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Commentary

THE need for more scientists and technologists in this country has been recognized for some time past and much thought has been given to ways and means of overcoming this shortage.

In the immediate post-war years this shortage was not particularly acute and it had certainly not become a national problem of the magnitude it is today. There were large numbers of men and women returning from the Services, many of whom had received sufficient scientific training during the war to enable them to take their place in industry and research with confidence.

These, with the considerable number of trained scientists and technologists in the various Government research establishments and elsewhere formed a large reserve of scientifically trained manpower which enabled this country to return to a peace time basis without too much difficulty.

Meanwhile the Universities and Technical Colleges were reorganized and expanded to deal with those men and women whose studies were interrupted by the War.

However, development in this scientific age has been so unexpectedly rapid in the post war years that this reserve of specialist manpower has been completely absorbed and a limit to future progress is now being set by the rate at which the scientists and technologists of the future are forthcoming. If this country is to remain in the forefront it will need many more potential scientists and technologists than are undergoing training at the moment and we must stimulate the interest of more boys and girls now at school in pursuing careers in science and technology.

But the necessary numbers of potential scientists and technologists of the requisite calibre cannot be created overnight and having created the interest of the schoolboy and schoolgirl it is imperative that this interest be maintained throughout and that the teaching and training facilities are of the best.

Much has already been done by the Ministry of Education and the local education authorities to raise the standard of science teaching in the State aided schools throughout the country although it must be acknowledged that a very great deal remains to be done.

The independent and direct grant schools have of course been excluded from such assistance and it is now encouraging to learn that their lack of suitable facilities in this respect is to be remedied by a grant of £3m from the Industrial Fund for the Advancement of Scientific Education in Schools.

This fund was set up in November 1955 by a small group of firms engaged principally in the chemical, electrical and mechanical engineering industries with the object of improving the quality and increasing the number of scientists and technologists by helping to provide better accommodation and equipment at school level. An executive committee was appointed to administer the fund and a preliminary study brought to light the fact that the Universities and Technical Colleges as now expanded would be able to offer more places to potential scientists and technologists than the schools would be able to fill.

Since the State aided schools were the responsibility of the Ministry of Education and the local education authorities it was decided that the Fund should be "directed only, and purposely, to the independent and direct grant schools which cannot receive the assistance from public funds for capital works".

A report of the committee under the chairmanship of Sir Hugh Beaver has now been published and it records that a hundred and forty companies have subscribed just over three million pounds most of which money has now been committed in the form of grants for building and equipment.

Before these grants were made, visits were paid to some 503 independent and direct grant schools who appealed for assistance and it was revealed that the science facilities were woefully inadequate. Although the number of boys reading science was three times as great as in 1935 they were being taught in accommodation which had hardly been increased at all. Grants to 187 schools have now been approved for building and it is anticipated that 92 of the buildings will be completed by the end of this year while grants for scientific equipment have been approved for 328 schools and the Fund hopes that this expenditure will make it possible to expand the amount of science accommodation by one half and raise the number of boys and girls in the science and mathematics sixth forms from the present level of some twelve thousand to seventeen thousand when the buildings are completed.

The operation of this Fund has without question brought much encouragement to both schools and science masters and its effect will be felt far outside the schools dealt with by the Fund. It will ultimately be of benefit to industry, which is what the Fund set out to do, and it may, which is of great importance, encourage science graduates to take up teaching as a result of the improved facilities.

THE INDUCTOSYN

and its Application to a Programmed Co-ordinate Table

(Part 1)

By L. H. R. Harrison*, B. A. Horlock*, A.M.Brit.I.R.E., Grad.I.E.E., and F. D. Hunt*, A.M.Brit.I.R.E.

The Inductosyn is a new control element capable of indicating angular position to an accuracy of 5sec of arc with a repeatability of 1sec in its rotary form, and capable of a positional accuracy better than 0·0001in with a repeatability of 25μin in its linear form.

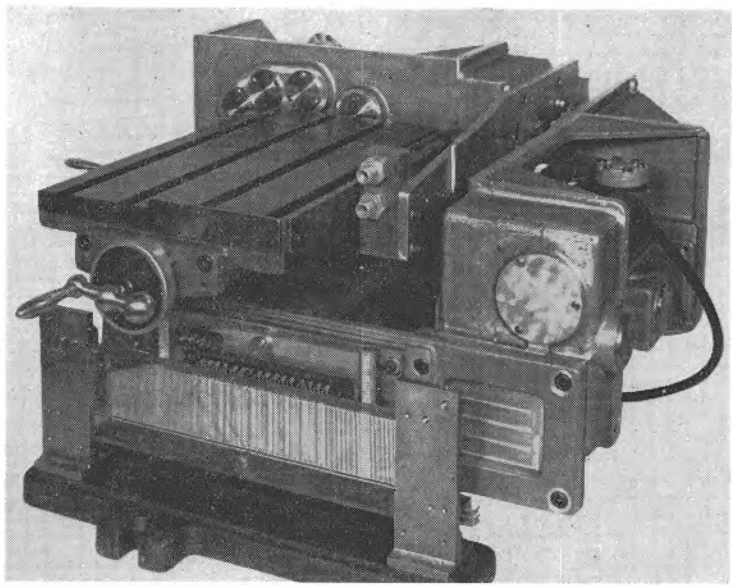
A description is given of the application of the linear Inductosyn to position control with particular reference to a programmed co-ordinate table. This table, measuring 10in by 8in, may be attached to any machine tool and positioned in the one plane to an accuracy of ±0·0002in.

The co-ordinate information is in the form of punched cards and any number of x, y co-ordinates can be specified. The program unit is described in some detail.

THE development of the rotary Inductosyn was undertaken for the American Air Force by the Farrand Optical Co., Inc., New York, and as a result of four years development the present form, with an accuracy of 5sec of arc, repeatability of 1sec of arc and sensitivity of 0·25sec of arc, has been accomplished.

In the rotary form it is used primarily for high accuracy angular measurement and may be likened to a resolver in that the quantity that varies as a function of mechanical motion is the electromagnetic coupling between two conductor formations. In a resolver these conductor formations are coils of wire wound on suitable iron cores and one coil is arranged to be rotated with respect to the other windings and provides a unique value of magnetic coupling for each angular position. In its simplest form there are only two poles, hence for one mechanical revolution the coupling varies over one complete cycle, conforming to a sine law. In the Inductosyn the conductors are formed by printing hairpin type windings on glass plates, and the magnetic coupling between two plates is through a small air-gap. A large number of poles are provided and in the case where there are 180 poles then one cycle is completed for every 4° of angular motion. This therefore performs the same function as a standard resolver which has been geared up 90 to 1. When used as an analogue-digital convertor and binary numbers are used it is convenient to use an Inductosyn with 128 poles resulting in 64 cycles per revolution, but for decimal notation 100 or 200 poles would be most suitable.

There are many uses to which this rotary form may be put, for example, an electronic servo controlled dividing head, an electronic gear ratio free from backlash, precision theodolite, etc. As an element for use in controlling linear motion of machine tools there is a disadvantage in that the rotary Inductosyn would be mounted on the lead-screw and, while the angular position would be defined accurately, the linear position of the machine table would be dependent on the accuracy of the lead-screw. It was to overcome this drawback that the Farrand Company developed the linear Inductosyn as a Company project.



Experimental installation of linear Inductosyn for position control

A scale corresponding to the rotor of the rotary form consists of a single hairpin winding, and a slider analogous to the stator has two space phase windings spaced 90 electrical degrees apart. These are illustrated in Fig. 1. Pole spacing represents one-tenth inches and this dimension represents 360 electrical degrees. There are 32 pairs of poles and the outputs from all pairs of poles are averaged so that any random error in pole spacing is reduced. It is possible to control positioning of the Inductosyn to an electrical angle of one milli-radian which gives the linear Inductosyn an accuracy of approximately 16μin.

The scale and slider windings are printed on glass blocks; the dimensions of the blocks are approximately 10in by 2½in by ⅜in and 5in by 3in by ⅜in respectively. Ruling of the masters used in the printing process has been accomplished using an extremely accurate dividing engine, capable of ruling 14 000lines/in to give the necessary accuracy.

In use it is often required to control movements greater than 10in and these dimensions can be readily accommodated by placing several scales end to end and connecting them electrically. The scales are made a few mils short of 10in length but there is no discontinuity at the gap between adjacent ends of the scales because the gap occurs in a space between the end bars. The scales can be very precisely phased, one with respect to the other, and of course the averaging effect minimizes any minor error.

* *The Plessey Company Ltd.*

Many machine tools comprise of a work table moving linearly past a fixed part of the machine, hence it is convenient to affix the scales to the table, and the slider to the fixed part of the machine. The air-gap between scales and slider is not very critical but it is obvious that the output from the scale is dependent upon the spacing (Fig. 2). A spacing of 0·005in is recommended and this should be constant over the whole length of travel.

Digital to Analogue Conversion

The shaft position of a resolver can be uniquely defined by impressing a voltage V_1 on one winding proportional

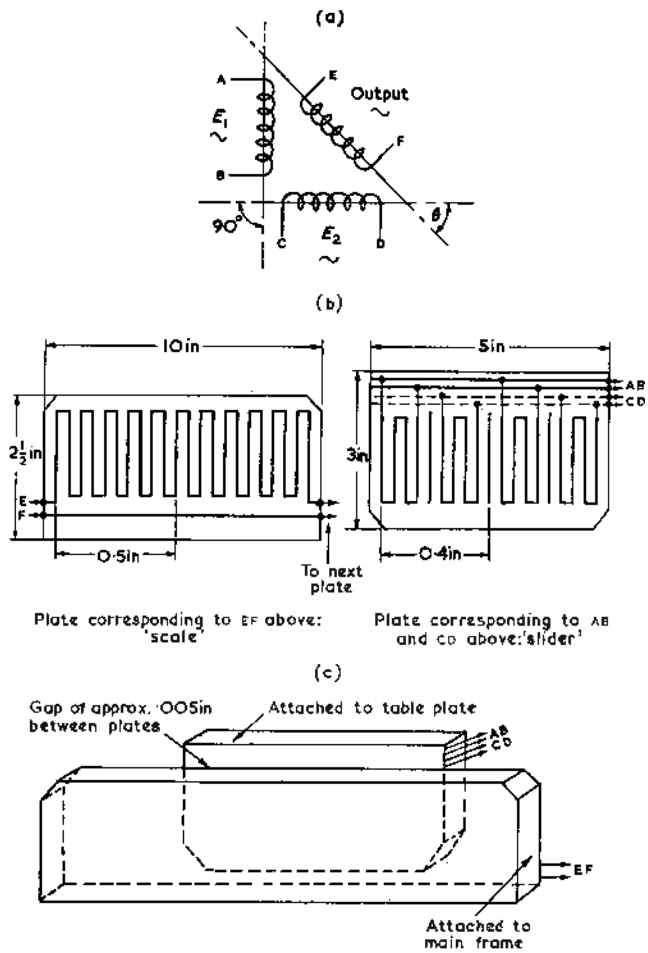


Fig. 1. The linear Inductosyn

to sin θ and a voltage V_2 on the quadrature winding proportional to cos θ, when its position will be at angle θ. Such a representation of position may be used with the linear Inductosyn. The 360 electrical degrees representing 0·1in is divided into 1 000 parts, i.e. each part of 0·36 electrical degrees is equivalent to one ten thousandth inch. Fig. 3 shows the arrangement for 3 digit representation in decimal form. Voltages are added trigonometrically and are impressed on the two windings of the slider as in a resolver, that is $V_1 \equiv \sin\theta$, $V_2 \equiv \cos\theta$.

The addition conforms to:

$$V_1 \equiv \sin(a+b+c) = \sin c\,(\cos a \cos b - \sin a \sin b) + \cos c\,(\sin a \cos b + \cos a \sin b) \quad \ldots\ldots \quad (1)$$

$$V_2 \equiv \cos(a+b+c) = \cos c\,(\cos a \cos b - \sin a \sin b) - \sin c\,(\sin a \cos b + \cos a \sin b) \quad \ldots\ldots \quad (2)$$

In equation (1) and considering the decimal number 111 where each unit is 0·36°, each ten is 3·6°, and each hundred is 36°, then assuming that 1V is equivalent to sin 90° or cos 0°.

$$V_1 \equiv \sin 39{\cdot}96 = \sin(36+3{\cdot}6+0{\cdot}36)$$
$$= \sin 0{\cdot}36\,(\cos 36 \cos 3{\cdot}6 - \sin 36 \sin 3{\cdot}6) + \cos 0{\cdot}36\,(\sin 36 \cos 3{\cdot}6 + \cos 36 \sin 3{\cdot}6)$$
$$= 0{\cdot}6413\text{V}$$

Similarly

$$V_2 \equiv \cos 39{\cdot}96 = \cos(36+3{\cdot}6+0{\cdot}36) = 0{\cdot}7665\text{V}$$

For very small angles the sine is equal to the angle and the cosine remains very nearly unity. Consequently for the least significant figure the cosine transformer does not require to be tapped and the sine transformer tapping may be linear.

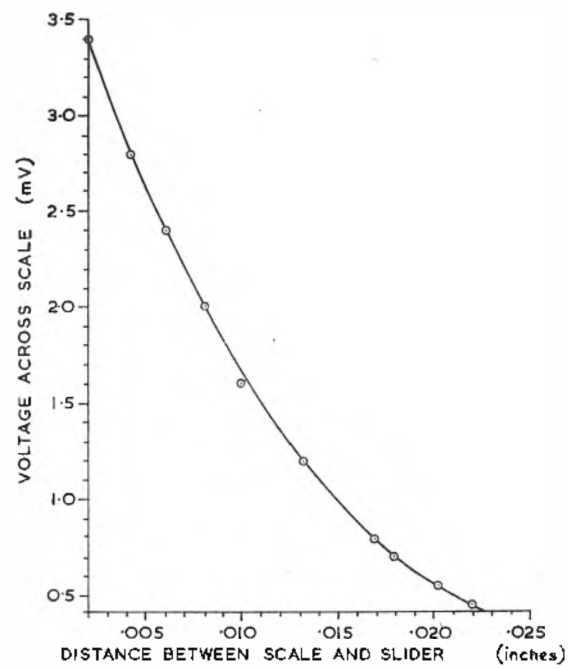


Fig. 2. Relationship between scale and slider spacing and output voltage

As the inputs to the windings of the Inductosyn are required to be in the ratio of the sine and cosine of the represented angle the tangents function can conveniently be used for the small angles. Fig. 4 shows the arrangement using the small angle conventions, and this shows that for the small angle transformers only one secondary winding need be tapped.

When the dimension is such that the represented angle exceeds 90° then centre-tapped transformers must be used in order to allow for the changing sign of the sine and cosine functions in all four quadrants. This is shown in both Figs. 3 and 4.

It is possible to use a resolver to provide the inputs to the Inductosyn slider, the rotor is energized with a 10kc/s signal and the outputs from the stator windings are used to provide the sine and cosine voltages to the Inductosyn. One complete revolution of the resolver shaft is then equivalent to a movement of 0·1in of the Inductosyn. When the dial on the resolver shaft is divided into 100 parts each division corresponds to a movement of 0·001in, provision of a vernier scale together with this scale permits settings of 0·0001in to be made.

The construction of the transformers may take many

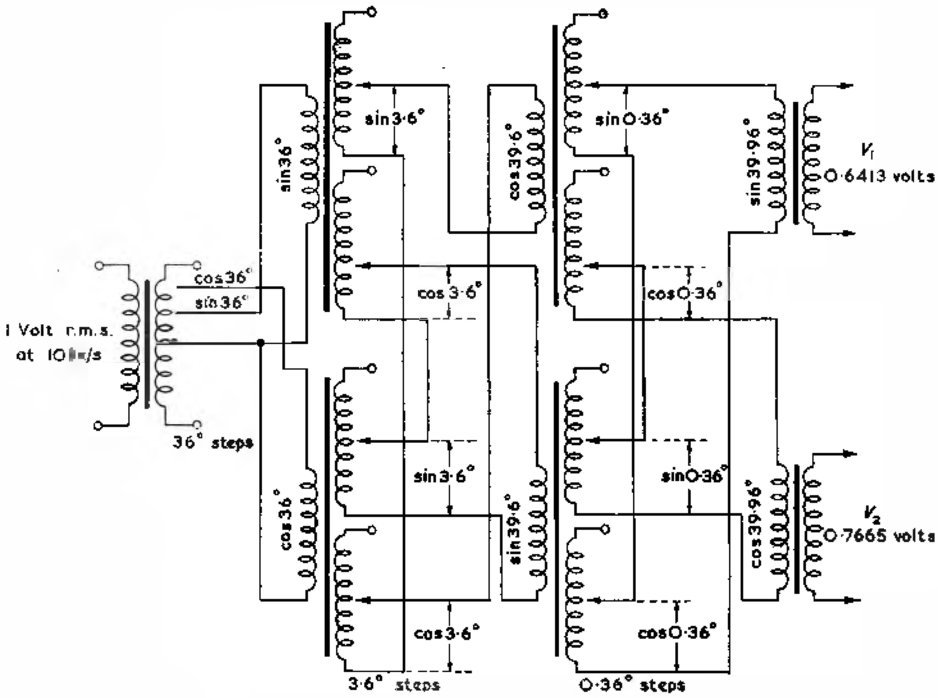
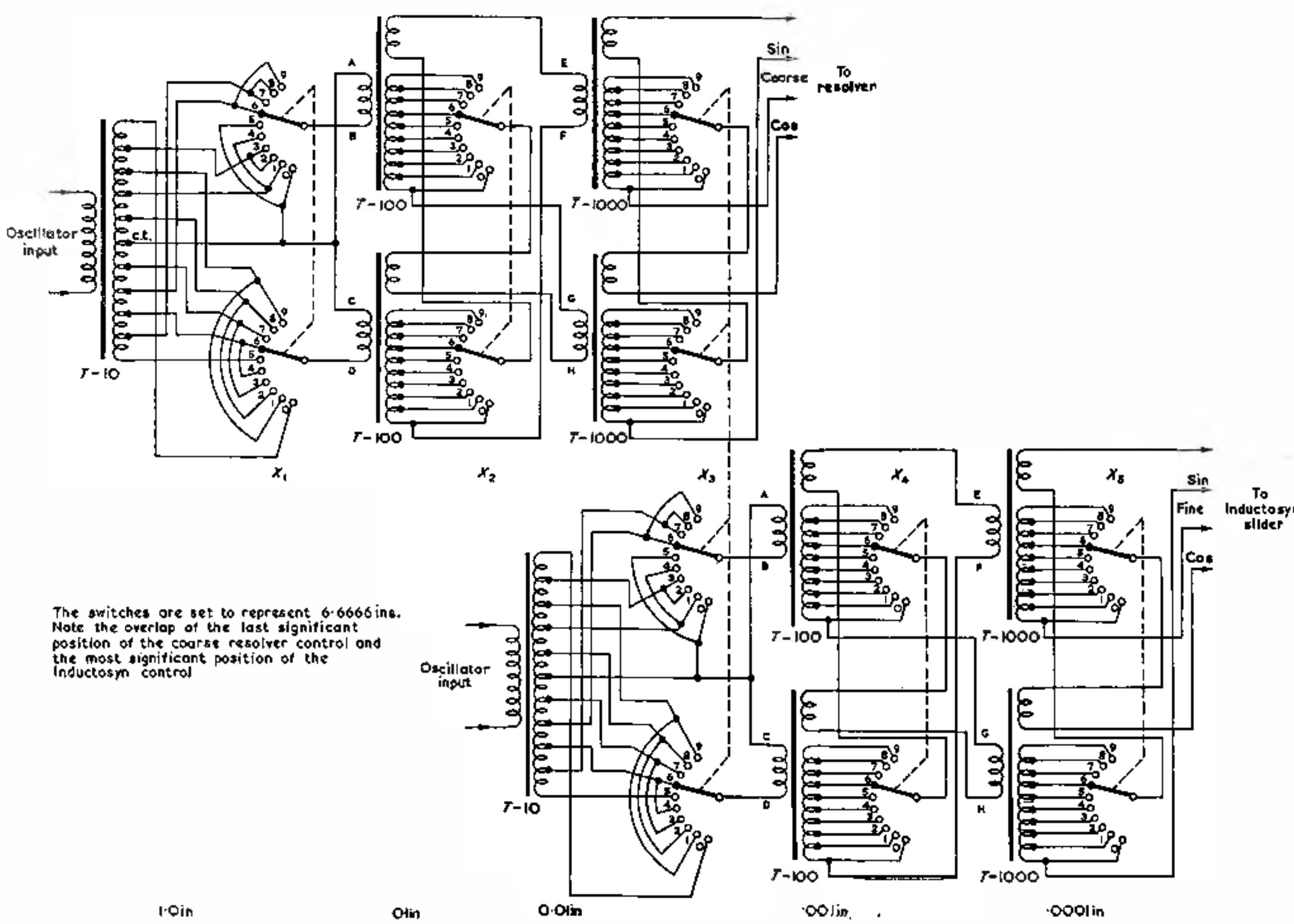


Fig. 3. Transformer digital to analogue convertor

Fig. 4. Schematic diagram of digital/analogue convertor



forms provided that the core material is suited to 10kc/s operation. 'C' cores using 0·004in strip would seem well suited electrically and also allow easy manufacture. Errors due to incorrect number of turns must be avoided and Table 1 gives some indication as to the tolerances that can be allowed. Measurements of the voltage ratios obtained are best made on each individual transformer using normal bridge techniques or simply by comparing with a known standard winding.

TABLE 1
Voltage Ratio Measurements

FUNCTION	VOLTAGE RATIO LIMITS	MEASURED RATIO		
		TRANSFORMER		
		1	2	3
sin 72 = 0·95106	0·95194 0·95015	0·9515	0·9515	0·9512
cos 36 = 0·80902	0·80730 0·81072	0·8095	0·8098	0·8092
sin 36 = 0·58778	0·59014 0·58544	0·5880	0·5889	0·5878
cos 72 = 0·30902	0·30625 0·31178	0·3094	0·3106	0·3094
−cos 72 = 0·30902	0·31178 0·30625	0·3085	0·3090	0·3097
−sin 36 = 0·58778	0·58544 0·59014	0·5869	0·5869	0·5892
−cos 36 = 0·80902	0·81072 0·80730	0·8087	0·8087	0·8090
−sin 72 = 0·95106	0·95015 0·95194	0·9511	0·9511	0·9510

Voltage ratio limits shown are 10' either side of relevant angle and 10' angular deviation represents approximately 0·00005in travel of the Inductosyn.

Application of the Linear Inductosyn to Position Control

A description of the various units required to be connected together to form a system for position control in one axis only follows:

Digital/Analogue Conversion

The digital information must first be converted into information suitable for feeding the Inductosyn element. As stated previously a trigonometric voltage analogue is used and a unit comprising of sine/cosine transformers and switches accomplishes this conversion. The same method can conveniently be used for feeding a resolver which acts as the coarse position element (Fig. 4).

The transformers may feed the resolver direct but a matching unit must be employed to couple the transformers to the Inductosyn element.

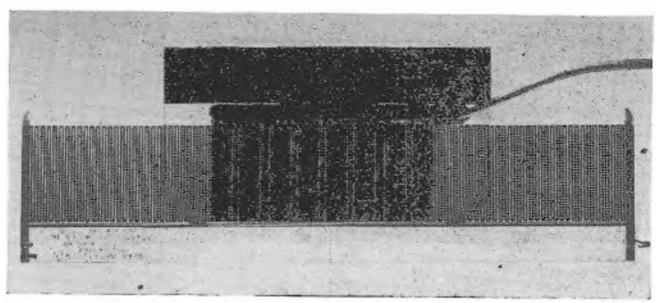
Coupling Digital/Analogue Transformers to Inductosyn

Two voltages, whose ratio is that of the sine and cosine of the represented angle form the output from the digital/analogue transformers. Since the voltage output of the transformers is about 1V for representing sin 90° or cos 0° and the optimum voltage input to the Inductosyn winding under these conditions is required to be approximately 100mV, and as the output impedance of the transformers is approximately 600Ω while the input impedance of the slider windings is approximately 0·5Ω, a matching unit is required having a voltage ratio of 10 to 1 and an impedance ratio of 1 200 to 1. Two of these units are required per axis, one for the sine winding and one for the cosine winding, with the important consideration that the final ratio of outputs be exactly that of the transformer outputs. A negative feedback amplifier fulfils these requirements satisfactorily (Fig. 5).

Inductosyn Element

The voltage output from the scale is dependent on its position from the electrical null of the slider. For a movement of 0·0001in from the null the output is about 6μV r.m.s.

The inductances of the Inductosyn scale and slider are only a few microhenries and consequently the impedance is mainly resistive at 10kc/s, therefore the output from the scale lags the input to the slider by almost 90°.



The Inductosyn electronic scale for linear micro positioning

Error Amplifier

Amplification of the 10kc/s signal from the scale has to be carried out to a suitable level to enable its phase and amplitude to be detected. This level is dependent on the type of detector employed but for a conventional phase detector employing transformers and selenium diodes it would be of the order of 10V.

Because of the very low impedance of the scale a large amount of amplification may be obtained in the input transformer itself and a factor of 1 000 to 1 may be realized. For consideration of noise the input stage should be tuned to a narrow pass-band around the signal frequency, and by using a low noise pentode considerable amplification may be obtained in this first stage. In the interests of valve type economy this input stage could well be a double triode valve, of the type used throughout the equipment, connected as a cascode amplifier.

This input stage is followed by further amplification and eventually fed into a cathode-follower which feeds the primary winding of the phase detector transformer.

Stability of the error amplifier is not a serious problem as the whole system is a null seeking device, ability to provide low noise at the null is by far the most important requirement as the ultimate resolution of the system is dependent upon the discrimination between the 10kc/s signal and noise.

Coarse-Fine Switching

A null voltage is produced by the Inductosyn every one twentieth of an inch travel between scale and slider. For measurements greater than this a coarse element is used, and it has been stated that a resolver is best suited for this purpose.

The resolver can be geared to the moving table and made to represent any desired length, the limit being set by the usable resolution of the resolver. That is, the resolution must be somewhat greater than the distance between nulls of the fine control. Using the Inductosyn as the fine control demands the overall resolution of gear train and electrical error of the resolver to be about 0·020in.

It is convenient to gear the resolver so that one complete revolution represents 10in of table motion. The sine/cosine supply to the resolver can then take exactly the same form as that supplied to the Inductosyn (Fig. 4). Resolvers may be obtained to work at 10kc/s or greater but the small frame size 15 resolvers for use at 400c/s have been found eminently suitable.

Further resolvers may be suitably geared and used for increasing the range of table movement still further.

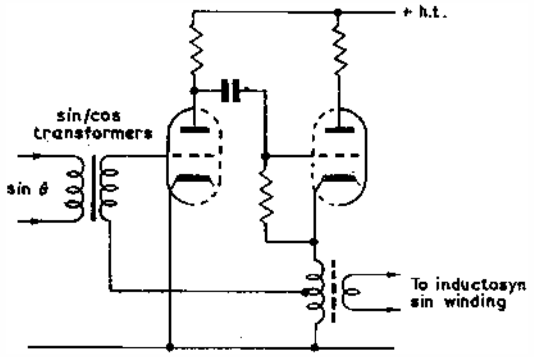


Fig. 5. Feedback amplifier coupling sin/cos transformers to Inductosyn slider

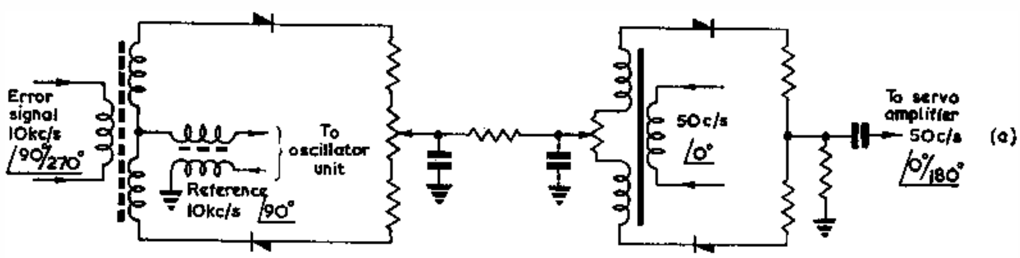


Fig. 6(a). 10kc/s phase detector and 50c/s modulator

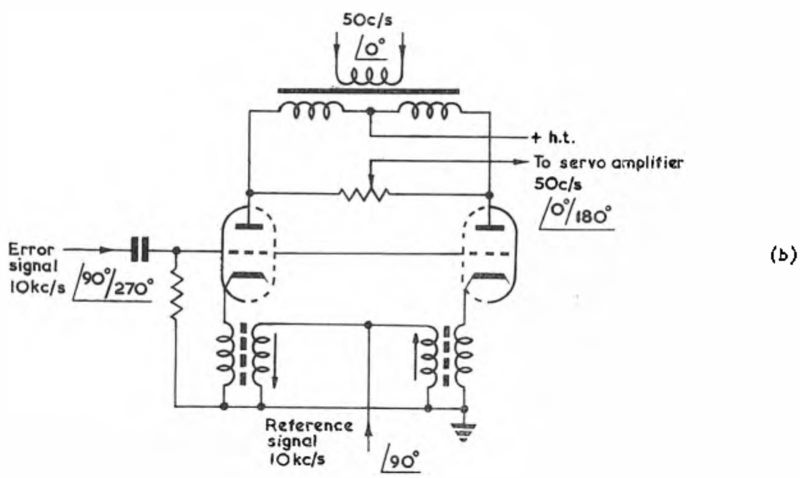


Fig. 6(b). Alternative 10kc/s detector and 50c/s modulator

The input to the 10kc/s phase detector must come from the most significant resolver until the table is moved to within the range of the Inductosyn. An electronic switch is used to accomplish this, and it is inserted between the error amplifier and the phase detector in the form of a diode switch; the most significant signal being used to bias off the least significant until some predetermined take-over point is reached.

Phase Detector

This may take one of many forms, but ultimately the output is a d.c. voltage, the polarity being determined by the relative phases of the signal and reference voltages, and the magnitude determined by the amplitude of the signal input (Fig. 6(a)).

The control signal applied to the servo system, from the detector may require to be a d.c. signal as above or if a 50c/s a.c. servo is used then a 50c/s signal, whose phase changes with respect to some reference signal when the 10kc/s signal phase changes, and whose amplitude is proportional to the 10kc/s signal, is required. The d.c. may be used to control a 50c/s modulator, as has been employed. A circuit is shown (Fig. 6(b)) where rectification of the 10kc/s signal and conversion to a 50c/s signal takes place in a double triode valve.

10kc/s Oscillator

The supply to the Inductosyn is nominally at 10kc/s but the actual frequency is of no real significance except that it should be a frequency giving good mutual characteristics between scale and slider and being low enough to be used without the need of undue screening of cables, components, etc. Frequency stability should be good, of the order of 0·5 per cent, and the harmonic content of the generated signal should be low. These requirements can be met by an *RC* oscillator fed from a stabilized h.t. supply (Fig. 7).

The generated 10kc/s has to provide two outputs per axis, one provides the input to the sine-cosine transformers, the other the reference voltage for the phase detector. Provision must be made for adjusting the relative phases of these two outputs to cater for the phase change of the signal input as mentioned earlier. A variable phase amplifier is incorporated in the oscillator unit for each axis.

Servo-system

The size of the machine to be controlled determines the power requirements of the servo-system. Electronic or magnetic amplifiers are perhaps best suited for power up to a few kilowatts but for moving very large masses quickly then a hydraulic servo has much to offer.

The electronic or magnetic amplifiers may be classified as follows:

(1) *A.C. Valve Amplifier*

This would normally be required to work at 50c/s in industrial conditions, and for low powers, up to about 100W, a reliable and reasonably efficient unit can be achieved. It would feed a 50c/s two-phase motor. The range of such motors is rather limited and confined mainly to low power machines of up to 50W.

(2) *D.C. Valve Amplifier*

Large powers may be obtained using d.c. amplifiers feeding split-field d.c. motors. Large stabilized power supplies are required and problems of drift have to be overcome.

(3) *Magnetic Amplifiers*

These are probably the most reliable, being completely static windings, and again large power outputs are obtainable. Where only 50c/s excitation is available the ultimate size is usually considerable. Difficulties are also encountered in time lag in response to the control signal but for a purely positioning device these may not be great.

(4) *Rotary Magnetic Amplifiers*

These are also used to control a split field d.c. motor and would seem to be best suited for power greater than 100W. They may be controlled by very small d.c. input signal powers, of the order of a few milliwatts for each 100W controlled. The overall size is somewhat smaller than the equivalent static magnetic amplifier.

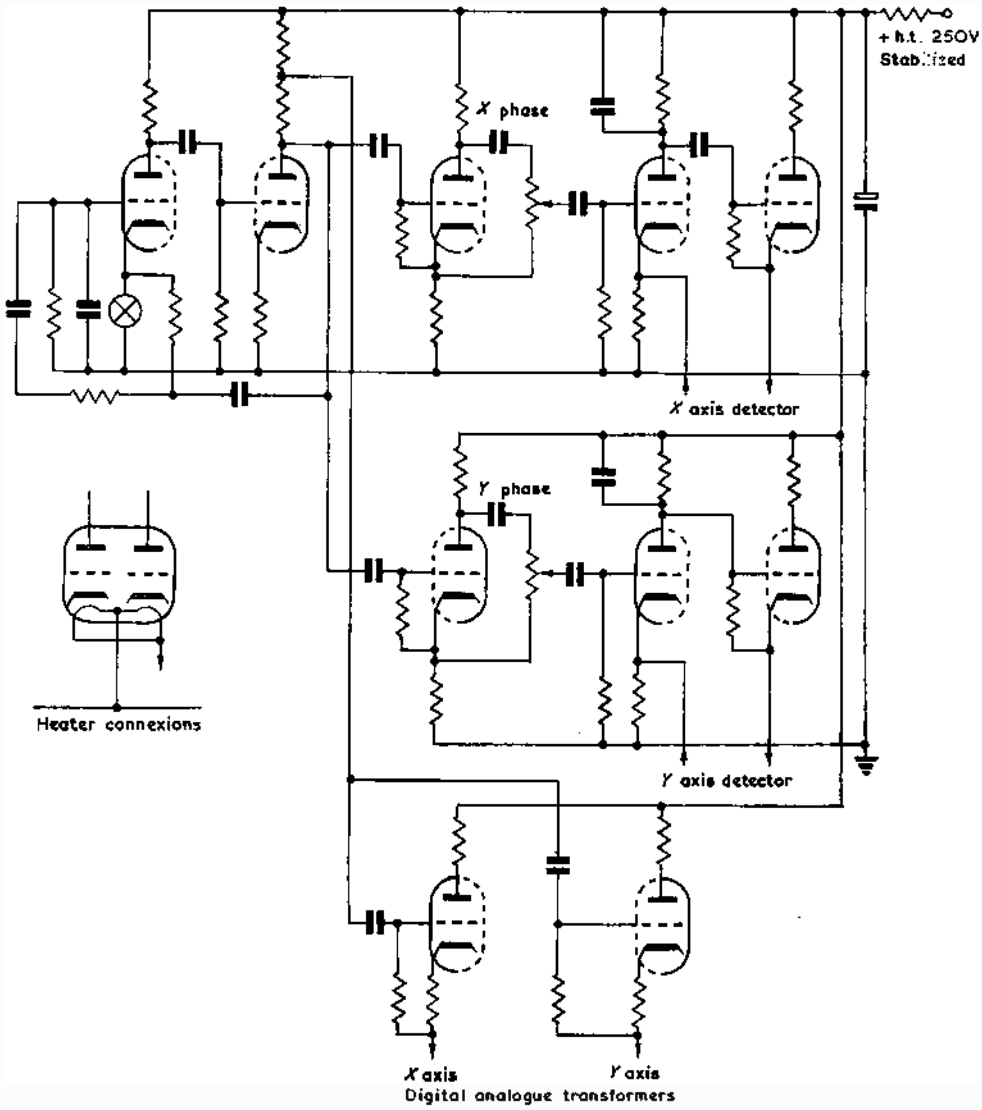


Fig. 7. Oscillator chassis

Whatever servo system is employed it would be controlled by the error signal appearing at the phase detector. In order to stabilize the system's overall response and to prevent undesirable overshoot and subsequent hunting it is expedient to provide negative feedback. A tachogenerator coupled to the motor can be used to provide negative feedback which is proportional to the angular velocity of the generator and hence proportional to the velocity of the machine table, this feedback signal can be introduced into the system in series with the error signal.

(To be continued)

Selective Amplification at Sub-Audio Frequencies

F. J. Hyde*, M.Sc., A.M.I.E.E.

Methods of obtaining selectivity and selective amplification at sub-audio frequencies are reviewed. Direct transmission systems discussed are the resonant galvanometer and the double CR (low-pass, high-pass) filter. Three of the feedback systems described are based respectively on the twin-T filter, the Wien bridge and the double CR network. Another is based on the zero-phase-shift oscillator and a final one is due to Schneider.

Advantages and disadvantages of the various systems are considered.

SELECTIVE amplifiers for use at sub-audio frequencies are now required for many applications. Some examples are in (a) detection of thermopile radiation, (b) electro-encephalography, (c) servomechanisms and (d) the study of very low frequency noise spectra of thermionic valves and of semiconductor devices.

In signal applications selectivity is used to improve the signal-to-noise ratio of the output of an amplifier, or to prevent unwanted signals from producing an output response. In the study of noise spectra selectivity is required if the noise spectral density (i.e. mean square noise current, voltage or power in unit bandwidth) is to be determined. The accuracy with which the spectral density at any frequency may be specified depends on its rate of change with frequency. In practice the spectral density usually varies approximately inversely as the frequency at low frequencies, so that if Q-factors of 5 or more are used there is negligible error in locating experimental points with respect to the frequency scale. If the Q-factor is made too high, however, then the statistical fluctuations of the noise in the narrow pass-band become large. For noise work, therefore, Q-factors between 5 and 20 are generally used.

The systems which are used to provide selectivity at sub-audio frequencies are of two types. These ar.:

(a) Direct transmission systems, in which the selectivity is determined directly, through maximum transmission of the basic frequency-sensitive network at its resonant frequency.

(b) Feedback systems, the nature of the feedback depending on whether the basic selective network has maximum or minimum transmission at the resonant frequency. In the first case appropriate positive feedback will increase the overall selectivity above that of the basic network, while in the second, negative feedback has the same effect.

An additional method which has been used[1] is to record the signals or noise being investigated on a slowly-moving magnetic tape, play back the tape at normal speed and then analyse the record using audio-frequency selective amplifiers.

Before dealing with systems (a) and (b) in detail, some general remarks concerning selectivity will be made.

Selectivity and Bandwidth

Inductors, which are frequently used in conjunction with capacitors to produce selectivity at audio-frequencies, become prohibitively large at frequencies much below 10c/s, so that, in general, resonant systems utilizing inductance are little used at frequencies below this. It is usual, however, to define the selectivity of 'ideal' single-tuned-stage sub-audio frequency amplifiers in terms of the quality factor Q in general use for inductance-capacitance resonant systems. In terms of this parameter, which in an ideal amplifier is invariant for the 'practical' range of frequencies on either side of resonance, the amplifier gain $A(\omega)$, defined as voltage output/voltage input, is of the form

$$A(\omega) = \frac{A_r}{1 + jQ[(\omega/\omega_r) - (\omega_r/\omega)]} = \frac{A_r}{1 + jQu} \quad \dots\dots (1)$$

Here $A(\omega)$ is the gain at angular frequency ω, and A_r is the gain at the resonant angular frequency ω_r.

The c.w. bandwidth B_{cw} is usually defined as the difference between the frequencies at which the modulus of $A(\omega)$ is reduced to $A_r/\sqrt{2}$; this is known as the 3dB bandwidth. In terms of Q, provided that Q is somewhat greater than unity, $B_{cw} = f_r/Q$, where $f_r = \omega_r/2\pi$.

The noise bandwidth B_N differs from the above because the individual frequency components, which make up the composite noise voltage, add on a power or square-law basis. Accordingly, the noise bandwidth is defined by

$$B_N = \int_0^\infty |A(\omega)/A_r|^2 \, d\omega \quad \dots\dots\dots\dots (2)$$

For a single ideal tuned-circuit B_N has the value $\pi f_r/2Q$ irrespective of the magnitude of Q.

In dealing with cascaded, synchronously-tuned stages, designed to have no interaction between them, the overall response is of the form

$$A_n(\omega) = \left\{\frac{A_r}{1 + jQ[(\omega/\omega_r) - (\omega_r/\omega)]}\right\}^n \quad \dots\dots\dots (3)$$

where n is the number of stages. For systems for which $n > 1$ the overall response cannot be rigorously defined in terms of an invariant Q-factor because the expansion of the right-hand side of equation (3) is no longer of the 'ideal' form of equation (1). It is sometimes convenient, however, in dealing with cascaded systems and also with non-ideal single stages, to define the selectivity in terms of an effective Q-factor, $Q' \equiv f_r/B_{cwn}$, where B_{cwn} is the difference between the 3dB-down frequencies. For cascaded systems B_{Nn} is defined in a similar manner to B_N in equation (2), the only difference being that the exponent 2 is replaced by $2n$.

Both the c.w. and noise bandwidths decrease progressively as n is increased, as may be seen from Table 1. The figures shown there are the coefficients which, when multiplied by f_r/Q and $\pi f_r/2Q$, yield B_{cwn} and B_{Nn} respectively.

Resonant Systems

Direct Transmission

Resonant Galvanometer

A resonant galvanometer affords a simple means of achieving selectivity, but lacks versatility because it is essentially a single-frequency system; the resonant fre-

* *Radio Research Station, Slough*

TABLE I

Reduction of Bandwidth with Increase in Number of Stages n.

Number of stages n	1	2	3	4	5	6
C.W. coefficient	1·0	0·64	0·51	0·44	0·39	0·35
Noise coefficient	1·0	0·5	0·37	0·31	0·27	0·25

quency is determined by the mechanical constants of the suspension and the moment of inertia of the armature plus coil assembly.

The general equation relating the angular deflexion θ of a galvanometer of internal resistance R_o, connected in series with an external resistance R, to an alternating e.m.f. $v = v_{pk} \exp(j\omega t)$, if inductive reactance is ignored, is

$$d^2\theta/dt^2 + 2\alpha_o(d\theta/dt) + \omega_o^2\theta = \omega_o^2 k_o i = \omega_o^2 k_o \left(\frac{v - 2g d\theta/dt}{R + R_o}\right) \quad \ldots\ldots\ldots\ldots (4(a))$$

or alternatively

$$d^2\theta/dt^2 + 2\alpha(d\theta/dt) + \omega_o^2\theta = \frac{\omega_o^2 k_o v}{(R + R_o)} \quad \ldots\ldots (4(b))$$

Here α_o is the damping constant of the open-circuited galvanometer, $\alpha \equiv \alpha_o + g k_o \omega_o^2/(R + R_o)$ is the damping constant under dynamic conditions, ω_o is the 'natural' frequency of the suspension, $2g$ is the back e.m.f. induced in the galvanometer per unit angular velocity, while k_o is the direct-current sensitivity, defined as θ_o/i_o, where θ_o is the steady deflexion resulting from a steady direct current i_o. The a.c. sensitivity, defined by

$$k(\omega) = (R + R_o)\ |\theta_{pk}|/|v_{pk}|$$

is found to be

$$k(\omega) = \frac{\omega_o^2 k_o}{\left\{(\omega_o^2 - \omega^2)^2 + (2\alpha\omega)^2\right\}^{\frac{1}{2}}} \quad \ldots\ldots\ldots (5)$$

which, by differentiation, is seen to have a maximum value at a frequency $\omega_r = (\omega_o^2 - 2\alpha^2)^{\frac{1}{2}}$ provided that $\omega_o > \sqrt{2}\alpha$. The maximum value of the sensitivity k_r is then $\omega_o^2 k_o / 2\alpha(\omega_o^2 - \alpha^2)^{\frac{1}{2}}$.

The Q-factor may be derived by inspection from equation (4(b)), by recognizing the analogy between this equation and that arising for a series tuned circuit. Accordingly

$$Q = \omega_o/2\alpha \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (6)$$

In Fig. 1 is shown a typical resonance curve for a commercial instrument for which $R_o = 1\text{k}\Omega$, $\omega_o = 0{\cdot}63$rad/sec, $\alpha_o = 0{\cdot}029$ and $k_o = 930\text{mm}/\mu\text{A}$. The ordinate axis is expressed in metres per microampere. The Q-factor is 11.

From the practical point of view the response of the galvanometer to direct current is troublesome if such a current is associated with the a.c. to be measured. This is the case for instance when excess noise is being measured in current-carrying semiconductor components.

A deflexion due to the direct current flowing can be avoided by two methods. (a) The response to d.c. is prevented by the inclusion of a large capacitance C (e.g. 1 000μF) in series with the galvanometer. Provided that $\omega C(R + R_o) \gg 1$, the formulæ given above for sensitivity are still valid. If not, a correction must be made. (b) A second method is to incorporate the noisy component being investigated as one arm of a Wheatstone bridge, the other arms of which are wire-wound resistors. The bridge is balanced for d.c. so that the galvanometer, which is arranged to be the detector at the bridge output terminals, experiences no steady deflexion. It will respond, however, to the noise e.m.f. developed at these terminals.

The galvanometer deflexion may be observed directly on a scale if the operating frequency is sufficiently low. A more convenient method, which may be used with a reflecting type galvanometer, is to photograph the focused, reflected light beam. A record of excess noise from a reverse-biased point-contact germanium diode obtained by this method is shown in Fig. 2. The waveform is quasi-resonant because the series capacitor which was used to prevent steady d.c. deflexion also discriminated against the very low frequency noise components.

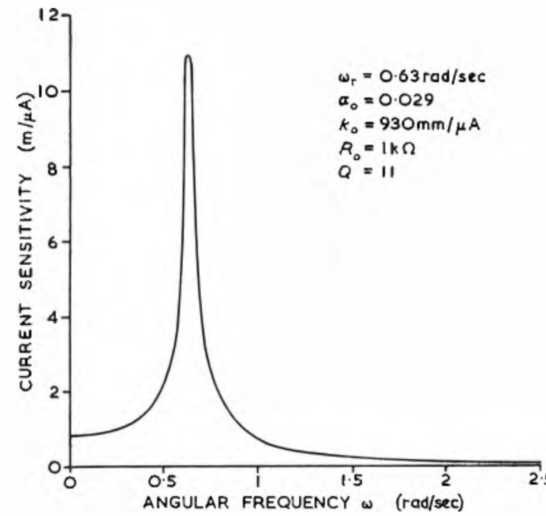


Fig. 1. Response of resonant galvanometer

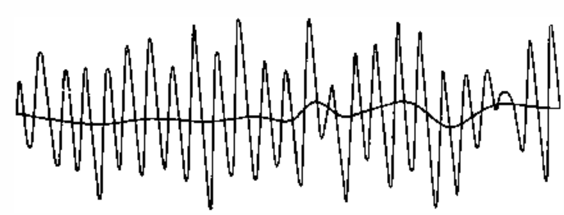


Fig. 2. Photographic record of quasi-resonant response of reflecting type galvanometer, to noise generated in a reverse-biased germanium diode.
Bandwidth 0·014c/s centred on 0·1c/s

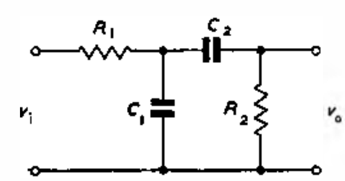


Fig. 3. Double *CR* selective network

Double CR Network

The four-element network of Fig. 3 which consists of a low-pass filter (R_1C_1) followed by a high-pass filter (R_2C_2) may be considered as representative of a class of selective *RC* networks having a maximum transmission at some frequency ω_r. The transmission is given by

$$\beta = \frac{j\omega C_2 R_2}{(1 + j\omega C_2 R_2)(1 + j\omega C_1 R_1) + j\omega C_2 R_1} \quad \ldots\ldots (7)$$

If $C_1R_1 = C_2R_2$ this may be put into the ideal form of equation (1). The parameter Q is then equal to $(2 + R_1/R_2)^{-1}$. If $R_1 \ll R_2$ then Q approaches its asymptotic limiting value of $\frac{1}{2}$. The transmission at the resonant frequency $\omega_r = (1/CR)$ is real and has a magnitude of $\frac{1}{2}$ in these circumstances.

A multi-channel noise spectrum analyser employing six synchronously-tuned cascaded stages, each comprised of one of the above four-element networks fed from a cathode-follower has been described by Nielson and van der Ziel[2]. Although the minimum resonant frequency used was 10c/s, the system can be used at much lower frequencies before the values of the resistors and capacitors become prohibitively large. Reference to Table 1 will show that the noise bandwidth of the six-stage filter is one quarter of that for a single-tuned stage, or from the c.w. standpoint the effective quality factor $Q' = 1{\cdot}4$, which is low.

Continuous tuning of such an amplifier would be best accomplished by varying the resistances, in view of the number of stages used, however, this would be cumbersome, so that stepped-tuning would be better. The whole system is wasteful, because the voltage attenuation of the six-stage filter is $1/2^6 = 1/64$ so that additional compensating amplification is required, over and above that which would normally be used.

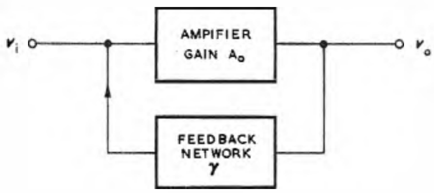


Fig. 4. Schematic circuit of basic feedback amplifier

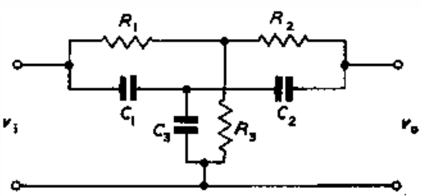


Fig. 5. Twin-T filter

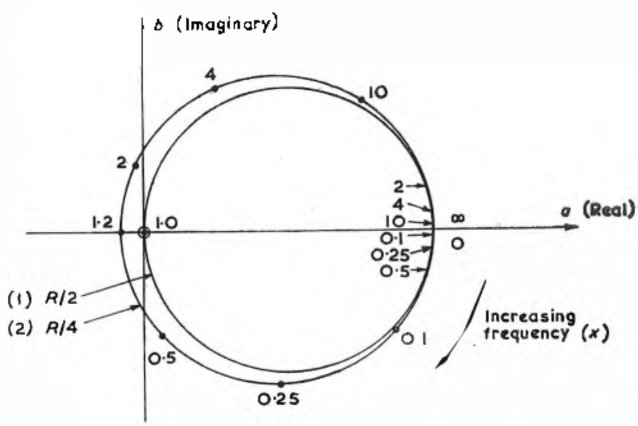


Fig. 6. Locus of transmission vector β of symmetrical twin-T filter
Shunt resistance arm (1) R/2 (2) R/4. $x = \omega/\omega_r$ is parameter

Feedback Systems

The basic feedback system is illustrated in Fig. 4. A fraction γ of the output voltage of an amplifier having internal gain A_0, is fed back in series with the input voltage. The overall gain of the system is then

$$A = \frac{A_0}{1 - \gamma A_0} \quad \ldots\ldots\ldots\ldots\ldots (8)$$

If γA_0 is positive we have positive feedback and vice versa. The overall selectivity is predominantly determined by that of γA_0 when γ is suitably chosen. Either A_0 or γ (or both) may be frequency-dependent. For sub-audio frequency working such frequency-dependence is produced by capacitance-resistance networks. Selective systems built around the twin-T filter, the Wien bridge and the double CR network respectively are described below, and also two others, the first due to Schneider and the second based on the 'zero-phase-shift' oscillator.

Use of Twin-T Filter

The twin-T filter is illustrated in Fig. 5. It was first described by Scott[3]. Provided that

$$\frac{C_1 + C_2}{C_3} = \frac{R_1 R_2}{R_3(R_1 + R_2)}$$

the transmission is zero at a resonant frequency defined by

$$\omega_r = \frac{1}{\{(R_1 + R_2)R_3C_1C_2\}^{\frac{1}{2}}} \quad \ldots\ldots\ldots (9)$$

The transmission β is given by

$$\beta = \frac{j(u/\delta)}{1 + j(u/\delta)} \quad \ldots\ldots\ldots\ldots\ldots (10)$$

where the attenuation factor $\delta = \{\omega_0 C_2(R_1 + R_2) + (C_2 + C_1)/\omega_0 C_1 C_2 R_2\}$.

For simplicity, balanced symmetrical filters are generally used. For these $C_1 = C_2 = C_3/2 = C$ and $R_1 = R_2 = 2R_3 = R$, so that the resonant frequency and transmission become respectively

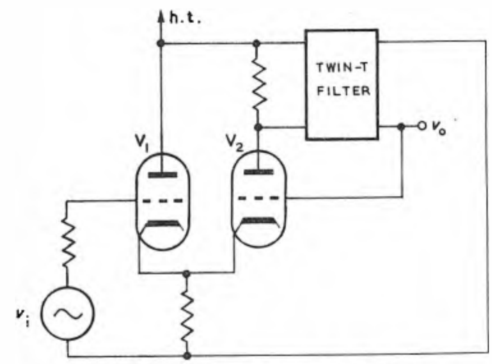


Fig. 7. Series feedback through twin-T filter

$$\omega_r = (1/CR) \quad \ldots\ldots\ldots\ldots\ldots\ldots (11)$$

$$\beta = \frac{j(u/4)}{1 + j(u/4)} \quad \ldots\ldots\ldots\ldots\ldots (12)$$

Plotted in the complex plane i.e., as $a + jb$, the locus of the tip of the transmission vector β is a circle (curve (1) in Fig. 6) tangential to the imaginary axis at the origin, with its centre $(\frac{1}{2}, 0)$ on the real axis.

If such a twin-T filter is connected into the basic circuit of Fig. 4 in such a way that β of equation (12) is equal to $-\gamma$ of equation (8), A_0 being positive and invariant with frequency, then the overall gain A is given by

$$A/A_0 = \frac{1 + j(u/4)}{1 + j(A_0 + 1)u/4} \quad \ldots\ldots\ldots (13)$$

The schematic circuit by means of which the feedback through a twin-T filter is applied in series with, and in phase opposition to the input voltage is shown in Fig. 7. The input signal is fed to the cathode and the feedback signal to the grid of V_2. Ideally the source resistance presented to the filter should be low and the output resistance into which it works should be high.

Although equation (13) does not have the ideal form of equation (1) it is apparent that if $A_0 \gg 1$, the denominator is very much more frequency-dependent than the numerator. Accordingly the numerator may be considered as invariant throughout the 'practical' range of frequencies on either side of ω_r. By inspection and comparison with equation (1) one then finds

$$Q' = \frac{A_o + 1}{4} \quad \ldots\ldots\ldots\ldots\ldots\ (14)$$

For circuit details of practical single-stage amplifiers based on the schematic circuit of Fig. 7 reference may be made to Sturtevant[4] and Fleisher[5]. Brown[6] has described an amplifier having two synchronously-tuned stages.

The tuning of amplifiers in which the feedback network is of symmetrical form may be accomplished by ganged variation of all three resistors or capacitors. (The resistors are the more suitable elements to vary at low frequencies). Q' is independent of the resonant frequency and may be controlled by variation of A_o.

If an asymmetrical filter is used in a feedback amplifier, then tuning may be affected by ganged variation of two circuit components[7]. The effective quality-factor is a function of the resonant frequency in this case.

The selectivity and resonant gain of a symmetrical twin-T resonant amplifier may be enhanced if either the shunt resistance arm of the filter is made less than $R/2$ or the shunt capacitance arm is made greater than $2C$. The reason[8] is that the locus of the transmission vector β passes into the second and third quadrants in the neighbourhood of the resonant frequency, so giving rise to selective positive feedback. The overall selectivity increases as (for example) the shunt resistance arm of the filter is reduced. If the value of this resistance is made not less than $R/4$, then the locus of β remains approximately circular; the actual locus for this particular value of resistance is shown as curve (2) in Fig. 6. In these circumstances the locus of the overall transmission is also a circle, although the distribution of the individual frequency points around it differs from that generated by the 'ideal' equation (1). An effective Q-factor must therefore be used to specify the selectivity as for the balanced symmetrical filter.

Curves relating Q' to the fraction $\dfrac{\text{actual shunt resistance}}{R/2}$ are given[8] for different values of A_o. The enhancement of selectivity that is obtained by unbalancing the filter is accompanied by a rise in the resonant frequency above the value $(CR)^{-1}$rad/sec. This may be allowed for in design.

Use of Wien Bridge

The Wien bridge is illustrated in Fig. 8. The resonant frequency, which may be defined as that frequency other than 0 or ∞ rad/sec, at which the transmission is real, is given by

$$\omega_r = \frac{1}{(C_1R_1C_2R_2)^{\frac{1}{2}}} \quad \ldots\ldots\ldots\ldots\ldots\ (15)$$

The transmission is

$$\beta = \frac{l}{1 + jqu} - k \quad \ldots\ldots\ldots\ldots\ldots\ (16)$$

where $l = C_1R_2/(C_1R_1 + C_2R_2 + C_1R_2)$, $k = R_4/(R_3 + R_4)$ and $q = l(C_2R_1/C_1R_2)^{\frac{1}{2}}$. q has a maximum value of $1/(2+\alpha)$, where $\alpha = R_2/R_1 = C_1/C_2$ when $C_1R_1 = C_2R_2$. If $k = l$ then β is zero at the resonant frequency. Plotted in the complex plane (curve (1) of Fig. 9) the locus of the transmission vector is a circle, tangential to the imaginary axis at the origin with its centre $(-l/2, 0)$ on the negative real axis. If l is kept fixed but k is reduced, then the circle diameter remains as l but the centre moves to $\left\{-(k-l/2), 0\right\}$ so that the transmission vector passes into the first and fourth quadrants (curve (2) of Fig. 9) in the neighbourhood of the resonant frequency, which remains unchanged.

It is evident that in a feedback amplifier the feedback must be negative as the extremes of frequency are approached; that is γA_o of equation (8) should be negative towards zero and infinite frequency. This may be brought about by having A_o positive and $\gamma = \beta$. Under these circumstances the feedback will be negative at all frequencies if $l < k$, although it will be positive over a range of frequency in the vicinity of ω_r if $l > k$. The range of frequencies over which positive feedback occurs will increase as k becomes increasingly less than l.

Putting $\beta = \gamma$ in equation (8) leads to the following dependence of gain on frequency;

$$A/A_o = \frac{\dfrac{1 + jqu}{1 - A_o(l - k)}}{1 + jq\left(\dfrac{1 + A_o k}{1 - A_o(l - k)}\right)u} \quad \ldots\ (17)$$

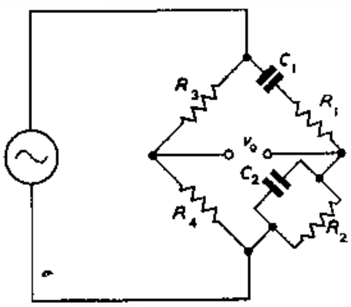


Fig. 8. Wien bridge

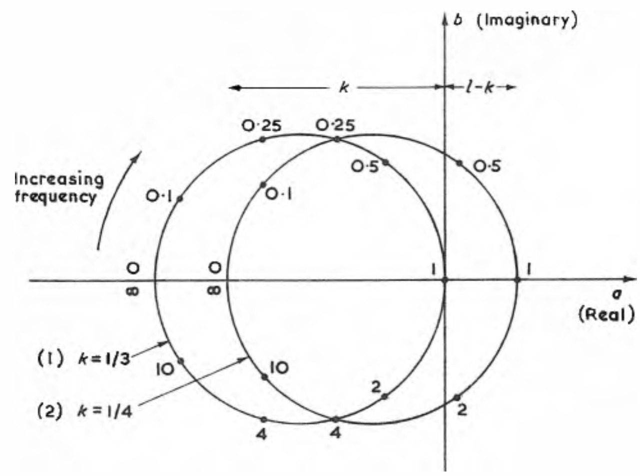


Fig. 9. Locus of transmission vector β of Wien bridge
(1) $q = l = k = 1/3$; (2) $q = l = 1/3$, $k = 1/4$. $x = \omega/\omega_r$ is parameter

This is obviously not of the ideal form of equation (1), but if $(1 + A_o k)/\left\{1 - A_o(l - k)\right\} \gg 1$ then, as for the twin-T filter amplifier, the denominator will be very much more frequency-dependent than the numerator. Under these circumstances equation (17) does behave similarly to equation (1) in the neighbourhood of the resonant frequency and by inspection the selectivity may be specified by $Q' \equiv q(1 + A_o k)/\left\{1 - A_o(l - k)\right\}$. It is evident therefore that for a fixed value of l the value of Q' can be controlled by variation of k leaving the resonant frequency unaffected. Oscillation will occur if $A_o(l - k) \geqslant 1$. As an example of a selective system we may take $q = k = l = \frac{1}{3}$, $A_o = 600$, in which case $Q' \simeq 70$.

A practical circuit of a Wien bridge selective amplifier for the frequency range 40 to 600c/s is given by Picchi[9]. Selective amplification may be obtained at much lower frequencies than this. A disadvantage of this system stems from the bridge structure of the selective circuit, which prevents one terminal of the input and output from being common (i.e. earthed) as is possible with the twin-T filter.

Tuning may be effected by variation of R_1 and R_2. If these resistances are ganged then l is made independent of the resonant frequency so that Q' is invariant with frequency. Q' may be varied independently through variation of the bridge ratio k.

Use of Double CR Network

The system is due to Beattie and Conn[10]. Use is made of the selectivity of the double CR network shown in Fig 3, in which $C_1 = C_2 = C$ and $R_1 = R_2 = R$. For these conditions the transmission β becomes

$$\beta = \frac{1/3}{1 + \mathrm{j}u/3} \quad \text{.................} \quad (18)$$

The above network is connected in cascade with an amplifier that has uniform gain A' throughout the practical frequency range. A fraction γ of the output voltage $v_o = A_o v_i = A'\beta v_i$ of the cascaded system is fed back in phase to the input (positive feedback) so that the expression for overall gain is

$$A/A' = \frac{\beta}{1 - A'\beta\gamma} = \frac{1/(3 - A'\gamma)}{1 + \mathrm{j}[\mu/(3 - A'\gamma)]} \quad \text{......} \quad (19)$$

By inspection it is seen that the Q-factor of the system, whose transmission is of the ideal form, is $1/(3 - A'\gamma)$†. It is evident that the Q-factor depends on A' and on γ. The resonant frequency is independent of either of these parameters and is given by $\omega_r = (1/CR)$. Accordingly both Q and ω_r may be varied independently. In the circuit described by Beattie and Conn Q was made continuously variable from 5 to 20 by varying $A'\gamma$ while the resonant frequency $f_r(=\omega_r/2\pi)$ was continuously variable from 1 to 3c/s, by ganged variation of the resistive branches.

Schneider's System

A 'New Type of Electrical Resonance' was described by Schneider[11] in 1945. The schematic circuit (Fig. 10), comprises (a) an amplifier having invariant stage gain A_o over the whole of the practical frequency range and giving phase-reversal; (b) an input circuit consisting of resistors R_1R_2 and capacitor C_1 and (c) a feedback capacitor C_2 connected between the final anode and the input side of C_1. The stage gain A_o referred to above is that between the first grid and the final anode with C_2 removed. Provided that $A_oR_1/\omega C_2 \gg R_3/\omega C_1$, R_3R_1 and that the leakage resistances of the capacitors are much larger than their reactances at the lowest frequencies involved, the overall gain approximates to the ideal form of equation (1). (R_3 is the output resistance of the amplifier with C_2 removed).

The resonant frequency and Q'-factor are respectively

$$\omega_r = 1/(A_1R_1C_1R_2C_2)^{\frac{1}{2}} \quad \text{............} \quad (20)$$

$$Q' = (A_1^{\frac{1}{2}}/g)\,(R_2C_2/R_1C_1)^{\frac{1}{2}} \quad \text{............} \quad (21)$$

where $A_1 = A + 1 + (R_3/R_2) + (R_3/R_1)$, $g = 1 + (R_2/R_1) + (R_2 + R_3)C_2/R_1C_1$.

The system is inherently stable because the feedback is degenerative; the phase reversal produced by the internal amplifier is an essential feature.

The most interesting property of the amplifier is contained in equation (20); the resonant frequency is approximately inversely proportional to the square root of the aperiodic stage gain. To achieve very low frequency resonance therefore, in addition to making the component values C_1, C_2, R_1 and R_2 large, the stage gain A_o may also be made large. To obtain minimum amplifier phase shift and preserve the approximately ideal form of the response, it is advisable to use direct coupling between valves if more than one are used.

† *The symbol Q used by Beattie and Conn has twice this value. Their definition is not the usual one for this symbol.*

The selectivity is enhanced by an increase in A_o, being approximately proportional to $\sqrt{A_o}$. Q' may be maximized if it is arranged that $R_2/R_1 = C_1/C_2 = \sqrt{(R_3/R_1)}$, when it has the value $\sqrt{A_1}/2\ [1 + \sqrt{(R_3/R_1)}]$. It is evident that R_3/R_1 should be kept as small as possible. Tuning without changing Q' is readily effected by switching C_1 and C_2 in pairs and maintaining their ratio. Continuous tuning may be achieved by ganged variation of R_2 and R_1: provided that $R_3 < R_1, R_2$ then Q' is not greatly affected. A step-tuned amplifier between frequencies of 5×10^{-3}c/s and 10^{-1}c/s has been described[12] with $Q' \simeq 8$ and also a band-pass amplifier[13] comprised of several basic resonant units, having staggered resonant frequencies, in parallel with one another.

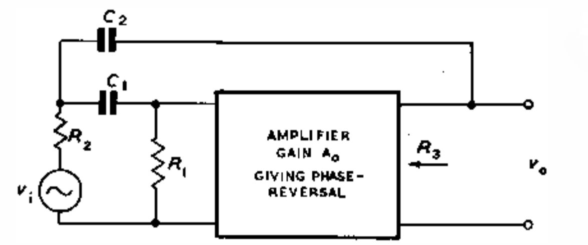


Fig. 10. Schematic circuit of Schneider's resonant system

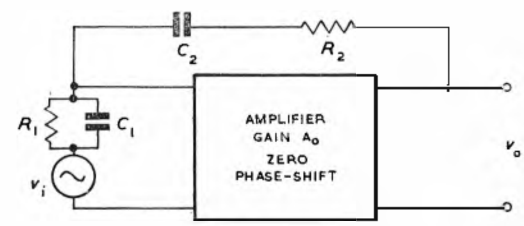


Fig. 11. 'Zero-phase-shift' resonant system

'Zero-Phase-Shift' Feedback System

This system which also utilizes two resistors and two capacitors is illustrated in Fig. 11. Unlike the previous system, however, regenerative feedback is used. The principle has already been used by Terman *et al*[14] in the construction of an oscillator.

If $C_1 = C_2 = C$ and $R_1 = R_2 = R$ and additionally R_2 includes the output resistance seen at the final valve anode of the internal amplifier, then a simple expression for the overall voltage gain is

$$A = \frac{u + \mathrm{j}/2}{u + \mathrm{j}(3 - A_o)} \quad \text{.............} \quad (22)$$

The resonant frequency is defined by $\omega_r CR = 1$. The expression for A is non-ideal, but approaches the ideal if A_o is less than, but approximately equal to 3. (If $A_o > 3$ the system is oscillatory). Under such circumstances the selectivity is described by

$$Q' = \left\{\frac{2 - (3 - A_o)^2}{2(3 - A_o)^2}\right\}^{\frac{1}{2}} \simeq \frac{1}{3 - A_o} \quad \text{......} \quad (23)$$

Although the system is regenerative, it may still possess a high degree of stability because the stage gain required is very low; this lends itself to the use of strong internal negative feedback.

Because Q' is determined solely by A_o, while the resonant frequency depends on C and R, there will be no interaction between Q' and ω_r provided that change of A_o to produce variable Q' does not give rise to a change in output impedance of the amplifier.

A practical amplifier of this type has been described by Morris and Dawe[13].

Stability

Provided that good quality components with small temperature coefficients are used in the frequency-sensitive networks, then the overall stability will be determined by that of the basic amplifier gain A_0. Variation of A_0 may be minimized in the usual way by use of internal negative feedback in the basic amplifier.

Conclusions

Of the systems considered, apart from the galvanometer, whose resonant frequency is determined by mechanical constants, the resonant frequencies of all the capacitance-resistance systems except that due to Schneider are of the form

$$f_r = 1/2\pi CR \qquad (24)$$

Bearing in mind the requirements imposed by intervalve coupling, a maximum value for R of about 2MΩ may be taken. For the capacitors, 20μF is a reasonable maximum figure considering both bulk and the requirement that $\omega CR_L \gg 1$ at all frequencies, where R_L is the shunt leakage resistance. From these figures a minimum resonant frequency of about 4×10^{-3}c/s is calculated. Lower resonant frequencies than this may be achieved using Schneider's system, in which an additional multiplying factor of $(1/A)^{\frac{1}{2}}$ occurs on the right-hand side of equation (22).

With the double CR, the balanced twin-T filter, the Wien bridge and the 'zero-phase-shift' systems, both the Q-factor and the resonant frequency may be varied independently with no interaction between them. Using unbalanced twin-T filters there is interaction, a rise in Q-factor being accompanied by a rise in resonant frequency. In Schneider's system, although the frequency may be varied independently an increase in selectivity achieved by increasing A_0 will also result in a lowering of the resonant frequency.

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A V.H.F. Radio Telephone Link for Needles Lighthouse

For nearly a hundred years keepers on the Needles lighthouse, which guards the important shipping lanes around the Isle of Wight, have relied upon lamp signalling to maintain communication with the shore. In bad weather poor visibility often made the passing of operational and emergency messages to the mainland extremely difficult, if not impossible.

Now this problem has been overcome by the installation of a v.h.f. radio-telephone link—the first of its kind to be employed by Trinity House for communication between one of its rock lighthouses and the shore.

Using the new equipment, which provides a normal telephone service, the Needles' crew will in future be able to contact the shore station at St. Catherine's Point—fourteen miles to the east—within seconds of an emergency arising.

No special operating procedure is necessary, and a person making a call would normally be unaware that the conversation was being transmitted by radio.

The equipment developed by the Automatic Telephone & Electric Company Group is designed to meet these special needs. In order that power consumption should be reduced to the minimum, one of the Company's battery-operated 'Country Sets' is being used at the Needles.

The equipment is intended for use in places where there are no mains supplies, and the conservation of power is essential, and a 12V car battery can last for up to a month without re-charging. This extremely low consumption has been achieved by various economy features, including a cyclic switching device which automatically turns the set on for approximately three seconds in every thirty to 'listen-out' for in-coming calls.

At St. Catherine's Point, the mainland end of the link, where power supply problems are not so severe, A.T.E's type RLS mains equipment is being used. This has been designed so that in an emergency calls can be made through the G.P.O. telephone system. At both the Needles and St. Catherine's Point, aerials mounted high on the lantern galleries of the lighthouses are beamed towards each other to achieve an optical path.

A second V.H.F. installation similar to that at the Needles will shortly link Flatholm, an island in the Bristol Channel, with the coastguard station at Barry.

A New Loudspeaker System

A new loudspeaker system which embodies new design principles and represents an advance in high-quality sound reproduction has been developed at the Research Laboratories of The General Electric Co. Ltd.

The system has been called the 'periphonic' loudspeaker system. It exploits fully the low distortion qualities of the metal-cone loudspeaker.

In the normal way loudspeakers are mounted inside the loudspeaker cabinet. In the periphonic system the sound is radiated through a small slot into the cabinet from the peripheries of two metal-cone loudspeakers mounted, one inside the other, in a V-shaped enclosure on the outside of the cabinet. The subsequent 'air-coupling' of the two loudspeakers reduces the distortion which occurs with a single speaker, particularly at the bass frequencies, by over 60 per cent.

The complete unit consists of the periphonic speaker system and six presence units. The speaker system is mounted on a cabinet which is designed to present the correct acoustic loading to the small slot, and to reproduce faithfully the lower frequencies from 2kc/s down to 30c/s. The presence units cover the frequency range above 2kc/s. They are completely sealed at the back so that no sound is radiated from them into the cabinet.

In spite of its large size, the cabinet is so constructed as to eliminate all resonances. The structure is reinforced with substantial struts, and in addition diaphragms are placed within the enclosure to break up any resonances of the air columns.

The system takes advantage of the wide arc of distribution of the higher frequencies which is associated with the presence unit. By the use of three separate pairs of units mounted in the front and on either side of the cabinet, an effect of spaciousness is given to the sound reproduction.

A Logarithmic Wattmeter

By J. A. Bennet*, Ph.D.

Some of the limitations of the conventional dynamometer type of wattmeter for the measurement of arc loss in mercury-arc rectifiers are first discussed. An electronic wattmeter using logarithmic and anti-logarithmic circuits is then described. The anti-logarithmic circuit uses a thermionic valve biased negatively into the exponential region, while the logarithmic circuit is a linear amplifier in conjunction with an anti-logarithmic negative feedback path. Finally the application of the instrument to the measurement of mercury-arc rectifier arc loss is described.

If a dynamometer wattmeter is used to measure the power dissipated in the arc of a mercury arc rectifier two factors arise which make the measurement more difficult than in ordinary a.c. circuits, namely the high reverse voltage between anode and cathode of the rectifier during the non-conducting part of the cycle, and the d.c. component of the arc current which prevents the use in the ordinary way of a current transformer between the rectifier anode circuit and the current circuit of the wattmeter. Various methods of arc-loss measurement have been described in the literature from time to time. In some cases the measurement is carried out indirectly and the above difficulties avoided, while in other cases special circuits are used in order to overcome them. The first part of this article contains a description of two original protective circuits of this type, which have been constructed and which overcome the former difficulty, while the second part describes an electronic wattmeter developed and used for arc loss measurement, in which both difficulties are overcome.

Protective Circuits

Conventional Protective Circuit using Diode Shunt

This is shown in Fig. 1, and consists of a diode shunted across the wattmeter potential circuit so that it conducts when the particular anode of the mercury-arc rectifier to which the wattmeter is connected is not conducting, so preventing excessive rise of voltage across the wattmeter potential circuit. Resistor R is used to limit the diode current to a safe value. Either the diode must have a current rating several times that of the wattmeter potential circuit, or the resistance of R must be so large that it seriously reduces the wattmeter sensitivity, and in practice the arrangement is only moderately satisfactory.

Single-Triode Protective Circuit

In an attempt to obtain more satisfactory operation than with the diode-shunt circuit, the protective arrangement shown in Fig. 2 was used. The ratio of the potential divider $R_{1,2}$ is adjusted to give a voltage gain of unity from the anode and cathode of the mercury arc rectifier to the potential circuit of the wattmeter, when the anode in question is conducting. During the remainder of the cycle the flow of grid current through R_3 prevents all but a small signal reaching the triode grid, so that there is no excessive rise in current through the wattmeter potential circuit. The current which does flow during this part of the cycle does not affect the wattmeter reading since no current is then flowing in the wattmeter current circuit.

In operation, S is first closed, removing the signal from the triode grid, and the wattmeter reading noted. S is then opened, and the wattmeter reading falls to a new value which is again noted. The difference of the two readings gives the arc power dissipated for the particular anode concerned.

The single-triode protective circuit has the merit that the wattmeter is more fully protected than with the diode-shunt circuit, since it cannot be damaged by failure of the valve heater circuit, and breakdown of the valve insulation is less likely on account of the much lower voltages applied to

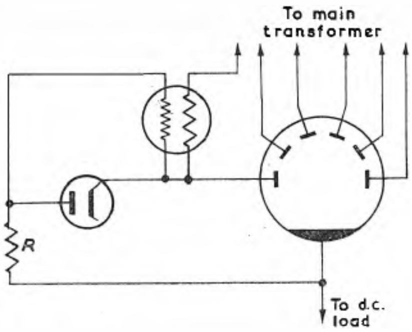

Fig. 1. Conventional protective circuit

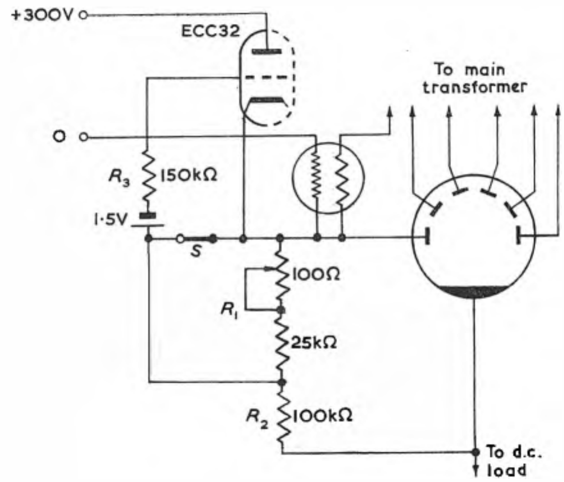

Fig. 2. Single-triode protective circuit

the valve. Moreover, the current rating of the triode valve need not be greater than that of the wattmeter potential circuit, although there is no reduction in the sensitivity of the instrument. However, the circuit has the following serious disadvantages:

(a) the operating characteristic shows significant non-linearity over the working range,

(b) the voltage gain of the valve circuit is sensitive to changes in supply voltage,

(c) the arc loss has to be calculated as the difference of t o wattmeter readings, each of which is altered when any change is made in the load on the mercury-arc rectifier, and

(d) only a limited part of the wattmeter scale can be usefully employed, as is explained below.

* *University of St. Andrews, Dundee.*

DOUBLE-TRIODE PROTECTIVE CIRCUIT

The disadvantages of the single-triode circuit are largely overcome with the circuit shown in Fig. 3. V_1 and V_2 function essentially as cathode-followers, but to compensate for the voltage attenuation from grid to cathode of V_1, a small signal is taken from the anode of V_1 and applied to the grid of V_2, so that there is unity voltage gain from the anode and cathode of the mercury-arc rectifier to the potential circuit of the wattmeter, which is connected between the cathodes of V_1 and V_2. VR_1 and VR_2 are used for zero setting and for adjusting the gain to unity respectively. Provided the voltages dropped across the anode and cathode resistors of V_1 and V_2 are suitably proportioned, the gain of the circuit is not sensitive to changes in supply voltage, and the power loss is given directly by the wattmeter reading, and not as the difference of two readings.

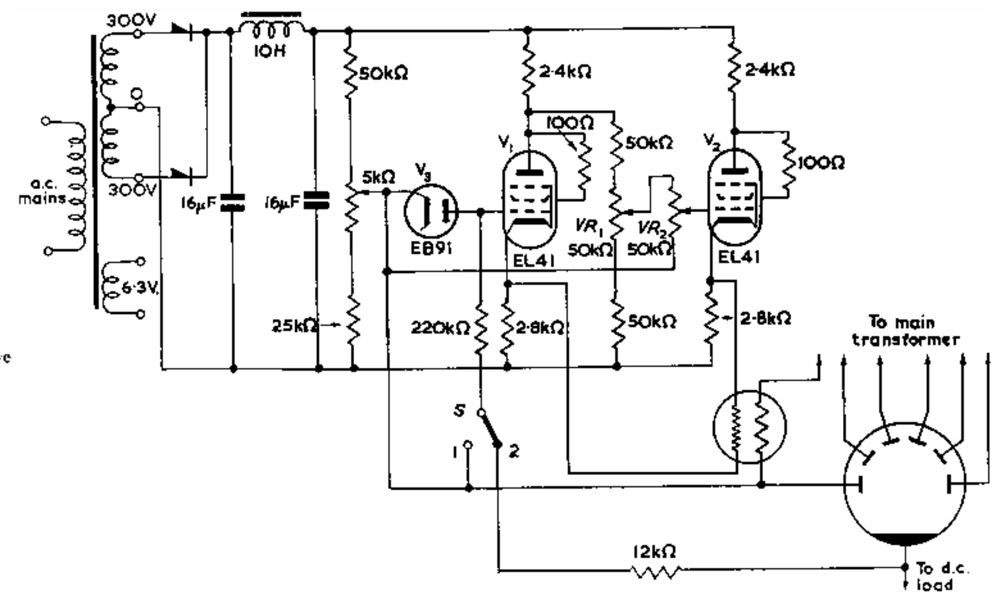

Fig. 3. Double-triode protective circuit

Wave Factors of Arc Current

If a multi-anode mercury-arc rectifier is operated in such a way that each anode carries current during one third of each cycle, the current to any one anode will approximate to a series of rectangular pulses as shown in Fig. 4. If I denotes the direct-current output of the rectifier, the average and r.m.s. values of rectifier anode current will be:

$$(0{\cdot}5/3)\,I = 0{\cdot}167\,I \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

and

$$\sqrt{[(0{\cdot}5\,I)^2/3]} = 0{\cdot}289\,I \text{ respectively} \quad \ldots\ldots \quad (2)$$

and the peak and form factors of the wave will each be

$$\sqrt{3} = 1{\cdot}732 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

(In practice the anode current does not start and stop abruptly as indicated, which results in the peak and form factors being rather less than the figures calculated above).

Usable Fraction of Wattmeter Scale

DIODE-SHUNT CIRCUIT

If I_w and E_w represent the ratings of the current and potential circuits respectively of the wattmeter, and full-scale deflexion corresponds to I_wE_w watts; and if the waveform of the anode current of the mercury-arc rectifier is as shown in Fig. 4, the wattmeter current circuit will be fully loaded when the r.m.s. value of the rectifier anode current is

$$I_w = (0{\cdot}5\,I)/\sqrt{3} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

If the wattmeter potential circuit is also operated at its full rating, the p.d. across it being approximately $+E_w$, while the anode is conducting and $-E_w$ during the remainder of the cycle, the instrument reading will be

$$E_w \times (0{\cdot}5\,I/3) = E_w \times (\sqrt{3}\,I_w/3) = 0{\cdot}577\,E_wI_w \quad \ldots \quad (5)$$

The assumption that the p.d. across the wattmeter potential circuit has the magnitude E_w throughout the cycle represents about the best conditions obtainable with an ordinary portable substandard wattmeter having a potential circuit rating of about 10mA, with a thermionic diode.

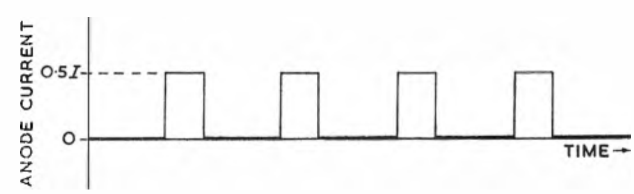

Fig. 4. Approximate waveform of mercury-arc rectifier anode current

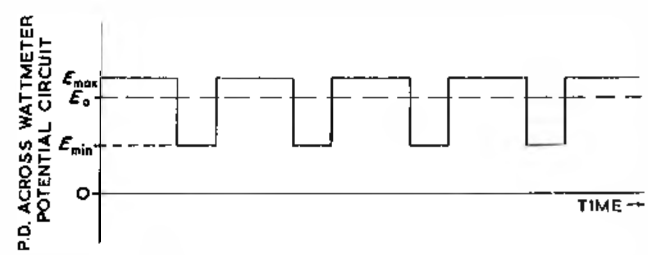

Fig. 5. Approximate waveform of p.d. across wattmeter potential circuit, with single-triode protective circuit

Better results could probably be obtained with a germanium diode. It has been assumed in the above discussion that the full-scale deflexion of the wattmeter is equal to the product of the potential- and current-circuit ratings. Instruments with more sensitive movements are constructed, and are useful in measurements of this type, but they tend to be less robust. It has been further assumed that the wattmeter circuit ratings correspond exactly with the magnitudes of the p.d. and current to be measured. This will seldom be the case in practice, and the instrument deflexion will be subject to further reduction accordingly. It is possible to operate the wattmeter circuits at their short-time ratings by using a press-button switch to connect them only when a reading is actually being taken; but this is not always a convenient arrangement in practice, where a continuous indication may be required over a lengthy period.

Single-Triode Protective Circuit

When switch S is closed, the p.d. across the wattmeter potential circuit will have the steady value E_0, which will depend on the no-signal anode current of the valve. When S is opened the p.d. will fall to E_{min} during the conducting part of the cycle and rise to E_{max} during the remainder of the cycle as indicated in Fig. 5. Thus, assuming each anode of the mercury-arc rectifier conducts for one-third of each cycle, the r.m.s. value of the p.d. may be about 15 per cent greater with S open than with S closed, and the wattmeter potential circuit must therefore not be operated at more than about 85 per cent of its full rating when S is closed, if it is to be capable of continuous operation with S open.

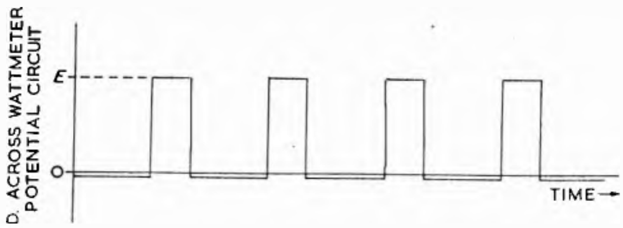

Fig. 6. Approximate waveform of p.d. across wattmeter potential circuit, with double-triode protective circuit

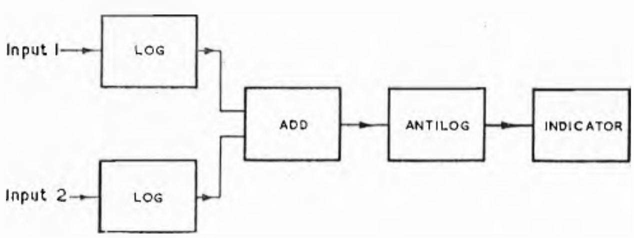

Fig. 7. Block diagram of logarithmic wattmeter

The maximum instrument deflexion obtainable with S open is therefore, from equation (5),

$$0{\cdot}85 \times 0{\cdot}577\, E_w I_w = 0{\cdot}490\, E_w I_w \quad \ldots\ldots \quad (6)$$

The value E_{min} to which the p.d. falls when S is opened cannot be allowed to go below about one half of E_0, because of the non-linearity of the triode characteristic at low values of anode current. Hence the wattmeter deflexion with S open cannot be less than half that with S closed, i.e.

$$0{\cdot}5 \times 0{\cdot}490\, E_w I_w = 0{\cdot}245\, E_w I_w \quad \ldots\ldots\ldots \quad (7)$$

Hence the usable operating range of wattmeter deflexion is

$$0{\cdot}490\, E_w I_w - 0{\cdot}245\, E_w I_w = 0{\cdot}245\, E_w I_w \quad \ldots\ldots \quad (8)$$

This is considerably less than the usable range with the diode-shunt circuit.

Double-Triode Protective Circuit

Referring to Fig. 3, when S is in position 1, the p.d. across the wattmeter potential circuit is zero, and when S is moved to position 2 the p.d. rises to E, the arc drop of the mercury-arc rectifier, during the conducting part of the cycle, and reverses to less than $0{\cdot}1E$ during the remainder of the cycle. Fig. 6 shows the idealized waveform of the p.d., and its r.m.s. value will be given by

$$\sqrt{\left[\frac{E^2 + 2(0{\cdot}1\,E)^2}{3}\right]} = 0{\cdot}583\, E \quad \ldots\ldots\ldots\ldots \quad (9)$$

The potential-circuit rating of the wattmeter, E_w, may therefore be 0·583 times the arc p.d., while conditions in the current circuit are the same as described above. Hence the maximum usable range of wattmeter deflexion will be, from equation (5).

$$\frac{0{\cdot}577\, E_w I_w}{0{\cdot}583} = 0{\cdot}990\, E_w I_w \quad \ldots\ldots\ldots\ldots \quad (10)$$

Electronic Wattmeter Used for Arc-Loss Measurement

The wattmeter consists essentially of a novel type of logarithmic multiplier giving an output signal whose instantaneous value is proportional to the product of two input signals, one of which is proportional to the anode current

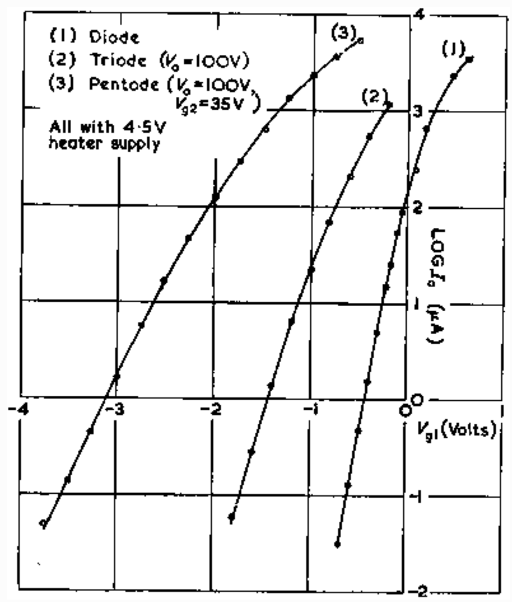

Fig. 8. Characteristics, for very small values of anode current, of Mullard EF42 connected as diode, triode and pentode

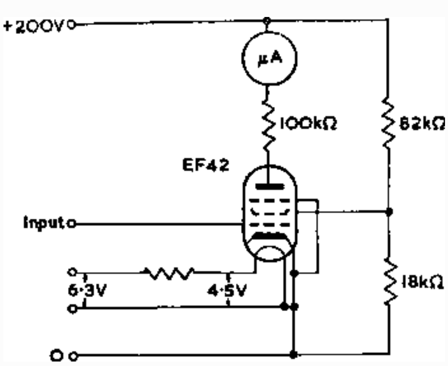

Fig. 9. Circuit of anti-logarithmic unit

and the other to the anode-to-cathode p.d. of the mercury-arc rectifier. The mean output of the multiplier is thus a measure of the mean power dissipated for the particular anode of the mercury-arc rectifier concerned, while the variation in power dissipation during the cycle can be investigated by applying the multiplier output to a cathode-ray oscillograph. Fig. 7 is a block diagram of the arrangement, while brief descriptions of the individual units, together with some experimental results obtained with the complete multiplier are given below.

Anti-Logarithmic Unit

It is well established that the initial velocities of the electrons emitted from the cathode of a thermionic valve are distributed in accordance with the Maxwell-Boltzmann distribution laws for the velocities of the particles of a perfect gas, and hence it can be deduced that the current, I, which passed the potential minimum in the cathode-anode space will be related to the potential E, at the minimum, as follows:

$$I \propto e^{BE} \qquad (11)$$

where e is the base of natural logarithms, and $B = q/kT$ where q is the electronic charge, k is Boltzmann's constant, and T is the absolute temperature of the electron gas.

I is equal to the anode current in a diode or triode valve, and is proportional to the anode current in a pentode valve, and although E is not normally the potential of any valve electrode, it will become the anode potential in a diode if the anode is made so negative that the potential minimum in the cathode-anode space is at the anode. Similarly if the control-grid potential in a triode or pentode is made sufficiently negative, E becomes proportional to the control grid potential. Fig. 8 shows experimental results obtained with a Mullard EF42 connected as diode, triode and pentode, and insofar as these graphs are rectilinear, the exponential relationships

$$I_A \propto e^{C_1 E_A} \quad \text{for the diode} \qquad (12)$$

$$I_A \propto e^{C_2 E_{G1}} \quad \text{for the triode} \qquad (13)$$

$$I_A \propto e^{C_3 E_{G1}} \quad \text{for the pentode} \qquad (14)$$

are obtained, where I_A and E_A are the anode current and potential respectively, E_{G1} is the control-grid potential, and C_1, C_2 and C_3 are constants. In the logarithmic multiplier the pentode arrangement is used, as this allows an anode load resistor to be added without appreciably affecting the anode current. The antilogarithmic unit of the multiplier simply consists of a pentode used in this way and is shown in Fig. 9. The input to the unit is applied to the control grid, and the output voltage obtained from the anode is then proportional to the anti-logarithm of the input. If only the mean value of the output of the anti-logarithmic unit is required, this can be read on a moving-coil microammeter connected so as to measure the anode current.

The resistor in series with the valve heater serves to reduce the heater supply voltage from 6·3V to 4·5V. The emission at the lower voltage is adequate since the cathode current is less than 1mA, but the lower cathode temperature results in reduced grid current and reduced variation in grid-cathode contact potential.

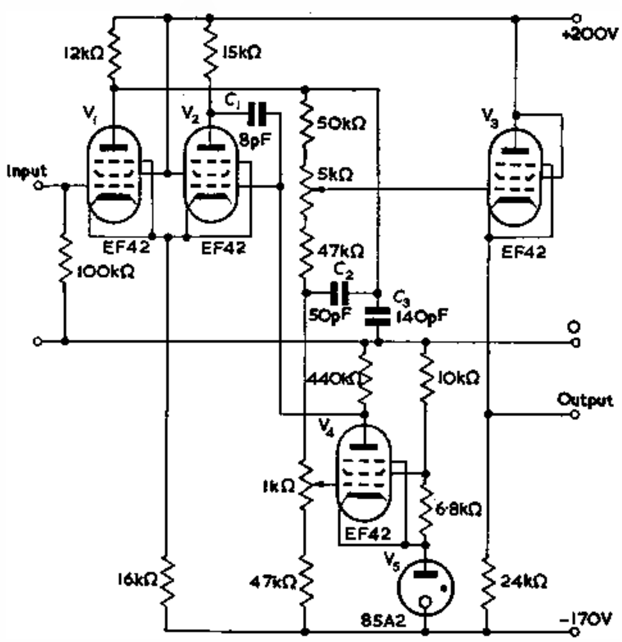

Fig. 10. Circuit of logarithmic unit

[The constant C_3 in expression (14) is dg_m/dI_A where g_m is the mutual conductance of the valve, and $1/C_3$ is the amount by which E_{G1} must be changed to alter I_A by the factor e.]

Logarithmic Unit

The circuit of the logarithmic unit is shown in Fig. 10.

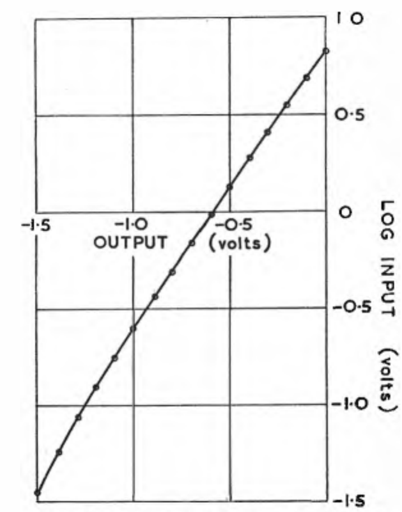

Fig. 11. Characteristic of logarithmic unit

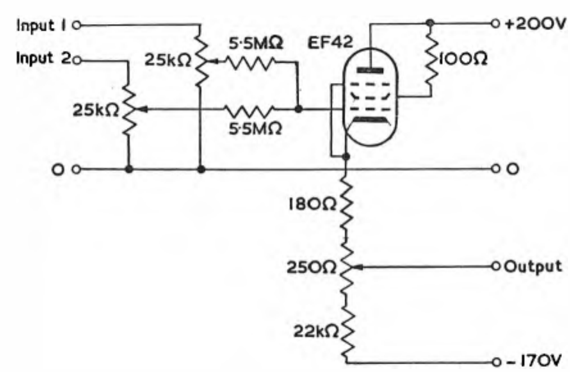

Fig. 12. Circuit of adding unit

V_1 is a linear amplifier which operates in conjunction with an anti-logarithmic negative-feedback path V_4 and V_2. V_4 functions in the same way as the anti-logarithmic unit described above and V_2 serves as a cathode-follower, coupling the anode of V_4 to the cathode of V_1. Capacitors $C_{1,2,3}$ are added to prevent self-oscillation.

The gain G, of any feedback amplifier whose gain without feedback is A, and the gain of whose feedback path is β, is given by

$$G = \frac{A}{1 - A\beta} \qquad (15)$$

In the anti-logarithmic feedback path used, β is a function of signal amplitude such that

$$\beta E_o = e^{C' E_o} \qquad (16)$$

where E_o is the output of the logarithmic unit and C' is a constant. Equation (15) may be rewritten

$$E_o/E_i = \frac{A}{1 - A\beta} \qquad (17)$$

where E_i is the input to the logarithmic unit. Combined with equation (16) this yields

$$C' E_o = \ln\left[-E_i(1 - (G/A))\right] \qquad (18)$$

The polarity of the signal is reversed in passing through the logarithmic unit, as this equation implies, and the circuit functions with E_i negative and E_o positive.

The ideal characteristic for the logarithmic unit would be

$$C'E_o = \ln(-E_i) \qquad (19)$$

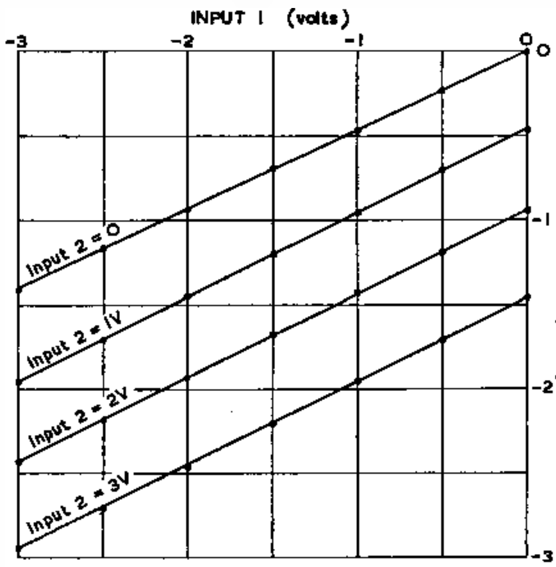

Fig. 13. Characteristics of adding unit

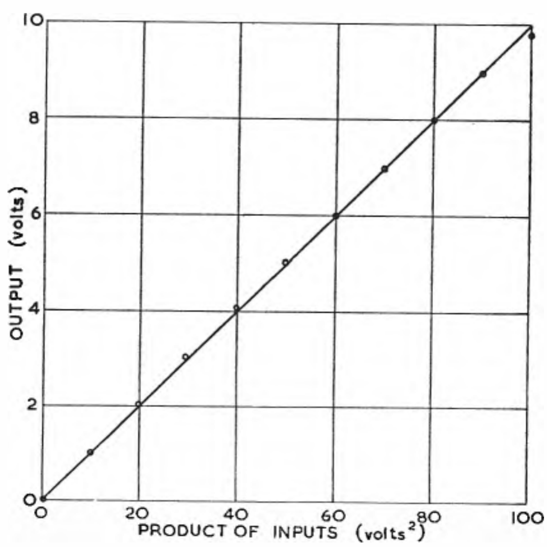

Fig. 14. D.C. characteristic of complete wattmeter

and if the error, referred to E_i, arising from the difference between the actual and the ideal characteristics, is not to exceed x per cent, it is necessary that

$$G/A \geqslant (x/100) \qquad (20)$$

$$\therefore \frac{1}{1 - A\beta} \geqslant (x/100) \qquad (21)$$

$$\therefore \frac{100 - x}{x} \geqslant -A\beta \qquad (22)$$

This relationship can only hold for input signals greater than some small but finite value, since β tends to zero as the magnitude of E_i is reduced. It is therefore possible to design the complete multiplier to respond with reasonable accuracy to input signal magnitudes within a limited range only, say, 100: 1, but this is sufficient for the present application.

Fig. 11 is a graph showing the results of a calibration test on the logarithmic unit, and it can be seen that this has the desired form over the range 1 per cent to 100 per cent of maximum input signal magnitude.

Adding Unit

The function of the adding unit is to provide an output

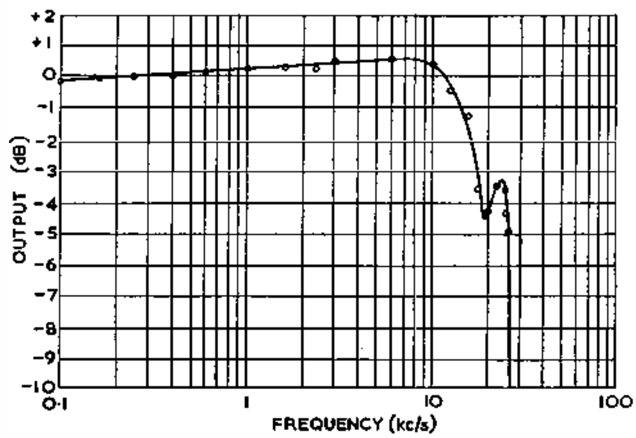

Fig. 15. Response of complete wattmeter as a function of frequency

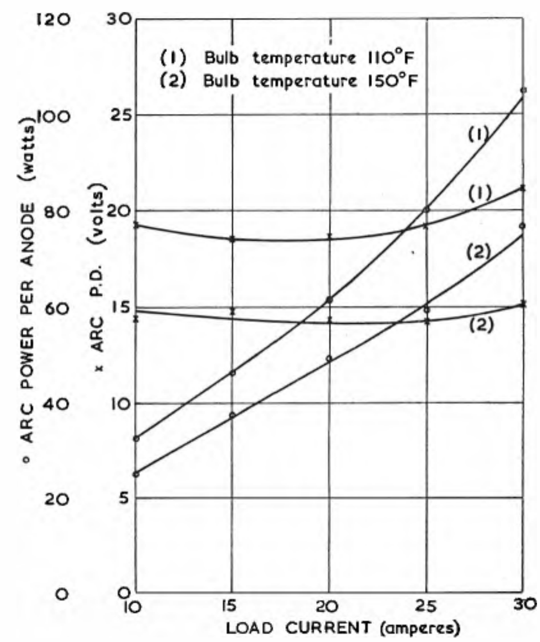

Fig. 16. Arc power loss and arc p.d. in a mercury-arc rectifier, plotted from results obtained with logarithmic wattmeter

proportional to the sum of the two independent signals fed to it from the logarithmic units. The circuit diagram of the unit is shown in Fig. 12. The two input signals are added by means of the resistance network shown, and a signal proportional to their sum applied to the control grid of the cathode-follower valve, from whose cathode circuit the output of the unit is obtained. Fig. 13 shows the results of some experimental tests on the unit.

Performance of Complete Multiplier

Fig. 14 shows the results of a d.c. test on the complete multiplier with one input held constant at the maximum rating and the other varied in steps from zero to the maximum rating. This shows an error which rises to about 2 per cent at the maximum output.

Fig. 15 shows the response of the multiplier plotted to a base of frequency. These results were obtained with both inputs receiving the same signal, which consisted of a direct voltage of one half the maximum input rating of the multiplier superposed on a sinusoidal voltage of peak value equal to the direct voltage. The bias component is necessary since the multiplier can handle inputs of only one polarity. Ideally the mean value of the output of the multiplier should be

$$0{\cdot}5^2 + \left(\frac{0{\cdot}5}{\sqrt{2}}\right)^2 = 0{\cdot}375$$

where unity corresponds to direct voltages of maximum

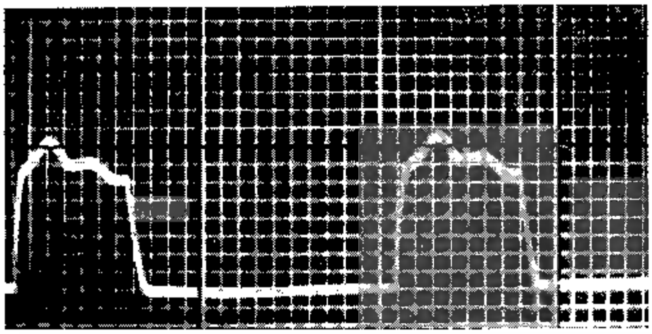

Fig. 17. Oscillogram of arc current for one anode of mercury-arc rectifier

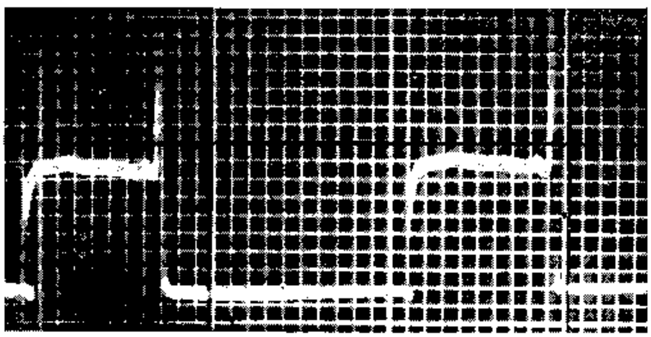

Fig. 18. Oscillogram of arc p.d. for one anode of mercury-arc rectifier

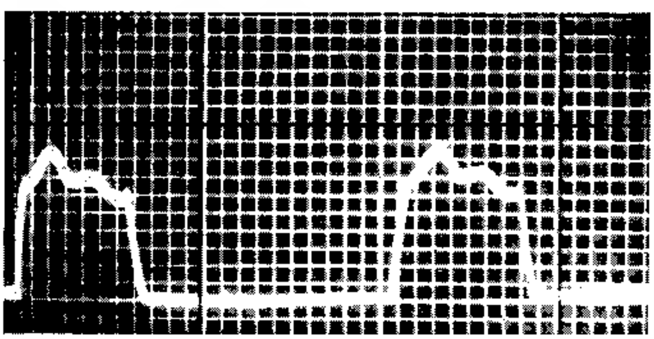

Fig. 19. Oscillogram of arc power for one anode of mercury-arc rectifier obtained with logarithmic wattmeter

rating applied to both inputs. Between zero and 13kc/s the greatest error is about 10 per cent at 6kc/s, but no special attention was given in designing the instrument to the performance above 1kc/s, and the maximum error to this frequency is about 5 per cent.

Fig. 16 shows results obtained when using the instrument to investigate the power dissipation in the arc of a 600V, 30A, 6-phase star-connected mercury-arc rectifier. The values of mean arc p.d. were obtained by dividing the values of arc power loss by the corresponding values of mean load current.

The logarithmic multiplier enables the waveform of arc power loss to be displayed on a cathode-ray oscilloscope, and Figs. 17, 18 and 19 show cathode-ray oscillograms of arc current, arc p.d., and arc power respectively for one anode of the rectifier when delivering full-load current.

When used for measurement of arc loss in a mercury arc rectifier, an auxiliary circuit is needed with the logarithmic multiplier, as with a dynamometer wattmeter, to protect it from damage by the high cathode-anode p.d. during the non-conducting part of the cycle. The conventional diode-shunt protective circuit was used for the above tests, but the diode current and the power dissipated in the resistor used to limit the diode current to a safe value are much less, due to the higher input impedance of the logarithmic multiplier, than they would be with a dynamometer wattmeter. The circuit is shown in Fig. 20. With switch S in position 1, the multiplier indicates the power dissipated in the shunt as well as that dissipated in the arc to the anode concerned. On moving the switch to position 2 the power dissipated in the shunt alone is measured, and the difference is then the power dissipated in the arc. This

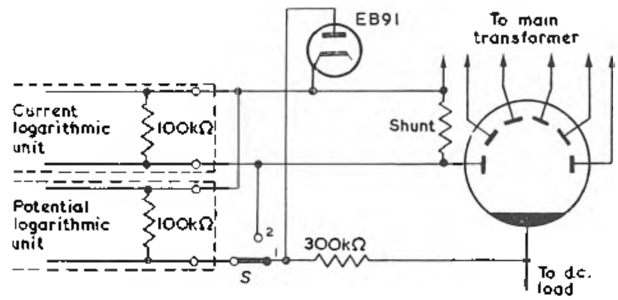

Fig. 20. Protective circuit used with logarithmic wattmeter

correction is required since about 7V peak is needed across the shunt to fully load the multiplier.

With the 6-phase star connexion used for the above tests each anode conducts for about one sixth of each cycle, and the value of the average product obtainable from the multiplier is then about one sixth of that obtainable with maximum d.c. signal applied to both inputs, assuming the potential divider and shunt are adjusted to fully load the multiplier inputs during the conducting period, but up to full-scale deflexion can be obtained in such cases by the use of a more sensitive moving-coil meter as indicating instrument. The possible percentage error in the reading which can result from departure from the ideal in the characteristics of the multiplier circuit or indicating instrument, apart from the frequency error discussed above, will be of the same order as when the multiplier is operated with maximum d.c. signals applied to both inputs.

The main difficulty encountered with the logarithmic multiplier has been that of drift of characteristic. It was found necessary to allow a considerable time after switching on for temperature readjustment to take place, and to feed the equipment from a thermistor-bridge stabilized a.c. supply before reasonable freedom from drift could be obtained.

Acknowledgment

The author wishes to record his indebtedness to the Sir James Caird's Travelling Scholarships Trust for funds, and to the Governors of Robert Gordon's Technical College, Aberdeen, for facilities which made possible the work described in this article.

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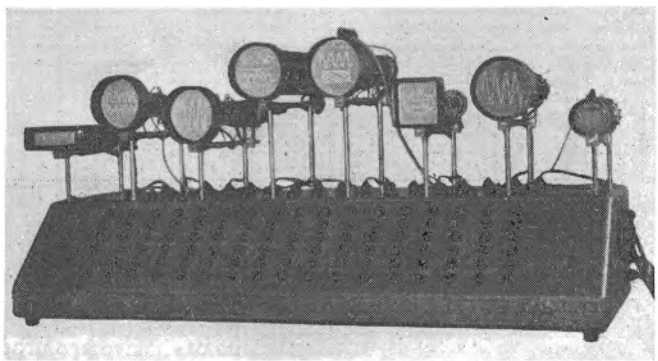
A Seventeen Channel Demonstration Oscilloscope

By C. J. Bargery*

A description is given of an equipment designed to demonstrate simultaneously up to seventeen separate c.r. tubes or the equivalent number of multi-gun tubes. The equipment is self-contained, incorporating power supplies, scan amplifiers and a waveform generator supplying a series of harmonically related waveforms with facilities for mixing.

A REQUIREMENT arose recently for an equipment capable of demonstrating to the best advantage a representative selection from a wide range of electrostatically deflected cathode-ray tubes. It was desirable that a large variety of tube types could be demonstrated simultaneously and that the equipment should be able to accommodate new types which would from time to time be introduced.

It was decided that it would be reasonable to operate up to eight cathode-ray tubes simultaneously on one piece of equipment. However, as these eight tubes would at any time consist of an unpredictable combination of single, double and four gun-tubes, approximately sixteen separate channels would be needed to operate them. The final design incorporated seventeen channels, as for practical reasons, it was necessary to choose an odd number.



The complete instrument

The following features have formed the main basis of the design:

(a) All tubes are operated at 4kV e.h.t. to produce traces of adequate brightness for viewing in fully lit rooms. Additional potentials up to a maximum of 16kV are available to operate tubes with post deflexion acceleration.

(b) Controls are made available to enable all tubes to be examined qualitatively in respect of brightness, spot size, evenness of focus, orthogonality of axes, degree of scan distortions, (trapezium etc.) and maximum useable screen area.

(c) Each channel incorporates an independent pair of identical scan amplifiers. The output from each amplifier is sufficient to fully scan the least sensitive tubes made, and a high degree of push-pull balance is maintained.

(d) All tubes operate 'in the open' rather than in a cabinet with only the faces showing, so that if necessary they may be examined physically.

In order to preserve the greatest possible flexibility the equipment is designed as if it were to operate seventeen independent single-gun cathode-ray tubes of any type. The output of each channel is brought out to sockets on the top of the equipment, so that in use, the connexions to a single-gun tube are plugged directly into any one channel and a multi-gun tube into two or more adjacent channels. Electrodes which are internally connected in multi-gun tubes (e.g. anodes No. 1 or 3 in four-gun tubes) are supplied from any channel, the corresponding connexion to the remaining channels being left blank.

A waveform generator which supplies a series of harmonically related outputs forms an integral part of the equipment, and provision is made to feed any waveform to any one or more of the deflexion amplifiers. The range of waveforms available are of relatively low frequency and are free from any form of distortion which might appear to an observer as being due to a defect in the tube.

To make the equipment reasonably portable, it is constructed as one main display unit, with two separate power supply units. In use, this arrangement enables the power units to be placed well away from the tubes and so removes the possibility of magnetic interference from the power transformers.

The C.R.T. Display Unit

This unit consists of a shallow cabinet 5ft 6in long by 12in high with a sloping front panel carrying all the controls. The cathode-ray tubes are mounted above, on adjustable brackets. The unit may be divided into the following four main sections.

The C.R.T. Supply Networks

It would obviously have been possible to operate seventeen cathode-ray tubes from one stabilized power supply by providing each gun with its own bleeder network incorporating series brightness and focus controls as in conventional oscilloscope practice. This arrangement, however, is extremely inflexible and uneconomical when large numbers of tubes are involved. Bleeder networks become very bulky as they have to be made up from high voltage resistors (or large numbers of low voltage ones) and have to be supplied with and dissipate considerable wasted power.

In the equipment being described, all guns are operated with their cathodes connected directly to a 4kV stabilized negative e.h.t. supply. One 2mA bleeder resistance chain situated in the main power unit is used for all tubes, and the total e.h.t. load is less than 5mA when they are working at normal brightness. The bleeder chain is tapped at its centre point to feed a rail +2kV with respect to the cathodes. All anodes No. 1 are connected directly to this rail, but as these electrodes draw negligible current, no interaction between tubes is produced.

As tubes with widely differing focus (anode No. 2) voltages are to be connected to the common system, the focus

* 20th Century Electronics Ltd.

controls are supplied from a separate power supply of 850V at 10mA. All components of this supply are insulated to 4kV and the negative terminal is connected to the -4kV (cathode) rail. The positive terminal feeds a bus bar which runs adjacent to a row of seventeen focus potentiometers (each 500kΩ), and resistors to a total value of 1·5MΩ (including the potentiometer) are connected from this bar and from the negative e.h.t. bus bar to either end of each potentiometer. By changing these resistors (when the equipment is set up) the limits of the voltage excursion of the individual focus controls can be varied to suit any combination of tubes. The bias for the tubes is derived in a similar manner from seventeen potentiometers which are fed in parallel from a stabilized source of 85V. This supply is insulated to 4kV and has its positive terminal connected to the -4kV rail.

A variety of waveforms, described later, is available for display on any gun, and facilities are provided to enable a flyback blanking pulse which is supplied from the waveform generated to be switched on or off from any grid. Fig. 1 gives details of the circuit associated with the c.r.t. guns. It will be seen that a cathode-follower and d.c. restorer at cathode potential are inserted between the coupling capacitor C_1 and the blanking pulse rail. This circuit is necessary for three reasons:

(a) Capacitor C_1 must be at least 4kV working, pass negligible leakage current and have low capacitance to earth. A practical limit is therefore set to its maximum value, and it is impossible to make the coupling time-constant C_1R_3 sufficiently long to prevent differentiation of the pulse.

(b) The load on the blanking pulse rail will vary over wide limits for different combinations of waveforms being displayed. The rail must, therefore, be fed from a low impedance source.

(c) As each type of cathode-ray tube has a different cut-off voltage, it is not possible to d.c. couple the blanking pulse rail to the c.r.t. grids, and as the grid resistors must be restricted to 1MΩ, it is necessary to limit the amplitude of the pulse in order to reduce positive overshoot to an absolute minimum.

The cathode-follower circuit is constructed on a small sub-chassis mounted at one end of the high voltage distribution panel. This panel, which consists of a long strip of insulating material, is mounted along the upper back part of the unit. It carries seventeen ceramic octal sockets, and seventeen semi-rotary switches (S_1). Each octal socket, which forms the output for all potentials necessary for the operation of one c.r.t. gun, is connected to a series of bus bars running the length of the unit. The three heater bars (0-4-6·3V a.c.) are fed from a 20A transformer situated at one end of the equipment, but all other potentials are obtained from the separate power supply unit. The seventeen blanking pulse switches are deeply recessed within insulating bushes and are intended to be operated with an insulated screwdriver only when the equipment is being set up. The ceramic valve sockets have been modified by having the outer metal mounting ring and two contacts (No. 4 and 6) removed to enable pin No. 5 to be operated at 2kV with respect to the remainder. The sockets are cemented into the distribution panel with epoxide resin and the whole panel is mounted on insulators to withstand 4kV to earth. The plugs from the c.r.t's. are inserted into the octal sockets through clearance holes of adequate size in the top panel of the unit.

The Deflexion Amplifiers

Each amplifier (Fig. 2) consists of two pentodes in a d.c. coupled see-saw circuit, with the d.c. level referred to a -150V stabilized line. The anodes are fed from a $+550$V rail and the screens from $+250$V stabilized supply. The

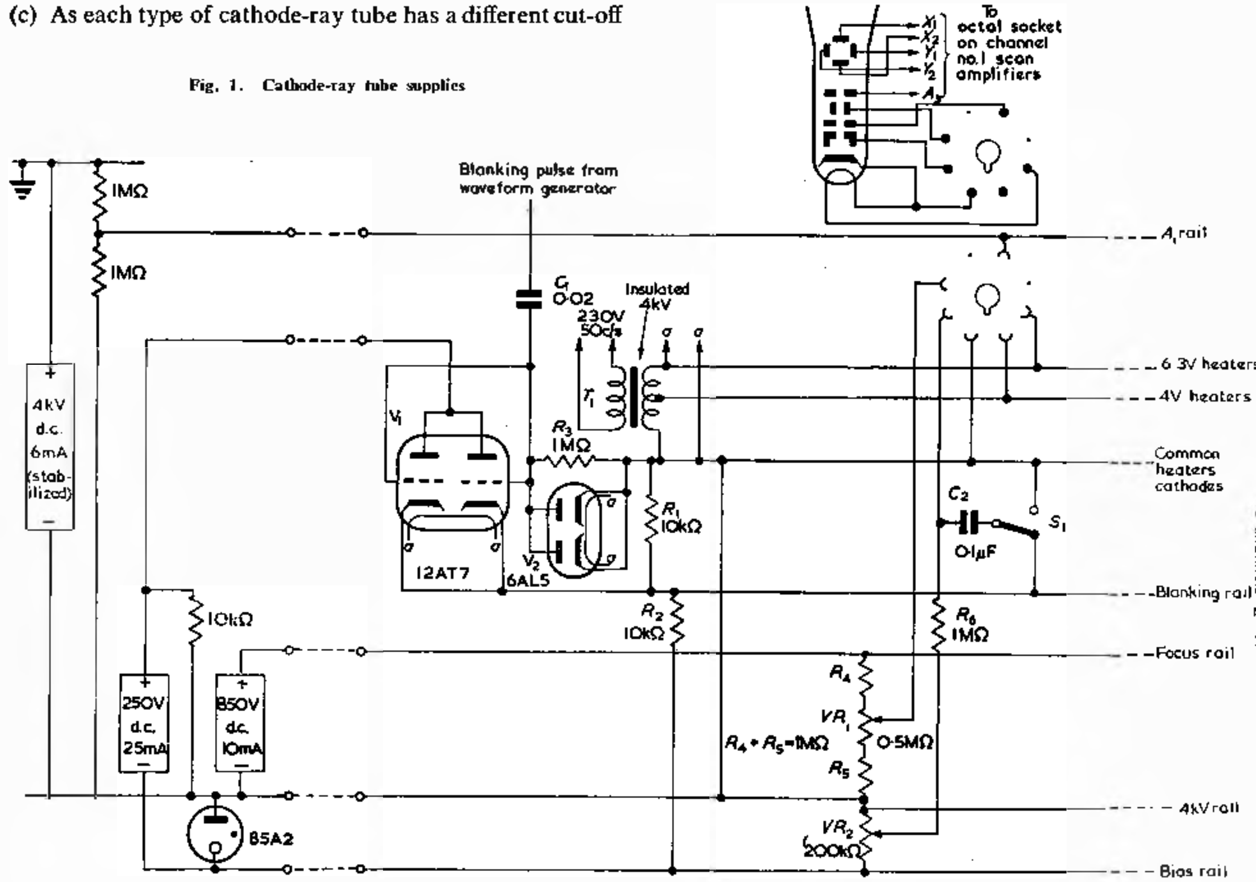


Fig. 1. Cathode-ray tube supplies

values of $R_{15,16,20}$ are chosen so that the mean potential of the c.r.t. deflector plates which are directly coupled to the anodes of V_3 and V_4 is equal to +250V at all times. The final anodes of all the c.r.t. guns are connected to the +250V rail, thus accurate balance is always maintained in all deflexion systems. D.C. shift is introduced via the third arms of the input see-saws R_8. $R_{7,9}$ are chosen such that the full output of ±400V between anodes is obtained with an input across the gain control VR_3 of ±20V. The preset resistors VR_5 in the common cathode circuits are adjusted so that in the absence of an input signal the grid of V_3 is at earth potential. Under this condition R_7 carries no direct current and adjustment of the gain controls produce no d.c. shift.

All the resistors in the see-saw circuits are high stability to prevent drift of the mean deflector plate potentials. This

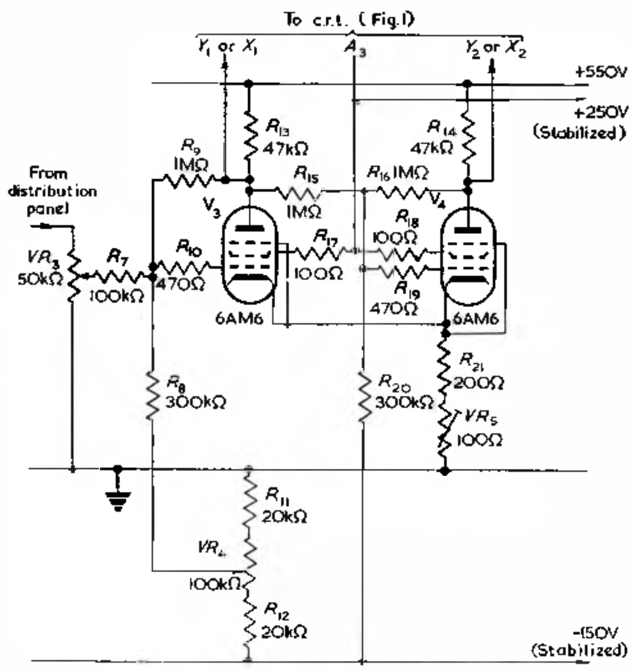


Fig. 2. The deflexion amplifiers

ment at any time be required to work in an appreciable 50c/s magnetic field. The twelve basic waveforms available are as follows:

(a) 50c/s sine (b) 50c/s cosine (c) 50c/s linear sawtooth
(d) 100c/s sine (e) 150c/s sine (f) 200c/s sine
(g) 250c/s sine (h) 250c/s square wave (1 : 1)
(i) 250c/s differentiated square wave (k) 1 250c/s sine
(l) 1 250c/s linear sawtooth
(m) variable 0·1-1·0c/s linear sawtooth.

Each output is low impedance from a cathode-follower and approximately 200V peak-to-peak.

A 70-0-70V r.m.s. 50c/s transformer T_2 feeds seven *RC* phase shift networks in parallel. The outputs from two of these networks (connected in reverse) are used after passing

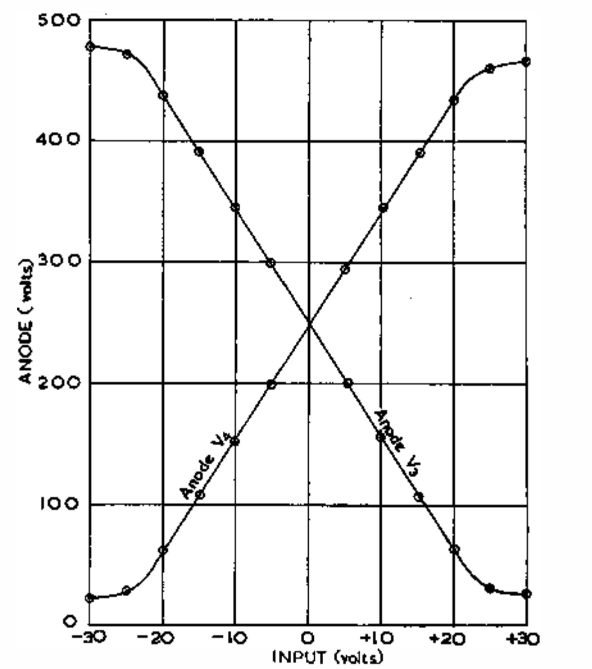


Fig. 3. Transfer characteristic of deflexion amplifiers

precaution, together with the high degree of balance obtained from the amplifiers (Fig. 3) provides the conditions for optimum focus of the c.r.t.'s.

Seventeen sub-chassis which each carry two identical amplifiers are arranged along the front of the display unit with the two shift and two amplitude controls linked to the front panel by flexible shafts. These, together with brightness and focus form the six controls available for each c.r.t. gun. An international octal socket is mounted on the top face of each amplifier and is accessible from the top of the unit. This socket carries the output connexions from each pair of amplifiers for the tube deflector plates, and a connexion to the +250V rail for anode No. 3 of the c.r.t. The width of the amplifier chassis is such that the output sockets are directly in front of the corresponding high voltage sockets at the rear of the top panel, and directly above the relative vertical row of controls.

The Waveform Generator

To provide the greatest variety of patterns for display on the c.r.t.'s, eleven of the basic waveforms provided by the generator bear a fixed harmonic relationship to each other, and most are independently variable in phase. The lowest frequency is 50c/s synchronized to the mains, as this prevents the possibility of spurious beats should the equip-

a single stage harmonic filter to provide outputs (a) and (b). The third variable phase 50c/s is used to generate the main linear sawtooth. The sweep generator (Fig. 4) is based on the well-known circuit by Puckle[1], and although it uses one more valve than would a transitron, has the advantage of being free from any annoying 'Miller step.' By using a type 6AM6 as the constant current pentode, good linearity is obtained. A blanking pulse is taken from the trigger amplifier and after being inverted by V_{6b}, (which also steepens the leading edge) is fed to the cathode follower input of Fig. 1.

The outputs from the remaining four phase-shift circuits drive four frequency multipliers to provide 100, 150, 200 and 250c/s sine waves. The 250c/s output subsequently drives an additional multiplier to provide a 1 250c/s sine wave. All the multiplier circuits are identical except for component values, and only one need be described.

The circuit used (Fig. 5) consists of a symmetrical multivibrator which is very lightly locked to the fundamental, followed by a harmonic filter consisting of a parallel-T *RC* network in the feedback loop of a pentode amplifier[2]. When the natural frequency of the multivibrator is adjusted to be very close to the harmonic required, very little phase shift occurs during synchronizing, and a very pure sine wave is obtained from the filter. In this application, the

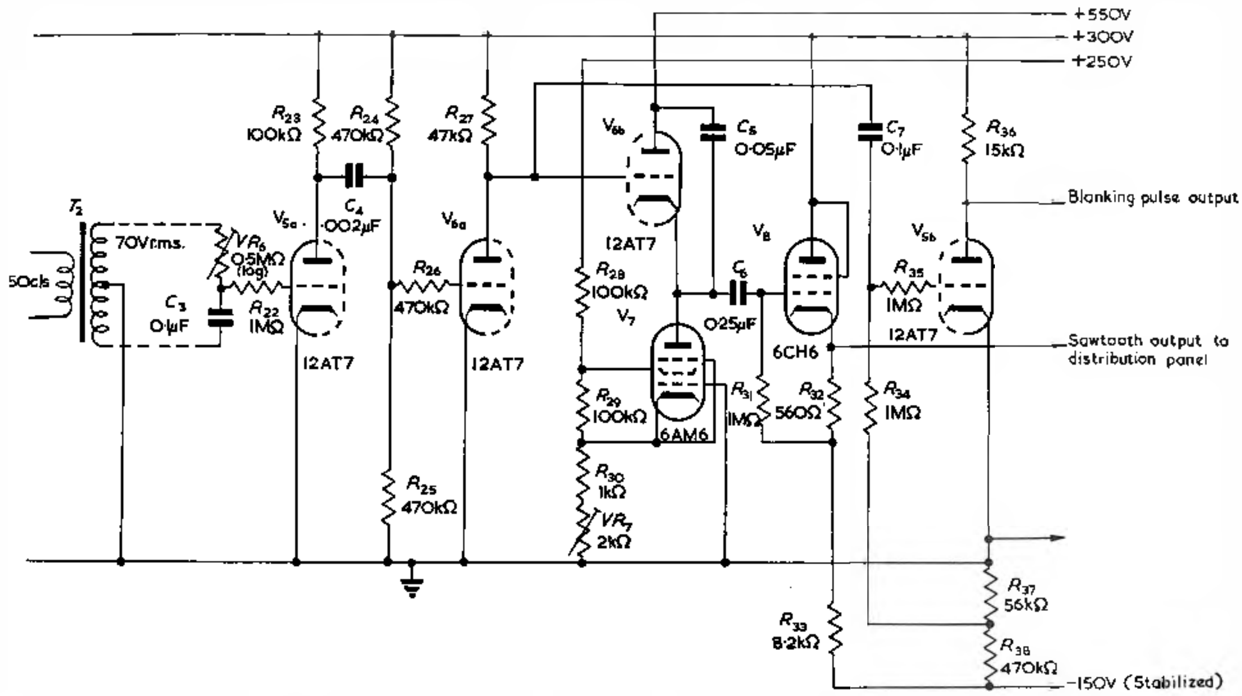


Fig. 4. Circuit of sweep voltage generator

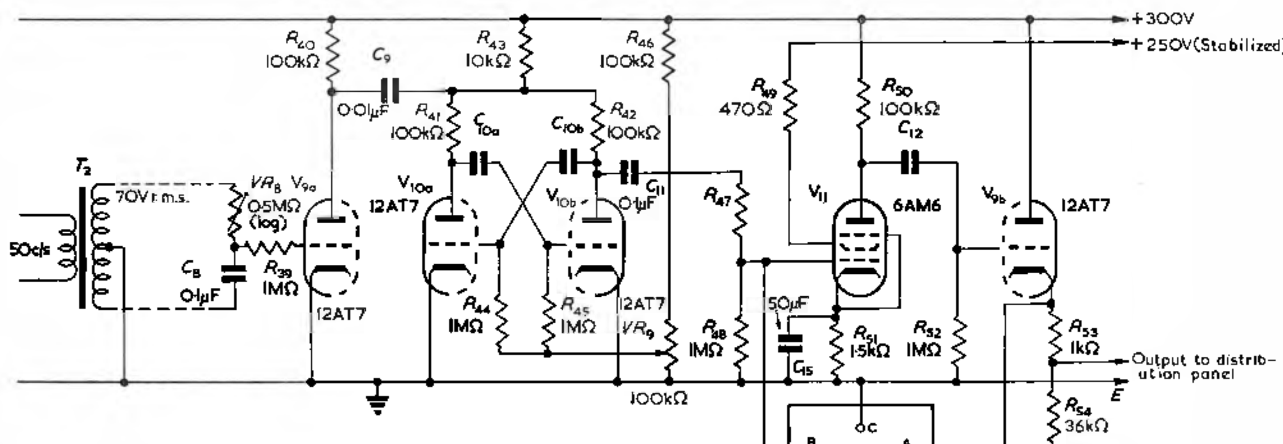
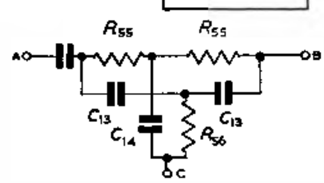


Fig. 5. Circuit of frequency multipliers

F (c/s)	C_{10} (μF)	C_{12} (μF)	C_{13} (μF)	C_{14} (μF)	R_{55} (kΩ)	R_{56} (kΩ)	R_{47} (MΩ)
100	0·004	0·25	0·02	0·04	75	40	3·3
150	0·003	0·25	0·01	0·02	90	56	4·7
200	0·002	0·1	0·01	0·02	75	45	4·7
250	0·002	0·1	0·01	0·02	56	33	1 5
1250	0·0005	0·1	0·002	0·004	56	33	2·0



presence of as little as 2 per cent of the fundamental can produce noticeable amplitude modulation of the required harmonic, and as the modulation is synchronized with the time-base, it can, depending on phase appear on the display as trapezium, pincushion or barrel distortion in the c.r.t's. The first triode section V_{9a} squares the input sine wave, giving edges sufficiently steep to provide a synchronizing pulse via the differentiating circuit C_9R_{43}. The multivibrator $V_{10a,10b}$ is a conventional a.c. coupled symmetrical circuit with the potentiometer VR_9 providing preset control of the natural frequency. The frequency selective stage consists of a high gain amplifier (V_{11}) followed by a cathode-follower output stage (V_{9b}), with a parallel-T network connected between output and input. This arrangement provides the necessary low impedance input to the network and enables the amplifier to be operated with a high anode load, thereby maintaining a high loop gain and, therefore, a high selectivity.

The potentiometer formed by $R_{47,48}$ attenuates the input to the frequency selective stage, and R_{47} is chosen in each case so that the output is approximately 200V peak-to-peak. The resistor R_{53} in the cathode circuit of V_{9b} offsets the bias voltage so that if there were no input signal to V_{11} the output terminal would be at earth potential. This is necessary because the scan amplifiers are directly coupled and if d.c. exists at the input an undesirable shift occurs as the amplitude controls are operated.

The output from the 250c/s multiplier stage, in addition to being available to feed the scan amplifiers and to synchronize the 1 250c/s multiplier, drives a triode squaring valve with cathode-follower output and limiter. Positive feedback is provided between the output and the triode cathode in order to steepen the edges of the output square wave. This square wave is available for displaying and also feeds a differentiating circuit and cathode-follower to provide a further output waveform.

The sine wave output from the 1 250c/s multiplier is available for display and also drives a linear sweep circuit which is identical to Fig. 4 except for component values and the omission of the blanking pulse amplifier. The output from this circuit, when used in combination with a 50c/s linear sweep gives a 25-line synchronized raster.

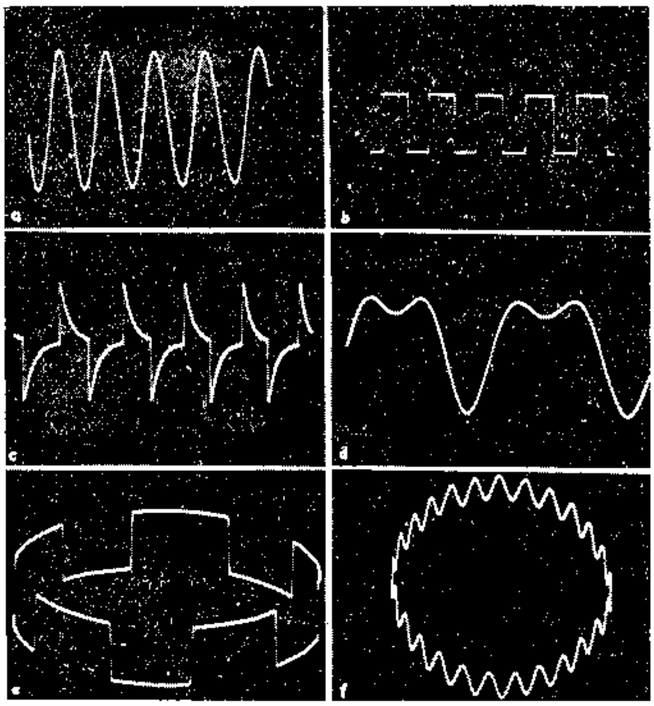


Fig. 6. Typical display waveforms
(a) 250c/s sine wave
(c) 250c/s square wave
(b) Differentiated square-wave 250c/s
(d) Mixed 100 + 200c/s sine waves
(e) Mixed outputs
(f) Mixed outputs

The above circuits provide eleven outputs (excluding the blanking pulse) which are all harmonically related. The twelfth output is from a very low frequency linear sweep circuit (Puckle[1]) which is similar to Fig. 4 except that it is self-triggering. The sweep frequency is variable between 1·0 and 0·1c/s and is unrelated to any other waveform. The output cathode-follower is d.c.-coupled to the charging capacitor and as it is unlikely that the output will ever be required to be mixed with any other waveform (see next section) the peak-to-peak amplitude is restricted to 40V.

The complete waveform generator is constructed on one chassis which is situated at the rear of the equipment. All components are directly available for service by removing the back panel and all valves are replaceable from above. Thirteen preset controls (seven phase, five synchronizing and one v.l.f. sweep) are linked to the front panel (for screwdriver adjustment) via flexible shafts which run from back to front at the bottom of the unit.

The Waveform Distribution Panel

A removable panel at the left-hand end of the unit gives access to the waveform distribution and mixer panel. The twelve outputs from the waveform generator are connected via coaxial leads to twelve groups of coaxial outlet sockets. The sockets in each group are connected in parallel and vary in number according to the probable demand for each waveform (e.g. 50c/s linear sweep, 12 outlets; 50c/s sine wave, 6 outlets; 250c/s sine wave, 3 outlets, etc.).

The coaxial leads to the inputs of the seventeen pairs of scan amplifiers are connected sequentially to two rows of coaxial sockets (one for horizontal and one for vertical), one row at each end of the panel. Immediately parallel with each row, and close to them are two more identical rows of sockets. Each socket in the second row is linked to the corresponding one in the first by a 150kΩ resistor. When viewed from the 2nd sockets, these resistors, together with the input resistance of the amplifiers form attenuators reducing the input to the amplifiers by 5:1. A series of coaxial jumper links are provided so that any output (except v.l.f.) from the waveform generator (200V peak-to-peak) may be fed into any amplifier using sockets in the 2nd rows as inputs and the amplifiers will receive full drive. The v.l.f. waveform is fed directly to the amplifiers via sockets in the first rows.

In the centre of the distribution panel are five sockets which are each connected to a number of surrounding sockets by resistors of varying value (150kΩ to 4·7MΩ). By connecting either of these five sockets to the direct input of any amplifier, these resistors, together with the input resistance of the amplifier form a resistance mixing network. Two or more waveforms fed into the surrounding sockets may thus be mixed with varying relative amplitudes to form a large variety of patterns.

General

The cathode-ray tubes are supported by eight demountable brackets which fit into the top of the unit and are independently adjustable in height from 4 to 16in above the top panel. Each separate gun is treated as a separate c.r.t. and the leads from the tube bases and side contacts terminate in plugs which fit into the octal sockets in the top panel.

In addition to the facilities described, three high voltage resistor chains are mounted on a long insulated panel immediately below the top panel and taps are brought out via high voltage sockets. These resistor chains are fed from an auxiliary power unit to provide the necessary additional positive potentials to operate tubes with post deflexion acceleration.

Power Supply Units

The power supplies for the complete display are provided from one main power unit, which is connected by two multiway cables. The design of this unit follows normal practice and need not be described.

The post deflexion acceleration potentials are supplied from a separate smaller unit consisting of a 3·5kV r.m.s. 50c/s transformer which feeds a Cockroft-Walton voltage multiplier made up of pencil type selenium rectifiers. Outputs are taken from the multiplier at approximately 4·3, 8 and 11·5kV and after being smoothed separately, are fed by three high voltage leads to the main display unit.

Acknowledgements

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A Time Scaler For Nuclear Track Length Measurements

By Kurt Enslein*

The instrument described in this article generates, under semi-manual control, the distribution function of nuclear track lengths by converting lengths to time intervals. It is based on straightforward circuits using glow tubes and coincidence circuits for timing with electro-mechanical registers for indication. The resolving time of the present instrument is of the order of 100msec, but can be decreased to any desired value by the incorporation of a suitable time-base.

THIS instrument was designed to measure the length of nuclear particle tracks on emulsions. This is accomplished by moving the track on a microscope stage at a uniform rate, and noting the beginning and end of each track. In this way the track lengths are converted to and measured as a series of time intervals. These intervals are measured in a slow speed counter; a series of outputs are taken from each stage of the counter so that an output tenth output is fed to the $\times$ 1 000 decade. As long as the key is depressed, pulses will be available from the output of the gate and the decades will continue to count. The ten outputs from each decade are connected to a set of selector switches. The selected outputs are connected to a set of four-fold coincidence circuits. When any one of these four-fold coincidence circuits is triggered, the register which is associated with it advances by one count. The selector

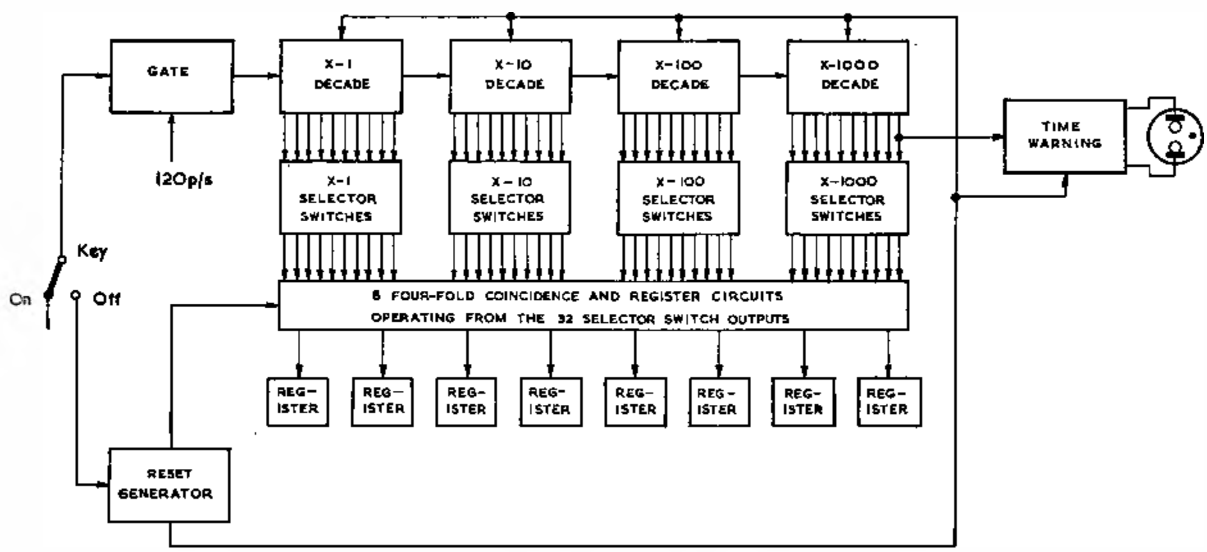


Fig. 1. Block diagram of complete equipment

channel register records one count on each occasion that the track 'time' equals or exceeds the setting of the input controls for the channel concerned. By choosing suitable settings for the input controls of the eight channels available, it is possible to obtain from the channel registers a distribution function for the track lengths.

General Principle of Operation

The instrument described is essentially a slow-speed scaler with a resolving time of the order of 8msec, using rectified 600c/s line voltage as timing information†. Referring to Fig. 1, the normal position of the key is in the 'Off' position; it is operated as one end of the track passes a reference mark in the microscope and released when the other end passes. When the key is moved to the 'On' position, the gate circuit is energized. The gate then allows 120c/s pulses to be fed through to the first decade counter. The tenth output from the first decade is fed to the second decade, whose tenth output is fed to the third, whose

switches permit the selection of any of the 9 999 combinations which are obtainable from the outputs of the four decades and it is therefore possible to set each of the eight channels included in the equipment to respond to any of the combinations. When the $\times$ 1 000 decade leaves its ninth position, (on the count of 10 000) a pulse is sent to a time-warning circuit which lights a neon lamp to indicate the fact that timing will now start again from zero.

When the key is returned from the 'On' to the 'Off' position, a negative pulse is produced and fed to the reset generator. This first disables the register circuits and then produces a reset pulse for the four decade counters and the overtime warning circuit, resetting these circuits back to zero.

The system then consists of a time scaler with eight channels operating at a basic rate of 120p/sec. There might be some question concerning the accuracy of this scaler in that the information is derived from the power line. Investigation with the local power company showed that for essentially 364 days of the year the accuracy of the 60c/s line is determined by a master clock which is crystal-controlled and one would expect long-term accuracies, of the order of 0·01 per cent.

* *University of Rochester, New York.*

† *A mains frequency of 50c/s instead of 60c/s will cause the smallest time increment to be 100 msec instead of 83·4 msec, no other changes will be entailed.*

Gate and Input Circuit

On the left-hand side of Fig. 2 are shown V_1 and V_{2a}, as well as the input key, which together form the gate and input circuit. This consists of a 5915 dual control-grid valve with both grids normally biased beyond cut-off. When the key contact is grounded, the bias is removed from g_3. The 120c/s input pulses are large enough to drive g_3 into conduction. Negative output pulses are then produced. These amplifier and inverted. The inverted output drives the next monostable multivibrator through V_{6a}, the m.m.v. then operating the next glow-counter decade. Each of the 10 outputs of V_4 is brought to one stator contact of eight, ten-position, single-pole selector switches, and the same is done for the other three decade counters. In other words, there are thus thirty-two, ten-position, single-pole selector switches in this system. The rotors of these selector switches

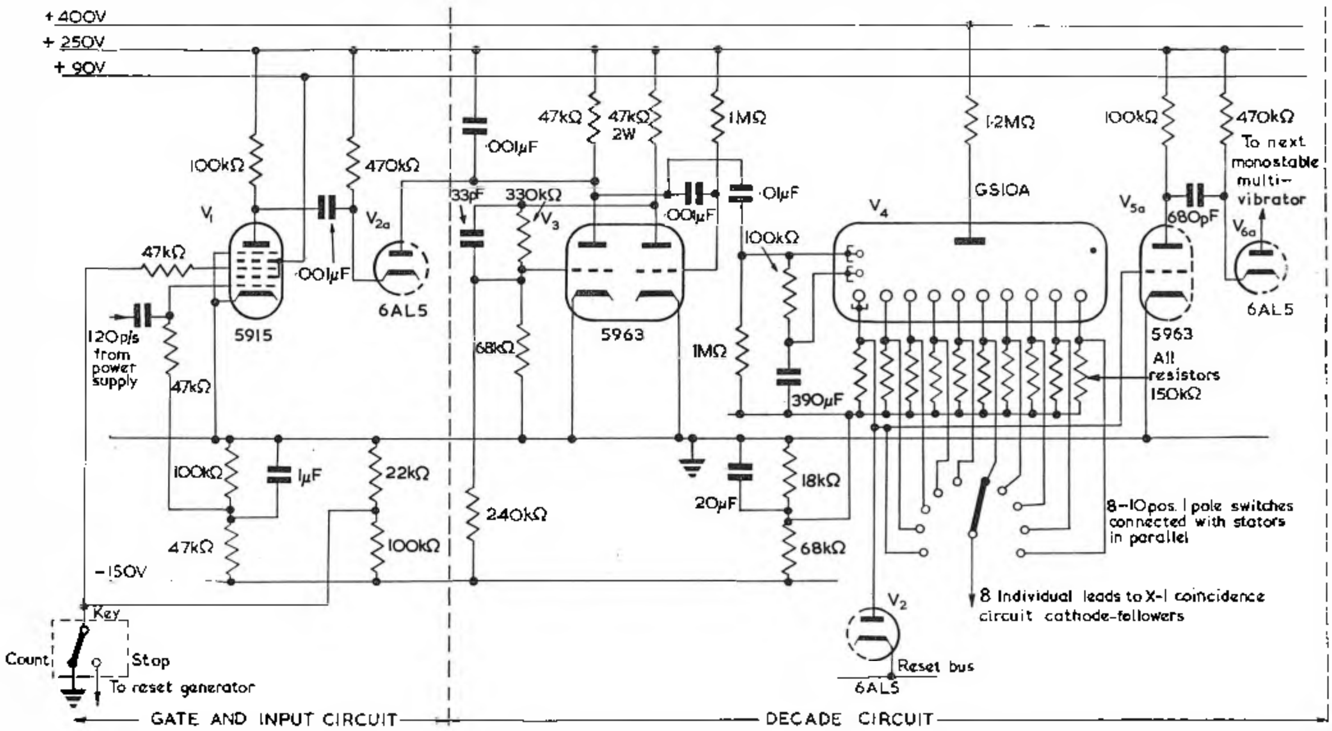
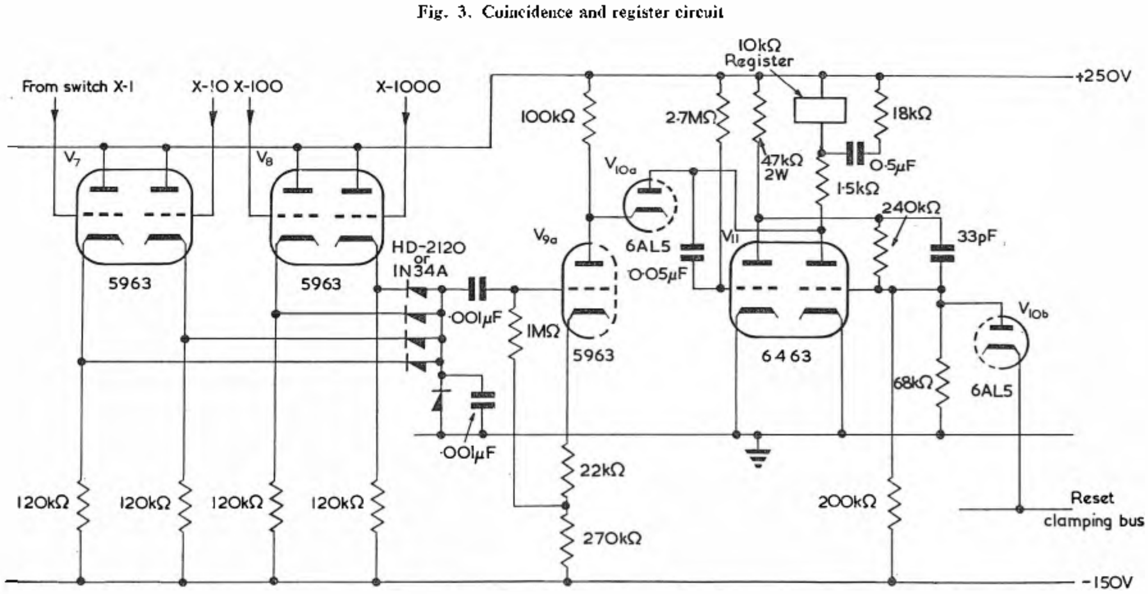


Fig. 2. Gate and input circuit and decade circuit

Fig. 3. Coincidence and register circuit



pulses are fed through the cathode of V_{1a} to the first decade counter circuit which appears on the right-hand side of Fig. 2. The negative pulses are used to drive a monostable multivibrator, V_3, the output of which is used to drive a GS10A glow counter tube, V_4. The network between the two guide electrodes of the GS10A is a phase-shifting network. The tenth output from V_4 is then fed to a triode are connected to the cathode-followers which feed the coincidence circuits which are now described.

Coincidence and Register Circuit

Referring to Fig. 3, it can be seen that the output from each ten-position selector switch is fed to a cathode-follower. The output of each one of these cathode-followers drives one input of a four-fold coincidence circuit. One each of each group of cathode-followers is fed from the first, second, third and fourth decade so that counts up to 9 999 can be selected.

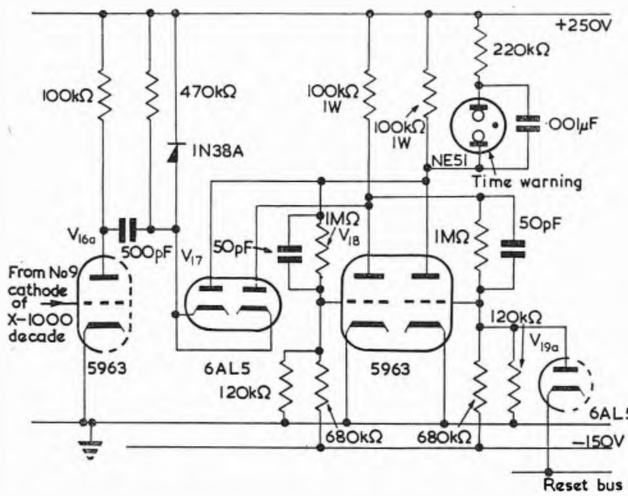


Fig. 4. Warning circuit

the positive part of it is absorbed by the 1N38A diode. The negative pulse is then fed through V_{17} to trigger V_{18} which is an Eccles-Jordan flip-flop, so that a NE51 warning light will be switched on.

Reset Generator

Referring to Fig. 5, the reset generator is divided into two parts. In the upper part of the figure it can be seen that when the input key is moved to the 'stop' contact, a negative voltage will be impressed on the output lead. This negative step is differentiated and fed through V_{11a} to V_{12}, which is a m.m.v. When this multivibrator has gone through its quasi-stable state, a negative output is obtained from its right-hand anode. This pulse is differentiated and sent to V_{11b}. The output of V_{11b} is used to trigger V_{13} which is another m.m.v. The output is fed to the reset bus-bar which, through all the diodes shown on the previous diagrams, resets the decade counters and the time warning circuit.

From the lower part of Fig. 5, it can be seen that the negative pulse which is generated at the time the key is released is fed to V_{14a} which is diode-connected. The output of V_{14a} is used to trigger V_{15}, which is a m.m.v. The output of V_{15} is used to drive V_{14b}, to provide a low impedance pulse to clamp the 6463 register circuits in such a manner that no additional count can be recorded during the reset-

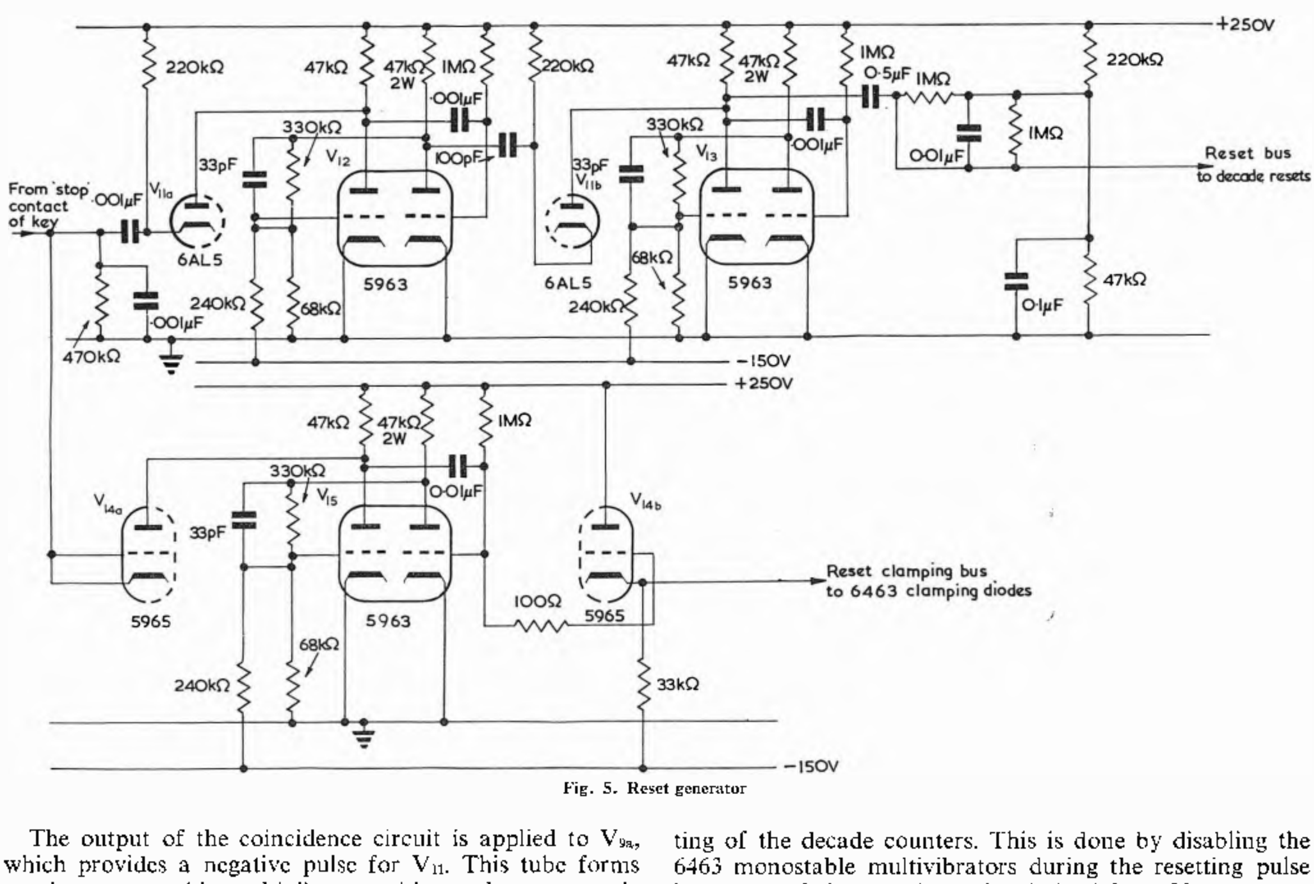


Fig. 5. Reset generator

The output of the coincidence circuit is applied to V_{9a}, which provides a negative pulse for V_{11}. This tube forms another monostable multivibrator with an electromagnetic register as its load. Each time a pulse is obtained from the coincidence circuit, this m.m.v. will go through its cycle and the register will advance by one digit.

Time Warning Circuit

Referring to Fig. 4, the input to V_{16a} is obtained from the ninth cathode of the × 1 000 decade. When the cathode glow leaves the ninth cathode, a negative pulse is produced from this cathode. This will cut off V_{16a} and produce a positive pulse at its output. This pulse is differentiated and

ting of the decade counters. This is done by disabling the 6463 monostable multivibrators during the resetting pulse by means of the negative pulse derived from V_{14b}.

Power Supply

Fig. 6 shows the power supply for the time scaler. This supply is of a very conventional type and supplies the various levels required in the rest of the equipment. The 120c/s pulses are derived from a full-wave rectifier.

Detailed Sequence of Operation

Fig. 7 shows a timing diagram of the sequence of operations in the scaler. On line 1 is shown the voltage on g_1

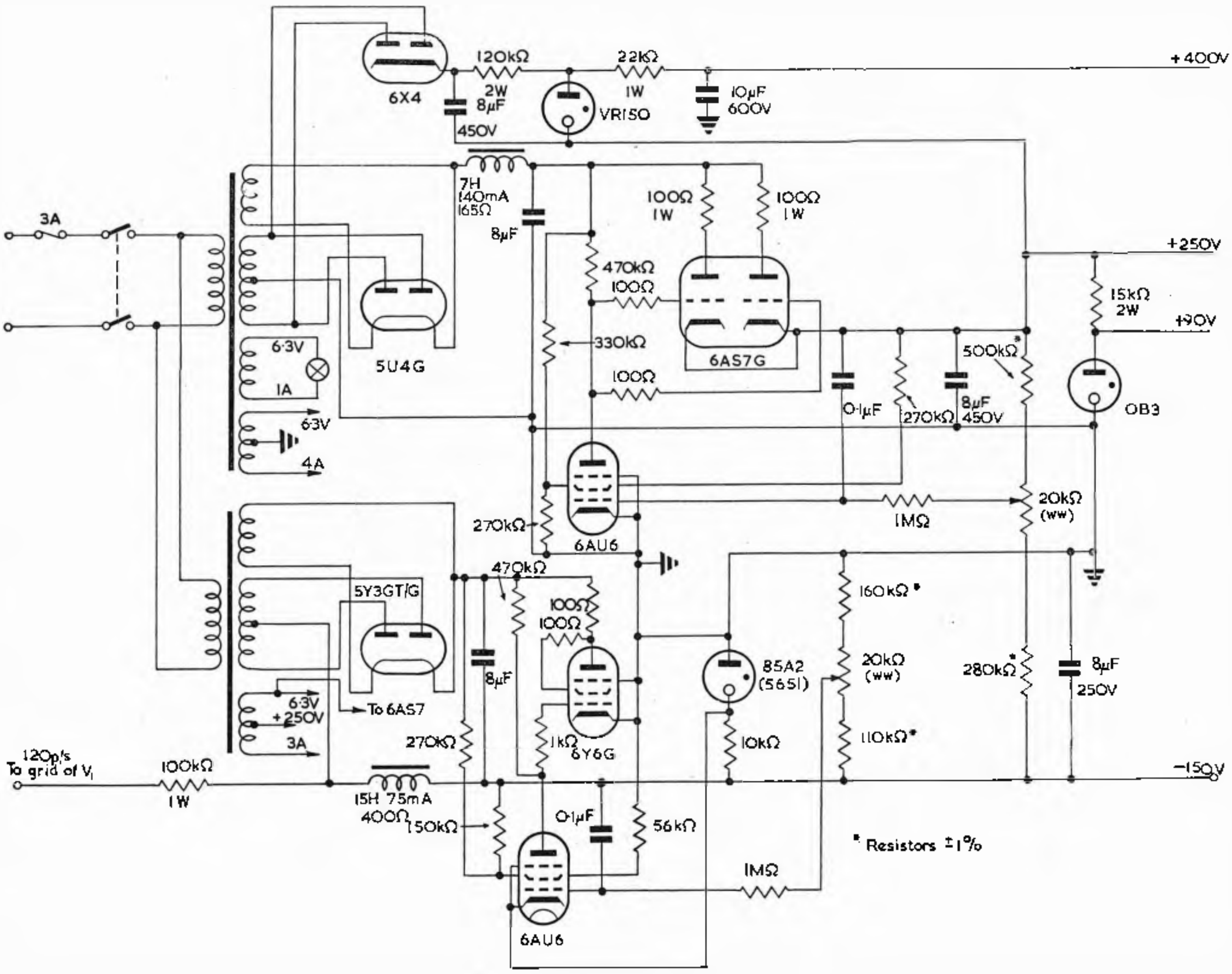
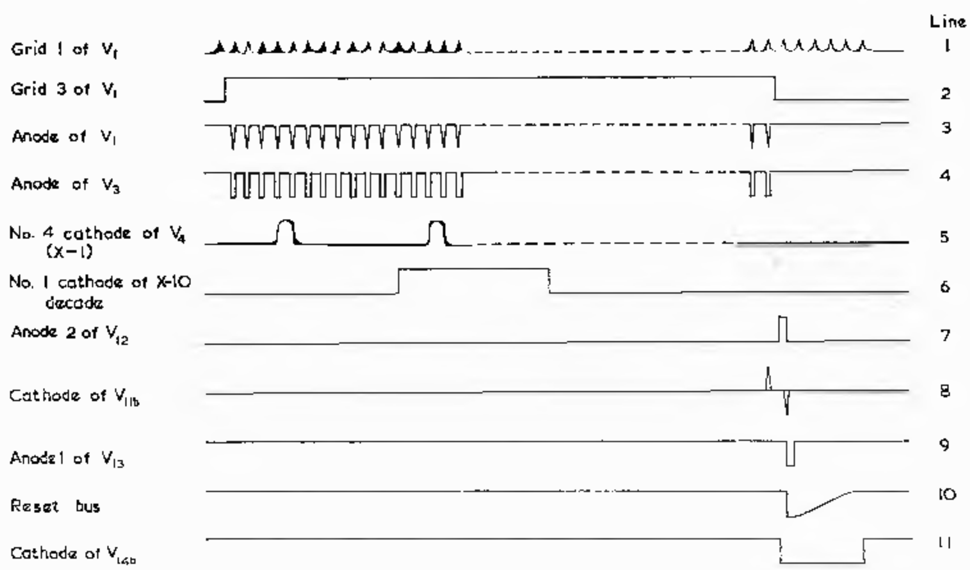


Fig. 6. Power supply circuit

Fig. 7. Timing diagram



of V_1. These pulses are shown to be triangular, but actually consist of peaks of sine waves. Line 2 shows the input derived from the input key to g_3 of the gate tube and is shown as a pulse of unknown length representing the time to be measured. Line 3 shows the output from the anode of V_1. This consists of negative-going pulses of 120c/s repetition rate. The first pulse occurs the first time that a 120c/s pulse is available after g_3 of V_1 has been raised to zero. From then on, pulses are available from V_1 until the key is

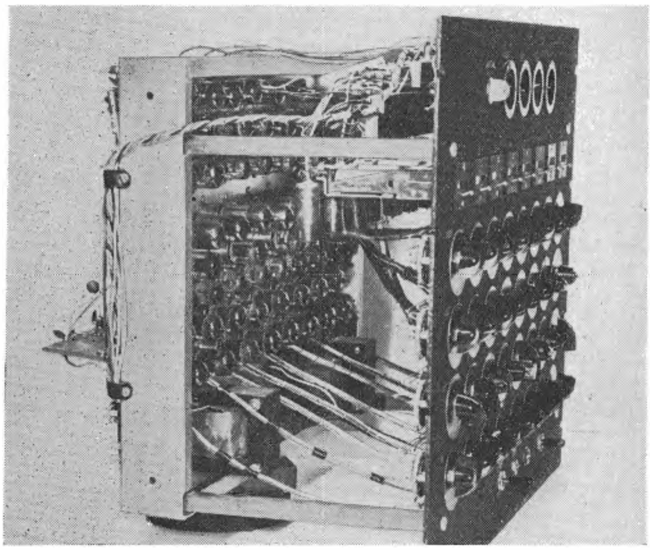


Fig. 8. View of complete equipment, covers removed

released and g_3 of V_1 returns to beyond cut-off. The last pulse is the one that occurs just prior to the re-opening of the key.

Line 4 shows anode a_1 of the V_3 m.m.v. which is used to drive the first scale-of-ten. Line 5 shows the output from a typical cathode of the $\times 1$ decade (in this case the fourth). A positive pulse is produced after four input counts have been received. This pulse lasts until the next count is received, at which time the glow leaves the fourth cathode and proceeds to the fifth cathode. This then reduces the current of the fourth cathode to zero again and it remains there until the glow returns to it again ten counts later.

Line 6 shows a typical output from the $\times 10$ decade (in this case the first cathode). This glow lasts for 10 periods of the input waveform, as shown on line 6. This output will repeat every hundred input counts, since for every hundred input counts 10 input counts are sent to the $\times 10$ decade. While the $\times$ 100 and $\times$ 1 000 decade outputs are not shown, it is quite evident that the same type of waveform, but occurring every thousand and ten thousand counts, will be obtained from the cathodes of the decades.

The remainder of the timing diagram for the registry of the pulse is not shown. When the $\times 1$, $\times 10$, $\times 100$, and $\times$ 1 000 decade outputs selected by a set of four selector switches are in coincidence, a pulse will be sent to the monostable multivibrator and its corresponding registry circuit and one count will be rung up in that particular channel. The remainder of the timing diagram illustrates the resetting procedure. Line 7 shows the pulse obtained at a_2 of V_{12} when the output key is released. This output is differentiated and sent to V_{13}, the output of which is shown on line 9. Line 10 shows the pulse which has been lengthened in order to allow sufficient time for the glow-counters to return to zero from whatever position they were at previously. Line 11 shows the output from the cathode of V_{14b} and it will be noted that this pulse starts before the reset pulse so that the 6463 register circuits are disabled before the resetting is actually accomplished.

Conclusion

A pictorial view of this instrument is shown in Fig. 8. It is quite compact and has been relatively reliable for the fairly large number of valves which were needed to complete the desired functions. It is not claimed that this is an optimum design, but it was a design which could be made to work in relatively little time. It is quite probable that considerable economy in components could be accomplished by the use of more semiconductor diodes and less thermionic valves.

If it should be necessary to obtain better timing accuracy than that available from the 60c/s power line, it would be entirely possible to introduce a high stability oscillator, instead of the 120c/s pulsating d.c. The accuracy should then be adequate up to the stability of the oscillator. The particular glow counting tubes that were used in this application have a maximum resolution which is considerably beyond anything needed in this type of instrument.

A Linear Accelerator for Industrial Radiography

A 5 million electron-volt linear accelerator has recently been installed at a Ministry of Supply establishment. This machine is believed to be the first of its kind designed specifically for industrial radiography.

The outstanding advantage of linear accelerators is that they give more X-ray output and greater energy than machines previously used for radiography, including Van der Graaf machines, resonant transformers, and cascade generators. The machine also compares favourably with multi-curie radio-isotope sources such as Cobalt 60. For example, a half-hour exposure using the linear accelerator will give a radiograph of the best obtainable definition of a steel specimen 9 or 10 inches thick, whereas the same exposure using a Cobalt 60 source of reasonable size would only give about 5 inches penetration.

The 5MeV linear accelerator has been designed to give extremely good definition. Despite the high energy and the large output of over 500 roentgens per minute at 1 metre the electron beam which creates the X-rays has a diameter of only 2mm when it strikes the target. In addition, the polar diagram of the output tends to be flatter than is normally associated with such a high energy beam because of a special new magnetic focusing device associated with the X-ray head.

The main part of the machine has been kept small, the overall length being only 9 feet. This has made it possible to obtain a very high degree of mobility so that the accelerator can be moved into the best position for obtaining a radiograph, instead of moving the specimen, which may be bulky and unwieldy. The following movements are possible:

(1) The whole accelerator can be tilted from vertical to horizontal.

(2) The X-ray head rotates.

(3) The accelerator turns laterally through 180°.

(4) The whole machine can be raised or lowered 8 feet.

(5) The machine is mounted on an overhead rail so that transverse and longitudinal movements are obtainable over the room which houses it.

The accelerator was designed and constructed by Mullard Research Laboratories on behalf of the contractors, Philips Electrical Limited.

Double-Lock Synchronizing Method for Frequency Division

J. B. Earnshaw*, B.Sc.

A method of synchronization is suggested where both the maximum and the minimum of the timing waveform of a multivibrator are locked to an incoming signal. The design of a simple multivibrator in which the double-lock principle can be used is discussed. Experimental results show that the method produces a large increase in the stability of the division ratio.

THE ideal timing waveform for frequency division is a linear sawtooth with zero flyback time. Such a sawtooth waveform with positive synchronizing pulses (sync pulses) superimposed is shown in Fig. 1. If n_1 is the frequency of the sync pulses and n_2 the sawtooth frequency, the optimum size of sync pulse e, is given by

$$e = \frac{E}{(n_1/n_2 + \frac{1}{2})} \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where E is the voltage excursion of the timing waveform. The final sync pulse on the linear rise exceeds the on-level

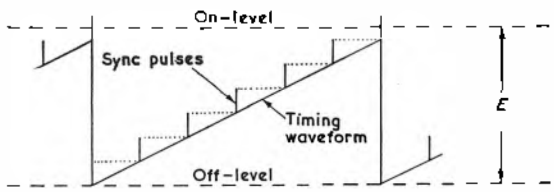


Fig. 1. An ideal timing waveform with optimum sized sync pulses superimposed

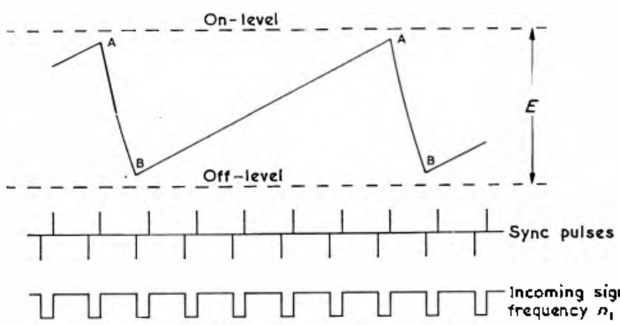


Fig. 2. Actual timing waveform and sync pulses for double-lock synchronization

by $e/2$ so that a variation of either the total excursion voltage, or the rate of rise of the sawtooth, can be accommodated. The percentage total variation permissible in each case is $100e/E \simeq 100/n$ (for $n > 5$ say) where $n = n_1/n_2$, the division ratio.

In practice the flyback is never instantaneous and the flyback voltage waveform rarely linear, so that jitter at the off-level can also cause slipping of the division ratio from n to $n \pm 1$. Consequently slipping takes place before the variation of $100/n$ per cent is reached.

Double-lock Principle

The double-lock principle involves the synchronization of both the on-level and the off-level limits of the timing waveform by the incoming signal. This is achieved by using both positive and negative sync pulses; the positive to lock points A (Fig. 2) and the negative to lock points B. These sync pulses are obtained by differentiation of the rectangular waveform whose frequency is to be divided. Fig. 2 shows the relationship between the sync pulses and the sawtooth timing waveform with double-lock synchronization. Some control of the flyback time is of course necessary so that points A and B coincide with positive and negative sync pulses (not necessarily successive ones).

Suitable Circuit

The asymmetrical multivibrator whose circuit appears in Fig. 3 is suitable for adaption to the double-lock technique. It has previously been used for single-lock frequency division by Attew[1] and by Bland and Cooper[2]. It is basically a Schmitt[3,4] trigger circuit with a large backlash. The operation of the circuit can best be described by initially

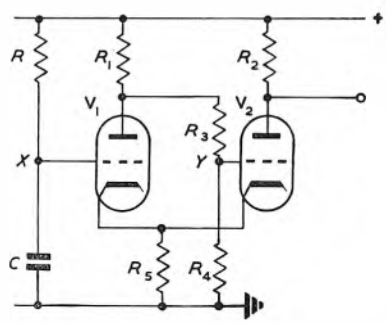


Fig. 3. Asymmetrical multivibrator for frequency division

considering the potential at X to be zero. Thus V_2 is conducting, with its grid at potential Y_1 (the on-level potential) and continues to do so until C charges sufficiently that the potential of X approaches Y_1. Then, with sufficient loop-gain a regenerative action takes place where V_1 is switched on and V_2 off. Due to the high value of R_1, V_1 cannot maintain the current previously passed by V_2 and grid current flows causing C to discharge. As C discharges the potential of X and the common cathode falls until it approaches Y_2 (the off-level potential) when regeneration again takes place restoring the circuit to its original condition whereupon the cycle recommences.

The potential waveforms at various points in the circuit are shown in Fig. 4. A suitable output, for continued division, can be obtained from the anode of V_2, and the sync pulses can be injected at Y, the grid of V_2.

Practical Considerations

(1) For regenerative action the loop-gain must be greater than 1 . . . .

$$\text{i.e. } \tfrac{1}{2}\mu \frac{R_1\,(R_3 + R_4)}{R_1\,(R_3 + R_4) + r_a\,(R_1 + R_3 + R_4)} \cdot \frac{R_4}{(R_3 + R_4)} > 1 \quad (2)$$

(2) V_1 must be unable to maintain a potential of Y_2 at its cathode . . . .

$$\text{i.e.} \qquad R_3/R_a > R_4/R_5 \quad \ldots\ldots\ldots\ldots\ldots\ldots (3)$$

where R_a is the d.c. resistance of V_1 when just drawing grid current.

* *Auckland University College, Auckland, New Zealand*

(3) If V is the h.t. voltage $Y_1 = VR_4/(R_1 + R_3 + R_4)$ .. (4) and

$$Y_2 = VR_4\,(R_2 + R_5)/[R_1(R_2 + R_3 + R_4 + R_5) + (R_2 + R_5)(R_3 + R_4)] \quad \ldots\ldots \quad (5)$$

(4) If v is the valve cut-off bias voltage, the timing waveform voltage excursion is $E \simeq (Y_1 - Y_2) - 2v$ .... (6)

(5) During the charging of C, the potential of X attempts to rise from $(Y_2 + v)$ to h.t. with a time-constant RC. The deviation from linearity is less than 5 per cent if

$$[V - (Y_2 + v)] > 10E \quad \ldots\ldots\ldots\ldots\ldots \quad (7)$$

During the discharge of C, the potential of X attempts to fall from $(Y_1 - v)$ to $V(r + R_5)/(r + R_5 + R) \simeq 0$. With a time-constant of $CR(r + R_5)/(r + R_5 + R) \simeq R_5C$. ($r$ is the forward resistance of the diode formed by the grid and cathode of V_1.)

(6) The output signal amplitude is Y_2R_2/R_5 if taken from the anode of V_2.

The following component values, although not claimed to be the optimum ones, satisfy all the above conditions

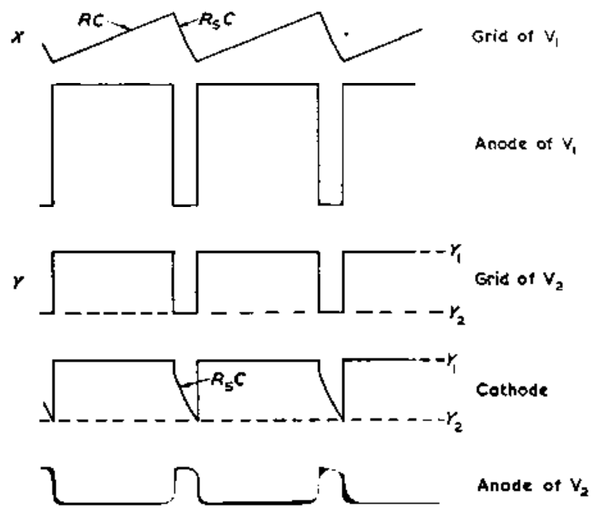


Fig. 4. Waveforms of the multivibrator, showing approximate d.c. voltage levels

and give adequate performance with an ECC91 double triode.

$R_1 = 56\text{k}\Omega$	$R_4 = 22\text{k}\Omega$
R_2 as desired	$R_5 = 10\text{k}\Omega$
$R_3 = 22\text{k}\Omega$	h.t. = 330V

With these components the value of E is 30V.

Synchronization

If the sync pulses are superimposed upon the 'on' and 'off' levels instead of upon the timing waveform itself, i.e. injected at Y, points A will be locked by negative and points B by positive pulses, so that an incoming signal of positive going rectangular pulses would be preferred. Hence the multivibrator described is most suitable for continued division as the output from one stage provides an ideal sync waveform for the following stage.

Setting-up Procedure

The adjustments to C and R for double-locking can be carried out using a single beam oscilloscope connected to the common cathode. Since the cathode waveform follows the potential of Y when V_2 is conducting, the sync pulses can be seen on the horizontal sections of the oscilloscope trace (Fig. 5). For Fig. 5, the division ratio $n = 8$ and the points A coincide with negative sync pulses while the points B coincide with the next but one positive sync pulse.

Stability

With a constant input frequency and a fixed value of C, a division ratio of 10 was set up. The value of R was then varied to alter the division ratio from 10 to 10 ± 1. The ratio of the total variation of R to the average value of R gives the fractional change in the natural frequency of the multivibrator which could be 'pulled in' by the synchronizing signal. This change in frequency can be brought about by many causes, one of which, the variation in R, is convenient for experimental purposes. The procedure was repeated for several values of C and a graph

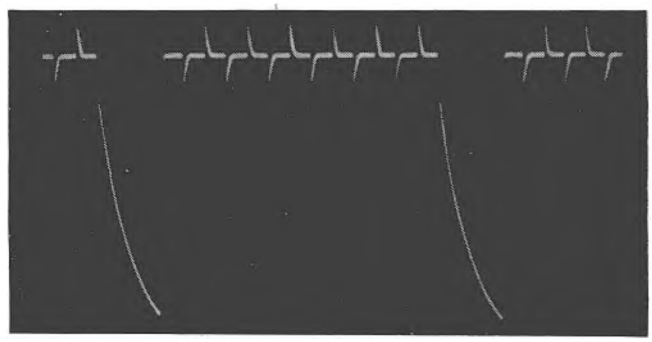


Fig. 5 Waveform observable at the common cathode showing incoming sync pulses (n = 8)

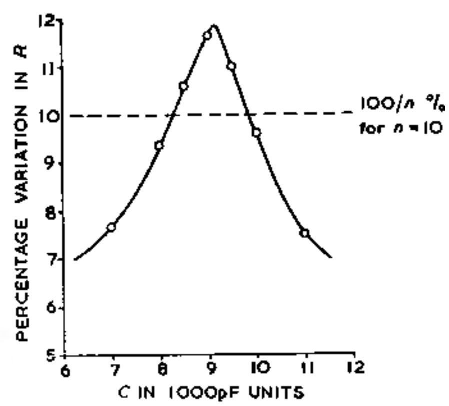


Fig. 6. Percentage variation in R as a function of C for n = 10

of the percentage variation in R versus C was plotted (see Fig. 6). The peak of the graph, representing maximum stability, corresponds to the value of C at which double-locking takes place.

Conclusion

An increase in stability of the division ratio by nearly 70 per cent is obtained when using the double-lock synchronization principle. The peak value of the graph in Fig. 6 is even greater than the permissible variation of $100/n$ for the 'ideal' single-locking case.

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A Zero Correcting Servo For use with D.C. Amplifiers

By B. Shackel*, M.A., A.B.Ps.S., and M. Beaney*

This device can correct the external 'zero' level input to a high performance d.c. amplifier at least to within $\pm 5\mu V$ dead zone. It is designed primarily to null drift voltages which have developed in the external input circuit; it will incidentally also null amplifier drifts.

WHILE investigating very small potentials with a d.c. recording system, it is often necessary to be able to back-off to zero a standing or slowly drifting base potential upon which the required signals are superimposed. When this has to be done simultaneously on several channels, the experimenter's task is not easy. In the course of developing an eye movement measuring apparatus for the technique known as electro-oculography[1] a relatively simple, semi-automatic, zero correcting device was found to be desirable.

The problem was that, although the d.c. amplifier drifts by less than $\pm 20\mu V$ in half-an-hour, the contacting electrodes and the subject's skin, even with due precaution[2], may produce drifts of the order of 100 to $200\mu V$ after several minutes. It is these drifts which it was desired to balance out at will on two separate channels just by pressing a push-button. It is understood[3] that d.c. electroencephalographic recording may have the same requirement, and it is presumed that this general problem occurs elsewhere.

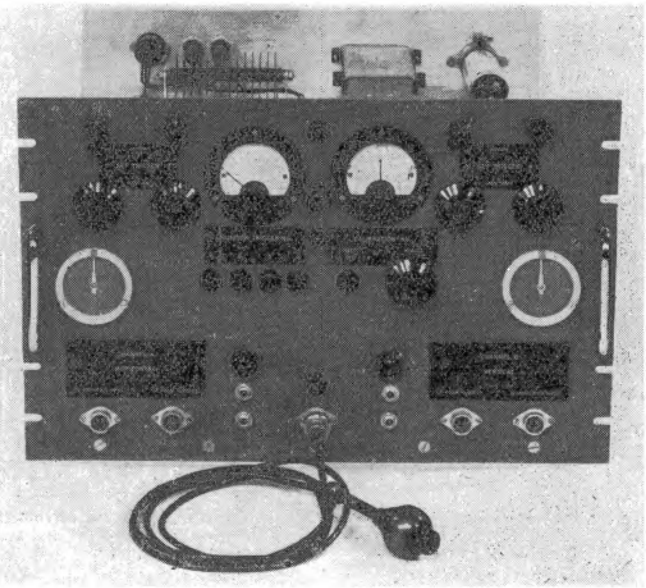
Some form of servo system, closing the loop from the main amplifier output to input and switched into circuit at the appropriate moment, was clearly indicated. For simplicity and relative cheapness a 'bang-bang' system operated by sensitive trigger elements was chosen, once a preliminary study had shown that reasonable performance could be expected.

The general form of the final system is shown in the block diagram (Fig. 1). When the push-button is closed, with the input signal theoretically at zero but in fact with a superimposed drift, the amplifier output voltage will operate the appropriate trigger element. The trigger relay contacts apply current to the motor to drive the helipot voltage source in the right direction to counteract the drift which has occurred. When the correction brings the amplifier output voltage below the relay element trigger voltage, the trigger reverses and the relay contacts apply armature braking to the motor. The push-button is then released but the backing-off voltage, of course, remains in circuit.

The Trigger Element

It was desired, by means of this system, to correct the 'zero' level at the amplifier input at least to within a dead zone $\pm 10\mu V$ of absolute zero. With the amplifier gain of 10 000 it was therefore necessary to be able to set the trigger elements to operate at about ± 100mV. More important, for satisfactory operation it was clearly essential that the trigger element 'backlash', and any drift of the triggering point, should not be much more than 10 per cent of the main 'zero' dead zone. Therefore a trigger unit was required which could be set accurately to within say $\pm 2\frac{1}{2}$mV of the nominal operating point, which would not drift by much more than 10mV, and which had a 'backlash' zone not much wider than 10mV. The development of a suitable, sensitive, trigger element was found to be necessary.

A circuit[4] was investigated, which had been found promising in another application, although it had been described as "Not designed for rapid operation and the point of regeneration is very unstable". By a suitable choice of circuit values, and by using a standard constant voltage transformer and a standard stabilized sub-unit to supply heaters and h.t. respectively, good operating stability was achieved. By using a suitable valve it was found that sufficient current change occurred on triggering to operate a standard post-office type 3000 relay. The final circuit is given in Fig. 2.



Front panel of the complete 2 channel servo unit with the four trigger setting and bias controls at the top left and right, the anode milliammeter and selector switches in the left centre and the calibration meter and controls in the right centre. The helipot position indicators are at the left and right sides, with the motor indicator lamps to the left and right of centre near the push-button.

By means of the 50Ω potentiometer the positive feedback is increased until a small range of settings is reached at which the system will trigger from one state to the other, i.e. minimum or maximum current through V_{1b}, for a voltage change at the input of between 10 and 15mV. This triggering occurs with the grid about 0·6V positive with respect to earth. By means of a voltage source in series with the input the trigger point, or rather the 10 to 15mV triggering zone, can be set at any absolute voltage with respect to earth.

The $1k\Omega$ preset variable resistor in V_{1b} cathode is to allow for change of valves. The very small range 50Ω potentiometer, in conjunction with the anode circuit meter, ensures easy setting up for minimum trigger 'backlash';

* *E.M.I. Electronics Ltd.*

but the change in characteristics between different valves may put the trigger point outside the range of this 'fine' control.

The anode current of V_{1b} changes between 3 and 8mA for the two trigger states; a type 3000 relay with a 2·5kΩ coil and four changeover spring sets was found to be optimum for these limits. High wattage resistors are used in the cathode, because a slow but steady drift of the triggering point was found to occur when the trigger was operated intermittently and was left passing 8mA in the meantime; the drift was caused by temperature change in resistors chosen with a less conservative wattage rating.

Nine different 12AT7 valves were tested in this circuit; each was given ten minutes to warm-up and then the triggering points in both directions were measured at every minute for half-an-hour. The trigger 'backlash' zone never exceeded 11mV, and the trigger operating points never drifted by more than 15mV over the whole test period.

The Servo

The complete circuit of the zero correcting servo based upon these trigger units is shown in Fig. 3; the triggers and the standard power supply units are not detailed for simplicity. The power supplies for both triggers and motors, and the calibration section, are of course common to however many separate servo channels are built in one unit.

The effectiveness of the armature braking was such that the triggering points could safely be set at + and −50mV of true zero instead of + and −100mV. With supplies cut and braking applied at this point the motor runs in and

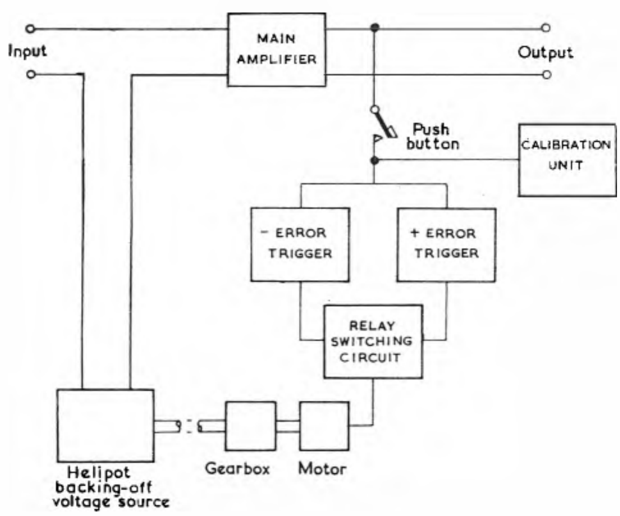


Fig. 1. The complete system

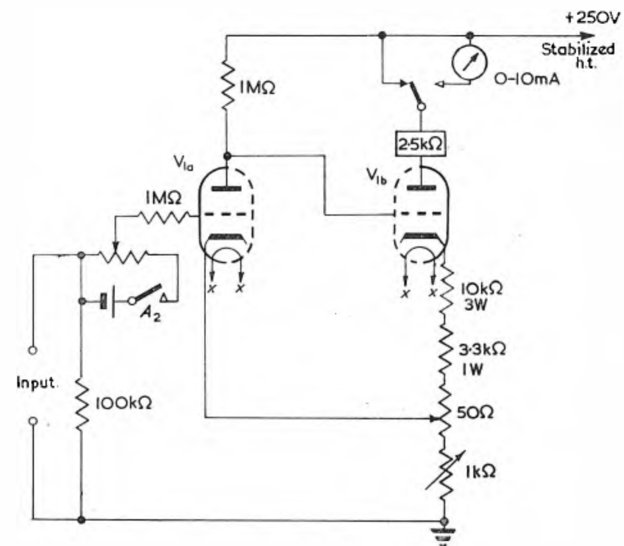
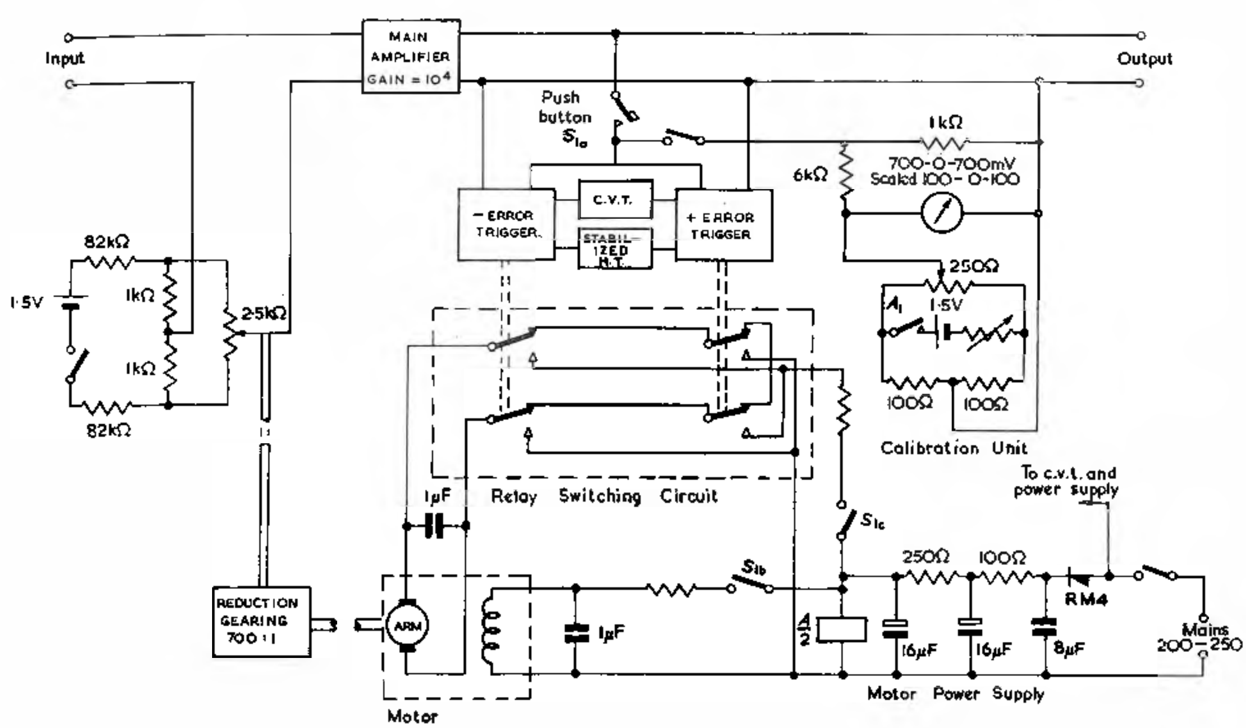
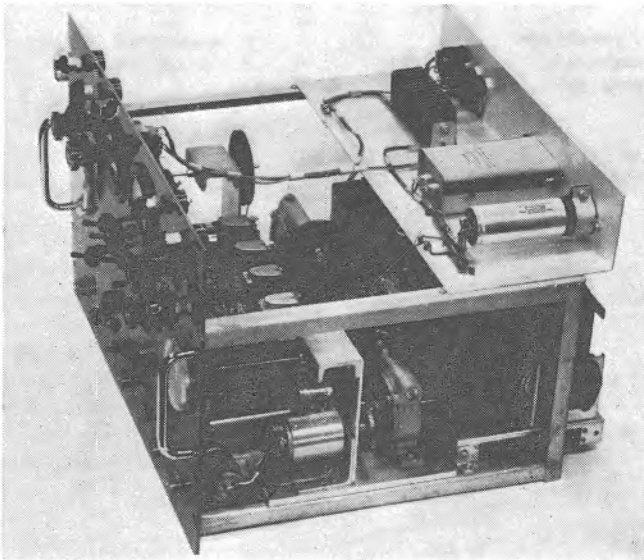


Fig. 2. The trigger circuit

Fig. 3. Circuit diagram of complete servo (1 channel)



stops almost exactly in the centre of the input zero dead zone. The gearbox reduction ratio, the motor speed, and the trigger operating point, of course all interact in this respect and can be suitably adjusted to achieve the optimum performance with minimum response time and no overshoot. The armature and field coil series resistors must be above some minimum value to suit the motor used, and



An interior view of the servo unit. The helipot driven via the gearbox is nearest the camera; behind it are the relays and trigger circuits. The stabilized power supplies are at the bottom rear of the chassis with the motor supply above.

thereafter can be adjusted finally to set the motor speed and braking.

An adequate range of backing-off voltage from the motor-driven potentiometer is needed to avoid the nuisance and delay of having to run back away from the end stops during an experimental session. However a minimum discrimination of about $2\mu V$ per potentiometer turn is needed if the servo is to work satisfactorily to a $10\mu V$ dead zone. Therefore a ten revolution helical potentiometer is used to satisfy both these requirements. With over 5 000 turns on the potentiometer adequate discrimination and a total backing-off voltage range of 10mV is achieved. In our application the initial potential difference between a pair of electrodes is nulled with the usual hand operated control on the amplifier, and thereafter the servo is used; drifts during a normal session have never accumulated to more than ± 5mV, and thereafter the total backing-off range of 10mV available has been found entirely adequate.

No difficulties with hum pick-up have been experienced (the bandwidth of the amplifier in use is 0 to 15c/s and the response is about 40dB down at 50c/s). The normal precautions of screening the leads in the helipot circuit, and placing the helipots at distant convenient points on the chassis from the motors and power supplies, have of course been adopted.

The calibration unit enables up to ± 100mV to be supplied at will to the input of each trigger unit to ensure easy and accurate setting of the operating point. The centre zero 100—0—100 meter reads full scale for ± 700mV, and this is of course reduced by an accurate potentiometer network to give the required ± 100mV output.

Three extra refinements, for clarity not shown on the circuit, have been added to aid the operator. A 10:1 reduction pulley from the helipot shaft indicates via a pointer the helipot wiper position, so that there is no risk of the wiper inadvertently being driven against the end stops. A panel lamp, lit via other contacts on the trigger relays when the motor is energized, shows when the motor is running and when the zero correction is finished. A push-button on the end of a long lead, connected in parallel with the panel-mounted push-button, allows the experimenter much greater freedom of movement and control during the experimental session.

The two channel unit, which has been built to this design, operates without any overshoot or hunting and corrects a $200\mu V$ drift signal in 2·5sec, a $500\mu V$ drift in about 6sec. The complete unit was tested for three consecutive days, being switched on in the morning and afternoon and off at midday and at night. None of the setting-up controls was altered throughout the test period. The triggering points and trigger backlash zones were checked fairly frequently at random intervals. The smallest and largest trigger backlash zones which occurred were 5 and 15mV respectively. The maximum drift of any triggering point during any one session was 14mV. The maximum deviation of any triggering point from the original setting-up level over the whole three days was 25mV.

The results with this unit in regular use are entirely satisfactory. The ultimate accuracy and minimum zero dead zone obtainable are more likely to depend upon the noise level and gain of the main d.c. amplifier than upon the limitations of the servo. By changing the backing-off voltage range and thus the discrimination per turn of the helipot, and by setting the triggering points differently, this unit can probably be generalized to a large number of applications.

Acknowledgments

The authors wish to thank the Directors of E.M.I. Electronics Ltd. for permission to publish this article.

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An Electronic Data-Processing System for Norwich

The City of Norwich has recently installed in the City Hall a comprehensive data-processing system (National-Elliott 405) consisting of an electronic computer supported by magnetic film storage and punched paper tape input and output devices. The first task for which the computer will be used is the preparation of rate demand notes and the maintenance of ratepayers' personal accounts.

Two sets of films have been prepared. In the first set are seven films, one for each district, containing details of all properties. The second set is restricted to properties for which a person other than the occupier is liable for the rates, and consists of two films: a special reference is stored with the record of property in the first set if it is also included in the second set.

The information stored on these films has been recorded on them by the computer, acting on information received from hand-punched paper tape.

The preparation of rate demand notes is preceded by the punching on paper tape of the amount of the rate in the pound to be levied: this short length of tape is used to inform the computer of the current rate factor. Each reel of magnetic film is then placed on the computer in turn, the rate due calculated and a record, in punched paper tape form, produced containing all the details required for the printing of a rate demand note. This paper tape is used to control electric typewriters which produce the finished product on continuous stationery. In those cases in which a person other than the occupier is responsible for the rates, separate schedules of the properties are produced to be attached to the rate demand notes which will then show only the total amount due.

During the preparation of demand notes, the amount due from each ratepayer is stored on the magnetic film and a separate operation is carried out to produce a tabulation to form a ratebook.

A 16kc/s Amplitude Modulator Employing Envelope Feedback

By K. G. Corless*, B.Eng.

The modulator described in this article permits the amplitude of a 16kc/s carrier to be varied with an accuracy of 1 per cent in conformity with a modulating voltage. This accuracy is preserved for all modulating frequency-components within the range 0 to 30c/s.

THE techniques of amplitude modulation, which apply to the variation in amplitude of a carrier-frequency signal in conformity with a slowly varying modulating-frequency signal, are by now well established and widely used, as in radio and line-telephone equipment. Among these techniques is that of envelope feedback[1-3], used to improve the linearity of the resultant modulation in a manner basically similar to that of negative feedback in audio amplifiers. The linearity is made dependent primarily on the degree of feedback and on the linearity of the demodulator, rather than on the assumption of linear valve characteristics.

The circuit to be described employs envelope feedback, but differs from those mentioned hitherto in that a departure from linearity of about 1 per cent of output voltage was specified, and therefore a greater degree of feedback is used in order to achieve this accuracy. The carrier frequency is 16kc/s and the range of modulating frequencies anticipated in practice extended from zero to about 20c/s.

As an ancillary requirement, the output impedance of the final stage in the modulator was required to be about 1Ω, such that it could be considered as a constant-voltage source.

General Description

The block diagram of Fig. 1 shows the interconnexion of the constituent circuits which results in envelope feedback. The complete circuit diagram is shown in Fig. 2 where it should be noted that, since the range of modulating frequencies extends down to zero, both the modulating-frequency (m.f.) amplifier and the demodulator are directly coupled.

The equation for the instantaneous value of voltage for an amplitude modulated carrier is

$$e(t) = (n + m \cos \omega_m t) \cos \omega_c t \quad \ldots\ldots\ldots (1)$$

where ω_m = angular modulating frequency

ω_c = angular carrier frequency

n and m are time variable amplitude coefficients related to the non-periodic and periodic variations in carrier amplitude respectively. ($n > m$).

If equation (1) is expanded trigonometrically, it becomes

$$e(t) = n \cos \omega_c t + (m/2) \cos(\omega_c + \omega_m)t + (m/2) \cos(\omega_c - \omega_m)t \quad \ldots\ldots\ldots\ldots (2)$$

Equation (2) shows that for amplitude-modulation to obtain, the resultant output waveform shall contain no periodic terms other than those which involve the carrier-frequency and the sum and difference frequencies, or side-bands. In other words frequency-selective circuits are required in the modulated-carrier amplifiers, which will effectively suppress, in particular, the modulating frequency.

In common with all negative feedback circuits care must

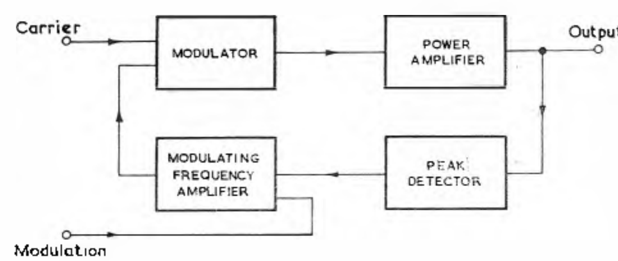
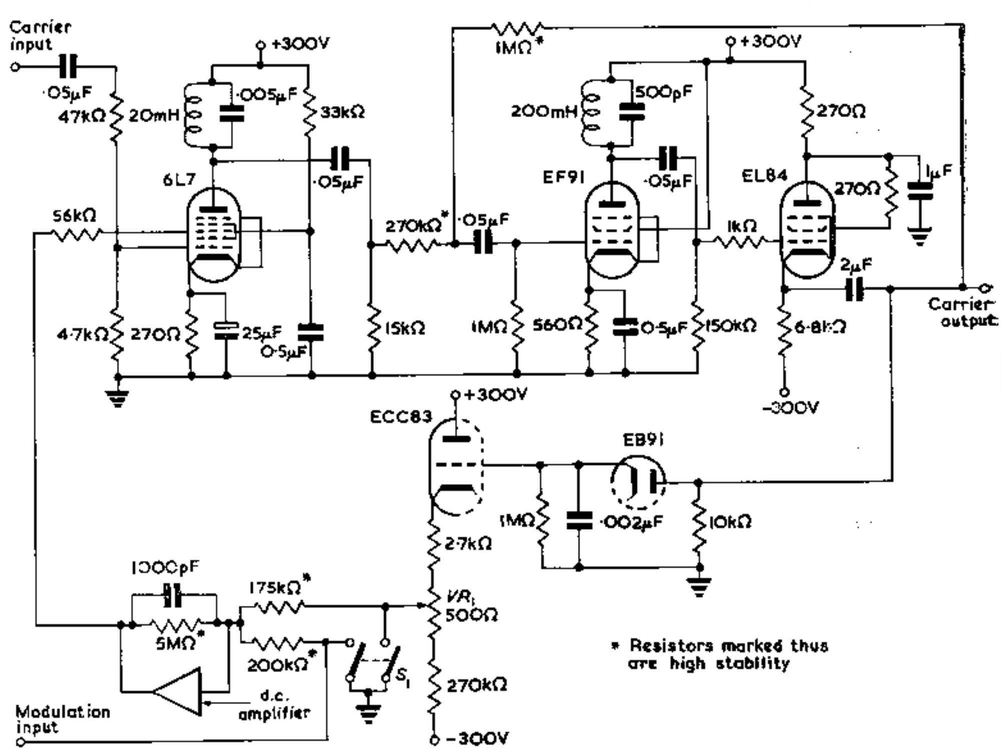


Fig. 1. Block diagram

Fig. 2. Circuit diagram



* The University of Liverpool

be exercised to ensure that the loop phase-shift does not exceed 180° at any frequency for which the loop gain exceeds unity. Normally the feedback loop will incorporate some form of low-pass filter, the cut-off frequency for which will be less than the carrier frequency. This must be so since negative feedback is required over the range of modulating frequencies only, and not at carrier frequency. In this modulator the feedback loop includes the carrier-frequency circuits, which can affect the loop phase characteristic if they introduce any envelope phase-shift. Therefore the parallel-tuned circuits used in the carrier-frequency amplifiers are damped to produce bandwidths which ensure a negligible envelope phase-shift for the anticipated m.f. range.

The Low-Output-Impedance Amplifier

Control of amplifier output impedance can be effected by employing the technique of positive-current and negative-voltage feedback[4]. By this method, positive, zero and negative output impedances can be realized. However, in this case, where the steady-state output power was low, recourse was made to the reduction in output impedance

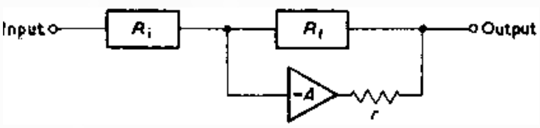


Fig. 3. Scaling circuit

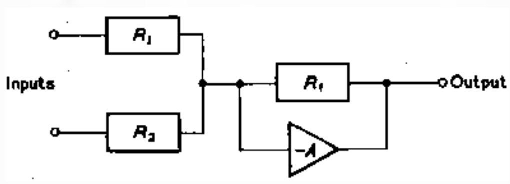


Fig. 4. Summing and scaling circuit

which results from the type of negative-voltage feedback used in the scaling amplifier circuits of analogue computers[5], and shown in Fig. 3, where r represents the output impedance of the amplifier. It can be readily shown that, with feedback applied, the effective output impedance is equal to r divided by the reduction in overall gain caused by the feedback. Hence, if a cathode-follower is used as the final valve, with a value for r of, say, 200Ω, then a gain reduction of 200 times will produce an effective output impedance of 1Ω.

The circuit of this amplifier uses a single stage of tuned amplification followed by an EL84 output-pentode, triode connected, as cathode-follower. Gain without feedback is approximately 500 times and with an overall gain with feedback of about three times, the output impedance is reduced from 220Ω to the order of 1Ω.

Whereas the large dissipation in the cathode load of the EL84 is a disadvantage, the simplification in circuit design is considerable.

During normal operation, the load on this amplifier is about 1kΩ with an r.m.s. load voltage of 10V.

The Modulator

Linear modulation has been preferred for simplicity and is achieved by the use of a heptode in which the transconductance of the first grid is a near-linear function of the third grid potential. A 6L7 has been used in this capacity but there is no obvious reason why the miniature type 6BE6 should not be satisfactory.

In the circuit of Fig. 2, a stage gain of about 12 times is obtained with a third grid static potential of −4V with respect to earth. This gain is sufficient to produce the required swing in load voltage while ensuring that the first grid is not overloaded. The 56kΩ series resistor to the third grid helps to limit the current drawn when the driving voltage for this grid goes excessively positive.

The Peak Detector

A series diode circuit, of conventional design, is direct-coupled to a cathode-follower which fulfills the role of providing a low loading on the detector and of permitting the output voltage to be set to zero for zero a.c. input to the detector. This latter facility also permits the spurious voltage caused by diode splash current to be backed off with negligible error because of the large voltage drop across the load resistor of the cathode-follower.

The Modulating-Frequency Amplifier

This forms the most important feature of the loop so far as accuracy and loop stability are concerned, for it is in this amplifier that the loop gain-frequency response is controlled. There are actually two other factors which affect this loop characteristic, namely the envelope-gain versus modulating-frequency response of both the carrier-frequency circuits and the detector. However, the tuned circuit bandwidths and detector load time-constant have been chosen such that the associated gains can be considered constant throughout the range of modulating frequencies for which the loop gain is appreciably greater than unity, a typical figure for 3dB bandwidth being 3kc/s. Therefore the form of the gain-frequency curve of the loop corresponds approximately to that of the m.f. amplifier; there being a constant multiplying factor given by the envelope gains of the carrier-frequency circuits and demodulator.

It is evident from feedback theory that a loop gain of 100 is required if the error is to be limited to 1 per cent.

Analogue computer practice has, again, been adopted in the design of this circuit, where a three-stage direct-coupled amplifier has been used in the summing and scaling circuit[5] of Fig. 4. The internal gain of this d.c. amplifier is about 800. If the polarity of the direct-voltage from the demodulator is made opposite to that of the reference voltage, then, for equal input resistors, the output will be zero when the magnitudes of the two voltages are equal. More generally, for zero output voltage, the ratio of the input voltages will be equal in magnitude to the ratio of the corresponding input resistors.

By suitable adjustment of the feedback resistor the overall gain, and hence loop gain, can be varied at will. If the feedback resistor is shunted by a capacitor it can be simply shown that the gain/frequency response of the amplifier has the same form as that of the shunt RC combination alone, with an asymptotic slope of 6dB per octave. Adjustment of this shunt capacitor is therefore a convenient way of controlling the gain-frequency response of the complete loop.

In common with all direct-coupled amplifiers there is provided a means of adjusting zero. In this case, however, the zero-set control is used to set the static bias on the third grid of the 6L7 to −4V, which value has been found to result in optimum control of modulation.

Circuit Behaviour with Feedback

As mentioned earlier, loop stability has been achieved simply by adjustment of the feedback capacitor in the summing amplifier. It has been found that, for a zero-frequency loop gain of about 100, the resulting gain frequency characteristic for the stabilized loop was quite satisfactory for the intended application; the loop gain falling to about 60 at 50c/s.

The measurement of the gain-frequency response on open loop (Fig. 5) was hindered by the distortion introduced by diagonal clipping in the demodulator at modulating frequencies in excess of 200c/s.

Because the ratio of demodulator output-voltage to peak input-voltage is less than unity (approx. 0·94) it has been necessary to compensate for this at the input of the summing amplifier, by suitable adjustment of the appropriate resistors as previously explained. Fig. 6 is the resulting static plot of peak a.c. load voltage as a function of direct voltage reference; the error is about 1 per cent of output voltage over the range 4 to 28V.

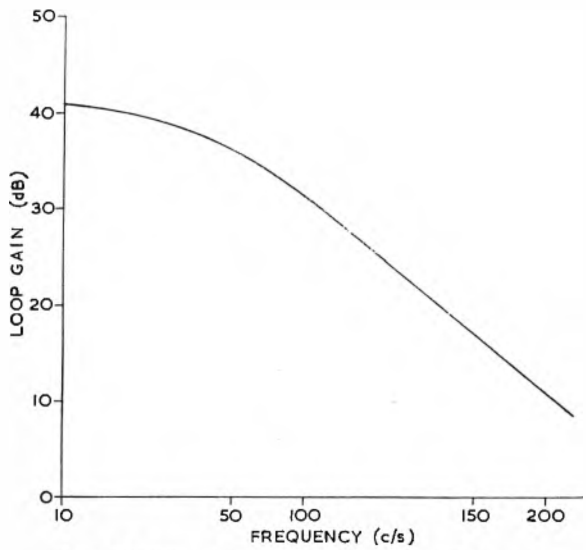


Fig. 5. Variation of loop gain with modulating frequency

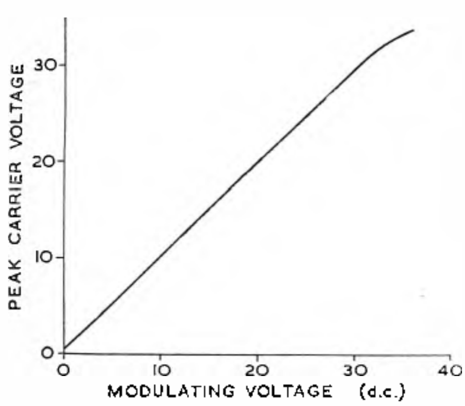


Fig. 6. Modulator linearity

The complete circuit is, by nature, self-stabilizing against changes in amplitude of the 16kc/s input signal. However, the extent of input amplitude variations which can be accommodated is limited by the fact that the envelope gain (and hence loop gain) increases as the input signal increases, as is shown in the Appendix. The limits are imposed by loop oscillation arising from excess loop gain on the one hand, and loss of accuracy caused by reduction in loop gain on the other. This trouble, if pernicious, would require some elementary form of amplitude stabilization of the input signal.

Setting-up Procedure

This involves the adjustment of a few pre-set controls which, once adjusted, will thereafter require only periodical checking. Following normal d.c. amplifier practice it is advisable to allow the whole circuit to warm up thoroughly before setting these controls, but subsequently the circuit can be used immediately after valve warm-up, as these control settings determine only the optimum operating conditions of the various constituent circuits. Overall accuracy is dependent primarily on the input resistances to the summing amplifier, for which high stability resistors will normally be used.

The procedure can be summarized as follows:

(a) The a.c. drive and reference voltage are reduced to zero, and S_1 is closed, which shorts to earth the inputs to the summing amplifier.

(b) The -4V bias level is adjusted by means of the d.c. amplifier set-zero control.

(c) S_1 is opened and the -4V level is reset by means of VR_1 in the demodulator cathode-follower.

(d) The (negative) reference voltage is set to a value corresponding to an approximate centre of the working range of amplitude variation and the a.c. drive is increased until the -4V bias is restored.

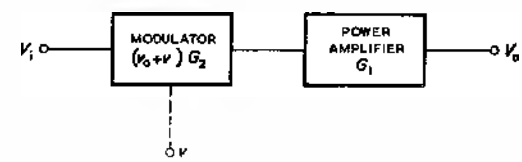


Fig. 7 (above). Carrier frequency amplifier

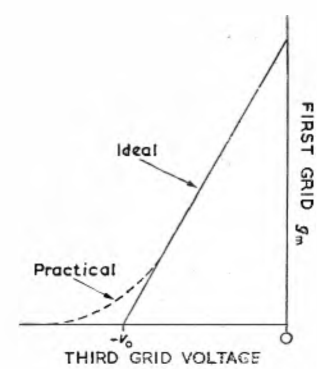


Fig. 8 (right). 6L7 control characteristic

Under these conditions the peak output will equal the reference voltage at the chosen centre value.

Conclusion

It is perhaps fortuitous that the desired circuit performance has been achieved with such an elementary form of loop stabilization. A specification which involved a greater range of modulating frequencies, would undoubtedly necessitate a closer investigation of the circuit performance limitations, including the use of more elaborate interstage networks[1].

Acknowledgments

The author wishes to thank Mr. A. S. Aldred for the many helpful discussions on the subject of this article, and also Prof. J. M. Meek for laboratory facilities provided in the Department of Electrical Engineering at the University of Liverpool.

Appendix

Dependence of Envelope Gain on A.C. Input Voltage

Fig. 7 is a block diagram of the carrier-frequency amplifier, in which the modulator and power amplifier are separated for convenience, and where

$-v_0$ = bias on the third grid of the 6L7 necessary to reduce the mutual conductance to zero for the idealized characteristic of Fig. 8.

v = actual bias on the third grid

V_i = a.c. input voltage

V_o = a.c. load voltage

G_1 = gain of the output stage

$(v + v_0)\,G_2$ = gain of the 6L7 stage with the assumed idealized characteristic

Then, for any given value of v within the working range, the total a.c. gain G_t is given by

$$G_t = (v + v_o)\ G_1G_2 \quad \ldots\ldots\ldots\ldots \quad (3)$$

$$\therefore \quad V_o = G_tV_i$$

$$= (v + v_o)\ G_1G_2V_i \quad \ldots\ldots\ldots\ldots \quad (4)$$

The envelope gain, G_e is now defined as

$$G_e = \frac{\partial V_o}{\partial v} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (5)$$

$\therefore$ from (4) and (5),

$$G_e = G_1\,G_2\,V_i \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

That is, envelope gain is a linear function of a.c. input voltage. Moreover, for a constant value of this voltage, envelope gain is independent of v and V_0 and therefore, once set up, accuracy is maintained throughout the working range. Also, from (4) and (6),

$$G_e = \frac{V_o}{(v + v_o)} \quad \ldots\ldots\ldots\ldots\ldots \quad (7)$$

from which it is apparent that, for a constant output voltage, the envelope gain varies rapidly as v approaches the value $-v_o$, although the curvature of the practical characteristic (Fig. 8) will modify equation (7) in this region, as has been found in practice.

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MANUFACTURING TECHNIQUES FOR WAVEGUIDES

A Symposium held at the Institution of Electrical Engineers

A symposium on "Manufacturing techniques for waveguides" was held at the Institution of Electrical Engineers on 1-2 May under the auspices of the Radio Components Research and Development Sub-Committee on Waveguides. The symposium gave an opportunity to a wider audience to hear and discuss some of the problems of waveguide manufacture which are regularly dealt with by the sub-committee.

Dr. A. F. Harvey, who is the chairman of the sub-committee, provided a general introduction to the papers which followed, describing briefly the various methods of manufacture of microwave components which were to be presented in greater detail by subsequent speakers. E. M. Wareham's paper on "The Economic Aspects of Design and Manufacture of Microwave Components" was a bold attempt to reduce to simple economic terms the complicated problem of the relationships between design development and testing as applied to waveguide equipment. The author stressed the three S's—Standardization, simplification and specialization—as a step to increased productivity. Mr. Jamieson read an excellent paper on the subject of "Integrated R.F. Heads". He described the milled block technique for producing complete waveguide assemblies—a method which he has pioneered—and produced convincing arguments to show the advantages of this system. His paper was accompanied by a film describing the numerical control of machine tools, a process which ideally lends itself to the manufacture of integrated r.f. heads. On the subject of inspection methods, Mr. Byrne was both simple and lucid. He restricted his attention to three topics, the mechanical inspection of waveguide tubing, the mechanical inspection of a waveguide flange and the problem of mechanical interchangeability as particularly applied to flanges. This latter was well illustrated by a specific example. One feels that Mr. Alison's paper on "Dimensional Tolerances in Waveguide Structures", while recording the results of the careful tolerancing of reactive structures, was not of very great assistance when one comes to tolerance a general structure.

Dr. Benson's paper on "Attenuation and Surface Finish" was of necessity only a brief summary of the vast amount of work which the author has carried out on the subject.

The second day's proceedings were opened by Mr. Andrews on the subject of the "Manufacture of Waveguide Components by Metal Spraying". This is a relatively new technique which can produce complicated microwave structures simply and accurately. "Hobbing" is a method of metal forming in which a hardened steel 'hob' is pressed into a billet of metal. Mr. Shearman dealt with the particular application of this technique to the manufacture of small magnetron anode blocks with a very high degree of dimensional accuracy.

Mr. Gardner gave a complete description of the present state of knowledge on the fabrication of aluminium components by brazing.

Three papers were read on the application of casting methods to the manufacture of microwave structures. Firstly, Mr. Mercer spoke on the subject of precision casting of waveguide components using accurate plaster cores. This technique is suitable for quantity production, particularly of light alloy items. "Lost Mercury Investment Casting" was described by Dr. Scholefield. Frozen mercury is built up from dies, etc. to the shape which the product is to take. The frozen mercury is then covered with a suitable material. When the mercury is melted and poured out, a mould is formed from which the final structure can be cast.

In an excellent colour film introduced by Mr. Cole three quite different manufacturing processes were shown. Firstly the application of conventional pressure die casting to waveguide structures, secondly the pressure forming of light alloy units and thirdly the manufacture of complicated microwave assemblies in light alloy by dipping into molten flux at brazing temperature a unit built up from preformed metal sheet coated with a layer of braze metal. The sheets are located and held together prior to dip-brazing by twisted tabs fitted through punched slots (the "German toy" technique).

Dr. Layton gave an introductory paper on electroforming and while his remarks were sound electrochemically, it was obvious from subsequent discussion that his views on the applicability of electroforming were not shared by a number of speakers. Mr. Morrison restricted his paper to a description of the difficulties he had encountered in producing a complicated Q-band waveguide assembly and the methods he developed to overcome them. Mr. Balmer gave a most interesting account of aluminium electroforming which has now become well established on the laboratory scale.

In a paper well illustrated by slides, Mr. Dukes discussed the photo-etching process for the production of microwave printed circuits from copper clad laminates. Mr. Buttery gave a detailed description of the 'spark-erosion' process, with particular emphasis to its use for the accurate cutting of slots in waveguides. Mr. James, discussing "Waveguide for High Pressure" gave some interesting facts and figures on the deformation of standard waveguide when internal pressure is applied in order to increase its power handling capacity. Various methods of counteracting the deformation were proposed, the simplest of which appears to be the use of waveguides whose walls are considerably thicker than standard. Mr. Stevens gave an account of a new way of fabrication, together with interesting information on tolerances.

The symposium was accompanied by a diverse and most interesting collection of samples illustrating the many methods of manufacture which the speakers had described.

The British Post Office Letter Sorting Machine

The letter sorting machine demonstrated at the recent Instruments, Electronics and Automation Exhibition at Olympia is a prototype machine designed and built by the Post Office Engineering Department. Twenty machines based on the present design are being built under contract to the Post Office by The Thrissell Engineering Company Ltd of Bristol for extended field trials.

The machine measures 16ft long by 2½ft wide by 6ft high, and weighs 2¾ tons. It is designed for control by a single operator and is used for sorting letters and postcards to 120 different destinations by the operation of a keyboard comprising two groups of twelve keys. To operate the machine two keys, one in each group, are depressed simultaneously.

The letter path through the machine can be considered as consisting of four main parts, namely:

(1) *The letter feed conveyor* is a band upon which the unsorted mail is stacked. A photocell at the head of the stack controls the feed of the conveyor to maintain the top letter of the stack in the correct position for entry to the presentation unit.

(2) *The presentation unit* controls the transit of the letter from the unsorted stack, via viewing windows to the main conveyor. The unit operates in response to the operation of a pair of keys, feeding one letter into the main conveyor, presenting a second letter to the operator's view, and picking a third letter from the top of the unsorted stack.

(3) *The main conveyor* carries all letters from the presentation unit to the start of the selective conveyor system, and includes a synchronizing gate which, if necessary, halts the letter for up to ½sec in order to bring the letter into synchronism with the selective conveyor memory system. (A feature of the machine is the masking of the machine rhythm from the operator which allows him to 'key' a letter away at any instant.) This necessarily means that letters enter the main conveyor at random time with respect to the memory system of the machine, and have to be synchronized before entering the selective conveyor system.

(4) *The selective conveyor* consists of five horizontal conveyors situated over five rows of stacking boxes, and a vertical distributor feeding these five horizontal conveyors. Divertor blades are interspersed along the conveyors to divert the letter to the correct horizontal run and thence to the correct stacking boxes. The divertors are operated at the appropriate time by linkages from pin wheel memories which store the information keyed by the operator until the letter reaches the diversion point.

The drive for the letters on the main conveyor and on the selective conveyor system is by means of pairs of rollers comprising rubber tyred idlers sprung against driven rollers. Every sixth idler is equipped with a centrifugal switch which stops the conveyor should a letter jam in the track.

The control for the machine is partly electronic and partly mechanical. The control system can conveniently be considered as four units:

(1) *The cold-cathode store* is provided to retain the information keyed by the operator until the main pin-wheel memory is in the correct position to permit pins being set. The storage time in the cold-cathode store can be up to ½sec, which is the time interval between successive pins on the main pin wheel memory. The information in the cold-cathode store is transferred to thyratrons which operate pin setting magnets upon being supplied (at the correct time for pin setting) with an anode pulse provided by a switch operated by a cam driven in synchronism with the pin-wheel memory.

(2) *The main pin-wheel memory* consists of twenty-four separate pin wheels on a common shaft, each pin-wheel being associated with a particular key on the operator's keyboard. The main pin-wheel memory stores the keyed information in code form while the associated letter travels between the presentation unit and the synchronizing gate. The letter is then held at the synchronizing gate until the pins on the main pin memory which have been set for it operate 'reading-off' switches. The letter is held at the synchronizing gate for a maximum of ½sec, the waiting period at the gate for any particular letter being equal to the time that the coded information was held in the cold-cathode store.

(3) *Thyratron translation field.*—In addition to controlling the synchronizing gate, the 'reading-off' switches also operate into the thyratron translation field which consists of a 12 by 12 matrix of thyratrons. Thyratrons with two grids are used and each thyratron has one grid in a column and one grid in a row. The left-hand part of the key combination acts on the grids in a row, and the right-hand part of the key combination acts on the grids in a column. By this means, for any given two-key combination, only the thyratron at the crossing point of the row and column associated with the two keys is fired. This performs the translation from code form to an individual selection.

(4) *Divertor pin-wheel memories.*—In the anode circuit of each thyratron is a pin setting electro-magnet associated with a stacking box divertor pin-wheel, and a pin setting electromagnet associated with the appropriate 'level divertor' (the divertor which routes the letter to the required horizontal conveyor) pin-wheel.

Thus the letter is released from the synchronizing gate at the same time as a pin is set on the appropriate level divertor pin-wheel memory and on the stacking box pin-wheel memory. From this time the letter travels in synchronism with the pin-wheels so that as the letter approaches the level divertor and the box divertor these are operated mechanically by the pins which have been set to route the letter firstly on to the correct horizontal level and then into the correct stacking box. The pin-wheels associated with the stacking boxes have too small a storage capacity to 'mark' a letter from the synchronizing gate to the most distant stacking box and it is necessary for the setting of pins on the more distant stacking box pin-wheels to be delayed until the letter has progressed some distance from the synchronizing gate. In order to provide this delay, a second bank of 'reading-off' switches is provided on the main pin-wheel memory which operates (about five seconds after the first set of reading-off switches) into a second 12 by 12 matrix of thyratrons. These thyratrons operate pin setting magnets to set pins on the memory wheels associated with the more distant stacking boxes.

Although basically the control system of the machine is capable of sorting to 144 different selections, only 120 selections are used on this machine.

In addition to the 24 'code' keys on the operator's keyboard a 'Feed' key and a 'Cancel' key are also provided. The feed key is thumb operated and causes the presentation unit to perform one cycle of operation, feeding a letter into the conveyor system and routing it to a permanently open 'overflow' box at the end of the top horizontal conveyor. This facility prevents the operator from being delayed by having to remove by hand any letter which he is unable to sort. The 'Cancel' key is provided to enable the operator to cancel his previous keying operation to divert a letter to a separate box should he realize that he has keyed incorrectly.

Ferrites at Microwaves

By P. E. V. Allin*, B.A.

This article explains the principles of some of the non-reciprocal microwave ferrite devices that are now becoming available to systems engineers. Some of the applications of these devices are also discussed and their performance given.

DURING recent years the use of ferrites at microwaves has made possible many new, useful and interesting devices. Practical passive non-reciprocal elements have become available and in fact, ferrites have properties which lead to components exhibiting reciprocal or non-reciprocal phase change, attenuation or field displacement.

Gyromagnetism

Ferrites are ceramics formed by sintering iron oxide with other selected oxides. Among the many possible ferrites some members of the magnesium-manganese and the nickel-zinc systems have attractive microwave properties. The important difference between ferrites and the ferromagnetic metals which makes the former useful at microwaves is their high resistivity. The resistivity of iron is very low, of the order of $10^{-5}\Omega$-cm, whereas a ferrite suitable for microwave purposes has a resistivity of at least $10^{+5}\Omega$-cm. A material with a low resistivity will have a small depth of penetration, or skin depth, at high frequencies; for example the skin depth of most metals at 10kMc/s is about 10^{-4}in. Therefore a microwave field incident on iron would see a reflector, but if the field were incident on a ferrite it would see a magnetic dielectric.

The magnetic properties of a material can be described in terms of spinning electrons and for the present purpose one can consider an electron as a negatively charged sphere which, since it is rotating, will behave like a circulating current and therefore have a magnetic moment. In passing through a ferrite a microwave field can interact with these spinning electrons; it is this interaction which gives rise to the gyromagnetic properties of ferrites.

Since an electron has mass it has angular momentum and will behave in a magnetic field as a gyroscope does in a gravitational field. One can say therefore that the magnetization vector **M**, which is the sum of the magnetic moments of the electrons associated with magnetism, behaves like a gyroscope. If the vector **M** is displaced from equilibrium in a steady magnetic field H_{dc}, it will not re-align itself with the direction of H_{dc}, but will precess about that direction. The frequency of precession is determined by the magnitude of H_{dc}.

If an alternating field is applied at right-angles to the steady magnetic field, the magnetization vector will be driven in a precessional resonance when the driving frequency is the same as the natural resonance frequency. As with all resonant systems a damping term must be included and power is absorbed by the system at resonance. Off resonance, the effective permeability seen by the driving force is a function of H_{dc} and frequency.

Consider a sample of ferrite magnetized to saturation in the $+z$ direction, with an alternating magnetic field applied in an arbitrary direction. The alternating field is assumed to be small compared with the steady field and damping is neglected. One then has:

$$\left.\begin{aligned} \mathbf{H} &= \mathbf{a}_x h_x + \mathbf{a}_y h_y + \mathbf{a}_z (h_z + H_z) \\ \mathbf{B} &= \mathbf{a}_x b_x + \mathbf{a}_y b_y + \mathbf{a}_z (b_z + B_z) \\ \text{and } \mathbf{B} &= \mu_0 \mathbf{H} + \mathbf{M} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots (1)$$

where H_z and B_z are steady fields,

and h_x, b_y etc. are alternating components

$\mathbf{a}_x \mathbf{a}_y$ and $\mathbf{a}_z$ are unit vectors

μ_0 is the permeability of free space.

The rationalized m.k.s. system of units is used.

An electron of magnetic moment **m** and angular momentum **J** under the influence of a magnetic field experiences a torque $\mathbf{m}\times\mathbf{B}$ where **B** is the magnetic induction it encounters. The component of **B** due to magnetization is parallel to **m** and drops out of the vector product. Therefore the torque

$$\mathbf{T} = \mathbf{m}\times\mathbf{H} \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

This torque accelerates the electron according to the usual law.

$$\mathbf{T} = (d\mathbf{J}/dt) \quad \ldots\ldots\ldots\ldots\ldots\ldots (3)$$

By combining equations (2) and (3) and multiplying by the density of electrons we get

$$(d\mathbf{M}/dt) = \gamma\mathbf{M}\times\mathbf{H} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

where $\gamma = (m/J)$, the gyromagnetic ratio of an electron. Note that since an electron has a positive mass and a negative charge γ is negative; and has a value of 22×10^4rad/sec/A/m (or 2·8Mc/s/oersted).

A more detailed analysis would include a damping term in equation (4).

By splitting equation (4) into its components and solving, one gets

$$\left.\begin{aligned} b_x &= \mu h_x - jkh_y \\ b_y &= jkh_x + \mu h_y \\ b_z &= \mu_0 h_z \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ldots\ldots (5)$$

where $\mu = \mu_0 - \dfrac{\gamma M \omega_0}{\omega_0^2 - \omega^2}$

$$k = \frac{\gamma M \omega}{\omega_0^2 - \omega^2}$$

$$\omega_0 = -\gamma H_z \text{ and } j = \sqrt{-1}$$

Equation (5) describes the r.f. permeability of a ferrite saturated in the z direction. The permeability described is not a scalar since b_x depends not only on h_x but also on h_y: in fact equation (5) describes a tensor permeability.

In order to convert the tensor permeability into a more familiar scalar permeability, let us consider circularly polarized r.f. magnetic fields. A positive circularly polarized field vector is defined as a vector which rotates in a clockwise direction when viewed along the direction of the steady magnetic field,

$$\left.\begin{aligned} &\text{i.e.} \quad h_y = -jh_x \text{ defines } h+ \\ &\text{and} \quad h_y = +jh_x \text{ defines } h- \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

By substituting h_+, h_- and b_- in equation (5) one finds that the permeability seen by a circularly polarized field is scalar.

$$\left.\begin{aligned} \mu_+ &= \frac{b_+}{h_+} = \mu - k = \mu_0 - \frac{\gamma M}{\omega_0 - \omega} \\ \text{and } \mu_- &= \frac{b_-}{h_-} = \mu + k = \mu_0 - \frac{\gamma M}{\omega_0 + \omega} \end{aligned}\right\} \quad \ldots\ldots (7)$$

* *Mullard Research Laboratories*

In Fig. 1, μ_+ and μ_- are plotted against H_z for a constant frequency, in the form in which they would appear if a damping term were included in equation (4). $\mu'\pm$ is the real part of $\mu\pm$ and $\mu''\pm$ is the imaginary or loss part of $\mu\pm$.

From Fig. 1 we see that there is a resonance for μ_+ at $H_z = H_{res}$ with its associated loss but no resonance for μ_-. This leads to one type of non-reciprocal device, the resonance isolator, which will be discussed later.

Point-Field Analysis

The solution of Maxwell's equations, particularly for partially filled waveguides, presents mathematical problems which have been solved in only a few cases. However, the

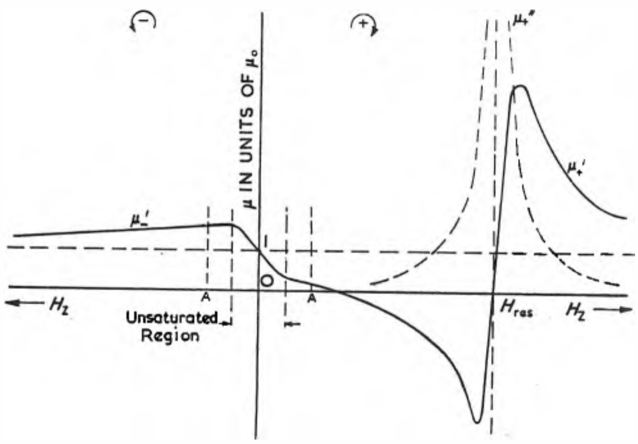


Fig. 1. The variation of μ_+ and μ_- with magnetic field. (After Fox, Miller and Weiss)

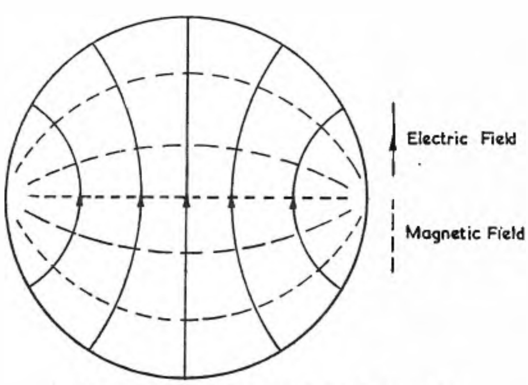


Fig. 2. The TE_{11} mode in cylindrical waveguide

use of a 'point-field analysis' introduced A. G. Fox, together with Fig. 1 is all that will be required to explain and predict the behaviour of many microwave devices. Fox assumed that "the ferrite-loaded waveguide contains a perturbed wave bearing a strong resemblance to one or more of the well-known modes in an empty waveguide." The field which the wave produces in every part of the ferrite can then be determined and the result of gyromagnetic interaction predicted.

Cylindrical Waveguide Components

PHASE CHANGE IN CYLINDRICAL WAVEGUIDE

Fig. 1 readily explains the behaviour of circularly polarized waves in ferrite loaded waveguide. The dominant mode in a cylindrical waveguide, the TE_{11} mode, is shown in Fig. 2. The direction of linear polarization is defined as the direction of the electric field vector at the centre of the waveguide; if this vector rotates at the radio frequency the wave is said to be circularly polarized. Consider a circularly polarized TE_{11} wave incident on a ferrite-filled section of waveguide immersed in an axial magnetic field ($\leqslant H_{res}$). The velocity of propagation of an electromagnetic wave through a medium is a function of the permeability of the medium. Therefore the phase change through the ferrite-loaded section of waveguide will be a function of μ_+ or μ_-. Since ferrites have a large dielectric constant even at microwaves (say 10-20) it is difficult to match ferrite-filled sections to empty waveguide. For this reason and to avoid multimoding, tapered pencils of ferrite placed along the axis of a waveguide are usually employed.

FARADAY ROTATION OF A TE_{11} MODE IN CYLINDRICAL WAVEGUIDE

If a linearly polarized TE_{11} wave is transmitted through a ferrite-loaded cylindrical waveguide immersed in an axial

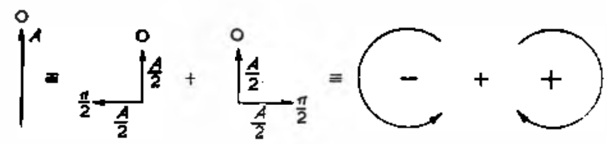


Fig. 3. The equivalence of a linearly polarized wave and two contra-rotating circularly polarized waves

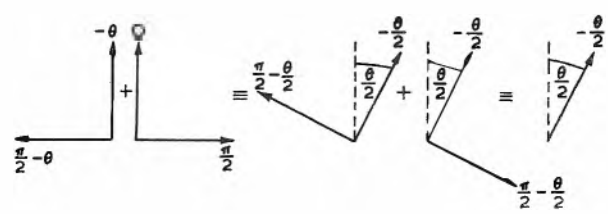


Fig. 4. The equivalence of phase shift of one circularly polarized component of a linearly polarized wave and Faraday rotation

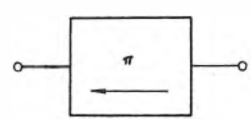


Fig. 5. The circuit symbol of the gyrator

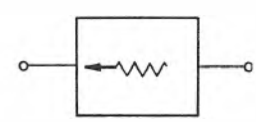


Fig. 6. The circuit symbol of the isolator

magnetic field, the plane of polarization will be rotated. The direction of rotation is independent of the direction of propagation and depends only on the direction of H_{dc}. This effect is known as Faraday rotation. To explain this in terms of Fig. 1 one has to resolve the linearly polarized wave into two equal oppositely rotating circularly polarized waves. Fig. 3 shows the equivalence of a linear wave and two circular waves.

If a small value of magnetic field is used, say that denoted by A in Fig. 1, it can be seen that the effective permeabilities are different for the two components of the linearly polarized wave. The negative circularly polarized wave, in traversing the ferrite section more slowly, lags in phase with respect to the positive wave at the output of the ferrite section. If this lag is, say θ, a rotation of the plane of polarization of $\theta/2$ is obtained, see Fig. 4.

Since the positive and negative waves were defined with respect to the direction of the steady field and not to the direction of propagation, we see that Faraday rotation is independent of the direction of propagation. This fact is very important and is the basis of many non-reciprocal devices.

THE GYRATOR

The gyrator is defined as a transmission element having no phase shift for one direction of propagation and a phase shift of π radians in the opposite direction. The circuit symbol of a gyrator is shown in Fig. 5. An element giving a Faraday rotation of 90° is one type of gyrator.

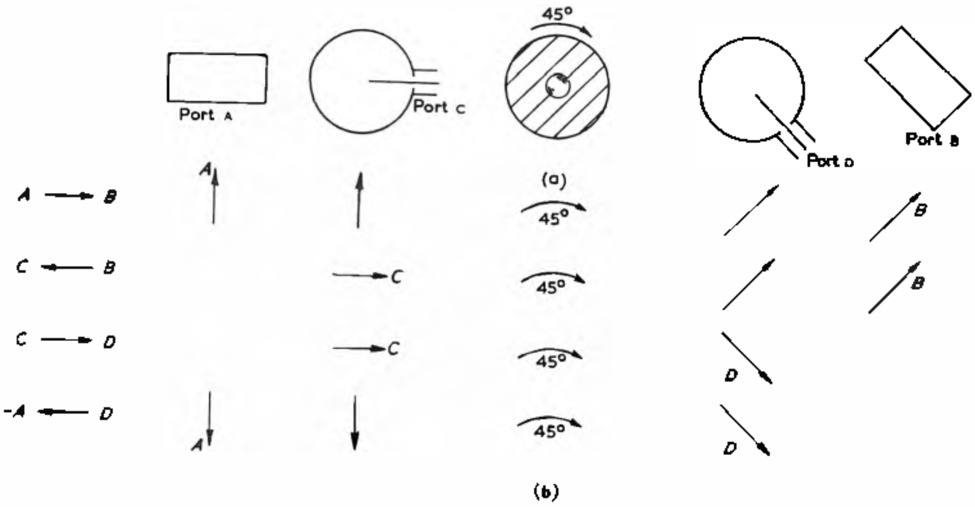


Fig. 8. The Faraday rotation four-port circulator

The Faraday Rotation Isolator

The isolator is an element which attenuates a wave propagating in one direction but not one propagating in the opposite direction. The circuit symbol of an isolator is shown in Fig. 6. In practice there is some attenuation in both directions, for example, one may have an isolator with a forward attenuation of 0·5dB and a reverse attenuation of 20dB. This can be considered as a reciprocal attenuator of 0·5dB in series with an isolator of 19·5dB. An isolator can be constructed from a ferrite section giving a Faraday rotation of 45° and two polarization selective junctions. Such a device is shown and explained in Fig. 7. Fig. 7(b) shows sections taken at intervals along the overall view, Fig. 7(a). The operation of the isolator is explained by the polarization diagram Fig. 7(c). Solid vectors represent waves travelling from left to right and dashed vectors waves travelling from right to left. The wave travelling from left to right is rotated by the ferrite section to an angle such that the rectangular waveguide at the output end of the device will accept it. The wave travelling in the opposite direction is rotated until it is parallel to the attenuating vane and is absorbed; some portion of this wave will be reflected at the first attenuating vane; this portion will traverse the ferrite once more and will be absorbed by the second vane.

Fig. 9. The circuit symbol of the four-port circulator

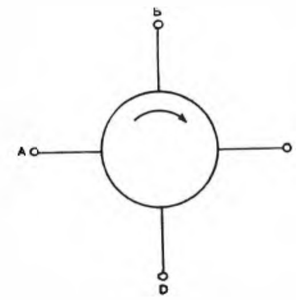
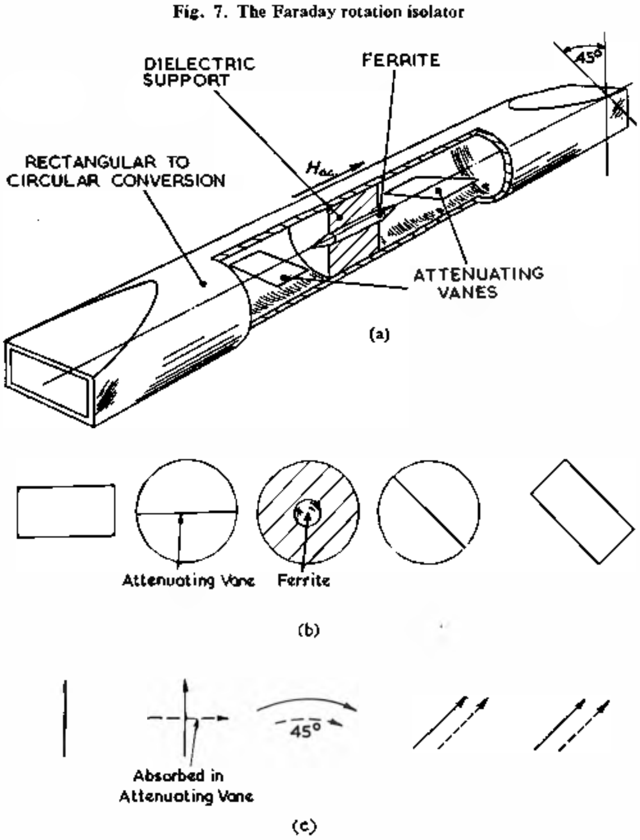


Fig. 7. The Faraday rotation isolator



The Circulator

If the attenuating vanes in the isolator of the previous section are replaced by pick-off transmission lines which accept polarizations at right-angles to those accepted by the circular to rectangular conversions, the isolator becomes a four-port circulator. The operation of the circulator is explained by Fig. 8. As shown by the polarization diagrams in Fig. 8(b), power flows from port to port in the order $A \rightarrow B \rightarrow C \rightarrow D \rightarrow -A$. The circuit symbol of the four-port circulator is shown in Fig. 9. An X-band four-port circulator of the type described above is shown in Fig. 10; the forward attenuation between a pair of ports is 0·2dB and the reverse attenuation 40dB. For special applications the attenuation between one pair of ports can be increased over a narrow bandwidth to about 70dB, but then the reverse attenuation between other pairs suffers slightly. The circulator shown uses a tapered pencil of a magnesium-manganese ferrite.

The Switched Circulator

It is but a small step from the circulator to a waveguide switch. If the field H_{dc} applied to the circulator is reversible power incident on port A can be switched between ports B and D.

Rectangular Waveguide Components

The TE$_{10}$ Mode in Rectangular Waveguide

Rectangular waveguide of the usual dimensions can support waves of only one polarization, hence the effects of ferrite in rectangular waveguide are different from those in cylindrical waveguide. However, point-field analysis and

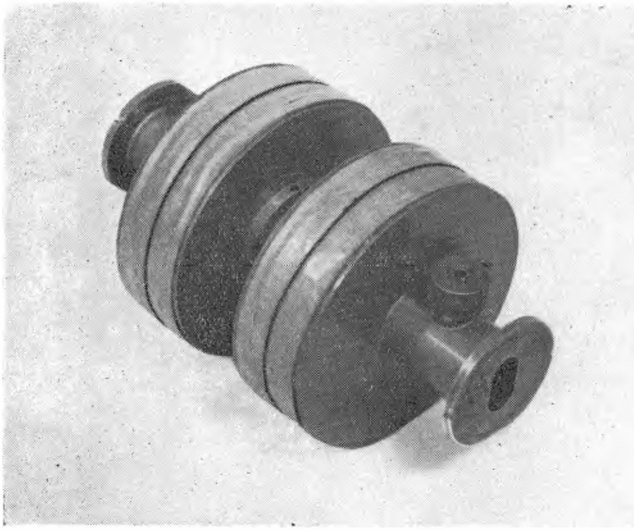


Fig. 10. An X-band circulation

Fig. 1 can still be used. The magnetic field pattern of the TE_{10} mode in rectangular waveguide is shown in Fig. 11. The pattern is viewed through a broad face.

An observer viewing the wave passing a point such as X or Y will see an elliptically polarized magnetic field. At X he will see an anticlockwise rotation for a wave travelling from left to right and a clockwise rotation for the other direction. At Y he will see rotations of the opposite sense to those seen at X. At some point between the centre of the waveguide and a narrow face he will find that the magnetic field is circularly polarized. If a steady magnetic field is applied in a direction parallel to the narrow sides of the waveguide, the circularly polarized waves will be positively and negatively polarized as defined for μ_+ and μ_-.

The Non-Reciprocal Phase Changer in Rectangular Waveguide

If a thin plate of ferrite is placed in a rectangular waveguide at the position where the magnetic field is circularly polarized and subjected to a steady magnetic field parallel to the narrow dimension of the waveguide, as shown in Fig. 12, the effective permeability of the ferrite-loaded section will depend on the direction of propagation. Therefore the phase-shift through the section of waveguide containing the ferrite will be different for the two directions of propagation. If this difference is adjusted by changing H_{dc}, for example, to π radians the element is a gyrator (with some reciprocal phase shift).

The Transverse-Field Resonance Isolator

So far we have been concerned with elements using magnetic fields small compared with the field required for resonance. This leaves us the interesting region of gyromagnetic resonance to exploit. As seen in Fig. 1 there is a resonance for μ_+ and not for μ_-. Thus if the steady magnetic field applied to the non-reciprocal phase changer of the preceeding section was increased to the resonant value, about 2×10^5A/m at X-band, the ferrite section would exhibit non-reciprocal attenuation due to the loss associated with resonance. An X-band resonant isolator using 'Ferroxcube B5', a nickel-zinc ferrite, is shown in Fig. 13; the forward attenuation is 0·5dB and the reverse attenuation 20dB.

The Gyrator as a Building Block

As an example of the way in which gyrators can be used as building blocks for more complicated devices, we will consider a circulator using one gyrator and two hybrid junctions.

The hybrid or magic-T is shown in Fig. 14. If microwave power is fed into arm C it will split, in-phase, and emerge from arms A and B; conversely if equal amounts of power are fed into arms A and B, in-phase, they will combine and emerge from arm C. Power fed into arm D will split, in anti-phase, and emerge from arms A and B; equal amounts of power fed into arms A and B in anti-phase will combine and emerge from arm D. There is no direct connexion between arms C and D.

It can be seen that if a gyrator is connected between two hybrid junctions as shown in Fig. 15 the result will be a four-port circulator with power flow $C \rightarrow C' \rightarrow D \rightarrow D' \rightarrow -C$.

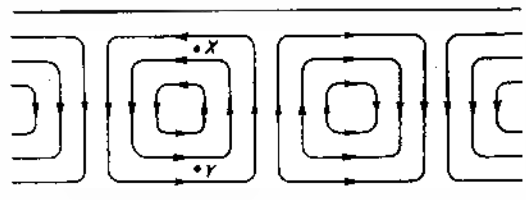
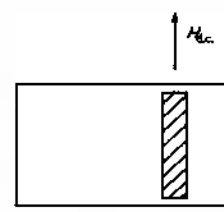


Fig. 11. The pattern of the magnetic field of the TE_{10} mode in rectangular waveguide

Fig. 12. An asymmetrically loaded waveguide



Other Methods of Obtaining Non-Reciprocal Properties

Not all the ways, by any means, in which non-reciprocal phase shift or attenuation can be achieved using ferrites at microwaves have been discussed; and non-reciprocal field displacement has not been dealt with at all.

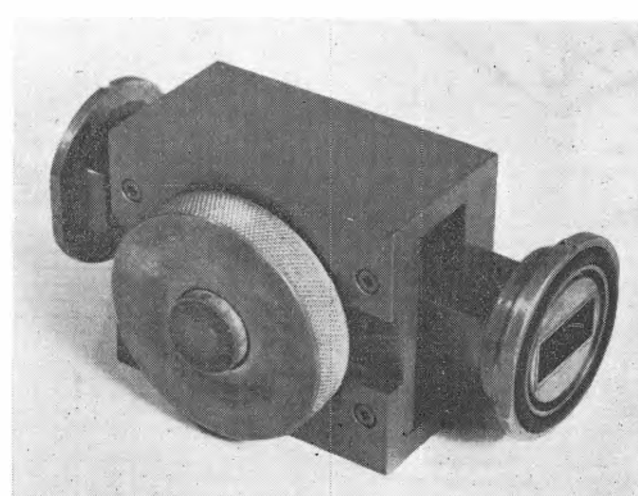
Applications

There are many possible applications of these devices and microwave engineers will find it easy to see ways in which ferrites can be incorporated advantageously in new equipment or can improve existing equipment.

Isolators

Isolators with a front to back ratio of 30dB are easily

Fig. 13. An X-band resonant isolator



obtained and ratios of up to 200 or so are possible. A typical simple resonance isolator might have an attenuation of 0·5dB in the forward direction and 20dB in the reverse direction over a bandwidth of 500Mcs at X-band. With the field supplied by permanent magnets these isolators are compact devices requiring no adjustment. If the field is adjustable the centre frequency can be chosen to be anywhere in the band for which the waveguide is designed. An obvious use is to replace attenuating pads used to protect klystron oscillators in measuring equipment. In order to protect a klystron against variations in loading it is necessary to isolate it from the load; this has been done in the past by attenuators of up to 20dB. If the 20dB attenuator is replaced by an isolator with 1dB forward attenuation and 39dB reverse attenuation, the isolation of the load is the same as before, but the useful output of the klystron is increased by 19dB or approximately 80 times in power.

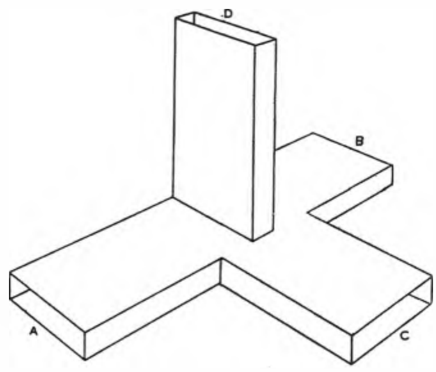


Fig. 14. A waveguide hybrid junction

If designed to handle high power, isolators can be used to protect magnetrons against long-line-effect pulling in radar systems, etc.

Circulators

Circulators have many possible applications one of which is in t.r. systems. If a three-port circulator is connected as shown in Fig. 16, it can replace the normal gas cell type of t.r. system. It can be used in low power transmitters where the peak power of the transmitting valve is insufficient to ignite a gas cell. The simple system shown in Fig. 16 can only be used if the aerial is a very good match, since reflexion from the aerial will pass to the receiver. This can be overcome by putting a switched circulator on the terminal marked 'Receiver'.

Valve Tuning

An interesting application of phase-changing by ferrites is the tuning of microwave oscillators. If a ferrite is placed in the resonator of a reflex klystron the electrical size of the resonator and therefore the frequency of oscillation can be made to be a function of an applied d.c. magnetic field.

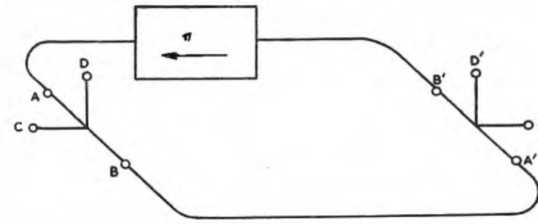
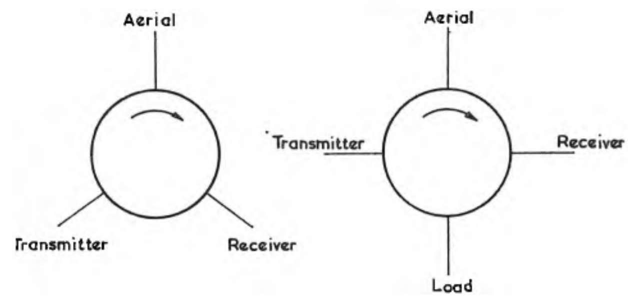


Fig. 15. A four-port circulator using a gyrator and two hybrid junctions



(a) Three-port (b) Four-port

Fig. 16. A circulator used to make a simple *t.r.* network

Acknowledgments

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A Colour Vectorscope

The extensive amount of research work which is taking place into colour television techniques has underlined the need for specialist test equipment for use in that field.

To this end, Marconi's Wireless Telegraph Company Ltd., have produced a new test instrument known as a Vectorscope, which is specially designed to display the chrominance component of the N.T.S.C. type of colour television signal.

The Vectorscope has proved to be of considerable value, not only for the correct setting up of coding systems of the N.T.S.C. type, but also for measurements of amplitude and phase relationship in a colour signal at any point in a television distribution system. A further application lies in the monitoring of actual colour camera signals, since its display gives an objective indication of the hue and saturation of the colour components. It can thus help in matching the characteristics of colour cameras and prove useful as an aid to programme directors in choosing colours for costumes and backgrounds.

The chrominance information is carried on a sub-carrier which is modulated in amplitude, representing the colour saturation, and in phase, representing the hue. The display is presented on a cathode-ray tube, the radial distance of the spot from the centre indicating the amplitude modulation or saturation, while the phase or hue is displayed as the angle subtended from a fixed phase reference on the screen.

The Vectorscope employs a pair of quadrature demodulators similar to those used in colour monitors and receivers, and a burst-controlled oscillator for deriving the reference sub-carrier from the colour synchronizing bursts contained in the signal under test. The outputs of the two demodulators are applied after suitable filtering and amplification to the horizontal and vertical plates of the cathode-ray tube.

When used in conjunction with the colour bar test signal, the Vectorscope produces a pattern of bright dots corresponding to the tips of the various colour vectors and a pattern of lines corresponding to the transitions between the colours. 'Boxes' indicating phase and amplitude tolerance limits are drawn on a transparent scale to provide a very convenient indication of the quality of the signal, although it must be remembered that these tolerances refer only to the sub-carrier information, since the luminance information has been removed by a filter.

Short News Items

The University of Birmingham Department of Electrical Engineering are holding a Graduate Course in Information Engineering commencing in early October for twelve months. The first nine months are spent in studying, by means of lectures, reading, tutorials and laboratory work, for a formal written examination. The remaining three months are occupied by work on an individual project, the report on which is the basis of the final assessment of the student's work. This advanced course is available to Honours graduates in electrical engineering who have preferably had a year or two of engineering experience outside the university. Students with other qualifications may also be considered for admission to the course. In accordance with the general regulations of the university, graduates who satisfy the examiners for the course will receive the degree of M.Sc. Details of fees, residential accommodation, and of financial assistance which may exceptionally be provided by the university if the student is not supported from other sources, may be obtained on application to the Graduate Course Supervisor, the University of Birmingham, Edgbaston, Birmingham 15.

The British Council is organizing a fortnight's course on Digital Computers to be held from 27 October to 9 November. The basic lectures will be given in London, where visits to various computing establishments will also be arranged. These visits will provide an opportunity not only of seeing computing equipment in use, but also of discussing practical problems with workers in the field. It is hoped that visits may also be arranged to the Universities of Cambridge and Manchester, and to the Computational Laboratory, Birkbeck College, University of London. There are vacancies for 15 members on this course and the fee will be £42. Further details may be obtained from The British Council, 65 Davies Street, London, W.1.

A Summer School in programme design for automatic digital computing machines will be held in the University Mathematical Laboratory at Cambridge during the period 16-27 September. It will be along the same lines as those held in previous years. A detailed syllabus and form of application for admission may be obtained from Mr. G. F. Hickson, Secretary of the Board of Extra-Mural Studies, Stuart House, Cambridge, to whom the completed application form should be returned not later than 15 June.

The Instruments, Electronics and Automation Exhibition is to be repeated in 1958. The dates will be from 16-25 April and the location will again be Olympia.

H.M. The Queen has consented to be patron of the National Radio and Television Exhibition to be held at Earls Court from 28 August to 7 September.

A Conference on Magnetism and Magnetic Materials will be held in Washington, D.C., from 18-20 November this year, by the American Institute of Electrical Engineers, in co-operation with the American Physical Society, the American Institute of Mining and Metallurgical Engineers, the Institute of Radio Engineers, and the U.S. Office of Naval Research. Authors should submit titles of proposed papers by 1 July and abstracts by 15 August. Further information may be obtained from L. R. Maxwell, U.S. Naval Ordnance Laboratory, White Oak, Silver Spring, Maryland, U.S.A.

An International Conference on Ultra High Frequency Circuits and Antennas will be held in Paris from 21-26 October. Further information may be obtained from Congres Circuits et Antennas Hyperfrequences, Societe des Radioelectriciens, 10 Avenue Pierre-Larousse, Malakoff (Seine), France.

Marconi's Wireless Telegraph Co Ltd have received an order from the Danish Posts and Telegraphs Department for transmitting and aerial equipment to a value of £90 000 for three new television stations which are shortly to be built at Aalborg, Vestjylland and Naestved. The order, which includes monitoring and test equipment and flying spot slide scanning units, was obtained for Marconi's by their agent in Denmark, Sophus Berendsen Ltd.

Pye Telecommunications Ltd have secured a contract from the Indian Department of Civil Aviation for communication equipment for a modernization programme for thirty-nine Indian airports. The equipment is to provide Aerodrome Approach Control and Runway Control communication.

The formation of a new company to manufacture transistors and other semiconductors in England was announced recently. To be known as Semiconductors Ltd, the new company has been formed by the Plessey Co Ltd and Philco Corporation of U.S.A.

The Rubber and Asbestos Corporation of Bloomfield, New Jersey, U.S.A. announce that they are now in a position to effect shipment of a new grade of Plymaster Adhesive Coated Copper specifically developed for use on epoxy glass laminates to produce a "copper clad" which has outstanding performance during and after exposure to silver and gold cyanide plating baths. This material is known as Plymaster Type C. Further information about this product may be obtained through agents in the United Kingdom, Omni (London) Ltd, 35 Dover Street, London, W.1.

The Radio Industry Council has announced that representatives from 28 countries were among the 24 000 who visited the Radio and Electronic Component Show which was held in April for four days.

The British Scientific Instrument Research Association announce that at a recent council meeting Mr. J. E. C. Bailey, C.B.E., was elected Chairman of the Association, his term of office to extend until July 1960. Mr. Bailey is Chairman and Managing Director of Baird & Tatlock (London) Ltd.

Vickers Ltd have announced that Mr. C. L. Old, Principal of the College of Technology, Wolverhampton, has been appointed Group Education Officer. His primary responsibility will be the co-ordination of all the educational and training activities of the Vickers Group Companies. It is expected that he will take up his appointment at the beginning of September.

Errata. The caption under Figs. 11-14 on page 222 of the May issue (Part 2 of article "Dynamic Methods of Testing Semi-Conductor Rectifier Elements and Power Diodes" by A. H. B. Walker and R. G. Martin) should read as follows.

"FORWARD CHARACTERISTICS
A = Full dynamic test.
B = Short-circuit bridge test. High impedance supply. Mean voltage measured.
C = Short-circuit bridge test. Low impedance supply.
D = Full wave rectified sine wave test.
E = Half sine wave test.
F = Loaded bridge test."

The Reverse Characteristics remain unaltered.

On page 250 of the May issue, the publishers of the book "Transistors in Radio and Television" are the McGraw Hill Book Co Ltd and not Chapman & Hall Ltd, as stated.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

Circuit Magnification Measurements

DEAR SIR,—The circuit shown in Fig. 1 enables Q, or circuit magnification measurements, to be made when used in conjunction with a valve-voltmeter and a signal generator having an output of 1V.

It consists, essentially of a high slope double triode, a calibrated variable capacitor for tuning the inductance under test, and some close tolerance silvered mica capacitors.

Both triodes act as cathode-followers and allow the use of a single valve-voltmeter for calibration and Q measurement without impairing the circuit. The injection system is not new.

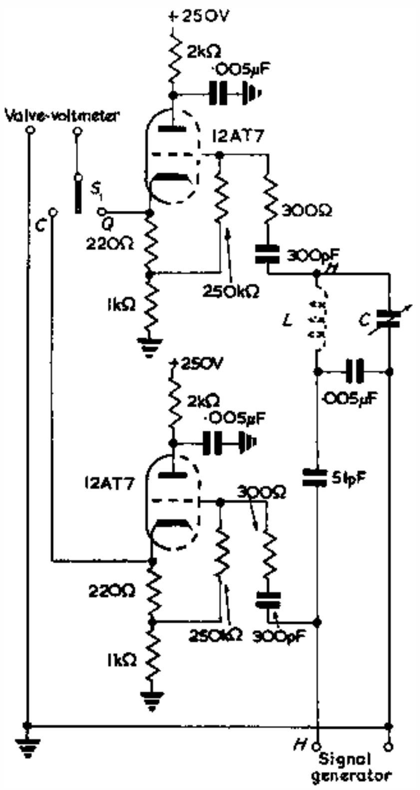


Fig. 1. The circuit for Q measurement

The variable capacitor can be calibrated with the aid of a battery-fed capacitance bridge and a push-pull input cathode-ray oscillograph. The ·005μF capacitor is taken into account.

The cathode-follower may be balanced by removing the variable capacitor from the circuit and connecting the H terminals of 'L' and the signal generator together. Balancing is done by switching the valve-voltmeter from C to Q and adjusting the cathode resistors.

With a signal generator output of exactly 1V the circuit magnification is 100 times the meter reading. Thus one can read a 'Q' of 500 on the 5V scale.

This arrangement is very useful for smaller laboratories and for production testing since it does not necessitate the constant use of expensive Q-meters and the operation is very simple.

Yours faithfully,

I. LUDCKX,

Antwerp,
Belgium.

Step Function Pulse Technique for Ultrasonic Measurement

DEAR SIR,—While agreeing with Mr. W. V. Richings that the article on the above subject in your February issue ignored the resonance method of thickness measurement, we cannot accept the opinion that configuration of the object to be tested is not a serious problem.

The history of the development of the method illustrates this. For a number of years we had been receiving enquiries from organizations who, for many reasons were unhappy with the resonance method. According to our records: corrosion pitting, internal flaws, wedged and radiused shapes provided a number of situations in which the use of resonance methods was found difficult if not impossible. Finally a demand for the inspection of very small flaws in multi-layer bonded plate led to the step function technique solution.

We would not argue that the resonance method is now outmoded but simply state that a method of ultrasonic inspection has now been developed to tackle a wider field of problems than hitherto.

Yours faithfully,

R. G. S. CLARKE,

Ultrasonoscope Co. (London) Ltd.,
Brixton Hill,
London, S.W.2.

The correspondent replies:

DEAR SIR,—Your correspondent, Mr. R. G. S. Clarke, may be interested to learn that further development of the resonance method of thickness measurement has largely overcome the early limitations to which he refers.

Thus by using small area crystals, higher frequencies and more sensitive indicating equipment, measurements can now be made on wedged and radiused shapes and small flaws detected under conditions which were previously quite impossible.

Corrosion and pitting have never been a very serious problem, for example portable equipment which has been available for over ten years is widely used for checking rates of corrosion of pipes and storage vessels in oil refineries.

The major advantage of the resonance method is the discrimination of 0·05 per cent to 0·1 per cent which can be achieved with the latest circuit designs. The accuracy of thickness measurement is set by the variation of the velocity of propagation of the ultrasonic wave between samples of a given material. This is, of course, a fundamental limitation of any ultrasonic method.

Improvements to both the resonance and pulse method of ultrasonic inspection are continually being made and the step function pulse technique described in your February issue is certainly a valuable contribution.

Yours faithfully,

W. V. RICHINGS,

Chief Development Engineer,
Dawe Instruments Ltd.,
Ealing.

Output Impedance of Anode-Follower Circuits

DEAR SIR,—If it is not too late I would comment upon a letter from Mr. D. McDonnell published in the December 1954 issue.

The writer of this letter sought to prove that the expression $Z_0 \simeq 2/g_m$ for an anode-follower was inaccurate and that the expression $Z_0 \simeq R/A\beta$ was more appropriate; the circuit in question is reproduced for convenience (Fig. 1).

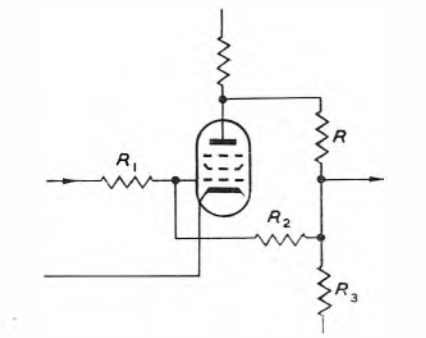


Fig. 1. The circuit analysed

A subsequent letter from Mr. E. F. Good appeared in February 1955 accepting this result; the correspondence culminated in May 1955 with a letter from Mr. A. W. Ward indicating that $Z_0 \simeq 2/g_m$ was correct for the more usual arrangement having $R_1 = R_2$ and with the output taken from the anode.

Mr. McDonnell's result is useful in drawing attention to the modification of the effect of R on output impedance; it is, however, inaccurate in that, by omission, it is assumed that the output impedance of the amplifier proper is zero.

It may be shown, quite readily, that the output impedance at the anode is, making the usual assumptions,

$$Z_{01} = 1/g_m\beta$$

Hence when $\beta = 0{\cdot}5$, the usual paraphase case,

$$Z_{01} = 2/g_m$$

Mr. McDonnell gives

$$Z_{02} = R/A\beta$$

Combining

$$\begin{aligned} Z_0 &= Z_{01} + Z_{02} \\ &= 1/g_m\beta + R/A\beta \\ &= 1/\beta\,(1/g_m + R/A) \end{aligned}$$

Putting

$$A = g_m R_L$$

$$Z_0 = (1/g_m\beta)\left\{1 + \frac{R}{R_L}\right\} \quad \ldots\ldots\ldots(1)$$

$$= R/g_m\beta\left\{\frac{1 + R_L/R}{R_L}\right\} \quad \ldots\ldots\ldots(2)$$

Equations (1) and (2) state the general case.

When $R = 0$

$$Z_0 = 1/g_m\beta = Z_0 \quad \ldots\ldots\ldots(3)$$

When $R \gg R_L$

$$Z_0 = \frac{R}{g_m\beta\ R_L} = Z_{02} \quad \ldots\ldots\ldots(4)$$

Equations (1) to (4) indicate the interesting properties of this circuit with respect to output impedance.

It is not always appreciated that, when $R = 0$, this circuit has an important advantage over a cathode-follower in that the rise and fall times are substantially equal, being determined by the output impedance. A cathode-follower, of course, has unequal rise and fall times since the output impedance to a negative going change is determined by the load impedance and not by the valve parameters.

Yours faithfully,

T. G. Clark,

East Molesey, Surrey.

Out-of-Channel Radiation from Mobile F.M. V.H.F. Transmitters

Dear Sir,—We would congratulate Mr. Rowles for his excellent article (March 1957) on the subject of adjacent channel interference radiated from f.m. transmitters. This has for some time past been recognized as one of the major factors affecting the close channel spacing of frequencies and legislation has been proposed by the F.C.C. to make compulsory the fitting of suitable low-pass filters.

It is perhaps a pity that this examination was confined to deviations which have been established for some long time and did not embrace deviations of the order of 4 to 5kc/s now to be used for channel spacings of 20 to 30kc/s.

It is somewhat misleading to suggest that unwanted audio sidebands are the only form of energy which is radiated in adjacent channels. When audio sidebands have been reduced by methods suggested by Mr. Rowles the limiting factor becomes the 'sideband noise' radiation which has been described in an I.R.E. paper by Gifford of the General Electric Co., U.S.A.

Measurements show that sideband noise level is only some 70 to 80dB less than the carrier amplitude and the bandwidth it occupies may run into many hundreds of kc/s. This 'sideband noise' which in contemporary transmitters is usually introduced by the phase modulator is now measured as a standard characteristic of f.m. transmitters and is included in specifications written by the U.S. and Canadian RETMA.

The successful elimination of this sideband noise will be a major factor in the close spacing of f.m. channels.

Yours faithfully,

J. R. Humphreys,

Chief Engineer (VHF), Pye Telecommunications Ltd.

The author replies:

Dear Sir,—As Mr. Humphreys appreciates, the maximum deviation appropriate to a channel spacing of, say, 25kc/s depends upon a number of factors. However, employing the same reasoning and criteria as used in the article for the established 'standard' deviations, the necessary post-limiter-filter slope may be found for any specified deviation.

For example, the sideband envelope of a signal frequency modulated to $\pm 4\frac{1}{2}$kc/s deviation by a 3 000c/s pure tone intersects the -93dBW level at approximately 23kc/s from the carrier. At this sideband frequency the power envelope of a 10dB-clipped 3 000c/s tone would be some 57dB greater.

In order to suppress the effects of these unwanted sidebands, the post-limiter low-pass filter requires a slope, extending from the knee at 3 000c/s, given by:

$$\alpha_0 = \frac{-57}{3{\cdot}33\log(23/3)} = -19\tfrac{1}{2}\text{dB/octave.}$$

While this figure affords a useful guide to the greater demands of closer-channel working in this particular respect, factors with which it is difficult to deal theoretically in quantitative terms become of increased importance. Resort has to be made to practical experiment, with consequent loss of generality.

Yours faithfully,

A. L. Rowles,

Marconi's Wireless Telegraph Co. Ltd Chelmsford.

A Transistor Demodulator

Dear Sir,—With reference to the article by Mr. H. Sutcliffe, in the March issue of your journal a transistor demodulator similar to a commonly employed valve demodulator was developed which has the advantage of not requiring d.c. supplies to the collectors. The circuit (Fig. 1) uses two GET 1 point contact transistors. Junction transistors were not easily available at the time of the development, but the circuit could be suitably modified for modern transistor types.

For large signals the base-collector junction behaves like a diode whose forward direction is from collector to base, and the emitter-base junction behaves like a diode whose forward direction is from emitter to base. If an a.c. voltage is applied between collector and

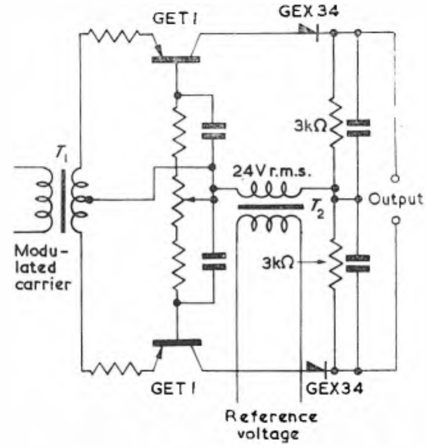


Fig. 1. Demodulator circuit using point contact transistor

base, transistor action occurs for the collector negative with respect to the base. To prevent large pulses of current flowing during the other half cycle when the collector is positive with respect to the base, a crystal diode is inserted in the collector lead with the appropriate orientation.

During the negative half cycle of the reference voltage, the transistor amplifies the current in the emitter lead. If the transistors are exactly alike no output appears if there is no in-phase voltage in the signal windings of the transformer T_1. To compensate for variations in transistors, a by-passed resistor is placed in the base leads to adjust the operating points of the emitters.

Yours faithfully,

N. G. Davies,

London. W.7.

The author replies:

Dear Sir,—Freedom from d.c. power supplies is the valuable feature of the circuit described by Mr. Davies, while in my article the emphasis was directed towards a circuit which was independent of transistor characteristics and required very little signal power. A combination of

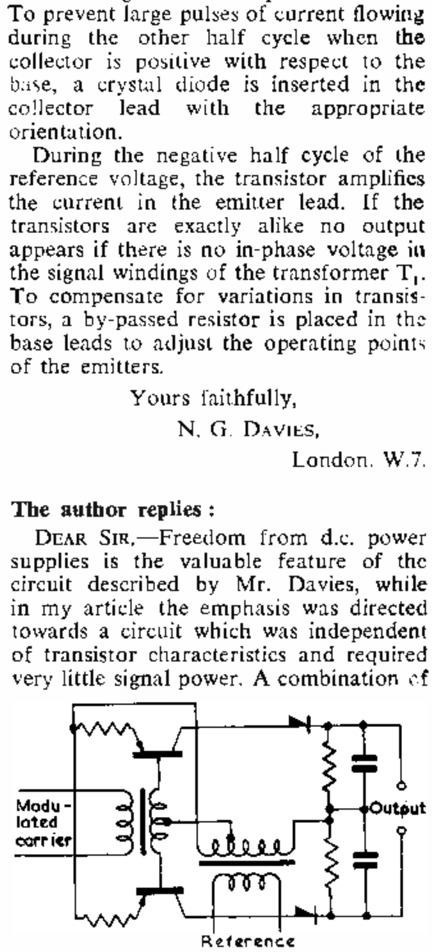


Fig. A. Re-arrangement of circuit for junction transistors

both approaches leads to the circuit shown in Fig. A in which the signal feeds the bases of a pair of junction transistors. The arrangement appears on paper to possess the advantages of both the original designs and I look forward to a practical trial.

Yours faithfully,

H. Sutcliffe,

Department of Electrical Engineering, University of Bristol.

BOOK REVIEWS

Handbook of Semi-Conductor Electronics

Edited by Lloyd P. Hunter. 400 pp. 120 figs. Royal 8vo. McGraw-Hill Book Co. Ltd. New York and London. 1956. Price 90s.

MR. HUNTER opens his preface by stating "In this book an attempt is made to collect in one place all the major principles of the field of semiconductor electronics." He and his co-authors have certainly produced a book which contains a lot of useful material but in parts of the book some of the major principles have been omitted and a lot of minor detail included.

Part I written by Mr. Hunter deals with the physics of transistors, diodes and photocells.

After giving a brief description of all the known types of point contact and junction transistors, electronic conduction in solids is described by comparing the energy levels in insulators, conductors and semiconductors. The effect of the addition of donor and acceptor impurities to the semiconductor leads to a description of rectification in solids, of which the major portion is devoted to the pn junction.

The next section on transistor action is devoted mainly to the junction transistor triode, but also includes a description of the action of other transistors such as the field effect transistor and the junction transistor tetrode. It is in this section that the first serious omissions occur. The first is that the effect of space charge layer widening described by Early, which is the major factor determining the small signal output impedance of the transistor, is not discussed. The second is that only the approximate equation for the alpha cut off frequency is given and no mention is made of the more exact transmission line equations. The third is that no mention is made of noise in transistors.

The final section in Part I is a brief but useful introduction to photoconductivity and photovoltaic cells.

Part II, entitled "Technology of Transistors, Diodes and Photocells," gives a detailed description of the problems facing a manufacturer of semiconductor devices and their practical solution. Mr. Priest deals with the preparation of the material from the source to the single crystal used for the fabrication of the transistor. Mr. Dunlop Jr. then describes the preparation of pn junctions by rate growing, alloying and diffusing. Mr. Sittner discusses metal to semiconductor contacts including the ohmic contact. Finally there is a description of encapsulation techniques by J. B. Little and a number of illustrations of device designs by L. P. Hunter.

It is unfortunate that the reader is then plunged straight into Part III on circuit design and applications without any section giving the tie up between the physics and the applications. A large number of equivalent circuits are used in Part III without any indication of how they are linked together by the transistor.

The first section of Part III is a very detailed discussion by H. J. Woll on the design of low frequency amplifiers. He uses the h parameter system for analysis and bases his designs on the low frequency T equivalent circuits. This is followed by a description of h.f. and video amplifiers by J. B. Angell in which the commonly used hybrid π common emitter equivalent circuit is omitted.

In the next section on directly coupled amplifiers C. Hurtig points out the temperature dependence of d.c. coupled transistor amplifiers and gives some methods of compensation. He briefly mentions chopper type amplifiers and describes some constant voltage power supplies using combinations of pnp and npn transistors. E. Eberhard then gives a general description of a number of types of oscillator employing point contact and junction transistors.

The first part of the final section on switching circuits by J. C. Logue is devoted to point contact transistors. His approach is based on the negative resistance characteristic and how this may be modified by catching diodes. No reference is made to the excellent work on the design of point contact switching circuits by F. C. Williams, G. B. B. Chaplin and E. H. Cooke-Yarborough. The remainder of the section is confined to the use of junction transistors in computer building blocks using d.c. type logic. All types of building blocks are discussed and the circuits are based on d.c. coupled amplifying gates and invertors together with the Eccles-Jordan circuit.

The final section of Part III describes circuits using special semiconductor devices including the double base diode, the field effect transistor and the tetrode transistor.

Part IV entitled "Reference Material" contains a section on graphical analysis of non-linear circuits, and a section on matrix methods both by H. Fleisher. Then came two sections on measurements by G. Knight Jr. and J. C. Logue and the book ends with a really excellent bibliography compiled by W. A. Kalenick.

E. WOLFENDALE.

Communication Engineering

By W. L. Everitt and G. E. Anner. 644 pp. 349 figs. Medium 8vo. 3rd Edition. McGraw-Hill Book Co. Ltd. New York and London. 1956. Price 67s. 6d.

THERE can be few communications engineers whose book-shelf does not contain Everitt's *Communication Engineering*. First published a quarter of a century ago it was revised five years later and has since established itself as a classic among text-books. With the co-operation of Professor G. E. Aner it has been modified in an attempt to bring it into line with current literature in the field and a new edition is now available.

As a solution to the problem of what to omit and how much to add in the new edition the authors have concentrated on the fundamentals of linear network theory pertinent to the science of electrical communication. Much of the first half of the previous edition can still be recognized and many of the original chapter headings have been retained. Among the major changes is a more extensive treatment of the mathematical formulation and solution of linear network problems. The approach is that of the Fourier integral and all networks are specified in terms of the sinusoidal steady-state. The circuital and nodal methods of setting up network equations are explained, but no reference is made to the case of networks with non-planar linear graphs. Nodal analysis by the datum-node method incurs no difficulty when the network graph is non-planar, but this is not generally true for circuital analysis. Similarly the reader is left to discover for himself how to deal with magnetically coupled branches when using nodal analysis.

The chapter on resonance now includes a more general treatment of reactance networks and an account of Cauer's extension to Foster's reactance theorem has been introduced. Bridge networks now occur earlier in the book and a detailed analysis of the parallel-T network has been added. Wave-filter design is still dealt with in the classical manner although passing reference is made to the modern approach. A new chapter is devoted to iterative 2-ports; the discussion of transmission lines has been extended to include lines of low loss; the treatment of impedance transformations has been considerably expanded; and the chapter on equalization now contains a section on delay equalizers. In the previous edition the latter half of the book was concerned with the use of the thermionic valve in communication systems, but very little of this has survived the revision. The valve as a linear amplifier is the subject of only one chapter, but both time and frequency responses are considered. Like the previous edition, the book concludes with a chapter on electromechanical coupling which is substantially the same as the original.

In the preface the authors indicate that the two principal topics of the book are the fundamentals of linear network analysis and synthesis. Apart from the emphasis on the sinusoidal steady-state the techniques of modern network analysis are fairly well represented, although no great use is made of them in the study of specific systems, but the book contains little that can properly be called modern network synthesis. The Foster-Cauer method of realizing driving-point reactance functions was the first significant contribution to modern network synthesis and is the only method mentioned in the book. No indication is given of recent developments in the subject by such workers as Brune, Gewertz, Darlington, Oono, Bayard, Belevitch, Fialkow, and Gerst.

The standard of the book is extremely high and there can be no doubt that it will appeal to communications engineers of the old school, but a young student preparing for a communications engineering career could make a better choice.

S. R. DEARDS.

Television Engineering Principles and Practice Volume 3. Waveform Generation

By S. W. Amos and D. C. Birkinshaw. 226 pp. 132 figs. Demy 8vo. Iliffe and Sons Ltd. 1957 Price 30s.

THIS is the third volume of a textbook on television engineering written by members of the BBC Engineering Division, primarily for the instruction of the Corporation's staff. The work is intended to provide a comprehensive survey of modern television principles and practice. It gives the application in television of sinusoidal, rectangular, sawtooth and parabolic waves and shows the mathematical relationship between them. The main part of the text is devoted to the fundamental principles of the circuits commonly used to generate such signals.

Volume 1 deals with fundamental television principles, camera tubes, television optics and electron optics. Volume 2 describes the fundamental principles of video-frequency amplifiers, and Volume 4, which will complete the series, will deal with a wide range of circuit techniques.

Radio Telemetry

By M. H. Nichols and L. L. Rauch. 461 pp. 90 figs. Demy 8vo. 2nd Edition. John Wiley & Sons, Inc, New York. Chapman & Hall, Ltd. London. 1957. Price 96s.

THIS is the second edition of a book originally published for the United States Air Force. It contains a comprehensive treatment of the basic theory of radio telemetry together with a review of current equipment.

The book has four parts, the first with 216 pages containing a semi-qualitative presentation of the fundamentals. After an introduction giving the historical background, there is a chapter on general methods introducing the concept of multiplexing. The third chapter considers noise and errors dividing them into environmental noise, due to acceleration, temperature changes, etc., and inherent noise, due to radio link noise, cross-talk, friction in pick-up devices, etc. Methods of reduction of environmental noise are briefly discussed here, inherent noise being dealt with in detail elsewhere in the book. The fourth chapter first considers regular pulse sampling of a continuous function without loss of information. Then follows a discussion on the information capacity of a single channel and its dependence on noise and overload levels and on bandwidth. The information capacity is determined for different types of continuous and pulse modulation. A comprehensive treatment of cross-talk, noise improvement, bandwidth, etc., for different methods of multiplexing and modulation is given in the next four chapters which contain a minimum of mathematics, the detailed analysis being confined to parts 2 and 4. A chapter on interpretation of the output of multiplexing systems is followed by a brief review of the fundamentals of radio links with particular emphasis on aspects of importance for air-to-ground links. After a comparison of methods with reference to specific telemetry problems, the first part ends with a review of instruments used in flight testing of rockets and aircraft.

The second part with 34 pages comprises two mathematical chapters which may conveniently be omitted by non-mathematical readers. The first deals with modulation and multiplexing and the second with analysis of the modulated signals.

Part 3 with 112 pages describes systems in recent and current use. Liberal use is made of photographs, waveforms and circuit diagrams.

Part 4 with 74 pages contains the appendices, giving detailed mathematical treatment of various topics arising throughout the book. An extensive bibliography is followed by a glossary of terms and a full index.

The book is well printed with clear figures and follows a logical pattern which enables this complex subject to be readily followed. It forms a first-class text book for those entering the telemetry field and a valuable reference book for those already so engaged. In addition the book should not be overlooked by those interested in the allied field of speech multiplex systems.

D. W. ELSON.

Electronic Musical Instrument Manual

By Alan Douglas. 247 pp. 216 figs. Demy 8vo. 3rd Edition. 1957. Sir Isaac Pitman & Sons, Ltd. Price 35s.

THE third edition of this authoritative work has been enlarged to include the latest circuits and constructional details of British, German, French and American musical instruments, in addition to many new experimental ideas for the amateur constructor.

V.H.F. Radio Manual

By P. R. Keller. 216 pp. 194 figs. Demy 8vo. George Newnes Ltd. 1957. Price 30s.

THIS book covers the principles and practice involved in the design of v.h.f. systems for all the more important communication applications, together with details of the circuits used in v.h.f. transmitters and receivers for domestic entertainment. It will be of interest to those concerned with maintenance and servicing, engineers engaged in planning and development, and also to the advanced amateur experimenters.

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

RELAY TESTERS

(Illustrated below)

Quality Electronics Ltd., 47-49 High Street, Kingston-on-Thames

TIMING TESTER

THIS tester is designed for the following tests:—

(1) Measuring the lag of a relay by means of a meter and timing scale over three ranges and up to a maximum of 0·1sec.

(2) Measuring the operate or release lag of the relay.

(3) Using a make or break action on the relay.

(4) Testing the relay with either battery, earth or loop conditions connected to the coil and contact.

(5) Using an external source of d.c. power of either 24, 40, 50 or 60V.

Accuracy of calibration ± 5 per cent or ± 1msec (which ever is the greater).

The components are tropicalized where possible.

The instrument is fitted into a teak case 15in × 12in × 6¼in and has a formica panel, which is hard wearing and of pleasing appearance.

CURRENT TESTER

This tester is designed for the following measurements and tests:—

(1) Applying a series of current tests, viz. saturate, hold, release, non-operate and operate, by means of appropriate keys in the correct sequence.

(2) Separately controlling or presetting each test supply by a potentiometer.

(3) Indicating the value of test current on a 5-range milliameter (up to 500mA max.), the saturate current automatically being switched to the highest range only.

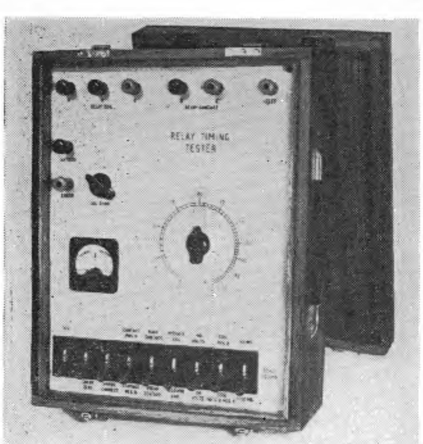
(4) Safeguarding the meter and potentiometer against overload.

(5) Testing relays, having coils connected (a) to battery, (b) to earth, or (c) as a loop.

The tester requires an external supply of d.c. nominally 50V at not more than ¼A.

Accuracy is determined by the meter which is to B.S.89, First Grade. The components are tropicalized where possible.

The instrument is fitted into a teak case, 15in × 12in × 6¼in and has a formica panel, which is hard wearing and of pleasing appearance.

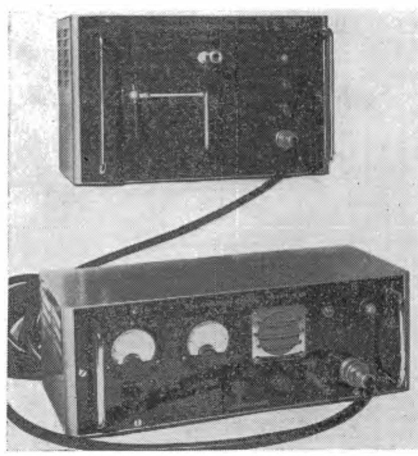


IONIZATION TESTER

(Illustrated below)

Airmec Ltd., High Wycombe, Buckinghamshire

THE use of ionization tests for determining the quality of insulation in components, cables, etc, is now firmly established. Ionization tests, do not damage the components under test and therefore offer many advantages over 'flash' tests which can only indicate poor insulation in a destructive manner. These advantages include (1) the facility for grading components such as capacitors, (2) the possibility of checking for dampness in motor armatures, windings, etc., (3) the ability to determine the safe working voltage of components, (4) the provision of information on the insulating properties of a dielectric.



The 20kV ionization tester type 209, which provides a reliable means of making these tests, indicates the ionization threshold voltage of insulation up to 20kV.

A direct voltage, continuously variable from about 2·5kV to over 20kV, is available from the instrument for applying to the insulation under test. Control of the voltage is effected by adjustment of a spring loaded potentiometer which, for safety purposes, returns to the minimum position directly it is released. In test conditions the voltage is gradually raised and if, at a particular voltage level, ionization currents start to flow within the dielectric, they are amplified by a very high gain amplifier to provide an audible indication by means of a loudspeaker.

The threshold voltage at which ionization commences is thus immediately indicated and the test may be discontinued before rupture of the dielectric occurs. Furthermore, with some components, the type of noise produced by the ionization will enable an operator with experience to interpret faults in the insulation.

If direct leakage currents flow during the test, their value is indicated on a microammeter mounted on the instrument. Since the test voltage is displayed on an adjacent, direct reading voltmeter, an accurate assessment of insulation resistance can be made.

The tester is constructed in two separate units, the e.h.t. unit which provides the test voltage and the control unit which contains the controls, indicators and mains power pack. Connexion between the units is made by 12 yards of flexible cable which contains low voltage lines only.

This arrangement enables the e.h.t. unit to be operated remotely or enclosed in a safety cage if desired, and an electrical interlock circuit is provided to ensure that, in the latter case, no test voltage can be fed to the e.h.t. terminal while the cage doors are open. Additional safety is afforded to both the operator and the component under test by the very high impedance source from which the test voltage is obtained.

To ensure maximum reliability of the e.h.t. unit, and minimum physical size, the rectifiers and smoothing capacitors and resistors are all mounted in an oil-filled container. In view of the high test voltage which can be obtained, a special anti-corona terminal is provided for the e.h.t. connexion and an earthing switch is fitted to enable the tested component to be fully discharged before removal.

Both units are contained in standard steel cases suitable either for forward mounting on a standard 19in rack or for bench use.

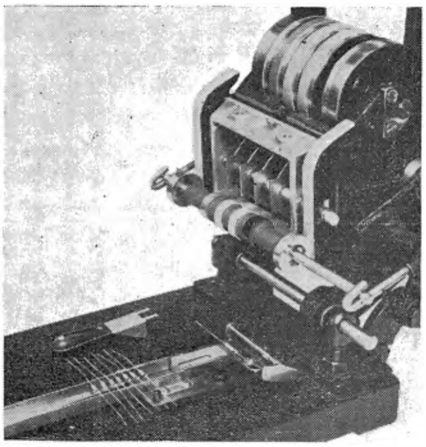
COLOUR CODING MACHINE

(Illustrated above right)

Rejafix Ltd., 81-83 Fulham High Street, London, S.W.6

REJAFIX Ltd announce the introduction of a new machine which will put up to four different colour bands and one line of print in one operation on fuses, resistors (with or without side wires) and such like electrical components. The machine is most simple and efficient in its operation and can be put in the charge of the factory's youngest employee. The four different colour inked bands, etc., are automatically transferred to the surface of the printing pad, over which the article is simply rolled, picking up the colour coding, etc., in the process.

The smallest distance between the bands can be 1/32in and can be ex-



tended to ⅜in or more. The distance between the bands is adjustable and takes one minute only. The article can be up to ⅝in diameter and up to 5in long.

Colour banding parts can be supplied according to customers' requirements, each thickness and combination of thicknesses made up in sets complete as one unit. Thickness of bands can be from hairline up to ⅛in or more.

The machine is offered in two models, hand operated or full automatic. The photograph shows the hand operated model.

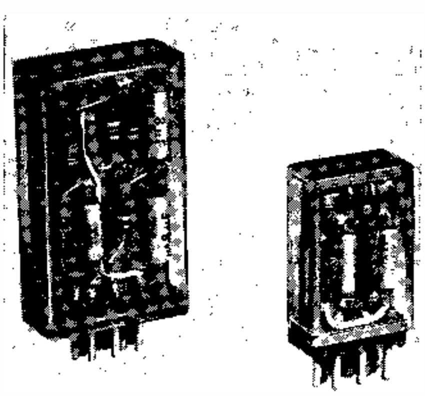
MINIATURE AMPLIFIERS

(Illustrated below)

Venner Electronics Ltd., Kingston By-Pass, New Malden, Surrey

VENNER Electronics Ltd have recently produced two types of transistorized amplifier The first (TS3) is intended for industrial applications and has a response with 3dB from 15c/s to 125kc/s; the second (TS4) is intended for frequencies in the audio range and has a response flat within 3dB from 120c/s to 10kc/s The voltage gains of the amplifiers are similar, being between 900 and 1 000. Each consists of two stages which may be used separately or in cascade, the individual stage gains being of the order of 60 The amplifiers are assembled in plastic cases and are mounted on small 8-pin bases. The pin connexions are identical.

The units are small and lightweight, amplifier TS3 being 1⅛in × 1⅜in × ⅝in and weighing ¾oz; amplifier TS4 being 1⅜in × 2¼in × ¾in and weighing 1½oz.



Both amplifiers are RC coupled and employ temperature compensation circuits to give satisfactory working from − 10°C to + 50°C. The amplifiers are designed to work from a nominal 10V supply but will function with supplies from 1·5V d.c. to 12V d.c. With a 10V supply the maximum undistorted output is 2V r.m.s.

The applications of the amplifiers are numerous and they may be used with other stages in the Venner transistorized plug-in range, which includes oscillators counting, and scaling units, etc.

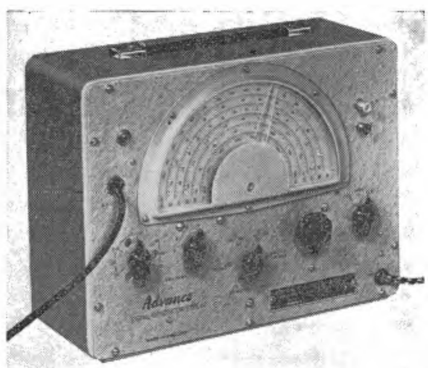
GLASS SEALING BEADS

The English Glass Co. Ltd., Empire Road, Leicester

SEALING beads are now being manufactured in this country by the English Glass Co.

'Englass' sealing beads are made by an accurate moulding process, and can be held to close limits on diameter and bore. Various shapes can be made to suit users' design, and application. Both borosilicate and soda glasses, clear and coloured, are available.

Due to relatively low tool costs small quantities can be produced for experimental purposes.

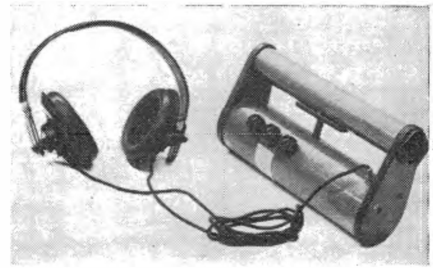


WIDE BAND SIGNAL GENERATOR

(Illustrated above)

Advance Components Ltd., Roebuck Road, Hainault, Ilford, Essex

THE type 62 is an r.f. signal generator with the wide frequency range of 150kc/s to 220Mc/s, thus covering all sound-radio and television frequency bands. The instrument employs a Colpitts oscillator and the frequency calibration is displayed on a scale with a total range of 50in. A band spreading device is incorporated which gives an open scale at the higher frequency ends of the bands. The overall frequency accuracy is − 1 per cent and an adjustable cursor is provided to give closer accuracy when checked against standard frequencies, e.g., broadcast carrier signals or a crystal calibrator. The output signal is available as a continuous wave or may be amplitude modulated 30 per cent at 400 c/s. The output level is controlled by an attenuator system which permits reliable and repeatable measurements down to low signals levels. When the r.f. oscillator is modulated an audio frequency signal of 400c/s is available at separate output terminals.



PORTABLE METAL DETECTOR

(Illustrated above)

Metal Detection Ltd., Moseley Street, Birmingham 12

METAL Detection Ltd. now offer a portable 'lightweight' detector incorporating transistors. The equipment comprises a hand unit and ear-phones, and enables hidden metals to be easily located.

The unit—only 4lb in weight—is passed over the surface or object suspected of metal, the oscillator volume increasing in the head-phones when metal is directly beneath the hand position.

The apparatus is very simple to operate, and can be used for detecting metal-reinforcing in concrete, metal in timber, hidden metal reeo-supports in mines, etc.

IN-LINE INDICATORS

(Illustrated below)

Hilger and Watts Ltd., 98 St. Pancras Way, London, N.W.1

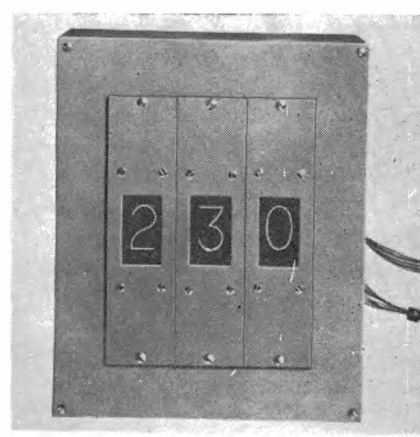
IN-LINE Indicators display a desired number in large clear figures without any mechanical movement in the indicacator. Such a number may be obtained from a 'count' or from a 'measurement'.

They are made for panel mounting and indicators can be assembled in banks so that as many decades as required are displayed in adjacent windows and the line of figures appear as they would normally be written.

Each unit (one decade) comprises a stack of ten perspex plates engaged individually for 0 to 9; each plate being separately edge-illuminated by its own lamp.

Five lamps are held in a common holder (lamp block) at the top of the unit and five at the bottom. Each holder can be readily slipped out to enable rapid replacement of lamps, and is located in the correct position by two pins.

As many indicators as required can be mounted side-by-side, the width of each indicator being 1¼in: in the illustration three units are shown.



Notes from ——— NORTH AMERICA

Artificial Respiration

A new experimental electronic system that enables a polio victim to use respiratory muscles that are still intact to control the breathing of his iron lung was described at the Institute of Radio Engineers' annual convention.

The system was described by Dr. L. H. Montgomery, of the Vanderbilt University School of Medicine, who stated that although the equipment is still in an experimental stage, the National Foundation for Infantile Paralysis plans to make it available to polio centres "as soon as it is practical to do so".

Even in the severest cases of paralysis, Dr. Montgomery reported, the victim is left with the control of a few muscles which contract when he tries to breath. This contraction generates minute voltages that can be detected by sensitive electrodes placed on the skin. The voltages are then fed to electronic equipment which amplifies them and uses them to control the flow of air to and from the iron lung or other type of respirator in accordance with the patient's needs.

The new respirator is expected to be a great boon to patients morale and comfort. In previous types he is compelled to eat, drink and talk in rhythm with a motor driven air pump, whereas, now he will be able to control his own breathing to suit himself.

High Speed Communications

Bell Telephone Laboratories have announced a new experimental device which will enable business machines to 'talk' directly with one another at a speed of about 1 000words/min.

Printed material could thus be sent over lines like those used for telephone purposes, but at a speed sixteen times that of conventional teletypewriter systems.

This new device, tentatively called a data subset, is designed to give accurate transmission among wide varieties of information processing machines, such as electronic computers, electric typewriters, or adding machines. Transmission of the data is automatic, and voice communication can precede or follow the operation.

Information is placed on magnetic tape by an electronic computer, electric typewriter or other machine. In contrast to holes punched in paper tape used in many other automatic machines, characters are placed on the magnetic tape in the form of magnetized spots. A group of seven such spots identifies the particular character. Some spots are magnetized in such a way that they represent 'zero' and others in a way that represent 'one'. All information put on the tape is translated into a code using combinations of these two numbers. The chief advantage of magnetic tape is that its speed is easily adapted to match the speed with which the signals can be sent over telephone wires. It also has the advantage of quiet operation.

After its preparation, the tape is then inserted into the data subset machine. The person on the 'sending' end talks by telephone with the person on the 'receiving' end to notify him that the message is to be transmitted. The recipient turns on a receiver mechanism and the sender then starts transmission. An automatic signal indicates when the taped message is completed. The person on the receiving end removes the tape from the machine and inserts it in the office machine for decoding. On an electric typewriter, for example, the mechanism could be arranged to type out the message automatically from the magnetic tape.

To provide essentially error-free transmission, extra error-checking equipment can be added. In the event of a disturbance in transmission, the receiving equipment will then reject the portion of the transmission affected and automatically cause it to be retransmitted.

The present laboratory models are about the size of office typewriters. If they were engineered for maximum compactness, they could be made considerably smaller. The small and convenient size is made possible by the use of transistors, semiconductor diodes and ferroelectric crystals.

Broadband Disk Bolometers and Thermistors

Two new broadband disk bolometers for coaxial detectors, models N603 and N603-4,5 covering the frequency range of from 500 to 10 000Mc/s and a new broadband disk thermistor, model N335, covering the same frequency range, have been developed by the Narda Corporation.

The new model 603 bolometers, which are suited for attenuation measurements and all relative power measurements, consist of two 100Ω Wollaston wire bolometer elements mounted on a mica disk. The mica disks incorporate the r.f. bypass capacitor for the ungrounded end of the bolometer as well as blocking capacitor at the central junction between the 100Ω elements. This provides r.f. contact to the transmission line centre conductor while blocking the bias current.

The elements are in series to present a 200Ω resistance to the external d.c. or a.f. bias circuits and are arranged to be in parallel in the r.f. transmission line, thus acting as a 50Ω termination.

The new bolometers are manufactured for either of two bias currents, 4·5mA or 8·75mA.

Alignment holes are provided to assure proper assembly in suitable bolometer mounts. When mounted in the Narda model 561 bolometer mount no tuning is required and the v.s.w.r. is less than 1·5 from 500 to 10 000Mc/s. The bolometers may be used with all commercial v.s.w.r. amplifiers and microwave power meters.

Square law response error is less than 1 per cent for power levels of 0·2mW with 8·75mA bolometers or 0·1mW with 4·5mA bolometers. Microwave power levels of 0·01mW to 10mW may be measured with 8·75mA bolometers and power levels of 0·01mW to 4mW may be measured with 4·5mA bolometers

The model N335 thermistor is specifically designed for use with the Narda model 561 coaxial bolometer mount for microwave power measurements. Two thermistor elements are mounted on a mica disk connected in series for d.c. and a.f. bias, and in parallel for r.f. When biased with approximately 13mA the d.c resistance is 200Ω; the r.f. impedance is 50Ω. Construction is similar to the Narda N603 bolometer.

The new broadband disk thermistor is used with microwave power meters capable of supplying the necessary bias (12 to 15mA) to measure power levels from 0·01 to 10mW. Because of the long time-constant (approximately one second) the new thermistor is particularly suited to measurement of pulsed signals. It is not susceptible to burnout, but in any case is readily replaceable and does not require tuning.

V.H.F. Communication over Mountains

Methods for using high mountains between a radio or television transmitter and receiver to improve, rather than hinder, the reception of ultra-high-frequency signals were described at the opening session of the Institute of Radio Engineers' four-day national convention.

R. E. Lacy of the Signal Corps Engineering Laboratories, Fort Monmouth, N.J., described an extensive series of tests made at forty different locations in California which verified that sharp mountain peaks blocking the transmission path will actually strengthen the signal on the other side by as much as 100 million times compared to what it would be with no mountain in the way. The tests were conducted over a wide range of frequencies above 50Mc/s.

This phenomenon, known as 'obstacle gain', was first noticed during the Korean War when it was discovered that radio reception was unaccountably improved in the mountainous terrain of Korea. It was later deduced that ultra-high-frequency radio waves, which act much like light waves, are bent down toward the ground when they pass over sharp mountain ridges, just as light waves are diffracted when passing by the edges of opaque objects.

The information gained from the California tests, Mr. Lacy reported, makes it possible to compute the 'obstacle gain' accurately and to locate transmitting and receiving sites to take advantage of the phenomenon.