennter Schwebungsgenerator vorhanden ist.

Die Ausgangsleistung des AD460 ist ½ kW, jedoch ist auch eine Ausführung für 1 kW—Typ AD470—lieferbar. Umfangreiche Testeinrichtungen zum schnellen Prüfen in allen Betriebsarten sind eingebaut.

EE 74 763 *für weitere Einzelheiten*

MARCONI INSTRUMENTS LTD

St. Albans, Hertfordshire

AM-Messender

(Abbildung Seite 716)

Der AM-Messender TF801D/SM1 entspricht im allgemeinen dem Standardmodell TF801D/1 (10 ... 470 MHz), jedoch sind für das Prüfen gewisser Navigationshilfen notwendige Sondereinrichtungen vorhanden, und die Frequenzüberdeckung wurde zur Erzielung eines besseren Unterscheidungsvermögens in den entsprechenden Bändern geändert. Der Ausgang gibt eine von 0,1 μV ... 1 V kontinuierlich regelbare EMK ab, und die Frequenzdrift ist geringer als $\pm 5 \times 10^{-5}$/10 min.

Die durch externe Modulation bei 30 Hz bewirkte Phasenverschiebung wird vernachlässigbar klein gehalten, um VOR-Anforderungen zu genügen, und die von einem internen Detektor demodulierte Wellenform kann man an der Frontplatte für Vergleichszwecke entnehmen; das UKW-Trägerfrequenzband liegt ungefähr in der Mitte des vierten Teilbereiches des Messenders (70 ... 150 MHz).

Es gibt auch einen internen quarzgesteuerten Oszillator, der ein genaues 75-MHz-Signal abgibt, das z.B. für Tests von Markierungsfunkfeuerempfängern gebraucht wird. Üblicherweise wird er zusammen mit dem dritten Teilbereich des Senders (48 ... 110 MHz) benutzt, wobei der durchstimmbare Oszillator abgeschaltet wird, der Verstärker aber weiterhin funktioniert. Dadurch erhält man ein Signal mit hohem Leistungspegel, das tief moduliert werden kann. Der Verstärker des vierten Teilbereiches ist auch für denselben Zweck geeignet, wenn das bequemer ist.

EE 74 764 *für weitere Einzelheiten*

Abstimmbares Sperrfilter

(Abbildung Seite 717)

Ausser seiner Verwendung als Zubehör für den Marconi-Wellenformanalysator TF2330, mit dem zusammen es das Messen relativer Amplituden, die mehr als 75 dB voneinander abweichen, erlaubt, hat dieses Filter auch andere Anwendungsmöglichkeiten, bei denen eine Einzelfrequenz unterdrückt oder eliminiert werden muss.

Typ TF2334 gibt bei Minimumschwächung der Harmonischen über 80 dB Unterdrückung der Grundwellen im Bereich 20 Hz ... 20 kHz. Es ist im Grunde genommen ein hochohmiges Doppel-T-Filter, das für Quellen mit weniger als 1000Ω Widerstand ausgelegt ist. Ein hochohmiger oder 600-Ω-Eingang stehen zur Verfügung, und bei Benutzung des letzteren kann eine Kompensation für Abschwächung der Harmonischen bis zur vierten eingeschaltet werden, so dass man deren Werte direkt am Wellenformanalysator ablesen kann.

Da keine aktiven Bauelemente verwendet werden, wird keine Stromversorgung gebraucht. Das Gerät ist als Tischinstrument oder für Gestelleinbau lieferbar.

EE 74 765 *für weitere Einzelheiten*

PLESSEY-UK LTD

Ilford, Essex

Flugdaten-Aufzeichnungssystem

Diese Ausrüstung wurde von BEA für alle Flugzeuge und von Aer Lingus für alle ihre BAC 111 bestellt und genügt völlig der Auflage des britischen Luftfahrtministeriums für Erstellung eines Protokolls, das Stoss, Feuer und chemische Angriffe überlebt.

Ausser den sechs vom britischen Zivilluftfahrt-Ministerium vorgeschriebenen Parametern—angezeigte Eigengeschwindigkeit, angezeigte Bodennähe, missweisende Peilung, senkrechte Beschleunigung, Längsneigungslage, Zeit—ermöglicht die Konstruktion durch Bereitstellung von Reservekanälen für Extra-Samplingschaltungen noch Spezialuntersuchungen. Digitalaufzeichnung aller Signale gewährleistet die Aufrechterhaltung der Genauigkeit während des gesamten Betriebsablaufes, von der Aufzeichnung während des Fluges bis zur Datenverarbeitung und Darstellung auf dem Boden, und erlaubt auch Einbau automatischer Prüfeinrichtungen.

Das System besteht aus drei Hauptgeräten: dem Kommandogerät in der Flugzeugkanzel, einem Verarbeiter und dem Aufzeichnungsgerät, an die eine Anzahl Messwertgeber angeschlossen sind. In dem Aufzeichnungsgerät—eine Sonderentwicklung der Firma S. Davall & Sons Ltd—ergibt magnetische Aufzeichnung auf nichtrostendem Stahldraht einen viel grösseren Widerstand gegen die beim Absturz auftretenden Gefahren, als andere Informationsträger.

Das eigentliche Aufzeichnungsgerät wird im Heck der Maschine eingebaut und ist so robust konstruiert, dass automatischer Auswurf nicht erforderlich ist. Es ist z.B. für Belastungen bis zu 1 Tonne, Beschleunigungen bis zu 100 g, Temperaturen bis zu 800°C für 15 Minuten und vollständiges Eintauchen in Seewasser für bis zu 30 Stunden bemessen.

Jedes Gerät hat eine Laufzeit von 200 Stunden ohne Wartung und wird automatisch bei Beginn und Ende jedes Fluges durch die Messwertgeber des Fahrtmessers ein- und ausgeschaltet.

EE 74 766 *für weitere Einzelheiten*

Anschlagwerkzeuge

(Abbildung Seite 717)

Es werden jetzt kraftbetätigte Anschlagwerkzeuge eingeführt, bei denen es sich um druckluftgetriebene Ausführungen des Hand-UNIcrimp handelt, ein universelles, mit einer Hand betätigtes automatisches Werkzeug, das fehlerfreie rechteckige Quetschverbindungen an Draht oder Kontakten aller Abmessungen bis zu 12 AWG (2 mm) oder 40 × 0,19 mm herstellen kann.

Die kraftbetätigte Ausführung wird in zwei Formen angeboten: ein dem Hand-UNIcrimp ähnliches, tragbares Modell und ein neues Tischmodell mit schwerer Grundplatte für Verwendung in der Werkstatt. Die kraftbetätigten Modelle können sowohl vorisolierte Fahnen wie auch Steckkontakte anschlagen. In allen drei Ausführungen ist UNIcrimp eine narrensichere, vollerprobte Fertigungsmaschine für ungelernte Arbeitskräfte.

EE 74 767 *für weitere Einzelheiten*

RANK BUSH MURPHY ELECTRONICS

Welwyn Garden City, Hertfordshire

Funkfeuer

(Abbildung Seite 717)

Ein als Hilfe für die Luftunterstützung von vorgeschobenen Bodentruppen entwickeltes Funkfeuer wurde erstmalig auf der Luftfahrtausstellung Farnborough im September gezeigt. Das Gerät ist die jüngste Ergänzung des international berühmten Navigationssystems Eureka-Rebecca der Rank Bush Murphy Electronics.

Es handelt sich um ein mit Eureka MR 343 bezeichnetes, transistorisiertes Kleinfunkfeuer, das für Fallschirmabwurf mit Luftsturmtruppen entwickelt wurde, um für unterstützende Flugzeuge Entfernungs- und Zielflugsignale abzugeben und als Landehilfe auf vorgeschobenen Feldflugplätzen zu dienen.

Die erzielte Reduktion in Abmessungen, Gewicht und Leistungsbedarf ist beachtlich, und das Gerät kann von nichttechnischen Mannschaften bedient werden. Es hat eine eingebaute Testeinrichtung für Empfänger und Sender.

EE 74 768 *für weitere Einzelheiten*

REDIFON LTD

Broomhill Road, London, S.W.18

Ortsveränderliches Funkfeuer

Das MF-Funkfeuer TG.142R wurde speziell für schnelles Aufstellen an neuen Standorten konstruiert. Die komplette Ausrüstung besteht aus acht transportierbaren Einheiten und kann von zwei Monteuren mittels eines zur Anlage gehörigen 6-m-Ladebaums und einschliesslich einer 18 m hohen vertikalen Mastantenne aufgestellt werden.

Die Ausrüstung besteht aus einem Sender, einem Antennenmast in sechs Schüssen von je 3 m Länge, 12 radialen Erdleitern von 33,53 m Länge, einem Antennenankopplungsgerät, zwei Reservebehältern, einem Zelt und einem Ladebaum.

Das Funkfeuer arbeitet im Frequenzbereich 200 ... 420 kHz im CW- oder tonmodulierten CW-Betrieb mit örtlicher oder Fernsteuerung. Zur Wahl der Signalinformation einschliesslich Rufzeichen wird eine Spezial-Tastscheibe, die sich leicht ändern lässt, benutzt.

EE 74 769 *für weitere Einzelheiten*

S. SMITH & SONS (ENGLAND) LTD

Kelvin House, Wembley Park Drive, Wembley, Middlesex

FLUGSTEUERUNGSSYSTEME

Das System Serie 6 ist das jüngste im Smith-Programm für Flugsteuerungsausrüstungen und wurde bereits für das Fokker-Transportflugzeug mit ZTL-Triebwerk, die Fellowship F28, spezifiziert.

Die Serie 6 umfasst den Flugregler SEP.6, das Flugsystem SFS.6, das Kompassystem SCS.6 und die Gashebelautomatik STS.6. Das nach dem Baukastenprinzip konstruierte System kann von einem einfachen automatischen Stabilisator auf die für Verkehr in reduzierten Wettermindestbedingungen erforderliche vollautomatische und Instrumentensteuerung ausgebaut werden.

Die Ausrüstung Serie 6 stimmt mit den ARINC-Empfehlungen überein und wird den Flugsteuerungsanforderungen von Maschinen verschiedenster Grösse vom Geschäftsflugzeug bis zur grössten Strahl-Grossverkehrsmaschine genügen.

Bei der Entwicklung des Flugsteuerungssystems, das allen Anforderungen der Zivilbehörden für Allwetterverkehr und Blindlandungen genügt, standen Sicherheit und Zuverlässigkeit im Vordergrund.

Ausser Blindlandung gibt das System die weiteste Auswahl von Flugregler- und Flugkommandobetriebsarten, einschliesslich automatisches Einkurven auf einen vorgewählten Kurs, automatisches Einfliegen auf eine vorgewählte Höhe, Höhenhaltung, geregelte Absinkrate, ILS- und VOR-Kupplung mit automatischer "über Flughafen"-Meldung, automatische oder manuelle Erfassung des Gleitweges und Absinken entlang dem Gleitweg.

Das ursprünglich in Duplexausführung in der Hawker-Siddeley Trident eingebaute Flugsteuerungssystem Serie 5 gibt diesem Flugzeug die Möglichkeit, automatisch auszuschweben. Die Trident ist die einzige in dieser Weise ausgestattete Maschine der Welt. Das von der Kgl. Luftwaffe bestellte neue Short-Frachtflugzeug Belfast mit Blindlandeanlage hat eine Triplex-Ausrüstung der Serie 5.

EE 74 770 *für weitere Einzelheiten*

THE SOLARTRON ELECTRONIC GROUP LTD

Farnborough, Hampshire

DIGITALDATENERFASSUNG

(Abbildung Seite 718)

Der "compact logger" ist ein kleines Digitaldatenerfassungssystem mit 20 Eingangskanälen und einer Höchstabtastgeschwindigkeit von 3 Messstellen/Sekunde. Das integrierende Digitalvoltmeter LM1420 mit einer Messunsicherheit von 0,05% bei einer Empfindlichkeit von 10 μV per Ziffer gehört zum System. In diesem Voltmeter wird Spannungs-Frequenzumsetzung benutzt, die vollisolierte Eingangsschaltungen erlaubt. Dadurch wird eine Gleichtaktunterdrückung von 150 dB und bei Benutzung des Eingangsfilters eine Serienunterdrückung von 60 dB erreicht.

Der "compact logger" ist in Modultechnik ausgeführt und kann—je nach den gewünschten Einrichtungen—in einem Zwei—oder Dreiebenengehäuse geliefert werden.

Als Einschubmoduln sind u.a. ein Bereichkommandogerät für den 20-Kanal-Abtaster des Digitalvoltmeters, Digitaluhr, Stanzverschlüssler, Druckwerktreiber, Schreibmaschinentreiber, Grenzwerterkennung, Linearisierung und Bezugsthermoelemente lieferbar. Die Einschübe lassen sich je nach Bedarf einbauen, und wo keine gebraucht werden, setzt man Leerfelder ein. Jeder Einschub hat seinen bestimmten Platz im Gehäuse.

Die Abbildung zeigt die Ausführung in drei Ebenen mit einem vollen Einschub-Komplement.

EE 74 771 *für weitere Einzelheiten*

VIDEOKARTE MIT HOHEM AUFLÖSUNGSVERMÖGEN

Die Solartron-Videokarte mit hohem Auflösungsvermögen kann in jedes Standard-Radarsystem integriert werden. Die Hersteller glauben, dass es zur Zeit die genauste greifbare Karte ist. Die Punktgrösse der Elektronenstrahlröhre ist $38,1 \times 10^{-3}$ mm; ein Leuchtstoff der Type "A" findet Verwendung. Die Röhre ist am Hals mit einer Spezialmanschette festgeklemmt, was zur Herstellung korrekter Beziehungen zur Schirmebene führt und das Fokussieren erleichtert.

Absichtliche Verzerrungen in den Ablenkverstärkern kompensieren für die durch den Planschirm der Röhre hervorgerufene Nichtlinearität, was die Kartenwiedergabe erleichtert. Zur Erzielung der gewünschten Genauigkeit werden hohe Anforderungen an Zeichenkunst und Fotografie gestellt.

Die technischen Daten sind wie folgt: Höchstbereich: 40 ... 400 Seemeilen (154 ... 1540 km) nach Kundenwunsch. Auflösungsvermögen: 1×10^{-3} (erlaubt fehlerfreie 20fache Dehnung). Genauigkeit: auf einer Karte mit 185,2 km Radius wird ein Punkt in 18,5 km Entfernung mit einer Unsicherheit von 183 m positioniert.

EE 74 772 *für weitere Einzelheiten*

ZWEISTRAHLOSZILLOGRAF

(Abbildung Seite 718)

Der neue Zweistrahloszillograf CD1400 ist mit einer 13-cm-Oszillatorröhre mit zwei Strahlsystemen bestückt und gibt eine grosse, helle Sichtanzeige. Sowohl X- wie Y-Systeme werden in das Gerät eingeschoben und treiben die Ablenkplatten direkt, so dass man ohne eingebaute Verstärker im Hauptgestell auskommt. Zur Zeit sind vier Einschübe lieferbar, andere in der Entwicklung. Der Breitbandverstärker CX1441 hat eine Bandbreite von 0 ... 15 MHz mit einem Frequenzgang von −3 dB und einem Ablenkfaktorbereich von 100 mV/cm ... 50 V/cm. Zusätzliche 10fache Verstärkung gibt es bei reduzierter Bandbreite.

Der Differenzverstärker mit hoher Verstärkung CX1442 hat von 0 ... 75 kHz einen Frequenzgang von −3 dB und einen Ablenkfaktorbereich von 1 mV/cm ... 5 V/cm. Zusätzliche, nur wechselspannungsgekoppelte 10fache Verstärkung gibt es bei reduzierter Bandbreite.

Das Zeitablenkgerät CX1443 bietet einen Geschwindigkeitsbereich von 0,5 μs/cm ... 200 ms/cm; ungefähr 5fache Dehnung steht zur Verfügung. Es gibt umfangreiche Triggereinrichtungen und HF-Synchronisierung von Y_1, Y_2 oder externe mit Polaritätswahl.

Das Zeitablenkgerät mit Verzögerung CX1444 entspricht dem CX1443, hat jedoch eine eingebaute Schriebverzögerung mit maximaler Verzögerungszeit von 100 ms. Amplituden und Zeit können in den meisten Bereichen mit 5% Unsicherheit gemessen werden.

Die Abbildung zeigt den CD1400 mit zwei Breitbandverstärkern und einem Zeitablenkgerät ausgerüstet.

EE 74 773 *für weitere Einzelheiten*

FS-SIMULATOR

(Abbildung Seite 718)

Ein Flugsicherungs-Simulator für vier Ziele wurde als Luftraumüberwachungs-Radarpanoramagerät vorgeführt. Es handelt sich um einen Simulator der Serie SY 2020, die für bestehende Aufträge gefertigt wird. Rationalisierte Modulkonstruktion und Serienherstellung führten zu wesentlich reduzierten Kosten, und gleichzeitig wurden durch technische Verbesserungen Genauigkeit und Zuverlässigkeit erhöht. Als Beispiel sei hier die Methode zur Integrierung der Zielposition genannt. Der neue elektronische Integrator ermöglicht erstmalig, den Standort eines Zielechos in bis zu 139 km Entfernung mittels SRE zu bestimmen und es mit ausreichender Gesamtgenauigkeit für Nahanflug mit Feinführungsradar bei realistischer Darstellung heranzubringen.

Die Abbildung zeigt die Verfahrensausbildung für die Präzisionsanflugradaranlage des Simulators der Kgl. Luftwaffe in Shawbury. Das Zielsteuergerät ist hinter dem Arm des Ausbilders zu sehen.

EE 74 774 *für weitere Einzelheiten*

STANDARD TELEPHONES & CABLES LTD

Connaught House, 63 Aldwych, London, W.C.2

SEKUNDÄR-ÜBERWACHUNGSRADAR

Der SGR.1 ist das Boden-Abfragegerät eines Sekundär - Überwachungsradarsystems. Es kann entweder unabhängig oder in Verbindung mit einem Primärradar eingesetzt werden und gibt im letzteren Falle eine in Wechselbeziehung stehende Sichtanzeige für die Flugsicherung.

Zwei Grundanordnungen sind für das System vorgesehen, je nachdem, ob die Antenne koaxial mit einer Primärradarantenne installiert wird oder als getrenntes System mit eigenem Schwenkgetriebe. Für die letztere Anordnung sind ein Schwenkmotor, ein Motorgenerator und eine Servosteuerungsanlage vorgesehen.

Die Abfrageausrüstung ist in einem Einzelschrank untergebracht, der einen Sender, einen Empfänger, die zugehörige Stromversorgung, Steuer- und Kontrollgeräte enthält.

Im Sender wird die über das Gitter mit Impulsen beaufschlagte, hochverstärkende Wanderfeldröhre zur Verstärkung von Impulsen aller Art benutzt, die dann in einen koaxialen Hochleistungs-Halbleiterschalter gespeist werden. Der Schalter, der eine sehr kurze Arbeitszeit hat, speist die Einzelimpulse über vorgegebene HF-Abschwächer je nach Bedarf entweder in den Abfrage- oder Steuerteil der Antenne.

Dieses Verfahren zur Ableitung der ausgestrahlten Abfrage- und Steuerimpulse von einem Sender über passive Netzwerke gewährleistet langfristige Impulshöhenkonstanz und macht das Abgleichen getrennter Sender überflüssig. Im Interesse langfristiger Konstanz und Zuverlässigkeit finden mit Ausnahme der Wanderfeldröhre durchweg Halbleiter Verwendung.

Das Antennenssytem besteht aus zwei in einer Gruppe zusammengebauten Antennen.

Die Ausrüstung kann entweder als Einzel- oder Doppelsystem geliefert werden. Im letzteren Falle sind Umschaltvorrichtungen zur Herstellung der Verbindungen zwischen Antenne und Abfrageschränken, für örtliche Steuerung und Überwachung, Fernsteuerung und -überwachung, sowie Antennenantrieb und Synchronisierung vorgesehen.

EE 74 775 *für weitere Einzelheiten*

MIKROMINIATUR-FUNKHÖHENMESSER

(Abbildung Seite 719)

STC hat zwecks Verbesserung der Zuverlässigkeit des FM-Tiefflug-Funkhöhenmessers für Blindlandesysteme ein Labor-Modell in Mikrominiaturtechnik hergestellt. Obwohl diese Entwicklung in erster Linie die Verbesserung der Zuverlässigkeit als Aufgabe hatte, ergaben sich bedeutende Einsparungen an Grösse, Gewicht und Leistungsaufnahme.

Die Fehlerlosigkeit der bestehenden STC-Funkhöhenmesser ist wohlbekannt. Sie wurden für Erstellung der Höheninformation in allen Allwetterlandesystemen verwendet, und zwar seit der Bildung der Blindlandungs-Versuchsabteilung seitens des britischen Luftfahrtministeriums bis zum Einbau in die Trident, VC.10, Belfast und zahlreiche andere Zivil- und Militärflugzeuge in jüngster Zeit. Innerhalb dieser Zeitspanne hat die Firma unaufhörlich danach gestrebt, Konstruktionsmethoden, Fertigungsverfahren sowie die Bedingungen, unter denen die Ausrüstungen hergestellt werden, zu verbessern. Diese Bemühungen haben in dem auf der Ausstellung vorgeführten miniaturisierten Funkhöhenmesser ihren Höhepunkt gefunden.

Der Höhenmesser besteht aus einem 4 300-MHz-FM-Festkörpersender, einem Videoverstärker und zugeordneten Schaltungen, sowie Frequenzzählschaltungen. Das Gerät bildet ein ideales Anwendungsbeispiel für die Mikrominiatur-Elektronik.

Das Modell wurde in einem kurzen 1/4ATR-Zwerggehäuse ausgestellt, das jedoch nicht notwendigerweise die endgültige Form darstellt. Es dient zur Veranschaulichung dessen, was mit modernen Methoden erreicht werden kann.

Das Foto zeigt den Mikrominiatur-Höhenmesser neben einem herkömmlich konstruierten Höhenmesser.

EE 74 776 *für weitere Einzelheiten*

VENNER ELECTRONICS LTD

Kingston By-Pass, New Malden, Surrey

BREITBAND-OSZILLOGRAF

(Abbildung Seite 719)

Der Breitband-Oszillograf TSA 625 überdeckt 10 Hz ... 1 MHz in fünf geschalteten Bereichen. Konstruktionsmerkmale des Gerätes sind u.a. Drucktasten-Bereichwahl und Batterieprüfung, Abschwächereinstellungen von 2,5 mV Vollausschlag bis zu 2,5 V und eine Ausgangsimpedanz von besser als 600Ω auf allen Bereichen.

Sowohl Sinus- wie Rechteckwellen sind vorhanden.

EE 74 777 *für weitere Einzelheiten*

WESTLAND AIRCRAFT LTD

Saunders-Roe Division, Strain Gauge Department, East Cowes, Isle of Wight

DEHNSTREIFEN-LÄNGUNGSMESSER

(Abbildung Seite 719)

Die Dehnungsmessstreifen-Abteilung des Geschäftsbereiches Saunders-Roe der Westland Aircraft Ltd hat einen Dehnstreifen-Längungsmesser einfacher Ausführung für Oberflächenmontage auf grossen Strukturen entwickelt. Er besteht aus einem Streifen aus Stahl mit kugelförmigen Enden oder einem ähnlichen geeigneten Werkstoff, der zwischen zwei auf der Struktur befestigten Endstücken eingesprungen wird. Jede Längenänderung im Bereich des Längungsmessers verursacht eine Änderung in der Durchbiegung des Streifens, die mittels in dessen Mittelpunkt angebrachten Dehnstreifen gemessen wird. Der Dehnstreifenausgang wird direkt als Änderung der effektiven Stahlstreifenlänge geeicht. Die Nennlänge des Längungsmessers ist 12 Zoll (30,48 cm); für spezielle Anwendungszwecke ist er jedoch in anderen Längen lieferbar.

Diese Ausführung hat Vorteile gegenüber fest angebauten Längungsmessern, da sie nur durch die Endbelastung der Kugelenden des Streifens mit der Strukturoberfläche in Berührung steht. Die Einflüsse von Enddrehung, örtlichen Verzerrungen und Schub werden dadurch auf ein Kleinstmass reduziert und nur die wahre lineare Bewegung zwischen den Endkontaktpunkten gemessen.

Die den Streifen positionierenden Endstücke können mittels Kleber, Schrauben in Gewindelöchern oder—bei Beton- oder Ziegelstrukturen—Steindöbeln befestigt werden. Für die Anzeige kann man jedes Dehnungsmessgerät oder Galvanometer verwenden, für den Aussendienst ist dagegen ein tragbarer, batteriegespeister Dehnungsanzeiger am bequemsten, wie z.B. der von Saunders-Roe erhältliche.

EE 74 778 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Entwurf einer Transistor-Widerstand-Logikschaltung mit grafischen Mitteln von C. W. M. Barrow

Zusammenfassung des Beitrages auf Seite 660-665

Eine grafische Lösung wird für den Entwurf einer NOR-Schaltung mit Transistor und Widerstand beschrieben. Die NOR-Schaltung erfordert, dass der Transistor gesperrt ist, wenn alle Eingangspegel der Eingangszuleitungen an Masse liegen, jedoch auf den unteren Grenzwert gebracht wird, wenn der Pegel einer der Eingangszuleitungen negativer wird als ein gewisser Schwellenwert. Die Berechnungen berücksichtigen den Einfluss der Grenzwerte von Widerständen, Stromversorgungen und Transistor-Parametern. Die Ergebnisse werden in der Berechnung eines mehrstufigen Zählers mit NOR-Schaltungen angewendet.

Elektronische Bestimmung des Weidenwachstums von F. J. Hyde und J. T. Lawrence

Zusammenfassung des Beitrages auf Seite 666-670

Entwurf und Konstruktion eines Messgerätes für zerstörungsfreies Messen des Graswachstums wird beschrieben. Die Kapazität eines Messkopfes wird geändert, wenn er auf Gras gelegt wird, und ruft eine Frequenzänderung im Transistor-Oszillator hervor; mit dem er verbunden ist. Die Differenz zwischen dieser Frequenz und der eines identischen 3-MHz-Festfrequenzoszillators wird in den Ausschlag eines Gleichstrom-Mikrometers umgewandelt. Es wird gezeigt, dass der Ausschlag des Messgerätes mit dem Wassergehalt des Grases unter dem Messkopf in Beziehung steht. Messungen an wachsendem Gras wurden im Labor ausgeführt, um Licht auf die Mechanik des Graswachstums zu werfen.

Ein zuverlässiger Gleichstromverstärker für Thermoelemente in Vakuummessern von U. Tallgren und U. Kracht

Zusammenfassung des Beitrages auf Seite 671-675

Eine einfache Schaltung zur Verstärkung des von einem Thermoelement in einem Vakuummesser abgegebenen Signals wurde entwickelt. Das Ausgangssignal ist gross genug zum Treiben mehrerer reihengeschalteter 1-mA-Messgeräte. Dem Verstärker liegt Verwendung eines kleinen Messwandlers als Umsetzer in die zweite Harmonische mit nachgeschaltetem wechselstromgekoppelten Verstärker zugrunde. Der Ausgang speist einen Gleichrichter durch eine transformatorgekoppelte Endstufe. Zwischen Ein- und Ausgangsschaltungen wird eine komplette galvanische Isolation erreicht. Die Linearität des Verstärkers hängt hauptsächlich von der Symmetrie des Messwandlers ab und kann besser als ±3 Prozent sein.

Für den Heizerstrom von zwei der üblichsten Vakuummeter mit Thermoelementen werden Konstanthalterschaltungen gegeben und ausserdem Schaltungen für die Betätigung eines Relais bei vorgewähltem Ausgangspegel des Thermoelementverstärkers.

Zwei Schaltungsarten werden beschrieben, und zwar eine für Netzanschluss und die andere für Betrieb mit einer 48-V-Bereitschaftsbatterie.

Die Schaltungen in diesem Beitrag dürften auch für andere Aufgaben der Instrumentierung, z.B. die Temperaturregelung mit Thermoelementen, von Interesse sein.

Grundentwurf eines SECAM-Farbfernsehers von P. Cassagne und G. Melchior

Zusammenfassung des Beitrages auf Seite 676-681

Der Prototyp eines kommerziellen SECAM-Farbfernsehers wurde entwickelt, und in diesem Beitrag werden einige der wichtigeren Konstruktionsmerkmale besprochen. Besondere Aufmerksamkeit wurde den Bedienelementen des Empfängers sowie der einfachen Bedienung gewidmet. Es wird daraus gefolgert, dass ein zufriedenstellender SECAM-Empfänger hergestellt werden kann, dessen Fertigungskosten zu denen eines vergleichbaren NTSC-Empfängers in einem günstigen Verhältnis stehen.

Messen der Dielektrizitätskonstante und Leitfähigkeit bei Temperaturen bis zu 500°C von R. H. A. Miles

Zusammenfassung des Beitrages auf Seite 682-687

Ein neu entwickeltes Gerät gibt ein halbquantitatives Mass für Änderungen der Dielektrizitätskonstante und Leitfähigkeit von Materialien über den Temperaturbereich von 20° . . . 500°C. Bei einigen Materialien, z.B. in der Ferroelektrik-Keramik, kommen grosse Änderungen dieser Parameter vor, was ein automatisches System für die Bereichumschaltung bedingt, wenn man die Ergebnisse auf einem herkömmlichen Registriergerät mit der erforderlichen Genauigkeit von ca. 10 Prozent darstellen will. Der Gesamtbereich der von dem Gerät überstrichenen Kapazität ist 1,0 pF . . . 1000 pF bei Frequenzen von 100 kHz . . . 1 MHz und der Registrierbereich für Gleichstromwiderstand 10 Ω . . . 500 MΩ. Für Prüflinge mit nennenswertem dielektrischen Verlust wird die Kapazitätsanzeige als Funktion der Kapazität sowie des Verlustwiderstandes gegeben. Trotzdem in diesem Fall auf keine absoluten Werte für die Parameter geschlossen werden kann, erhält man doch ausreichende Information, um festzustellen, ob sich Anwendung eines hochgezüchteten Messverfahrens lohnt.

Wahre Abstimmung auf die dritte Harmonische von F. D. Bate

Zusammenfassung des Beitrages auf Seite 688-691

Unter der Voraussetzung, dass die beiden vorhandenen Frequenzen f und f_3 sind, wird eine Analyse der Abstimmung auf die dritte Harmonische gegeben, die z.B. in den Hochspannungs-Stromversorgungen für den Zeilenrücklauf Anwendung findet.

Verschiedene Kurven wurden erstellt, mit Hilfe deren eine Konstruktion für optimale Abstimmung auf die dritte Harmonische entworfen werden kann.

Nichtlineare RC-Schaltungen mit Siliziumkarbid-Varistoren von W. G. P. Lamb

Zusammenfassung des Beitrages auf Seite 692-693

In diesem Beitrag wird gezeigt, dass das Einschwingverhalten nichtlinearer Widerstand-Kapazität-Reihenschaltungen für die Funktionserzeugung benutzt werden kann, wenn nichtlineare Siliziumkarbidwiderstände in Verbindung mit Funktionsverstärkern Verwendung finden.

Aktivieren einer elektromagnetischen Bremse durch die Muskeln der Augenbraue von L. Vodovnik

Zusammenfassung des Beitrages auf Seite 694-695

Nach einer Besprechung der Vorgänge, die eingeleitet werden, wenn ein Fahrer entscheidet zu bremsen, wird die Möglichkeit in Betracht gezogen, einen Teil der Reaktionszeit zu kürzen. Ein versuchsmässiges System mit einer durch die Aktionspotentiale der Augenbraumuskeln aktivierten elektromagnetischen Bremse wird beschrieben. Einige Verbesserungsmöglichkeiten und weitere Anwendungsgebiete, z.B. für Amputierte, werden vorgeschlagen.

Der Eingangswiderstand transistorisierter Gegenkopplungsverstärker von E. de Boer

Zusammenfassung des Beitrages auf Seite 702-706

Dieser Beitrag beschreibt eine allgemeine Theorie der Gegenkopplung, die sich statt auf Verstärkung mehr auf Wellenwiderstand der Schaltung konzentriert. Einfache sowie komplexe Gegenkopplungssysteme werden untersucht und bereits bei einfacher Betrachtung nützliche Einsichten gewonnen. Regeln für die Anwendung der Theorie sind einfach und logisch. Wenn Gegenkopplung einem Punkt parallel entnommen oder zugespeist wird, so wird angenommen, dass er unendlichen Innenwiderstand hat. Wo andererseits Gegenkopplung Unterbrechung einer Schleife erfordert, wird angenommen, dass die Schleife einen Innenwiderstand von Null hat. Irgendwelche interne Eigenwiderstände solcher Punkte werden in die Quelle oder Last absorbiert. Eine Anzahl erläuternde Beispiele, einschliesslich Entwurf eines Zeilentransformators für Betrieb innerhalb vorgeschriebener Wellenwiderstände, zeigen die einschlägigen Punkte der Theorie.

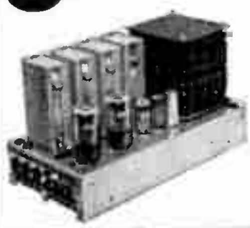
EE 74 141 *for further details*

B Series rugged reliable modules

The B series is probably the most advanced open chassis design available. Extremely tough construction inter-services approved reliable valves and silicon junction rectifiers give durability coupled with the high efficiency of series pentode circuits. All units have an Impedance of 0.1Ω and a stability of 0.02%.

Model	DC output (Floating) Volts	Current	AC outputs	Ripple mV	Size W	Inches D	H	UK price
B series rugged reliable modules								£
B101/100	100	30-150mA	2×6.3V3A CT	1.0	7½	9¾	6½	55
B101/150	150	30-150mA	2×6.3V3A CT	1.0	7½	9¾	6½	55
B101/200	200	0-150mA	2×6.3V3A CT	1.0	7½	9¾	6½	55
B101/250	250	0-150mA	2×6.3V3A CT	1.0	7½	9¾	6½	55
B101/300	300	0-150mA	2×6.3V3A CT	1.0	7½	9¾	6½	55
B101/350	350	0-100mA	2×6.3V3A CT	1.0	7½	9¾	6½	55
•B101/400/450/500	•	0-150mA	2×6.3V3A CT	1.0	7¾	10¾	7¾	75
•B101/550/600/650/700/750	•	0-150mA	2×6.3V3A CT	2.0	7¾	11½	8¼	95
•B101/800/850/900/950/1000	•	0-150mA	2×6.3V3A CT	3.0	7¾	14½	8¾	120
B102/100	100	30-300mA	2×6.3V5A CT	1.0	7½	14¾	6½	75
B102/150	150	30-300mA	2×6.3V5A CT	1.0	7½	14¾	6½	75
B102/200	200	0-300mA	2×6.3V5A CT	1.0	7½	14¾	6½	75
B102/250	250	0-250mA	2×6.3V5A CT	1.0	7½	14¾	6½	75
B102/300	300	0-250mA	2×6.3V5A CT	1.0	7½	14¾	6½	75
B102/350	350	0-200mA	2×6.3V5A CT	1.0	7½	14¾	6½	75
•B102/400/450/500	•	0-300mA	2×6.3V5A CT	1.0	7¾	14½	8¾	95
•B102/550/600/650/700/750	•	0-300mA	2×6.3V5A CT	2.0	8	16¾	9½	120
•B102/800/850/900/950/1000	•	0-300mA	2×6.3V5A CT	3.0	8¾	17¾	9½	140
B103/100	100	30-500mA	2×6.3V5A CT	1.0	8	16¾	6½	95
B103/150	150	30-500mA	2×6.3V5A CT	1.0	8	16¾	6½	95
B103/200	200	0-500mA	2×6.3V5A CT	1,0	8	16¾	6½	95
B103/250	250	0-500mA	2×6.3V5A CT	1,0	8	16¾	6½	95
B103/300	300	0-500mA	2×6.3V5A CT	1.0	8	16¾	6½	95
B103/350	350	0-400mA	2×6.3V5A CT	1.0	8	16¾	6½	95
•B103/400/450/500	•	0-500mA	2×6.3V5A CT	1.0	8¾	17¾	9½	120
•B103/550/600 650/700/750	•	0-500mA	2×6.3V5A CT	2.0	11¾	17¾	10	150
•B103/800/850/900/950/1000	•	0-500mA	2×6.3V5A CT	3.0	13	17¾	10	175
B104/100	100	100mA-1A	2×6.3V5A CT	1.0	11	17¾	8½	160
B104/150	150	100mA-1A	2×6.3V5A CT	1.0	11	17¾	8½	160
B104/200	200	0-1A	2×6.3V5A CT	1.0	11	17¾	8½	160
B104/250	250	0-1A	2×6.3V5A CT	1 0	11	17¾	8½	160
B104/300	300	0-1A	2×6.3V5A CT	1.0	11	17¾	8½	160
B104 350	350	0-800mA	2×6.3V5A CT	1.0	11	17¾	8½	160

S Series single an d twin regulated supplies



The S series embodies the same advanced design as the B series but originally included only twin units. Owing to their popularity they are now also offered on their own for single output voltages.

Model	DC output (Floating) Volts	Current	AC outputs	Ripple mV	Size W	Inches D	H	UK Price
S series open chassis single output units								£
S101/100	100	30-150mA	2×6.3V3A CT	1.0	6¾	10¾	7 3/16	48
S101/150	150	30-150mA	2×6.3V3A CT	1.0	6¾	10¾	7 3/16	48
S101/200	200	0-150mA	2×6.3V3A CT	1.0	6¾	10¾	7 3/16	48
S101/250	250	0-150mA	2×6.3V3A CT	1.0	6¾	10¾	3/16	48
S101/300	300	0-150mA	2×6.3V3A CT	1.0	6¾	10¾	7 3/16	48
S101/350	350	0-100mA	2×6.3V3A CT	1.0	6¾	10¾	7 3/16	48
S102/100	100	30-300mA	2×6.3V5A CT	1.0	6¾	14½	7 3/16	66
S102/150	150	30-300mA	2×6.3V5A CT	1.0	6¾	14½	7 3/16	66
S102/200	200	0-300mA	2×6.3V5A CT	1.0	6¾	14½	7 3/16	66
S102/250	250	0-250mA	2×6.3V5A CT	1.0	6¾	14½	7 3/16	66
S102/300	300	0-250mA	2×6.3V5A CT	1.0	6¾	14½	7 3/16	66
S102/350	350	0-200mA	2×6.3V5A CT	1.0	6¾	14½	7 3/16	66
S103/100	100	30-500mA	2×6.3V5A CT	1.0	6¾	17¾	8¾	88
S103/150	150	30-500mA	2×6.3V5A CT	1.0	6¾	17¾	8¾	88
S103/200	200	0-500mA	2×6.3V5A CT	1.0	6¾	17¾	8¾	88
S103/250	250	0-500mA	2×6 3V5A CT	1.0	6¾	17¾	8¾	88
S103/300	300	0-500mA	2×6.3V5A CT	1.0	6¾	17¾	8¾	88
S103/350	350	0-400mA	2×6.3V5A CT	1.0	6¾	17¾	8¾	88

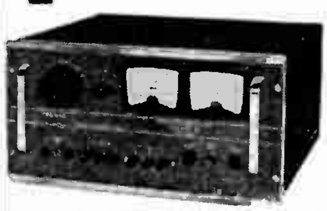
Model		Size W	Inches D	H	UK Price
S Series front panel twin floating output units					£
2 S101/volts S101/volts	Electrically identical to their single voltage counterparts.	19	10¾	7	136
2 S102/volts S102/volts	When ordering please state voltages required.	19	14½	7	172
2 S101/volts S103/volts	Thus, a 300V, 500mA and 250V, 150mA twin unit would be	19	17¾	7	216
2 S101/volts S102/volts	2 S103/300 S101/250.	19	14½	7	154
2 S102/volts S103/volts	There is no need to state the required polarity as both sections are floating and	19	17¾	7	194
2 S103/volts S101/volts	completely, independent from each other	19	17¾	7	176

M Series regulated supplies

This series employs a chokeless design offering small size, excellent high frequency response (with stable operation down to at least 1 cps), reliability at high temperatures and safe operation with mains voltage changes up to ±10%. Stability 0.02%, Resistance 0.1Ω, Ripple Voltage 1.0 mV.

Model	DC output (Floating) Volts	Current	AC outputs	Rip ple mV	Size W	Inches D	H	UK Price
M series open chassis units								£
M22	150	0- 20mA	nil	1.0	4¾	3¾	4½	26
M22A	200	0- 20mA	nil	1.0	4¾	3¾	4½	26
M23	150	0- 50mA	6.3V2A & 6.3V1A	1.0	6	5½	5¾	27.10
M23A	200	0- 50mA	6.3V2A & 6.3V1A	1.0	6	5½	5¾	27.10
M33	250	0- 50mA	6.3V2A & 6.3V1A	1.0	6	5½	5¾	27.10
M33A	300	0- 50mA	6.3V2A & 6.3V1A	1.0	6	5½	5¾	27.10
M14A	100	0-100mA	2×6.3V3A CT	1.0	10¾	7 3/16	6½	63
M24	150	0-100mA	2×6.3V3A CT	1.0	6¾	7¾	6½	33.10
M24A	200	0-100mA	2×6.3V3A CT	1.0	6¾	7¾	6½	33.10
M34	250	0-100mA	2×6.3V3A CT	1.0	6¾	7¾	6½	33.10
M34A	300	0-100mA	2×6.3V3A CT	1.0	6¾	7¾	6½	33.10
M25	150	0-200mA	2×6.3V3A CT	1.0	8½	8¾	7	52.10
M25A	200	0-200mA	2×6.3V3A CT	1 0	8½	8¾	7	52.10
M35	250	0-200mA	2×6.3V3A CT	1.0	8½	8¾	7	52.10
M35A	300	0-200mA	2×6.3V3A CT	1.0	8½	8¾	7	52.10
M26S	350	30-500mA	2×6.3V5A CT	1.0	14 3/16	11½	8 3/16	88
M26SA	200	0-500mA	2×6.3V5A CT	1.0	14 3/16	11½	8 3/16	88
M365	250	0-500mA	2×6.3V5A CT	1.0	14 3/16	11½	8 3/16	88
M365A	300	0-500mA	2×6.3V5A CT	1 0	14 3/16	11½	8 3/16	88
M28	150	30-350mA	2×6.3V5A CT	1.0	13 3/16	8¾	7½	66
M28A	200	0-350mA	2×6.3V5A CT	1.0	13 3/16	8¾	7½	66
M38	250	0-350mA	2×6.3V5A CT	1.0	13 3/16	8¾	7½	66
M38A	300	0-350mA	2×6.3V5A CT	1.0	13 3/16	8¾	7½	66
M series front panel units								
M26	150	30-500mA	2×6.3V5A CT	1,0	19	12½	8¾	97.10
M26A	200	0-500mA	2×6.3V5A CT	1.0	19	12½	8¾	97.10
M36	250	0-500mA	2×6.3V5A CT	1.0	19	12½	8¾	97.10
M36A	300	0-500mA	2×6.3V5A CT	1.0	19	12½	8¾	97.10
M27	150	50mA-1A	2×6.3V5A CT	1.0	19	16½	8¾	165
M27A	200	0-1A	2×6.3V5A CT	1.0	19	16½	8¾	165
M37	250	0-1A	2×6.3V5A CT	1.0	19	16½	8¾	165
M37A	300	0-1A	2×6.3V5A CT	1.0	19	16½	8¾	165
M29	150	100mA-2A	2×6.3V5A CT	1.0	19	18½	10½	255
M29A	200	0-2A	2×6.3V5A CT	1 0	19	18½	10½	255
M39	250	0-2A	2×6.3V5A CT	1.0	19	18½	10½	255
M39A	300	0-2A	2×6.3V5A CT	1.0	19	18½	10½	255

V Series variable voltage regulated supplies



The V series of laboratory units offers several exceptionally stable and one moderate specification design.

Model	Main DC output (Floating) Volts	Current	Negative Rail Volts	Current	Size Inches W D H	UK Price
V series variable voltage regulated supplies						£
V50-20D	0 to 500	0-200mA (restricted to 150mA below 200V)	0 to 85 via 100k pot.		19 10 7 Panel height 7"	79
V50-20R	200 to 500	0-200mA	nil		19 10 7 Panel height 7"	79
V50-50	0 to 500	0-500mA	250	0-50mA	19¾ 14½ 12 Panelheight 10½"	155
2V35-25	(a) 0 to 350	0-250mA (restricted to 150mA below 200V)	250	0-40mA	19¾ 16¾ 10½ Panel height B¾"	
	(b) 0 to 350	0-250mA (restricted to 150mA below 200V)				198
V100-25	200 to 1000	0-250mA	nil		19¾ 18½ 10½ Panel height 8¾"	170
V200-12	200 to 2000	0-125mA	nil		19¾ 21 15½ Panel height 14"	265

roband

POWER

SUPPLIES

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Charlwood Works
Charlwood . Horley
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Telephone:

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ROBAND ELECTRONICS LIMITED
Charlwood Works,
Lowfield Heath Road,
Charlwood, Horley, Surrey.
Crawley 20172.

SILICON

Variable voltage single and twin output instruments

These all-silicon units have much reduced size and offer complete electronic protection for any output current down to one tenth of the full rated value. 300V units are fully protected by fuses and series resistors.

Output voltage is smoothly controlled from zero to maximum in one continuous sweep and a fine control is included. All units above 1A have four terminal systems, the twin voltage units having floating, independently controlled outputs.

Model	DC output (Floating) Amps	Volts	Ripple mV R.M.S.	Resistance Ohms	Size Inches W	D	H	UK Price
T series variable voltage single output instruments								£
T244	0.33	10-300	2.0	0.5	12¼	10	6¼	125
T245	0.5	10-200	2.0	0 5	12¼	10	6¼	125
T246	1	0-100	2.0	0.2	12¼	10	6¼	125
T247	2	0-50	1.0	0.05	12¼	10	6¼	125
T248	3	0-30	1.0	0.03	12¼	10	6¼	125
T229	0.5	10-300	2.0	0.5	19¼	10	6¼	145
T230	1	10-150	2.0	0.2	19¼	10	6¼	145
T236	2	0-75	2.0	0.08	19¼	10	6¼	145
T237	3	0-50	2.0	0.03	19¼	10	6¼	145
T238	5	0-30	1.0	0.02	19¼	10¼	6¼	145
T231	1	10-300	2.0	0.3	19¼	12¼	6¼	180
T232	2	10-150	2.0	0.2	19¼	12¼	6¼	180
T233	3	0-100	2.0	0.1	19¼	12¼	6¼	180
T234*	5	0-60	2.0	0.02	19¼	12¼	6¼	180
T235*	10	0-30	1.0	0.01	19¼	12¼	6¼	180
T239	1.5	10-300	2.0	0.2	19¼	14	8¼	265
T240	3	10-150	2.0	0.1	19¼	14	8¼	265
T241*	5	0-100	2.0	0.02	19¼	14	8¼	265
T242*	10	0-50	2.0	0.01	19¼	14	8¼	265
T243*	15	0-30	2.0	0.007	19¼	14	8¼	265
T series variable voltage twin output instruments								
T249*	0.5/0.5	10-300/10-300	2.0	0.5	19¼	15¼	10¼	255
T250*	1/1	10-150/10-150	2.0	0.2	19¼	15¼	10¼	255
T251*	2/2	0-75/0-75	2.0	0.08	19¼	15¼	10¼	255
T252*	3/3	0-50/0-50	2.0	0.03	19¼	15¼	10¼	255
T253*	5/5	0-30/0-30	1.0	0.02	19¼	15¼	10¼	255
T254*	1/1	10-300/10-300	2.0	0.3	19¼	17	12	325
T255*	2/2	10-150/10-150	2.0	0.2	19¼	17	12	325
T256*	3/3	0-100/0-100	2.0	0.1	19¼	17	12	325
T257*	5/5	0-60/0-60	2.0	0 02	19¼	17	12	325
T258*	10/10	0-30/0-30	1.0	0.01	19¼	17	12	325

*Normally supplied for 200-250V mains or for 100-125V on request. All other units operate from both mains groups

Single preset voltage, open chassis units

Operating to at least 55°C ambient these units are reliably protected against overloads and short circuits, particularly where the output exceeds 30V. They can be preset to any voltage within their range, with other outputs available by adjustment within the unit. The T180 will operate in 65°C ambient and is therefore suitable to operate in close proximity to vacuum tube equipment. It also has two a.c. outputs, each 6.3V 5A.

Model	DC output (Floating) Amps	Volts	Effective Resistance ohms	Size Inches W	H	D	UK Price
T series open chassis single floating output units							£
T295	0.5	6-30	0.08	2¾	5¼	4 7/16	39
T296	0.5	6-50	0 1	4¼	7¼	5¼	52
T297	0.5	6-100	0.15	5¼	8¼	6¼	79
T298	1	6-15	0.04	2¾	6¼	5¼	47
T299	1	6-30	0.05	4¼	7¼	5¼	49
T300	1	6-50	0.08	5¼	8¼	6¼	65
T301	1	6-100	0.1	6	9	6	89
T302	3	6-15	0.04	6	9	6	74
T303	3	6-30	0.04	6	9	6	78
T190	3	6-50	0.05	6¼	11	7	85
T194	3	6-100	0.07	7	13	8	128
T304	5	6-15	0.02	6	13	6¼	97
T305	5	6-30	0 02	6	13	6¼	104
T191	5	6-50	0.03	6¼	11	7¼	115
T195	5	6-100	0.04	8¼	13¼	8¼	158
T306	10	6-15	0.01	6¼	17	7	137
T307	10	6-30	0.01	7	17	8	147
T192	10	6-50	0.01	8¼	13¼	8¼	185
T196	10	6-100	0.02	10¼	15¼	11	256
T180	0.5	100-300	0.1	4¼	11¼	6¼	95

Twin preset voltage, open chassis units

Operating to at least 55°C ambient, these units are reliably protected against overloads and short circuits, particularly where the output exceeds 30V. Complete stability is maintained even when mains are ±10% nominal.

Because these units have two independently controlled floating outputs there is no need to specify polarity when ordering. All sections rated above 1A have four

Model	DC output (Floating) Amps	Volts	Effective Resistance ohms	Size Inches W	D	H	UK Price
T series open chassis twin floating output units							£
T308	0.5/0.5	6-30/6-30	0.08/0.08	6	9	6¼	74
T309	0.5/0.5	6-50/6-50	0.1/0.1	6	9	6¼	99
T310	0.5/0.5	6-100/6-100	0.15/0.15	6¼	11	7	148
T311	1/1	6-15/6-15	0.04/0.04	6	9	6¼	88
T312	1/1	6-30/6-30	0.05/0.05	6	9	6¼	93
T313	1/1	6-50/6-50	0 08/0.08	6¼	11	7	116
T225	1/1	6-100/6-100	0.1/0.1	6¼	14¼	7¼	166
T314	3/3	6-15/6-15	0.04/0.04	6¼	11	7	140
T315	3/3	6-30/6-30	0.04/0.04	6¼	14¼	7¼	146
T221	3/3	6-50/6-50	0.05/0.05	7	13	8	154
T226	3/3	6-100/6-100	0.07/0.07	8¼	15	8¼	234
T316	5/5	6-15/6-15	0.02/0.02	6¼	14¼	7¼	182
T317	5/5	6-30/6-30	0.02/0.02	7	17	8	194
T222	5/5	6-50/6-50	0.03/0.03	8¼	15	8¼	208
T227	5/5	6-100/6-100	0.04/0.04	10¼	17	11	298
T318	10/10	6-15/6-15	0.01/0.01	9¼	17	8¼	252
T319	10/10	6-30/6-30	0.01/0.01	10¼	17	9¼	268
T223	10/10	6-50/6-50	0.01/0.01	10¼	17	11	345

EE 74 143 *for further details*

T SERIES POWER SUPPLIES

GERMANIUM

Single preset voltage, open chassis units

These units will operate at full load, at 45°C ambient, even when mains are permanently ±10% nominal. They can be preset to any voltage within their range with other outputs available by adjustment within the unit. All units have safeguards against overload and short circuits, those above 1A have four terminal outputs to eliminate lead resistance.

Model	DC output (Floating) Amps	Volts	Effective Resistance ohms	Size Inches W	D	H	UK Price
T series open chassis single floating output units							£
T201	0.5	0.06	0.04	2⅞	5½	4 7/16	32
T98	0.5	6-30	0.08	2⅞	5½	4 7/16	32
T202	0.5	6-50	0.1	4½	7¼	5¾	40
T149	0.5	0-50	0.1	4½	7¼	5¾	45
T203	0 5	6-100	0.15	5½	8½	6¾	62
T197	1	0-6	0.04	2⅞	5½	4 7/16	35
T165	1	6-15	0.04	2⅞	6½	5¾	38
T181	1	6-30	0.05	4½	7¼	5¾	40
T100	1	0-30	0.05	4½	7¼	5¾	45
T189	1	6-50	0.1	5½	8½	6¾	49
T108	1	0-50	0.1	5½	8½	6¾	55
T193	1	6-100	0.1	6	9	6	76
T169	2	0-15	0.04	4½	7¼	5¾	47
T170	2	0-30	0.05	5½	8½	6¾	55
T109	2	0-50	0.05	5½	10½	7½	70
T198	3	0-6	0.04	4½	7¼	5¾	54
T166	3	6-15	0.04	6	9	6	62
T182	3	6-30	0.04	6	9	6	65
T137	3	0-30	0.04	5½	10½	7½	75
T161	3	0-50	0.05	7	12	6¾	93
T162	4	0-40	0.04	7	12	6¾	93
T199	5	0-6	0.02	5½	8½	6¾	69
T167	5	6-15	0 02	6	13	6½	79
T183	5	6-30	0.02	6	13	6½	84
T110	5	0-30	0.02	8½	13½	7½	98
T141	6	0-50	0.03	9	15	8½	135
T163	7.5	0-40	0.02	9	15	8½	135
T200	10	0-6	0.01	6	13	6½	98
T168	10	6-15	0 01	6½	17	7	115
T184	10	6-30	0.01	7	17	8	120
T114	10	0-30	0.01	10	17	8½	135
T146	12	0-50	0.01	12	17½	9½	180
T164	15	0-30	0.01	12	17½	9½	180
T142	20	0-30	0.005	16½	17½	12	245
T135	25	0-20	0.005	16½	17½	12	245
T143	50	0-20	0.002	33	10	22	475
T157	50	0-20	0.002	17½	20	22	475
T158	100	0-20	0.001	17½	20	22	665

Twin preset voltage, open chassis units

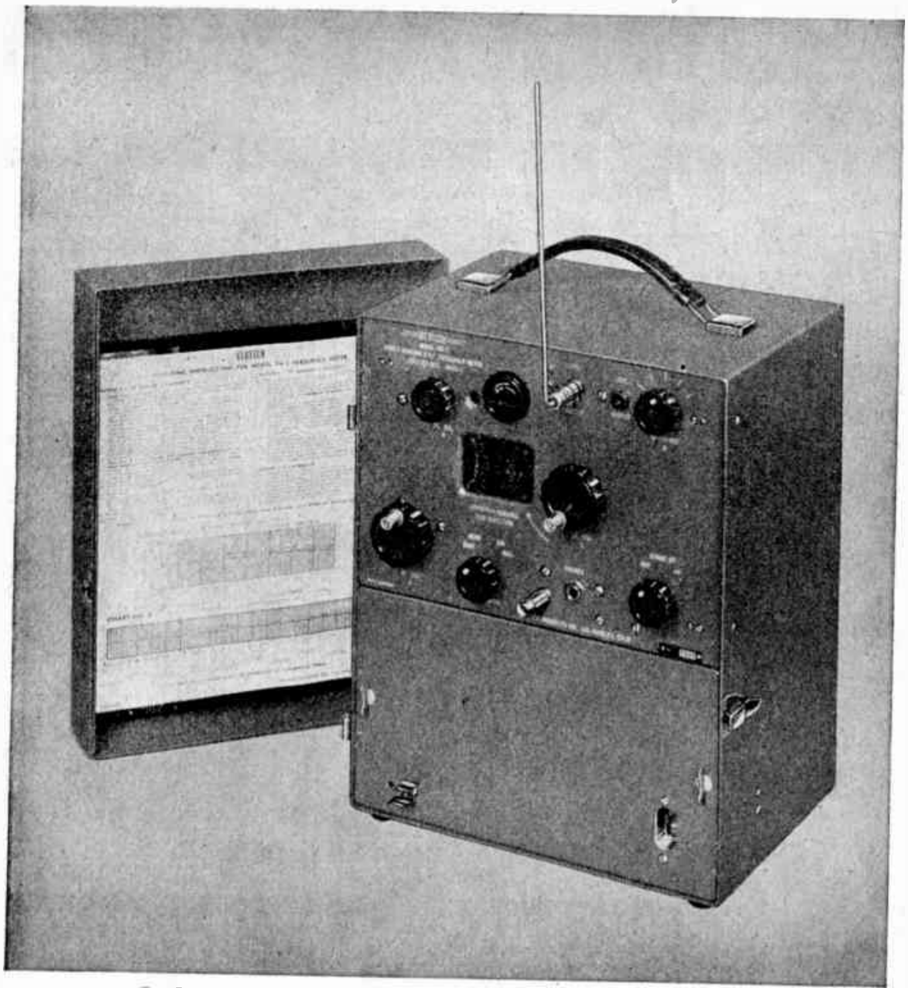
Because these units have two independently controlled floating outputs there is no need to specify polarity when ordering. Thus, for example, a requirement for −18V 3A and +14V 3A would be met by the unit T116/18/15. All sections rated above 1A have four terminal networks to eliminate lead resistance. All units have safeguards against overload and short circuits.

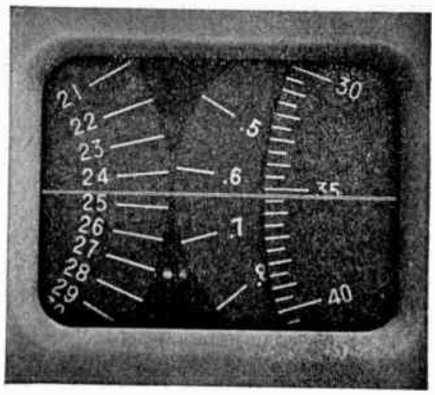
Model	DC output (Floating) Amps	Volts	Effective Resistance ohms	Size Inches W	D	H	UK Price
T series open chassis twin floating output units							£
T204	0.5/0.5	0-6/0-6	0.04/0.04	4½	7½	5¾	62
T209	0 5/0 5	6-15/6-15	0.06/0.06	6	9	6½	62
T214	0.5/0.5	6-30/6-30	0.08/0.08	6	9	6½	62
T219	0.5/0.5	6-50/6-50	0.1/0.1	6	9	6½	76
T224	0 5/0.5	6-100/6-100	0 15/0.15	6½	11	7	118
T205	1/1	0-6 0-6	0.04/0.04	4½	7½	5¾	65
T210	1/1	6-15/6-15	0.04/0.04	6	9	6½	69
T115	1/1	0-20/0-20	0.04/0.04	6	10½	6	87.10
T185	1/1	6-30/6-30	0.05/0.05	6	9	6½	72
T220	1/1	6-50/6-50	0.1/0.1	6½	11	7	88
T151	1/1	0-50/0-50	0.1/0.1	8½	12½	7	120
T206	3/3	0-6/0-6	0.04/0.04	6	9	6½	102
T211	3/3	6-15/6-15	0.04/0.04	6½	11	7	114
T116	3/3	0-20/0-20	0.03/0.03	8½	13¾	7¾	130
T186	3/3	6-30/6-30	0.04/0.04	6¾	14½	7½	122
T152	3/3	0-50/0-50	0.05/0.05	10	16½	9	178
T207	5/5	0.6/0.6	0.02/0.02	6½	11	7	126
T212	5/5	6-15/6-15	0.02/0.02	6¾	14½	7½	144
T117	5/5	0-20/0-20	0.02/0.02	10	17½	8½	176
T187	5/5	6-30/6-30	0.02/0 02	7	17	8	158
T153	5/5	0-50/0-50	0.03/0.03	14	17½	9½	245
T208	10/10	0-6/0-6	0.01/0.01	8½	17	8½	182
T213	10/10	6-15/6-15	0.01/0.01	9½	17	8½	210
T118	10/10	0-20/0-20	0.01/0.01	14	17½	9½	245
T188	10/10	6-30/6-30	0.01/0.01	10¾	17	9½	218
T144	20/20	0 30 0-30	0 003/0.003	33	10	22	475
T159	20/20	0-30/0-30	0 003/0.003	17½	20	22	475
T145	25/25	0-20/0-20	0.003/0.003	33	10	22	475
T160	25/25	0-20/0-20	0.003/0.003	17½	20	22	475

Variable voltage single and twin output instruments

These precision instruments are available in either single or twin outputs, the latter having floating independently controlled outputs. Having 19 inch front panels, these units can also be supplied without the casing for rack mounting. Adequate safeguards against overloads and short circuits are built in.

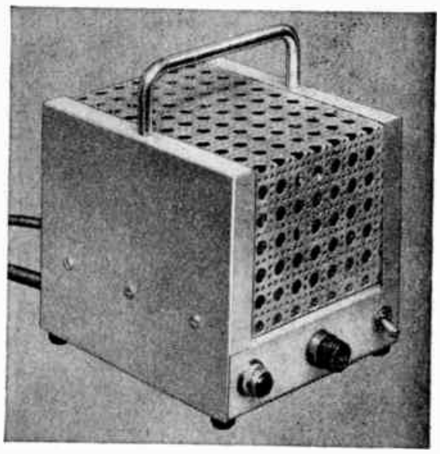
Model	DC output (Floating) Amps	Volts	Ripple mV	Effective Resistance ohms	Size Inches W	D	H	19" front panel height	UK Price
T series variable voltage single output instruments									£
T99	0.5	0-30	1.0	0.1	9¾	5½	5	—	64.10
T155	0.5	0-50	1.0	0.2	11½	8½	6	—	83
T103	1	0-30	1.0	0.05	13½	10½	9¾	—	85
T101	1	0-50	1.0	0.1	13½	10½	9¾	—	93
T106	1	0-100	2.0	0.1	19¾	14½	10½	8¾	149
T102	2	0-30	1.0	0.05	16¾	10½	10¾	—	100
T105	2	0-50	1.0	0 05	16¾	10½	10¾	—	118
T107	2	0-100	2.0	0.1	19¾	14½	10½	8¾	183
T171	3	0-30	1.0	0 03	16¾	10½	10¾	—	118
T111	5	0-50	2.0	0.01	19¾	18½	10½	8¾	155
T172	10	0-30	2.0	0 01	19¾	18½	10½	8¾	168
T112	10	0-50	2.0	0.01	19¾	20½	12	10½	245
T173	15	0-30	2.0	0.01	19¾	20½	12	—	265
T113	20	0-30	2 0	0.01	19¾	21	15¾	14	315
T320	30	0-30	3.0	0 005	19¾	21	19	17½	385
T321	50	0-30	5.0	0.002	21½	25½	38½	31½	595
T322	100	0-30	6.0	0.001	21½	25½	38½	31½	825
T series variable voltage twin output instruments									
T104	1/1	0-30/0-30	1.0	0.05	16¾	10½	10¾	—	129
T134	1/1	0-50/0-50	1.0	0.1	19¾	14½	10½	8¾	166
T174	2/2	0-30/0-30	1.0	0.05	16¾	10½	10¾	—	143
T175	3/3	0-30/0-30	1.0	0.04	19¾	14½	10½	8¾	166
T136	3/3	0-50/0-50	1.0	0.05	19¾	18½	10½	8¾	220
T176	5/5	0-30/0 30	1.0	0.01	19¾	18½	10½	8¾	220
T139	5/5	0-50/0-50	1.0	0 01	19¾	20½	12	10½	275
T177	10/10	0-30/0-30	2.0	0.01	19¾	20½	12	10½	275





Frequency is read directly from the dials. Illustration shows 24.6353 Mc/s.

A regulated power supply can be used instead of batteries, housed separately (as illustrated) or within the FM-3.



20–1,000 Megacycles within 0.001%

FM-3 GENERATES AND MEASURES WITHOUT CALIBRATION CHARTS

DIRECT READING

The FM-3 VHF frequency meter has a unique frequency adjustment system providing readings to six significant figures. The display embodies three directly-calibrated dials covering the full fundamental frequency range of 20 to 40 Mc/s. The precision mechanism associated with the dial display, in conjunction with the advanced circuit techniques, provides a resetability of 5 parts in 10^6.

WIDE RANGE

The instrument can be used for measuring and generating frequencies between 20 and 1,000 Mc/s. Frequencies below 20 Mc/s can be measured by using a harmonic of the external signal. Frequencies above 40 Mc/s and up to at least 1,000 Mc/s can be measured or generated by using harmonics present in the FM-3 signal. Facilities are provided for internal amplitude modulation of the signal at 1,000 c/s to a depth of approximately 30%.

HIGH STABILITY

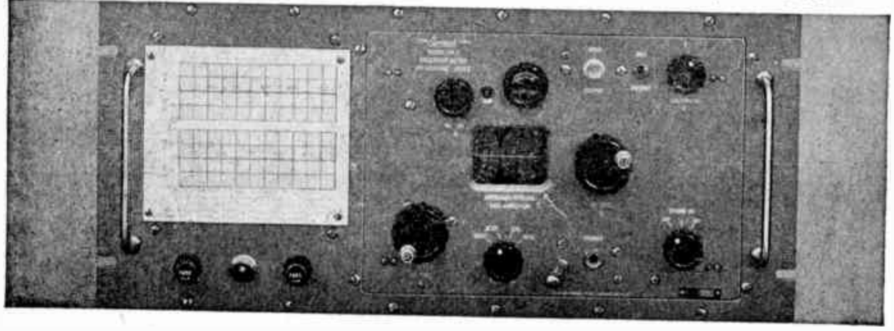
Stability and accuracy of the output frequency are within $\pm$ 10 parts per million over the full range of the instrument. This unusually high accuracy is achieved by means of a patented Multiple Oscillator Principle and is maintained without corrections from 20 to 30°C. The accuracy is maintained to within $\pm$ 0.0025% even when the ambient temperature variations extend from —15 to +55°C. The FM-3 has a built-in thermometer and individually drawn temperature correction curves are supplied with the instrument.

OUTPUTS AVAILABLE

A screened socket serves a dual purpose as the output from the generator or the input for measurement purposes. A rod aerial is provided and, for transportation, this is housed in the lower compartment. A separate socket provides precision frequency markers at 1 Mc/s intervals throughout the range 1 to 20 Mc/s and a continuously variable output from 1 to 2 Mc/s is available at the auxiliary socket.

ALTERNATIVE VERSIONS

The FM-3 has a separate compartment in the base for housing batteries. These can be removed and a regulated power supply unit (PS-3) substituted, or the power supply can be used as a separately housed unit (PS-3C). The frequency meter FM-3 is provided with a detachable lid protecting the front panel. A rack mounting version (illustrated below) is constructed on a standard 19″ x 8¾″ panel. A metal cabinet can be supplied as an optional extra for the instrument in this form.



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The Wayne Kerr Laboratories Limited Coombe Road, New Malden, Surrey. Telephone: MALden 2202.

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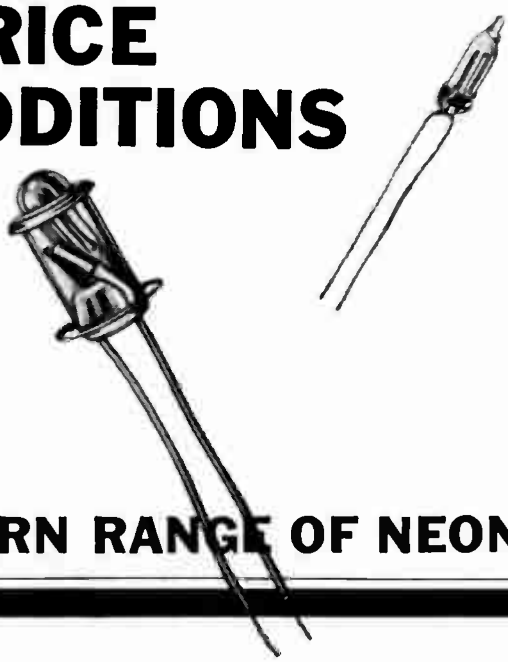

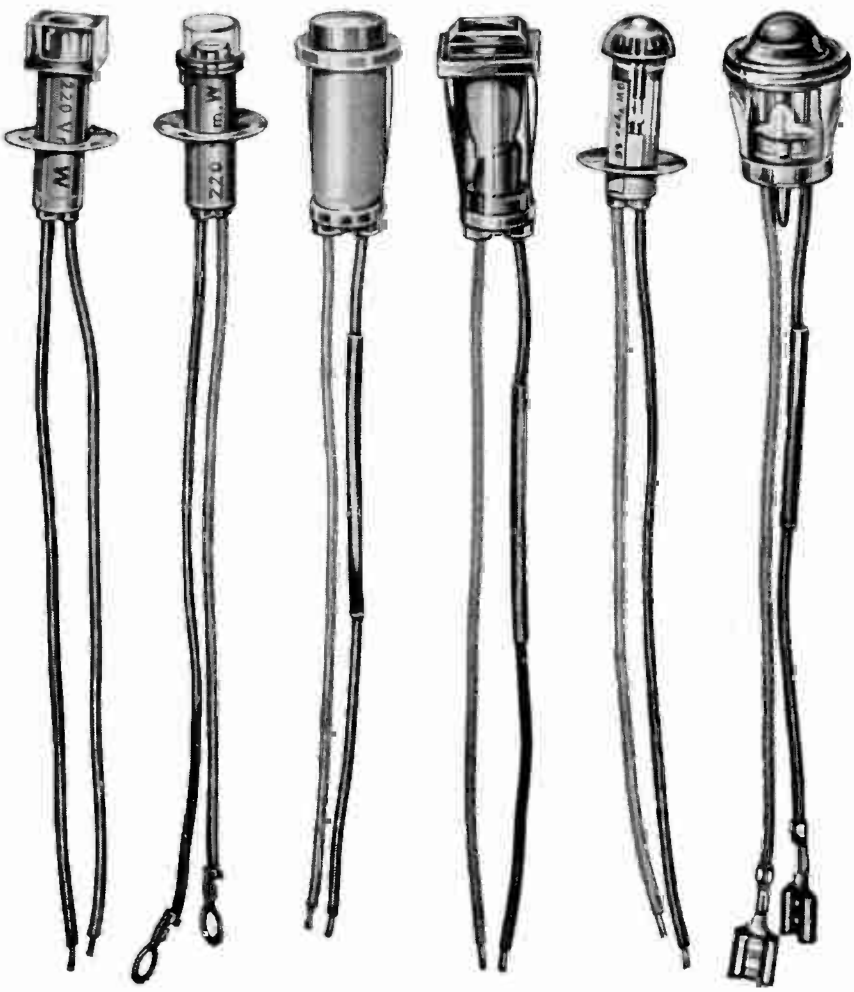

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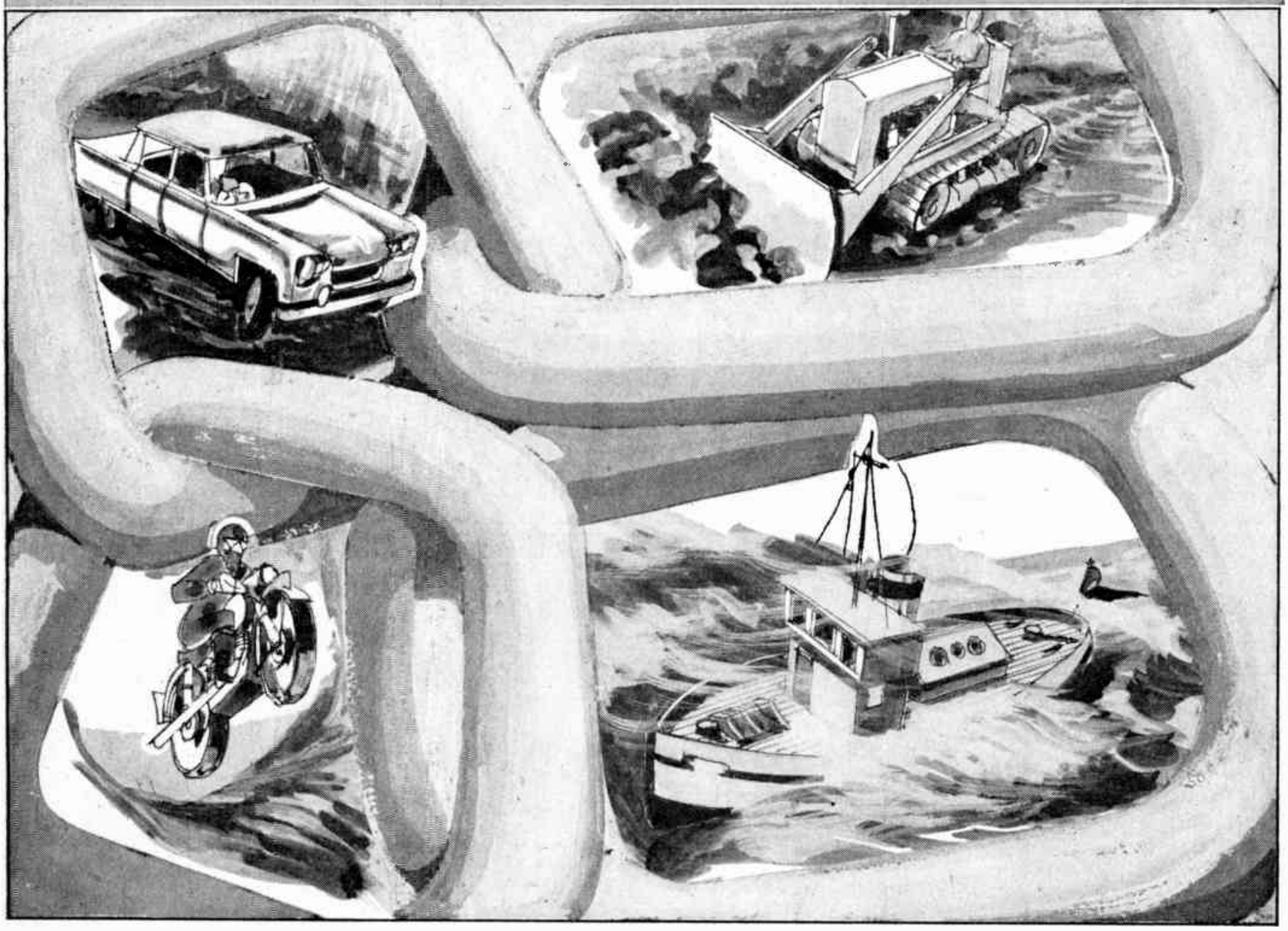
COMMUNICATIONS LINK



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Special Features:

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