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Commentary

A NOTEWORTHY feature of the Society of British Aircraft Constructors Flying Display and Exhibition held at Farnborough recently is the increasing evidence of the role played by the electronic industry of this country. It is now accepted as a truism that no aircraft can be designed and tested, much less flown and controlled without invoking the aid of electronics.

We have become familiar with the vast array of computers of all kinds which are nowadays the essential tools in the design stage of a modern aircraft. To resolve the analytical problems involved in the design it has become necessary to take measurements running into hundreds of thousands and their correlation and interpretation would be impossible without the computer. It has been stated for example, that in tests lasting for one hour on a large modern bomber, the data recorded may run into seven million separate factors which may require at least three months to reduce them to a useable form.

Once designed and put into operation, the modern aircraft must depend on the many navigational devices which have been developed by the British electronic industry and which are among the best in the world.

Mention should be made of the Doppler navigational system which, being completely airborne, renders it completely independent of ground based aids.

It has now been extensively tested under the most adverse conditions over a total distance of 1 500 million miles—equivalent to 6 000 times the distance between the earth and the moon—and it has proved itself to be a device of great accuracy and reliability. With it an aircraft can now fly in any part of the world and be instantly presented with data on its speed and position and so on.

However, one of the most outstanding problems of the moment has been the lack of a suitable aid to enable an aircraft to land at its intended destination irrespective of weather conditions. Despite advances in the military field, the absence of a safe and reliable landing system which can bring an aircraft down in "zero zero" visibility has been one of the most outstanding problems confronting civil airlines today.

Landing aids such as the Instrument Landing System, in conjunction with an automatic pilot system, allow an approach to be made by flying down a radio beam and descending with safety through cloud to a height of some 150ft above ground. At this height, the approach lighting system and other visual aids are usually sufficient for the pilot to take over the final landing.

Under conditions of zero zero visibility, this last stage has always been fraught with danger for although this i.l.s. system is accurate enough for approach purposes, the tolerance allowed by the International Civil Aviation Organization specification renders it unsuitable for the whole landing procedure.

Considerable interest therefore has been aroused in a fully automatic landing system developed by the Royal Aircraft Establishments' Blind Landing Experimental Unit at Bedford and demonstrated for the first time last October.

It is designed to eliminate the final manoeuvre so that the whole procedure of approach and landing can be carried out entirely automatically and safely without intervention on the part of the pilot, and under a wide variety of wind and visibility conditions.

"Autoland" as it has been named is relatively simple and makes no great demands in the way of bulk and complexity of apparatus. Assuming that the aircraft is already fitted with an autopilot, the only additional airborne items required are a Sequence Indicator, Leader Cable Receiver, Radio Altimeter and an Automatic Speed Control Unit—in all an additional weight of 100lb.

The ground-based installation requires no costly reconstruction of the airport and consists merely of cables laid on either side of the runway and remote control equipment which is small enough to be housed in an existing building.

With the Autoland system an aircraft now makes its approach down to a height of some 300ft using its autopilot and the existing i.l.s. system. At this height, the aircraft enters the fields of the Leader Cables and the localizer signals derived from the i.l.s. and which have been fed into the autopilot are automatically replaced by signals from the Leader Cables which are then used for azimuth guidance for the remainder of the landing.

At about 100ft, a height determined by the Radio Altimeter, the i.l.s. Glide Path Signal is switched out and the aircraft pitch attitude is maintained constant at a datum established during the approach.

The next phase occurs at a height of about 40ft when a flare out signal derived from the Radio Altimeter is fed into the autopilot and is maintained until touchdown, and at some 20ft above ground the final switch sequence takes place. This is designed to remove any drift due to crosswind. The wings are levelled and the heading is set so that the touchdown is made with the aircraft close to and in line with the centre of the runway.

This 'Autoland' system, a description of which is given elsewhere in this issue, has now reached the stage of development where a 'no hands' landing is more than a possibility. The Air Registration Board, which is responsible for the safety of civil flying in Britain is at present evaluating the system for application to civil air lines and the NATO air authorities are making a similar appraisal for military purposes.

A DIGITAL REMOTE POSITION CONTROL

By K. G. Hilton*

This article describes an investigation into the problems involved in constructing a digital remote position control. Shaft position is measured digitally and the servomotor control circuits operate digitally. The article concerns itself with the logical arrangements of the circuits and methods of stabilizing a digital remote position control are discussed.

A coarse system is described which is accurate only to 45° since this was considered to be sufficient to show the principles of the system and the problems involved in designing it.

(Voir page 570 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 575)

THE purpose of the investigation was to consider the problems involved in constructing a digital remote position control. Digital methods and transistor circuits were to be used throughout the device.

The potential advantage which a digital system has over an analogue system is that its accuracy increases with the number of digits being used. For example, a good analogue system can have an accuracy of about 0·1 per cent. This corresponds to a binary system using 10 digits. If 14 digits are used, the possible accuracy of the binary system becomes 1 part in 16 384, or 0·0061 per cent. Also, binary devices are two-state devices, being either 'on' or 'off' and are consequently more reliable than analogue systems which have ideally an infinite number of states.

The bulk of published work on digital control systems considers the problem from the basis of operating servomechanisms from digital data. Information in digital form can be transmitted over long distances and suffer severe distortion without losing accuracy. Much work has been and continues to be done on digital data transmission. This approach leads naturally to a hybrid system in which an error signal is obtained in digital form and converted to an analogue form before it is applied to the servo amplifier and motor. Effectively this is an analogue system fed with information from a digital source. These devices have been investigated in some detail, in particular, the effects of the quantization of the input have been considered.

The conversion from digital to analogue form is accompanied by a loss of accuracy. This is relatively unimportant if the accuracy of the analogue system is equal to or better than that of the digital input. When the digital input is of high accuracy the use of analogue control leads to waste of information. The logical conclusion is to use the digital signal to operate the servo directly. Such a system is described in this article and its block diagram is shown in Fig. 1.

Measurement and Control

MEASUREMENT

The first requirement of a control system is a means of measuring the quantity to be controlled, and of converting this measurement into a form acceptable to the system. In the system being considered, this means measuring the angular position of a shaft and expressing it in digital form. Digital encoding of shaft position has received much attention, particularly in the United States and numerous codes have been devised to eliminate ambiguities and errors which can occur. In this work, a simple binary code was used. In general, the use of other codes requires translating circuits to convert the information to pure binary form.

A circular binary scale of eight digits was printed photographically on to a Perspex disk. Behind each digit ring was mounted a germanium photo-electric cell. The position of the shaft with reference to the radial line containing the photocells is read by projecting a line of light at the photocells. This passes through the disk and the outputs of the cells give the position of the motor shaft as a binary number. Thus the outputs of the cells give a continuous indication of the binary number representing shaft position as the shaft rotates.

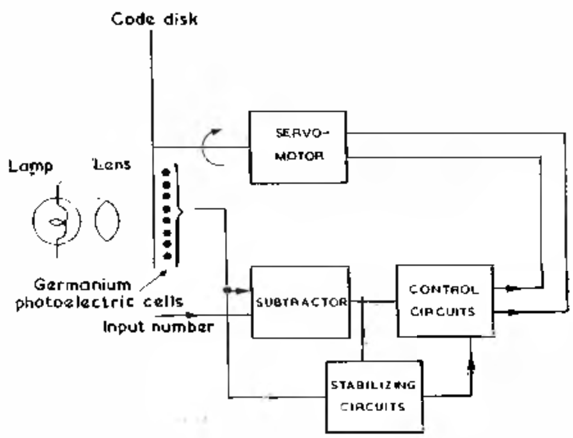

Fig. 1. Block diagram of system

A drawing of the coded disk is shown in Fig. 2 and a table of the angles represented by the binary numbers generated by the disk is given in Table 1. The symbols used to denote the various logical elements used in the system are shown in Fig. 3.

CONTROL

If an analogue system is not to be used, the control circuits cannot be linear and an 'on-off' system must be used. Three states are required, forward, off and reverse. Ideally these should be velocity states since when there is a positive error, a forward velocity is required, when there is no error, no velocity is required and when there is a negative error a reverse velocity is needed. The problem of operating a high accuracy system by this means is that the control circuit states are power states, not velocity states. To avoid overshoot or excessively slow operation, some form of stabilization must be provided.

Serial System

The first approach to the problem was a serial relay system without carry. A serial system works by consider-

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ing each digit separately and then proceeding to the next less significant digit. In each digit the demand is compared with shaft position. A subtraction is performed and the direction of acceleration governed by the sign of the result.

If two or more digits are used, there is a probability that a less significant digit will demand a rotation opposite to that demanded by the most significant. This must be prevented. Each digit must inhibit the operation of succeeding ones until it reaches coincidence.

Minimum Angle Operation

It is essential that a circular position control shall operate in minimum angle, that is, the system shall take the shortest way to cancel an error. Otherwise, if the system is not dead beat and an overshoot occurs, the system will run away.

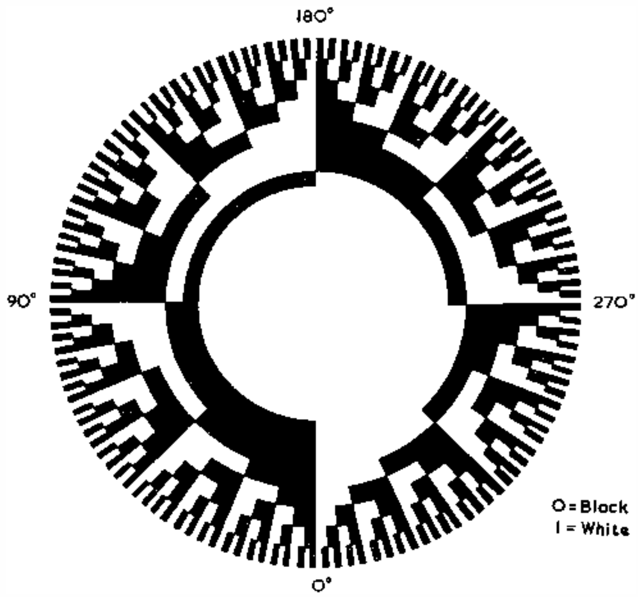

Fig. 2. Diagram of binary code disk

TABLE 1
Four Digit Binary Numbers and their Equivalent Sectors

NUMBER	POSITION (degrees)	NUMBER	POSITION (degrees)
0000	0 to 22·5	1000	180 to 202·5
0001	22·5 to 45	1001	202·5 to 225
0010	45 to 67.5	1010	225 to 247.5
0011	67·5 to 90	1011	247·5 to 270
0100	90 to 112·5	1100	270 to 292·5
0101	112·5 to 135	1101	292·5 to 315
0110	135 to 157·5	1110	315 to 337.5
0111	157·5 to 180	1111	337·5 to 360

Expressed arithmetically, this means that the system must recognize four states of error ϵ viz.

$$0 < \epsilon < \pi$$
$$\pi < \epsilon < 2\pi$$
$$0 > \epsilon > -\pi$$
$$0 > \epsilon > -2\pi$$

or, in terms of two binary digits

$$\epsilon = 01$$
$$\epsilon = 10,\ 11$$
$$\epsilon = -01$$
$$\epsilon = -10,\ -11 \qquad \text{where } \epsilon = \text{error}$$

Suppose that, to cancel a small positive error, the shaft must turn forward, then for positive errors greater than π, reverse rotation is required Similarly a small negative error requires a reverse motion; large negative error a forward motion. A circular system cannot distinguish between $-\theta$ and $2\pi - \theta$, so that a negative error less than π is equivalent to a positive error greater than π.

Thus the direction of motion is governed by whether or not the error is greater than π, that is, by the first binary digit. In order to achieve minimum angle operation, therefore, the presence of a first digit must cause reverse motion.

The serial system previously described does not fulfil this requirement since there is no carry. Considering two digits, the difficulty arises at the 01, 10 boundary. The

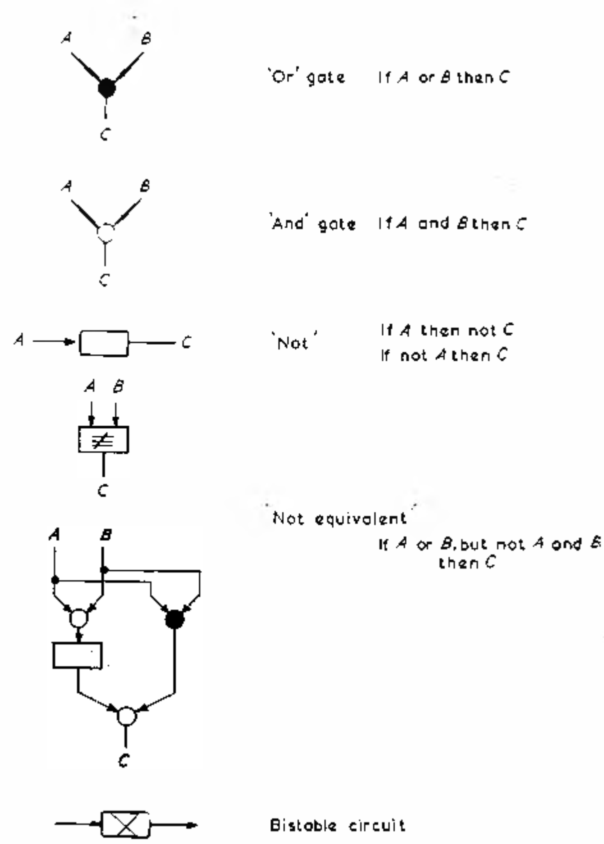

Fig. 3. Table of logical elements

reason for this can be seen by considering the action of the subtractor. Since the subtraction process is non-carrying and each digit subtraction is a separate process, the answer is a three-state or ternary number. When a ternary number is used in a binary system, each ternary number is not unique. If the three states are 1, 0 and T representing 1, 0 and -1, this can be shown as follows. The ternary numbers 01 and $1T$ are both equal to the binary number 01. Applying this to the system proposed, the number 01, having a significant second digit, calls for forward rotation whereas the number $1T$, having a significant first digit, calls for reverse rotation. This is the fundamental reason for the inability of this system to operate in minimum angle.

A method of avoiding this is to perform a carrying binary subtraction of the complete numbers in parallel form. This removes the ambiguity since the subtraction will always translate the ternary number into the corresponding binary number because of the carry.

Parallel System

The requirements of the subtractor for the parallel system are that:

(1) It must perform a binary subtraction of the numbers fed into it.

(2) It must discriminate between answers which are

(a) less than π

(b) greater than π.

The subtractor must provide two items of information. Firstly it must indicate when there is an error, and secondly it must indicate to the servomotor what direction of rotation is required to cancel that error.

The first piece of information must put the motor driving circuits into operation if the output of the subtractor is non-zero. Thus by gating the input to the motor with the output of the subtractor, the motor can be held off until an error is present.

It has been seen that the presence of a first digit signal indicates that the motor should rotate in a reverse direction. To achieve this, a circuit is required which has two possible outputs, one of which is given when there is no signal input, and gives a forward motion, and the other when a signal is applied from the first digit of the subtractor, giving a reverse motion. The logical arrangement of the circuit which performs this is shown in Fig. 4.

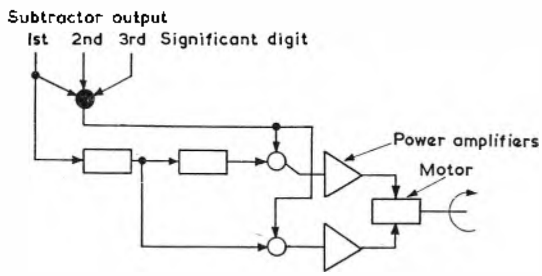

Fig. 4. Logical diagram of control unit

The Subtractor

The general digit circuit of a parallel binary subtractor must examine the numbers to be subtracted, taking into account whether there is a carry from a less significant digit, and also be able to set up a carry to the next more significant digit.

Considering the subtractor in two parts, there is a non-carrying subtractor giving a ternary output, and a convertor which takes 'carrys' into consideration and gives a binary output.

The ternary subtractor gives outputs according to the following table.

$$0 - 0 = 0$$
$$0 - 1 = T$$
$$1 - 0 = 1$$
$$1 - 1 = 0$$

This process is performed on each digit independently.

The binary convertor takes the output of the ternary subtractor, together with any carrys from less significant digits, and gives an output which is a binary number and also a carry on to the next more significant digit. This circuit gives outputs according to the following table.

Ternary subtractor Output	*Carry in*	*Output*	*Carry Out*
0	0	0	0
1	0	1	0
T	0	1	1
0	1	1	1
1	1	0	0
T	1	0	1

Combining the two circuits, the output of the general digit circuit of the subtractor is as follows.

		carry out
no carry in	$0 - 0 = 0$	0
	$0 - 1 = 1$	1
	$1 - 0 = 1$	0
	$1 - 1 = 0$	0
carry in	$0 - 0 = 1$	1
	$0 - 1 = 0$	1
	$1 - 0 = 0$	0
	$1 - 1 = 1$	1

The least significant digit circuit is appreciably simpler than this since there can be no carry in and the most significant digit circuit dispenses with the carry out facility.

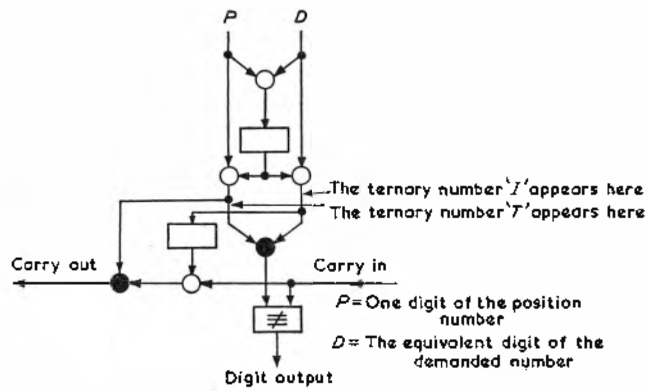

Fig. 5. Logical diagram of the general digit circuit of the subtractor

The logical diagram of the subtractor is shown in Fig. 5.

Stabilization

The system as it stands is damped solely by the mechanical losses in the motor. For the three-digit system studied in this article, an error of π gives about 14 overshoots before the system comes to rest. This is clearly undesirable and some method must be devised of improving the performance of the system.

The approach which has been used is to reverse the motor field excitation at some point during the run in, chosen so as to stop the motor when it reaches coincidence. Two methods have been considered:

(1) Fixed point changeover.

(2) Changeover at half the initial error.

In order to carry out either of these schemes, two further units are required, a velocity unit and a unit to give an output equal to the modulus of error such that it is always less than π. This will be referred to as a modulus of error unit.

Modulus of Error Unit

The purpose of the unit is to ensure that whatever the error and however it is derived, there is available a number representing error which decreases as the angular error decreases. For example, if the initial difference between demand and position number is 011, as the motor turns, the number will change in the sequence 011, 010, 001, 000. This is satisfactory but, if the initial difference is 101 the sequence will be 101, 110, 111, 000. In this instance the value of the difference does not decrease with angular error. In order to produce a number which does decrease in that way, if the difference is greater than or equal to π (i.e. 100 or more), it is complemented and 001 is added to it. This can be seen from the following table.

Number	*Complement*	*Complement* + 1
100	011	100
101	010	011
110	001	010
111	000	001

If the complement alone is used, the answer 000 will be given when in fact the system is 001 from the required position.

The logical arrangement of this is shown in Fig. 6.

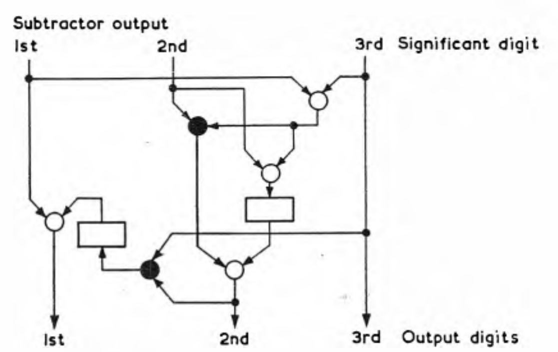

Fig. 6. Logical diagram of modulus of error unit

Velocity Circuit

When the motor first comes to rest after an error has disappeared, the reverse field condition must be cancelled and the system effectively started from a fresh position with a new initial error. In order to do this, a signal must be available which indicates whether the shaft is turning or not.

This is obtained by means of a sampling technique. The edge track of the disk, containing 256 binary digits, is scanned by a photocell and the resulting waveform is gated with a rectangular waveform with an 'on' period of 100msec and an 'off' period of 10msec. The number of edge track digits which appear in the successive sampling periods are counted and the resulting binary number gives a measure of shaft angular velocity.

The sample is counted in a three-stage counter chain which is reset at the beginning of the sampling period. During the 'off' period, the counters hold their states. These are gated with the sampling waveform inverted. The outputs of these three gates give, therefore, at 100msec intervals, the value of the sample as a three-digit number. In order to obtain a continuous signal, these pulses are fed into three 'count holding' bistable circuits, each holding one digit, which are reset at the beginning of each 'off' period of the sampling waveform. Thus from each of these circuits a signal can be taken which changes every sampling period and maintains the value it then takes up until the end of the next sampling period. In this particular application, the value of the velocity is only of secondary interest; the requirement is for an indication that the shaft is turning. This can be obtained by combining the outputs of the three circuits in an 'OR' gate. A schematic diagram is shown in Fig. 7.

Fixed Changeover Scheme

The first method which was used to solve the problem of damping the system was the application of reversed field excitation at a fixed angle from the required position. If the initial error is less than this fixed angle, the system operates in an undamped manner. Ideally, the system should come to rest within the desired digit, but, if it does not, the circuits must return to normal operating con-

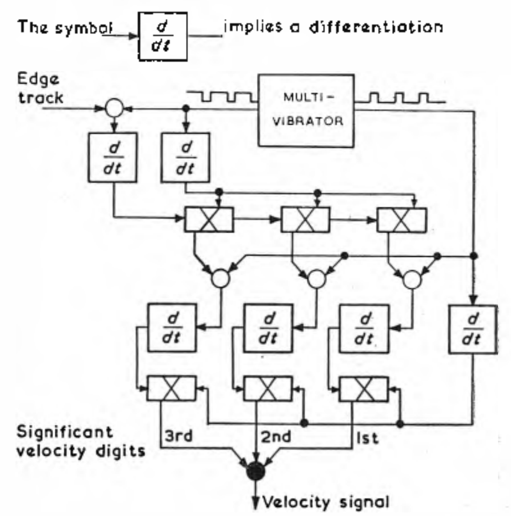

Fig. 8. Logical diagram of fixed changeover stabilizing circuits

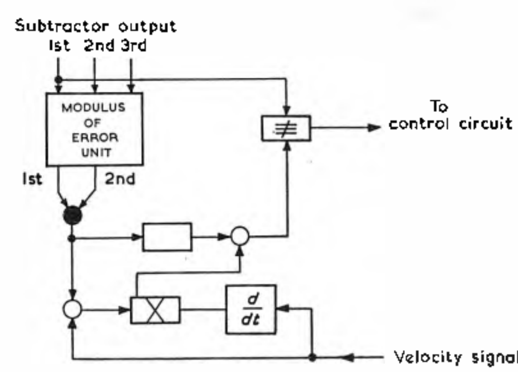

Fig. 8. Logical diagram of fixed changeover stabilizing circuits

ditions and examine the error again. The circuits must perform the following functions:

(1) Detect whether or not the initial error E is greater than the changeover value E_c.

(2) Store this information.

(3) Detect when the error falls to E_c.

(4) Cause the motor field excitation to reverse if $E > E_c$.

(5) Observe when motion ceases and return all circuits to normal.

(6) Render the field reversing circuits inoperative if $E < E_c$.

These functions will be referred to by number.

The changeover value was taken as 001 as a compromise between the values required for different values of E. This has the convenience that the first two digits only of the error need be considered. If either a first or second digit is present the instantaneous error $\epsilon > E_c$ and when both disappear $\epsilon < E_c$. This satisfies requirements (1) and (3).

A bistable circuit, which is triggered when $E > E_c$ and reset when the motor shaft stops, satisfies (2) and also (6). An output is taken from this circuit and is combined with the first digit of the subtractor output and is applied to the motor control described previously in place of the subtractor first digit output. This satisfies (4) and the velocity circuit described satisfies (5). The schematic diagram of this part of the system is shown in Fig. 8.

This system, with fixed changeover reverse field stabilization, is a noticeable improvement on the undamped system. It does not, however, fulfil its expectations. The main reason for this is that a fixed point changeover gives dead beat operation for only one value of E and, also, that once the system gets within 000 $\pm$001 it is undamped. The rather coarse three-digit system which was used does not show the performance of any changeover system to the best advantage, since there can be 45° difference in angular position for the same binary number. Using this system, the number of overshoots was reduced by about four. The response to an initial error of 010 which has seven overshoots undamped has three overshoots when fixed changeover stabilization is used; similarly the response to

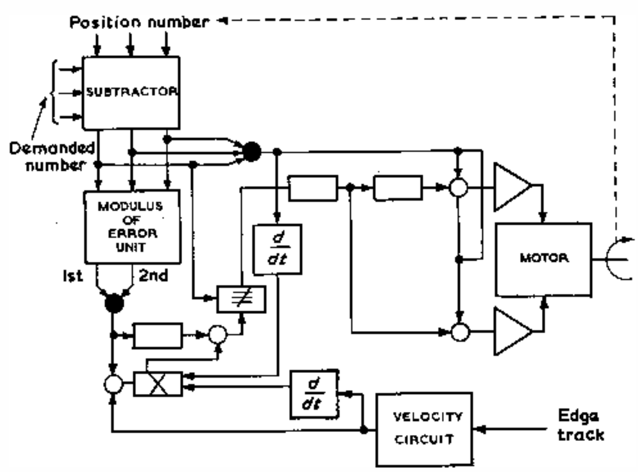

Fig. 9. Complete scheme with fixed changeover stabilization

an initial error of 100 has fourteen overshoots in the undamped system and nine when damping is applied. The improvement which can be obtained by increasing the number of digits is due to better positioning of the shaft and the more closely specified choice of E_c which can be made. However, a problem arises in that with more digit tracks the operation when $E < E_c$ becomes more important and does not improve.

The complete logical diagram of the fixed changeover system is given in Fig. 9.

Half-Initial-Error Changeover

The quickest way for a given system to proceed from point A to point B is to accelerate at a constant maximum rate for half the distance and decelerate at the same rate for the remainder of the distance, assuming that maximum acceleration and deceleration are equal. Here, this means that run in is normal until $E/2$ is reached and then field reversal takes place, causing a reversal of motor torque and hence deceleration.

The division of a binary number by 2 is comparable with the division of a decimal number by 10, i.e. by movement of the point, e.g. 1101011 divided by 2 = 110101·1.

In a practical system, division by 2 is performed by omitting the least significant digit and shifting the remaining digits once. The value of the initial error must be obtained from the modulus of error unit. This number is divided by 2 and stored. Division and storage must occur as soon as possible after an error appears, since any appreciable delay could result in an incorrect number being stored, due to a change in the position of the disk before the circuits had operated. Having obtained and stored this number which represents the point at which reverse field is to be applied, it is compared with the instantaneous value of error. When the difference between these two quantities falls to zero, reverse field is applied. This state must persist until the error is reduced to zero. If overshoot takes place, the field will reverse due to the normal operation of the motor circuit and, consequently, the reverse field which has been applied to stabilize the system must be removed when the system reaches coincidence. This can be done by arranging that, when $\epsilon = E/2$ and reversal of field is called for, the signal indicating equality is used to change the state of a store. This store is reset either by motion ceasing or by arrival at coincidence. If

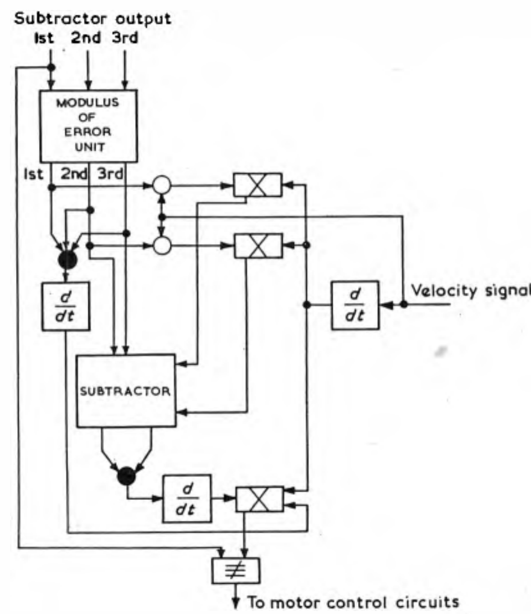

Fig. 10. Logical diagram of $E/2$ changeover stabilization circuit

arrival at coincidence is used to cancel the field reversal, there will be a sector through which the system idles. For a small number of digits this is a disadvantage, but as the accuracy of the system is increased, the sector narrows and becomes correspondingly less important. Using this technique for stabilization, the logical arrangement for which is shown in Fig. 10, much better performance can be expected.

Conclusions

The problem of arranging a digitally operated control system is relatively straightforward. The real problem is to devise methods of stabilizing such a system. In the absence of the conventional analogue stabilizing techniques, even such as are used in 'bang-bang' servos, some such method as has been described, depending on the application of a braking torque at a certain distance from coincidence, appears to be necessary. Other systems could be based on relating shaft velocity, a measure of which exists in the device, to instantaneous error. This method could be used to provide a digital device corresponding very closely to a 'bang-bang' servo system. (A 'bang-bang' servo system is one in which small deviations to one side or the other of the required position produce full motor torque in the appropriate direction. Such devices commonly include relays to control the power applied to the servomotor.)

The biggest single item in the system which was constructed was the subtractor, which required 24 transistors and 40 diodes for three digits. For a 10-digit machine, which would have a comparable accuracy to an analogue system the subtractor would require 101 transistors and 166 diodes and rather more again for the stabilizing circuits.

Such a system as this has a potentially higher order of accuracy than an analogue machine, but at the price of much greater complexity.

Apart from the obvious applications of a position control system to machine tools, ballistics and other power handling applications, this type of system could have uses in remote direct instrumentation. There are many occasions when an indication of the behaviour of a quantity is necessary in order to control it, and in addition an indication is required elsewhere either for recording purposes or to assist in general plant control. A system to perform this function need not be more than 1 per cent accurate, or 7 binary digits, depending upon the accuracy of the master instrument and would be less complex and expensive than a control system attempting to compete with analogue systems. Furthermore, by using a time-sharing method, the same instrument could be used to display the behaviour of a number of different quantities within a plant.

In a practical machine it would be desirable to use a non-ambiguous disk code. There are several such codes in existence, most of which have been developed for data transmission work. Many of these codes are not amenable to arithmetical manipulation, and they must be translated to simple binary form before the numbers can be processed.

Considerable simplification both in circuits and in logical arrangements could possibly be achieved using ternary devices in place of the binary ones used in this work.

Acknowledgments

The author wishes to thank Sir Willis Jackson, F.R.S., Director of Research and Education, Metropolitan-Vickers Electrical Co. Ltd, for permission to publish this article.

Appendix 1

Electrical Details

B.T.H. type GT3 junction transistors were used for all inverting and pulse shaping circuits. These were operated between voltages of +5V, earth and −15V. Mullard type OC76 junction transistors were used to control the motor field currents, with a negative line voltage of −30V. A position-indicating system was devised in which Mullard type OC72 junction transistors switched on indicator lights to show the position of the disk.

Gating circuits were constructed with S.T.C. type Q8/2 Unistors, these being miniature selenium rectifiers.

Circuits were standardized to a large extent but it did not prove possible to standardize fully owing to differing input and output conditions from one circuit to another and, to a certain extent, to transistor variations.

The servomotor used was a reversing motor generator type XC331 with a split field winding. The centre point of the field winding was taken to the negative supply and the two ends of the winding were fed separately from separate OC76 amplifier systems.

Appendix 2

Error Halving Stabilization System

Given an initial error E, then assuming that both field systems are symmetrical and have equal currents passed through them, acceleration under positive field drive is equal to deceleration under negative applied field (where positive and negative indicate field polarity relative to direction of motion). Therefore, provided the motor never achieves maximum speed, to ensure accurate operation with a minimum of overshoots field polarity should be changed at $E/2$.

Graphically, this gives results, shown in Fig. 11, assuming uniform acceleration. This is true as a first approximation, neglecting the effect of armature reaction.

If motion ceases when either the system has not reached coincidence or when it has overshot, the process repeats itself until coincidence is reached.

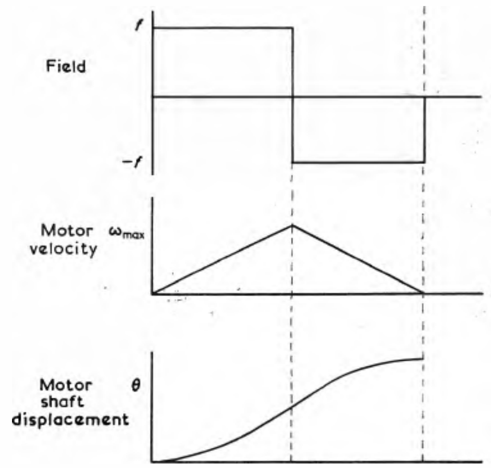

Fig. 11. Motor velocity and displacement, error-halving stabilization

Appendix 3

Time Lags

(1) *Circuit Lags*

Under the circuit conditions which obtain, an approximate figure of 0·5μsec can be given for the turn on time of a transistor, and approximately 6μsec for turn off and hole storage times. These figures can only be considered as orders of magnitude.

In the general digit circuit of the subtraction, there are five transistors in the digit path and five in the carry path. If half of these are switched on and half switched off, there is a time lag of the order of 20μsec, since the digit and carry paths are in parallel, although under certain conditions it could be as low as 5μsec.

The self-capacitance of a Q8/2 rectifier is 500pF. These are usually associated with a 47kΩ resistor in a gate. The time-constant is therefore of the order of 25μsec.

There are three gates in the digit path and four in the carry path of the general digit circuit. Since all these gates will not necessarily operate at any given time, 75μsec could be taken as an order of magnitude. The total gating lag in the carry line can be taken as 150μsec.

The switching circuit has two gates and four transistors one of which is on and three are off. If the motor is being started from rest, there will be a time lag of 7μsec for the transistors and 25μsec for the gates, assuming that only one of the two operates.

The total lag due to the circuit is therefore of the order of 200μsec.

(2) *Motor Lag*

The motor armature is fed continuously.

Field

Field coil resistance	=	$1\,250\Omega$
Resistance in external circuit (i.e. Z_o of OC76)	$\simeq$	10Ω
$\therefore$ Total resistance	=	$1\,260\Omega$
Field inductance	=	11H
$\therefore$ Motor field time-constant	=	8·8msec
$\therefore$ Total lag to 67% full power, including circuit lags	=	9msec

(3) *Velocity Lag*

Period of sampling waveform	=	110msec
Net sampling period	=	100msec

At changeover, maximum speed varies from 010 to 111. These correspond to edge track digits, of which there is a total of 256. The counter records positive digits only, so there are 128 velocity digits in the circle.

If there is a velocity count of 7 in 100msec, this is 1/16rev per 110msec which is approximately equal to the time taken by one 4th significant digit.

(4) *Inertia Lags*

The principal lags in the system are the two introduced by the motor and load inertias. The moment of inertia of the motor armature is 0·56 lb.in^2. The load which the motor drives is the binary coded disk whose moment of inertia is 8·6 lb.in^2. The total load inertia is therefore approximately 9·2 lb.in^2.

(5) *General*

The lags due to the switching of transistors are small and can be neglected by comparison with the lags due to the operation of gates. These are cumulative as the number of digits increases, due to the carry path. Even so, the gate lags are small compared with the motor time-constant. However, the effect of this is small compared with that of the inertia load on the motor.

Appendix 4

Analysis

The system is only solvable analytically if it is considered in a piecewise linear manner. The system will be considered firstly to have uniform equal acceleration and deceleration and secondly uniform unequal acceleration and deceleration.

(1) *The Undamped Condition*

Assume uniform acceleration f

Let s be initial distance from coincidence

Let t_1 be time of acceleration.

At distance s from coincidence $v = 0$

At distance 0 $\quad v = \sqrt{(2fs)}$

$$v = ft$$

Now $\quad s = \frac{1}{2} ft^2$

$$t_1 = \sqrt{(2s/f)}$$

Now the motor idles for a distance $001 = \sigma$

$\therefore$ at distance -001 $\quad v = \sqrt{(2fs)}$

$$\therefore\ t_1 + t_2 = \sqrt{(2s/f)} + \sqrt{(\sigma^2/2fs)}$$

At $v = 0$ a further distance s has been covered. Thus, the system will oscillate with a period

$$\tau = 2\left\{2\sqrt{(2s/f)} + \sqrt{(\sigma^2/2fs)}\right\}$$

$$= \frac{2(4s + \sigma)}{\sqrt{(2fs)}}$$

(2) *Fixed Changeover Operation*

Let changeover occur at distance a from coincidence.

If $0 < s < a$ the system proceeds as in (1)

If $s > a$ the system behaves differently.

At s, $\quad v = 0$

At distance a, $v = \sqrt{[2f(s - a)]}$

$$2(s - a) = ft_1^2$$

$$\therefore \quad t_1 = \sqrt{\left(\frac{2(s-a)}{f}\right)}$$

The system is now distance a from coincidence.

There are three possibilities.

(a) $a < s < 2a$

(b) $s = 2a$

(c) $s > 2a$

(a) If $a < s < 2a$

The system will travel a further $(s - a)$ and come to rest in a time

$$t_2 = \sqrt{\left(\frac{2(s-a)}{f}\right)}$$

$$\text{total time } t = 2\sqrt{\left(\frac{2(s-a)}{f}\right)}$$

This does not bring the system to 0 and motion restarts and proceeds as in the undamped case.

(b) If $s = 2a$

a further $(s - a)$ brings the system to 0, so it comes to rest at coincidence and

$$t = 2\sqrt{(2a/f)}$$

(c) If $s > 2a$

the system is still moving at distance 0 and the investigation must proceed further.

The system decelerates for a distance a from changeover, taking a time

$$t_2 = \sqrt{(2a/f)}$$

the velocity at distance 0

$$= \sqrt{[2f(s - a) - 2fa]}$$

$$= \sqrt{[2f(s - 2a)]}$$

The system then idles for a distance $\sigma = 001$, taking a time

$$t_3 = \frac{\sigma}{\sqrt{[2f(s - 2a)]}}$$

The system now decelerates for a distance $(s - 2a)$ and comes to rest at a new $s' = -(s - 2a)$, having taken a total time

$$t = \sqrt{\left[\frac{2(s-a)}{f}\right]} + \frac{\sigma}{\sqrt{[2f(s-2a)]}} + \sqrt{\left[\frac{2(s-2a)}{f}\right]} + \sqrt{(2a/f)}$$

$$= \frac{2\sqrt{[(s-a)(s-2a)]} + \sigma + 2(s-2a) + 2\sqrt{[a(s-2a)]}}{\sqrt{[2f(s-2a)]}}$$

$$= \frac{2(s-2a) + \sigma + 2\sqrt{(s-2a)}\,[\sqrt{(s-a)} + \sqrt{a}]}{\sqrt{[2f(s-2a)]}}$$

The system then restarts with the new value of s' and repeats until either $s' = 2a$ when the system will start and come to rest at 0 or $s' < a$ when the system reverts to (1). The period of one half cycle is

$$t = \frac{2(s-2a) + \sigma + 2\sqrt{(s-2a)}\,[\sqrt{(s-a)} + \sqrt{a}]}{\sqrt{[2f(s-2a)]}}$$

(3) *E/2 Changeover Operation*

If changeover occurs at $s/2$, the condition in (2) that $s = 2a$ is automatically fulfilled, and the system comes to rest at $s = 0$ after a time

$$t = 2\sqrt{\left(\frac{2(s - s/2)}{f}\right)}$$
$$= 2\sqrt{(s/f)}$$

(4) *Unequal Acceleration and Deceleration*

In all the foregoing it has been assumed that acceleration and deceleration have been equal and constant. In practice, this state of affairs will not hold and they will be neither equal nor constant. Since the highest speed which the motor achieves is low, the effects of windage and falling off in the torque-speed curve can be neglected, and the error in assuming the acceleration to be constant is small. A closer approximation to the physical state of affairs can be obtained by superimposing on the acceleration and deceleration produced electromechanically, a constant deceleration term f'.

Thus, considering the undamped condition of (1) at distance 0

$$v = \sqrt{[2(f - f')s]} = (f - f')t$$
$$t_1 = \sqrt{\left(\frac{2s}{(f - f')}\right)}$$

During the region in which there is no applied acceleration, the system decelerates at f'. The speed on entering this section is

$$v = \sqrt{[2(f - f')s]}$$

and the section extends for a distance σ.

The speed on leaving this section will therefore be

$$v^2 = -2f'\sigma + 2(f - f')s$$
$$\therefore v = \sqrt{[2(fs - f'(\sigma + s))]}$$

The time spent in this region

$$= \frac{v - u}{f'}$$
$$= t_2 = \frac{\sqrt{[2(f - f')s]} - \sqrt{[2(fs - f'(s + \sigma))]}}{f'}$$

Deceleration commences at $-(f + f')$. For a distance

$$s' = \frac{v^2}{2(f + f')}$$
$$= \frac{(fs - f'(\sigma + s))}{(f + f')}$$

and for a time

$$t_3 = \sqrt{\left[\frac{2(fs - f'(s + \sigma))}{(f + f')^2}\right]}$$

$\therefore$ total time for half a cycle

$$t = \sqrt{\left(\frac{2s}{f - f'}\right)} + \frac{\sqrt{[2(f - f')s]} - \sqrt{[2(fs - f'(s+\sigma))]}}{f'} + \sqrt{\left[\frac{2(fs - f'(s + \sigma))}{(f + f')^2}\right]}$$

$$t = \frac{f(f + f')\sqrt{(2s)} - f\sqrt{[2((f - f')^2 s - f'(f - f')\sigma]}}{f'(f + f')\sqrt{(f - f')}}$$

The system now stands at a new s'

$$= \frac{(fs - f'(s + \sigma))}{(f + f')}$$

and restarts. The process repeats until the system comes to rest at 0.

Considering the fixed changeover system of (2) and considering the case when $s > 2a$.

Let changeover occur at distance a from coincidence. At distance a

$$v = \sqrt{[2(f - f')(s - a)]}$$
$$\therefore t_1 = \sqrt{\left[\frac{2(s - a)}{(f - f')}\right]}$$

Deceleration now takes place at $-(f+f')$ for a distance a and takes a time

$$t_2 = \sqrt{\left(\frac{2a}{f + f'}\right)}$$

The velocity at this point

$$= \sqrt{[2(f - f')(s - a) - 2a(f + f')]}$$

The system now decelerates for a distance σ at f'. The velocity on leaving this section will be

$$v^2 = -2f'\sigma + 2(f - f')(s - a) - 2a(f + f')$$
$$= 2f(s - 2a) - 2f'(s + \sigma)$$
$$\therefore v = \sqrt{[2f(s - 2a) - 2f'(s + \sigma)]}$$

and takes a time

$$t_3 = \frac{v - u}{f'}$$
$$= \frac{\sqrt{[2(f - f')(s - a)]} - \sqrt{[2f(s - 2a) - 2f'(s+\sigma)]}}{f'}$$

The system now decelerates at $(f + f')$ until $v = 0$. It covers a distance

$$s = \frac{f(s - 2a) - f'(s + \sigma)}{(f + f')}$$

and takes a time

$$t_4 = \sqrt{\left[\frac{2f(s - 2a) - 2f'(s + \sigma)}{(f + f')^3}\right]}$$

$\therefore$ the total time taken to complete a half cycle

$$t = t_1 + t_2 + t_3 + t_4$$

$$= \sqrt{\left[\frac{2(s - a)}{(f - f')}\right]} + \sqrt{\left[\frac{2a}{(f + f')}\right]} + \frac{\sqrt{[2(f - f')(s - a)]} - \sqrt{[2f(s - 2a) - 2f'(s + \sigma)]}}{f'} + \sqrt{\left[\frac{2f(s - 2a) - 2f'(s + \sigma)}{(f + f')^2}\right]}$$

$$= \frac{f(f + f')\sqrt{[2(s - a)]} - f\sqrt{[2^2(s - 2a) - 2ff'(2s + \sigma - a) + 2f'^2(s + \sigma)]} + f'\sqrt{[2a(f - f'^2)]}}{f'(f + f')\sqrt{(f - f')}}$$

The system then restarts from the new value s' and proceeds until either $s' = 2a$ or $s' < a$. The behaviour when $s' < a$ was discussed in (4). If $s' = 2a$ the system will ideally stop at a distance short of 0 by

$$\frac{2f'a}{(f + f')}$$

after a time

$$t = \frac{\sqrt{2a[(f + f') + \sqrt{(f^2 - f'^2)}]}}{(f + f')\sqrt{(f - f')}}$$

after which it restarts and comes to rest at 0.

Waveguide as a Long Distance Communication Medium

By A. E. Karbowiak*

A discussion is given of a communication system utilizing waveguide as the communication medium. All the more important aspects of such a communication system are analysed and the properties of various types of waveguide are discussed in some detail. It is shown that although plain metallic waveguides are unsuitable for long distance communication, waveguides of special construction (either dielectric coated or helical waveguides of appropriate proportions) can be designed to fulfil any reasonable practical requirements and that a communication system of extremely wide bandwidth can be constructed.

(Voir page 570 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 575)

A LONG distance waveguide communication system consists of a number of well defined functional elements (Fig. 1). The individual communication channels are grouped together and passed through a coder (C), where the information to be transmitted emerges coded in p.c.m., the reason for this procedure will become evident later on. The coded signal is made to activate an on-off modulator (M), operating on a suitable carrier frequency (microwaves). The potential capacity of the waveguide, however, is considerably in excess of the basic arrangement described and, therefore, a number of modulators (M_1 to M_n) operating on different frequencies (f_1 to f_n respectively)

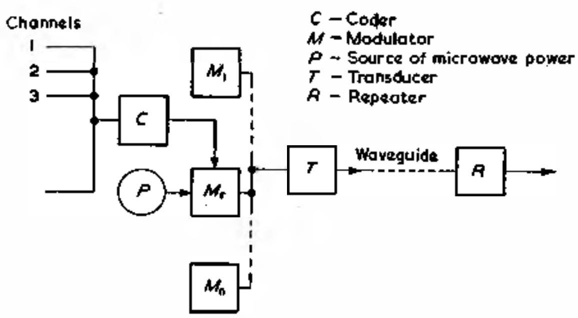

Fig. 1. A possible communication system using waveguide

are connected in parallel. Their outputs are combined and passed through a transducer (T) to the waveguide.

At intervals (of say 20 miles) along the waveguide, regenerative repeaters of appropriate construction are inserted. In this way, the accumulation of signal distortion—which would take place if non-regenerative repeaters were employed—is largely prevented and communication over long distances becomes a possibility[1,2].

The principal features of the waveguide, the medium of communication, are tabulated in Table 1 for comparison with other systems currently in operation. An examination of Table 1 reveals that the real advantage of a waveguide communication system lies in its potential capacity for simultaneous handling of a very large number of communication channels and to some extent because it is a screened system and because of its low attenuation. All other features of the waveguide communication system are really of minor importance.

There are two principal sources of signal distortion after transmission through a length of waveguide:

(1) Delay distortion[3].

(2) Mode conversion-reconversion[1,2].

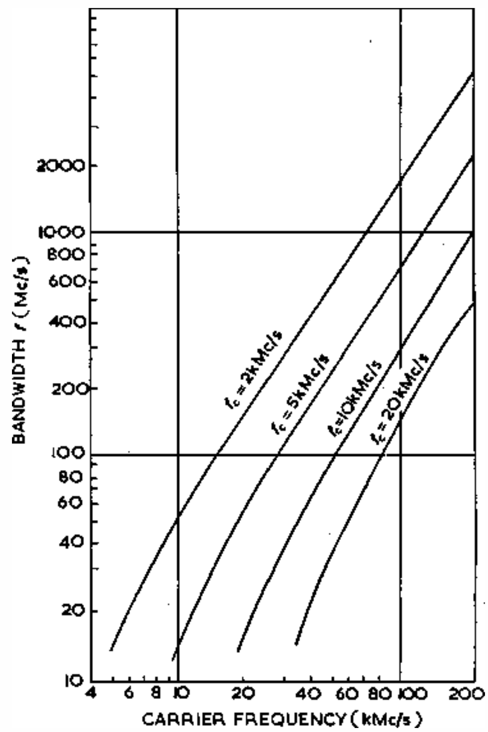

Fig. 2. Bandwidth as a function of frequency

f_c = cut-off frequency

$\delta\phi$ = I length 30km

Delay distortion is a direct result of the curvature of the $\beta(\omega)$ characteristic (or velocity of propagation/frequency) of the waveguide. It simply means that the high frequency components of the signal travel faster than the low frequency components (dispersion). Consequently a typical waveguide cannot transmit, without substantial distortion,

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intelligence occupying a bandwidth of more than 100Mc/s (see Fig. 2) over distances in excess of about 30km. The phenomenon of dispersion is an inherent feature of the waveguide and cannot be successfully compensated by suitable waveguide design, and must, therefore, be taken care of in the design of the terminal equipment.

The mode conversion-reconversion takes place because the waveguide can support a large number of modes apart from the desired mode of transmission (H_{01}) and also because of the irregularities in the waveguide structure. In effect, if the waveguide is excited at one end by a pure H_{01} mode, the wave on its passage down the waveguide becomes scattered by all the minute waveguide irregularities (small dents, changes in the diameter and cross-sectional shape, small bends and kinks, etc.) and partially transformed into the parasite modes. This process results in the gradual diffusion of energy into the parasitic modes, thus effectively increasing the attenuation of the H_{01} wave: the process is known as mode conversion. This, however, is not the complete phenomenon for, by reciprocity, it cannot exist on its own but of necessity is accompanied by the process of 'reconversion'. By reconversion is understood the process whereby the energy now partially carried in the parasitic modes becomes reconverted back into the H_{01} mode, at every waveguide irregularity. But since the group velocities of the parasitic modes, in general, differ from the group velocity of the H_{01} wave, the reconversion is not in phase with the energy carried by the H_{01} wave, which results in signal distortion. Moreover, since the phase of the reconverted signal depends on the distance between the reconversion points, expressed in terms of the wavelength, the magnitude of the mode conversion-reconversion is frequency sensitive.

Mode conversion-reconversion effects are of such a character that it is inconceivable that successful terminal equipment could be invented to nullify its harmful effects and must, therefore, be combated by judicious waveguide design.

Yet even if these precautions are taken, it is still necessary to code (e.g., in p.c.m.) all the information to be transmitted and regenerate it at as frequent intervals as needed, to avoid an overall limitation on the total circuit length.

Waveguides suitable for long distance transmission cannot, by their very nature be bent to very small radii of curvature (of the order of a few metres) and retain good transmission characteristics. This, however, need not be put on the debit side of the waveguide because normally bending radii of the order of a few hundred metres is all that would be required from a commercial waveguide—special waveguide when properly designed can certainly achieve this[2]. Bends having radii smaller than about 100m are required only very infrequently and in such situations

TABLE 1

Comparison of Communication Media

TYPE		OPERATING FREQUENCY	CHANNEL CAPACITY	REPEATER SPACING/ ATTENUATION
Open wire	One pair	36-84kc/s (go) 92-140kc/s (return)	12+3+(1) = 15+(1)	60-70 miles
	Maximum 16 pairs	As above	240+(16)	(0·4dB/mile)
Balanced pair	One pair	12-252kc/s	60	12-30 miles
	Maximum 24 pairs	As above	1440	(4dB/mile)
Coaxial G.B.		Up to 4Mc/s	960 (One super channel)	6 miles
		Up to 12Mc/s	3 super channels	3 miles
Coaxial U.S.A.		Up to about 3Mc/s	720	8 miles
		L.3 system Up to about 8Mc/s	1 Television+ 600 speech channels	4 miles
Single wire transmission line		About 100-1 000Mc/s	Probably a few television channels	About 2-10 miles
Microwave Links		Below 500Mc/s	Maximum 60	40-50 miles
		500-1 000 Mc/s	Up to 120	40-50 miles
		2 000Mc/s	240 × 6 or 1 television × 6	30-40 miles
		4 000Mc/s	600 × 6 or 1 television × 6	25-30 miles
		6 000-8 000Mc/s	600 × 6 or 1 television × 6 Max. 2 television × 6	25-30 miles
		11 000Mc/s	Less than 600	Less than 20 miles
Long Haul Waveguide (H_{01})		30 000Mc/s up to about 100 000Mc/s	1 000 super channels ≡ several hundreds of television channels ≡ several hundred thousands of speech channels	20-40 miles (2 to 4dB/mile)

NOTE: *One super channel ≡ 4Mc/s*

bends and corners of special designs having excellent transmission characteristics can be employed.

It will be shown that a waveguide, when properly designed, will give satisfactory service and will handle successfully a bandwidth of 10Mc/s or more, which is effectively equivalent to a total intelligence bandwidth of more than 1 000Mc/s, a capacity not offered by any existing communication system.

Uniform Waveguides and Their Principal Properties

The performance of a waveguide as a communication medium (its phase and attenuation characteristics) depends on four factors:—

(1) Its geometry (cross-section and length) in relation to wavelength/frequency.

(2) Its mode of operation

(3) Nature and magnitude of the surface impedance of its walls.

(4) Tolerances on its cross-section, their nature and variation along the length of the waveguide.

Circular waveguide geometry is chosen because such waveguides can be made to high tolerances and also because in a circular waveguide the desired wave, H_{01}, can be made to propagate with small attenuation. Furthermore, circular waveguides can be constructed in such a way as to discriminate between various undesirable modes of propagation.

The particular mode of operation, H_{01}, is chosen for long distance communication for a number of reasons. The wave has a relatively small attenuation in a convenient frequency band (20 to 100kMc/s) and has no radial plane of polarization. Furthermore, the H_{01} mode belongs to the H_{0n} family of modes, for which special waveguides can be constructed to minimize mode conversion-reconversion phenomena. Of all the H_{0n} modes, the H_{01} suffers smallest attenuation and smallest dispersion. This mode is also relatively insensitive to waveguide irregularities and particularly to imperfect joints.

The design of a waveguide communication link is—like any other engineering undertaking—full of compromises decided on in the light of customers' requirements and economy.

The diameter of the waveguide is a compromise between the permissible level of mode conversion-reconversion on one side and the magnitude of attenuation and delay distortion on the other. The larger the diameter the smaller the delay distortion and attenuation, but the larger the mode conversion-reconversion and *vice versa*.

Modulus of overmoding, M (= the number of possible modes in the waveguide), is a geometrical parameter and is given by

$$M = 2{\cdot}55\,(D/\lambda_0)^2 \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where D is diameter of the waveguide and λ_0 the free-space wavelength, both measured in the same units.

The formula is valid provided $M > 10$. Typically M is in excess of 100 (Fig. 3), and such waveguides are termed overmoded The phase (V_{ph}) and group velocities (V_g) of the individual modes are given by the simple expressions

$$\frac{V_g}{V_0} = \frac{V_0}{V_{ph}} = \sqrt{(1 - F^2)} \quad \ldots\ldots\ldots (2)$$

where $F = \lambda_0/\lambda_c = f_c/f_0 = h_0/k_0 \quad \ldots\ldots\ldots (3)$

= the ratio of the cut-off frequency to the carrier frequency.

For homogeneous waveguides the attenuation[4] of the various modes are given by*

$$\alpha_{mn}^{(e)} = \frac{8{\cdot}686}{s} \frac{R_s}{\sqrt{(1 - F^2)}} \quad \ldots\ldots\ldots (4)$$

for E- waves and

$$\alpha_{mn}^{(h)} = \frac{8{\cdot}686}{s} \frac{F^2}{\sqrt{(1 - F^2)}}\left[1 + \frac{1}{F^2} \frac{(m/\chi_{mn}^{(h)})^2}{1 - (m/\chi_{mn}^{(h)})^2}\right] R_s \quad \ldots\ldots\ldots (5)$$

for H-waves.

The attenuation is given in decibels/metre if s (the waveguide radius) is measured in metres, R_s is the normalized (with respect of $Z_0 = 377\Omega$) surface resistance of the waveguide walls and is given by[4]

$$R_s = \sqrt{(\pi f \mu/\sigma)} \div \sqrt{(\mu_0/\epsilon_0)} = R_m \quad \ldots\ldots\ldots (6)$$

where σ is the conductivity of the metal, f carrier frequency, μ permitivity and $\sqrt{(\mu_0/\epsilon_0)} = Z_0 = 377\Omega$.

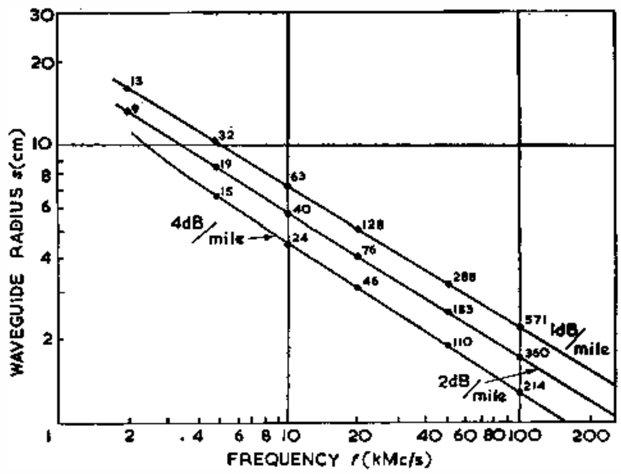

Fig. 3. Waveguide radius plotted against frequency

The numbers denote the number of propagating modes
Attenuation figures refer to copper waveguide

At a free space wavelength of 8·7mm R_m for a copper surface ($\sigma = 5{\cdot}8 \times 10^7$ mhos/m) is $1{\cdot}28 \times 10^{-4}$.

The attenuation of the H_{01} wave is given by

$$\alpha_0 = \frac{8{\cdot}686}{s} \frac{F^2}{\sqrt{(1 - F^2)}} R_m \quad \ldots\ldots\ldots (7)$$

which for a copper pipe of 7cm diameter is 1·2dB/mile.

Commercial metallic pipes, however, have surface impedance which is higher by some 20 to 50 per cent than the values given above, because of the physical state of the surface (surface roughness).

It can be shown from the above expressions that because F^2 is normally a small quantity (typically 0·02) the attenuation of the E-waves is some 50 times larger than that of H_{01} mode (α_0), but that the attenuation of some of the H-modes† notably the H_{02}, H_{12} and H_{13} is only about 3 to 6 times larger than α_0. Furthermore the velocities of H_{01} and E_{11} modes are identical and the modes H_{11}, H_{21}, H_{31}, H_{41}, H_{22} differ in velocity from the H_{01} mode by only one per cent or less. These modes tend to be particularly troublesome.

If the waveguide is filled with a dielectric of high di-

* *The quantity $\chi_{mn}^{(h)}$ is the n^{th} root of $J_m'(x) = 0$.*

† *Typically for a 7cm diameter waveguide operated at $\lambda_0 = 8{\cdot}7$mm the attenuations of the more troublesome modes (expressed as multiples of α_0) are: H_{02}—3·4, H_{12}—3·5, H_{13}—5·8, H_{22}—7·4, H_{03}—7·5, H_{23}—9·3, H_{32}—12, H_{33}—13·2, H_{11}—13·2, H_{24}—14·8, H_{42}—16·7, H_{43}—17·9, H_{15}—18·5, H_{11}—18·6, etc.*

electric constant then the group velocities of all the modes will be increased and so will the number of propagation modes. If the dielectric were perfect then the attenuation of some of the modes would decrease (notably that of the H_{01} mode) and of others would increase. Since, however, the loss angle of available dielectrics is in excess of 0·0001, the waveguide attentuation is dominated by the loss in the dielectric. this is orders of magnitude higher than the loss in the waveguide walls. For this reason alone dielectric filled waveguides are impractical.

Similarly elliptic waveguides have little to offer that could be of commercial importance.

On the other hand, waveguides whose surface impedance is anisotropic[4,5] are of extreme interest. The propagation of such waveguides can be most easily obtained by modifying the parameters of a perfect waveguide, as follows.

The propagation coefficient of a perfect waveguide (whose walls are perfectly conducting) is given by

$$= \mathrm{j}\,\beta_o = \mathrm{j}\,2\pi/\lambda_o\,\surd(1-F^2) \quad \ldots\ldots\ldots\ (8)$$

where F^2 is the ratio of the cut-off frequency to the carrier frequency of the mode concerned.

The propagation coefficient of an anisotropic waveguide is, in general, given by

$$\gamma = \gamma_o + \delta\gamma = \mathrm{j}\,\beta_o + (\alpha + \mathrm{j}\delta\beta) = \alpha + \mathrm{j}\beta \quad \ldots.\ (9)$$

where the quantity $\delta\gamma$ $(= \alpha + \mathrm{j}\delta\beta)$ is composed of the attenuation coefficient (α) and a quantity $\delta\beta$ modifying the original phase change coefficient β_o. In general, α is proportional to the surface resistance components and $\delta\beta$ is proportional to the surface reactance components of the waveguide surface.

The quantity $\delta\gamma$ is given—depending on the mode concerned and the surface impedance components* Z_ϕ and Z_z —by†

$$\alpha = 1/s \quad R_z/\surd(1-F^2) \quad \ldots\ldots\ldots\ (10)$$

for E-waves,

$$\alpha = 1/s\,\frac{F^2}{\surd(1-F^2)}\left[R_\phi + R_z\frac{1-F^2}{F^4}\left(\frac{m}{2\pi}\ \frac{\lambda_o}{s}\right)\right]\left[1-\left(\frac{m}{2\pi}\ \frac{\lambda_o}{s}\right)^2\right]^{-1} \ldots\ (11)$$

for H_{mn} waves in general and

$$\alpha_o = \frac{1}{s}\ \frac{F^2}{\surd(1-F^2)}R_z \quad \ldots\ldots\ldots\ (12)$$

for H_{on} waves in particuar.

In the above formulæ F^2 is the ratio of the cut-off frequency to the carrier frequency, λ_o is the free-space wavelength. If X_z, X_ϕ are substituted respectively for R_z, R_ϕ then the quantity $\delta\beta$ will be obtained.

There are three types of waveguides of particular importance (1) dielectric coated, (2) ring structures and (3) helical waveguides[2].

TABLE 2

The Surface Impedance of a Dielectric Coated Metal Surface

Surface Impedance	Real Part	Imaginary Part
$Z_z = R_z + \mathrm{j}X_z$	$R_z = tk_o(\tan\delta/\epsilon_r)$	$X_z = tk_o(1-1/\epsilon_r)$
$Z_\phi = R_\phi + \mathrm{j}X_\phi$	$R_\phi = \frac{1}{3}(tk_o)^3(\epsilon_r\tan\delta)$	$X_\phi = \frac{1}{3}(tk_o)^3(\epsilon_r - 1)$

These values must be augmented by $Z_m = R_m + \mathrm{j}X_m$ $(R_m = X_m = \surd(\pi f\mu/\sigma)\ \surd(\epsilon_o/\mu_o))$, *the surface impedance of the metal.*

* *Z_ϕ is the surface impedance presented to the currents flowing in the waveguide walls in the circumferential direction, and Z_z is the surface impedance presented to the currents flowing in the waveguide walls in the axial direction.*

† *These attenuations are given in nepers/metre if s is measured in metres.*

(1) Dielectric coated waveguide

This is a cylindrical waveguide having on its inner surface a thin layer of a dielectric. If the thickness of the dielectric is t (Fig. 4) its relative permitivity ϵ_r and its loss-angle $\tan\delta$, then its surface impedance components are simply as given in Table 2.

It will be observed that by choosing suitable materials and thicknesses of dielectric coating it is possible to adjust the surface impedance components within wide limits and, therefore, the attenuation and phase change coefficients of the E and H modes relative to the H_{on} modes can be modified to fulfil any reasonable requirements. As an illustration consider a 7cm diameter copper pipe provided with a dielectric coating (constants $\epsilon_r = 2$, $\tan\delta = 0{\cdot}15$) thickness $t = 0{\cdot}046$mm at a free-space wavelength $\lambda_o = 8{\cdot}7$mm.

From the formulae shown in Table 2 the following values for the surface impedance (normalized with respect to $Z_o = 377\Omega$) will be obtained.

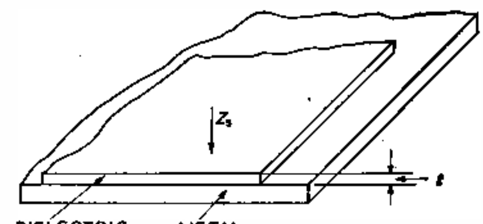

Fig. 4. Dielectric coated surface

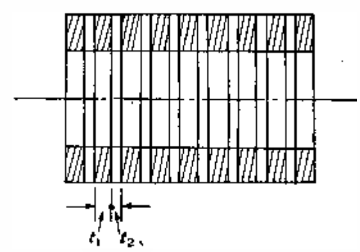

Fig. 5. Disk waveguide

$$Z_z = 2{\cdot}5 \times 10^{-3} + \mathrm{j}\,1{\cdot}7 \times 10^{-2} + Z_m$$

$$Z_\phi = 3{\cdot}7 \times 10^{-6} + \mathrm{j}\,3{\cdot}3 \times 10^{-4} + Z_m$$

The copper intrinsic impedance, Z_m, is

$$Z_m = 1{\cdot}25 \times 10^{-4}(1 + \mathrm{j})$$

These values can be substituted in equations (10) to (12) to obtain the propagation parameters. It transpires that at the expense of a 3 per cent increase in the attenuation of the H_{01} mode the attenuation of E and H-modes (other than H_{on}) has been raised by a factor of 20. Simultaneously because of the 140-fold increase in surface reactance (X_z) the phase velocities of E-waves and higher order H-waves have been reduced.

(2) Ring Structures

A periodic array of disks (Fig. 5) of different surface property, or any periodic cylindrical structure (e.g. corrugated waveguide) is—if proportions are chosen suitably—a promising structure. The guiding properties of such structures can be deduced[2,4,5] using formulae (10), (11) and (12) provided the surface impedance components are known. If the period of the structure is small in comparison with the wavelength (say a factor of 10) then the surface impedance components can be deduced as follows.

Let Z_1 and Z_2 be the surface impedances of the elements (thickness t_1 and t_2 respectively) of the periodic structure then the effective surface impedance components of the entire waveguide are given by

$$Z_z = \frac{Z_1 t_1 + Z_2 t_2}{t_1 + t_2}$$
$$Z_\phi = \frac{Z_1 Z_2 (t_1 + t_2)}{Z_2 t_1 + Z_1 t_2} \quad \ldots\ldots\ldots\ldots\ldots \quad (13)$$

In most applications it is desirable to choose $Z_2 = Z_1$ and $t_2 < t_1$. Under such conditions this reduces to

$$Z_z \simeq Z_2 \frac{t_1}{t_1 + t_2}$$
$$Z_\phi \simeq Z_1 \frac{t_1 + t_2}{t_1} \quad \ldots\ldots\ldots\ldots\ldots \quad (14)$$

An estimate of the value of the quantities Z_1 and Z_2 can be made by considering the elements of the structure. Thus, for the important structure shown in Fig. 6, the quantity Z_1 is the surface impedance of the metal rings while the quantity Z_2 can be taken equal to the input impedance of a transmission line length l (the depth of the rings) terminated in the impedance R. For such a waveguide the surface impedance components are, therefore, given by

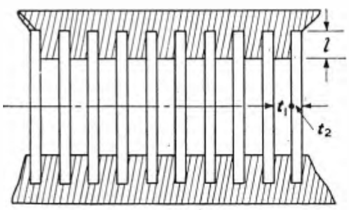

Fig. 6. Corrugated waveguide

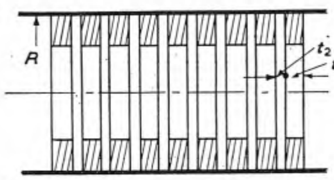

Fig. 7. Ring structure with an absorbing layer

$$Z_\varphi = Z_m \frac{t_1 + t_2}{t_1}$$
$$Z_z = R \frac{t_2}{t_1 + t_2} + j\,2\pi\,(l/\lambda_0) \frac{t_2}{t_1 + t_2}$$

where it has been assumed that Z_m/R and l/λ_0 are small quantities.

Here again as with dielectric coated waveguides the surface impedance components can, within wide limits, be adjusted to any desired value by choosing suitable material constants and waveguide proportions.

(3) Helical Waveguide

A ring structure although satisfactory from a theoretical point of view is not easy to construct and for this reason helical waveguides are preferred. A helical waveguide consists essentially of a helix of conducting wire supported in a suitable structure (Fig. 7.) The performance of a helical waveguide[5], provided the helix angle is small, is similar to that of a ring waveguide. To a first approximation the surface impedance of a helical waveguide (Fig. 8) with an absorbing layer, for example, is the same as that of a ring structure with an absorbing layer (Fig. 6) where $l = t =$ the wire diameter and $t_2 =$ space between the wires. The equations (10) to (12), (14) and (15) are applicable. To a better order of approximation allowance can be made for the lay angle of the helix and for the fact that the helix is made of a round wire rather than rectangular[6,7].

As an example consider a helical waveguide with the following constants

$$t_1 = l = \lambda_0/10,\ t_2 = \tfrac{1}{2} t_1,\ R_m = 10^{-4}$$

(approximately intrinsic impedance of copper), $R = 10^{-2}$ (approximately 4Ω). Neglecting the lay angle of the helix the surface impedance components become

$$Z_z = 3 \times 10^{-3} + j\,2{\cdot}1 \times 10^{-2}$$
$$Z_\phi = 1{\cdot}5\,(1 + j) \times 10^{-4}$$

while if it were a plain metallic waveguide its surface impedance components would be

$$Z_z = Z_\varphi = Z_s = Z_m = 10^{-4}\,(1 + j).$$

Consequently at the expense of a 50 per cent increase in attenuation of the H_{01} mode the attenuation of E-modes and higher order H-modes is increased 30 times while the surface reactance is increased by a factor of 200, with a consequent change in the phase velocities of the modes concerned.

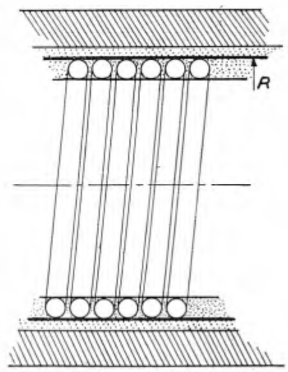

Fig. 8. Helical waveguide with an attenuating layer (round wire)

Special Waveguides, their Uses and Limitations

The reasons for requiring special waveguides for long distance transmission are numerous, but broadly speaking, these waveguides are necessary to limit the mode reconversion process to a tolerable level. The process of mode conversion as discussed previously cannot be prevented since it is related to the waveguide tolerances, which for technological reasons it is impracticable to improve beyond a certain limit (say 0·001in on the diameter and a fraction of an inch on straightness). Besides, the process of mode conversion is, in itself not a particularly harmful phenomenon, because it only leads to a slight increase in attenuation, which is substantially frequency independent. The combined phenomenon of mode conversion-reconversion, however, is frequency dependent and must be avoided at all costs, if wide bandwidth handling capacity is to be preserved. Special waveguides will effectively suppress mode conversion-reconversion provided the surface impedance of the waveguide is chosen correctly.

The alternative solution, to insert mode filters at frequent intervals in a plain metallic waveguide run, is undesirable for a number of reasons. In the first place mode filters cannot be perfect and even the best types have attenuation for some of the mode less than 1dB/ft and, therefore, for effective functioning mode filters about 10ft long would have to be used. Also, successful mode filters are of the helical waveguide construction and at the junction to the plain metallic waveguide there is a considerable mismatch for modes other than H_{0n} and it is therefore necessary to employ special tapered transitions which further increase the length and cost of the filters. Furthermore, the waveguide sections between the filters must be made uneven (preferably random) to avoid extremely harmful resonances. For these reasons the proportion in length of the mode filters to the remainder of the plain metallic pipe is 1:4 or more, and therefore the saving as compared with an all helix construction is not likely to be significant, particularly since jointing forms a substantial proportion of the cost of the whole scheme. In addition the waveguide-mode-filter assembly is considerably inferior to an all helix construction in negotiating bends.

The signal distortion due to mode conversion-reconversion is a function of the quantity

$$N = \left[\frac{c}{(\beta_n - \beta_0)l}\right]\frac{1}{\alpha_n - \alpha_0} \quad \ldots\ldots \quad (16)$$

where c^2 is the mean square value of the coupling coefficient between the H_{01} wave (having attenuation α_0 and phase change coefficient β_0) and a parasite mode (having attenuation α_n and phase change coefficient β_n). l is a typical distance between the waveguide irregularities responsible for coupling between the modes.

The smaller the quantity N the smaller will be the mode conversion-reconversion effect. Consequently the larger the difference $(\beta_n - \beta_0)$—i.e. the difference between the wavelengths of the modes concerned—and the larger the difference $(\alpha_n - \alpha_0)$ the smaller will be the signal distortion.

It transpires that for typical waveguide assemblies the most harmful effects result from mode conversion-reconversion due to small kinks (changes in waveguide direction). The modes produced by such irregularities are E_{11} and H_{1n} modes. In plain metallic waveguides it is the E_{11} and H_{12} modes that are particularly troublesome: for the H_{12} mode the quantities $(\beta_n - \beta_0)$ and $(\alpha_n - \alpha_0)$ are small and for the E_{11} mode although $(\alpha_n - \alpha_0)$ is not a particularly small quantity $(\beta_n - \beta_0)$ is extremely small.

With special waveguides, however, the quantity $(\alpha_n - \alpha_0)$ for the H_{12} mode can be easily raised 10-fold or more by making R_z larger than R_m by this factor. In a similar manner by increasing the surface reactance, X_z, by a suitable factor, the quantity $(\beta_n - \beta_0)$ may be increased in proportion (see numerical examples given for the dielectric coated and the helical waveguides).

The quantity $(\beta_n - \beta_0) = 2\pi/\lambda_n - 2\pi/\lambda_0$ where λ_n is the guide wavelength of the interfering mode and λ_0 is the guide wavelength of the H_{01} mode, may be put in the form

$$(\beta_n - \beta_0) = 2\pi/\lambda_0 \frac{\lambda_0 - \lambda_n}{\lambda_n} = 2\pi/\lambda_s$$

where $\lambda_s = \lambda_0 \dfrac{\lambda_n}{\lambda_0 - \lambda_n}$

which is usually very much larger than λ_0. The quantity λ_s is an extremely important parameter for it can be shown that at frequencies for which λ_s is of the order of the periodicities of the waveguide irregularities, considerable signal distortion takes place. For this reason waveguide joints or any other likely sources of irregularities must be placed at random. For the same reason periodically spaced mode filters are a likely source of trouble.

If all irregularities are distributed at random then the only manifestation of mode conversion-reconversion is a kind of crosstalk noise whose level is proportional to

$$c^2(\beta_n - \beta_0)^{-2}(\alpha_n - \alpha_0)^{-1}$$

and whose intensity increases with distance, eventually setting a limit to the maximum repeater spacing. With practical waveguide lines the irregularities are not distributed at random but, due to a number of causes some correlation in the waveguide irregularities is bound to exist resulting in irregular $\alpha(\omega)$ and $\beta(\omega)$ characteristics with a consequent signal distortion. The level of signal distortion is a function of the quantity N and provided that by suitable waveguide design the quantity N can be kept small the effects discussed will have a tolerable magnitude for distances of the order of 20 miles.

In conclusion it should be stated that a substantial amount of surface reactance X_z is necessary if the waveguide is to negotiate bends successfully. Thus, a plain metallic waveguide cannot be made to negotiate bends having radii smaller than several miles or substantial increase in attenuation and signal distortion will take place. In contrast a dielectric coated or a helical waveguide having its surface reactance X_z enhanced say 1 000-fold can be made to follow bends having radii of the order of a few hundred feet with little effect on the waveguide performance.

Although mode conversion-reconversion can be largely combated by using suitably designed waveguides the delay distortion cannot be successfully compensated by modification of the waveguide construction.

For an overmoded waveguide, the delay distortion $(\delta\phi)$ and the bandwidth f are related approximately by[3]

$$f = 19/F\,(f_0/l)^{1/2}\,(\delta\phi)^{1/2} \quad \ldots\ldots\ldots\ldots\ldots \quad (17)$$

where f is the h.f. bandwidth in Mc/s

F = cut-off frequency/carrier frequency

l = the length of the waveguide in km

f_0 = carrier frequency in kMc/s

$\delta\phi$ = the delay distortion, representing the permissible delay* (measured in radians) between the side-band and the carrier.

Any equalization of the waveguide characteristics is best carried out at the terminal equipment. But, the necessary amount of equalization is a function of a number of factors: permissible signal-to-noise ratio, particular type of coding, methods of demodulation etc.

Because of the two phenomena: mode conversion-reconversion and delay distortion, simple methods of modulation, e.g. a.m., are unsuitable for long distance transmission by waveguide. In fact it is necessary to use equalizers to divide the messages into suitable bands, to code the information in p.c.m., regenerate at suitable intervals and employ waveguides of special design for the entire installation. But, provided these measures are taken, a communication system of transmission capacity unequalled by any existing systems will result.

Acknowledgments

Acknowledgment is made to Standard Telecommunication Laboratories Ltd for permission to publish this article. The author is indebted to the Institution of Electrical Engineers, London, for permission to draw on the material in his recent paper[8].

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* *E.g., if $\delta\phi = \pi/2$ an inversion of sidebands takes place; a condition of severe distortion.*

Millimicrosecond Digital Computer Logic

By Prof. N. F. Moody*, A.M.I.E.E. and R. G. Harrison†, B.A.(Cantab)

A system of fast pulse logic is described which combines the efficiency of transformer coupled stages with digit delay tolerances approaching that of d.c. coupled systems.

Logical circuits for OR, AND, INVERTOR *and* RECLOCK *are shown, together with a driver which permits a 'fan out' factor of 5. Transistor circuits are used throughout.*

(Voir page 571 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 576)

THE advent of drift transistors with power gain at 300Mc/s and above has stimulated the design of binary logic circuits with digit intervals less than 100mμsec. Transmission delays become a significant factor at these speeds and the circuit design must be directed to tolerate them. The ability to 'fan out' from one logical unit to many is an important factor, and may become a key criterion where parallel methods are employed. As always, the design compromise must seek to minimize power consumption and circuit complexity.

Direct coupled non-saturating circuits can be fast and yet tolerant to transmission delays. However their 'fan out' factor is critically dependent on maximum transistor dissipation, and the power consumption is high.

Saturated direct coupled circuits reduce the circuit complexity and power consumption but introduce a severe speed loss.

An important class of a.c. coupled circuits uses transformers. By this means the effective pulse current amplification of a transistor and its associated coupling network can be enhanced and, equally important, the isolation provided by the secondary winding permits the localization of pulse circulating currents. Such currents can inject spurious signals across common 'earth' impedances—often a source of difficulty in high speed systems. In the interest of simplicity such systems, including the one described, use transformers of time-constant shorter than the digit interval. Then the transformer does not remember its history, but a digit is necessarily restricted to a pulse short on the digit interval time scale. A system using this method[1] shows the poor tolerance to transmission delays which tends to result.

The System Outline

The system to be described seeks to exploit the impedance matching capabilities of transformers to enhance current amplification; the polarity reversal which is possible; the localization of earth currents; and the low secondary impedances which exist outside the transformer pass-band. By means of an artifice this system is made tolerant to digit delays and, further, transistors may be saturated without significant loss in speed. The compromise described promises low power consumption, moderate circuit complexity, good 'fanning out', and a straightforward extension to large logical systems. Existing logical units will handle $1{\cdot}2 \times 10^7$ digits a second, and a figure of $2{\cdot}0 \times 10^7$ appears within reach. The speed of a logical unit, it will be seen, is one half the potential data handling speed of the system, and hence $4{\cdot}0 \times 10^7$ digits a second should be possible.

Consider a logical element, whose digit interval, T, is defined in terms of clock pulses C_A, C_A' (Fig. 1 (a)), and which is operated by pulses D_A which are short compared to T. It is a feature of the system that transmission delays are permitted such that any D_A may lie between the bounds C_A and C_B ($=T/2$)*. There is a forbidden time zone in each digit period from C_B to C_A' where D_A pulses must not lie: a period used for recovery processes in the logical unit. The time $C_B - C_A'$ is not lost however. A sufficient succession of logical operations on pulses arising in the A period would cause the resultant output to lie in the forbidden region for A pulses. Before the delays have been accumulated to this value, however, the logical chain is clocked by a C_B pulse. The resultant D_B pulse, (Fig. 1(b)) then

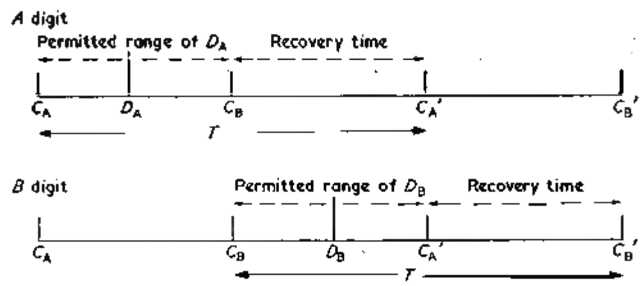

Fig. 1. System timing

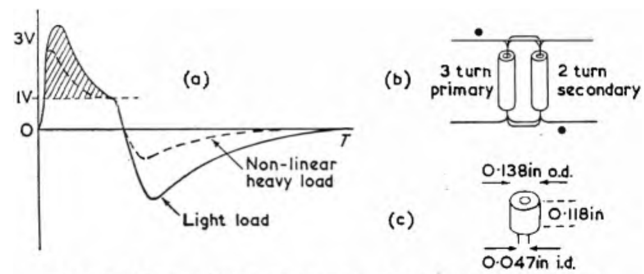

Fig. 2. The coupling transformers and their output waveshapes
(a) Typical digit or clock pulses
(b) Transformer construction
(c) Transformer core dimensions
Material: Ferroxcube Corp. of America 4B ferrite 225T118 E1-4B

utilizes the period C_B to C_A' and the period $C_A'C_B$ becomes forbidden to the B pulses. Logical operations can be performed on A pulses, or on B pulses, but an A pulse can be compared with a B pulse only after having been clocked to a B interval. Delayed B pulses may, in turn, be reclocked to A pulses, and then a time delay T equal to the digit handling time of a logical unit has resulted from the double clocking.

The two binary states of a digit are represented by the presence or absence of a D pulse. In the present system each C or D pulse appears at the output of a transformer, which may have several secondary windings, and deliver either polarity. The level is 2 to 4V and will feed a 50Ω transmission line or several (up to 5) logical circuits via unterminated lines. Fig. 2(a), which shows the typical waveshapes,

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† Defence Research Board of Canada.

* *The concepts given here are those of an ideal system in which the pulses are of zero duration. In a practical system, it will be seen, some timing latitude is lost.*

indicates that the 'forbidden' periods are required in part for transformer overshoot decay. Only the shaded part of the waveform is transmitted as intelligence, so that disturbing signals of less than 30 per cent amplitude are ignored*. Transformer construction is seen to be very simple from Figs. 2(b) and 2(c).

Logical Circuits

The Driver

The various techniques employed are best introduced by means of selected circuits. These will all be based on the use of Philco 2N501 transistors and a value of $T = 80\text{m}\mu\text{s}$. The discussion will commence by describing the driver stage, a unit whose sole purpose is to permit fan out. With reference to Fig. 3(a), S_1 represents the output transformer

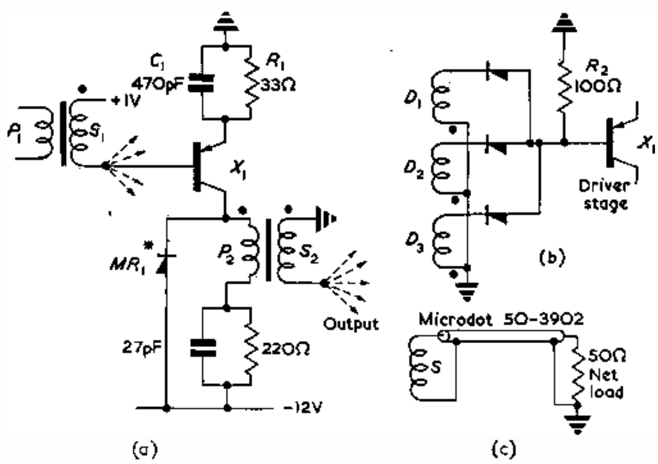


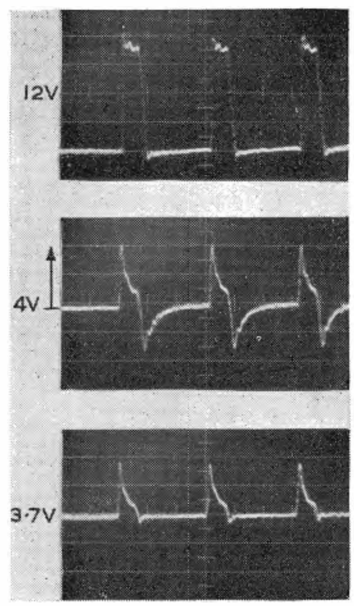
Fig. 3. (a) Driver
(b) OR gate
(c) Transformer feeding terminated cable showing correct earthing
Note: Transistors Philco 2N501
** Transistron S570G*
Diodes not asterisked Qutronics Q5-100

secondary of any logical unit, here shown as capable of supplying digits to 5 stages (not all units will supply this number). With the transformer secondary arranged to apply the digit pulse as an initially negative going waveform, the transistor is switched to a heavy initial surge current which is rapidly limited to 50mA by saturation. As the transformer secondary S_1 overshoots, the transistor is rapidly cleared of carriers, and a pulse such as that of Fig. 2(a) appears at S_2. Amplitude limiting is provided by transistor saturation, and a degree of pulse shaping by the networks in emitter and collector. The diode MR_1 is included to protect the transistor collector from excess voltage. Waveform photographs of a driver are shown in Fig. 4, and other performance figures for the various logical units are to be found in Table 1.

* *In practice parasitic circuit elements usually superimpose some high frequency 'ringing'. The system is very tolerant to such distortion of waveshape.*

The 'or' Gate

Assuming, for the purposes of circuit description, that a binary 1 is represented by the presence of a pulse, then the OR function arises simply by connecting the transformer secondaries of several logical units together via diodes to the input terminal (base of X_1) of a driver as shown in Fig. 3(b). The diodes prevent interaction between the various secondaries, and this is necessary since the same secondaries may be required to feed other logical units.



Transistor collector

S_2, Output transformer secondary. Load is diode and 100Ω resistance in series.

Output waveform. Load is diode and 100Ω resistance in series. Waveform developed across 100Ω only approximates that of current to transistor base.

Fig. 4. Driver waveforms
23mμsec per division

The diodes would prevent clearance of carriers from the driver transistor and therefore a resistor, R_2 of Fig. 3(b), is added. This resistor, in conjunction with the emitter network R_1C_1 of the driver, provides the clearance mechanism. The OR configuration may be applied to units other than a driver.

Transmission delays may cause the digits applied to an OR circuit to be displayed with respect to each other by a time theoretically limited to $T/2$. Then the subsequent circuit can be pulsed several times per digit. This does not disturb the logic, but all circuits are protected against excess dissipation from this cause by means of the network in the collector of X_1 (see Fig. 3(a)), which takes

TABLE 1

Data on Units Designed for $T \sim 80$ mμsec[1]

UNIT	PULSE RISE[2] TIME (mμsec)	TRANSMISSION[3] DELAY (mμsec)	PULSE WIDTH[4] (mμsec)	DELAY TOLERATION (mμsec)	FAN OUT[5] FACTOR	AVERAGE POWER[6] CONSUMPTION (mW)
DRIVER	2	2	20-25	No limit	5	100
OR	As input	~0	—	No limit	1	Nil
AND (4 input)	6	10	10-25	35	2-3	300
INVERTOR	3·5	$T/2$	~14	$(T/2)$- 7	2	250

NOTE: *Any unit will drive, or may be driven from, any other unit.*
(1) Based on the use of Philco 2N501 transistors and Qutronics Q5-100 diodes
(2) 10-90 per cent
(3) Point on waveform which is the actuating level.
(4) and (5) Somewhat load dependent.
(6) Based on equal numbers of zeros and digits.

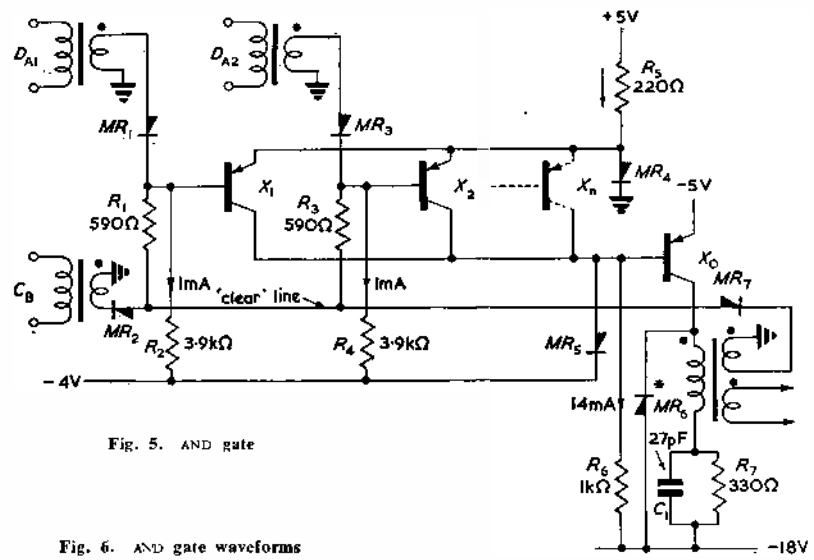
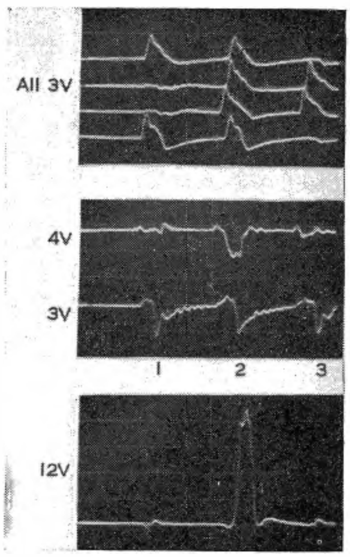


Fig. 5. AND gate

Fig. 6. AND gate waveforms
23mμsec per division



Bases of transistors X_1, X_2, X_3, X_4.

Collectors of transistors X_1—X_4.

Clearing line: Pulses 1 and 3 are clock pulses.
Pulse 2 is clock pulse and self-clear pulse superimposed.

Collector of transistor X_0.

a time of the order of $T/2$ to recover. In Fig. 3(c) one output is shown feeding a terminated transmission line, as may be necessary when inter-unit connexions exceed 6in. A transformer may feed only one terminated line which, in turn, should operate not more than two logical units.

Under favourable conditions each driver can fan five times, so that a fan out from 1 to 25 outputs requires six transistors.

The 'AND' Gate

For the AND function an artifice is required to achieve operation with narrow pulses whose relative delays may be up to the limit $T/2$ so that they do not necessarily overlap. By some means all input pulses should be extended to the $T/2$ boundary. When a pulse is transmitted via a diode to the base of a transistor charge is delivered to or removed from the transistor according to transformer polarity. Such trapped charge can set and maintain a transistor operating condition for times as long as $T/2$.

In the AND circuit shown in Fig. 5 pulse stretching is achieved by reverse charge storage in the emitter-collector transition and stray capacitances. Some 40pC is stored in this way. Thus in Fig. 5, n transistors, X_1 to X_n, ($n \leqslant 5$) are used, one for each input to be compared. These transistors share a common emitter current, injected by resistor R_5 and have a common collector load R_6. In the quiescent state all transistors conduct by virtue of current in each resistor, R_2, R_4 etc., the collectors are clamped below saturation level on diode MR_5, and transistor X_0 is cut off by about 1V. Since any one of the transistors X_1 to X_n can carry the total emitter current, and thus maintain diode MR_5 clamped, X_0 will conduct only when X_1 AND X_5 AND . . . X_n are cut off. This is the condition required for an output.

The pulses D_{A1} . . . D_{An} to be logically compared are delivered from reversed secondaries so that the initial swing is positive. On the arrival of each pulse, charge is trapped on the appropriate transistor base, which is raised some 1·5V, and that transistor is maintained in a cut off state. When all n pulses have been received, the emitter source current is diverted to diode MR_5, and the resistor R_6 draws 13mA from the base of X_0. The latter, which has the usual form of collector network, delivers an output pulse at the secondary of its transformer.

As described, the base current of X_0 can be maintained for some time, and the output pulse may become undesirably wide and slow in fall time. A second secondary is added to the output transformer to provide a feedback mechanism which limits pulse width. On excitation of X_0 this winding drives the 'clear' line negative, whereupon the resistors R_1, R_3 etc. bring the transistors X_1 to X_n into conduction after a short delay, and current in X_0 is shut off.

At digit intervals T of 80mμsec and less, it is quite possible for the storage of one or more of the transistors X_1 to X_n (no AND having occurred) to 'remember' an input and carry it over to the next digit period. This undesirable effect is prevented by injecting a positive clock pulse into the clear line, at the time $T/2$, from the winding C_B. Thus the memory is destroyed at time $T/2$.

The AND gate may be used to clock A pulses against a B clock. Then one logical input is made a C_B pulse. In this event the C_B pulse to the clear line must be delayed a few millimicroseconds, which may be done by passing it through a few feet of cable.

For digit periods $T < 60$mμsec the transition capacitances provide adequate reverse storage to allow proper circuit operation: for longer periods the capacitances should be augmented by additional base-emitter capacitance.

Waveforms of the AND unit are shown in Fig. 6.

The Invertor

The addition of an invertor, shown in Fig. 7, to the other circuits permits all logical operations to be performed. In this circuit the alternate clock and digit pulses reverse the stored charge at the base of X_1. In the absence of a digit therefore, X_1 is cut off at all times; but in the presence of a digit conducts during the time interval D_A to C_B. When D_A is absent, and X_1 is cut off, the circuit is required to deliver an output. With X_1 cut off diode MR_1 is in conduction and the transistor X_2 is biased off by one volt. Then the C_B pulses applied to the emitter, which are 2 to 4V amplitude, can switch X_2 thus delivering output from the usual collector circuit.

In the presence of a digit at the input, X_1 is conductive and X_2 is biased off by 5V. There is then no output as a result of the emitter clock pulse.

It might be supposed that the clock pulse to X_1 should be delayed with respect to the clock pulse to X_2. In practice it has been found that the transmission delay in X_1 proves sufficient for the purpose. Waveforms* of the invertor are shown in Fig. 8.

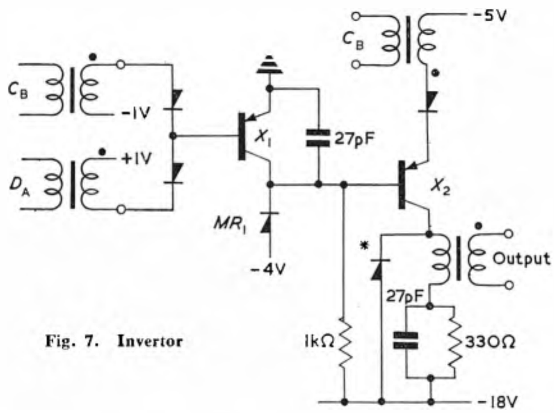


Fig. 7. Invertor

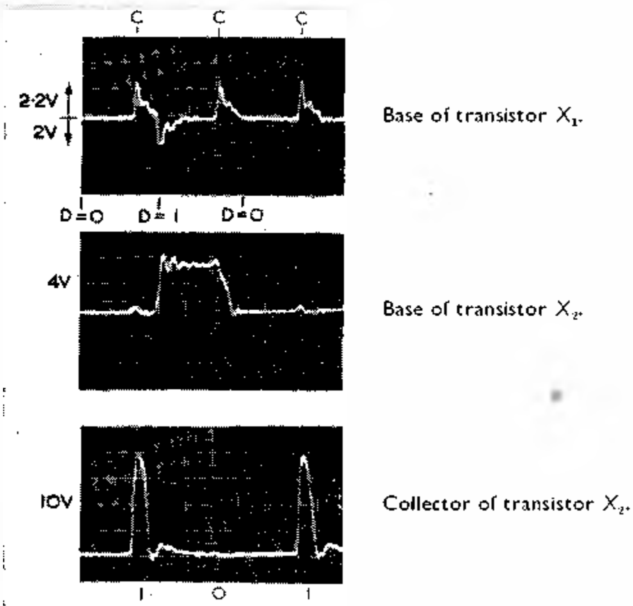


Fig. 8. Invertor waveforms
23mμsec per division

The basic invertor circuit may also be used for clocking without inversion. To do this the D_A pulse is applied instead to the C_B input transformer, and the C_A pulse is applied instead to the D_A transformer. When the circuit is used in this way the clock applied to the base of X_1 must be delayed for a few milliseconds with respect to the clock applied to the emitter of X_2.

Conclusions

A photograph of some experimental units is shown in Fig. 9. They are constructed on boards of the type used for printed wiring, in which the copper is left on one side to provide a good ground plane. Where unit inter-connexions exceed an inch or so, coaxial cable is employed†. This cable (Microdot 50-3902, $Z_0 = 50\Omega$) which is used generally throughout the system, is easy to handle and fits in with the miniature construction since its diameter is about 1/16in.

It is good practice to earth a cable only at the unit it is driving (see Fig. 3(a)); and if several units, not in proximity, are being driven from a single transformer, a separate secondary should be provided for each cable.

H.F. interaction between supply lines is prevented by slipping transformer core beads (Ferroxcube 3B material) on the leads of each logical unit, with ceramic by-pass capacitors to provide decoupling.

It must be emphasized that the system described is experimental and has been tested only on very small scale assemblies. Component values have not been fully optimized and a variety of voltage supply lines have been tolerated, some of which can be eliminated. Any logical unit may be called upon to deliver current to widely differing non-linear loads, which will modify output pulse shape. An attempt has been made to design the circuits so that any combination may be coupled, but reclocking may be required in certain instances solely to regain pulse shape.

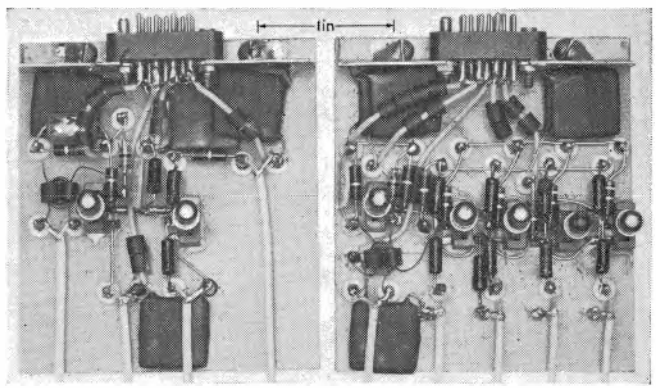


Fig. 9. Invertor (left), AND gate (right)

By use of the principles outlined it will be found simple to develop special purpose units such as EXCLUSIVE OR in a form more economical than is provided by assemblies of the elementary units. In the interests of space these are not shown here.

In the foregoing discussion it has been assumed that the presence of a digit represents a binary one, and the performance figures given in Table 1 are based on this mode of operation. If, instead, the absence of a digit is made to represent binary one, the functions of the AND and OR gates are interchanged. Due to the relative complexities of the two types of gate, a given logical design may require less components when the alternate convention is employed.

Recently some of the circuits have been built using the Motorola 2N695 transistor. With the circuit time-constants appropriately reduced it appears that a value of $T = 40$mμsec is feasible.

Acknowledgment

The work described is published by permission of the Defence Research Board of Canada, in whose laboratories it was undertaken.

** It is not usual to clock the digit waveform with its own phase of clock as shown in Fig. 8. However substitution of the alternate phase of clock causes no alteration to the waveforms other than a change in timing.*

† Cables of less than 6in length are left unterminated.

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A TRANSISTOR STIMULATOR FOR PHYSIOLOGICAL USE

By R. E. George*

A stimulator is described for measurement of the refractory period which follows excitation in a nerve. It produces pairs of rectangular pulses separated by an adjustable time interval. The duration of both pulses can be varied independently.

The maximum amplitude available is 20V from an impedance not exceeding 2kΩ. The pulse durations are variable in steps from 30μsec to 30msec. The sequence repetition rate may be varied from once in 3sec to 330 times/sec.

(Voir page 571 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 576)

THE application of transistor circuit techniques to the problem of electro-physiological stimulation, makes it possible to fulfil the long standing requirement for a comparatively isolated stimulus from a portable apparatus capable of operation for long periods on an internal battery.

Some of the possibilities have already been exploited in a circuit[1] producing a pulse of fixed duration, designed about a single transistor, having facilities for variation of pulse amplitude and repetition time.

When an impulse greater than the threshold is applied via electrodes to a nerve, an action potential is propagated along it. A propagated impulse can be detected with a second pair of electrodes connected to an amplifier and oscilloscope (Fig. 1). A second identical stimulus cannot be propagated if the time lapse between the stimuli is less

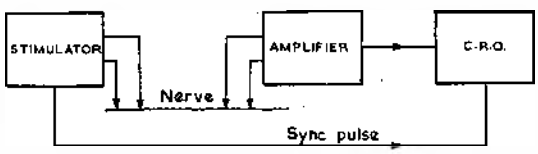

Fig. 1. Experimental arrangement

than a critical time, called the refractory period, which is commonly of the order of a few milliseconds. Since a nerve trunk contains more than one type of nerve fibre each having its own stimulation threshold, the experiment may be elaborated by selective stimulation of one of the types and then by application, within the refractory period, of a second stimulus of different amplitude and duration, to obtain a response originating from another type of fibre not excited by the first stimulus.

The specification of pulse widths and output impedance of the stimulator to be described follows as far as is practicable that previously adopted for electro-diagnostic stimulators[2]. The main difference is that the maximum voltages need not be so high in applications to muscle and nerve preparations, and normally need not exceed 20V. Even with small metal electrodes, the tissue resistance may be as low as 1kΩ with a shunt capacitance of 0·01μF. The low impedance negative going voltage pulse readily obtained from a pnp transistor satisfies the physiologist's requirement of constant voltage cathodic stimulation.

It is helpful to bear in mind the analogies between transistor circuits and the more familiar valve circuits; for such problems as sharpness of rise-time, trigger, threshold, change in d.c. level, and pulse amplification are considered as for valve circuits. The functional arrangement of the circuit (Fig. 5) is similar to that of the more familiar valve physiological stimulator[3].

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Basic Switching Circuit

The circuit comprises two identical pulse generators triggered from delay circuits timed from a common free-running initiator. The two pulses are available either at separate terminations or combined at one of them (Fig. 2).

Two transistors X_A and X_B (Fig. 3) with emitters connec-

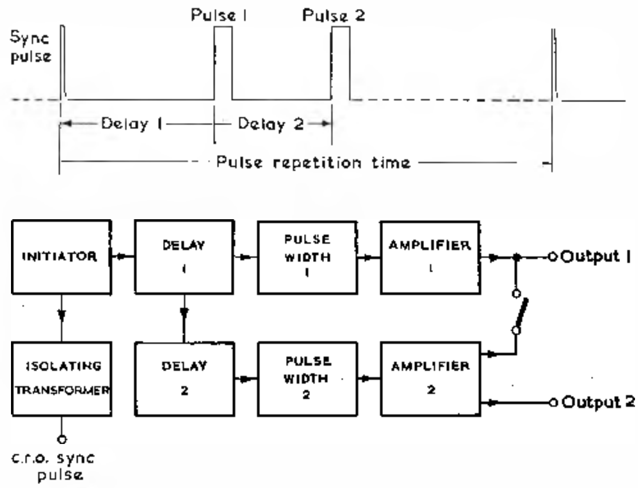


Fig. 2. Basic switching arrangement and pulse diagram

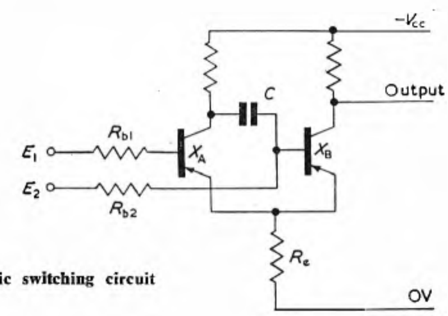


Fig. 3. Basic switching circuit

ted to a common load R_e comprise a multivibrator having the single capacitor C. The mode of operation is substantially the same as if valves were used but with an important difference concerned with the leakage and bias currents which affect the switching times. By making the base return potentials, E_1 and E_2 equal, the circuit becomes a free running oscillator, the capacitor charging and discharging at different rates. While the collector of X_A is in its more positive resting state and X_B is cut off, the discharge of C is controlled mainly by R_{b2}. When the collector of X_A is at $-V_{cc}$ and the base of X_B is negative, C charges at a rate approximately determined by R_{b2} in parallel with the input resistance of X_B, which is approximately $R_e/1-\alpha$.

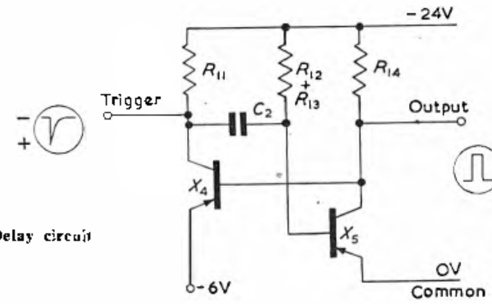


Fig. 4. Delay circuit

Throughout the charge and discharge cycles, the small steady collector leakage current I_{co2} flows through R_{b2}, offsetting the base return potential E_2 by $I_{co2} \times R_{b2}$. On the discharge cycle, modified leakage currents from the emitter and collector of X_B flow through the base, giving the circuit different base return potentials in alternate states.

If $E_2 > E_1$, the circuit has one stable state, the triggering threshold being $(E_2 - E_1)$. A positive going trigger pulse may be applied at either the collector of X_A or the base of X_B.

In the interests of thermal stability, the leakage current of each junction should not exceed 2 or 3μA at 22°C. This applies especially to X_B, for a 1μA change in I_{co} will produce a 0·1V change across the 100kΩ base resistor R_{b2}, changing the circuit firing threshold by this amount.

Small-signal amplifying transistors with an f_α of 0·5Mc/s are suited to switching applications where a rise-time of 3μsec is acceptable. Care should be taken not to exceed the manufacturer's rating for V_{cb}, or alternatively, transistors should be selected for high turnover potential. For the pulse generating circuits, an emitter-base turnover potential exceeding 6V is essential; while the second transistor in the delay circuit (Fig. 4) requires a turnover not less than 18V. Although the collector-base turnover potential may greatly exceed 24V, some samples have been found to exhibit a slow increase in the leakage current I_{co}. Since this leads to a change in the working points of the switching circuits, care must be taken to choose a transistor without 'creep'[4] in I_{co}.

Initiator

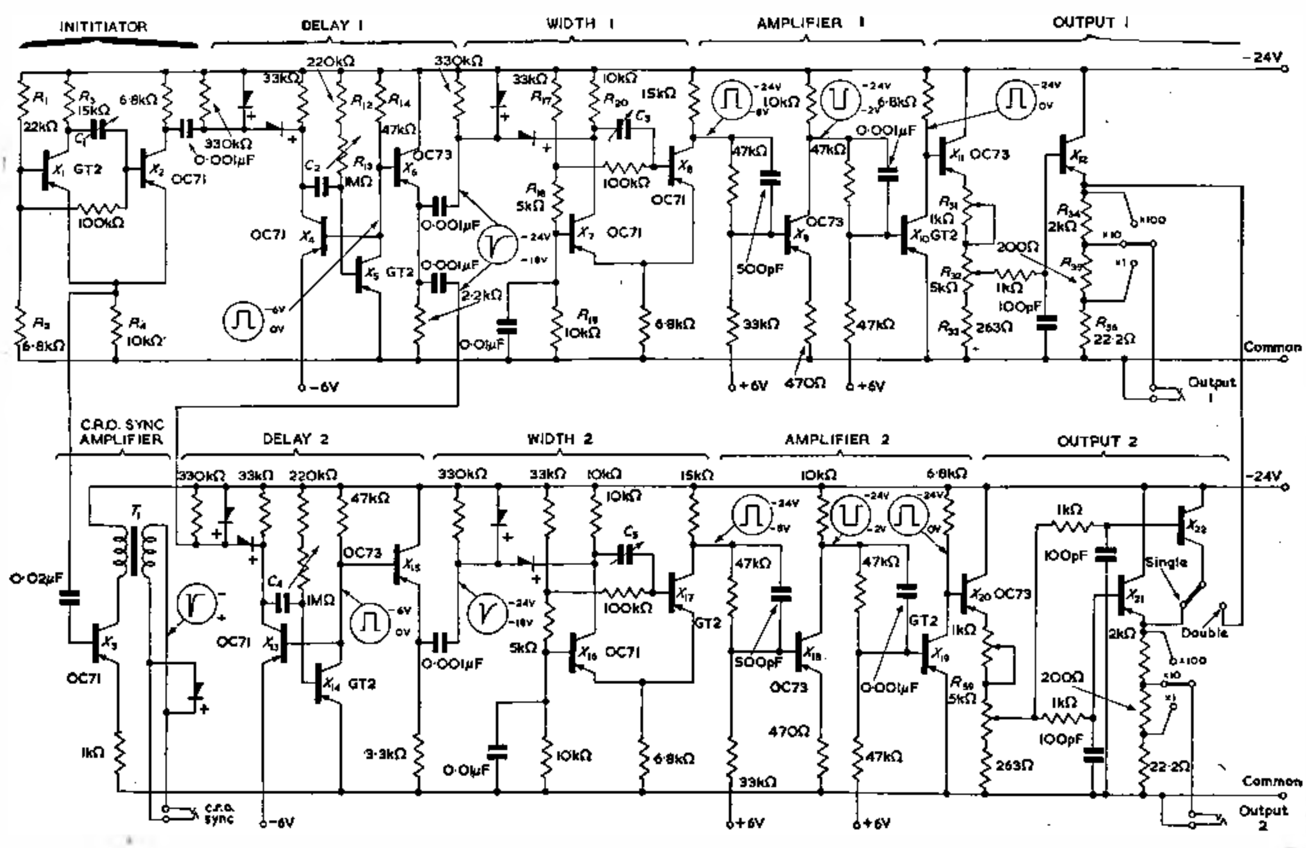
When E_2 is equal to E_1 in the basic switching circuit, the multivibrator becomes free running and is used to produce the master waveform from which the other parts of the circuit are timed. This unstable condition is obtained by returning both base bias paths to the same point on the potential divider R_1, R_2 (Fig. 5). When X_1 is 'on', the collector current flowing (ignoring base current) is roughly given by $-V_{cc} \times R_2/[R_1 + R_2] \times 1/R_4$ provided R_3 is small enough to prevent bottoming with this current.

The transition to the other state occurs when C_1 has discharged sufficiently to allow X_2 to start passing increased emitter current, a corresponding amount being lost by X_1. As V_{cb1} remains relatively large, junction capacitance is small and a rapid transition follows, the base of X_2 being driven negative. The emitter follows the base potential, cutting off X_1. The second transition may be kept as well defined as the first, by ensuring that X_2 collector does not bottom.

Delay Circuit

To produce a continuously adjustable delay time, a different type of circuit is used (Fig. 4). Considering the first delay circuit (X_4 and X_5 of Fig. 5) in the resting state, X_5 remains bottomed provided $\alpha/1-\alpha > [R_{12}+R_{13}]/R_{14}$. When a positive impulse is applied to the base of X_5 via capacitor C_2, cut-off occurs, the collector potential rising (in the case of pnp transistors, going negative) until caught at -6V by

Fig. 5. Complete circuit of stimulator



the base of X_4, which now begins to conduct. The action becoming regenerative, holds X_5 in the cut-off condition, while C_2 discharges through $R_{12} + R_{13}$ and also through the base with a leakage current, which has been found by measurement to have a value intermediate between I_{co} and $I_{co} + I_{eo}$. The base of X_5 is driven to +18V at the beginning of the exponential decay. When this has fallen to zero, X_5 begins to conduct, cutting-off X_4, the action again becoming regenerative via C_2, driving X_5 well into the 'on' and stable state. The X_4, X_5 combination produces a steep sided negative going rectangular pulse of duration governed mainly by C_2, with $R_{12} + R_{13}$, which can be adjusted continuously in the ratio of 5 to 1 giving adequate overlap on the switched settings of C_2.

The emitter-follower X_6, feeds two separate differentiating and diode clipping networks identical to the one prior to the first delay, each selecting the positive going trailing edge of the pulse to trigger both the first pulse generator and the second delay.

Pulse Generator

This circuit (X_7 and X_8 in Fig. 5) differs from the initiator in one important detail. The base return to the second transistor X_8, is taken to a higher return point on the bias divider R_{17-19}, allowing X_7 to remain cut-off while the circuit is in the resting state. A positive going impulse applied to the collector of X_7, if it exceeds the difference between the base potentials, will bring X_7 into the 'on' state, when the action becomes regenerative. The charge applied to C_3 depends on the voltage swing developed across R_{20}. Reliable operation can be obtained without X_7 or X_8 bottoming. The potential across R_{18} sets the trigger threshold, while both R_{18} and R_{20} affect the pulse width. Between pulses the output level is −8V.

Pulse Amplifier

Two grounded emitter amplifiers in cascade provide a pulse at a constant amplitude of 24V and at low impedance (4·7kΩ) for driving the output circuit. The second stage of pulse amplification (X_{10}) is alternately 'bottomed' and cut off by the pulse applied to its base. In this way the pulse excursion is made to depend only on the supply voltage. The base returns are taken to a +6V line so that switching may be effected from the 'off' to the 'on' state in a direct coupled circuit. To ensure a rapid transition between the states, small coupling capacitors have been added.

Output Circuit

The first emitter-follower X_{11}, applies almost the whole of the −24V supply to the potentiometer network R_{31-33} in the form of negative going pulses, from which a proportion, variable over a range of twenty to one, is taken to drive the output emitter-follower X_{12}. This double emitter-follower arrangement keeps the impedance at the top of the output potential divider R_{34-36}, substantially constant, at about 100Ω, over the whole range of settings of R_{32}. Allowance for changes in battery voltage is made by adjustment of R_{31}, so as to obtain −20V at the output on maximum setting.

When the output is set to its maximum of −20V, the base-line error between pulses due to the residual current in the emitter-followers is −40mV or 0·2 per cent of the pulse height; while at the minimum setting of −1V, the base-line is −12mV up, the proportion having risen to 1·2 per cent.

Using transistors with an f_α of 0·5Mc/s, a pulse rise-time of 2 to 3μsec is obtained from the emitter-follower X_{11}, but observed at the output of X_{12} the rise-time is slightly longer. On resistive loading the waveform approaches the ideal; but with capacitive loading on the highest range, when direct connexion is made, considerable overshoot and ringing (Fig. 6(a)) were experienced on the pulse leading edge. This may be attributable to hole storage effects.

To minimize this overshoot, two methods were tried. Firstly, a phase retarding network (Fig. 6(b)) at the input of X_{12} (Fig. 5) significantly reduced it. The other alternative was to substitute another transistor with a higher f_α. Satisfactory results (Fig. 6(c)) were obtained by combining both methods. The 1kΩ resistor to the base also serves to limit the maximum current which can be drawn by the transistor in the event of accidental short-circuit on the maximum range, although damage may not be prevented if the overload is prolonged.

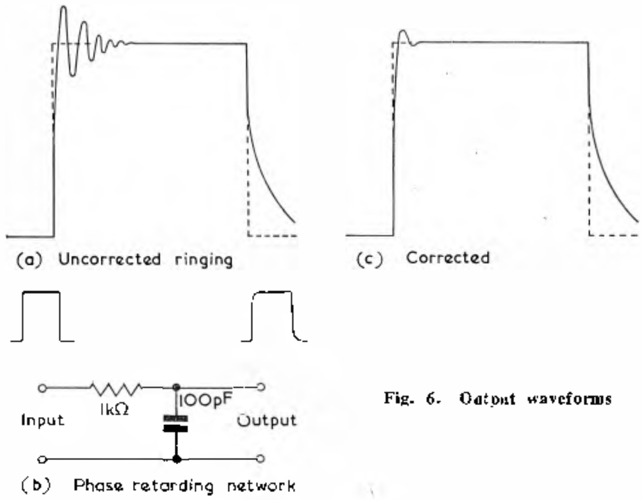


Fig. 6. Output waveforms

Adding Network

A single transistor X_{22} takes the output from the potentiometer R_{59} of channel two, and at the turn of a switch, applies it directly in shunt with the range network of channel one. The two trains of pulses are then superposed in time at a single termination. It should be noted that in this condition, as each pulse appears at output (1), via X_{12} or X_{23}, so will it appear between the emitter and base of the transistor that happens to be non-conducting. Due to the circuit arrangement, the limitation to the 'double pulse' condition is that the maximum ratio between the amplitudes of the two pulses, on any of the three ranges, is set by the potentiometer network of X_{11} and cannot exceed 20 to 1.

C.R.O. Synchronizing Impulse (Pre-Pulse)

A damped oscillatory circuit may be used for the production of short impulses[5]. Current impulses derived from the transient part of the waveform at the common emitter point of the initiator (X_1, X_2), and applied to the transformer primary, set the secondary into trains of damped oscillation at about 80kc/s. A diode shunted across the winding, removes everything except the initial positive going wave, which has a duration of 10μsec and a peak excursion of a little more than 20V, provided any additional damping resistance is not less than 200kΩ. These impulses are available at an insulated jack socket for c.r.o. triggering.

Ideally, a ring core would have been used, but for the sake of ease in winding, a Mullard LA3 Ferroxcube pot core assembly was chosen. 280 turns of 38 s.w.g. enamelled copper wire gave the primary winding an inductance of 11mH, while the secondary was wound from 660 turns of 42 s.w.g. A spacer of ⅛in Perspex was used to reduce the capacitance between the windings.

Isolation of Stimulus

To reduce stimulus artefact, it is very desirable to isolate the output pulse. This may be obtained in one of three ways.

(1) The stimulus itself can be transferred via a pulse transformer (Fig. 7(a)) having a fast transient response time to accommodate the edges of a rectangular pulse, and a long time-constant to avoid pulse droop. An instrument incorporating such a transformer passing a 100μsec rectangular pulse is in regular clinical use[6]. But, since the design must satisfy the contrary requirements of a high inductance winding with a low self-capacitance, the realization of a transformer to handle pulses of longer duration is doubtful.

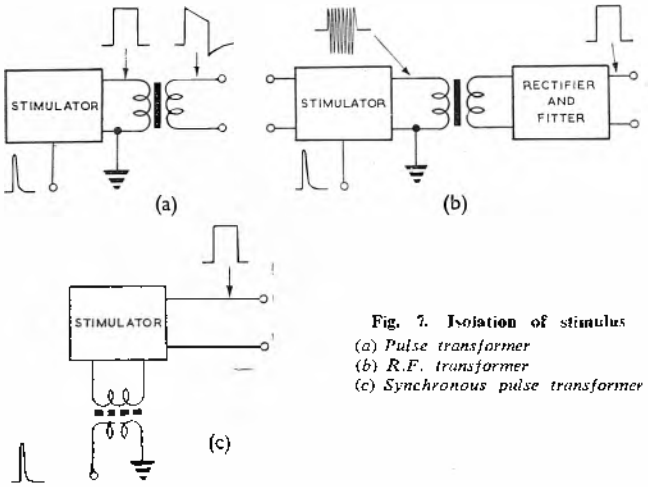


Fig. 7. Isolation of stimulus
(a) Pulse transformer
(b) R.F. transformer
(c) Synchronous pulse transformer

(2) The highest degree of isolation is obtained with the radio-frequency transformer[7], in which the stimulation waveform amplitude modulates a high frequency oscillation. A capacitance as small as 5pF between secondary winding and ground is claimed (Fig. 7(b)).

(3) The method used here was to 'float' the whole apparatus. The transformer in the connexion to the trigger circuit of the associated display c.r.o. discussed in the previous section allowed direct connexion to be made to the preparation. Isolation to the extent of 50pF was obtained (Fig. 7(c)).

Power Supply

The stimulator contains its own large layer type batteries designed specially for transistor equipment. The current consumption on the -24V line varies from 14 to 18mA, according to pulse times.

It was considered desirable to incorporate a pilot lamp as an aid to conserving battery life. A low consumption lamp (3·5V 0·15A) runs from a separate battery.

Provision is made for operation from an external 24V supply, when another identical pilot lamp fed via a series resistor R_{67} (Fig. 8) from the -24V line, comes into use, raising the consumption to around 160mA.

The $+6$V line, from which only 0·5mA is drawn, is obtained from a separate battery; and is switched on whether the -24V is supplied internally or externally.

Performance

The pulse times depend to some extent on supply voltage. In going from -24 to -20V, the following decreases were observed in pulse times; initiator 7 per cent, delay 2 per cent, width 3 per cent.

Pulse times also depend to some extent upon temperature; but this has not been found to be a serious disadvantage in this type of equipment, and in practice the pulse and delay times may be observed on the oscilloscope. In going from 22 to 30°C, the initiator showed no change, while the delay decreased by 7 per cent, and the width decreased by 9 per cent. Reliable operation has been obtained within the temperature range 16 to 30°C.

Redesign of the pulse circuits about the new pnp silicon transistors such as type OC200, should result in a higher order of stability, since the value of I_{co} is only about 0·01μA.

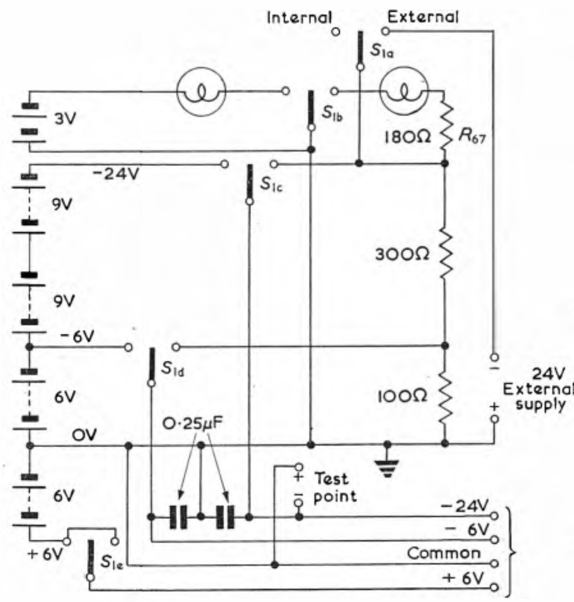


Fig. 8. Power supply switching arrangement

In the experimental arrangement to measure the refractory period, since only the high frequency components of the propagated nerve action potential are required, a low noise pre-amplifier with differential input transformer[8], in conjunction with a commercial oscilloscope is adequate to display the propagated impulse.

With this equipment, the stimulus artefact was comparable in size to the displayed action potential, serving as a convenient time reference.

The circuit described achieves the simplicity and compactness with economy of operation possible with transistor circuits. Apart from the transistors, there are few miniature components. The space saved by the use of small plug-in units, has been used to obtain a chassis layout with easy access to all parts of the circuit. The case measures 15¼ × 9 × 8in and the complete stimulator weighs 22¾lb.

Acknowledgment

Thanks are due to Mr. F. Huthwaite of the Physics Department, Guy's Hospital, for the construction and high standard of finish of the instrument; and to Dr. J. N. Hunt of the Physiology Department, Guy's Hospital, to whose specifications the instrument was designed.

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'AUTOLAND'

A Blind Landing System for Aircraft

A NEW automatic landing system has recently been announced by the Blind Landing Experimental Unit (B.L.E.U.) of the Royal Aircraft Establishment and units comprising the system have been displayed at the recent Flying Display and Exhibition at Farnborough.

Named 'Autoland', this system has been under development for some time past at the B.L.E.U. Bedford in conjunction with Murphy Radio, Smiths Aviation Division and Standard Telephones and Cables. To date some 4 000 landings have been made in perfect safety without damage to aircraft equipped with the system.

At the present time the more important airports of the world are equipped with automatic approach systems such as the Instrument Landing System (i.l.s.) and Ground Control Approach (g.c.a.) which together with an automatic pilot system enable an aircraft to approach the runway and start its final descent on instruments alone to the so called 'break off' point when the pilot assumes manual control and completes the landing operation.

Under conditions of good visibility the changeover from the instrument landing procedure to manual control until the final touch down presents no particular difficulty to the skilled pilot.

In poor visibility this final landing manoeuvre is particularly hazardous and there has arisen in consequence the need for a landing system which would enable a completely blind landing to be made under all conditions of wind and visibility.

The 'Autoland' system as now developed is one in which the aircraft is landed entirely automatically and its performance is such that under a wide variety of wind and visibility conditions the scatter of the touchdown point is less than half that obtained when the aircraft is landed manually in good visibility.

The system employs a modified Smiths S.E.P.2 civil autopilot into which guidance information is fed. A coupling unit computes the manoeuvre required to keep the aircraft on the desired flight path and the manoeuvre is executed by the autopilot servos operating the aircraft controls.

The landing is preceded by a standard automatic approach which is already in use in many civil airlines. The guidance information is derived from the internationally accepted Instrument Landing System (i.l.s.). Although this system is accurate enough for approach purposes, the tolerance allowed by the I.C.A.O. specification renders the i.l.s. unsuitable for the whole landing and two new guidance elements have to be substituted when the aircraft is close to the ground. These are Magnetic Leader Cable for azimuth guidance and a Radio Altimeter for height guidance.

Radio Guidance Equipment

The Magnetic Leader Cable installation developed by Murphy Radio consists, on the ground, of two cables situated about 250ft on either side of the runway centreline and extending about 5 000ft beyond the threshold in the undershoot area to as far along the runway as azimuth guidance is required during the ground roll. Each cable is fed from an alternator with a current at a different frequency, the current being kept constant by an electronic control circuit. The magnetic fields from these cables are detected in the aircraft by a small rotating loop, and their value is compared in a simple receiver. This allows position relative to the runway centreline to be determined to better than 5ft.

The Radio Altimeter is of the frequency modulated type. It is a variant of the Standard Telephones & Cables S.T.R.30, called S.T.R.30B1, specially modified by S.T & C. to a B.L.E.U. specification. It provides not only a very accurate measurement of height above ground, but also can detect a change of height of not more than 2ft and can therefore produce a signal accurately proportional to rate of change of height. These properties are used for guidance during the flare-out to control pitch attitude so as to make the rate of descent proportional to height, i.e. the lower the height the smaller the rate of descent.

The landing sequence is therefore as follows. As has already been said, it starts with a standard approach with the aircraft coupled through its autopilot to the i.l.s. beams. This continues down to a height of about 300ft when the aircraft enters the fields of the Magnetic Leader Cable. When the airborne receiver detects a sufficiently large signal, the Magnetic Leader Cable is switched in place of the i.l.s. Localizer signal into the autopilot azimuth channel, and is used for azimuth guidance for the remainder of the landing.

At some installations the i.l.s. Glide Path becomes unusable below about 100ft, and so at this height the Glide Path signal is automatically switched out and for a few seconds the aircraft attitude is kept constant at the datum established during the whole of the Glide Path phase. This continues until a height of about 60ft, when it is necessary to start the flare. As has already been explained, the flare is executed by using the information from the accurate Radio Altimeter, and its guidance is used in the pitch plane all the way to touchdown.

At a height of about 15 to 20ft the final switching sequence takes place, which removes any drift due to cross wind. During this manoeuvre the wings are levelled and the aircraft heading changed by application of rudder so that a touch-down is made with the aircraft not only close to but also aligned with the runway centreline.

After touchdown the pilot disengages autos and brings the aircraft to a stop manually using as guidance either the runway centreline lights or information from the Leader Cable displayed on his flight instruments.

One other item must be mentioned, and that is Automatic Speed Control. A throttle servo is fitted, using as inputs airspeed and pitch attitude, which maintains the aircraft speed constant during the approach and which, during the flare, automatically closes the throttles so that at touch-down the engines are at safe idling speed.

Although the landing system is entirely automatic, the pilot fulfils the vital function of monitor. A feature of the system is that during all the phases of the landing, monitoring signals are displayed on his normal flight instruments. These form part of the Smiths' Flight system. The only additional instrument is a sequence indicator which displays the successive phases of the landing.

Leader Cable Ground Equipment

The equipment can be conveniently housed in a small building at one end of the runway, and provision is made for remote control by an operator in the airfield control tower. The operation of the ground equipment is continuously monitored and in the event of any fault arising which might impart misleading information to an approaching aircraft, the equipment is automatically switched off and a warning given to the operator.

The ground installation consists of the two Leader Cables laid parallel to and equi-distant from the runway centre line; the cable termination boxes at the remote end of the runway; the Alternator Unit generating high-frequency a.c. for the Cables; the Power Unit and monitoring and switching equipment mounted together in a Console;

and the Remote Control Unit enabling the operator to initiate and interrupt the working of the equipment from a distance by means of a G.P.O. land line.

The Monitor Unit mounted at the top of the Console displays on its front panel the Cable alternating current and voltage conditions by means of two cross-pointer meters, while a third meter indicates any departure from the ideal balanced-current condition.

The Alternator Unit provides an output of 250V 4A at 1 070c/s and 1 750c/s for the Leader Cables. Its outputs are controlled by regulating circuits in the Monitor Unit which respond to an increase of Cable current by reducing the field-current supply to the alternator, so that the Cable current is maintained constant within narrow limits.

For correct operation of the system it is essential to minimize any unbalance between the two Cable currents, which would give rise to unequal magnetic fields at the runway centre line. The alternating currents in the two Cables are taken through current transformers whose outputs are rectified and applied to a centre-zero micro-ammeter in series with a sensitive centre-stable relay. In the event of the unbalance exceeding $2\frac{1}{2}$ per cent the relay is energized and gives an alarm by the illumination of a red lamp on the front panel of the Monitor Unit.

The a.c. flowing in each Cable is rectified at the Termination Box at the remote end and returned as d.c. to the Monitor Unit; this serves to check the continuity and insulation of the Cables. The d.c. in each Cable is compared with a standard current, drawn from a stabilized 150V negative line, in the windings of a two-coil centre-stable relay. If the Cable d.c. does not lie within predetermined limits the relay is energized and gives an alarm as in the previous case, while simultaneously another lamp is illuminated to show which Cable is at fault and whether its current is excessive or deficient.

Airborne Equipment

The airborne installation consists of a Control Unit located in the pilot's position, and a Receiver Assembly mounted in the aircraft rack. The Aerial Assembly is mounted externally on the fuselage of the aircraft, and is housed in a bullet shaped nylon cover to reduce drag. This housing contains the two motor driven aerial loops which rotate at 1 000 rev/min in the magnetic field present from the Leader Cables. The aerial housing is protected against icing by a heater muff which operates from the 400c/s aircraft power supply.

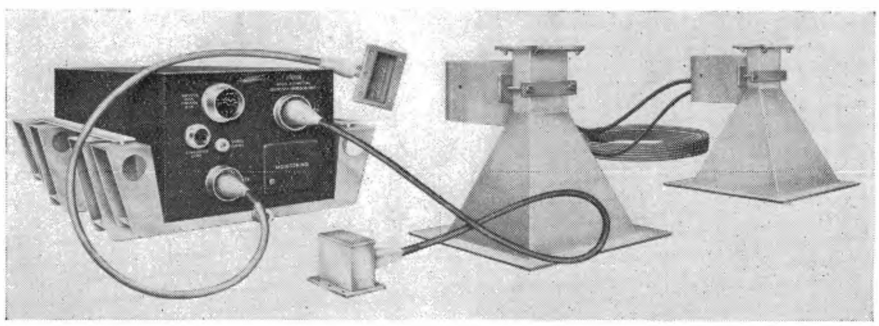
The Control Unit is normally housed in the cockpit and carries in addition to the displacement meter, the main On/Off switch and a Test switch which enables the correct functioning of the equipment to be checked prior to a landing commitment.

The magnetic fields developed by the alternating currents in the Leader Cables induce an e.m.f. at the same frequency in each of the loops of the Aerial Assembly mounted on the aircraft. These e.m.f.'s are fed via band-pass filter circuits to the primary of a transformer on the Receiver Unit; the secondary of this transformer is fed through a two-stage amplifier and filtered at each stage of amplification to minimize the noise content of the resulting signal. This is then resolved into its separate components at 1 070c/s and at 1 750c/s by narrow band-pass filters. Each component is then detected by the d.c. and i.f.c. unit and its magnitude controls the current, and hence cathode potential, of a cathode-follower comparator stage in that unit. In the event of the aircraft being displaced from the runway centre line, the two signal components and resultant cathode potentials will be unequal, and a voltage proportional to the inequality will be fed to the automatic pilot system of the aircraft to initiate corrective action. The meter on the Control Unit will also indicate the magnitude and direction of the displacement.

Radio Altimeter

The installation consists of a counter unit and a power unit, a transmitter/receiver unit fitted close to the two horn-type aerials and a control unit and either one or two indicators as required.

A frequency-modulated signal is radiated by one of the horn aerials and after reflection from the ground is received by the other aerial. It is then mixed with a signal direct from the transmitter and, since a brief period of time has elapsed since transmission, a difference frequency (or 'video signal') is produced in the mixer. This frequency is proportional to the clearance of the aircraft from the ground, to the sweep and to the sweep frequency.



The combined Transmitter-Receiver and horns of the Radio Altimeter as used on 'Autoland'.

After pre-amplification in the transmitter/receiver unit, the video signal is fed by a coaxial line to the counter unit. There, after further amplification, a d.c. voltage proportional to the video frequency is produced. This voltage drives the indicators.

A 'confidence switch' on the control unit operates as follows:—It energizes waveguide switches which divert power from the aerials into a length of delay cable. This cable provides an artificial signal at a realistic level for the video frequency produced, and thus checks the sensitivity and accuracy of the complete equipment.

The transmitter is a coaxial line oscillator valve which is coupled into a resonant cavity and operates in the range of 4 200 to 4 400Mc/s. Power is taken from this, via a coaxial feeder, to the radiating aerial. The resonant cavity is frequency-modulated at 300c/s by a hysteresis motor.

The receiver includes a hybrid coupler with balanced mixers. The video amplifier has a variable gain/frequency characteristic to suppress unwanted low frequency interference. After limiting, the signal is fed to the counter circuit. This circuit is so designed that 'half-counts' are impossible.

The indicator has two height ranges of 0 to 500 and 0 to 5 000ft and errors have been found not to exceed 3ft on the lower range and 30ft on the higher.

A Method of Reducing the Time Lag of Transducers which have an Exponential Response

By L. Whitlow* and M. J. Porter*

A method of reducing the time lag of thermocouples, or other transducers which have a simple exponential response, is described. The output voltage from a thermocouple is added to its amplified derivative, this corrects the response for both phase and amplitude. The limiting factor, in reducing the time lag, is the ratio of the input signal to the eletrical noise generated in the equipment.

Design details and the performance of a drift corrected d.c. amplifier and a highly stable power supply are also given.

(Voir page 571 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 576)

In the recording of fluctuating gas temperatures using thermocouples, the time lag of the thermocouple is often of the same order or even longer than the period of the fluctuation being recorded. Several methods of overcoming this have been described by Shepard and Warshawsky[1]. These are all based on feeding the output of a thermocouple to a network which has a transfer function that is the inverse of the transfer function of the thermocouple. In all the detailed methods described, the output of a thermocouple is fed to a passive network comprised of resistances and reactances which give an inverse transfer function to that of the thermocouple. With the exception of a method using transformers the output of the thermocouple is attenuated, this is impracticable when small variations of temperature have to be recorded. Furthermore, varying the network parameters to allow for variation of the time-constant of the thermocouple—which changes with gas pressure—is no easy matter.

In the method used here, the output of a thermocouple is added to its derivative which has been multiplied by a constant. If a step change of temperature is applied to a thermocouple, which has a simple exponential response, then the change of the output voltage v is given by $v = V[1 - \exp(-t/\tau_1)]$ where V is the steady state output voltage of the thermocouple for the applied temperature change, t is the time in seconds and τ_1 is the time-constant of the thermocouple. The derivative of this is $dv/dt = V/\tau_1 \exp(-t/\tau_1)$. If this is multiplied by a constant coefficient equal to τ_1 and then added to the original function then:

$$v + \tau_1 dv/dt = V[1 - \exp(-t/\tau_1)] + V \exp(-t/\tau_1) = V$$

This is a step change of voltage proportional to the step change of temperature applied to the thermocouple.

In practice the differentiating circuit will have a time-constant τ_2, which is very much shorter than that of the thermocouple. This modifies the output to $v = V[1 - \exp(-t/\tau_2)]$. A more rigorous proof along with the effects of input and output impedance on the differentiating circuit is given in Appendix 1.

Exponential Lag Corrector

A block diagram of the Exponential Lag Corrector, as the equipment is called, is given in Fig. 1. The output of a Chromel/Alumel thermocouple, which has a thermo-electric power of 40mV/1 000°C, is first of all amplified by the drift-corrected d.c. amplifier. This has a gain of 900, the output of this amplifier is then fed into a mixer stage and also to a resistance-capacitance differentiating circuit. This differentiating circuit is the input circuit of an uncorrected d.c. amplifier which has a gain of 10 to 68 times. This multiplies the differentiated signal by the correct coefficient, and applies it to the mixer stage. In the mixer stage the original signal and its amplified derivative are added together; this combined signal is then fed to an output stage. This output stage is designed to operate a 65c/s recording galvanometer. Provision is made for varying the sensitivity

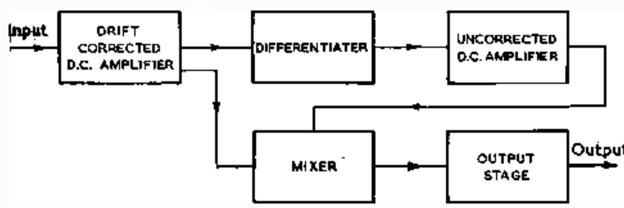


Fig. 1. Exponential lag corrector

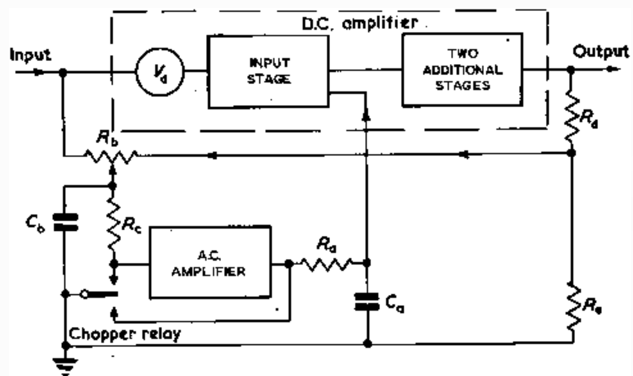


Fig. 2. Drift corrected d.c. amplifier

of the galvanometer and also for supplying a 'backing off' current, so that small changes in temperature anywhere in the range 0 to 1 000C can be recorded.

Drift Corrected D.C. Amplifier

The output voltage of d.c. coupled amplifiers will drift with time, this drift which is random, is due to changes in temperature, cathode emission, grid current, etc., and it cannot be reduced by negative feedback. To reduce drift, a drift correcting circuit proposed by Goldberg[2] is used. The principle of this circuit is that the output of the d.c. amplifier is attenuated by a factor equal to the amplifier gain. This is then subtracted from the input voltage, the only voltage now remaining is that due to drift in the d.c. amplifier. This drift voltage is amplified by a 'chopper' type amplifier, and then applied to the input of the d.c. amplifier in such a polarity as to oppose the drift. In this way the drift voltage is reduced by a factor equal to the gain of the 'chopper' amplifier, which in this case is 900. In the block diagram of the drift corrected d.c. amplifier

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(Fig. 2) the drift at the output is represented by the voltage generator, V_d, in series with the input of a driftless d.c. amplifier. This gives the same voltage at the output, as that due to drift in the amplifier being considered.

The output of the d.c. amplifier, which is of opposite polarity to the input, is attenuated by R_d and R_e and applied to one side of R_b, the d.c. amplifier input voltage being applied to the other side. The slider of R_b is adjusted so that the voltage on it, due to any voltage applied to the input socket, is always zero. The voltage at the slider is thus solely due to drift in the d.c. amplifier. This voltage is 'chopped' by a relay, and then amplified by an a.c. amplifier. The output of the a.c. amplifier is rectified by the same relay acting as a synchronous rectifier. This rectified output is then smoothed by R_a and C_a, before being applied to the input stage of the d.c. amplifier, to oppose the original drift.

The detailed circuit diagram of the drift corrected d.c. amplifier is shown in Fig. 3. V_1, V_3 and V_4 form the d.c. amplifier, with V_2 as a drift control valve, while V_5 and V_6 form an a.c. amplifier for the 'chopped' drift voltage. V_1 has a common cathode load with the drift control valve V_2. The output of V_1 is fed to V_3 and then V_4, the operating potentials of the d.c. amplifier are set by RV_1. The output voltage of V_4 is fed to R_{15}, R_{16}, R_{17} and R_{18}, which reduce the datum level of the signal. One tapping from this chain is taken to the differentiator and the mixer, while another tapping is used to apply negative feedback to V_3. Also from this same point the output is applied to an attenuator which reduces the signal to the same level as the original input in order to obtain the drift voltage. A 'hum bucking' voltage obtained from RV_2, which is across a 6·3V a.c. heater winding on the mains tranformer, is also applied to this point. The screen voltages for V_1, V_2 and V_3 are obtained from a 150V stabilizer in the power supply unit. To prevent the amplifier oscillating at a high frequency, due to the phase change round the amplifier loop being zero when the gain is not less than unity, the capacitors C_1 C_2 and C_3 are shunted across from the anode of one valve to the input of the next. These advance the phase of the higher frequencies and so ensure the circuit is stable.

The heater supply of V_1 and V_2 is obtained from a con-

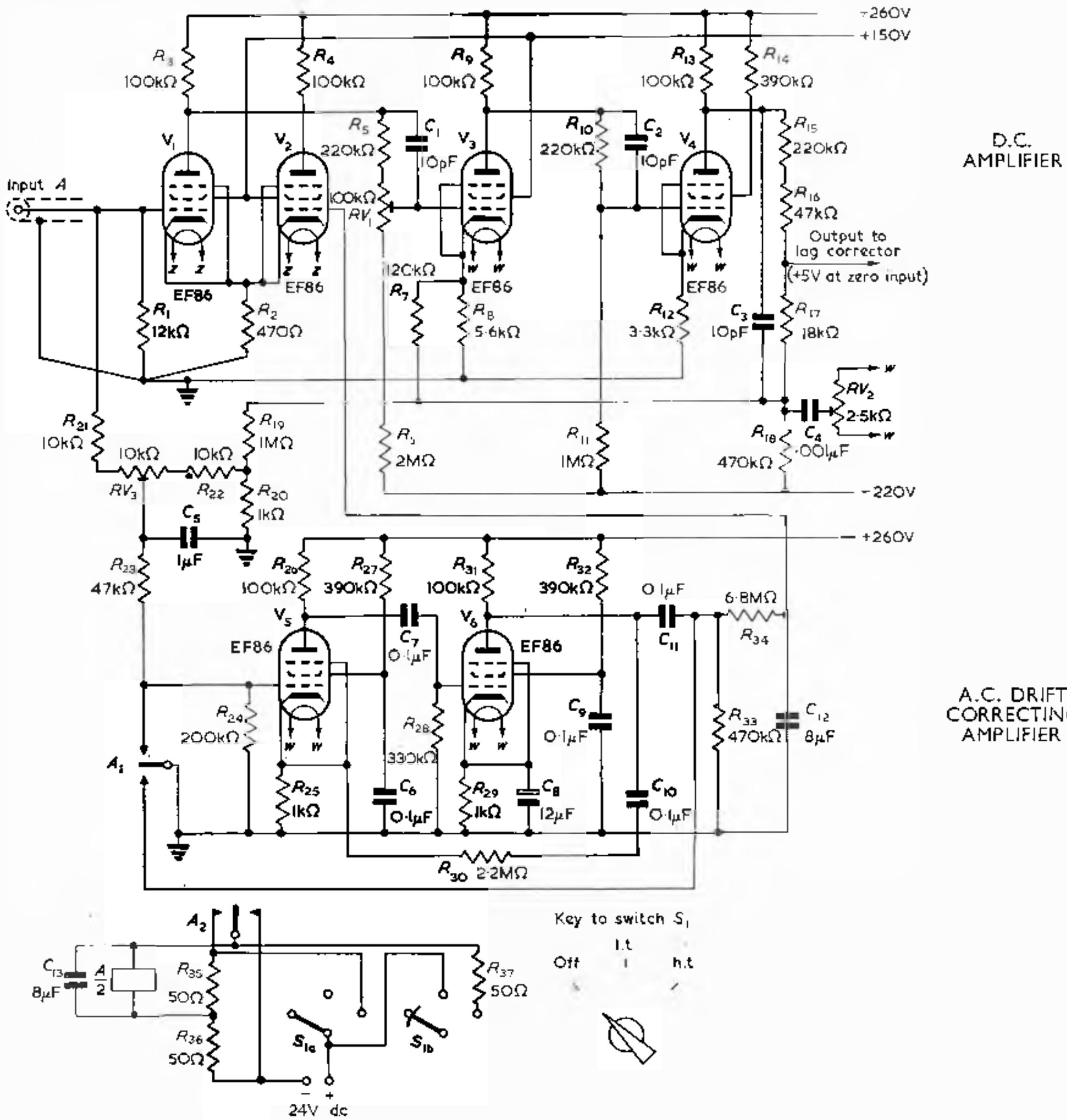

Fig. 3. Complete circuit of drift corrected d.c. amplifier

stant voltage transformer which helps to reduce the drift due to heater voltage changes. The output voltage at the junction of R_{17} and R_{18} is attenuated by R_{19} and R_{20}, and is applied to one side of the preset potentiometer RV_3. The input voltage, which is of opposite polarity, is applied to the other side of the potentiometer, and the slider is set so that the voltage on it due to any voltage applied to the input socket is zero. Any drift which occurs in the d.c. amplifier, however, will appear only on one side of the potentiometer, and so will not be cancelled out. This drift voltages charges C_{10}, which is periodically discharged through R_{23} and the relay contacts. The grid voltage of V_5 is thus a 'chopped' d.c. which is amplified by V_5 and V_6. These two valves and their associated circuit form an a.c. amplifier, negative feedback being applied by R_{30} and R_{25}, the gain being 900. The output of the a.c. amplifier is rectified by the relay contacts which function as a synchronous rectifier, the input and output of the a.c. amplifier being 'earthed' alternately. The rectified output can be either a positive or negative voltage by virtue of the synchronous rectification, the polarity being opposite to that at the slider of RV_3. This output is smoothed by R_{34} and C_{12} and then fed to the grid of the cathode control valve V_2. Since V_2 has a common cathode load with V_1 then any voltage applied to V_2 grid has the same effect as one of opposite polarity applied to V_1 grid. By this means it is possible to inject a voltage into the input in such a way as to counteract the drift.

To summarize, if the d.c. amplifier drifts so that the output goes positive say, then the drift voltage after being attenuated in the input/output comparison circuit is amplified by a 'chopper' amplifier and appears as a negative voltage at the grid of V_2. This has the same effect as a positive voltage on V_1 grid, and causes the output voltage to go negative and so correct the drift at the output. Since the drift correction is essentially an error-operated device some drift must always be present, but this is inversely proportional to the gain of the a.c. amplifier. This is shown in Appendix 2.

All the valves used in the drift-corrected d.c. amplifier are low noise, low hum, and low microphony, pentodes. V_1 is selected for the lowest noise figure and the heater transformer supplying V_1 and V_2 has an adjustable centre-tap to reduce hum. These, along with the 'hum bucking' circuit, reduce the noise and hum to less than 5μV peak-to-peak when referred to the input. The input impedance of the drift corrected d.c. amplifier is 7kΩ and the output impedance is 100kΩ. The gain and phase characteristics are shown in Fig. 4. The drift voltage referred to the input is less than 25μV.

The 'chopper' relay A is operated by the 24V d.c. supply, the current through the coil being reversed by the contacts A_2 (3). The frequency of operation is about 160c/s, being determined by R_{35}, R_{36} and C_{13}. C_{13} also filters out any noise due to the current changeover in the coils. The relay is normally at rest with the wiper in the centre position, both contacts being open. To start the relay vibrating a set of contacts on the main switch, S_1, has a make-before-break contact, so that in switching from 'heaters on' to 'h.t. on' relay contacts A_3 are momentarily connected together to energize the relay initially and so start it operating.

Differentiator, Uncorrected D.C. Amplifier Mixer and Output Stage

The output of the drift corrected d.c. amplifier is applied to the differentiating circuit formed by C_{14}, C_{15}, RV_4, RV_5, R_{38} and the output impedance of the drift corrected d.c. amplifier (Fig. 5). The time-constant of the differentiating circuit is 9·2msec. The differentiated signal at the slider of

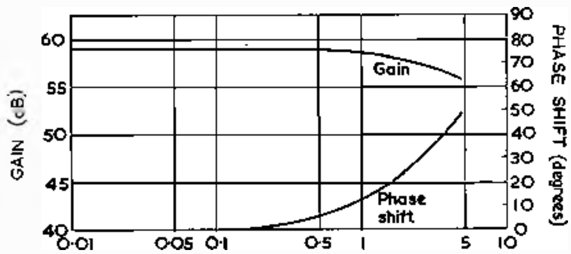

Fig. 4. Drift corrected d.c. amplifier characteristics

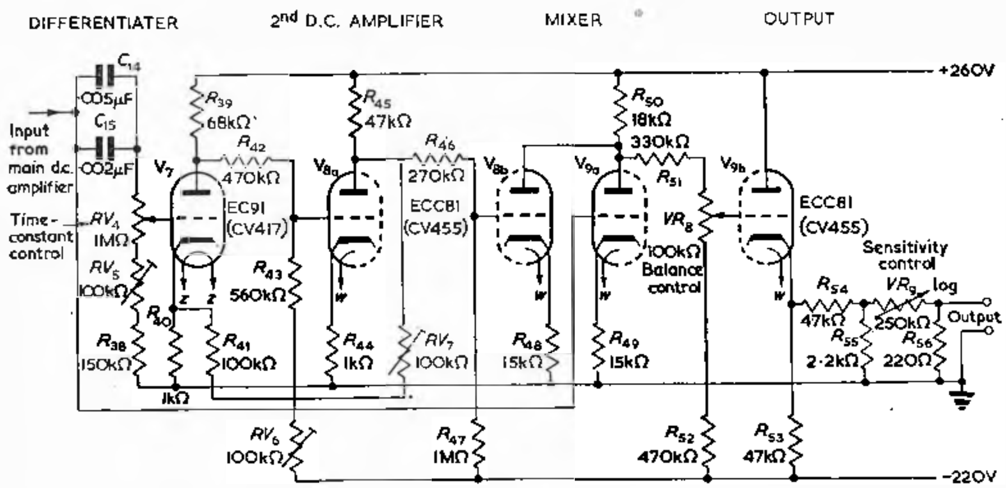

Fig. 5. Uncorrected d.c. amplifier, differentiator, mixer and output stage

RV_4 is fed to the grid of V_7 which along with V_{8a} forms an uncorrected d.c. amplifier with a gain of 68; negative feedback being applied by RV_7, R_{41} and R_{40}. The amplification of the differentiated signal is adjusted by RV_4, which is brought out to a control knob on the front panel and calibrated in time-constants of 0·1 to 0·6sec. The minimum amplification of the differentiated signal is adjusted to 10 times by the preset control RV_5. The maximum amplification of 68 times is adjusted by varying the negative feedback of the d.c. amplifier by means of the preset control RV_7. The operating potentials of the d.c. amplifier are set by the preset RV_6.

The output of the differentiating circuit must be amplified by a d.c. amplifier, since the d.c. levels would be lost in an a.c. amplifier. This is because the mean signal level must always be zero in an a.c. amplifier. The mean level of the signal from the differentiator is seldom zero, and so distortion would take place when an a.c. amplifier made the mean signal level zero.

The output from V_{8a} anode is fed to V_{8b} grid, V_{8b} along with V_{9a} forming a mixer stage. The differentiated and undifferentiated signals are added together across the common anode load of V_{8b} and V_{9a}.

The output of the mixer, which is the corrected response, is fed to V_{9b} which is the output stage. This is designed to operate a 65c/s recording galvanometer which has an impedance of 50Ω and requires 100μA for full scale deflexion; the shunt resistance for critical damping is 220Ω.

V_{9b} is connected as a cathode-follower across the positive and negative supplies, the cathode being connected to earth through a network which feeds the recording galvanometer. R_{54} and R_{55} form an attenuator, while RV_9 is the sensitivity control, which varies the galvanometer current. The damping resistor is R_{56}. The 'balance' control, RV_8, sets the operating potential of V_{9b} and is normally set so that with the input to the exponential lag corrector earthed the galvanometer deflexion is zero. When small changes of temperature are being investigated RV_8 provides a means of 'backing off' the deflexion due to the steady temperature. RV_8 and RV_9 are both brought out to control knobs on the front panel, being labelled 'Balance' and 'Sensitivity respectively.

Power Supplies

The design of low drift d.c. amplifiers calls for a very stable power unit, and if very large voltages are to be avoided, positive and negative supplies are required.

For this equipment a positive supply of 260V and a negative supply of 220V are used. These are shown in Fig. 6. The design of these stabilized supplies are based on a circuit by Miller[3]. The operation of these two supplies is identical, except that the negative line is earthed on the positive supply, and the positive line is earthed on the negative supply. Only one unit will be described, since they are both identical in operation.

The output of the transformer T_1 is rectified by V_{10}, and smoothed by L_1, C_{16} and C_{17}, R_{57} acting as a 'bleeder' across the smoothing circuit. The voltage across C_{17} is approximately 500V and this is applied to the stabilizing circuit formed by V_{11}, V_{12a}, V_{12b}, V_{13a}, V_{13b} and V_{14}. The grid of V_{13b} is held at 85V by the voltage reference tube V_{14}, any noise appearing at the anode of V_{14} being smoothed out by R_{70} and C_{20} before being applied to V_{13b} grid. V_{13a} grid is held at a voltage slightly negative to this, the common cathode will thus be slightly above 85V. If any change occurs in the +260V line, then the grid of V_{13a} will vary in the same polarity. This variation will be amplified by V_{13a} and will appear at its anode in opposite polarity. This is fed to V_{12b} grid and since V_{12a} and V_{12b} have a common cathode load, then V_{12a} anode will vary in the same polarity. V_{11} grid is thus varied in such a way as to oppose the change of the 260V line.

Any change of the unstabilized 500V line is amplified and inverted by V_{12a} which causes V_{11} to oppose the change. RV_{10} sets the amount by which the change in the unregu-

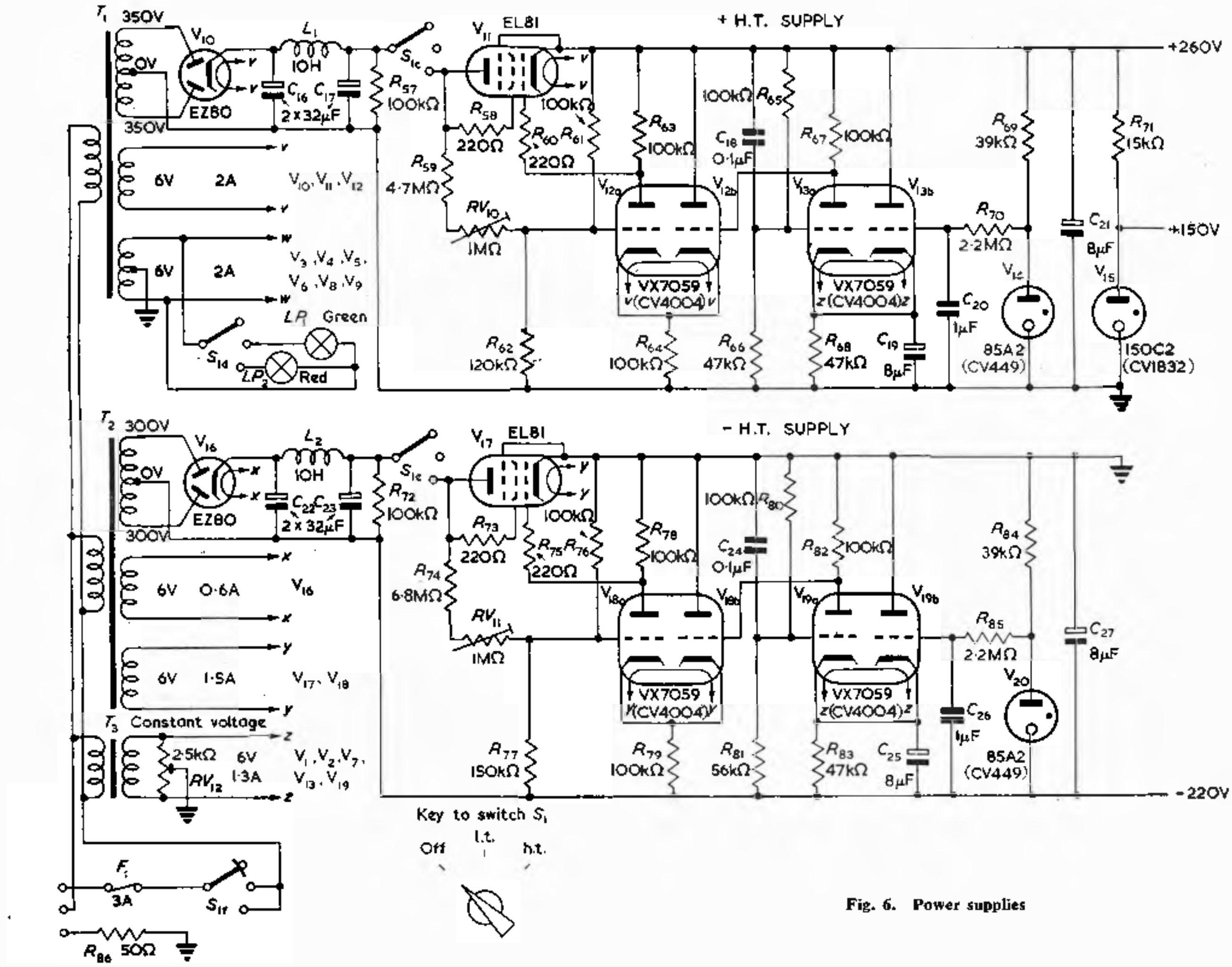

Fig. 6. Power supplies

lated supply is amplified. The detailed analysis of the power unit is given in Appendix 3, which also shows the effect of the settings of RV_{10}.

The heater supplies of V_{13} and V_{19} are obtained from a constant voltage transformer, which reduces the drift of the output voltage. A screen voltage supply for the drift corrected d.c. amplifier is obtained from the 150V stabilizer V_{15}. The mains switch is a multi-pole three-way switch, which in addition to starting the chopper relay also switches on the heaters and the h.t. supplies.

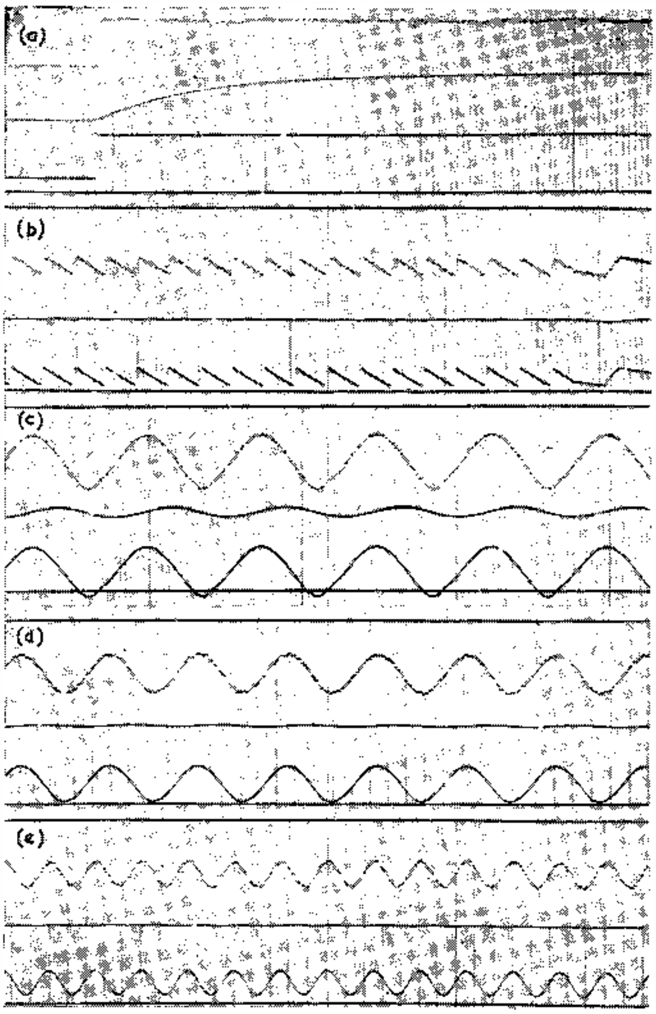

Fig. 7. **Performance of lag corrector for various input functions using a transducer with a 0·6sec time-constant and showing:**
Corrected output
Transducer output
Input function
(a) Step function
(b) 5c/s sawtooth wave
(c) 1·3c/s sine wave
(d) 5c/s sine wave
(e) 10c/s sine wave

The output impedance of the power supply is less than 0·25Ω and the hum is less than 100μV. A 15 per cent change in the mains supply causes an output voltage variation of less than 5mV.

Performance

The input impedance of the exponential lag corrector is 7kΩ and the maximum input signal is 40mV. It will reduce time-constants in the range of 100 to 600msec down to 10msec. The sensitivity is such that full scale deflexion of the recording galvanometer can be obtained for an input of 600μV. (This is equal to 15°C). The noise level referred to the input varies from 50μV peak-to-peak when used with a thermocouple of 100msec time-constant, to 340μV peak-to-peak when used with a thermocouple with a 600msec time-constant. The output drift is less than 200μV when referred to the input, this is mainly due to drift in the uncorrected d.c. amplifier.

Fig. 7 shows the output obtained from the exponential lag corrector for various input functions. The output that would be obtained under these conditions from the thermocouple is also shown.

Conclusions

This system can be adapted for use with other transducers or circuits which have an exponential response. The limiting conditions for reducing the time lag are the ratio of input signal to the self generated noise and the bandwidth of the amplifiers. In use with thermocouples it has been found that, when using it to record temperatures in a gas jet, fluctuations in the gas temperature due to turbulence are the limiting factor on the sensitivity and not the self generated noise in the equipment.

Acknowledgments

The authors would like to thank the Chief Scientist, Ministry of Supply for permission to publish this article.

Appendix 1

Principles of Lag Correction

Transfer Function of a Thermocouple

If a thermocouple is exposed to a varying gas temperature $T_o(t)$, the time rate of change dT_1/dt of the element temperature $T_1(t)$ is proportional to the temperature difference, when radiation or thermal conduction from the bead is negligible. This can be expressed as

$$\tau_1(dT_1/dt) = T_o - T_1 \text{ or } \tau_1(dT_1/dt) + T_1 = T_o \quad \ldots \text{(1)}$$

The output e.m.f. of a thermocouple can be expressed as being proportional to the difference between T_1 of the thermocouple and some reference temperature T_r and

$$e_1 = Q(T_1 - T_r) \quad \ldots\ldots\ldots\ldots\ldots \text{(2)}$$

T_r being the cold junction temperature and Q the thermoelectric power of the couple.

An e.m.f. e_o can thus be defined as the equivalent e.m.f. produced by the primary element at a temperature T_o

$$\text{i.e. } e_o = Q(T_o - T_r) \quad \ldots\ldots\ldots\ldots\ldots \text{(3)}$$

Substituting equations (2) and (3) in (1) gives

$$\tau_1(de_1/dt) + e_1 = e_o \quad \ldots\ldots\ldots\ldots\ldots \text{(4)}$$

By the use of the Laplace transform[4] this can be expressed as

$$e_1(1 + P\tau_1) = e_o \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(5)}$$

where P replaces the differential operator $d()/dt$.

The ratio

$$e_1/e_o = \frac{1}{1 + P\tau_1} \quad \ldots\ldots\ldots\ldots \text{(6)}$$

can be defined as the transfer function $Y_1(P)$, where $Y_1(P)$ is defined as the ratio of input quantity to output quantity.

Transfer Function of a Resistance Capacitance Differentiating Circuit

A resistance capacitance differentiating circuit is shown in Fig. 8.

If I is the current in the circuit $I = dq/dt$ where q is the instantaneous charge on the capacitor C. The voltage V_c across the capacitor is given by

$$V_c = q/C = (1/C)\int I\,dt$$

$$\therefore E_i = IR + V_c;\ \text{using the Laplace transform}$$

$$\bar{V}_c = \bar{I}/PC$$

and

$$\bar{E}_i = \bar{I}(R + 1/PC)$$

and

$$\bar{I} = \frac{\bar{E}_i PC}{PCR + 1}$$

the transfer function $Y_2(P) = \bar{E}_o/\bar{E}_i = \bar{I}R/\bar{E}_i$

$$= \frac{P\tau_2}{1 + P\tau_2} \quad \ldots\ldots\ldots\ldots\ldots\ (7)$$

where $\tau_2 = CR$; the time-constant of the resistance capacitance network.

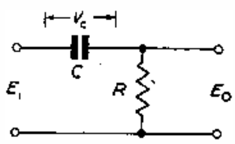

Fig. 8. Resistance capacitance differentiating circuit

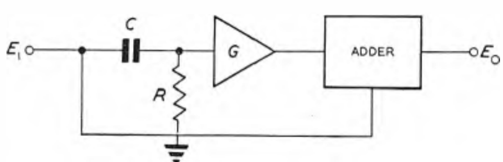

Fig. 9. Lag corrector

Lag Correction

If the output of the thermocouple is fed to the circuit shown in Fig. 9 then the transfer function

$$Y_3(P) = 1 + \frac{GP\tau_2}{1 + P\tau_2} \quad \ldots\ldots\ldots\ldots\ (8)$$

where G is the amplifier gain. The combined transfer function $Y_4(P)$ of a thermocouple and this circuit in cascade $= Y_1(P)\,Y_3(P)$

$$\therefore Y_4(P) = \frac{1}{1 + P\tau_1}\left\{1 + \frac{GP\tau_2}{1 + P\tau_2}\right\}$$

$$= \frac{1}{1 + P\tau_1}\left\{\frac{1 + P\tau_2(G + 1)}{1 + P\tau_2}\right\}$$

If the amplifier gain G is adjusted so that it is equal to $(\tau_1/\tau_2) - 1$, then

$$Y_4(P) = \frac{1}{1 + P\tau_1}\left\{\frac{1 + P\tau_2(\tau_1/\tau_2)}{1 + P\tau_2}\right\}$$

$$\therefore Y_4(P) = \frac{1}{1 + P\tau_2} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (9)$$

This equation is of the same form as equation (6), which is the transfer function of a thermocouple $Y_1(P)$.

The time-constant has now decreased from τ_1 to τ_2. The amplifier gain G has to be made equal to $G = K - 1$ where K is defined as the time-constant improvement factor and is equal to τ_1/τ_2.

Effect of Input and Output Resistance on the RC Differentiating Network

The *RC* network with the input resistance R_g and output resistance R_L is shown in Fig. 10. Since R_L is in parallel with R this can be replaced by R_t where $R_t = RR_L/(R + R_L)$. This is shown in **Fig. 11.**

From which it is seen that

$$\bar{E}_i = \bar{I}(R_g + R_t + 1/PC)$$

$$\therefore \bar{I} = \frac{\bar{E}_i PC}{1 + PC(R_g + R_t)}$$

and the transfer function $Y_5(P) = PC\,R_t/(1 + P\tau_3)$ where $\tau_3 = C(R_t + R_g)$.

The transfer function for the network containing R_g and R_L and the thermocouple is now

$$Y_5(P) = \frac{1}{1 + P\tau_1}\left\{1 + \frac{G'PC\,R_t}{1 + P\tau_3}\right\}$$

where G' is the amplifier gain

$$\therefore Y_5(P) = \frac{1 + P\tau_3 + G'PC\,R_t}{(1 + P\tau_1)(1 + P\tau_3)}$$

$$= \frac{1 - (\tau_3/\tau_1) - G'(CR_t/\tau_1)}{[1 - (\tau_3/\tau_1)]\ (1 + P\tau_1)} + \frac{1 - 1 - G'(CR_t/\tau_3)}{[1 - (\tau_1/\tau_3)]\ (1 + P\tau_3)}$$

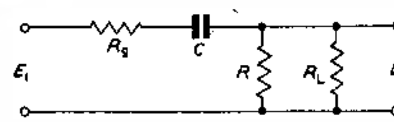

Fig. 10. Resistance capacitance differentiating circuit with generator and load resistances

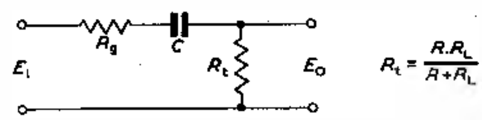

Fig. 11. Equivalent circuit of Fig. 10

The first term vanishes when the numerator is zero, i.e. $1 - (\tau_3/\tau_1) - G'(CR_t/\tau_1) = 0$.

$$\therefore G' = \frac{\tau_1 - \tau_3}{CR_t}$$

$$\therefore Y_5(P) = \frac{\left(\dfrac{\tau_3 - \tau_1}{CR_t}\right).\ CR_t/\tau_3}{\left(\dfrac{\tau_3 - \tau_1}{\tau_3}\right)(1 + P\tau_3)}$$

$$= \frac{1}{1 + P\tau_3} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ (10)$$

The only effect of the input and output resistances is to reduce the output time-constant from τ_2 to τ_3. This is due to the shunting effect of R_L on R. To compensate for this and also for the attenuation due to R_g the gain G' has to be increased from $\dfrac{\tau_1 - \tau_2}{\tau_2}$ to $\dfrac{\tau_1 - \tau_3}{CR_t}$.

Appendix 2

Drift Correction of the D.C. Amplifier

A block diagram of the d.c. amplifier with drift correction is shown in Fig. 12. From this it is seen that

$$V_a = -(V_i + V_d - V_c)\,\alpha \quad \ldots\ldots\ldots\ (1)$$

also

$$V_b = -(V_i + V_d - V_c) \quad \ldots\ldots\ldots\ (2)$$

and

$$V_b = (V_a + V_i)$$

substituting this in equation (2) gives $V_b = -(V_d - V_c)$. Since $V_c = -\beta V_b$

$$\therefore V_c = (V_d - V_c)\,\beta \quad \ldots\ldots\ldots\ldots\ (3)$$

$$= \frac{V_d\beta}{1 + \beta} \quad \ldots\ldots\ldots\ldots\ldots\ (4)$$

substituting this in equation (1) gives

$$V_o = -\left(V_i + V_d - \frac{V_d\beta}{1+\beta}\right)\alpha$$

$$= -\left(V_i + \frac{V_d}{1+\beta}\right)\alpha \quad \ldots\ldots\ldots\ldots\ldots\ldots (5)$$

Thus the drift voltage V_d is reduced by a factor $1/(1+\beta)$.

In the practical circuit there are two time-constants which must be selected to give critical damping. These are shown in Fig. 13.

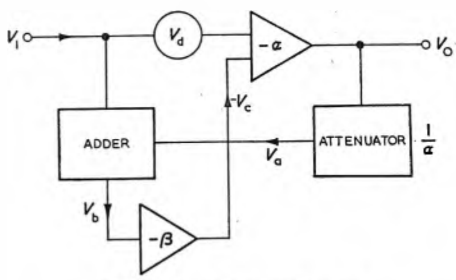

Fig. 12. Drift correction circuit

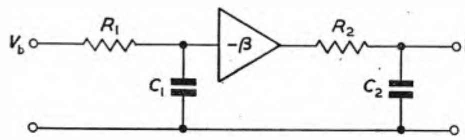

Fig. 13. Chopper amplifier showing input and output time-constants

Equation (3) is now modified to

$$\overline{V}_c = \frac{(\overline{V}_d - \overline{V}_c)\beta}{(1+P\tau_1)(1+P\tau_2)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

where

$\tau_1 = C_1R_1$ and $\tau_2 = C_2R_2$,

$$\overline{V}_c = \frac{\overline{V}_d\beta}{(1+P\tau_1)(1+P\tau_2)} \cdot \frac{(1+P\tau_1)(1+P\tau_2)}{(1+P\tau_1)(1+P\tau_2)+\beta}$$

$$\therefore\ \overline{V}_c = \frac{\overline{V}_d\beta}{(1+P\tau_1)(1+P\tau_2)+\beta} \quad \ldots\ldots\ldots\ldots\ldots (7)$$

substituting equation (7) in equation (1) gives

$$\overline{V}_o = -\alpha\left\{\overline{V}_i + \overline{V}_d - \frac{\overline{V}_d\beta}{(1+P\tau_1)(1+P\tau_2)+\beta}\right\} \ldots\ldots (8)$$

$$= \frac{-\alpha[\overline{V}_i + \overline{V}_d]\,\{(1+P\tau_1)(1+P\tau_2)+\beta\} - \overline{V}_d\beta]}{(1+P\tau_1)(1+P\tau_2)+\beta} \quad (9)$$

For critical damping the term $(1+P\tau_1)(1+P\tau_2)+\beta$ in the denominator of equation (9) must be of the form $(P+A)^2$.

$$(1+P\tau_1)(1+P\tau_2)+\beta = P^2\tau_1\tau_2 + P(\tau_1+\tau_2) + (1+\beta)$$

$$= P^2 + P\frac{(\tau_1+\tau_2)}{\tau_1\tau_2} + \frac{1+\beta}{\tau_1\tau_2} \quad \ldots\ldots\ldots\ldots (10)$$

This is equal to

$$\left(P + \frac{\tau_1+\tau_2}{2\tau_1\tau_2}\right)^2 + \frac{1+\beta}{\tau_1\tau_2} - \left(\frac{\tau_1+\tau_2}{2\tau_1\tau_2}\right)^2$$

which is of the required form if

$$\frac{1+\beta}{\tau_1\tau_2} = \left(\frac{\tau_1+\tau_2}{2\tau_1\tau_2}\right)^2$$

i.e. $$1+\beta = \frac{\tau_1^2+\tau_2^2+2\tau_1\tau_2}{4\tau_1\tau_2}$$

$$\therefore\ \beta = \frac{\tau_1^2+\tau_2^2+2\tau_1\tau_2-4\tau_1\tau_2}{4\tau_1\tau_2}$$

$$= \frac{(\tau_1-\tau_2)^2}{4\tau_1\tau_2}$$

If $\tau_2 \gg \tau_1$ as it is in practice then

$$\beta = -(\tau_2/4\tau_1) \quad \ldots\ldots\ldots\ldots\ldots (11)$$

In the circuit used τ_1 is comprised of C_5 and either R_{21} in series with half of RV_3, or R_{22} in series with the other half of RV_3, being dependent on whether the drift is negative or positive. τ_2 is comprised of R_{34} and C_{12}, the ratio of τ_2/τ_1 being 54·4/0·015. This gives a gain β of 54·4/0·06 = 900.

Because of the effect of the input and output time-constants the gain of the 'chopper' amplifier, and hence the drift of the main d.c. amplifier, is limited by the required speed of response of the drift correction.

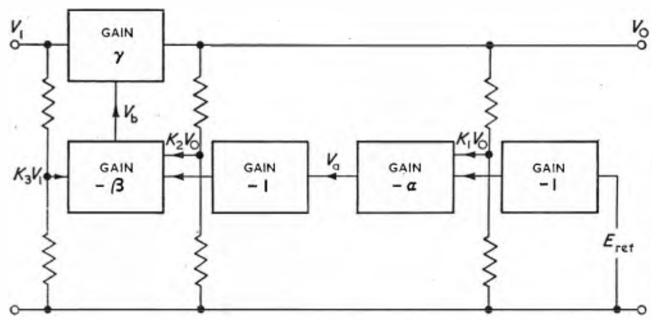

Fig. 14. Power supply stabilizer circuit

Appendix 3

D.C. Stabilizer for Power Supplies

A block schematic diagram of the stabilizer circuit is shown in Fig. 14. Using the symbols defined in the figure

$$V_i + \gamma V_b - V_o = 0 \quad \ldots\ldots\ldots\ldots\ldots (1)$$

$$V_b = -\beta(K_3V_i + K_2V_o - V_a) \quad \ldots\ldots\ldots (2)$$

$$= -\beta\,\{K_3V_i + K_2V_o + \alpha(K_1V_o - E_{ref})\} \ldots (3)$$

substituting $V_o - V_i = \gamma V_b$ from equation (1) gives

$$V_o - V_i = -\beta\gamma\,\{V_o(K_2+\alpha K_1) + K_3V_i - \alpha E_{ref}\}$$

$$\therefore\ V_o\{1+\beta\gamma(K_2+\alpha K_1)\} = V_i(1-\beta\gamma K_3) + \alpha\beta\gamma E_{ref}$$

$$\therefore\ V_o = \frac{V_i(1-\beta\gamma K_3) + \alpha\beta\gamma E_{ref}}{1+\beta\gamma(K_2+\alpha K_1)} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

if K_3 is made equal to $1/\beta\gamma$ then the term containing V_i vanishes and under these conditions V_o is independent of changes of V_i.

This modifies (4) to

$$V_o = \frac{\alpha\beta\gamma E_{ref}}{1+\beta\gamma(K_2+\alpha K_1)} \quad \ldots\ldots\ldots\ldots (5)$$

Since $\beta\gamma(K_2+\alpha K_1) \gg 1$ and $\alpha K_1 \gg K_2$ this may be written as

$$V_o = E_{ref}/K_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

By adjusting RV_{19} on Fig. 3 it is possible to obtain the condition where V_o is independent of V_i, but this only holds for small changes of input voltage V_i.

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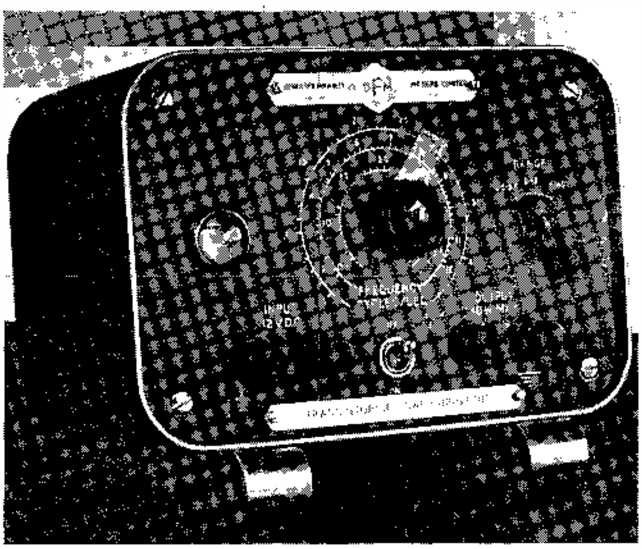
Transistors and Saturable-Core Transformers as Square Wave Oscillators

By G. C. Fleming*, A.M.I.E.E.

The use of transistors as switches for the d.c. supply to saturable core transformers is described and it is shown that by this means small and efficient convertors and invertors can be constructed. The common base, common emitter and common collector configurations are considered and methods of obtaining a multi-phase output are described.

(Voir page 571 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 576)

SMALL, efficient convertors and invertors without moving parts can be constructed using transistors to switch the d.c. supply to a saturable-core transformer. The square wave output obtained can be rectified and a minimum of smoothing is required to reduce the ripple on the resultant waveform. The main application of the convertor circuit is the provision of direct voltage supplies, e.g., for thermionic valve amplifiers from a low voltage d.c. source. The square wave a.c. output can be used to advantage as the supply source for magnetic amplifiers, fluorescent lighting, etc., when the only available supply is from accumulators.



Transistorized switching unit

History

This device appears to have been described in the first place by Uchrin and Taylor[1] in 1955, and later that year by Royer[2], both in the U.S.A. In this country the earliest work published is by Light and Hooker[3], who describe a variety of convertors including the ringing choke type. A later paper by Uchrin[4] describes a convertor capable of providing a 250W output. Further references are included in the Bibliography.

Circuits

Three basic circuits can be used and these are shown in Fig. 1; these are the common emitter, common base and common collector. The principle of operation is the same in each case but it should be noted that in the common base case the voltage is applied to N_1 and N_2 in series opposition. A longer transistor life is claimed by Campling[5] for the common base circuit. A feature of the common base circuit is the large variation of output voltage with load. This is a disadvantage for most applications but useful where a high voltage, low load device is required.

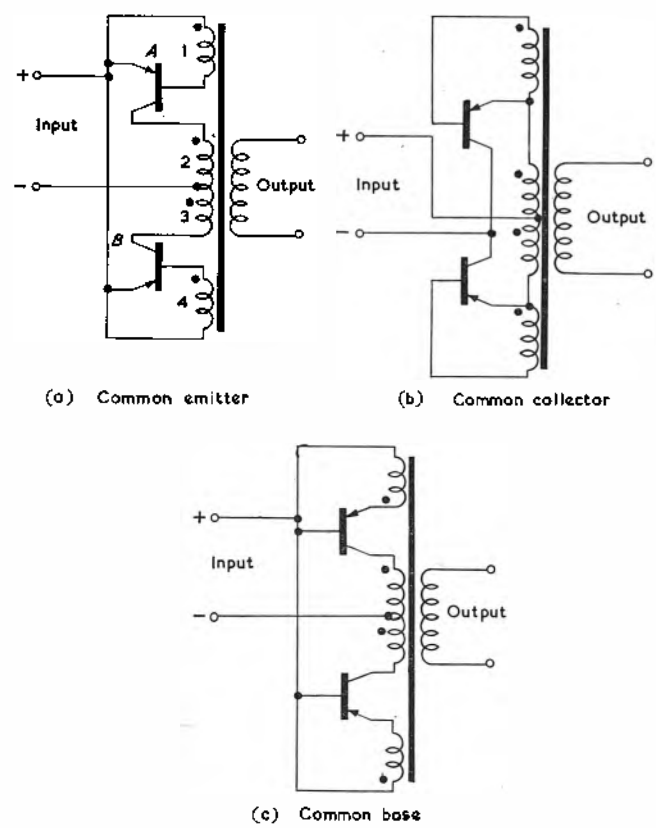


Fig. 1. Basic d.c. convertor and invertor circuits

An advantage is claimed for the common collector circuit when the negative of the supply is connected to earth, e.g., some aircraft. In this configuration all the collectors, which is frequently the case of the transistor, can be mounted directly to the chassis. A ready made heat sink is thus provided. The output can be connected in series with the input to advantage when the output voltage is up to two or three times the input voltage. A higher efficiency has been claimed for the common collector circuit. Efficiencies can be of the order of 80 per cent but depend on the load since there is a standing input on no load. Designs should therefore be aimed at operating with the load ap-

* *Denis Ferranti Meters Ltd.*

proaching the maximum output. Fig. 2(a) shows typical input current and Fig. 2(b) efficiency curves to a base of output current. In comparison with small rotating machinery or vibrators this is an appreciable saving, especially where weight and size are important factors.

Principle of Operation

The transistors are used as switches to operate alternately and so apply the input voltage across windings 2 and 3 in turn. The flux in the core is therefore driven to saturation in one direction and then the other so that a square wave a.c. output appears on winding 5.

In the common emitter circuit when transistor *A* is conducting, the supply voltage is applied to winding 2. The voltage induced in winding 1 provides base current for this transistor since the base is negative with respect to the emitter. While the flux is increasing the voltage induced in winding 4 has a polarity such that transistor *B* does not conduct, i.e., the base is positive with respect to the emitter. When saturation of the core is reached in this direction, the drive and hold-off voltages of transistors *A* and *B* respectively disappear. Then transistor *A* ceases to conduct and due to the reduction in core flux, transistor *B* starts to conduct. The base of transistor *B* is then negative to its emitter so that this transistor then conducts and causes the flux to reduce still further until saturation is reached in the opposite direction. This process is then repeated and the output voltage appears on winding 5 as a square wave with its frequency for a given core size determined by the applied voltage and the number of turns.

It is normal to make $N_2 = N_3$ and since $V_n = N_2 \dfrac{d\Phi}{dt}$

then the integral of the supply voltage over $\frac{1}{2}$ cycle is

$$\int_{-\Phi_{max}}^{+\Phi_{max}} N_2 \, d\,\Phi \text{ so that } f = \frac{V}{4N_2\,\Phi_{max}}$$

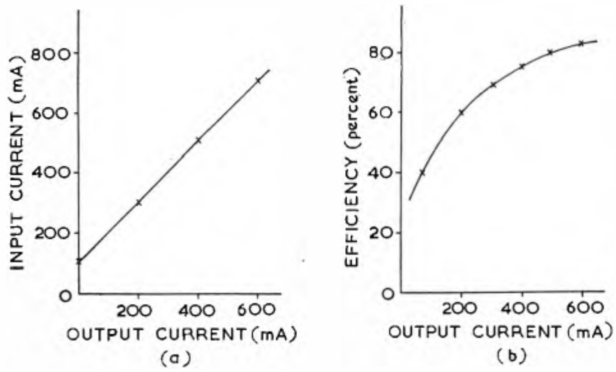


Fig. 2. **Typical convertor characteristics**
(a) Input to output current
(b) Efficiency to output current curve

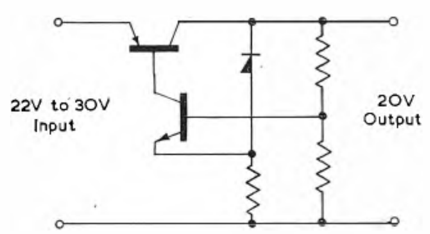
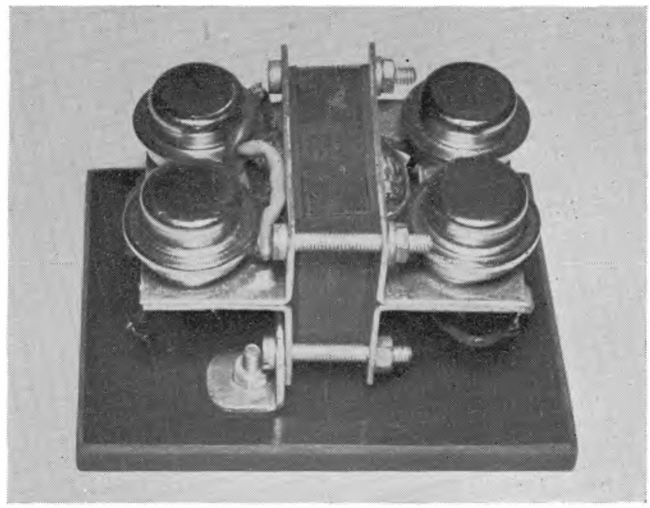


Fig. 3. **Typical transistor regulator**

For the common base (differential) circuit this expression should be modified to $f = \dfrac{V}{4(N_2 - N_1)\,\Phi_{max}}$

When a constant frequency supply is required from the transformer, it is therefore necessary to stabilize the input voltage. This can be achieved by a transistor controlled supply with a Zener diode reference as shown in Fig. 3 and described by Chase, Hamilton and Smith[6]. The frequency of operation can be chosen by the transformer design and has been known to range from a fraction of 1c/s to several kilocycles per second.

When a d.c. output is required the deciding factors for the intermediate frequency are the core size, the core losses, the rectifier losses and the transistor switching losses. The optimum is generally in the 600c/s to 1kc/s range.



A typical unit

Transistors

Since the transistors are operated as switches they can control comparatively large values of power. The effects of changes in transistor characteristic are reduced. The use of transistors as switches is fully described by Bright[7]. It should be remembered that as a result of the voltages induced in windings 2 and 3, the voltage applied to the transistors in the cut-off condition is twice the supply voltage. Selected transistors with a high turn-over voltage are therefore required for input voltages greater than 12V. Transistors rated at 13A 75V are now available and units with power outputs of several hundreds of watts are possible. The transistors can be protected against reverse polarity voltages, fed back into the circuit by inductive loads, by fitting diodes between the emitter and collector.

Transformers

The use of a high permeability material with a rectangular loop characteristic such as H.C.R. is recommended to promote rapid switching but reasonable performance can be obtained with silicone iron in C-core form and other materials. Toroids and overlapping E's are both in common use for the cores. Designs are largely based on the material and winding machines available. The saturating flux density for H.C.R. is 15 000 lines/cm² (1·5Wb/m²). So, if the frequency of operation has been selected, the cross-section, *A*, can readily be calculated.

Multi-Phase Output

Conventional capacitance phase shifting is not possible on account of the square wave output. Where a two-phase supply is required from a d.c. source, e.g., to operate an a.c. servo motor from a battery supply, this can be obtained by a pair of square-wave oscillators with phase locking as deescribed by Milnes[8]. The method used is to supply the sum of the two output voltages to a saturable choke and feed the rectified output across a resistor in the base circuit of one transistor. Similarly a three-phase supply can be obtained. A convertor supply used to drive the synchronous motor of a portable tape recorder is described by Caldwell and Wagner[9].

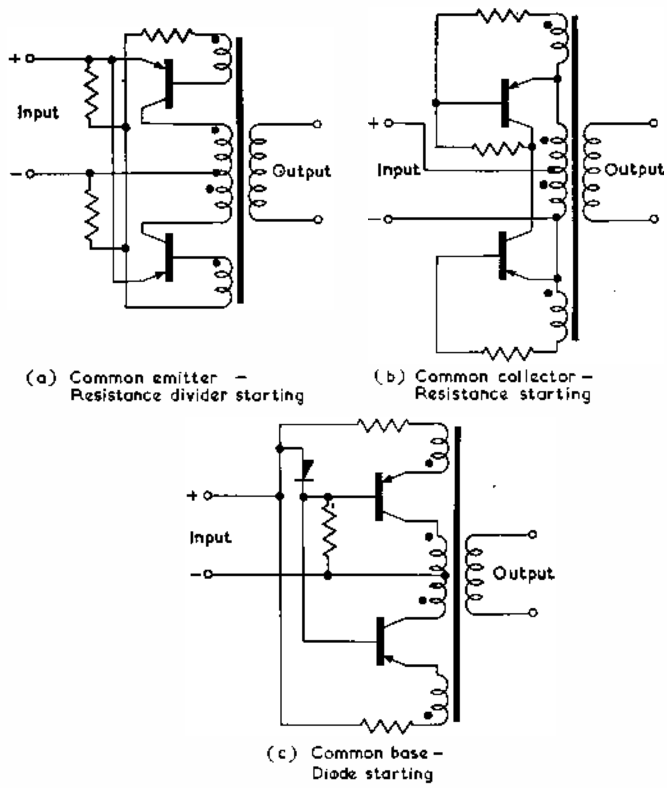


Fig. 4. Starting circuits for d.c. convertors and invertors

Output Voltage and Waveform

The output voltage can be fixed by normal transformer theory, i.e., $\frac{V_{in}}{V_{out}} = \frac{N_2}{N_1}$

neglecting regulation. Although the output waveform has so far been described as square, there are noticeable imperfections. Examples of these are (a) The peak at the leading edge of each half cycle, especially on no-load due to the leakage reactance of the primary of the transformer, (b) the slope of the leading and trailing edges due to the switching time of the transistors and (c) differences of the positive and negative half cycles caused by the unbalance of the drive windings or resistors. If necessary these effects can be eliminated by (a) fitting capacitors across the drive resistors, (b) reducing the switching frequency, (c) adjustment of drive resistors.

Starting Arrangement

Difficulty in starting especially on load and with minimum supply voltage, is common in most designs. An additional bias circuit is normally provided to overcome this difficulty. Typical circuits are shown in Fig. 4 for the resistance and diode methods. In some applications it is more convenient to disconnect the bias circuit after starting. This can be accomplished by direct switching or by relays (see Fig. 5).

Typical Units

The following units have been designed and demonstrate the wide range of applications of this device.

Purpose	*Input Voltage*	*Output Voltage*	*Output Power*	*Frequency*
Convertor, h.t. supply ..	12V	5kV	15W	1kc/s
Invertor, magnetic amplifier supply	20V	12V	0·5W	1·6kc/s
Transmitter, h.t. supply ..	6V	300V	30W	800c/s
Receiver, h.t. supply ..	6V	100V	2W	1kc/s
Spaced pulses, relay operation	50V	50V	2·5W	10c/s
Switching unit, variable frequency operation ..	12V	1V	0·5W	0·5 to 40c/s
Synchronous motor supply	20V	115V	20W	400c/s
Fluorescent lighting supply	12V	105V	30W	650c/s

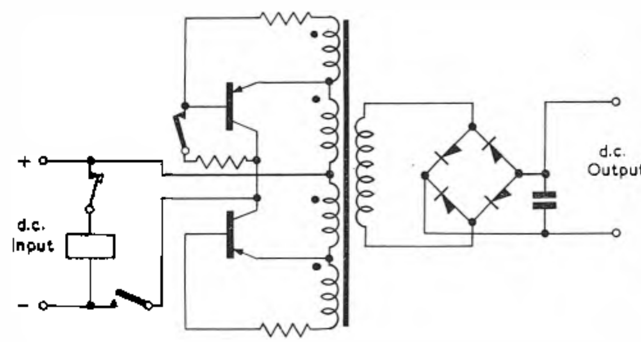


Fig. 5. Push-button relay starting circuit

Conclusion

Transistor convertors have the following advantages over other methods of stepping up the voltage of a low power d.c. source:

(a) Smaller size and reduced weight.

(b) Absence of contacts and moving parts.

(c) Reduced radio interference.

(d) Higher efficiency.

Without rectification the output can provide an ideal d.c. to a.c. invertor. It is also possible to obtain frequency conversion, e.g., 50c/s to 1kc/s by rectifying the mains supply for the input to a convertor. The square wave output is suitable for many applications and two- or three-phase supplies can readily be arranged.

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The Design Criteria of a

Common Aerial System for Simultaneous Transmission and Reception of V.H.F. Signals

By J. K. Grierson*, M.Sc.

A common aerial system that consists of five sections is described. The sections are a transmitter, a transmission coupling network, a receiver, a reception coupling network and an aerial. The transmission and reception coupling networks enable a single aerial to be used for transmission and reception of frequency-spaced signals without the necessity of time sharing. The coupling networks are considered in detail and practical values of components are assigned.

(Voir page 571 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 576)

THE common aerial system to be described performs the following functions. The transmitter produces a signal at a frequency f_1 which is fed via the transmission coupling network to the aerial where it is radiated. At the same time, the aerial receives a signal on a slightly higher frequency f_2 from a distant transmitter and feeds it to the receiver via the reception coupling network. The use of a single aerial gives advantages which include cost reduction, easier siting of aerials and, if the frequencies of transmission and reception are not too widely separated ($f_2 - f_1 \ll f_2, f_1$) close similarity of aerial patterns for transmission and reception. The techniques used in the construction of such a system depend on the frequency of the transmitter and the degree of separation between the receiver and transmitter frequencies. Two disadvantages of the system are a slight loss in radiated power and a slight deterioration of received signal-to-noise ratio.

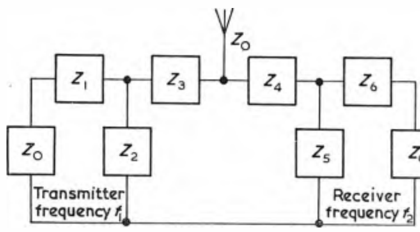

Fig. 1. The ideal system

The system to be described was originally a part of the Janet development programme[1]. It has been shown that observability of meteor scattered signals depends on antenna patterns[2] and it is obvious that such observability affects the operation of bi-directional systems based on Janet principles.

An Ideal System

An ideal system, based on the T-network configuration, is represented diagrammatically by Fig. 1, where both the input impedance of the receiver and the output impedance of the transmitter are represented by Z_0, a constant. The impedance of the antenna is also Z_0. The numbered impedances have the following characteristics; at the transmitter carrier frequency f_1, Z_1, Z_3, and Z_5 are zero while Z_2, Z_4 and Z_6 are infinite, at the receiver carrier frequeny, f_2, Z_1, Z_3, and Z_5 are infinite while Z_2, Z_4, and Z_6 are zero. Under these conditions, energy from the transmitter will be transferred to the aerial and radiated without interfering with the receiver, and at the same time energy will be transferred from the aerial to the receiver without interfering with the transmitter.

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In practice, the impedances Z_2 and Z_5 can be simple series-parallel tuned circuits, while Z_1, Z_3, Z_4, and Z_6, can be simulated by quarter wave transmission line sections. In the practical case no impedance can be made zero or infinity, but it can be made very small, $R_F = R_S \ll Z$, in the one case and very large, $R_F = R_p \gg Z$, in the other case, where R_F is the resistance of the filter in the general case, R_S is the resistance

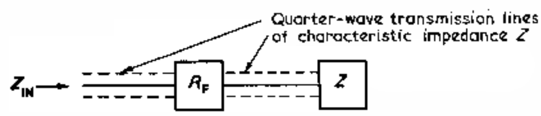

Fig. 2. Transmission line transformers

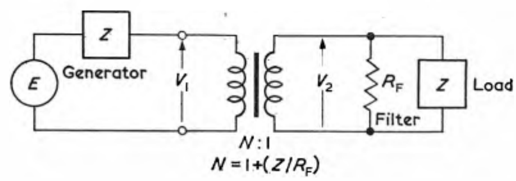

Fig. 3. The equivalent circuit

of the series resonant section, and R_P is the resistance of the parallel resonant section at their respective resonant frequencies. The theory which follows is not entirely original[3] but is useful as a collected set of ideas for understanding the complete system.

Transmission Line Transformers

As a preliminary review of the problem, a T-network will be considered, which replaces Z_5 by R_F and Z_4 and Z_6 by quarter wave transmission lines of characteristic impedance Z. In Fig. 2, the input impedance is given[4] by

$$Z_{in} = Z(1 + Z/R_F) \qquad \ldots\ldots\ldots (1)$$

At the transmitter frequency, f_1, $R_F \ll Z$, then Z_{in} becomes

$$Z_{in} \sim Z^2/R_s \qquad \ldots\ldots\ldots\ldots (2)$$

At the receiver frequency, f_2, $R_F \gg Z$, then Z_{in} becomes

$$Z_{in} \simeq Z \qquad \ldots\ldots\ldots\ldots (3)$$

Thus, the input impedance of this circuit is high at the frequency f_1 and equal to Z at the frequency f_2. Since the circuit is symmetrical, reversing Z to the input will give identical results for the output impedance. This circuit is then seen to be a simple way of contriving the reception network of Fig. 1 (Z_4, Z_5, and Z_6). The frequencies f_1 and f_2 must be sufficiently near together to consider the quarter wave sections of transmission line to be of constant electrical length. For the purposes of calculation, an equivalent circuit of Fig. 2 has been drawn up as Fig. 3. The generator, of output impedance Z and e.m.f., E, represents the aerial, and the load represents the receiver.

Rejection and Insertion Loss

The voltage across the load, V_2, in Fig. 3, is given by

$$V_2 = \frac{E}{Z/R_F + 2} \quad \ldots\ldots\ldots\ldots\ (4)$$

where the voltage at the input of the equivalent T replacement unit is shown as V_1. As $R_F \to \infty, V_2 \to E/2$. The ratio of the voltage across the load when R_F is infinite to the voltage across the load when R_F is finite is given by

$$\frac{(V_2)_\infty}{(V_2)_{RF}} = 1 + Z/2R_F \quad \ldots\ldots\ (5)$$

The attenuation of V_2 (in decibels) due to the insertion of the section of Fig. 3 between the generator and the load is then given by

$$A = 20 \log (1 + Z/2R_F) \quad \ldots\ldots\ldots\ldots\ (6)$$

The insertion loss is given by

$$I = 20 \log (1 + Z/2R_p) \quad \ldots\ldots\ldots\ldots\ (7)$$

where R_p is high at the frequency f_2.

The rejection of the unwanted signal at a frequency f_1 is given by

$$R = 20 \log (1 + Z/2R_s) \quad \ldots\ldots\ldots\ldots\ (8)$$

where R_s is low at the frequency f_1. The ratio of the rejection of the unwanted frequency to the insertion loss at the wanted frequency is given by

$$M = R/I = \frac{20 \log (1 + Z/2R_s)}{20 \log (1 + Z/2R_p)} \quad \ldots\ldots\ldots\ (9)$$

where M is the figure of merit of the inserted filter section, since M gives the value of rejection of the unwanted signal per unit of insertion loss of the wanted signal.

Stacking

The T replacement unit may be coupled to a similar unit. The coupling of successive units may be described as stacking. In stacking, however, only one of the two quarter wave transformers are used so that the sequence along any transfer network is generator, quarter wave line, filter, quarter wave line, load, etc. For a stack, certain modifications to equations (6), (7), (8) and (9) occur. For one section of an n section stack, the equations are used as given, for the remaining $n - 1$ sections, the equations are modified by writing R_F wherever $2R_F$ occurs in the original equations. A consideration of Fig. 3 will show why this is necessary. If the attenuation due to the insertion of the filter is given by the ratio of $\frac{V_2}{V_1}$ where V_1 is held constant, then equation (6) becomes

$$A = 20 \log (1 + Z/R_F) \quad \ldots\ldots\ldots\ldots\ (6a)$$

This will be true for $(n - 1)$ sections in an n section stack. The input impedance to a stack can be obtained by the repeated application of the principle which was used to obtain equation (1) and it will not differ very greatly from the input impedance of a single section. The rejection and insertion losses of the whole stack will be approximately the same as the sums of the rejection and insertion losses of the individual units, bearing in mind the remarks above (equation (6a)). These attenuations can be easily computed for the extended case of lossy lines by the use of the Smith chart[5].

Design of the Basic Filter

In the foregoing it has been assumed that a filter can be constructed such that

$$R_F = R_s \ll Z \text{ at a frequency } f_1 \text{ and}$$
$$R_F = R_p \gg Z \text{ at a frequency } f_2 > f_1.$$

The simple circuit illustrated in Fig. 4 has these properties: At the frequency f_1, given by

$$f_1 = \frac{1}{2\pi (LC_A)^{\frac{1}{2}}} \quad \ldots\ldots\ldots\ldots\ldots\ (10)$$

the series circuit will resonate and the impedance of the series branch will be R_s, a low resistance. At the frequency f_2, given by

$$2\pi f_2 L - \frac{1}{2\pi f_2 C_A} = \frac{1}{2\pi f_2 C_p} \quad \ldots\ldots\ldots\ldots\ldots\ (11)$$

the whole circuit will behave like a parallel tuned circuit. The Q of the series circuits is defined in the usual way by

$$Q_s = \frac{2\pi f_1 L}{R_s} \quad \ldots\ldots\ldots\ldots\ldots\ (12)$$

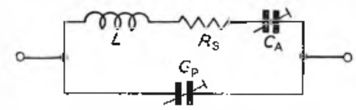

Fig. 4. The basic filter

The Q of the parallel circuit is given by

$$Q_p = \left(2\pi f_2 L - \frac{1}{2\pi f_2 C_A}\right) \Big/ R_s = \frac{1}{2\pi f_2 C_p R_s} \quad (13)$$

At the frequency f_2, the parallel circuit behaves like a high resistance R_P which is given by

$$R_p = Q_p X_L = Q_p X_c \quad \ldots\ldots\ldots\ldots\ldots\ (14)$$

where X_L and X_C are the effective inductive reactance of the series arm and the reactance of C_p at the frequency f_2. That is,

$$R_p = \left[\left(2\pi f_2 L - \frac{1}{2\pi f_2 C_A}\right) \Big/ R_s\right] \cdot \frac{1}{2\pi f_2 C_p} \quad (15)$$

Now equation (13) can be rewritten in the form

$$Q_p = Q_s/Q_k \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (16)$$

and equation (15) may be rewritten in the form

$$R_p = (2\pi f_1 L Q_s)/Q_k^2 \quad \ldots\ldots\ldots\ (17)$$

where Q_k is given by

$$Q_k = \frac{f_2 f_1}{f_2^2 - f_1^2} = \frac{f_2 f_1}{(f_2 + f_1)(f_2 - f_1)} \quad \ldots\ldots\ (18)$$

When $f_2 - f_1$ is small compared with f_2, equation (18) may be written

$$Q_k = \frac{f}{2(f_2 - f_1)} \quad \ldots\ldots\ldots\ldots\ (19)$$

Q_k is entirely fixed by the operating frequencies of the transmitter and receiver. The parameters open to choice, then, are, Q_s and L, since with a given Q_s and L, the components C_A, R_s, and C_p are determined. From equations (12) and (17), Q_s should be as high as possible in order that R_s be small and R_p large. The requirements for L are conflicting, since to make R_p large, L should be large, but to make R_s small, L should be small. The optimum L has to be determined experimentally since it is an unknown function of f and Q. Also the maximum Q_s is determined experimentally for a given L.

The foregoing has described the reception network. Similar reasoning applies to the transmission network except that the frequencies of series resonance (f_1) and parallel resonance (f_2) are reversed, hence it is necessary to change the series capacitor (C_A) to a new value (C_B) in order to series resonate L at the frequency f_2 and it is necessary to replace C_p with an inductance L_p in order to resonate the new series circuit in the parallel mode at the frequency f_1.

Since these changes necessitate rewriting some of the foregoing equations, a list of the equivalent equations for the transmission filter are given below:

$$f_2 = \frac{1}{2\pi(LC_B)^{\frac{1}{2}}} \qquad (10b)$$

$$2\pi f_1 L - \frac{1}{2\pi f_1 C_B} = -2\pi f_1 L_p \qquad (11b)$$

$$Q_s = \frac{2\pi f_2 L}{R_s} \qquad (12b)$$

$$Q_p = \frac{[-(2\pi f_1 L - (1/2\pi f_1 C_B))]}{R_s} = \frac{2\pi f_1 L_p}{R_s} \qquad (13b)$$

$$R_r = \frac{[-(2\pi f_1 L - (1/2\pi f_1 C_B))]}{R_s} \cdot 2\pi f_1 L_p \qquad (15b)$$

$$R_p = \frac{2\pi f_2 L Q_s}{Q_k^2} \qquad (17b)$$

Equations without equivalents are identical for both cases. It should also be noted that the distant station will have transmission and reception frequencies reversed.

Filter Component Ratings

The filter components must be able to withstand the feed conditions. It is necessary to know with what currents and voltages the individual components have to deal. Since the power output of the transmitter at the frequency f_1 and the insertion and rejection losses of the filters are known, it is easy to compute the power loss in each filter, and this need only be done at the transmitter frequency. Further, at the transmitter frequency every filter is nearly a pure resistance (the receiver filters will be slightly capacitive), the transmission filters will be parallel resonant and the reception filters will be series resonant. With these considerations in mind, if W_p is the required dissipation of a parallel filter, then V_p, the voltage across the filter is given by

$$V_p = \sqrt{(W_p R_p)} \qquad (20)$$

The feed current to the filter terminals is given by

$$i_f = V_p / R_p \qquad (21)$$

The circulating current in the parallel circuit is given by

$$i_c = Q_p i_f \qquad (22)$$

and the voltage rating of the series capacitor is given by

$$V_c = i_c/(2\pi f_1 C_B) \qquad (23)$$

Proceeding to the reception filter, if W_s is the power dissipation then the voltage across the circuit is given by

$$V_s = \sqrt{(W_s R_s)} \qquad (24)$$

The series current is given by

$$i_s = V_s / R_s \qquad (25)$$

and the voltage across the series capacitor is given by

$$V_c = i_s/(2\pi f_1 C_A) \qquad (26)$$

This concludes the theoretical discussion of the problem.

Experimental Filters

The design of a practicable filter has certain arbitrary features. Initially, the frequencies of transmission and reception will be assigned by the user or designer. Equipment will likewise fix Z. When a value of L has been decided, C_A, C_B, C_p and L_p will be determined. The only other independent variable, apart from maximum ratings, will be Q_s, which will be as high as possible. From the foregoing, the choice of L is the only serious decision to make, since everything else depends on this value.

The value of L must lie in the range such that C_A, C_B or L_p are not impracticably small nor C_p or L_p impracticably large. Having determined L, it remains to try to obtain the highest Q_s for the given frequency; this must be done mainly by experiment.

The filters which were constructed, had the following characteristics:

(a) *Reception Filter*

$L = 1{\cdot}5\mu$H, $C_A = 10{\cdot}3$pF, $C_p = 208$pF, $Q_s = 450$, $R_s = 0{\cdot}85\Omega$, $Q_p = 22$, $R_p = 409\Omega$, $R = 29{\cdot}7$dB, $I = 0{\cdot}5$dB, $M = 58$, R(for $n-1$ sections) $= 39$dB, Input Impedance $Z_{in} = 2990\Omega$, (at f_1) $Z_{in} = 56{\cdot}1\Omega$ (at f_2).

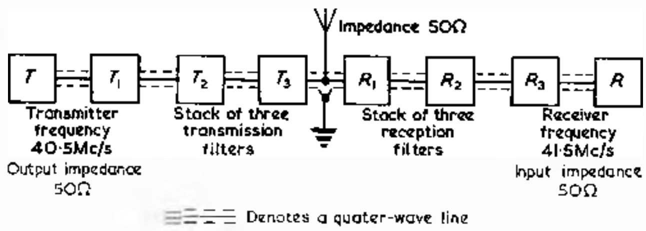

Fig. 5. A practical system

(b) *Transmission Filter*

$L = 1{\cdot}5\mu$H, $C_B = 9{\cdot}8$pF, $L_P = 75$mμH, $Q_s = 0{\cdot}85\Omega$, $Q_p = 22$, $R_p = 414\Omega$, $R = 30$dB, $I = 0{\cdot}5$dB, $M = 58$, R (for $n-1$ sections) $= 36$dB, Input impedance $Z_{in} = 56\Omega$ at f_1, $Z_{in} = 2990\Omega$ (at f_2).

(c) *Common Characteristics*

$f_1 = 40{\cdot}5$Mc/s, $f_2 = 41{\cdot}5$Mc/s, $Q_k = 22{\cdot}5$, $Z = 50\Omega$.

A Practical System

The system to be described is illustrated in Fig. 5. The transmitter, operating on 40·5Mc/s, is connected via a stack of three T sections incorporating transmission filters to the aerial. The receiver, operating on 41·5Mc/s, is connected via a stack of three T sections incorporating reception filters to the aerial.

The output impedance of the transmitter, the input impedance of the receiver, the aerial impedance and the characteristic impedance of the transmission lines, are all 50Ω. The analysis of the system will be divided into sections to assess the relevant properties.

(a) The Decrease of Radiated Power

Power available for radiation will be reduced due to (1) the insertion of the transmitter filter, (2) the connexion of the receiver filter, terminated in the receiver, to the aerial and (3) to the mismatch produced at the aerial and at the input to the transmitter filter. The power lost due to mismatching will be small, since the transmitter output impedance can be adjusted.

The input impedance to the receiver filter, terminated with the receiver input impedance, is approximately 3 000Ω at 40·5Mc/s, when viewed from the aerial. The output end of the transmitter filter looks into the 50Ω aerial impedance, shunted by 3 000Ω. This is approximately 49Ω. The input impedance to the transmitter filter, when viewed by the transmitter, is therefore 54·5Ω. The ratio of the output voltage at the aerial to the voltage at the transmitter output terminals is 1 : 1·25. Consequently, the power ratio is 1 : 1·40. This takes the relevant impedances into account. The ratio of the power radiated by the aerial to the input power to the transmitter filter is therefore 1 : 1·42, since the receiver filter absorbs some power. The loss of radiated power, due to the connexion of the filter system, is therefore 1·5dB. The bulk of this loss occurs in the transmitter filter, with the first section of the receiver filter, nearest to the aerial, accounting for most of the remainder. This calculation does not take account of the power dissipated in the transmitter, but the loss within the transmitter can be made small by readjusting the transmitter output impedance.

On the basis of a 1kW transmitter, 680W will be radiated; some 20W will be dissipated by the first section of the receiver filter, and each section of the transmitter filter will absorb some 100W. It can be seen that the figure of 1·5 insertion loss is consistent with the statement that the insertion loss of a stack of filters is approximately the sum of the individual losses due to each filter, since the transmission filter has an insertion loss of 0·51dB.

If the transmitter output impedance is considerably less than the load impedance an adjustment of the insertion loss of the filter (equation (7)) will be necessary.

(b) Spurious Output from the Transmitter

The transmitter will probably have a spurious output signal at 41·5Mc/s which will interfere with the receiver operation. This signal may take the form of wide band noise or spurious voltages due to non-linearities in the transmitter parameters. It is assumed that the transmitter has a spurious output signal at 41·5Mc/s which is at least 90dB below the rated output of 1kW at 40·5Mc/s. The transmitter filter will attenuate this output by a further 101dB. The spurious signal will then be 191dB down on 1kW at the aerial, that is, $7{\cdot}9 \times 10^{-17}$W. The sensitivity of a receiver is limited by noise. At 41·5Mc/s the predominant noise is cosmic in origin, which is considerably higher than all other sources of noise combined. Provided man-made noise is low in the area the cosmic noise input to the reciver is about $1{\cdot}38 \times 10^{-15}$W and the spurious signal should be 12dB below this limiting threshold noise.

(c) The Rejection of the Transmitter Signal

The transmitter signal at the antenna is 680W at 40·5Mc/s. This corresponds to an r.m.s. voltage of 187V. The receiver rejection filter reduces this by 98dB. The unwanted signal at 40·5Mc/s at the receiver input is then 2·4mV. This signal is sufficiently small to be rejected by the selectivity of the receiver.

(d) The Insertion Loss of the Receiver Filter

Cosmic noise fixes the threshold sensitivity of the receiver only if it is greater than the receiver noise. Once the receiver noise level is approached, further reduction of the cosmic noise and desired signal will seriously degrade the signal-to-noise ratio. Cosmic noise is approximately 9dB higher than receiver noise and the insertion loss of the receiver filter is approximately 1·5dB.

(e) The Component Ratings

Straightforward application of equations (20) to (26), together with the power dissipations of the transmitter and receiver filters given above, give the ratings below.

The transmitter filters have a dissipation of 100W. The voltage across the filter terminals is 203V r.m.s., the feed current is 0·49A, r.m.s., the circulating current is 10·8A, r.m.s., and the capacitor voltage is 4 130V r.m.s. (5 850V peak). The receiver filter adjacent to the aerial has a dissipation of 20W. The voltage across the filter is 4·1V r.m.s., the current through the filter is 4·9A r.m.s., and the voltage across the series capacitor is 1 850V r.m.s. (2 600V peak).

(f) Connecting Lines

A slight degradation of performance due to the losses of connecting lines will be experienced. This is dependent on the type of lines used and was found to be small in the experimental cases tested.

Conclusions

It has been shown that a receiver operating on 41·5Mc/s and limited by cosmic noise may use an aerial which is radiating 680W from a 1kW transmitter operating at 40·5Mc/s.

In the text a single aerial has been considered, in practice, an array may be used. Helix arrays[6] are very suitable for narrow beam applications. In the Janet System, Yagis were used.

Note: The complete system was not tested as such. Further experimental verification would be necessary to determine the complete validity of the design criteria. Most aspects of the design were tested in sections.

Acknowledgments

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PICTURES ACROSS THE ATLANTIC

Experimental transmission of news film between England and Canada has recently taken place. After consideration of the characteristics of the Atlantic cable, a film scanning apparatus was devised by the BBC Engineering Division which could be operated in conjunction with Standard Telephones and Cables Ltd Music-in-Band Transmission Equipment already installed in Oban, Montreal and New York.

STC music transmission equipment normally permits the operation of high-grade circuits which displace either 2 or 3 speech circuits in any one twelve circuit group operated over the carrier system. A music circuit displacing 2 speech circuits has a frequency bandwidth of 50 to 6 400c/s and one displacing 3 circuits, 50 to 10 000c/s.

The smaller of the two bandwidths was chosen for transmitting the pictures. In order to limit the variation in the group delay/frequency characteristic to a value which could be corrected, it was necessary to restrict the useable video bandwidth to 4·5kc/s.

Vestigial sideband transmission was used with a special form of negative-going amplitude modulation. The carrier was 5kc/s and the whole of the lower sideband was transmitted, the vestige of the upper sideband extending from 5kc/s to 5·5kc/s.

Local Feedback in Transistor Amplifiers

By H. Pfyffer*, dipl. Ing. ETH.

In this article the common emitter amplifier stage is first discussed. The effects of negative feedback on gain, impedance and cut-off frequency are then calculated, and theoretical calculations are compared with measured results. Some of these comparisons are illustrated graphically.

(Voir page 571 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 576)

In many applications transistors are used in the common emitter configuration. This has the advantage of a current and a voltage gain higher than unity. In contrast the common base stage produces a voltage gain which is higher than unity, and the common collector gives current gain only. The disadvantage of the common emitter stage is the considerable spread of the parameters among different transistors of the same type, specially the short-circuit current gain is likely to change over a wide range. Thus the other characteristics, such as the impedances and the voltage gain, are due to vary as well, as they depend on the short-circuit current gain.

The basic behaviour of the transistor is represented by the common base arrangement, to which the current gain α is normally related. Changes of α are normally confined to a few per cent. The effect of this small change is very large in the common emitter configuration. This can be seen from the following calculation:

$$\beta = \frac{\alpha}{1 - \alpha} \quad \ldots\ldots\ldots\ldots \quad (1)$$

$$d\beta/\beta = d\alpha/\alpha\ (1 + \beta) \quad \ldots\ldots\ldots \quad (2)$$

If a transistor is assumed to have an $\alpha = 0{\cdot}98$, it follows from the formulae $\beta = 49$, and the variation of β is approximately 50 times as much as it is for the common base circuit.

Negative feedback is normally used to stabilize an amplifier against change of transistors. Feedback may be applied as overall or local feedback or as a combination of both. The feedback may modify the gain, the impedances, the frequency characteristic and the bandwidth. The effects of the local feedback are discussed in this article.

Symbols Used

R_b	=	Base resistance
R_d	=	Collector resistance
R_e	=	Emitter resistance
R_F	=	Voltage feedback resistor
R_L	=	Load resistor
R_0	=	$R_L + R$
R_s	=	Source resistor
R	=	Current feedback resistor
Z_i	=	Input impedance with feedback
Z_{i0}	=	Input impedance without feedback
Z_o	=	Output impedance
α	=	Common base short-circuit current gain
β	=	Common emitter short-circuit current gain
γ	=	Current gain with load
μ	=	Voltage gain
f_α	=	Common base cut-off frequency
f_β	=	Common emitter cut-off frequency
f_γ	=	Cut-off frequency with feedback

* *The General Electric Co. Ltd., Coventry*

The Equivalent Circuit for the Common Emitter Stage

Source and Load

The transistor can be represented by a four terminal network which is fed at the input by a source with the source impedance R_s, and terminated at the output with a load R_L.

The two circuits of Fig. 1 are equivalent.

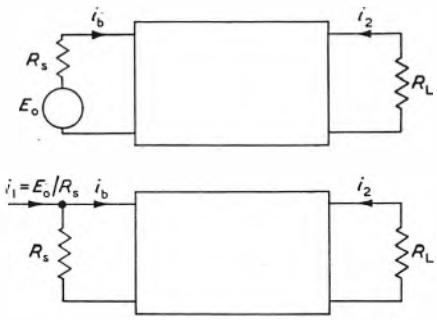


Fig. 1. The four terminal network with source and load

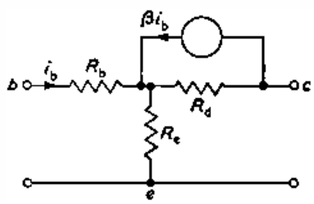


Fig. 2. The *T*- type equivalent circuit

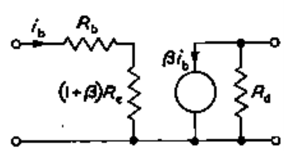


Fig. 3. Simplified *T*- type equivalent circuit

Definitions

Short-circuit current gain	:	$(i_2/i_1)_{V_2=0}$	$= \beta$
Current gain with load R_L	:	(i_2/i_1)	$= \gamma$
Voltage gain	:	(V_2/V_1)	$= \mu$
Power Gain	:	$\mu \cdot \gamma$	$= K$

The Transistor

One of the equivalent circuits of the transistor is given in Fig. 2.

The four parameters depend on the d.c. conditions, especially R_e which changes with the emitter current:

$$R_e = (12{\cdot}5/I_e) \text{ ohms } (I_e \text{ in mA}) \quad \ldots\ldots \quad (3)$$

In most cases $R_e \ll R_d$, even if the emitter resistor is increased by adding an external resistor to provide feedback. Thus, the circuit can be redrawn as shown in Fig. 3.

In a wide range of applications this simplification gives a result which can be regarded as a good approximation to the practical results. The following calculations are mostly based on the simplified circuit in Fig. 3.

Series Current Feedback

The Circuit

The block diagram for this type of feedback is represented as shown in Fig. 4. For the practical circuit this means the insertion of a resistor in the emitter lead, as it is shown by Fig. 5.

Fig. 6 represents the equivalent circuit diagram with feedback. Applying node and mesh analysis all the required formulae can be obtained.

THE CURRENT GAIN

Solving the equations for the circuit gives the expression for the current gain:

$$\gamma = i_2/i_1 = \frac{(R_o/R_d - \beta}{(1+(R_o/R_d)+R_L/R_d)\left(1+\dfrac{R_b+R_o}{R_s}\right)-(R_o/R_s)((R_o/R_d)-\beta)} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

where $R_o = R_e + R$.

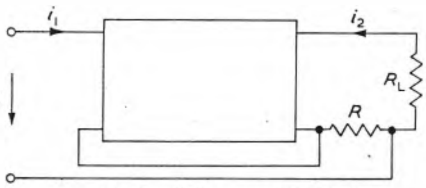


Fig. 4. Series current feedback

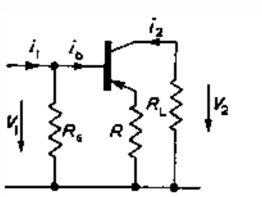


Fig. 5. Amplifier stage with series current feedback

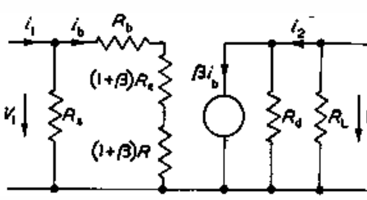


Fig. 6. Simplified equivalent circuit of the amplifier stage with current feedback

Assuming some simplifications, such as

$$R_L < R_d \text{ and } R_o/R_d \ll \beta \quad \ldots\ldots\ldots\ldots (4a)$$

(the first one is specially considered in wide band amplifiers) the expression can be rewritten:

$$\gamma \simeq \frac{-\beta}{1+(1+\beta)\ R_o/R_s\ +\ R_b/R_s} \quad \ldots\ldots\ldots\ldots (5)$$

$$\gamma_o \equiv \frac{-\beta}{1+(1+\beta)\ R_e/R_s\ +\ R_b/R_s} \quad \ldots\ldots\ldots\ldots (6)$$

and finally taking into account that $(1+\beta)R_e + R_b$ represents the input impedance without feedback Z_{io}, the formula becomes:

$$\gamma = \frac{-\beta}{(1+(Z_{io}/R_s))\left(1+\dfrac{(1+\beta)R}{R_s(1+Z_{io}/R_s)}\right)} \quad \ldots\ldots (7)$$

It can now easily be seen that there is always a certain amount of feedback, even in the case of $R = 0$, due to the finite value of R_e and to the fact that the source impedance is not of an extremely high value.

In the case of a very high value of β the current gain is then given by:

$$\gamma \simeq \frac{\gamma_o}{1-\gamma_o(R/R_s)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (8)$$

(γ_o is a negative value).

where γ_o is the gain which results by the use of a definite source with a transistor (see equation (6)).

THE VOLTAGE GAIN

Using the same methods as in the preceding case, it follows that:

$$\mu = V_2/V_1 = \frac{-\beta R_L}{Z_i(1+(R_L/R_d))} \quad \ldots\ldots\ldots\ldots (9)$$

where $Z_i = Z_{io} + (1+\beta)R$.

It is obvious that the voltage gain decreases with increasing feedback, since Z_i depends on the feedback resistor.

Assuming again R_L to be much smaller than R_d, a simplified formula is obtained.

$$\mu = \frac{-\beta R_L}{Z_i} = \frac{-\beta R_L}{Z_{io}(1+(\beta R/Z_{io}))} \quad \ldots\ldots\ldots\ldots (10)$$

THE POWER GAIN

The power gain can be represented as the product of the current and the voltage gain.

$$K = \gamma \cdot \mu$$

If the simplifications (4a) are justified, it follows that:

$$K = \frac{\beta^2 R_L}{Z_{io}(1+Z_{io}/R_s)\left[1+\left(\beta R/Z_{io}+\dfrac{2\beta R}{R_s}+\dfrac{\beta^2R^2}{R_sZ_{io}}\right)\left(\dfrac{1}{1+Z_{io}/R_s}\right)\right]} \quad \ldots\ldots\ldots\ldots (11)$$

where $Z_{io} \doteq R_b + (1+\beta)R_e$ represents the input impedance without feedback.

THE REQUIRED FEEDBACK RESISTOR

Comparing the expressions for the current and the voltage gain (8) and (9) respectively with the general formula for a feedback system:

$$G = \frac{G_o}{1+G_oF} \quad \ldots\ldots\ldots\ldots\ldots\ldots (12)$$

it is obvious that $(1+G_oF)$ must be equal to the factors

$$1-\gamma_o(R/R_s) \text{ and } 1+(\beta R/Z_{io})$$

If therefore a certain amount of feedback is required, the required value of R can be determined. If A is the factor, by which the gain has to be reduced (20 log A is the applied feedback in decibels), it follows that:

(a) Current gain: $$R = \frac{A-1}{\beta}(R_s+Z_{io}) \quad \ldots\ldots (13)$$

(b) Voltage gain: $$R = \frac{A-1}{\beta}Z_{io} \quad \ldots\ldots\ldots\ldots (14)$$

THE INPUT IMPEDANCE

The input impedance is increased by applying emitter feedback. The formula has already been used in the previous sections:

$$Z_i = Z_{io} + (1+\beta)R \text{ where } Z_{io} = R_b + (1+\beta)R_e \quad \ldots\ldots (15)$$

THE OUTPUT IMPEDANCE

The output impedance is not much affected by this type of feedback. According to Fig. 2 the formula can be obtained as follows:

$$Z_o = R_d + \frac{R_o(R_s+R_b+\beta R_d)}{R_o+R_s+R_b} \quad \ldots\ldots\ldots (16)$$

In most cases $R_s + R_b$ is much smaller than βR_d, therefore the impedance can be rewritten in the form:

$$Z_o = R_d\left(1+\frac{\beta R_o}{R_o+R_s+R_b}\right) \quad \ldots\ldots\ldots\ldots (17)$$

The second term in the brackets is likely to be smaller than unity and thus the impedance is only slightly increased.

Parallel Voltage Feedback

THE CIRCUIT

For this type of feedback the block diagram is shown in Fig. 7 and the practical circuit by Fig. 8.

The equivalent circuit which is used is shown in Fig. 9. The effect of this feedback can be calculated by applying node and mesh analysis.

The Current Gain

The resulting current gain is:

$$\gamma = \frac{1 - (1+\beta) R_F/(R_F + Z_{io})}{1 + (R_L/R_d) + \dfrac{(1+\beta)R_L}{R_F + Z_{io}}} \quad \text{(18)}$$

If $\beta \gg 1$ and $R_F \gg Z_{io}$:

$$\gamma \simeq \frac{-\beta}{1 + (1+\beta) R_L/R_F} \quad \text{(19)}$$

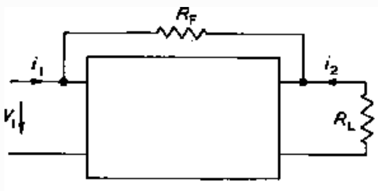


Fig. 7. Parallel voltage feedback

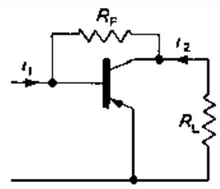


Fig. 8. Amplifier stage with parallel voltage feedback

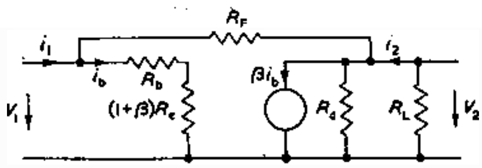


Fig. 9. Simplified equivalent circuit of the amplifier stage with parallel voltage feedback

The Voltage Gain

The voltage gain is obtained thus:

$$\mu = \frac{((Z_{io}/R_F) - \beta)}{Z_{io}(1/R_d + 1/R_L + 1/R_F)} \quad \text{(20)}$$

If $\beta \gg Z_{io}/R_F$ and $R_d \gg R_L$:

$$\mu \simeq \frac{-\beta R_L}{Z_{io}(1 + R_L/R_F)} \quad \text{(21)}$$

The Power Gain

The power gain is given by:

$$K = \frac{((Z_{io}/R_F) - \beta)\left(1 - (1+\beta)\dfrac{R_F}{R_F + Z_{io}}\right)}{Z_{io}(1/R_d + 1/R_L + 1/R_F)\left(1 + (R_L/R_d) + \dfrac{(1+\beta)R_L}{R_F + Z_{io}}\right)} \quad \text{(22)}$$

This formula can not easily be surveyed, since it is very complicated. But in many cases several simplifications can be assumed.

If $R_L \ll R_d$, $R_L \ll R_F$, $Z_{io} \ll \beta R_F$, $\beta \gg 1$, then

$$K = \frac{\beta\,(\beta\,(R_F/(R_F + Z_{io})) - 1)R_L}{Z_{io}\left(1 + \dfrac{\beta R_L}{R_F + Z_{io}}\right)} \quad \text{(23)}$$

If $Z_{io} \ll R_F$ and $\beta R_F/(R_F + Z_{io}) \gg 1$, the power gain can be rewritten as:

$$K \simeq \frac{\beta^2 R_L}{Z_{io}(1 + \beta(R_L/R_F))} \quad \text{(24)}$$

The Required Feedback Resistor

As mentioned previously, the resistor, which is required to give the wanted amount of feedback, can be calculated, using the equations (19) and (21).

(a) Current gain: $R_F = \dfrac{\beta R_L}{A - 1}$ .... (25)

(b) Voltage gain: $R_F = \dfrac{R_L}{A - 1}$ .... (26)

The Input Impedance

The input impedance is reduced by this type of feedback, as can be seen from the following formulae:

$$Z_i = \frac{Z_{io}(1 + R_L/R_F + R_L/R_d)}{1 + (1+\beta) R_L/R_F + Z_{io}/R_F(1 + (R_L/R_d)) + R_L/R_d} \quad \text{(27)}$$

Making the same assumptions as for power gain, the simplified expression is obtained:

$$Z_i \simeq \frac{Z_{io}}{1 + \beta(R_L/R_F)} \quad \text{(28)}$$

It is obvious that the impedance is reduced.

The Output Impedance

The formula for this is:

$$Z_o = \frac{R_d}{1 + R_d/R_F\left(\dfrac{\left(1 + \beta\dfrac{1}{1 + Z_{io}/R_s}\right)}{1 + Z_{io}/R_F \cdot \dfrac{1}{1 + Z_{io}/R_s}}\right)} \quad \text{(29)}$$

As it is not very useful in practice, two extreme cases are considered:

(a) Constant voltage control: $R_s = 0$

$$Z_o = \frac{R_d}{1 + (R_d/R_F)} \quad \text{(30)}$$

(b) Constant current control: $R_s = \infty$

$$Z_o = \frac{R_d}{1 + (R_d/R_F)\left(\dfrac{1+\beta}{1 + Z_{io}/R_F}\right)} \quad \text{(31)}$$

and if in addition $R_F \gg Z_{io}$ and $\beta \gg 1$ then:

$$Z_o = \frac{R_d}{1 + \beta(R_d/R_F)} \quad \text{(32)}$$

Additional Current Feedback

In practical circuits, where a finite source impedance is used, in addition to the voltage feedback some current feedback occurs. The transistor to which voltage feedback is applied, is regarded as a new device with a definite current gain and input-impedance. The effect of the current feedback can then be obtained in the same way as for a single transistor, by introducing the modified values of i_2/i_1 and Z_i into the formulae for the current feedback. Using the simplified expressions:

$$\gamma_F = \frac{\beta}{1 + \beta(R_L/R_F)},\quad Z_i = \frac{Z_{io}}{1 + \beta(R_L/R_F)} \text{ into } \gamma_c = \frac{\beta}{1 + (Z_i/R_s)}$$

formula for the combined feedback is obtained:

$$\gamma = \frac{\beta}{1 + \beta(R_L/R_F) + (Z_{io}/R_s)} \quad \text{(33)}$$

$$R_F = \frac{\beta R_L}{A - 1 - (Z_{io}/R_s)} \quad \text{(34)}$$

The Frequency Response

There are several ways of representing transistor behaviour as a function of the frequency. One method is to assume complex impedances (resistors and capacitors) in the equivalent circuit instead of pure resistors. In this case the parameters are assumed to be constant. Another method is to represent the frequency dependence by a variable current generator and pure resistances in the circuit*. This method will be discussed in this article. Although it has been found that the first method gives a better approximation to the effec-

** A capacitor across the emitter resistance, which is determined by $C_e = 1/(2\pi R_e f_\alpha)$, is neglected in this article, as the influence in the considered frequency range is negligible.*

tive behaviour at very high frequencies, (higher than cut-off frequency), the resulting formulae are very useful to the designer, as they are fairly easy to handle and are easily surveyed. A further justification is the fact, that the parameters vary considerably among different samples of the same type and the accuracy of the formulae is within the accuracy of the available data on the transistors.

As stated in the literature, the common base current gain in function of the frequency can be assumed as follows:

$$\alpha = \alpha_0 \frac{\exp(jmf/f_\alpha)}{1 + jf/f_\alpha} \ldots\ldots\ldots\ldots\ldots (35)$$

As $f < f_\alpha$ and $m < 1$, the influence of the numerator will be neglected in this article. f_α is the cut-off frequency for the common base circuit (3dB point or corner in the semi-infinite slope diagram). Using the relations:

$$\beta = \frac{\alpha}{1-\alpha},\ \alpha = \frac{\alpha_0}{1 + jf/f_\alpha}$$

one obtains the impression for the common emitter circuit:

$$\beta = \frac{\beta_0}{1 + jf/f_\beta} \ldots\ldots\ldots\ldots\ldots\ldots (36)$$

and

$$f_\beta = f_\alpha(1-\alpha_0) \simeq f_\alpha/\beta \ \ldots\ldots\ldots\ldots (37)$$

This relation will be used in the following sections.

Series Current Feedback

The Current Gain

Using the expression:

$$\gamma = \frac{-\beta}{(1+(Z_i/R_s)) + R_L/R_d\left(\frac{R_b + R_e}{R_s}\right) + R_e/R_d} = \gamma_1 \ldots\ldots\ldots\ldots (38)$$

for the current gain (which is obtained from equation (4) by assuming $R_e/R_d \ll \beta$ and the relation (36) for β), it can be obtained that:

$$\gamma = \frac{\gamma_1}{1 + jf/f_\beta\left[\frac{1 + R_b/R_s + R_L/R_d\left(\frac{R_b+R_e}{R_s}\right) + R_L/R_d + R_e/R_s}{1 + Z_i/R_s + R_L/R_d(1+(R_b+R_e)/R_s)}\right]} \ldots\ldots\ldots\ldots (39)$$

The cut off frequency is obviously affected by the feedback and is:

$$f_\gamma = f_\beta \cdot \frac{1 + Z_i/R_s + R_L/R_d(1+(R_b+R_e)/R_s)}{1 + \frac{R_b + R_e}{R_s} + R_L/R_d\left(1 + \frac{R_b + R_e}{R_s}\right)} \ldots (40)$$

As the numerator is always larger than the denominator, the bandwidth is increased. The simplified formula shows roughly the effect of feedback:

$$f_\gamma \simeq f_\beta \cdot \left(1 + \frac{\beta_0}{1 + (R_s/R_e)}\right) \ldots\ldots\ldots (41)$$

The maximum bandwidth is obtainable in the case of unity gain. This case is more of theoretical interest; it can be shown that:

$$f_{\gamma(\max)} = \frac{1+\beta_0}{2} \cdot f_\beta \ \ldots\ldots\ldots\ldots (42)$$

for an ideal transistor with $R_b = 0$ and $R_L = 0$. This value corresponds to half the common base cut-off frequency. In practical circuits this value is not obtainable.

If the last two conditions are true, the formulae for the gain and the bandwidth can be derived as follows:

$$\gamma = \frac{-\beta_0}{\left(1 + \frac{(1+\beta_0)R_e}{R_s}\right)\left(1 + jf/f_\beta \cdot \frac{1}{1 + \frac{(1+\beta_0)R_e}{R_s}}\right)}$$

$$= \frac{-\beta_0}{A(1 + j\,f/(f_\beta \cdot A))} \ldots\ldots\ldots\ldots (43)$$

$$f_\gamma = A \cdot f_\beta \ \ldots\ldots\ldots\ldots\ldots\ldots (44)$$

$$A = 1 + \frac{(1+\beta_0)R_e}{R_s} \ \ldots\ldots\ldots\ldots (44a)$$

A is the factor by which the gain is reduced (20log A is the feedback applied in decibels). It is at the same time the factor, by which the bandwidth is increased. But for a better approximation R_b has to be considered.

To illustrate the results of this section the gain/frequency response of a transistor with various amounts of feedback will be plotted.

The following data are assumed:

Transistor: $\beta = 100$, $R_e = 20\Omega$, $R_b = 100\Omega$, $f_\beta = 100$kc/s.
Source impedance: $R_s = 5\text{k}\Omega$.

Equation (8) is used for the gain and (41) for the cut-off frequency.

Fig. 10 shows the curves plotted as semi-infinite slope characteristics. The broken line represents the locus of the 3dB points.

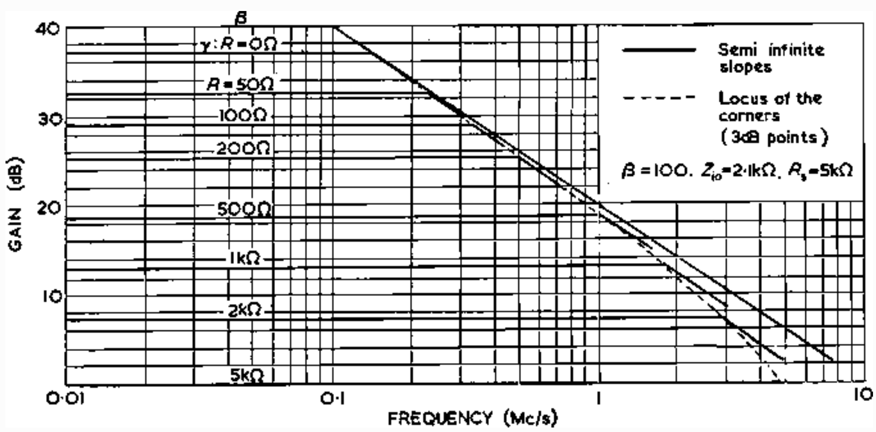


Fig. 10. Current gain with emitter feedback as a function of frequency

The Voltage Gain

The voltage gain as a function of frequency is:

$$\mu = \frac{-\beta_0}{Z_i(1 + (R_L/R_d))\left(1 + j\,f/f_b \frac{R + R_b}{Z_i}\right)} \ldots (45)$$

and the cut-off frequency is given by:

$$f_\mu = f_\beta \cdot \frac{R_b + (1 + \beta_0)R}{R_b + R} \quad \ldots\ldots\ldots\ldots (46)$$

For high values of R, f_μ tends to be equal to the common base value. But if R becomes high, the assumption $R \ll R_d$ is no longer true, and therefore this value is never obtained.

Parallel Voltage Feedback

General

Useful design formulae are obtained with the specifications:

$$R_d \gg R_L,\ Z_{io} \ll R_F$$

The Current Gain

The current gain as a function of the frequency can be represented by [(according to (18)) $\gamma_0 = \gamma$ given in (18)]

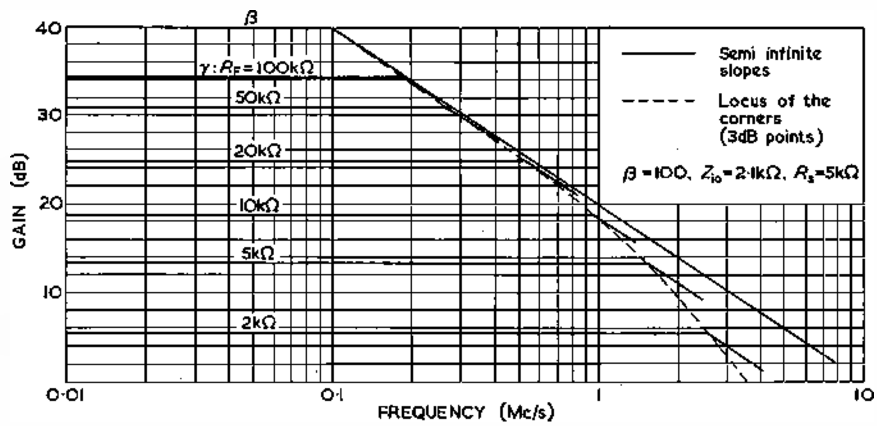


Fig. 11. Current gain with voltage feedback as a function of frequency

$$\gamma = \frac{\gamma_0 \left(1 - \mathrm{j}\, f/\beta_0 f_\beta \cdot \dfrac{(R_b + R_0)/R_F}{1 - (Z_{io}/\beta R_F)}\right)}{\left(1 + \mathrm{j}\, f/f_\beta \dfrac{1 + R_L/R_F + (R_b + R_0)/R_F}{1 + Z_{io}/R_F + (1 + \beta_0)R_L/R_F}\right)} \quad \ldots (47)$$

and the cut-off frequency:

$$f_\gamma = f_\beta \frac{1 + Z_{io}/R_F + \dfrac{(1 + \beta_0)R_L}{R_F}}{1 + R_L/R_F + \dfrac{R_b + R_0}{R_F}} \quad \ldots\ldots\ldots\ldots (48)$$

If R_L is much smaller than R_F, the formulae can be re-written in the form:

$$\gamma = \frac{-\beta_0}{(1 + (1 + \beta_0)\ R_L/R_F)\left(1 + \mathrm{j}\ f/f_\beta \dfrac{1}{1 + (1 + \beta_0)\ R_L/R_F}\right)} \quad \ldots\ldots\ldots\ldots (49)$$

$$\gamma = \frac{-\beta_0}{A(1 + \mathrm{j}\, f/f_\beta \cdot 1/A} \quad \ldots\ldots\ldots (50)$$

$$A = 1 + (1 + \beta_0)\ R_L/R_F \quad \ldots\ldots\ldots (50a)$$

where A represents the amount of feedback. (A itself is the gain reducing factor and $20\log A$ is the feedback in decibels).

In formula (47) a factor appears in the numerator of the form $(1 - \mathrm{j}X)$. It is represented in the semi-infinite slope method by an amplitude increase and a phase lag. Usually an amplitude increase is associated with a phase lead, but due to the negative imaginary part there occurs a phase lag. This form has its physical justification. Considering a very low feedback impedance R_F, it is obvious that the signal is going through R_F from the base to the collector instead of going through the transistor, and therefore the phase reversing properties of the transistor are then cancelled. In most practical cases this non-minimum phase behaviour may be neglected, as the corner for the factor $(1 - \mathrm{j}X)$ occurs at very high frequencies, but it has to be observed in multi-stage feedback amplifiers, which have to have phase margins.

As for the series case, a set of curves is plotted. The transistor is assumed to be the same. For the gain the formula (18) is used and for the bandwidth expression (48). (Fig. 11).

The Voltage Gain

For the voltage gain the expression is obtained as:

$$\mu = \frac{-\beta_0 R_L}{Z_{io}(1 + (R_L/R_F))\left(1 + \mathrm{j}\ f/f_\beta \dfrac{R_b + R_e}{Z_{io}}\right)} \quad \ldots (51)$$

and the cut-off frequency is:

$$f_\mu = f_\beta \cdot \frac{Z_{io}}{R_b + R_e} \quad \ldots\ldots\ldots\ldots\ldots (52)$$

The bandwidth is not affected by the feedback.

Influence of Current Feedback

The same method is used as previously to obtain the additional feedback influence of the source impedance. There follows:

$$f_\gamma = f_\beta(1 + (1 + \beta_0)\ R_L/R_F)\ \left(\frac{1 + (1 + \beta_0)R_L/R_F + Z_{io}/R_s}{1 + (1 + \beta_0)R_L/R_F + \dfrac{R_b + R}{R_s}}\right) \quad \ldots\ldots\ldots\ldots (53)$$

Examples

As a part of a wideband amplifier a single stage will be considered. The following specifications are to be met: Current gain $\gamma = 10 = 20$dB. Source impedance: $R_s = 600\Omega$. Load impedance $R_L = 600\Omega$.

The available transistor is assumed to have the following data:

$\alpha = 0{\cdot}975$, $R_b = 100\Omega$, $R_e = 25\Omega$, $R_d = 10\mathrm{k}\Omega$, $f_\alpha = 20$Mc/s.

The common emitter stage with current feedback will be considered.

For this type of circuit it is found:

$$\beta_0 = \frac{\alpha}{1 - \alpha} = \frac{0{\cdot}975}{0{\cdot}025} = 40 = 32\text{dB}.$$

$$f_\beta = f_a(1-\alpha) = 500\text{kc/s}.$$

As R_L is much smaller than R_d the simplified formula (5) is used for γ.

$$\gamma = \frac{\beta_o}{1 + \dfrac{R_b(1+\beta_o)R_o}{R_s}} = 20\text{dB}$$

The amount of feedback to be applied is therefore the difference between γ and β_o which is 12dB. Using the formula (13) to calculate the required feedback resistor it follows that:

$$R = \frac{A-1}{\beta_o}(R_s + R_b + (1+\beta_o) = 16{\cdot}5\Omega$$

The cut-off frequency which can be expected is, according to expression (40), $3{\cdot}25 \,.\, f_\gamma = 1{\cdot}6$Mc/s.

The same stage will now be designed by using parallel voltage feedback. For the first assumption the simplest formula for γ will be used.

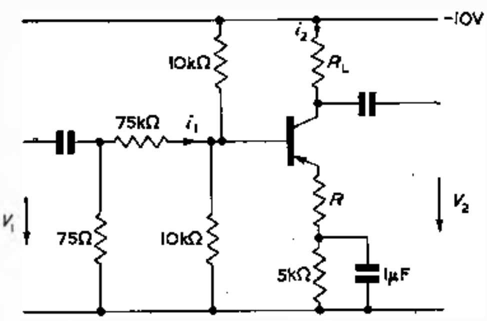


Fig. 12. Measured single stage amplifier with current feedback

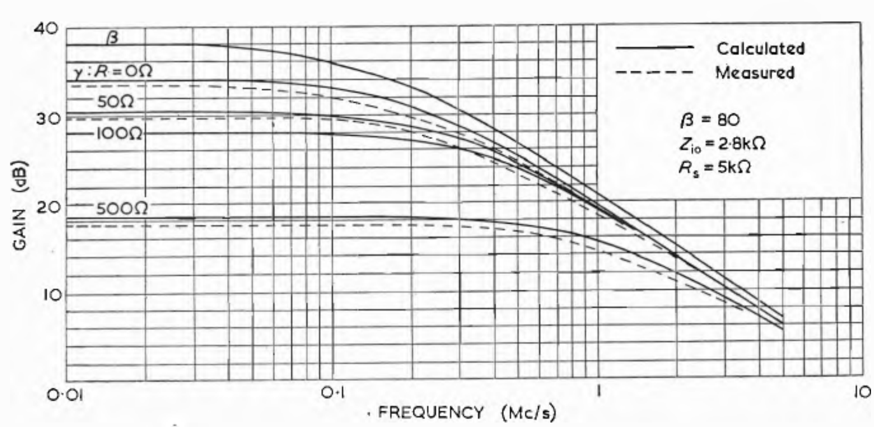
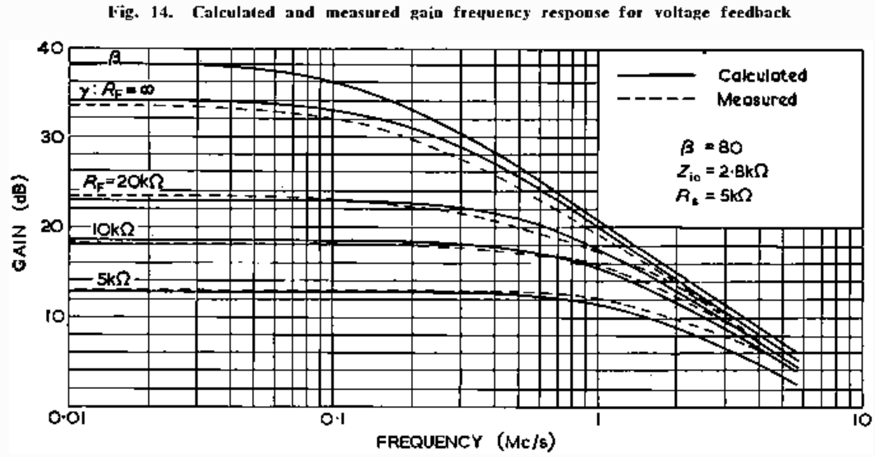


Fig. 13. Calculated and measured gain frequency response for current feedback

Fig. 14. Calculated and measured gain frequency response for voltage feedback



Therefore the formula (33) is used for γ and expression (34) to calculate the necessary feedback resistor.

$$\gamma = \frac{\beta_o}{1+\beta_o R_L/R_F + Z_{io}/R_s} \qquad R_F = \frac{\beta R_L}{A-1-(Z_{io}/R_s)} = 20{\cdot}6\text{k}\Omega$$

As Z_{io} is 1·1kΩ and $R_F = 20$kΩ, the use of the simplified formulae is justified. Using finally the relation (53) to get the cut-off frequency it follows that:

$$f_\gamma = f_\beta(1+(1+\beta_o)R_L/R_F)\left(\frac{1+(1+\beta_o)R_L/R_F + Z_{io}/R_s}{1+(1+\beta_o)R_L/R_F + \dfrac{R_b+R}{R_s}}\right)$$

$$= 3{\cdot}7 f_\beta = 1{\cdot}85\text{Mc/s}.$$

The input impedance is modified to 500Ω according to formula (28).

To show the correspondence between the formulae derived in this article and the practical results, a single stage amplifier has been measured and the results plotted together with the corresponding calculated responses.

The transistor used had a β of 80, $Z_{io} = 2{\cdot}8$kΩ; the d.c. operating was adjusted to $I_e = 0{\cdot}95$mA and $V_{ce} = 5{\cdot}2$V. The circuit which has been used is shown in Fig. 12 and has the following constants:

(a) $R_L = 75\Omega$ for current feedback.

(b) $R_L = 1\text{k}\Omega$ for voltage feedback.

$R_s = 5\text{k}\Omega$.

The current gain is then:

$$i_2/i_1 = \frac{V_2/R_L}{V_1/75\text{k}\Omega} = V_2/V_1 \,.\, \frac{75\text{k}\Omega}{R_L} \qquad B = 20\log\frac{75\text{k}\Omega}{R_L}$$

$B = 60$dB and 37·5dB for 75Ω and 1kΩ respectively.

As $Z_{io} = R_b + (1+\beta)R_e$, and R_e can be calculated when I_e is known, according to equation (3), R_b is then obtained as 500Ω.

(a) *Current Feedback*

Five curves have been plotted, β_o on its own, then the response corresponding to the feedback which is due to the finite input impedance only, and curves for three different values of emitter resistors, 50, 100 and 500Ω.

The same curves have been calculated using the formulae (5) and (41). Fig. 13 shows the two sets of curves. The correspondence is within 1dB.

(b) *Voltage Feedback*

As in paragraph (a) five curves are shown in Fig. 14, β_o, no external feedback and three different values of feedback resistors, 20kΩ, 10kΩ and 5kΩ. The formulae (18) and (48) were used to calculate the effect of feedback on the gain and the cut-off frequency. The curves, measured and calculated, correspond within 1dB as under (a).

Acknowledgment

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Short News Items

The Institution of Electrical Engineers will hold a series of discussion meetings on Wednesday and Thursday, 20 and 21 January, 1960, at Savoy Place, London, W.C.2, dealing with 'Managerial and Engineering Aspects of Reliability and Maintenance of Digital Computer Systems'. The programme will be divided and, on the first day, the British Computer Society will be responsible for arranging meetings which will deal with these subjects: Methods for Determining the Functioning Status of a System; Recording Techniques for Determining Operating Efficiency of a System; Programming Techniques for Protection against Transient Failures; and Management and Organization Problems. On the second day, the discussions will be concerned with such topics as Experience of System Reliability; the Influence of Engineering Design on Reliability; and Factors affecting the Reliability of Peripheral Equipment. These meetings will be organized by the Measurement and Control Section of the Institution with the support of the Council, and anyone wishing to contribute material is asked of submit a digest of approximately 500 words, for a 10 to 15 minutes' contribution, to the Secretary of the Institution not later than 30 October, 1959. The whole series of meetings is being held under the aegis of Group B, the British Group for Computation and Automatic Control, of the British Conference on Automation and Computation.

The London Electric Wire Company and Smiths Ltd have recently opened a new laboratory at their Leyton Works. It has been named the 'Wildy Laboratory' after the chairman and managing director of the Company and provides additional facilities for the development of wire coverings, wire covering plant and test equipment. One of the main topics of research is high temperature coverings for wire since most of the present known insulants, including p.t.f.e., have a number of serious disadvantages. Development is also continuing on bulk containers for fine wires: at present these can be supplied for wires down to ·0068in and it is envisaged that ultimately they will be able to be used for wires down to ·0044in, the weight of wire carried in them being about 60lb.

The Scientific Instrument Manufacturers' Association of Great Britain announced on September 1 the inauguration of SIMA's new Director—Captain Robert Alexander Villiers, C.B.E.

E.M.I. Electronics Ltd will supply an EMIDEC 1100 electronic data processing system to the Ministry of Labour and National Service. To begin with the computer will be largely employed in calculating statistics relating to employment, wages, hours of work, and the Retail Prices Index. In addition it will prepare the payroll for 20 000 employees at the Ministry. The computer will cost £140 000 and is due to be delivered at the end of 1960. This is the sixth computer to be ordered from E.M.I. Electronics in the past three months. The other five have been ordered by J. Sainsbury Ltd, British European Airways, the Royal Army Ordnance Corps, Barclays Bank and the Air Ministry.

The Independent Television Authority commenced low power test transmissions from the site of their eleventh transmitting station on 22 September, which was the fourth anniversary of the start of Independent Television programmes in this country. This station is now nearing completion at Church Hougham near Dover and it will serve an area stretching from the Medway towns to Beachy Head. The Authority hopes to start transmissions on full power from this station before the end of the year. The low power signals transmitted from the site of the Dover station will have an effective radiated power of approximately 1kW and will be radiated from a 75ft tower close to the recently completed permanent mast of 750ft. These low power transmissions will be sent out from 10 a.m. to 12.30 p.m. and from 2 p.m. to 5.30 p.m. on Mondays to Fridays and 10 a.m. to 12.30 p.m. on Saturdays: there will be no transmissions on Sundays. The Dover station will operate on channel 10 and it will transmit vertically polarized signals.

The British Standards Institution's Annual Report for 1958-1959 reflects the growing demand of industry for new standards specifications on which to base production and purchasing requirements. The report, which is for the year ended 31 March, says that although the total number of new standards published is slightly less than in the previous twelve months "there has been a sharp increase in new projects to be undertaken in the technical divisions: in the engineering division 144 as compared with 92 last year; in the building division 34 compared with 17; in the chemical division 48 compared with 16".

Imperial Chemical Industries Ltd have added an experimental silicone resin designed for flexible electrical insulation to their development range. This resin, identified as DP.139, is designed to impregnate glass-cloth at the significantly lower temperature range of 100 to 120°C with much improved coating speeds. Glass-cloth is very easily wetted by DP.139, and a tack-free film is formed after heating for five minutes at 100°C.

The Institution of Electrical Engineers, Electronics and Communications Section, have announced that with the support of the Council they are organizing the Third International Conference on Medical Electronics in London in the second half of July 1960, in association with the International Federation for Medical Electronics which was inaugurated this year.

Pye Telecommunications Limited announce that a contract has recently been signed with Yugoslavia for three Pye Instrument Landing Systems. The Pye system embodies all the basic facilities required for automatic 'hands-off' landing, as recently demonstrated at the Royal Aircraft Establishment, Bedford. A considerable quantity of v.h.f. and h.f. units for ground-to-air communication have also been ordered, as well as a fixed station and number of mobile radiotelephones for airfield vehicle control.

The Radio Show, Earls Court, London, which closed its ten-day run on 5 September, attracted a total attendance of 310 161, compared with 334 502 last year. The slight reduction being attributed by the organizers to the exceptionally fine weather this year. The attendance of buyers from home and overseas was considerably higher. There were 4 109 overseas visitors in all, including 679 classified as buyers, the latter figure being 25 per cent up on last year. Of more than 100 countries represented, most visitors were from: Union of South Africa, nearly 400; India, over 350; Australia, about 275; Pakistan, over 200; New Zealand, about 175; and Ceylon, over 150.

The Ministry of Supply announce that Mr. A. N. Christmas has been appointed Director of Guided Weapons Research and Development (Techniques).

Universal Winding Company, exclusive American licensee of the 'Hydrox' fuel cell, has announced that final arrangements are being made to extend its licensing agreement with the National Research Development Corporation of England to cover other types of fuel cells, including the 'Carbox'. The announcement was made by Mr. P. P. Johnson, executive vice-president of Universal Winding, at a luncheon given by the Company in New York to welcome from England Mr. F. T. Bacon and Dr. H. H. Chambers, inventors of the 'Hydrox' and 'Carbox' fuel cells respectively. Much of Messrs. Bacon and Chambers' work on fuel cells was done under the sponsorship of the National Research Development Corporation. The British scientists visited America in order to read papers at a Fuel Cell Symposium held at the American Chemical Society Convention in Atlantic City. The Patterson, Moos Research Division of the Universal Winding Company has already designed and built prototype units of the 'Hydrox' cell of such compact size and efficiency that the principle shows promise for use where space and weight are critical. Efficiencies of about 65 per cent have been achieved, compared with an overall efficiency of about 30 per cent for a modern power station. 'Hydrox' systems in the range of 1·5 to 10kW are now being made. Through a recent agreement, a research and development programme on this fuel cell is being carried out jointly by Universal Winding and the Pratt and Whitney Aircraft Division of United Aircraft Corporation.

The National Institute for Research in Nuclear Science announce that contracts amounting to a total cost of over £430 000 have been placed for the manufacture of the magnet coils required to energize the 7 000 ton electromagnet of the 7 000MeV proton synchrotron. This machine, which has been named NIMROD, is being built for the Institute by the United Kingdom Atomic Energy Authority at the Rutherford High Energy Laboratory, Harwell. Contracts have been awarded to British Copper Refiners Ltd of Prescot, Lancs, for the supply of over 300 tons of refined copper (from which the coils are to be made) in the form of cast billets; to James Booth and Co. Ltd of Birmingham for extrusion of the cast copper into hollow rectangular bars: and to Metropolitan Vickers Electrical Co. Ltd of Manchester for the manufacture of the finished coils from the extruded bars.

The Institution of Electrical Engineers have made arrangements for the Inaugural Address by the President, Sir Willis Jackson, F.R.S., to take place on Friday, 9 October, 1959, at 5.30 p.m., instead of Thursday, 8 October, as previously stated.

A Society of Environmental Engineers has been formed to provide a forum, by meetings, publications and visits, for the exchange of information and views among those engineers who are concerned with the development of equipment to withstand shock, vibration and other forms of environmental conditions, and who research in these fields. Membership and associate membership is open to scientists and engineers working on these subjects. Sponsor membership is open to firms similarly concerned. Details may be obtained from the Secretary, The Society of Environmental Engineers, 42 Manchester Street, London, W.1.

Marconi's Wireless Telegraph Company has been awarded a contract for the supply and installation of television and sound broadcasting equipment by the Royal Board of Swedish Telecommunications. The Board, which is responsible for all public sound and television transmitting stations in Sweden, has planned a major expansion of Sweden's television and f.m. sound broadcasting networks. The Marconi equipment ordered on this contract, costing approximately £292 000, will play a major part in the plans to give complete coverage to an area with nearly 7 million inhabitants.

The Central London Productivity Association is organizing a one-day conference, to be held on Thursday, 22 October, 1959, from 9.45 a.m. until 4.45 p.m. at the Council Chambers FBI, 21 Tothill Street, London, S.W.1. This conference is solely designed to introduce the smaller firm to the potentials offered by computers; to illustrate the use of the smaller computers; the progressive utilization of computers; service centres, joint ownership. Speakers from three leading computer manufacturers will be illustrating their papers with slides and case histories. Two other speakers will give information on computer centres, and user experience. There will be ample opportunity for questions during the latter part of the conference. Conference fee will be 45s. (members of the Association 35s.), and includes coffee, luncheon and tea. Programmes and tickets may be obtained from Mr. D. MacGlashan, Central London Productivity Association, c/o Stafford Allen & Sons Ltd, 20/42 Wharf Road, London, N.1.

The Ministry of Supply Signals Research and Development Establishment at Christchurch, Hants, has constructed an ammonia maser clock, which is to be installed in the Australian Post Office Research Laboratories at Melbourne by a team of British scientists. This clock, the first of its kind in the Southern Hemisphere, is accurate to within one ten thousandth of a second per day. One purpose for which the maser will be used is the measurement of the rate at which pulses sent out by a radio transmitter near Rugby are received in Melbourne, some 15 000 miles away. Although the transmitting rate is constant the pulses are not received at a constant rate, due to variations in the ionosphere. The maser will be used to measure these variations, providing more information about the ionosphere. An ammonia maser relies on the natural vibrations of ammonia gas molecules which produce electric pulses at the rate of 24×10^{9}/sec.

The Institute of Physics has recently published a 36-page revised edition of its booklet to assist less-experienced authors and to serve as a reference booklet for all those who wish to contribute to the Institute's publications. Under its new title "Notes for Authors", it gives hints on the preparation of scripts and diagrams, on the layout of mathematics, the correction of proofs and so on. In addition to a bibliography of reference books and works on technical writing, the pamphlet also contains lists of the spellings, symbols and abbreviations used by the Institute. Copies are obtainable from The Institute of Physics, 47 Belgrave Square, London S.W.1, price 3s 6d including postage.

The INTERKAMA — International Congress and Exhibition for Instrumentation and Automation, which was such a success in 1957, is to be repeated during 1960 from 19-26 October in Dusseldorf. The new INTERKAMA committee has already been formed and Professor Dr. Eduard Gerecke, lecturing professor at the Zurich Technical College, elected as its chairman. Once again INTERKAMA is subdivided into a congress and an exhibition section. The congress will be organized by the German technical and scientific associations. The first meeting of the congress advisory board, at which the competent scientific bodies were represented, has already been held in Dusseldorf. The exhibition advisory board also met in Dusseldorf and was attended by delegates from England, France, Netherlands and Switzerland.

The Council of the Institution of Electrical Engineers for the 1959/60 season will include the following: President: Sir Willis Jackson, F.R.S. Honorary Treasurer: E. Leete. Ordinary Members of Council: Prof. H. E. M. Barlow, C. O. Boyse, Prof. M. W. Humphrey Davies, L. Drucquer, J. M. Ferguson, Dr. J. S Forrest, R. J. Halsey, C.M.G., F. C. McLean, C.B.E., B. L. Metcalf, Dr. J. R. Mortlock, H. G. Nelson, H. V. Pugh, J. R. Rylands, Dr. G. A. V. Sowter, C. E. Strong, O.B.E.

The Minister of Supply, the Rt. Hon. Aubrey Jones, has given permission for Sir Owen Wansbrough-Jones, K.B.E., C.B., Chief Scientist to the Ministry of Supply, to resign from the Civil Service as from 30 September, 1959, to take up an appointment in industry.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

Response of a Capacitance-Resistance Divider

DEAR SIR,—I was interested in Mr. Turk's article on the response of *RC* voltage dividers (ELECTRONIC ENGINEERING, October 1958), and its generalization for arbitrary inputs by Mr. R. H. Evans (April 1959). The generalization can, however, be carried a good step further by allowing both the voltage wave input to the circuit and the circuit itself to be of arbitrary composition. Furthermore, what has to be sought is not the voltage waveform developed across any specified components of the chain, as this will not embrace all possible voltage divisions the circuit is capable of, but the far more informative *current response*, the calculation of the individual voltage waves by the simple $L\, di/dt$, $1/c\ {}_0{}^t i dt$ and Ri relations then presents no further difficulties.

The application of Duhamel's superposition integral to

$$h(t) = \frac{M(0)}{N(0)} + \frac{M(pk)}{pkN'(pk)}\ e^{pkt}$$

the well-known response of the series circuit to Heaviside's unit function—and all circuits capable of maintaining a continuous sequence of electrical events are of the series type—will yield the following general result:

$$I(t) = \frac{M(0)}{N(0)}\ V(t)$$

$$-\frac{M(pk)}{N'(pk)} \sum_{d=0}^{\infty} \frac{V(t)^{(d+1)}}{pk^{d+2}} - \frac{V^{(d)}{}_{(0)}}{pk^{d+1}}\ e^{pkt}$$

the notation being that used in Heaviside operational calculus. The summation has to be extended to all values of k, and d will have of course a finite maximum value, when the voltage wave $V(t)$ is presented in the form of a finite polynomial. The function is easily calculated for any particular case and holds under very general conditions. Being a current response function, it has a wider application than the voltage response functions of individual components—which it implicitly contains—for instance, one may read from it the necessary conditions, component values and tolerances for linear (or near-linear) currents in cathode-ray deflector coils, to give one instance. As to parallel branches within the series circuit, the usual procedure of replacing them by fictitious single impedances of the p-type should be employed.

A large selection of both current waveform and voltage division problems can thus be covered by the single $I(t)$ function quoted, and when the voltage waveform is presented in the form of an oscillogram, the polynomial

$$\sum_{1}^{m} (D_n/D)t^{m-1}$$

can be advantageously used for $V(t)$, D being the determinant

$$\begin{vmatrix} 1 & t_1 & t_1^2 & \dots & t_1^{m-1} \\ 1 & t_2 & t_2^2 & \dots & t_2^{m-1} \\ \vdots & & & & \\ \vdots & & & & \\ 1 & t_m & t_m^2 & \dots & t_m^{m-1} \end{vmatrix}$$

and D_n is obtained by replacing column n (containing the n-1 th powers of the observation times t_1, t_2 ... t_m) by the voltage readings V_1, V_2 ... V_m occurring at these instances, or, in other words, replacing a set of selected values on the X—deflexion axis by the corresponding values on the Y axis. The advantage of this algebraic representation of $V(t)$ is, that only a finite number of differentiations have to be carried out to obtain the current function.

Yours faithfully,

M. L. TELCS, Ph.D.,
Perdio Ltd.

The Correspondent Replies:

DEAR SIR,—Dr. Telcs has usefully pointed out that for each exponential term in the step function response of a circuit, the response to an arbitrary input contains a series of the form of equation (5) of my letter. Of course, in general, some of the zeros p_k of $N(p)$ will be conjugate complex pairs, corresponding to oscillatory terms in the response. For each conjugate pair $\alpha \pm j\omega$, the step function response $h(t)$ contains a term

$$2\,R\ \frac{M(\alpha + j\omega)}{(\alpha + j\omega)\ N'(\alpha + j\omega)}\ e^{(\alpha+j\omega)t}$$

which may be written

$$2\ \frac{|M(\alpha + j\omega)|}{(\alpha^2 + \omega^2)^{\frac{1}{2}}|N'(\alpha + j\omega)|}$$

$$\times\quad e^{\alpha t}\ \cos\ (\omega t + \phi_m - \phi_n - \theta)$$

where

$$\phi_m = \arg M(\alpha + j\omega)$$
$$\phi_n = \arg N'(\alpha + j\omega)$$
$$\theta = \arg\,(\alpha + j\omega)$$

The corresponding series in the expression for $I(t)$ becomes

$$-\frac{2\ |M(\alpha + j\omega)|}{|N'(\alpha + j\omega)|}$$

$$\times \sum_{d=0}^{\infty} \left\{ \frac{V^{d+1}(t)\ \cos\ [\phi_m - \phi_n - (d + 2)\theta]}{(\alpha^2 + \omega^2)^{(d+2)/2}} \right.$$

$$\left. - \frac{V^{d+1}(t) \cos\,[\phi_m - \phi_n - (d+2)\theta]}{(\alpha^2 + \omega^2)^{(d+1)/2}}\ e^{\alpha t} \right\}$$

Dr. Telcs' emphasis on the importance of representing the input $V(t)$ by means of a polynomial is very valuable, since the calculation of the coefficients by means of determinants will usually be much simpler than the numerical differentiation involved in the general series form of the solution, or the numerical integration involved in the superposition integral form.

Yours faithfully,

R. H. EVANS,
Farnborough, Hants.

Diffraction Grating Measurements

DEAR SIR,—The technique of linear measurement by the use of optical diffraction gratings is proving itself to be an extremely important one. Although the system has been used on machine tools in particular with great success, it is still in its infancy, and we may confidently expect great improvements in the near future. Perhaps the most important advances are at present being made in the direction of making the system an a.c. rather than a d.c. one. However, a sophistication of the conventional d.c. reading head which has been discovered at the N.P.L. might be of interest. It has not to our knowledge been published yet.

The usual technique for employing diffraction gratings for linear measurement involves the use of two signals spaced 90° apart on the intensity modulation pattern produced by movement between two similar gratings. For practical reasons each of these signals is usually produced by an antiphase pair of photo-transistors and it is necessary to form four 'phases' at 0, 90°, 180° and 270° to one another. If diffraction gratings were perfect many systems could be used to produce these signals, but in fact good commercial gratings suffer more or less severely from both 'progressive error' (for instance in 2 500 line/in gratings ±0·0002in is normally present) and 'fan error' (for instance in 2 500 line/in gratings one minute of arc is possible).

No commercially practical head has yet been devised that will correct for known errors on the gratings, although such schemes as cam correction are perfectly possible theoretically. However, a practical head may be made which optically compensates for the errors in such a manner as to present the information to the electronic counter in the best possible manner. There are three criteria for this type of reading head.

(1) Maintenance of absolute amplitude of quadrature signals.

(2) Maintenance of relative amplitude of quadrature signals.

(3) Maintenance of accurate phase quadrature.

Of these, the most important is the last, since it is employed for direction discrimination.

The conventional four-phase head employs a skewed cursor to select the four phases, and it may be seen Fig. 1, that it is vulnerable to fan error, this causing the relative phases to alter. To keep this effect to a manageable level

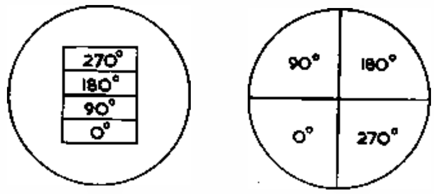


Fig. 1. Arrangement of pick-up heads

it is necessary to use a very small area of cursor.

The quadrant head picks up the signals from the grating as Fig. 1 shows, around four quadrants of a circle. Although the absolute magnitude of the signals is affected by both progressive error and fan error, the phase quadrature is maintained however severe these errors become within reason.

Quantitative analysis of both systems shows that the quadrant head has definite practical advantages over the conventional head.

Yours faithfully,
R. W. Wilde,
The Staveley Research Department,
Bedford.

Zener Diode Characteristics

Dear Sir,—As a result of a short series of measurements on Zener diodes, certain properties of these devices have become evident which, if confirmed as repeatable and stable with time, promise some interesting applications.

Zener diodes, with nominal voltages from 4·7 to 9·1V from three manufacturers, were found to conform to the single set of characteristics shown in Fig. 1. As the current through a particular

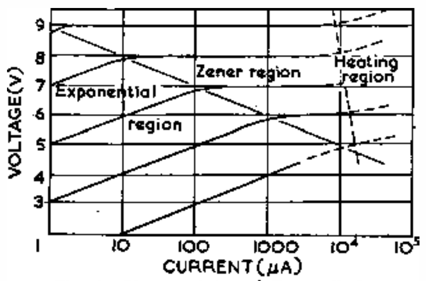


Fig. 1. Characteristics of Zener diodes

diode is increased from zero, the voltage across the device at first increases as the logarithm of the current (the 'exponential region'), is then essentially constant for an interval, and finally increases slightly again as the region of resistance and heating effects is reached. The boundaries indicated are approximate.

In the exponential region the constant relating change of voltage to current ratio reaches the high value of 0·75V to 1V per 10 times increase of current and in those diodes tested at elevated temperatures was independent of temperature.

The extent of the Zener region for 9·1V diodes was maintained for at least 90 from a batch of 100.

The extended exponential region of the lower voltage diodes appears suitable for analogue conversion of linear to logarithmic functions with a sensitivity approximately ten times greater than that offered by the forward characteristic of semi-conductor diodes.

Yours faithfully,
M. R. Nicholls,
Decca Radar Ltd,
Chessington.

Self-Quenching in Super-Regenerative Parametric Amplifiers

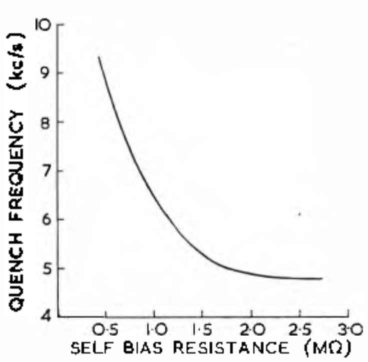
Dear Sir,—In our work on cavity-type parametric amplifiers in the u.h.f. range, using gold-bonded germanium diodes (Mullard OA47), it was noticed that the small signal theory of the three-frequency parametric amplifier[1] does not always agree with the experimental results. This is expected since the theory assumes certain idealizations which may not be fulfilled in practice.

The aim of this letter is to report on, and give a qualitative reasoning of a phenomenon observed when self-bias, or external bias with high internal resistance, is applied to the diode.

Under such bias conditions it was noticed that the gain of the amplifier increases and its bandwidth decreases as the pump power increases up to a certain level, when an increase in the pump power would not set the amplifier into self-oscillation as would be expected from a regenerative amplifier. Instead, the increase in the pump power changes the frequency response of the amplifier from that of a single-tuned circuit to a multi-peak response, similar to that of a super-regenerative valve amplifier operated in the coherent state. When the pump power is further increased, however, the side peaks give way to a single peak of very high gain. It was also noticed that this phenomenon would not occur if external bias of low internal resistance is used. This would suggest that the long time-constant of the diode capacitance and bias resistance is responsible for this effect, since the variation of the diode capacitance can no longer follow the voltage variation of the high frequency pump (S-band).

Further experiments supported this belief. For instance, when the signal was removed the amplifier worked as a squegging oscillator with quench frequency equal to the spacing between the multi-peaks of the coherent frequency response. The diode self-bias voltage

Fig. 1. Effect of diode bias resistance on the quench frequency



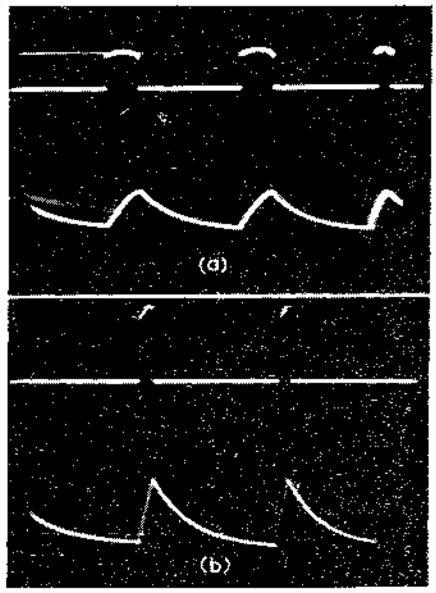


Fig. 2. Effect of the pump power on the quench cycle
(a) Pump power 100mW
(b) Pump power 120mW

(about −5V) was also found to be modulated with a quench frequency sawtooth oscillation (of amplitude in the order of 1 to 2V), which is similar to that of the grid bias of a self-quenching valve oscillator.

The quench frequency was found to vary with the diode bias resistance, as shown in Fig 1. The limitation on the lowest quench frequency is determined by the internal shunt resistance of the diode. On the other hand, the duration of the quench cycle, which seems to be responsible for the coherent operation of the amplifier, changes from almost a square waveform to very narrow pulses (of a few microseconds) as the pump power increases.

Figs. 2 (a) and (b) show the effect of the pump power on the pulse width of the self-quenching oscillation, and its synchronism with the diode bias voltage. The figures are photographs of the traces of a double beam oscilloscope. The upper traces are for the detected squegging oscillations, and the bottom traces are for the diode bias voltage.

Further work is being carried out to give quantitative treatment of the self-quenching in the diode parametric amplifiers.

The writer wishes to express his gratitude to J. H. Gibson of the microwave physics group for pointing out to him the advantages of the super-regenerative mode of operation and demonstrating his work on the separate quenching super-regenerative parametric amplifier.

Yours faithfully,
I. Hefni,
Marconi's Wireless Telegraph Co. Ltd,
Chelmsford.

REFERENCE

1. Heffner, H., and Wade, G. Gain Band Width and Noise Characteristics of the Variable Parameter Amplifier. *J. Appl. Phys.*, 29, 1321 (1958).

BOOK REVIEWS

Noise in Electron Devices

Edited by L. D. Smullin and H. A. Haus. 413 pp. 109 figs. Demy 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1959. Price 96s.

THE benefits to be gained in receiver sensitivity, by the reduction of the noise introduced by electron devices, are so great that interest in the subject has been at a high pitch for many years. Rapid advances have been made on the practical side, especially in travelling wave tubes, where the last decade has seen an average improvement in the best attainable noise factor of about 1dB per year, to a present figure of about 3dB.

It is fair to say that in general the theory of noise effects has lagged behind the practical achievements, and in very few types of device can theory be used to give an exact prediction of the noise performance, especially at very high and very low frequencies. Meanwhile new amplifying devices such as masers and reactance (parametric) amplifiers have been introduced, with lower noise factors than the older devices. Other aspects of operation will still often favour the latter and further improvements in their noise factor will still be sought. For this purpose, it is probable that more accurate theory will be needed.

It is very opportune therefore that Professors Smullin and Haus should have given us this up-to-date account of the theoretical and practical position of noise in electron devices which, of course, covers both valves and transistors. The book differs from a number recently published, in that it does not deal with the mathematics of random processes, but deals entirely with the noise characteristics of the devices themselves, and the way the noise output is influenced by the random emission velocity, drift motion or production of the charge carriers

The individual chapters are written by experts in their own fields, and are largely complete in themselves, with excellent bibliographies. Three chapters cover shot noise from cathodes, low frequency (flicker) noise, and noise in grid control valves, two cover noise in electron beams and the effects in low noise travelling wave tubes, and two are devoted to semiconductor and transistor noise. The whole is well edited, and where there is overlap of subject matter between the chapters, the slightly different emphasis adds rather than detracts from the value.

This is a book that can confidently be recommended to the device engineer or physicist. Much of the information in it was previously only available in scattered papers. It is to be hoped that when further work is done on the problems that are indicated to be as yet unsolved, the editors and authors will produce the later editions that will undoubtedly be wanted.

W. H. ALDOUS.

Basics of Missile Guidance and Space Techniques. Vol. 1 & 2

By M. Hobbs. 304 pp. 286 figs. Demy 8vo. J. F. Rider Publisher Inc., New York. 1959. Price $3.90 each.

THIS two volume 'picture book' course deals with the fundamentals of the important areas of electronics that will contribute to the conquest of space. Volume 1 covers the principles of control and guidance and in Volume 2, telemetering, space exploration by optics and electronics, satellite theory and practice, satellite monitoring and tracking as well as applications of earth satellites are covered.

Elementary Mathematical Programming

By R. W. Metzger. 246 pp. 81 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1959. Price 48s.

THIS book has been designed for the reader with a limited knowledge of mathematics, who wishes to understand the basic techniques and applications of mathematical programming. The various methods are presented via typical yet simple problems in a logical pattern. Further examples illustrate the techniques involved in formulation and analysis of solution methods.

Theory of Elliptic Functions

By H. Hancock. 498 pp. 76 figs. Demy 8vo. Dover Publications Inc., New York. 1959. Price $2.55.

THIS is an unaltered and unabridged republication of the first edition which was published in 1958.

In the 21 chapters, the author develops the theory of elliptic integrals beginning with formulas establishing the existence, formation and treatment of all three types (rational, simply periodic, doubly periodic) and concluding with a general description of these integrals in terms of the Riemann surface.

The Services Textbook of Radio. Volume 5: Transmission and Propagation

By E. V. D. Glazier and H. R. L. Lamont. 500 pp. 190 figs. Royal 8vo. Her Majesty's Stationery Office. 1959. Price 25s.

THIS volume deals with the basic theory used in radio and is one of a series of seven, covering the whole range of radio theory from the elementary principles of electricity and magnetism to the advanced techniques of radar and multi-channel communication systems. This series has been written primarily to meet the needs of the three fighting services for a comprehensive and authoritative work on the theory and technique of radio, including in this term radar and direction-finding as well as telecommunications, but is also well adapted for civilian needs.

Progress in Nuclear Energy —Economics

Edited by I. R. Maxwell, P. W. Mummery and P. Sporn. 419 pp. 109 figs. Medium 8vo. Pergamon Press Ltd, New York, London. 1959. Price 105s.

THIS book is the second volume of Series VIII of "Progress in Nuclear Energy", the edited proceedings of the Second International Conference on the Peaceful Uses of Atomic Energy at Geneva in September 1958. The whole review constitutes a remarkable record of the theory, development, applications and implications of nuclear energy which will be indispensable for those who are directly concerned with the subject.

For those less directly interested, this particular volume is perhaps the most appropriate of any because it presents 25 papers, by different authors, surveying the practical possibilities of adopting nuclear power in many parts of the world.

It is no secret that the "flourish of trumpets" which heralded nuclear power production has been followed by less flamboyant statements as time has shown some of the difficulties, economic and technical, inherent in its development for widespread use.

The book, which is beautifully produced, with many clear diagrams, has four parts: world demand for nuclear energy, nuclear fuel supplies and costs, the generation of nuclear energy and nuclear power programmes. It thus covers most of the aspects to be considered in assessing the economic future of nuclear power for different geographical, political and social circumstances.

That this form of power will eventually be used on a great scale cannot be doubted, but many factors govern its rate of adoption. Dr. H. J. Bhabha, of India, shows the great need for more energy to improve living standards in the under-developed countries of South Asia and the Far East where, in 1953, 775 million lived, but where the installed capacity of electrical generating plant averaged less than 22kW per thousand of population. More than 75 per cent of the world's population live in under-developed areas and need energy but how, and how soon, will they be able to pay for nuclear generating stations? Nuclear fuel supplies appear to be ample for many years to come and their costs of production will doubtless fall. Future technical advances in plant construction will certainly reduce capital costs but the elaborate organizations established for the development of large stations are faced with a rather uncertain prospect of further work.

Again, as P. Ailleret points out in his paper, there remain other sources of energy still to be developed—hydro power, tidal, geothermal, wind and sun and "the race between costs of future kinds of energy is far from run".

While, as in India, the costs of transporting fuels over long distances may make conventional thermal generation of power expensive enough for nuclear stations to compete, the abundance of hydro and other forms of cheap power as, for example, in Canada, U.S.A. or South Africa, must delay nuclear power developments. The present excess production of coal and the rapid strides made by the oil industry also have important influences. Nevertheless, this is not the complete story; much depends upon the characteristics of the loads to be supplied in relation to those of the non-nuclear power plant to be used. Several papers in the book therefore discuss the integration of nuclear stations in a mixed system in which they might play a very effective part.

G. W. GOLDING.

The Electronics of Microwave Tubes

By Dr. W. J. Kleen. 349 pp. 80 figs. Demy 8vo. Academic Press Inc., New York, London. 1958. Price 72s.

DURING a period of rapid technical advance, innovation and obsolescence may be almost contemporary. The last fifteen years have been such a period for microwave tubes. The basic ideas of synchronous interaction of a perturbed electron stream with travelling or rotating electromagnetic waves, have proved to be extremely fertile. In less than two decades, microwave tubes capable, separately, of generating megawatts of peak power, amplifying with power gains of tens of thousands, operating and electronically tuning over bandwidths of an octave or more, have ceased to be projects for research and development. They are available as engineered tubes, with specified performance and guaranteed life. The designer of centimetre wavelength systems has an embarrassing range of valve attributes at his disposal; his problems are those of selection, rather than extension. The present is essentially a period of consolidation for microwave tube techniques.

Dr. Kleen's book is therefore timely and the subject matter well chosen. Starting from basic principles, he unifies, in a clear and concise way, the theory of those microwave tubes which depend for their action on the presence of an electron discharge. This theory forms the foundations for the design of those power amplifiers and oscillators which will be in use at centimetre wavelengths for the next few decades. As far as low noise signal amplifiers are concerned, it is probable that the simpler travelling wave tube devices will be replaced in some applications by parametric type amplifiers.

This is an ideal book for the intending microwave specialist, who needs to supplement the few established but more general works on thermionic valve theory. The practising engineer will find this book, with its carefully selected bibliography, a palatable distillate of the abstracts and photostat copies he has collected over the last ten years.

The publishers should be thanked for realizing the need for such a book and the translators congratulated for their good sense in translating not only the descriptive text but also the mathematical presentation.

M. ESTERSEN.

Basic Electronics

By J. Daly and R. A. Greenfield. 221 pp. 254 figs. Demy 8vo. Heywood & Co. Ltd. 1959. Price 45s.

THIS is Part IV in the publisher's series 'Physical Processes in the Chemical Industry' and the scope is intended primarily for the young chemist or chemical engineer who, in these days, must have a sound basic knowledge of electricity leading to electronic devices and circuits. The level is intended for graduate and non-graduate students who are not yet acquainted with electrical matters.

For a book meant for non-electrical engineeers the work is remarkably refreshing in that it does teach correct practice in electrical terminology and symbols, unlike many such books where the authors appear to have no knowledge of standard terminology as covered by various B.S.I. Glossaries. Mathematics are simple: although the book is quantitative as well as qualitative it is by no means a simple descriptive book, with the exception of the last chapter. C.G.S. and m.k.s. units are mentioned in connexion with magnetism: $\mu = k\ B/H$, but do not appear to be otherwise used. The difference between this and a book for young electrical engineers will be noted.

There are twelve chapters covering: D.C.; Magnetism; A.C.; Hot- and cold-cathode devices; Amplifiers; Oscillators; Modulation and demodulation; D.C. amplifiers and magnetic amplifiers; Cold-cathode circuits; Semi-conductors; Power supplies; Input and output devices.

A few comments which might be made include the too-sketchy treatment of measuring instruments in Chapter 1 and in Chapter 2 permanent magnets are hardly treated at all. Chapter 4 includes the cathode-ray tube but does not refer to the cathode-ray oscillograph which is in Chapter 12, which latter is a collection of miscellaneous information including some electronic measuring instruments, photocells, strain gauges, pH measurement etc. Some of the ionization measuring devices are mentioned much too briefly to be of much use and absentees include the electron microscope, betatron, cyclotron and other electronic devices used increasingly for non-electrical applications.

Some chapters have a brief bibliography but not all are representative of the field.

In spite of these criticisms the book can be well recommended to the type of reader for whom the work is written—it could also be read with advantage by students of physics, biology etc.

E. H. W. BANNER.

Model-Radio Control

By E. L. Safford. 192 pp. 210 figs. Demy 8vo. Gernsback Library Inc., New York. 1959. Price $2.65.

THIS handbook of radio-control of model planes, boats trains etc., is a thoroughly revised and greatly enlarged version of a previous book by the same author. It covers all aspects of radio control from theory to construction of coders, decoders and other complex components as well as complete systems.

General Circuit Theory

By G. Newstead. 142 pp. 45 figs. Pott 8vo. John Wiley & Sons Inc., New York, Methuen & Co. Ltd. London. 1959. Price 15s.

THIS book develops the subject of circuit theory in a fundamental way with the minimum of assumption so that the results will apply to all the fields where the circuit concept is valid, although the language used is that of electric circuit theory. Emphasis is placed on the development of general relationships rather than the detailed analysis of particular circuits.

Linear Groups

By L. E. Dickson. 310 pp. Demy 8vo. 2nd Edition. Dover Publications, New York, Constable & Co., London. 1959. Price 16s.

THIS book on linear groups with an exposition of the Galois Field theory, is an unaltered and unabridged republication of the first edition, with a new introduction by Wilhelm Magnus, of the New York University.

Low Frequency Amplifiers

Edited by Dr. A. Schure. 88 pp. 38 figs. Demy 8vo. J. F. Rider Publisher Inc., New York. 1959. Price $1.80.

THIS book is Volume 30 in the Electronic Technology series edited by Dr. Schure and deals with the audio frequency band between 20c/s and 16kc/s. The book clarifies the application of thermionic valves as well as transistors to audio frequency amplifiers and special emphasis is given to considerations underlying the design of systems, showing and solving specific design problems.

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

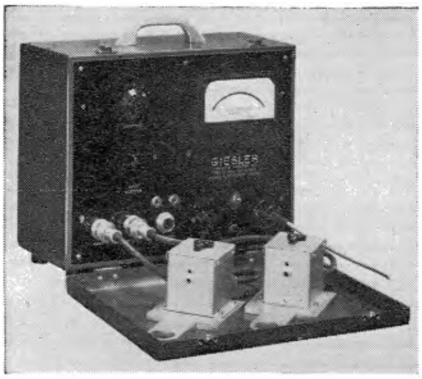
(Voir page 567 pour la traduction en français: Deutsche Übersetzung Seite 572)

VIBRATION MONITOR

(Illustrated below)

C.F.R. Giesler Ltd, Small Hall, River Place, Essex Road, Islington, London N.1

This portable equipment is designed for measuring vibration amplitudes on all types of rotating machinery. It has a frequency range of 6c/s to 300c/s which corresponds to a rotational speed range of 360 to 18 000rev/min. The equipment consists of an amplifier, amplitude meter, power supply unit and a seismic transducer. Provision is made for driving a single or double pen recorder. Two input sockets are provided so that two transducers can be connected, a changeover switch being used for selection.



The total range of amplitudes which may be read on the meter is from ·0001in to 0·1in. The full-scale sensitivity of the instrumentation is controlled by the 'sensitivity switch' and the meter has two scales, so proportioned that a reasonable pointer deflexion can be obtained with any vibration amplitude from ·00025in.

EE 13 751 *for further details*

TRANSFER FUNCTION ANALYSER

(Illustrated in next column)

Servo Consultants Ltd, 17 Woodfield Road, London W.9

This instrument has been designed to provide a stable and accurate means of testing of servomechanisms and control gear. Its main advantages are as follows:

(1) The electromechanical system employed is accurate and free from drift.

(2) The equipment is insensitive to harmonic distortion introduced by the system under test. Only the fundamental being taken into account.

(3) Direct indication of phase-angle and amplitude is given.

(4) A.C. operated servo systems can be tested directly.

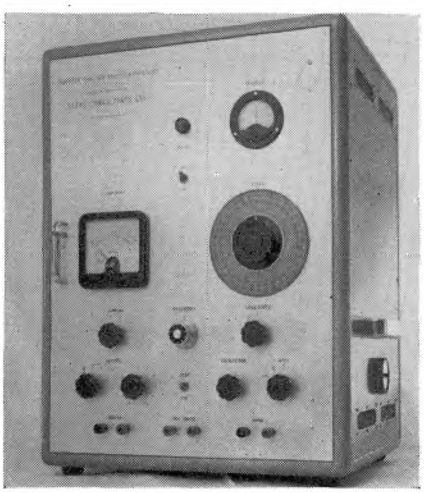
(5) Performance of modulators, rectifiers and magnetic amplifiers can be investigated.

(6) Possibility exists of direct coupling to mechanical and hydraulic control without the use of actuators.

(7) The equipment is completely self contained.

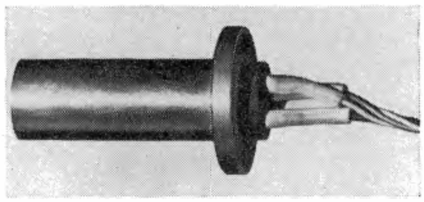
Two electromechanical modulators are driven by a variable speed motor, the speed of which is equal to the test signal frequency. The speed of this motor can be adjusted manually by means of a precision helical potentiometer.

Modulator 1 has a stationary winding which is supplied with a constant amplitude carrier signal. The output winding of this resolver, which is rotating, will therefore produce a modulated carrier signal, which in the case of an a.c.-operated servomechanism is then available directly as a test signal. The carrier supplied to the resolver can be derived either from a built-in 2kc/s generator or from any other source outside the instrument. In the case of a d.c. servo system this signal is demodulated to obtain the signal at the test frequency. In addition, a mechanical coupling can be introduced between the generator and the system under test. The output from the system under test can again be either a modulated carrier such as a 400c/s output from a synchro or a signal at the test frequency. Such signal is amplified and fed into the resolver 2, which is rotating in synchronism with the resolver 1. In the case of a d.c. system the output signal from the system under test is used to modulate the carrier, which is then applied to resolver 2. The stator of this resolver has a three-phase connexion to another resolver (resolver 3) coupled to the angle indicating dial. The angle between the stator and the rotor of this resolver will correspond to the phase difference between the outgoing and the incoming signals to the instrument when the signal from the output winding of this resolver is at a null. The resolver also carries another winding at right-angles to the first one. The output from this winding provides a voltage proportional to the amplitude to the input signal. The magnitude of this vector is indicated by a direct reading meter.



The instrument is housed in a strong cabinet constructed from aluminium alloy. The front panel, which carries all meters and controls, is hinged to give easy access to all the important parts of the equipment. Two vertical chassis which carry all the electronic circuits are accessible on both sides without being removed from the cabinet. The generator and resolver gear-box is mounted at the bottom of the cabinet. The output shaft is accessible from the right-hand side.

EE 13 752 *for further details*



PRESSURE TRANSDUCER

(Illustrated above)

The Solartron Electronic Group Ltd., Thames Ditton, Surrey

The Solartron Electronic Group have concluded with the Svenska Flygmotor Aktiebolaget a manufacturing and marketing agreement of world-wide coverage (except Scandinavia) for a new 'vibrating cylinder' pressure transducer, which is claimed to be unique.

Production has already begun in Britain and the transducer has Ministry of Supply approval for incorporation into missiles and other equipment during test.

The 'vibrating cylinder' employs an entirely new design principle and is small, accurate and extremely robust as a pressure transducer. The variable frequency output signal makes it especially suitable for magnetic tape recording, as for instance with the Solartron Data Recording Equipment; also for analogue-to-digital conversion by pulse counting. The equipment which uses the signal from the transducer can be located a considerable distance from the point of measurement, without loss of accuracy.

The pressure sensitive element of the transducer is a mechanical vibrating system, whose natural frequency is dependent upon the applied pressure. This vibrating element determines the frequency of an electronic feedback oscillator thus providing an output voltage of constant amplitude but with a frequency which varies with the pressure applied.

The transducer was developed as part of a system for recording ramjet data on magnetic tape, during flight.

For this purpose a transducer with a variable frequency output was found to be the most suitable. The variable frequency signal is easily convertible to digital form for further treatment in a digital computer.

These transducers are available for input pressures of 25, 50 100 and 600 lb/in^2.

EE 13 753 *for further details*

SENSITIVE VALVE-VOLTMETER

(Illustrated below)

Dawe Instruments Ltd, 99 Uxbridge Road, London W.5

The type 614C valve voltmeter is a compact, portable instrument for the measurement of voltages from 80µV to 300V at frequencies between 5c/s and 100kc/s. It is possible to measure alternating voltages with superimposed direct current, although a combined peak level of 1kV should not be exceeded.

The basic circuit comprises a high-gain amplifier which feeds a full-wave rectifying indicating meter through an attenuator. The two stages of this attenuator—one at the input and the other separating the twin amplifier stages—are both switched by a 13-position range switch to obtain full scale deflexions of 300µV, 1mV, 3mV, 10mV, 30mV, 100mV, 300mV, 1V, 3V, 10V, 30V, 100V, and 300V. The attenuator stage following the first amplifier section is brought in first to improve the signal-to-noise ratio at higher input levels. High calibration accuracy and stability are ensured by the application of negative feedback over the two amplifier stages and by the incorporation of a stabilized h.t. power supply.

The meter is calibrated in r.m.s. values for a sinusoidal input, the indications representing the full-wave average value. The logarithmic voltage scale gives equal reading accuracy (better than ±2 per cent of indication with a sinusoidal input) for all voltages. The frequency error is less than ±3 per cent from 5c/s to 100kc/s. In addition to the thirteen normal voltage ranges, an auxiliary linear decibel scale gives a range from −80 to +50dB with reference to 1V.

An output jack is provided for monitoring the signal or to enable the instrument to be used as an amplifier with a gain of up to 88dB. When used in this way, the output is approximately 8V r.m.s. at full-scale deflexion into a minimum load of 10Ω. The source impedance is roughly 20Ω.

The input impedance is approximately equivalent to a resistance of 10MΩ in parallel with a capacitance of 25pF up to the 10mV range and with about 10pF on higher ranges.

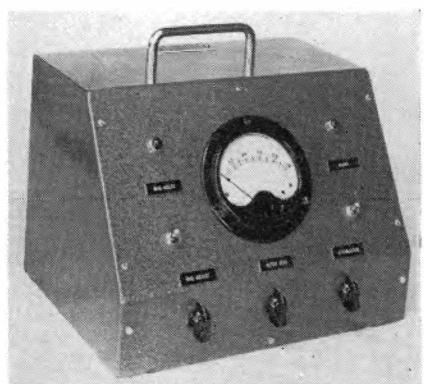
EE 13 754 *for further details*

SPARK INTENSITY INDICATOR

(Illustrated below)

Lexor Electronics Ltd, 25 Allesley Old Road, Coventry, Warwickshire

Incorporating a sensitive valve-voltmeter this instrument will indicate a very wide range of sparking levels and can



easily be calibrated to suit particular applications.

Typical uses include development and production testing of d.c. machines, switchgear, contactors, components, etc., and for radio interference suppression work. Models are available for permanent installation for continuous monitoring in connexion with maintenance and safety schemes.

The instrument will indicate sparking not normally visible, such as on the underside of brush-gear and on equipment undergoing tests in an altitude chamber. Models are available to suit either continuous spark or single pulses.

An input filter separates the spark pulses from standing potentials of up to 1kV d.c. and the final indication is displayed on a robust 3½in moving-coil meter. Maximum sensitivity is 50mV for f.s.d.

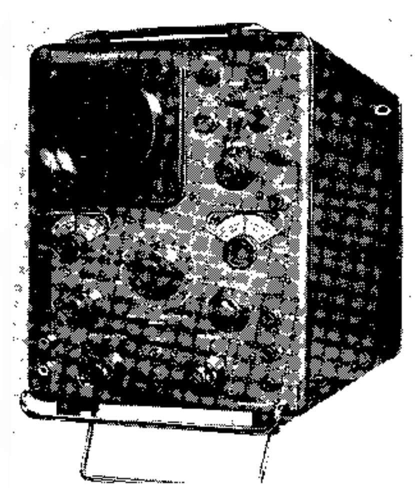
EE 13 755 *for further details*

GENERAL PURPOSE OSCILLOSCOPE

(Illustrated above right)

Marconi Instruments Ltd, St. Albans, Hertfordshire

The TF1330 is a high-grade, general-purpose oscilloscope with all the features required for detailed viewing and precision measurement of waveforms. Its frequency response extends



from d.c. to 15Mc/s, and sensitivity is variable in seven steps from 50mV/cm to 50V/cm; a signal of only 300mV will provide full coverage of the 6 × 10cm screen. A balanced lumped-constant delay line is incorporated in the vertical deflexion system to allow distortion-free observation of the whole of a signal under investigation.

The time-base sweep velocity is variable from 1sec/cm to 0·1µsec/cm in fifteen ranges, and can be increased up to 0·02µsec/cm using X expansion; each range may be easily standardized against an external source. Triggering can be applied internally or externally, or from an internal supply-frequency source. A selection of d.c. and a.c. coupling arrangements ensures stable triggering with signals from very low frequencies up to at least 15Mc/s and includes provision for television field synchronizing.

Directly calibrated potentiometers are used for the measurement of both signal amplitude and time: the calibration is accurate to within ±2 per cent and is independent of either X expansion or Y gain.

The 5in cathode-ray tube has a spiral post-deflexion accelerator and operates at an e.h.t. of 10kV, features that combine to give a linear high-brilliance display at fast writing speeds or low pulse repetition rates. To allow full use to be made of the very slow time-base speeds, a longer-persistence tube can be supplied in place of the normal type. The viewing hood around the screen of the tube may be removed to permit the attachment of a conventional oscilloscope camera unit.

Operational convenience is assured by colour-coded, functionally arranged, concentric controls. If the trace moves off the screen, in any direction, depression of a push-button switch immediately reveals its position; it is then a simple matter to return the display to the centre of the screen by means of the two shift controls. A tilting stand is fitted below the front of the instrument to allow easy viewing.

A dual-trace model, Type TF 1331, is also available. This has the same basic performance as the TF 1330, plus electronic beam switching for displaying two signals either on alternate, sweeps or at 100kc/s switching rate.

EE 13 756 *for further details*

COIL BONDING TIMER

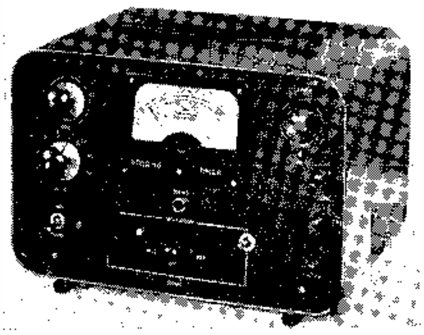
(Illustrated below)

Avo Ltd, Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

When using self-bonding winding wires, the quickest method of effecting the bonding of a coil, is to pass through it, for a short period, a current of suitable magnitude. Since the voltage required to do this may be quite high, it is first necessary to check the insulation between the wire and the mandrel on which the coil has been wound.

The Avo bonding timer carries out both these functions in the following manner:—

The two ends of the coil are attached to terminals on the device, and switches



operated simultaneously by both hands to give a direct reading in megohms of the value of the insulation between the coil and the mandrel. If the insulation value is satisfactory, the operation of a third switch allows heating current to pass through the coil for a predetermined period.

Because the operator's hands must both be employed while these tests are made, the apparatus is intrinsically safe. Timing of the bonding current is variable between 3/4sec to 2½ min, while insulation can be measured up to 500MΩ.

The instrument operates from 210 to 250V, 50c/s mains, but where the mains voltage is considered too high for bonding purposes, a lower voltage for bonding can be injected via terminals provided.

EE 13 757 *for further details*

ENCAPSULATED PAPER DIELECTRIC CAPACITORS

Dubilier Condenser Company (1925) Ltd, Ducon Works, Victoria Road, North Acton, London W.3

These new 'Dubilier' paper dielectric tubular capacitors, known as type 560, have been designed to meet the demand for high performance paper dielectric tubular capacitors having exceptionally good electrical characteristics coupled with long life. These capacitors may be operated in the temperature range of $-40°C$ to $+125°C$ on d.c., and $-40°C$ to $+70°C$ on a.c. without voltage derating.

The method of construction is as follows. A conventional paper and foil element provided with interconnexions between electrodes and wire terminations is impregnated with a high temperature plastic material to produce a solid capacitor element. The formed wire terminations are inserted in the interconnexions and these are formed on and soldered to the wire terminations, thereby providing robust electrical and mechanical joints. The assembled element is housed by encapsulation in a mineral loaded epoxy resin providing excellent resistance to shock and the entry of moisture.

They are available in a range of values between ·001μF and 0·1μF, the working voltage varying between 180V a.c. and 250V a.c., or 350V d.c. and 1kV d.c., depending on the capacitance.

EE 13 758 *for further details*

COMBINED WAVE-CHANGE SWITCH AND TUNING CAPACITOR

(Illustrated below)

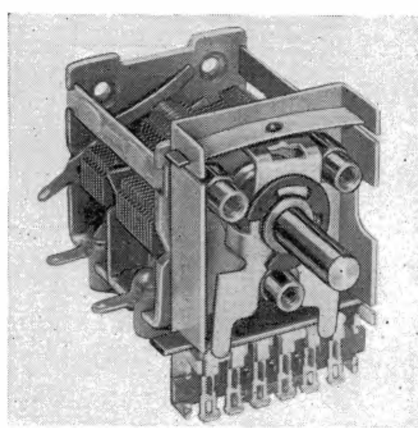
The Plessey Co. Ltd, Vicarage Lane, Ilford, Essex

The introduction of switched variable capacitors enabling normal wave change and tuning operations on any two waveband radio receiver to be carried out with only one control knob is announced by The Plessey Company Limited.

The new capacitors incorporate a conventional moving vane spindle fitted with a link motion that moves a Plessey slide switch in one direction when almost fully meshed, and in the other direction when almost fully unmeshed. It is recommended that the switching is so arranged that the small loss of usable swing occurs at the high frequency end of the long waveband and the low frequency end of the medium waveband. By this means, any loss of station coverage is unimportant.

The switch is a two pole changeover assembly but provision has been made, by the addition of a third pole, for dial lamp switching or similar operations; provision can also be made for mechanical waveband indication.

Plessey add that in addition to the facility of one-knob, two waveband control as a retail selling point for domestic receivers, a further economic



factor worthy of consideration is the simplification of circuits implied by use of the new capacitors.

EE 13 759 *for further details*

SENDING LEVEL STANDARD

(Illustrated below)

Radiometer, 72, Emdrupvej, Copenhagen NV, Denmark.

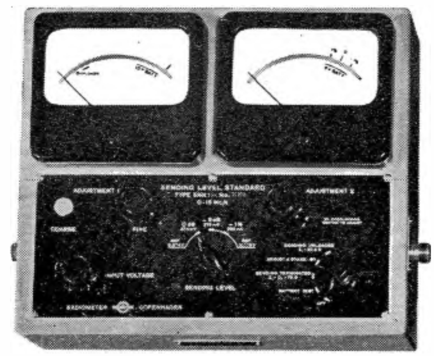
The type SNN1 sending level standard is a high-precision instrument for accurate calibration of level measuring instruments and exact measurement of their frequency response.

When connected to a suitable generator, the type SNN1 can deliver three output levels, either open-circuit or across a 75Ω load. The levels are

0dB/0·274V = 274mV

9dB/0·775V = 275mV

− 1N/0·775V = 285mV



The accuracy of the output levels is better than 1 per cent from d.c. to 25Mc/s with a 75Ω load, and better than 1 per cent to 15Mc/s with unloaded output. The frequency response is within ±0·2 per cent from 10c/s to 10Mc/s.

The instrument works on a saturated-diode principle which ensures fast response, 0·1 per cent incremental reading accuracy and true r.m.s.-indication. Another feature is an efficient overload protecting device.

The type SNN1 was developed for use in conjunction with carrier-frequency telephone equipment, but it can be used in many other applications.

EE 13 760 *for further details*

SPECTRUM ANALYSER

(Illustrated above right)

James Scott & Co. (Electrical Engineers) Ltd, 68, Brockville Street, Carntyne Industrial Estate, Glasgow, E.2.

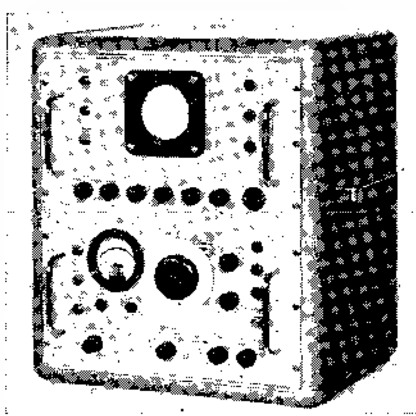
The type 190 spectrum analyser can be used at audio and supersonic frequencies to analyse the spectrum of a complex or simple signal. The frequency range of the standard instrument is from 500c/s to 90kc/s, but custom-built equipment for other frequency ranges is manufactured to order.

The analyser provides two alternative presentations of the spectrum. For rapid qualitative analysis, the spectrum is displayed on a cathode-ray oscilloscope. The X co-ordinate of the trace represents the frequency scale and the Y co-ordinate the power level at any frequency. Accurate calibration of the frequency scale is made possible by imposing a marker blip on the trace derived from a calibrated manually controlled oscillator.

In order to make an accurate analysis of the power distribution at any part of the spectrum, the marker blip is lined up on the point of interest and the analyser switched to 'manual' operation. The manual oscillator setting now determines the centre frequency at which the measurement is made by means of a thermocouple milliameter. The signal power level is determined by using an accurate calibrating oscillator, the meter indication always being set to a constant value by attenuator adjustment.

The automatically swept oscillator can be switched to scan either 1 to 7kc/s or 1 to 100kc/s across the tube at a rate which may be controlled by the user.

The power in a part of the spectrum can only be expressed with reference to



the bandwidth in which this is measured, in this equipment 1kc/s or 70c/s. The response of the narrow-band filter represents a close approach to the ideal 'plateau' as it incorporates no less than six staggered crystal filter elements.

In addition to its function as a spectrum analyser at frequencies up to 90kc/s, the instrument can also be employed for work at higher frequencies by the use of a suitable superheterodyne receiver in which the i.f. must naturally have a minimum bandwidth of 100kc/s.

EE 13 761 *for further details*

ZENER VOLTAGE REGULATORS

Texas Instruments Ltd, Dallas Road, Bedford, Bedfordshire

Texas Instruments have announced a new range of silicon Zener voltage regulators, types 1S501—1S516 as well as new silicon Zener voltage reference diodes.

The 1S501—1S516 series is stud-mounted and covers the voltage range of 22 to 91V in 16 steps. These new regulators are rated at 8W dissipation, are available with Zener voltage tolerances of 5 per cent or 10 per cent, and can be supplied with anode to stud, or as double anode clippers.

The 1S207—1S218 series of Zener voltage reference diodes covers the range 3·6 to 10V in 12 steps. They are available with Zener voltage tolerances of 5 or 10 per cent and are hermetically sealed in small molybdenum-glass sleeves, similar to the Texas range of 400mA silicon rectifiers.

These semiconductor devices are available in prototype quantities now and large-scale production is commencing at Bedford.

EE 13 762 *for further details*

HIGH TEMPERATURE CAPACITORS

Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, London E.C.1

Johnson, Matthey & Co. announce the development of a new range of silvered mica capacitors for defence applications.

The new capacitors are designed for use over the temperature range of −70°C to +250°C and are protected by a dip coating of special silicone rubber. Fired construction is employed, the individual silvered mica plates being bonded into a solid block.

Two standard sizes are manufactured and any capacitance value in the range 5pF to 0·05μF can be supplied to a minimum tolerance of ±½ per cent or 1pF, whichever is the greater. Each size is available rated at 200 or 350V peak for temperatures up to 100°C. These ratings are reduced to 100 and 200V peak at 250°C.

EE 13 763 *for further details*

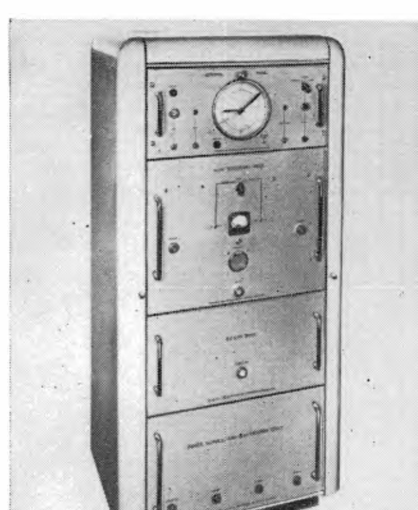
SEQUENCE TIMER

(Illustrated below)

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

This new 'programme recording and playback equipment' has been designed to control the sequence and timing of industrial processes and is capable of providing up to forty-nine on-off switching actions on independent or common circuits as required. It can be used to programme any process in which the sequence of events can be controlled by the timed operation of a number of contacts.

The flexibility of the system enables the timing of any part of the process to be altered at will and the sequence of events can be adjusted without major wiring modifications. The command



pulses are stored on standard magnetic recording tape. From a single master tape an unlimited number of copies can be made for the independent control of a number of machines or for use by the various branches or overseas subsidiaries of a manufacturer. The system enables the secrecy of unique processes to be safeguarded.

A total operating period of up to 2½ hours is standard but special units are available for longer periods.

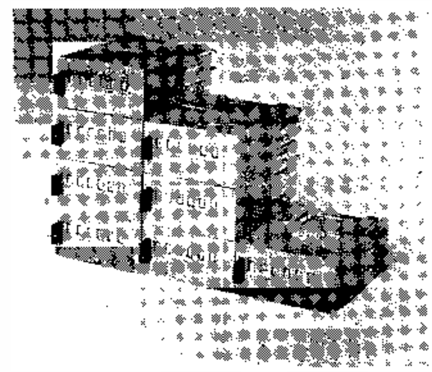
EE 13 764 *for further details*

ELECTROMAGNETIC COUNTERS

(Illustrated below)

Lancashire Dynamo Electronic Products Ltd, Rugeley, Staffordshire

Lancashire Dynamo Electronic Products Ltd have obtained the sole selling rights in Great Britain for the Hengstler electromagnetic counters types F43 and F43M2.



Capable of counting at speeds up to 2 400 impulses per minute and having a long life expectancy, the Hengstler F43 and F43M2 units are basically similar in design, both providing six digit indications: the F43 incorporates manual reset and the F43M2 has both manual and electrical reset. Both types are available in high speed or low speed versions and at a number of preferred d.c. operating voltages. Rectifier units for operating at certain preferred a.c. ratings will also be available in the near future.

As will be seen from the accompanying illustration, each counter is comprised of two parts, the counter itself and a socket box into which it is inserted, electrical connexions being made by means of a plug and socket at the rear. The socket box is a pressure die casting with provision for mechanically interlocking with adjoining boxes on all four sides so that, where required, large banks of counters can be assembled into a single structure. The incoming connexions are made on the rear of the socket box.

Lancashire Dynamo Electronic Products Ltd wish to emphasize that the arrangements outlined above apply to the marketing of the Hengstler high-speed counters types F43 and F43M2 only. Other mechanical and slow speed electro-mechanical counters produced by J. Hengstler K.G., will continue to be marketed in this country through their agents, Brian R. Morris & Co. Ltd.

EE 13 765 *for further details*

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

All London meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London W.C.1. at 6.30 p.m.

Computer Group Meeting

Date: 7 October.
Lecture: Some Reflections on Computer Design.
By: W. Renwick.

Students Meeting

Date: 14 October.
Lecture: The Use of Transistors in Communication and Control.
By: E. Wolfendale.

Medical Electronics Group Meeting

Date: 21 October.
Lecture: Aviation Medicine.
By: P. V. Byford.

BRITISH SOUND RECORDING ASSOCIATION

Date: 16 October. Time: 7.15 p.m.
Held at: Royal Society of Arts, John Adam Street, Adelphi, London W.C.2.
Lecture: Looking at Stereo (on a c.r.o.).
By: B. Shelley.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings will be held at Savoy Place, commencing at 5.30 p.m.

Ordinary Meeting

Date: 9 October.
Inaugural Address.
By: Sir Willis Jackson, President.

Electronics and Communications Section

Date: 28 October.
Chairman's Address: Development of Eurovision
By: M. J. L. Pulling.

Measurement and Control Section

Date: 13 October.
Chairman's Address: The Relationship of Physical Mechanisms to Psychological Processes.
By: Professor A. Tustin.
Date: 27 October.
Discussion: Future Trends in Memory Stores for High Speed Digital Computers.
Opened by: W. Renwick.

Cambridge Electronics and Communications Group

Date: 13 October. Time: 8 p.m.
Held at: Cavendish Laboratory, Free School Lane, Cambridge.
Lecture: New Amplifying Techniques.
By: C. W. Oatley.

East Anglian Sub-Centre

Date: 19 October. Time: 7.30 p.m.
Held at: Assembly House, Norwich.
Lecture: A Review of Work towards Nuclear Energy from Controlled Thermonuclear Reaction.
By: D. W. Fry.

East Midland Centre

Date: 20 October. Time: 6.30 p.m.
Held at: College of Arts and Crafts, Nottingham.
Lecture: The Recognition of Moving Vehicles by Electronic Means.
By: T. S. Pick and A. Readman.

North-Eastern Measurement and Electronics Group

Date: 19 October. Time: 6.15 p.m.
Held at: Rutherford College of Technology, Newcastle upon Tyne.
Chairman's Address.
By: C. C. Baxendale.

North-Western Electronics and Communications Group

Date: October 7. Time: 6.15 p.m.
Held at: Engineers' Club, Albert Square, Manchester.
Chairman's Address: Some Features of the Post Office Telephone and Television Networks in the North-West.
By: S. D. Mellor.

Western Utilization Group

Date: 26 October. Time: 6 p.m.
Held at: S.W.E.B. Demonstration Theatre, Colston Avenue, Bristol.
Section Chairman's Address.
By: T. E. Houghton.

West Wales (Swansea) Sub-Centre

Date: 8 October. Time: 6 p.m.
Held at: Conference Room, S.W.E.B. Showrooms, The Kingsway, Swansea.
Chairman's Address.
By: J. Harley.

Oxford

Date: 7 October. Time: 7 p.m.
Held at: Southern Electricity Board Service Centre, 37 George Street, Oxford.
Lecture: New Methods of Exploiting Electronic Computers in the Industry.
By: H. McG. Ross.

Sheffield Sub-Centre

Date: 21 October. Time: 6.30 p.m.
Held at: Grand Hotel, Sheffield.
Chairman's Address: Permanent Magnets and the Electrical Engineer.
By: F. G. Tyack.
Date: 26 October. Time: 6.45 p.m.
Held at: Angel Hotel, Brigg.
Chairman's Address as above.

South Midland Electronics and Communications Group

Date: 26 October. Time: 6 p.m.
Held at: James Watt Institute.
Lecture: Reliability of Electronic Equipment.
By: A. G. Field.

Southern Centre

Date: 7 October. Time: 6.30 p.m.
Held at: C.E.G.B. Offices, Portsmouth.
Chairman's Address
By: W. D. Mallinson.
Date: 20 October. Time: 6 p.m.
Held at: Technical College, Boundary Road, Farnborough.
Lecture: Space Research.
By: R. L. F. Boyd.
Date: 21 October. Time. 6.30 p.m.
Held at: Technical College, Brighton.
Lecture: Discrimination between H.R.C. Fuses.
By: E. Jacks.

Northern Ireland Centre

Date: 13 October. Time: 6.30 p.m.
Held at: David Keir Building, Queen's University, Stranmillis Road, Belfast.
Chairman's Address: The Provision of Commercial Television in Northern Ireland.
By: T. S. Wylie.

INSTITUTION OF PRODUCTION ENGINEERS

Scottish Section

Date: 7 October. Time: 7.30 p.m.
Held at: Windmill Hotel, Arbroath, Dundee.
Lecture: Numerical Control of Machine Tools.
By: Messrs. Brett and Tack.

South Eastern Section

Date: 22 October. Time: 7 p.m.
Held at: Royal Commonwealth Society, Northumberland Avenue, Strand, London W.C.2.
Lecture: Computers as Applied to Production Control.
By: B. L. J. Hart.

JUNIOR INSTITUTION OF ENGINEERS

Date: 16 October. Time: 7 p.m.
Held at: Pepys House, 14 Rochester Row, Westminster, London S.W.1.
Lecture: Programming a Computer.
By: A. C. Quaterman.

THE RADAR AND ELECTRONICS ASSOCIATION

Date: 27 October. Time: 7.30 p.m.
Held at: Royal Society of Arts, John Adam Street, Adelphi, London W.C.2.
Lecture: Electronics and Supersonic Flight (with special reference to the English Electric Aviation 'Lightning' aircraft).
By: F. W. Page.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

All London meetings will be held at Mansion House, 26 Portland Place, London W.1 at 7 p.m.

Data Processing Section

Date: 14 October.
Lecture: An Automatic Analogue Computer for Missile Homing Investigations.
By: J. G. Thomason.
Date: 27 October.
Lecture: Modern Developments in Optical Instruments.
By: J. H. Bach.

THE TELEVISION SOCIETY

Date: 22 October. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London W.C.2.
Discussion: New Television Standards: Their Effect on British Television.

PUBLICATIONS RECEIVED

INTRODUCTION TO BESSEL FUNCTIONS by F. Bowman is based on lectures given from time to time in the College of Technology, Manchester, and it is hoped that it may serve as an introduction to the larger treatizes on Bessel functions and their applications. Dover Publications Inc., New York. Price $1.35.

PROCEEDINGS OF THE FOURTH NATIONAL CONFERENCE ON TUBE TECHNIQUES is a report on the conference sponsored by the Advisory Group on Electron Tubes held from 10 to 12 September, 1958. New York University Press, Washington Square, New York. Price $7.50.

CAST ACRYLIC SHEET is the eleventh in a series of Ministry of Supply reports on plastics in the tropics and is published in London by Her Majesty's Stationery Office. Price 5s.

SYMPOSIUM ON 'POTTED CIRCUIT' TECHNIQUES was held at the Royal Radar Establishment of the Ministry of Supply, Great Malvern, on 15 and 16 April, 1959, and a report of the proceedings has now been issued. This Symposium was the first of its kind to be held in Britain in relation to the application of 'potting' for electronic equipment. Its purpose was to review and discuss current and new developments in encapsulation materials, applications, and more particularly, techniques to be applied to the various forms of 'potting'.

PROCEEDINGS OF THE EIA SYMPOSIUM ON NUMERICAL CONTROL SYSTEMS FOR MACHINE TOOLS is published by the Engineering Publishers, GPO Box 1151, New York 1, and deals with the symposium held in Los Angeles, California, on 17 and 18 September, 1957. This was sponsored by the Engineering Department of the Electronics Industries' Association with participation by Aircraft Industries' Association, National Machine Tool Builder Association, National Electrical Manufacturers' Association and Office Equipment Manufacturers' Institute.

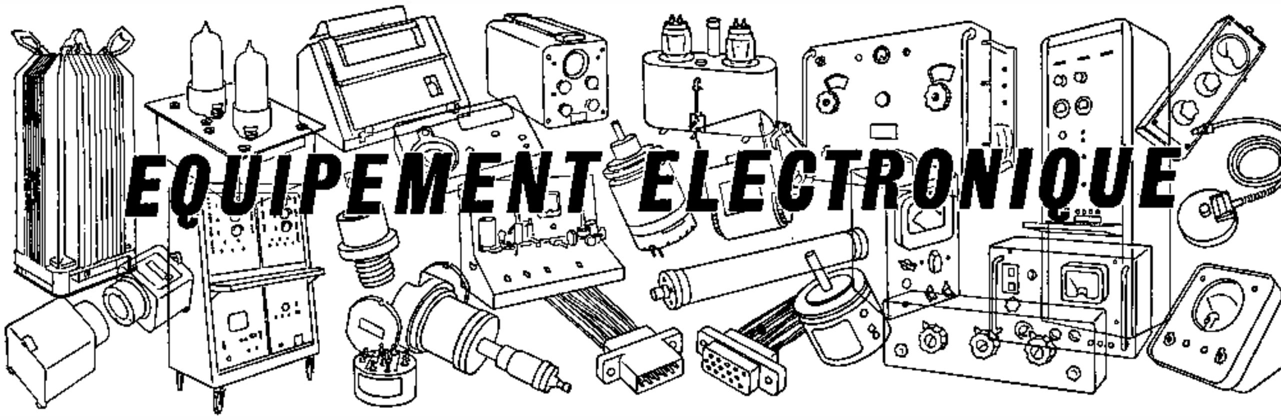
BRITISH STANDARD FOR PERIODICALS OF REFERENCE VALUE: FORM AND PRESENTATION (B.S. 2509:1959) is intended for those who edit or contribute to scientific, technical and other periodicals used for reference over a long period. This new publication brings up to date the previous standard, published in 1954 under the title 'Layout of Periodicals'. All the clauses of the 1954 edition have been retained with modifications incorporated in the light of comment and criticism over the past five years. The main change is in the clauses on 'Articles and other contributions' which now advocate internationally recommended practice; the new standard itself conforms in the main to a recommendation of the International Organization for Standardization (ISO/R8). British Standards Institution, Sales Branch, 2 Park Street, London W.1. Price 6s. Postage will be charged extra to non-subscribers.

NEL RELIABILITY BIBLIOGRAPHY SUPPLEMENT 2 should be inserted in the basic edition of the NEL Reliability Bibliography published in May, 1956, which was distributed to sections of the Department of Defence and to a selected list of appropriate government contractors. Now the basic volume and the supplements of the Bibliography are being made available at a nominal charge from the Office of Technical Services, Department of Commerce, Washington 25, D.C.

TRANSISTOR SOUND RECEIVER SR2/2 is an Application Report published by Siemens Edison Swan Ltd, Radio Division, 155 Charing Cross Road, London W.C.2, describing the design of a portable radio receiver employing semiconductor devices throughout. It operates on a 6V supply provided by four dry cells, and delivers an output of over 200mW. The mean sensitivity is 90µV/m for 5mW output, and the design has been arranged to accommodate the normal spread of transistor and component tolerances. The receiver embodies an internal ferrite rod aerial, and complete portability is obtained with an all-up weight, excluding cabinet, of 3½lb.

HUNT'S BULLETIN Volume 1, Issue 1 was published in June 1959 and was the first of a series intended to be published every two months containing articles of interest to the Electronic and Electrical industries. The Company is issuing a new general catalogue C273 which will be bound in loose leaf covers and the bulletins are designed for insertion in those covers. The first issue dealt with many new developments in ceramic capacitors largely effected by the Company's French associate Le Condensateur Ceramique of Montreuil, and in current use in the Company's ceramic production. A. H. Hunt ((Capacitors) Ltd, Bendon Valley, Garratt Lane, Wandsworth, London S.W.18.

(Pages 567 to 576 French and German section)



EQUIPEMENT ELECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 562 à 565

DÉTECTEUR DE VIBRATIONS

(Illustration à la page 562)

C. F. R. Giesler Ltd, Small Hall, River Place, Essex Road, Islington, London N.1

Cet appareil portatif a été calculé pour mesurer l'amplitude des vibrations dans tous les types de machines tournantes. Il a une gamme de fréquences de 6 Hz à 300 Hz, correspondant à une gamme de vitesses de rotation de 360 à 18000 tours par minute, et il comprend un amplificateur, un enregistreur d'amplitude, un bloc d'alimentation et un transducteur séismique. L'appareil a été prévu pour l'emploi d'un enregistreur à plume unique ou à double plume. Il est muni de deux prises d'entrée pour le branchement de deux transducteurs, un commutateur étant employé pour le choix.

La gamme complète d'amplitudes pouvant être lues sur l'enregistreur s'étend de 0,0001", à 0,1". La sensibilité totale de l'appareil est contrôlée par le 'commutateur de sensibilité', et l'enregistreur a deux échelles, proportionnées de façon à ce qu'une déviation d'aiguille normale puisse être obtenue sans amplitude de vibration, à partir de 0,00025".

EE 13 751 *pour plus amples renseignements*

ANALYSEUR DE FONCTION DE TRANSFERT

(Illustration à la page 562)

Servo Consultant Ltd, 17 Woodfield Road, London W.9

Cet équipement a été étudié afin de constituer un moyen stable et précis d'essai de servo-mécanismes et d'appareillages de commande. Ses principaux avantages sont les suivants:

(1) Le système électromécanique utilisé est précis et sans glissement.

(2) L'équipement est insensible à la distorsion non-linéaire causée par le système soumis à la vérification. Seule la fondamentale est accusée.

(3) Une indication directe est donnée de l'angle de phase et de l'amplitude.

(4) Les servo-systèmes fonctionnant sur courant alternatif peuvent être contrôlés directement.

(5) Les performances de modulateurs, de redresseurs et d'amplificateurs magnétiques peuvent être étudiées.

(6) La possibilité de couplage direct à des commandes mécaniques et hydrauliques sans faire usage de systèmes d'entraînement.

(7) L'équipement est entièrement autocommandé.

Deux modulateurs électromécaniques sont actionnés par un moteur à vitesse variable, dont la vitesse est égale à la fréquence du signal d'essai. Cette vitesse peut être réglée à la main au moyen d'un potentiomètre hélicoïdal de précision.

Le modulateur 1 a un enroulement stationnaire alimenté par un signal porteur d'amplitude constante. L'enroulement de sortie de ce dispositif de résolution, qui est rotatif, produira donc un signal porteur modulé qui, dans les cas d'un servo-mécanisme fonctionnant sur courant alternatif, servira alors directement de signal d'essai. Le courant porteur fourni au dispositif de résolution pourra être obtenu soit d'un générateur incorporé de 2 kHz soit de n'importe quelle source extérieure à l'équipement. Dans le cas d'un servo-système à courant continu, ce signal sera démodulé afin d'obtenir le signal à la fréquence d'essai. De plus, un couplage mécanique pourra être introduit entre le générateur et le système soumis à l'essai. La sortie du système mis à l'épreuve pourra encore être soit un courant porteur modulé, tel qu'une sortie de 400 Hz d'un synchro, soit un signal à la fréquence d'essai. Pareil signal sera amplifié et transmis au dispositif de résolution 2, qui tourne en synchronisme avec le dispositif 1. Dans le cas d'un système c.c. le signal de sortie du système en cours d'épreuve sera employé pour moduler le courant porteur, qui sera ensuite appliqué au dispositif de résolution 2. Le stator de ce dispositif est relié par une connexion triphasée à un autre dispositif de résolution (dispositif 3), accouplé au cadran indicateur d'angle. L'angle entre le stator et le rotor de ce dernier dispositif correspond à la différence de phase entre les signaux d'entrée et de sortie de l'équipement lorsque le signal de l'enroulement de sortie de ce dispositif est au zéro. Ce dispositif porte également un autre enroulement à angles droits avec le premier. La sortie de cet enroulement fournit une tension proportionnelle à l'amplitude, au signal d'entrée. La grandeur de ce vecteur est indiquée par un compteur à lecture directe.

L'équipement est logé dans un meuble robuste en alliage d'aluminium. Le panneau avant, qui porte tous les compteurs et les commandes est monté sur charnière afin de donner accès facilement à toutes les parties importantes de l'équipement. Deux châssis verticaux portant tous les circuits électroniques sont accessibles des deux côtés, sans qu'il soit besoin de les sortir du meuble. Le générateur et la boîte de vitesses du dispositif de résolution sont montés au fond du meuble. L'axe de sortie est accessible du côté droit.

EE 13 752 *pour plus amples renseignements*

TRANSDUCTEUR DE PRESSION

(Illustration à la page 562)

The Solartron Electronic Group Ltd, Thames Ditton, Surrey

La Société Solartron vient de conclure avec la Svenska Flygmotor Aktiebolaget un contrat de fabrication et de vente s'étendant au monde entier (à l'exception des pays scandinaves), pour un nouveau transducteur de pression à "cylindre vibrant" qui serait unique en son genre.

La production de cet appareil a déjà commencé en Grande-Bretagne et il a, en outre, été agréé par le Ministère des Approvisionnements en ce qui à trait à son incorporation dans des engins téléguidés et d'autres équipements en cours d'essai.

Le "cylindre vibrant" applique un principe de construction entièrement nouveau car il est petit, précis et extrêmement robuste en tant que transducteur de pression. Le signal de sortie à fréquence variable le rend particulièrement indiqué pour les enregistrements sur bande magnétique comme, par exemple, ceux effectués avec l'Enregistreur de Données Solartron (Solartron Data Recording Equipment), ainsi que pour les conversions de données analogiques en données numériques par comptage d'impulsions. L'équipment, utilisant le signal du transducteur, peut être placé à une distance considérable du point de mesure, sans qu'il y ait perte de précision.

L'élément sensible à la pression du transducteur est un système vibrant mécanique dont la fréquence naturelle dépend de la pression appliquée. Cet élément vibrant détermine la fréquence d'un oscillateur à réaction électronique, fournissant ainsi une tension de sortie d'une amplitude constante mais dont la fréquence varie selon la pression appliquée. Le transducteur a été prévu pour faire partie d'un système d'enregistrement sur bande magnétique de données d'engins à statoréacteurs en cours de vol.

A cet effet, un transducteur avec sortie à fréquence variable a été trouvé le plus indiqué. Le signal de fréquence variable est facilement convertible en signal numérique pouvant ensuite être traité par un calculateur numérique.

Ces transducteurs sont livrables pour des pressions d'entrée de 1,75 kg/cm², 3,50 kg/cm², 7 kg/cm² et 42 kg/cm².

EE 13 753 *pour plus amples renseignements*

VOLTMÈTRE ÉLECTRONIQUE SENSIBLE

(Illustration à la page 563)

Dawe Instruments Ltd, 99 Uxbridge Road, London W.5

Ce voltmètre électronique Type 614C est un instrument portatif et compact pour la mesure de tensions de 80 μV à 300 V, à des fréquences de 5 Hz à 100 kHz. Il permet aussi de mesurer des tensions alternatives avec courant continu superposé, quoi-qu'un niveau de pointe de 1 kV ne doive pas être dépassé.

Le circuit de base comprend un amplificateur à grand gain qui alimente, à travers un atténuateur, un compteur indicateur-redresseur biphasé. Les deux étages de cet atténuateur—l'un à l'entrée et l'autre séparant les étages d'amplification doubles—sont, tous deux, commutés par un commutateur de gammes à 13 directions de façon à obtenir des déviations totales de 300 μV, 1 mV, 3 mV, 10 mV, 30 mV, 100 mV, 300 mV, 1 V, 3 V, 10 V, 30 V, 100 V et 300 V. L'étage d'atténuation suivant la première section d'amplification est introduit en premier lieu afin d'améliorer le rapport signal/bruit aux niveaux d'entrée plus élevés. Une précision d'étalonnage très poussée et une grande stabilité sont assurées par l'application d'une réaction négative aux deux étages d'amplification et par l'incorporation d'une alimentation haute tension stabilisée.

Le compteur est étalonné en valeurs efficaces pour une entrée sinusoïdale, les indications représentant la valeur moyenne des deux alternances. L'échelle de tensions logarithmiques donne une précision de lecture égale pour toutes les tensions (supérieure à $\pm 2\%$ d'indication avec une entrée sinusoïdale). L'erreur de fréquence est inférieure à $\pm 3\%$, de 5 Hz à 100 kHz. En plus des 13 gammes de tensions normales, une échelle de décibels linéaires auxiliaire donne une gamme de -80 à $+50$ dB par rapport à 1 volt.

Un jack de sortie permet de contrôler le signal ou d'employer l'instrument comme amplificateur avec un gain atteignant jusqu'à 88 dB. Lorsqu'il est employé de cette façon, la sortie est d'environ 8 volts efficaces pour une déviation totale, dans une charge minimum de 10 ohms. L'impédance de source est d'environ 20 ohms.

L'impédance d'entrée équivant à peu près à une résistance de 10 megohms, en parallèle avec une capacité de 25 pF, jusqu'à la gamme de 10 mV, et avec une capacité d'environ 10 pF aux gammes supérieures.

EE 13 754 *pour plus amples renseignements*

INDICATEUR D'INTENSITÉ D'ÉTINCELLES

(Illustration à la page 563)

Lexor Electronics Ltd, 25 Allesley Old Road, Coventry, Warwickshire

Cet appareil qui comporte un voltmètre électronique, indique une série très étendue de niveaux d'étincelles et il peut être facilement étalonné pour répondre à des applications particulières.

Il est couramment utilisé pour le contrôle de production et de développement d'appareils à courant continu, de commutateurs, de contacteurs, de pièces détachées, etc. ainsi que pour les travaux de suppression de parasites. Il existe également des modèles pour installation permanente, pour des besoins de contrôle continu intéressant les systèmes d'entretien et de sécurité.

L'appareil peut indiquer des étincelles qui ne sont normalement pas visibles, comme celles se produisant au-dessous des porte-balais ou dans des instruments sous essai dans une chambre d'altitude. Il est livrable en modèle pour étincelles continues ou pour impulsions isolées.

Un filtre d'entrée sépare les impulsions d'étincelles de potentiels stationnaires allant jusqu'à 1 kV c.c. et l'impulsion finale est donnée sur un robuste appareil de lecture à cadre mobile de 89 mm. La sensibilité maximum est de 50 mV pour une déviation totale.

EE 13 755 *pour plus amples renseignements*

OSCILLOSCOPE UNIVERSEL

(Illustration à la page 563)

Marconi Instruments Ltd, St. Albans, Hertfordshire

L'oscilloscope universel TF 1330 est un appareil de haute qualité ayant toutes les caractéristiques voulues pour une présentation détaillée et des mesures précises de formes d'ondes. Sa réponse de fréquence s'étend du courant continu à 15 MHz, sa sensibilité est variable en sept paliers de 15 mV à 50 V/cm et un signal de 300 mV seulement suffit à couvrir tout l'écran de 6 $\times$ 10 cm. Une ligne à retard concentré équilibré est incorporée au système de déviation verticale pour permettre l'observation sans distorsion de l'ensemble d'un signal soumis à l'examen.

La vitesse du balayage de la base de temps varie de 1 sec/cm à 0,1 μsec/cm en quinze gammes et elle peut être augmentée jusqu'à 0,02 μsec/cm par l'étalement X. Chaque gamme peut être facilement normalisée contre une sortie extérieure. Le déclenchement peut être effectué intérieurement ou extérieurement, ou encoure à partir d'une source de fréquence d'alimentation intérieure. Un choix de dispositifs de couplage c.c. et c.a. garantit la stabilité du déclenchement avec des signaux de fréquences très basses jusqu'à, au moins, 15 MHz et prévoit, de même, la synchronisation de champ de télévision.

Les potentiomètres étalonnés directement servent à mesurer tant l'amplitude de signal que la durée. L'étalonnage est précis jusqu'à $\pm 2\%$ et il est indépendant de l'étalement X ou du gain Y.

Le tube cathodique de 127 mm a un accélérateur de post-déviation spirale et il fonctionne à très haute tension de 10 kV. Ces particularités s'allient pour donner une image de haute brillance linéaire à de grandes vitesses d'enregistrement ou à des régimes de répétition à faibles impulsions. Afin de permettre de tirer le plus grand parti possible des vitesses de base de temps très lentes, un tube cathodique à persistance prolongée peut être fourni à la place du tube ordinaire. Le capot de vision autour de l'écran du tube peut être enlevé pour pouvoir y fixer une caméra d'oscilloscope conventionnel.

La facilité du fonctionnement est assurée par des commandes concentriques, disposées fonctionnellement et à codage de couleurs. Si la trace disparaît de l'écran, dans n'importe quelle direction, sa position pourra étra immédiatement révélée en pressant un boutonpoussoir. Il sera ensuite fort simple de rétablir l'image au centre de l'écran, au moyen de deux commandes de glissement. Grâce à un support basculant, fixé sous l'avant de l'appareil, la vision est grandement facilitée.

Il existe aussi un modèle à double trace, le Type 1331, dont les performances de base sont les mêmes que celles du Type 1330, mais qui offre en plus la commutation des faisceaux électroniques pour la présentation de deux signaux, soit par balayages alternants, soit à un régime de commutation de 100 kHz.

EE 13 756 *plus pour amples renseignements*

COMPTE-TEMPS DE LIAISON ÉLECTRIQUE DE BOBINES

(Illustration à la page 564)

Avo Ltd, Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

Lorsqu'on emploie des fils d'enroulements auto-adhérent, la méthode la plus rapide d'effectuer la liaison d'une bobine est de faire passer au travers d'elle, pendant quelques instants, un courant d'une amplitude appropriée. Vu que la tension requise pour ce faire peut être assez élevée, il faut tout d'abord vérifier l'isolement entre le fil métallique et le manchon sur lequel la bobine a été enroulée.

Le compte-temps de liaison AVO effectue ces deux opérations de la manière suivante:

Les deux extrémités de la bobine sont fixées aux bornes du dispositif, tandis que des commutateurs, actionnés simultanément avec les deux mains, donnent une indication directe en mégohms de la valeur de l'isolement entre la bobine et le mandrin. Si la valeur d'isolement est satisfaisante, on actionne un troisième commutateur qui fait passer un courant de chauffage à travers la bobine pendant une période de temps prédéterminée.

Etant donné que l'opérateur emploie ses deux mains au cours de ces essais, l'appareil est d'une parfaite sécurité intrinsèque. Le minutage du courant de liaison varie entre trois-quarts de seconde et deux minutes et demie, cependant que l'isolement peut être mesuré jusqu'à 500 mégohms.

L'appareil fonctionne sur courant secteur de 210 à 250 volts, 50 Hz, mais lorsque la tension de secteur est considérée comme étant trop élevée pour le travail de liaison, une tension plus basse pourra être injectée à cet effet par les bornes.

EE 13 757 *pour plus amples renseignements*

CONDENSATEURS DIÉLECTRIQUES TUBULAIRES AU PAPIER

Dubilier Condenser Co. (1925) Ltd, Ducon Works, Victoria Road, North Acton, London W.3

Ces condensateurs diélectriques tubulaires au papier Dubilier, désignés sous le Type 560, ont été étudiés pour répondre à la demande de condensateurs tubulaires diélectriques au papier de grand rendement et ayant des caractéristiques électriques exceptionellement bonnes, alliées à une longue durée. Ces condensateurs peuvent être utilisés dans une gamme de températures de $-40°C$ à $+125°C$ sur courant continu, et de $-40°C$ à $+70°C$ sur courant alternatif, sans changement de tension.

Le principe de construction est le suivant:

Un élément classique de papier et de feuilles de métal, muni d'interconnexions entre les électrodes et de terminaisons à fils, est imprégné d'une matière plastique à haute température, afin de produire un élément de condensateur solide. Les terminaisons à fils formés sont insérées dans les interconnexions et ces dernières sont mises en forme et soudées aux terminaisons à fils, produisant ainsi de robustes jonctions électriques et mécaniques. L'élément assemblé est enchâssé dans une résine d'époxie à charge minérale, garantissant une excellente résistance aux chocs et à la pénétration de l'humidité.

Ces condensateurs sont fournis dans une série de valeurs allant de 0,001 μF à 0,1 μF, la tension de service variant de 180 V c.a. à 250 V c.a., et de 350 V c.c. à 1 kV c.c., selon la capacité.

EE 13 758 *pour plus amples renseignements*

CONDENSATEUR D'ACCORD ET COMMUTATEUR DE GAMMES D'ONDES COMBINÉS

(Illustration à la page 564)

The Plessey Co. Ltd, Vicarage Lane, Ilford, Essex

La Société Plessey vient d'annoncer l'introduction de condensateurs variables à commutation, autorisant le changement normal de gammes d'ondes et les réglages d'accord sur n'importe quel récepteur radio à deux gammes d'ondes, au moyen d'un seul bouton de comande.

Les nouveaux condensateurs comportent une broche à ailettes mobiles pourvue d'un mouvement de liaison qui actionne un interrupteur auto-nettoyeur Plessey dans une certaine direction lorsqu'il est presque entièrement engrené, et dans la direction opposée lorsqu'il est presque entièrement désengrené. Il est recommandé d'effectuer la commutation de telle manière que la faible perte de course utilisable se produise à l'extrémité à haute fréquence de la bande d'ondes longues et à l'extrémité à basse fréquence de la bande d'ondes moyennes. De cette façon, toute perte d'émissions est sans importance.

Le commutateur est un ensemble bipolaire, mais il permet, en ajoutant un troisième pôle, la commutation d'une lampe de cadran ou toute opération similaire; on peut également prévoir l'indication mécanique de bandes d'ondes.

Les constructeurs soulignent, enfin, le fait qu'en plus de l'avantage, du point de vue de la vente en détail de récepteurs radio, d'un seul bouton de commande de deux gammes d'ondes, la simplification des circuits qu'implique l'emploi des nouveaux condensateurs constitue un facteur économique supplémentaire méritant bien d'être pris en considération.

EE 13 759 *pour plus amples renseignements*

APPAREIL-ÉTALON DE NIVEAU DE TRANSMISSION

(Illustration à la page 564)

Radiometer, 72 Emdruprej, Copenhagen NV, Danemark.

L'étalonneur de niveau de transmission type SNNI est un appareil de haute précision pour étalonner les instruments de mesure de niveau et mesurer avec exactitude leur réponse de fréquence.

Lorsqu'il est relié au générateur de signaux qui convient, l'appareil type SNNI fournira trois niveaux de sortie soit à circuit ouvert soit à travers une charge de 75 ohms. Ces niveaux sont:

0 dB/0·274 V = 274 mV
9 dB/0·775 V = 275 mV
1 N/0·775 V = 285 mV

La précision des niveaux de sortie est supérieure à 1%, du courant continu à 25 MHz, avec une charge de 75 ohms, et elle est supérieure à 1% jusqu'à 15 MHz avec une sortie sans charge. La réponse de fréquence est en deçà de ± 0.2%, de 10 Hz à 10 MHz.

L'appareil fonctionne selon le principe des diodes saturées ce qui assure une réponse rapide, une vitesse de lecture incrémentale de 0.1% et une indication de valeur efficace réelle. Enfin, il est muni d'un disjoncteur à maxima des plus sûrs.

Quoique prévu pour l'emploi avec le matériel téléphonique à fréquence porteuse, l'appareil type SNNI peut, cependant, être également affecté à de nombreux autres usages

EE 13 760 *pour plus amples renseignements*

ANALYSEUR DE SPECTRE

(Illustration à la page 565)

James Scott & Co. (Electrical Engineers) Ltd, 68 Brockville Street, Carntyne Industrial Estate, Glasgow, E.2.

L'Analyseur de spectre type 190 peut être employé à des fréquences acoustiques et ultra-sonores pour analyser le spectre d'un signal simple ou composé. La gamme de fréquences de l'appareil standard va de 500 Hz à 90 kHz, mais des appareils pour d'autres gammes peuvent être construits sur commande.

L'Analyseur offre un choix de deux images différentes du spectre. Pour l'analyse qualitative rapide, le spectre est présenté sur un oscilloscope cathodique. La co-ordennée X de la trace représente l'échelle de fréquence et la coordonnée Y le niveau de puissance à n'importe quelle fréquence. L'étalonnage précis de l'échelle de fréquences s'effectue par l'imposition d'un top repère sur le trace, obtenu à partir d'un oscillateur étalonné à fonctionnement manuel.

Afin d'assurer une analyse exacte de la puissance répartie en n'importe quelle partie du spectre, on fixe le top repère sur le point d'intérêt et on branche l'analyseur sur le fonctionnement manuel. Le réglage de l'oscillateur manuel détermine alors la fréquence nominale à laquelle la mesure est effectuée au moyen d'un milliampèremètre thermique. On détermine le niveau de puissance du signal en utilisant un oscillateur d'étalonnage précis, l'indication de l'ampèremètre étant toujours placée à une valeur constante par réglage d'atténuation.

L'oscillateur à balayage automatique peut être branché pour le balayage de 1 à 7 kHz ou de 1 à 100 kHz à travers le tube, à un taux pouvant être contrôlé par l'usager.

La puissance dans une partie du spectre ne peut être exprimée qu'en fonction de la largeur de bande dans laquelle elle est mesurée: dans cet appareil, 1 kHz ou 70 Hz. La réponse du filtre à bande étroite constitue un pas décisif vers le "plateau" idéal car il ne comprend pas moins de six éléments à filtres piézoélectriques décalés.

En plus de ses fonctions d'analyseur de spectre à des fréquences allant jusqu'à 90 kHz, on peut également compleyer l'appareil pour des travaux à des fréquences plus élevées par l'utilisation d'une superhétérodyne appropriée, dans laquelle la fréquence intermédiaire doit, évilemment, avoir une largeur de bande minimum de 100 kHz.

EE 13 761 *pour plus amples renseignements*

RÉGULATEURS DE TENSION ZÉNER

Texas Instruments Ltd, Dallas Road, Bedford, Bedfordshire

La Société Texas Instruments Ltd. vient d'annoncer la réalisation d'une nouvelle série de régulateurs de tension Zéner au silicium, types 1S501—1S516, ainsi que de nouvelles diodes de référence de tension Zéner au silicium.

La série des régulateurs types 1S501—1S516, montés sur goujons, couvre la gamme de 22 à 91 volts en 16 plots. Ces nouveaux régulateurs ont une dissipation nominale de 8 W et sont livrables avec des tolérances de tension Zéner de 5% ou de 10%. Ils peuvent, en outre, être fournis avec anode au goujon ou comme séparateurs à double anode.

La série des diodes de référence de tension Zéner 1S207—1S218 couvre la gamme de 3,6 à 10 Volts en 12 plots. Elles sont livrables avec des tolérances de tension Zéner de 5% ou de 10% et elles sont scellées hermétiquement dans de petits manchons en verre au molybdène, semblables à la série des redresseurs au silicium Texas de 400 mA.

Ces dispositifs semi-conducteurs sont fournis en quantités prototypes jusqu'à présent et leur production sur une grande échelle commence actuellement à Bedford.

EE 13 762 *pour plus amples renseignements*

CONDENSATEURS POUR HAUTES TEMPÉRATURES

Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, London E.C.1

La Société Johnson, Matthey & Co. Ltd. vient d'annoncer la réalisation d'une nouvelle série de condensateurs au mica argenté pour des applications défensives.

Ces nouveaux condensateurs ont été étudiés pour être utilisés dans une gamme de températures de −70°C à +250°C et ils sont protégés par un revêtement par immersion en caoutchouc spécial au silicone. Ils sont fabriqués par mise au feu, les plaques individuelles de mica argenté étant jointes en un bloc solide.

Ils sont livrables en deux formats courants et n'importe quelle valeur de capacité dans la gamme de 5 pF à 0,05 pF peut être fournie pour une tolérance minimum de ±1/2% ou 1 pF, selon celle qui est la plus élevée. Les deux formats sont prévus pour une tension nominale de 200 V ou une tension de pointe de 350 V, pour des températures de 100°C. Ces tensions sont réduites à 100 V et 200 V respectivement pour des températures de 250°C.

EE 13 763 *pour plus amples renseignements*

DÉCLENCHEUR PÉRIODIQUE

(Illustration à la page 565)

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

Ce nouvel appareil d'enregistrement et de réenregistrement de programmes a été prévu pour commander le cadencement et le minutage de procédés industriels et il peut effectuer jusqu'à 49 commutations "arrêt-marche" sur des circuits indépendants ou communs, selon les besoins. On peut l'utiliser pour régler le programme de n'importe quel processus dont le cadencement des opérations peut être commandé par fonctionnement chronométré d'un certain nombre de contacts.

La souplesse du système permet de changer à volonté le minutage de n'importe quelle partie du processus, le cadencement des opérations pouvant être ajusté sans que cela entraîne d'importantes modifications de câblage. Les impulsions de commande sont "emmagasinées" sur bande magnétique ordinaire. Une quantité illimitée d'exemplaires peut être tirée d'une bande originale, soit par commande indépendante d'un nombre de machines, soit pour l'usage des diverses succursales ou filiales à l'étranger d'un fabricant. Le système permet, enfin, de sauvegarder le secret de certains procédés uniques.

La durée de fonctionnement courante est de 2 heures et demie, mais des modèles spéciaux peuvent être fournis pour de plus longues durées.

EE 13 764 *pour plus amples renseignements*

COMPTEURS ÉLECTROMAGNÉTIQUES

(Illustration à la page 565)

Lancashire Dynamo Electronic Products Ltd, Rugeley, Staffordshire

La Société Lancashire Dynamo Electronic Products Ltd. a obtenu les droits exclusifs de vente en Grande-Bretagne des compteurs électromagnétiques Hengstler Types F43 et F43M2.

Capables de compter à des vitesses atteignant 2400 impulsions par minute et assurés d'une longue durée de fonctionnement, les compteurs Hengstler F43 et F43M2 sont, fondamentalement, de conception semblable, les deux appareils fournissant des indications à six signes. Le type F43 est à réenclanchement manuel tandis que le F43M2 est à réenclanchement tant manuel qu'électrique. Les deux types sont fournis en version à grande vitesse et à vitesse réduite, ainsi que pour un nombre au choix de tensions de fonctionnement c.c. Des redresseurs pour le fonctionnement à certaines tensions alternatives pourront également être obtenus dans un proche avenir.

Ainsi qu'on le voit dans la gravure s'y rapportant, chaque compteur est formé de deux parties, soit le compteur proprement dit et une boîte raccord dans laquelle il est fixé, les connexions électriques s'effectuant au moyen d'une fiche et d'une prise à l'arrière. La boite raccord est une pièce moulée en coquille permettant l'enclenchement mécanique avec des boîtes contigües, sur tous les quatre côtés, de manière à pouvoir assembler, s'il y a lieu, de grandes rangées de compteurs en une seule structure. Les jonctions d'entrée s'effectuent à l'arrière de la boîte raccord.

La Société Lancashire Dynamo Electronic Products Ltd. tient à préciser que les susdites dispositions ne s'appliquent qu'à la vente des compteurs à action rapide Hengstler Types F43 et F43M2. Quant aux autres compteurs mécaniques et électromécaniques à action lente, fabriqués par la Maison J. Hengstler K. G., ils continueront à être vendus en Grande-Bretagne par leurs agents, soit la Société Brian R. Morris & Co. Ltd.

EE 13 765 *pour plus amples renseignements*

Résumés des Principaux Articles

Télécommande de position numérique par K. G. Hilton

Résumé de l'article aux pages 512 à 519

Cet article traite des problèmes que comportent la construction d'une télécommande de position numérique. La position de l'axe est mesurée numériquement et les circuits de commande fonctionnent numériquement. L'auteur considère la disposition logique des circuits et discute les méthodes employées pour stabiliser la télé-commande de positions numériques.

Il décrit, enfin, un système rudimentaire dont la précision ne dépasse pas 45°, cette précision semblant suffisante pour indiquer le principe de système et les problèmes que pose son élaboration.

Le guide d'ondes, moyen de télécommunication par A. E. Karbowiak

Résumé de l'article aux pages 520 à 525

Un système de communication, utilisant le guide d'ondes comme moyen de transmission, est discuté dans cet article, qui analyse tous les aspects les plus importants du système.

L'auteur y étudie en détail les propriétés des divers types de guides d'ondes. Il démontre que si les guides d'ondes en métal ordinaire ne se prêtent pas aux communications à grande distance, on peut,

cependant, construire des guides d'ondes d'un type spécial (soit à revêtement diélectrique soit hélicoïdaux) répondant à des besoins pratiques raisonnables, ainsi qu'un système de communication d'une largeur de bande extrêmement étendue.

Systèmes logiques de calculateurs numériques à millimicrosecondes par N. F. Moody et R. G. Harrison

Résumé de l'article aux pages 526 à 529

Les auteurs décrivent un système logique à impulsions rapides qui allie l'efficacité d'étages couplés par transformateurs à une tolérance de retard numérique voisine de celle des systèmes couplés c.c. Ils indiquent des circuits logiques pour "OU", "ET", "INVERSEUR" *et* "REMINUTAGE", *ainsi qu'une commande permettant un facteur de "déploiement" de 5. Ces circuits sont entièrement transistorisés.*

Un excitateur à transistors pour usage physiologique par R. E. George

Résumé de l'article aux pages 530 à 533

Cet article décrit un excitateur servant à mesurer la période réfractaire qui suit l'excitation d'un nerf. L'appareil dont il s'agit produit des paires d'impulsions rectangulaires séparées par un intervalle de temps réglable. La durée des deux impulsions peut être modifiée indépendamment.

L'amplitude maximum disponible est de 20 volts, d'une impédance ne dépassant pas 2 kΩ. Les durées d'impulsions sont variables par paliers de 30 μsec à 30 msec. On peut faire varier le régime de répétition périodique d'une fois par 3 secondes à 330 fois par seconde.

Une méthode de réduction du retard des transducteurs à réponse exponentielle par L. Whitlow et M. J. Porter

Résumé de l'article aux pages 536 à 542

Cette article décrit une méthode pour réduire les retards des couples thermoélectriques ou d'autres transducteurs à réponse exponentielle simple. La tension de sortie d'un thermocouple est ajoutée à sa dérivée amplifiée, ce qui corrige la réponse de phase et d'amplitude. Dans la réduction de ce retard, le facteur limiteur est constitué par le rapport du signal d'entrée au bruit électrique produit dans l'appareil.

Les auteurs donnent également des détails de construction et de rendement d'un amplificateur c.c. à dérive corrigée et d'une alimentation à haute sensibilité.

Transistors et transformateurs à noyau saturable en tant qu'oscillateurs à ondes carrées par G. C. Fleming

Résumé de l'article aux pages 543 à 545

L'auteur décrit l'emploi de transistors comme commutateurs pour l'alimentation en continu de transformateurs à noyau saturable et il démontre qu'on peut construire par ce moyen des petits convertisseurs et inverseurs fort efficaces. Il examine la configuration de la base commune, de l'émetteur commun et du collecteur commun et indique des méthodes pour obtenir une sortie multiphases.

Les critères d'étude d'un système d'antenne ordinaire pour la transmission et la réception simultanées de signaux hyperfréquences par J. K. Grierson

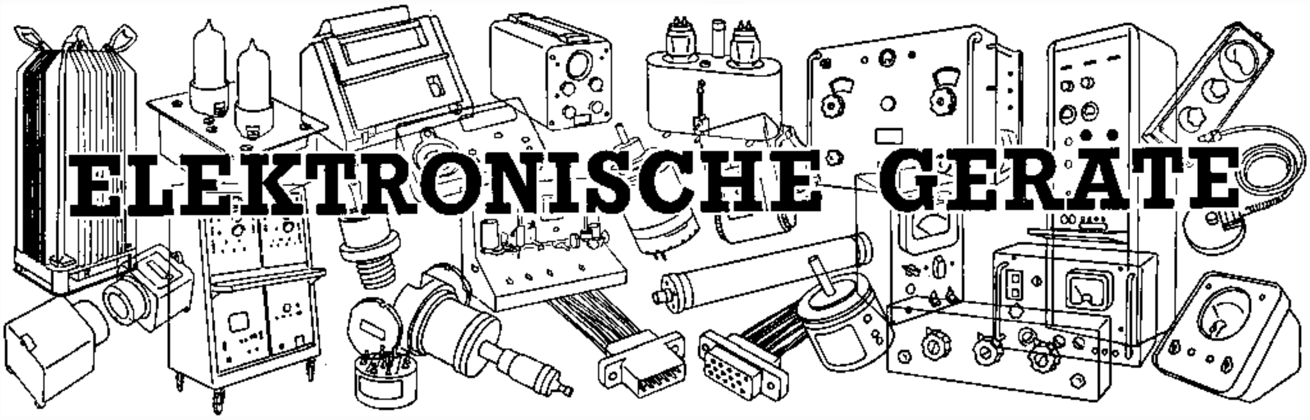
Résumé de l'article aux pages 546 à 549

Cet article décrit un système d'antenne commune à cinq sections, soit un émetteur, un réseau de couplage de réception et une antenne. Les réseaux de couplage de transmission et de réception permettent l'emploi d'une seule antenne pour la transmission et la réception de signaux à intervalles de fréquence, sans qu'il y ait nécessité de division de temps. L'auteur examine en détail les réseaux de couplage et indique les valeurs pratiques des composantes.

Rétroaction locale dans les amplificateurs à transistors par H. Pfyffer

Résumé de l'article aux pages 550 à 555

L'auteur discute d'abord les étages d'émetteurs amplificateurs courants. Il étudie ensuite les effets de la rétroaction négative sur le gain, l'impédance et la fréquence de coupure, puis il compare les calculs théoriques avec les résultats mesurés. Certaines de ces comparaisons sont expliquées graphiquement.



Beschreibungen neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben. *(Ubersetzung der Seiten 562 bis 565)*

VIBRATIONSMONITOR

(Abbildung Seite 562)

C. F. R. Giesler Ltd, Small Hall, River Place, Essex Road, Islington, London N.1

Dieses tragbare Gerät wurde zur Messung von Vibrationsamplituden an rotierenden Maschinen aller Art entwickelt. Der Frequenzbereich von 6 Hz—300 Hz entspricht einem Drehgeschwindigkeitsbereich von 360—18,000 U/min. Das Gerät besteht aus einem Verstärker, Amplitudenmesser und seismischen Umwandler. Ein- oder zweispurige Schreiber können angeschlossen werden. Zwei Eingangssockel sind für zwei Umwandler vorgesehen, die durch einen Umschalter wahlweise angeschlossen werden.

Amplitudenhöhen von 2,5 μm bis 2,5 mm können am Messgerät abgelesen werden. Die Instrumentempfindlichkeit bei Vollausschlag wird durch einen Empfindlichkeitsschalter geregelt. Das Messgerät hat zwei Skalen, die so bemessen sind, dass selbst eine Schwingungsamplitude von 6,5 μm einen angemessenen Ausschlag hervorruft.

EE 13 751 *für weitere Einzelheiten*

ÜBERGANGSFUNKTION-ANALYSATOR

(Abbildung Seite 562)

Servo Consultants Ltd, 17 Woodfield Road, London W.9

Dieses Gerät wurde als stabile und genaue Prüfmethode für Servoeinrichtungen und Regelorgane entwickelt und hat folgende Hauptvorteile:

(1) Das elektromechanische System des Gerätes ist genau und driftfrei.

(2) Das Gerät ist gegen harmonische Verzerrungen, die durch das zu prüfende System eingebracht werden unempfindlich und zieht nur die Grundfrequenz in Betracht.

(3) Phasenwinkel und Amplitude werden direkt angezeigt.

(4) Wechselstrom - Servoeinrichtungen können direkt geprüft werden.

(5) Die Leistung von Modulatoren, Gleichrichtern und magnetischen Verstärkern kann untersucht werden.

(6) Eine direkte Kupplung zu mechanischen und hydraulischen Reglern ist ohne Führungsglieder möglich.

(7) Das Gerät ist vollständig in sich abgeschlossen.

Zwei elektromechanische Modulatoren werden durch einen Regelmotor angetrieben, dessen Drehzahl der Testsignalfrequenz gleich ist. Die Motordrehzahl kann mittels eines Präzisionspotentiometers von Hand eingestellt werden.

Die Statorwicklung des Modulators 1 wird mit einem Trägerfrequenzsignal konstanter Amplitude gespeist. Die rotierende Ausgangswicklung des Stellgliedes gibt daher ein moduliertes Trägerfrequenzsignal ab, das im Falle einer Wechselstrom—Servoeinrichtung direkt als Testsignal Verwendung finden kann. Die dem Stellglied zugeführte Trägerfrequenz kommt entweder von einem eingebauten 2 kHz Geber oder von einer Fremdquelle. Für den Fall einer Gleichstrom-Servoeinrichtung wird dieses Signal gleichgerichet, um das Signal auf der Testfrequenz zu erhalten. Ausserdem kann zwischen dem Geber und dem zu untersuchenden System eine mechanische Kupplung eingeschaltet werden. Der Ausgang von dem zu prüfenden System kann wieder entweder eine modulierte Trägerfrequenz, wie z.B. die 400 Hz Abgabe eines Synchro-Motors, oder ein Signal auf der Testfrequenz sein. Solch ein Signal wird verstärkt und in das Stellglied 2 gespeist, das mit Stellgleid 1 synchron läuft. Im Falle eines Gleichstromsystems wird das Ausgangssignal von dem zu untersuchenden System dazu benutzt, den Träger zu modulieren, der dann dem Stellglied 2 zugeführt wird. Der Stator dieses Stellgliedes hat eine Dreiphasenverbindung zu einem weiteren Stellglied 3, das an eine Winkelanzeigeskala gekoppelt ist. Der Winkel zwischen dem Rotor und Stator dieses Stellgliedes entspricht der Phasendifferenz zwischen den Ausgangs- und Eingangssignalen des Gerätes, vorausgesetzt, dass das Signal von der Ausgangswicklung dieses Stellgliedes null ist. Das Stellglied hat eine weitere Wicklung, die rechtwinklig zur ersten angeordnet ist. Die Ausgangsspannung dieser Wicklung ist der Amplitude des Eingangssignals proportional. Die Grösse dieses Vektors kann direkt von einem Messgerät abgelesen werden.

Das Gerät ist in einem dauerhaften Leichtmetallgehäuse untergebracht. Die Frontplatte mit allen Bedienungsgriffen und Messgeräten hängt in Scharnieren, um alle wichtigen Teile des Gerätes leicht zugänglich zu machen. Zwei senkrechte Chassis mit allen elektronischen Kreisen sind von beiden Seiten zugänglich, ohne dass sie aus dem Gehäuse entfernt werden müssen. Der Geber und Getriebekasten für das Stellglied sind auf den Gehäuseboden montiert. Die abgehende Achse ist von rechts zugänglich.

Das Amplitudenanzeigegerät hat eine 102 mm Spiegelskala. Die Phasenanzeigeskala ist direkt in Winkelgraden des Phasenvorlaufs oder der Phasenverzögerung geeicht. Die Frequenzeinstellung erfolgt durch ein direkt geeichtes Potentiometer, das in einem Durchgang den ganzen Frequenzbereich durchläuft.

EE 13 752 *für weitere Einzelheiten*

DRUCKUMWANDLER

(Abbildung Seite 562)

The Solartron Electronic Group Ltd, Thames Ditton, Surrey

Ein Abkommen mit Svenska Flygmotor AB gibt der Solartron Electronic Group die Weltrechte ausserhalb Skandinaviens für Herstellung und Vertrieb eines als „Vibrierender Zylinder" bezeichneten Druckumwandlers, der einzigartig sein soll.

Die Fertigung hat in England bereits begonnen, und das englische Heeresversorgungsamt hat seinen Einbau in Fernwaffen und andere Geräte während der Tests genehmigt.

Für den kleinen, genauen und widerstandsfähigen „Vibrierenden Zylinder" Druckumwandler wurde ein ganz neues Konstruktionsprinzip angewandt. Die variable Frequenz des Ausgangssignals eignet sich besonders für Magnetbandaufzeichnung, z.B. mit dem Solartron Informations - Aufzeichnungsgerät, und auch Umwandlungen von Analogen zu

Digitalformen durch Impulszählen. Das die Umwandlersignale verarbeitende Gerät kann, ohne Genauigkeit einzubüssen, in beträchtlicher Entfernung vom Messpunkt aufgestellt werden.

Das druckempfindliche Element des Umwandlers besteht aus einem mechanisch vibrierenden System, dessen Eigenfrequenz vom angewandten Druck abhängig ist. Das vibrierende Element bestimmt die Frequenz eines elektronischen Rückkopplungsoszillators, wodurch eine Ausgangsspannung konstanter Amplitude erzeugt wird, deren Frequenz sich jedoch mit dem angewandten Druck ändert. Der Umwandler wurde für die Magnetbandaufzeichnung von im Fluge an Düsentriebwerken gemessenen Werten entwickelt.

Ein Umwandler mit variabler Ausgangsfrequenz ist für diesen Zweck besonders geeignet; sein frequenzvariables Ausgangssignal kann leicht in eine Form gebracht werden, in der es sich zur Bearbeitung in Digitalrechnern eignet.

Die Umwandler sind für Eingangsdrucke von 1,75 kg/cm², 3,5 kg/cm², 7 kg/cm² und 42 kg/cm² lieferbar.

EE 13 753 *für weitere Einzelheiten*

EMPFINDLICHES RÖHRENVOLTMETER

(*Abbildung Seite 563*)

Dawe Instruments Ltd, 99 Uxbridge Road, London W.5

Das Röhrenvoltmeter Modell 614 C ist ein kompaktes, tragbares Gerät zur Messung von 80 μV bis 300 V im Frequenzbereich 5 Hz bis 100 kHz. Wechselspannungen mit überlagertem Gleichstrom können damit gemessen werden, doch soll der summierte Spitzenpegel 1 kV nicht überschreiten.

Die Grundschaltung besteht aus einem Hochleistungsverstärker, der durch einen Abschwächer ein Anzeigegerät mit Vollweggleichrichter speist. Der Abschwächer besteht aus zwei Stufen: eine für den Eingang und die andere zur Trennung der Zwillingsverstärkerstufen. Ein Bereichschalter mit 13 Positionen schaltet beide Abschwächerstufen und gibt Vollausschlag bei 300 μV, 1 mV, 3 mV, 10 mV, 30 mV, 100 mV, 300 mV, 1 V, 3 V, 10 V, 30 V, 100 V und 300 V.. Die dem ersten Verstärkerteil folgende Abschwächerstufe wird eingeschaltet, um bei hohen Eingangspegeln den Geräuschabstand zu verbessern. Zur Erzielung hoher Eichgenauigkeit und Stabilität wird negative Gegenkopplung über beide Verstärkerstufen und ein stabilisierter Hochspannungs-Netzteil angewandt.

Das Messgerät ist in Effektivwerten für sinusförmige Eingaben geeicht, wobei die Anzeige den Mittelwert für die volle Wellenform darstellt. Die logarithmische Spannungsskala ermöglicht für alle Spannungen gleichförmige Ablesegenauigkeit, die für sinusförmige Eingaben innerhalb ±2% der Anzeige liegt. Zwischen 5 Hz und 100 kHz ist der Frequenzfehler geringer als ±3%. Ausser den 13 normalen Spannungsbereichen ist eine lineare dB Hilfsskala für den Bereich von −80 bis +50 dB (auf 1 V bezogen) vorgesehen.

Durch Einbau einer Ausgangsklinke kann entweder das Signal überwacht, oder das Gerät als Verstärker mit einem Verstärkungsfaktor von 88 dB verwandt werden. So eingesetzt, ist der Effektivwert der Ausgangsspannung bei Mindestbelastung von 10Ω für Vollausschlag ungefähr 8 V. Die Quellenimpedanz kann als 20Ω angenommen werden.

Bis zum 10 mV Bereich ist die Eingangsimpedanz ungefähr einem 10 MΩ Widerstand mit einer parallelgeschalteten Kapazität von 25 pF gleichwertig. Für die höheren Bereiche ist der Kapazitätswert 10 pF.

EE 13 754 *für weitere Einzelheiten*

FUNKEN-INTENSITÄTSANZEIGER

(*Abbildung Seite 563*)

Lexor Electronics Ltd, 25 Allesby Road, Coventry, Warwickshire

Das in dieses Gerät eingebaute empfindliche Röhrenvoltmeter kann Funkenintensität über einen grossen Bereich anzeigen und kann leicht für besondere Anwendung geeicht werden.

Typisch für seinen Einsatz ist die Verwendung für Entwicklung und Serienprüfung von Gleichstrommaschinen, Schaltgeräten, Kontaktgebern, Bauelementen u.s.w., sowie für Untersuchungen im Zusammenhang mit Radioentstörung. Zur Dauerüberwachung in Wartungs- und Betriebssicherheitsprojekten können Modelle zum Einbau in Anlagen geliefert werden.

Das Gerät zeigt Funkenbildung an, die normalerweise nicht sichtbar ist, wie z.B. auf der Unterseite von Bürstenarmen und in Geräten, die sich für Tests in Höhenunterdruckkammern befinden. Modelle für Dauerfunken- oder Impulsmessung sind lieferbar.

Bis zu 1 kV Gleichstrom wird der Dauerpegel durch ein Eingangsfilter von Funkenimpulsen getrennt. Die Endanzeige erfolgt an einem robusten 89 mm Drehspulmessgerät. Die Höchstempfindlichkeit für Vollanschlagist 50 mV.

EE 13 755 *für weitere Einzelheiten*

VIELZWECK-OSZILLOGRAF

(*Abbildung Seite 563*)

Marconi Instruments Ltd, St. Albans, Hertfordshire

Der TF 1330 ist ein hochwertiger Vielzweck-Oszillograf mit allen für direkte Beobachtung und Präzisionsmessung von Wellenformen erforderlichen Eigenschaften. Bei einem Frequenzgang von 0—15 MHz und einer in 7 Stufen von 50 mV/cm bis 50 V/cm veränderlichen Empfindlichkeit genügt ein Signal von 300 mV für die volle Austastung des 6 × 10 cm Schirmes. Eine in den Vertikalablenkteil eingebaute symmetrische Verzögerungskette mit punktförmiger Konstante ermöglicht eine verzerrungsfreie Beobachtung des ganzen zu untersuchenden Signals.

Die Zeitablenkung ist von 1 sec/cm—0,1 μs/cm in 15 Bereichen veränderlich und kann durch X Ausdehnung bis auf 0,02 μs/cm erhöht werden, wobei jeder Bereich leicht gegen eine Fremdquelle normalisiert werden kann. Triggerung erfolgt intern, extern oder durch eine eingebaute Netzfrequenzquelle. Die Auswahl verschiedener Gleich- und Wechselstromkopplungen ergibt eine stabile Triggerung für Signale sehr niedriger Frequenz bis zu mindestens 15 MHz und sieht Fernsehsynchronisierung vor.

Direkt geeichte Potentiometer werden zur Messung der Signalamplitude und Zeit verwandt. Die Eichung ist innerhalb ±2% genau und von der X Ausdehnung oder Y Verstärkung unabhängig.

Die 127 mm Elektronenstrahlröhre hat eine spiralenförmige Nachbeschleunigung und eine Betriebshochspannung von 10 kV. Zusammen geben diese Merkmale eine geradlinige Sichtanzeige grosser Helligkeit bei hoher Schreibgeschwindigkeit oder geringer Impuls-Wiederholungshäufigkeit. Statt der normalen Type wird eine Röhre mit längerer Nachleuchtzeit geliefert, um von der sehr niedrigen Ablenkgeschwindigkeit vollen Gebrauch machen zu können. Die Abschirmhaube für den Bildschirm kann abgenommen werden, um die Verwendung einer der üblichen Oszillografkameras zu erleichtern.

Funktionsmässig angeordnete, konzentrische Bedienungsknöpfe mit Farbkennzeichnung erleichtern den Gebrauch. Wenn die Spur in irgendwelcher Richtung aus dem Bildfeld herausgeht, wird der Druck auf einen Schalterknopf sofort die Position anzeigen. Es ist dann sehr einfach, die Spur mittels der beiden Ablenkregler in die Schirmmitte zu bringen. Das Gerät kann mit Hilfe eines Standes zur leichteren Beobachtung nach hinten gekippt werden.

Ein Zweispur-Modell TF 1331 ist auch lieferbar. Grundsätzlich ist die Leistung dieselbe wie die des TF 1330 mit zusätzlicher Elektronenstrahlschaltung zur Sichtanzeige zweier Signale durch wechselweise aufeinander folgende Ablenkung oder bei einer Schaltfrequenz von 100 kHz.

EE 13 756 *für weitere Einzelheiten*

SPULENBINDUNGSGERÄT

(*Abbildung Seite 564*)

Avo Ltd, Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

Ein Strom geeigneter Grösse, der für kurze Zeit durch eine Spule geht, die mit selbstbindendem Draht gewickelt ist, ist die schnellste Methode für deren Bindung. Die dafür benötigte Spannung mag ziemlich hoch sein, und es ist daher nötig, die Isolation zwischen den Drähten und gegenüber dem Dorn, auf dem die Spule gewickelt wurde, zu prüfen.

Mittels des Avo Bindungsgerätes werden die notwendigen Arbeiten folgendermassen durchgeführt:

Beide Spulenenden werden an die Klemmen des Gerätes angelegt. Dann werden Schalter mit beiden Händen gleichzeitig bedient und der Isolationswert zwischen Spule und Dorn direkt in Megohm abgelesen. Bei zufriedenstellendem Isolationswert wird durch Bedienung eines dritten Schalters der Heizstrom für eine vorgewählte Zeit eingeschaltet.

Da das Bedienungspersonal jeweils beide Hände für den Test benutzen muss, ist das Gerät vollkommen betriebssicher. Die Zeitsteuerung für den Bindestrom ist zwischen 3/4 sec und 2½ min veränderlich; Isolationsmessungen bis zu 500 MΩ können ausgeführt werden.

Das Gerät ist für 210—250 V 50 Hz Netzpannungen gebaut. Wenn diese Spannungen jedoch für den Bindevorgang als zu hoch betrachtet werden, können dem Gerät niedrigere Spannungen durch Klemmen zugeführt werden.

EE 13 757 *für weitere Einzelheiten*

EINGEBETTETE PAPIERKONDENSATOREN

Dubilier Condenser Co. (1925) Ltd, Ducon Works, Victoria Road, North Acton, London W.3

Diese neuen „Dubilier"-Papierwickelkondensatoren Type 560 wurden entwickelt, um der Nachfrage für Hochleistungs-Papierwickelkondensatoren mit aussergewöhnlich guten elektrischen Eigenschaften und langer Lebensdauer gerecht zu werden. Die Betriebstemperatur für diese Kondensatoren kann ohne Herabsetzung der Betriebsspannung bei Gleichstrom zwischen −40° C und +125° C und bei Wechselstrom zwischen −40° C und +70° C liegen.

Für den Aufbau wird ein herkömmlicher Papier- und Foliewickel mit Anschlusstreifen zwischen den Elektroden und Drahtanschlüssen versehen und mit einem Hochtemperatur-Kunststoff imprägniert, wodurch ein festes Kondensatorelement entsteht. Die Drahtanschlüsse werden in die Anschlusstreifen eingelegt, zusammengedrückt und verlötet, so dass widerstandsfähige elektrische und mechanische Verbindungen entstehen. Das zusammengebaute Element wird dann in ein Epoxydharz mit Mineraleinschlüssen eingebettet, das einen ausgezeichneten Widerstand gegen Stoss und Feuchtigkeitseintritt hat.

Bei Betriebsspannungen von 180 V bis 250 V Wechselstrom oder 350 V bis 1000 V Gleichstrom liegt die Wertreihe zwischen 0,001 μF und 0,1 μF.

EE 13 758 *für weitere Einzelheiten*

KOMBINIERTER WELLENSCHALTER UND ABSTIMMKONDENSATOR

(Abbildung Seite 564)

The Plessey Co. Ltd, Vicarage Lane, Ilford, Essex

Die Plessey Company Ltd. gibt die Einführung eines Drehkondensators mit Schalter bekannt, der für jeden Zweibandempfänger Einknopfbedienung der Wellenumschaltung und Abstimmung ermöglicht.

Die neuen Kondensatoren verwenden eine normale Rotorachse. Wenn die Platten fast vollständig eingedreht sind, betätigt diese Achse einen Plessey Schiebeschalter durch eine Gelenksteuerung in der einen Richtung, und wenn die Platten fast völlig ausgedreht sind, in der anderen Richtung. Es wird empfohlen, die dabei auftretende unausgenützte Kapazität am HF Ende des Langwellenbandes und am NF Ende des Mittelwellenbandes anzuordnen. Auf diese Weise wird der Verlust an einstellbaren Sendern unwichtig.

Der Schalter ist als zweipoliger Umschalter ausgebildet, kann aber für Skalenlampenschaltung und ähnliches mit einem dritten Pol versehen werden; auch Anordnungen für mechanische Bandanzeige sind möglich.

Plessey weist darauf hin, dass die vollständige Einknopfabstimmung eines Zweiband-Rundfunkempfängers nicht nur einen Verkaufsschlager darstellt, sondern dass die Verwendung der neuen Kondensatoren auch Schaltungsvereinfachungen zulässt, die von wirtschaftlicher Bedeutung sind.

EE 13 759 *für weitere Einzelheiten*

NORMALPEGEL FÜR PEGELSENDER

(Abbildung Seite 564)

Radiometer, 72 Emdrupvej, Copenhagen NV, Denmark.

Der Normalpegel SNNI für Pegelsender ist ein Laborgerät zur genauen Eichung von Pegelmessern und sorgfältigen Bestimmung ihres Frequenzganges.

Mit einem geeigneten Messender zusammengeschaltet, gibt das Modell SNNI entweder in einen offenen Stromkreis oder einen Belastungswiderstand von 75Ω folgende drei Ausgangspegel ab:

0 dB/0,274 V = 274 mV
−9 dB/0,775 V = 275 mV
−1 N/0,775 V = 285 mV

Die Pegelgenauigkeit ist bei einer Belastung mit 75 Ω im Bereich von 0-25 MHz besser als 1% und bei Abgabe in einen offenen Kreis bis zu 15 MHz ebenso besser als 1%. Der Frequenzgang von 10 Hz-10 MHz is ± 0,2%.

Das Gerät arbeitet nach dem Prinzip der gesättigten Diode und hat daher eine hohe Empfindlichkeit, 0,1% Genauigkeit bei Anzeigeänderungen und eine zuverlässige Effektivwertanzeige. Ein weiteres Merkmal ist der zuverlässige Überlastungsschutz.

Das Modell SNNI wurde für Trägerfrequenz-Nachrichtenaniagen entwickelt, kann aber auch auf vielen anderen Gebieten Andwendung finden.

EE 13 760 *für weitere Einzelheiten*

SPEKTRUMANALYSATOR

(Abbildung Seite 565)

James Scott & Co. (Electrical Engineers) Ltd, 68 Brockville Street, Carntyne Industrial Estate, Glasgow, E.2.

Der Spektrumanalysator Type 190 findet für die Analyse einfacher oder zusammengesetzter Signale im NF und Überschallbereich Verwendung. Der Frequenz-bereich des Gerätes in Normalausführung geht von 500 Hz bis 90 kHz; Geräte mit anderen Bereichen können im Sonderauftrag hergestellt werden.

Zwei wahlweise Methoden sind für die Spektrumdarstellung vorgesehen. Zur schnellen qualitativen Analyse wird das Spektrum auf einem Bildschirm wiedergegeben. Die X-Koordinate der Spur stellt die Frequenzskala dar und die Y-Koordinate den Leistungspegel der jeweiligen Frequenz. Der Spur können zur genauen Eichung Eichmarken aufgegeben werden, die von einem manuell einstellbaren Eichmarkengeber herrühren.

Zur genauen Analyse der Leistungsverteilung in irgendeinem Teil des Spektrums wird die Eichmarke auf die entsprechende Stelle eingestellt und der Analysator auf „Handbedienung" umgestellt. Die Handeinstellung des Markengebers bestimmt nunmehr die Mittelfrequenz, bei der die Messung mittels eines thermischen Milliamperemeters ausgeführt wird. Der Signalpegel wird mittels eines genauen Eichsenders festgestellt, wobei immer eine konstante Messanzeige durch Einreglung des Dämpfungsgliedes erreicht wird.

Der automatische Kippgenerator kann umgestellt werden, um entweder den Bildschirm von 1-7 kHz oder 1-100 kHz mit einer vom Benutzer bestimmbaren Geschwindigkeit zu bestreichen.

Die Leistung eines Teiles des Spektrums kann nur unter Bezugnahme auf die Bandbreite angegeben werden, in der sie gemessen wurde, und die für dieses Gerät entweder 1 kHz oder 70 Hz ist. Der Frequenzgang des Engbandfilters ist infolge der Vedwendung von sechs gestaffelten Quarzelementen eine gute Annäherung an die ideale „Hochebene".

Das Gerät kann nicht nur als Spektrumanalysator bis zu 90 kHz benutzt werden; mittels eines geeigneten Überlagerungsempfängers, dessen ZF-Bandbreite mindestens 100 kHz ist, kann es auch bei höheren Frequenzen eingesetzt werden

EE 13 761 *für weitere Einzelheiten*

ZENER-SPANNUNGSREGLER

Texas Instruments Ltd, Dallas Road, Bedford, Bedfordshire

Eine neue Silizium-Zener-Spannungsreglerserie, Typen 1 S 501 bis 1 S 516 und neue Silizium-Zener-Bezugsdioden wurden von Texas Instruments angekündigt.

Typen 1 S 501—1 S 516 sind mit Stehbolzen versehen und in 16 Stufen über den Spannungsbereich von 22—91 V verteilt. Die neuen Regler haben 8 W Verlustleistung und Zener-Spannungstoleranzen von 5% und 10%. Mit Anode am Stehbolzen oder doppelter Anodenfederklemme lieferbar.

Die 1 S 207—1 S 218 Reihe der Zener-Bezugsspannungsdioden liegt in 12 Stufen im Bereich von 3,6 bis 10 V. Sie sind mit Zener-Spannungstoleranzen von 5% und 10% lieferbar und luftdicht in Molybdänglas-Muffen eingeschmolzen, wie das auch für die Texas 400 mA Silizium-Gleichrichter üblich ist.

Diese Halbleiterelemente sind in Prototypmengen lieferbar und Massenfertigung läuft jetzt in Bedford an.

EE 13 762 *für weitere Einzelheiten*

KONDENSATOREN FÜR HOHE TEMPERATUREN

Johnson, Matthey & Co. Ltd. 73-83 Hatton Garden, London E.C.1

Johnson, Matthey and Co. geben bekannt, dass sie für Verteidigungszwecke eine Serie neuer versilberter Glimmerkondensatoren entwickelt haben.

Die neuen Kondensatoren können im Temperaturbereich von $-70°C$ bis $+250°C$ Verwendung finden und werden durch Tauchen mit einer besonderen Silikonkautschuk-Umhüllung versehen. Durch Zusammenschmelzen der einzelnen versilberten Glimmerblättchen ensteht ein fester Block.

Herstellung erfolgt in zwei Grössen und für jeden Kapazitätswert im Bereich von 5 pF bis 0,05 μF mit einer Mindesttoleranz von $\pm\frac{1}{2}\%$ oder 1 pF, wobei der höhere Wert gültig ist. Jede Grösse wird für 200 oder 350 V Spitzenspannung bei Temperaturen bis zu 100°C geliefert. Bei 250°C geht der Nennwert der Spitzenspannung auf 100 und 200 V zurück.

EE 13 763 *für weitere Einzelheiten*

PROGRAMMUHR

(Abbildung Seite 565)

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

Dieses neue „Programmaufzeichnungs- und Wiedergabegerät" wurde für Folge- und Zeitsteuerung industrieller Verfahren entwickelt und kann bis zu 49 „ein—aus" Schaltungen in Einzel- oder Gemeinschaftsschaltungen ausführen. Es kann zur Steuerung aller Verfahren eingesetzt werden, in denen die Operationsfolge durch zeitgesteuerte Bedienung einer Anzahl von Kontakten reguliert werden kann.

Durch die Anpassungsfähigkeit des Systems kann die Zeitsteuerung eines beliebigen Abschnittes des Verfahrens wunschgemäss geändert werden, und die Vorgangsfolge kann ohne grössere Verdrahtungsänderung umgestellt werden. Die Steuerimpulse werden auf normalen Magnettonbändern gespeichert. Eine unbeschränkte Anzahl Kopien des Originalbandes kann hergestellt werden, um entweder eine Reihe von Maschinen unabhängig voneinander zu steuern, oder Kopien an Filialen oder ausländische Tochtergesellschaften für deren Gebrauch weiterzuleiten. Dadurch kann die Geheimhaltung einzigartiger Prozesse gewährleistet werden.

In Normalausführung kann die Anlage bis zu $2\frac{1}{2}$ Stunden hintereinander arbeiten. Sonderausführungen für längere Zeitspannen sind lieferbar.

EE 13 764 *für weitere Einzelheiten*

ELEKTROMAGNETISCHE ZÄHLER

(Abbildung Seite 565)

Lancashire Dynamo Electronic Products Ltd, Rugeley, Staffordshire

Lancashire Dynamo Electronic Products Ltd. sind die Alleinvertriebsrechte der „Hengstler" elektromagnetischen Zähler Typen F 43 und F 43 M2 für England übertragen worden.

Sie können bis 2400 Impulse/min zählen und haben eine hohe Lebenserwartung. Die Hengstler F 43 und F 43 M2 Bausteine haben dieselben Konstruktionsgrundzüge und geben beide eine sechsstellige Anzeige: Typ F 43 hat Handrückstellung und Typ F 43 M2 hat sowohl manuelle, als auch elektrische Nullstellung. Beide Typen werden in schneller und langsamer Ausführung und für eine Reihe bevorzugter Gleichstromspannungen geliefert. In Kürze werden auch Gleichrichter für gewisse bevorzugte Wechselstromspannungen zur Verfügung stehen.

Aus der Abbildung ist ersichtlich, dass jeder Zähler aus zwei Teilen besteht: dem eigentlichen Zähler und einem Sockelgehäuse, in das er eingeschoben wird, wobei die elektrischen Verbindungen durch einen hinten liegenden Stecker und Sockel hergestellt werden. Das Sockelgehäuse ist ein Spritzgussteil, der mit anliegenden Gehäusen an allen vier Seiten mechanisch verkuppelt werden kann, so dass lange Zählerreihen, wenn nötig, in einen einzigen Baustein zusammenmontiert werden können. Alle Eingangsverbindungen werden an der Rückseite des Sockelgehäuses hergestellt.

Lancashire Dynamo Electronic Products Ltd. möchten betonen, dass die oben erwähnten Abmachungen sich nur auf den Vertrieb der Hengstler Schnellzähler Typen F 43 und F 43 M2 beziehen. Andere mechanische und langsame, von J. Hengstler KG hergestellte Zähler werden in England weiter durch deren Agenten Brian R. Morris and Co. Ltd. vertrieben.

EE 13 765 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Eine Digitale Ferneinstellungsregelung von K. G. Hilton

Zusammenfassung des Beitrages auf Seite 512-519

Der Beitrag beschreibt eine Untersuchung der bei der Konstruktion einer digitalen Ferneinstellungsregelung auftretenden Aufgaben. Die Welleneinstellung wird digital gemessen und die Regelstrecke arbeitet digital. Der Bericht beschäftigt sich hauptsächlich mit der logischen Schaltungsanordnung und bespricht die Stabilisierungsmethoden für eine digitale Ferneinstellungsregelung.

Die bei der Entwicklung zu beachtenden Grundsätze und Aufgaben werden mit Hilfe eines Grobeinstellungsverfahrens behandelt, dessen Genauigkeit von nur 450 für diesen Zweck als ausreichend betrachtet wird.

Hohlleiter: Ein Fernverkehrsmittel von A. E. Karbowiak

Zusammenfassung des Beitrages auf Seite 520-525

Die Verwendung von Hohlleitern als Übertragungsmittel im Fernverkehr wird behandelt.

Die wichtigsten Gesichtspunkte eines derartigen Nachrichtensystems werden analysiert und die Eigenschaften der verschiedenen Hohlleitertypen eingehend diskutiert. Trotzdem einfache Metallhohlleiter für den Fernverkehr ungeeignet sind, können Hohlleiter in Spezialausführung (entweder mit dielektrischer Beschichtung oder Wendelhohlleiter geeigneter Gestaltung) konstruiert werden, die vernünftigen praktischen Forderungen entsprechen. Auf diese Weise kann ein Nachrichtensystem äusserst grosser Bandbreite konstruiert werden.

Nanosekunden Digitalrechnerlogik von N. F. Moody und R. G. Harrison

Zusammenfassung des Beitrages auf Seite 526-529

Die beschriebene schnelle Impulslogik verknüpft den Wirkungsgrad transformergekoppelter Stufen mit einer Digitverzögerungstoleranz, die der gleichstromgekoppelter Stufen nahekommt.

"Oder"- und "Und"-Gatter, ein Umkehrer und die Wiederherstellung der Impulssynchronisierung werden zusammen mit einem Treiber behandelt, der gleichzeitig fünf Bausteine speisen kann. Transistoren finden für alle Schaltungen Verwendung.

Ein Transistor-Reizstromgerät für physiologische Untersuchungen von R. E. George

Zusammenfassung des Beitrages auf Seite 530-533

Ein Reizstromgerät zur Messung der refraktionären Pause, die der Nervenreizung folgt, wird beschrieben. Das Gerät gibt zwei Rechtecksimpulse ab, die durch ein einstellbares Intervall getrennt sind. Die Dauer beider Impulse kann unabhängig eingestellt werden.

Von einem Widerstand unter 2 kΩ können Amplituden bis zu 20 V abgegeben werden. Die Impulsdauer ist in Stufen von 3 μs . . . 30 ms einstellbar. Die Wiederholungsfrequenz ist von einmal in 3 s . . . 330/s variabel.

Methode zur Verzögerungsherabsetzung in Umformern mit Exponentialgang von L. Whitlow und M. J. Porter

Zusammenfassung des Beitrages auf Seite 536-542

Eine Methode zur Verzögerungsherabsetzung in Thermoelementen und anderen Umformern mit einfachem Exponentialgang wird behandelt. Zur Korrektur des Phasen- und Amplitudenganges wird die Ausgangsspannung des Thermoelementes ihrem verstärkten Differentialquotienten zuaddiert. Der Grenzfaktor für die Verzögerungsherabsetzung ist das Verhältnis des Eingangssignales zum elektrischen Rauschen im Gerät.

Konstruktionseinzelheiten und Leistung eines Gleichstromverstärkers mit Driftkorrektion und eines hochstabilen Netzgerätes werden auch angegeben.

Transistoren und Ubertrager, deren Kerne gesättigt werden können, als Rechteckgeneratoren von G. C. Fleming

Zusammenfassung des Beitrages auf Seite 543-545

Der Beitrag beschreibt die Verwendung von Transistoren als Schalter in der Gleichstromleitung zu Ubertragern, deren Kerne gesättigt werden können. Es wird gezeigt, dass dadurch kleine und leistungsfähige Umwandler und Zerhacker entwickelt werden können. Basis-, Emitter- und Kollektorschaltungen werden behandelt und Methoden zur Erzielung mehrphasiger Ausgänge beschrieben.

Entwicklungsmerkmale für eine Gemeinschaftsantennenanlage zum gleichzeitigen Senden und Empfangen von UKW Signalen von J. K. Grierson

Zusammenfassung des Beitrages auf Seite 546-549

Die beschriebene Gemeinschaftsantennenanlage besteht aus fünf Teilen: Sender, Senderankopplung, Empfänger, Empfängerankopplung und Antenne. Simultanbetrieb für Signale mit Frequenzabstand ist durch die Sender- und Empfängerankopplungskreise mit einer einzigen Antenne und ohne Antennenweiche möglich. Die Ankopplungskreise werden eingehend behandelt und Werte für die Bauelemente angegeben.

Gegenkopplung innerhalb einer Transistor-Verstärkerstufe von H. Pfyffer

Zusammenfassung des Beitrages auf Seite 550-555

In diesem Beitrag wird die Verstärkerstufe in Emitterschaltung behandelt. Der Einfluss der negativen Gegenkopplung auf Verstärkung, Impedanz und Grenzfrequenz werden berechnet und die theoretischen Berechnungen mit gemessenen Werten verglichen. Einige dieser Vergleiche sind illustriert.