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"To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering."
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ENGINEERING DEVELOPMENT

The Institution's contribution to engineering development is expressed in the object of "promoting the advancement of radio, electronics and kindred subjects by the exchange of information." There are various ways in which this object may be achieved, the principal means being by publication of papers in the *Journal*, and the holding of regular meetings in Great Britain and overseas.

New members may be hesitant in contributing to the Institution's proceedings because of uncertainty as to the value of their contribution. It is worth recalling, therefore, that in his 1953 presidential address, Mr. W. E. Miller referred to two important groups of papers—the highly specialised paper dealing with original scientific work, and the survey paper, of more general interest, referring to the latest developments in a particular sphere of radio engineering.

The paper describing original work is in a class on its own, and is often of greatest importance to the advancement of knowledge in a particular field. Such reports are usually the result of long and extensive research, most probably by a group or team in a university, government research establishment, or a large industrial organization. These contributions not only have an immediate value but are an acquisition to the reference library.

The survey paper, however, has an equal value by bringing together information on a much wider aspect of a problem. The extent of the interest will depend largely on whether the survey is intended to summarize in detail present practice in a comparatively limited field of radio or electronics, or whether its approach is more general and intended to help engineers to keep up to date in fields other than their own.

Comparatively few radio engineers are concerned with research into new work which will lead to the publication of really original papers; similarly the requirements of a comprehensive survey may often call for the knowledge and experience of a senior engineer.

There is a third class of paper, however, which is intermediate between the other two groups, namely the paper describing a complete equipment or a project. Such a paper may not be original since it is probably concerned with the application of principles already well known. In its demonstration of the application of engineering principles, however, a high degree of professional ability will be shown. It is of interest to note that papers falling into this category, which could be termed "the engineering development paper," frequently qualify for the Marconi Premium which is awarded for the most outstanding engineering paper.

The writing of an "engineering development" paper should be within the powers of the large body of *development* engineers who are engaged in shaping the equipment of the future. In most instances the engineer's work involves the preparation of internal reports and memoranda which, by suitable change of emphasis, and by stating the background requirements of the project, could be adapted to a form suitable for publication as a paper. The latter point is of considerable importance since the engineering principles involved, and their application, are intrinsically of interest to other engineers and the provision of information on the basic problem will enable them to gain the maximum value from the paper.

As a recent survey of corporate membership has shown, the Institution has within its ranks very many members who are able to prepare original contributions and an even larger number of members capable of adding their quota to Institution proceedings by preparing a survey or papers describing engineering development. It is hoped, therefore, that more members will be encouraged to assist in fulfilling one of the most important objects of the Institution.

NOTICES

Obituary

The Council has learned with regret of the deaths of the following members.

Squadron Leader **Raymond Harrison Stephenson,** O.B.E., Royal Air Force (Associate Member), was posted missing, presumed killed, on January 18th. At the time of his death he was on the Headquarters staff of Middle East Air Force and was engaged on a tour of R.A.F. stations in Arabia. The Anson aircraft in which he was a passenger crashed in the Persian Gulf.

Sqdn. Ldr. Stephenson entered the R.A.F. at the age of 16 years as an apprentice; from 1939-45 he served as a pilot and was commissioned in 1942. In 1946 he was appointed an instructor at the Empire Radio School, where he remained until 1949. He subsequently held various appointments as Signals Officer in the United Kingdom before being posted to the Middle East in August of last year. He was appointed an O.B.E. in Her Majesty's Birthday Honours List in 1954.

Elected first as an Associate in 1947, Sqdn. Ldr. Stephenson was transferred to Associate Member in 1949. He was 39 years of age.

* * *

Cheslyn Frank Lines (Associate Member) died on March 8th in a Glasgow hospital following a brief illness. He was 51 years of age.

Mr. Lines had been in the radio industry almost since its beginning and he first joined the Institution as an Associate in 1935, being transferred to corporate membership in 1943. He was one of the founder members of the Scottish Section and had served on the local Committee. From 1935 to 1944 Mr. Lines lectured in radio subjects and, during the period from 1941 to 1942, he was full-time lecturer to Army classes at Allan Glen's School and the Royal Technical College.

Mr. Lines was appointed an examiner for the Radio Servicing Certificate of the Radio Trades Examination Board in 1943 and he served as a representative on the Board from 1947-49.

* * *

Advice has just been received of the death on September 8th last of Second Lieutenant **Ramdas Yeshwant Oke,** Corps of Signals, Indian Army.

Second Lt. Oke, who was 27 years of age, received the B.Sc. degree of the University of Bombay in 1949 and subsequently took an advanced course at the School of Radio Physics and Electronics, Poona. He was commissioned in 1953 and at the time of his death was serving at the Signals Training Centre, Jubbulpore. He registered as a Student of the Institution in 1951.

Institution Dinner

The forthcoming General Election necessitates postponing the Dinner arranged for May 26th. Further details are being sent to members who have requested tickets.

Scottish Section Dinner

The annual dinner of the Scottish Section took place on March 25th at the St. Enoch's Hotel, Glasgow, and was attended by 70 members and guests. The toast of "The Institution" was proposed by Professor G. W. O. Howe, D.Sc., Ll.D., and the response was made by Mr. R. H. Garner, Chairman of the Scottish Section. Mr. A. A. M. Turnbull, a former honorary secretary of the Section proposed the toast of "The Guests," Mr. R. Newton replying.

A display of radio and electronic equipment was arranged for the entertainment of those attending the dinner and it included a closed-circuit television system, using an industrial television camera linked to a standard domestic receiver.

Subscription Renewal

All grades of members are reminded that subscriptions for 1955-56 were due on April 1st. It is requested that those who have not already remitted will do so as soon as possible in order to avoid the need for postal reminders.

The Annual Report of the Brit.I.R.E. Benevolent Fund for 1953-54 showed that, for the first time, donations from members had not increased compared with the previous year. The Trustees therefore make a special appeal to all members to include a donation to the Fund with their annual subscription.

Teachers' Vacation Course

As in previous years, a Vacation Course for teachers in radio and television is to be held jointly by the Ministry of Education and the Radio Industry Council from July 17th to July 27th at the Northampton Polytechnic.

Although the Course is concerned primarily with servicing, it has been opened this year to teachers who are not engaged on this aspect of the subject.

THE DEVELOPMENT AND DESIGN OF DIRECT-COUPLED CATHODE-RAY OSCILLOSCOPES FOR INDUSTRY AND RESEARCH*

by

M. J. Goddard, A.R.C.S., M.Sc., D.I.C.†

Read before the Institution in London on November 24th, 1954. Chairman: Dr. G. L. Hamburger.

SUMMARY

The wide field of application and range of requirements of the cathode-ray oscilloscope are shown. The problems of design are considered in relation to the limitations of the cathode-ray tube and the requirements of the time base and amplifier. The "Miller-Transitron" time base is considered in detail. The need for direct coupling in both time base and amplifier is discussed. "Straight" and carrier-frequency d.c. amplifiers are described. Drift is a serious problem in d.c. amplifiers and ways of reducing it by negative feedback and stabilization of h.t. supplies are mentioned.

Probes, high-discrimination input units, high-resistance input units, beam-blanking units and delay lines are discussed briefly.

1. Introduction

The cathode-ray oscilloscope is used for investigating repetitive electrical wave-forms by applying a repetitive time base to the X plates of the cathode-ray tube, and synchronizing it with the signal to be investigated. Fig. 1 shows a typical wave-form produced by the time-base generator. It is common practice to make the time duration BC of the fly-back very small compared with the duration AB of the sweep, but from the point of view of obtaining good synchronization there are advantages in avoiding a fly-back of very short duration relative to that of the sweep.[1] For the investigation of signals arising at irregular intervals, or at intervals long compared with the duration of each signal, or once only, the time-base generator must be triggered by the signal instead of producing a "continuous running" time base synchronized by the signal. Examples of such signals are radar pulses, spark discharges and the wave-forms arising at the closing of a circuit.

The cathode-ray oscilloscope now also serves as a measuring instrument for the time intervals and voltage amplitudes of electrical wave-forms. But the field in which it serves is much wider even than this. It is now possible to obtain electrical signals correlated with very many physical phenomena which are not normally directly associated with electrical signals by using the microphone, photo-electric cell, strain gauge, pressure gauge, thermocouple, etc. The signals obtained from such devices can be applied to a cathode-ray oscilloscope, and in this way the range of application of this instrument can be extended to cover almost all branches of science and engineering. It has also been realized that electrical signals are associated with the functioning of various organs of living things, with muscles, nerves, heart, brain. In the biological and medical fields, study of these signals is increasing, and for this study the cathode-ray oscilloscope has a very useful part to play.

2. The Problems of Design

The problems of design fall into two main groups: those associated with the time base, and those associated with the amplifier. It is here assumed that the oscilloscope is to be used for the direct observation of signals as opposed to observation of Lissajou's figures.[2] The design of the cathode-ray tube itself must be accepted as presented by the tube manufacturers, but this is a problem in itself and the limitations imposed by the tube must be considered.

2.1. *Limitations imposed by the Cathode-Ray Tube*

It is desirable, in order to be able to observe accurately the form and amplitude of the observed signals, to present the image at the

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surface of a flat graticule. This is clearly impossible if the screen of the cathode-ray tube is curved. The obvious answer to this problem is to employ a tube with a flat-faced screen.

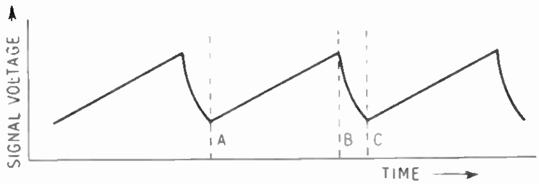


Fig. 1.—Typical time-base wave-form.

This, however, presents its own problems. It can be clearly seen from Fig. 2 that the electrons travelling along OB to the top of the screen have a longer path than those travelling along OA to the centre of the screen. Consequently the deflection in the X-direction produced on the electrons travelling along OB is greater than that produced on the electrons travelling along OA, with the result that the scan at B is larger than the scan at A. Similar reasoning shows that the vertical deflection is greater at the edges of the screen than at the centre. The effect is to produce pincushion distortion on the screen. This effect can be compensated by setting the graphite coating G of the tube to a different potential from the mean potential of the deflection plates P_X, P_Y, and final anode A3. An electron lens is thus produced where the electrons leave the deflection plates. The potential of the graphite layer can be adjusted experimentally to correct the pincushion distortion arising from the geometry of the tube. Compensation is, however, not perfect. This fact sets one limitation to the accuracy of the display on the screen.

This problem is even more acute in a multiple-gun cathode-ray tube. Consider, for example, the double-gun tube shown diagrammatically in Fig. 3. It will be immediately apparent that the path O C′ from the upper gun to the bottom of the screen is longer than the path O′B′ from this gun to the top of the screen. Consequently the inherent curvature of the scan is greater at C′ than at B′ for the upper gun and vice versa for the lower gun. The compensation required for either gun is therefore asymmetrical, and can only be provided in very limited degree by means of the graphite potential. Curvature of the trace is thus much more serious in multiple-gun tubes than in single-gun tubes.

Another distortion arising in almost all cathode-ray tubes is non-orthogonality of the axes, that is to say the deflection produced by the Y-plates is not at right angles to that produced by the X-plates. This distortion arises from inaccurate mechanical alignment of the deflection plates. The effect is most serious when the observed signal contains transients, for then a transient in one direction will appear to have a poor rise-time while a transient in the other direction may actually appear to have a negative rise-time!

A similar distortion which occurs in multiple-gun tubes is non-alignment of the axes of one

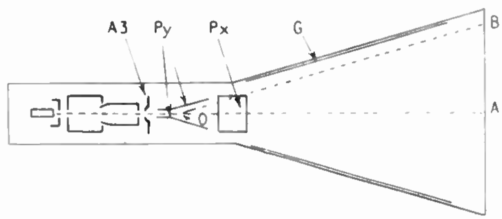


Fig. 2.—Pincushion distortion in flat-faced tubes.

gun relative to another. This means that the scan from one gun is not parallel to that from the other. It does not in itself produce distortion of the observed wave-form, but it makes comparison of the wave-forms from the two guns less convenient and is generally somewhat disturbing to the user.

A fundamental limitation set by the tube is the deflection sensitivity for a given intensity of

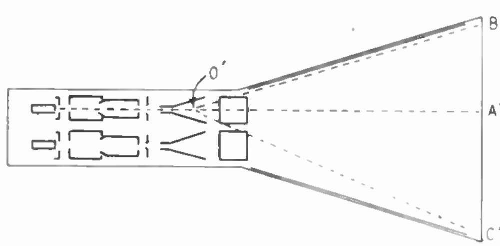


Fig. 3.—Distortion in a double-gun tube.

the trace. Reduction in the final anode potential of the electron gun provides increased deflection sensitivity but at the cost of reduced intensity and resolution of the trace. This is most serious when non-recurrent phenomena of short duration are to be observed. High deflection sensitivity is then desirable so that the writing speed of the time-base trace will be sufficiently high to display the phenomena adequately, but if high deflection sensitivity is obtained by reducing the final anode potential of the gun, then a single trace at high velocity may not be sufficiently bright to be visible.

A useful technique to improve the performance in such cases is to employ post-deflection acceleration of the electron beam. For this purpose a potential more positive than that of the deflection plates is applied to one or more graphite bands near the screen. In this way the final accelerating potential of the electrons can be increased without serious distortion of the trace to about three times the potential of the final anode of the electron gun, the loss of deflection sensitivity thereby produced being only some 30 per cent. The distortions produced by this technique are, however, more serious in double-gun tubes. Another effect which is liable to occur in tubes using post-deflection acceleration is a bright splash on the face of the tube produced by the electron beam striking the deflection plates when it is deflected beyond the edge of the tube-face. This effect can be avoided by suppressing the electron beam when it is deflected sufficiently to strike the deflection plates, but this involves increased complication in the time-base generator.

Another limitation is set by the fact that the electron image is defocused if the mean potential of the deflection plates changes. Consequently, to avoid defocusing, it is desirable in direct-coupled oscilloscopes that both time base and amplifier should have a push-pull output.

2.2. *The Range of Requirements*

The range of requirements for oscilloscopes at the present time is very great, and is continually extending. At one extreme, the techniques of radio-location and radio-control employ pulses which may be as short as 0·1 microseconds or less, and may repeat only at the order of 50 c/s or may even be non-recurrent. For such work the final accelerating potential of the cathode-ray tube must be at least of the order of 10 kV, and the time-base generator must be capable of sweeping the beam across the face of the tube at a velocity of the order of 10 in or more per microsecond. For this, considerable power output from the time base is required in order to charge or discharge the inherent capacitance of the deflection plates of the cathode-ray tube and the associated circuitry in the very small time of the sweep. A brightening pulse must also be applied to the grid of the cathode-ray tube of sufficient power to charge the inherent capacitance of the grid and its associated circuitry in a fraction of the sweep time. The amplifier usually does not need to have a very high gain, but it requires a rise-time of the order of 0·01 microseconds. Furthermore, as the time base does not start instantaneously at the application of a triggering signal, it is desirable that a delay network should be incorporated in association with the amplifier, thus delaying the signal passing through the amplifier sufficiently long to enable the time base to start before the signal from the amplifier is applied to the cathode-ray tube. This allows the start of the pulse to be seen.

At the other end of the range of requirements workers in the biological and chemical fields sometimes require to observe phenomena occurring in times exceeding a second, and for this work the time base requires to run at a fraction of a cycle per second. For such work it is sometimes most convenient to generate the time base by mechanical means.

Biological workers frequently require to observe signals of small amplitude, of the order of microvolts. For such purposes the amplifier requires to have a gain of the order of 100,000 to 1,000,000, while having at the same time adequate band-width to cover the frequency range of the signals to be observed, with a noise level lower than that of the signal to be observed and sufficient freedom from drift to enable the maximum gain of the amplifier to be employed without excessive drift of the trace on the screen. These requirements are very exacting indeed.

In practice it is necessary to employ a variety of models of oscilloscope to cover the very wide range of requirements, and, in order to put these to the most effective use, various pieces of auxiliary equipment are needed.

3. Time Bases

For some purposes it is convenient to use circular or spiral time bases.[4] The spiral form is particularly useful where a very long time base is required. The simplest and most convenient for most purposes, however, is the linear time base.[3] It is common practice to produce the time base by charging a capacitor by a constant current. The difficulty, however, is to produce the constant current to charge the capacitor.

3.1. *The "Miller" Time Base*

One very useful time-base circuit is that known as the "Miller run-down circuit," or "Miller time-base generator," the basic circuit of which is shown in Fig. 4. It consists essentially of a

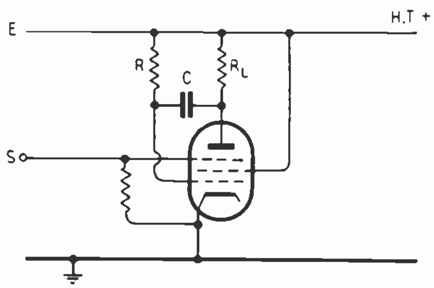


Fig. 4.—The "Miller" time base.

valve having a capacitor C connected between anode and control grid and the control grid connected through a resistor R to a positive potential E. In this circuit the capacitor C is charged through the valve in the same way as a capacitance AC would be charged through the resistor R from a potential AE if the valve were not present, where A represents the amplification of the valve. Thus we have the *effect* of a capacitor charged through a resistance to a potential E from a source of much higher potential AE, without the inconvenience of actually providing such a source of high potential.

In order to operate this circuit as a time-base generator, it is necessary to provide a switching mechanism. This may be provided in various ways. In Fig. 4 it is shown as applied to a terminal S connected to the suppressor grid of the valve. Before the instant at which the sweep of the time base is required to commence, the suppressor grid is held to a potential sufficiently negative to cut off the current to the anode, which is therefore charged to the supply potential (H.T.+). During this time the control grid is pulled positive by the connection to H.T.+ through the resistor R, but grid current holds the grid near cathode potential. The cathode thus passes the full "zero bias" current of the valve, and the bulk of this current is collected by the screen grid. At the instant at which the sweep of the time base is required to commence, the suppressor grid is suddenly brought to earth potential by the switching mechanism. Immediately the whole current goes through to the anode, but in so doing it forces the anode negative. This negative transient on the anode is communicated by the coupling capacitor C to the control grid, driving the control grid negative, thus sharply reducing the current. The result is that, provided that the effects of stray capacitances are negligible, the anode and control grid each receive equal negative transients, the amplitude of these transients being such that the negative bias on the grid is sufficient to reduce the anode current by that amount which, when multiplied by the load resistance R_L, will produce that amplitude of transient. The "Miller run-down" then commences at the anode. The action here is that the capacitor C discharges through the resistor R, thus allowing the grid to run positive, but in running positive the grid allows a larger current to pass to the anode, driving the anode negative and hence, through the capacitor C,

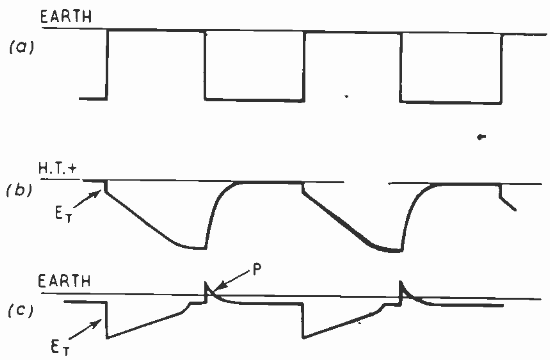


Fig. 5.—Wave-forms of the potentials at the suppressor grid (a), anode (b), and control grid (c), of the "Miller" valve in the circuit of Fig. 4.

feeding a negative-going signal back on to the grid tending to neutralize the effect of the discharge of the capacitor C through the resistor R. It is this negative feedback which produces the linearizing action similar to the charging of a capacitance AC from a potential AE. The run-down proceeds, provided that it is not interrupted by the action of the switching mechanism at S, until the anode reaches such a potential that any increase in valve current goes to the screen instead of to the anode. The anode then remains at a constant potential until the switching mechanism again switches the suppressor grid to a negative potential, when the current is cut off from the anode, the potential of which returns exponentially to HT+ by discharge of the capacitor C through the load resistor R_L, and the cycle of operations is complete.

The wave-forms of the potentials at the anode and control grid in this circuit are shown in Fig. 5b and c, assuming a square-wave type of signal applied to the suppressor grid as shown in Fig. 5a. The vertical scale for the control grid is larger than that for the anode, the transient E_T being actually of equal magnitude in each. The positive peak P in the control-grid wave-form is produced by the positive-going fly-back potential of the anode, which is communicated through the capacitor C to the control grid, pulling the control grid further into grid current than its normal equilibrium produced by the connection to H.T.+ through the resistor R alone.

3.2. *The Time Base in "continuous running," triggered or single-stroke operation.*

The cycle of operations described with reference to Figs. 4 and 5 is suitable for triggered operation of the time base provided that the switching mechanism is set in operation by the triggering signal. A mono-stable flip-flop, such as that shown in Fig. 6, forms a convenient switching mechanism for this.

The "Miller" valve V_M is connected in the same way as in Fig. 4, except that the suppressor grid is returned through a resistor R1 to a negative potential E_{n1} such that, in the quiescent state, current to the anode is cut off. The flip-flop is constituted by the valves V2 and V3. The valve V1 serves as a triggering valve. The grid of this valve is returned through a resistor R2 to a suitable negative bias potential E_{n2}. A positive-going triggering signal applied at S initiates the sweep of the time base. The fly-back is initiated by the leaking away of the negative charge on the grid of the valve V2 through the resistor R4. As soon as the valve V2 begins to conduct again, the anode is driven negative, thus switching the suppressor grid of the "Miller" valve and the control grid of the valve V3 back to a negative potential. This cuts off the anode current of the "Miller" valve, producing the fly-back. Conditions have now reverted to the quiescent state which prevailed before the arrival of the triggering pulse, and the circuit is ready to receive the next triggering pulse. The duration of the switching signal produced by the flip-flop is determined by the time constant C_2R_4. This must be so chosen as to be of the same length as or just a

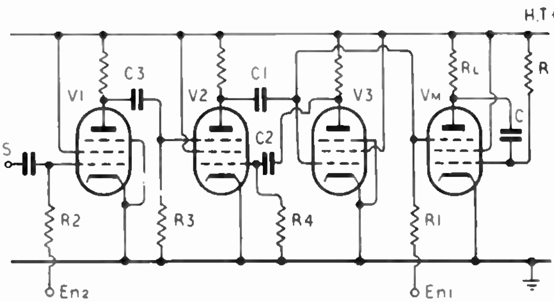


Fig. 6.—"Miller" time base operated by monostable flip-flop.

little longer than the sweep time required from the time base.

The requirements for a "continuous running" time base are different. Here the switching mechanism must oscillate continuously, although being capable of synchronization by an external signal. This condition may be obtained by converting the circuit of Fig. 6 into a multivibrator. There is available, however, a simpler method which obviates the need of the two valves constituting the multivibrator. This is by using the circuit known as the "Miller-Transitron," which is shown in Fig. 7. It will be seen that the suppressor grid of the "Miller" valve is returned to earth through a resistor R1 and the screen grid is connected to H.T.+ through a resistor R_S, these electrodes being coupled together by a capacitor C_S.

The operation of this circuit can most easily be considered with reference to the curves shown in Fig. 8. Curves a and b, referring respectively to the anode and control grid, are essentially of the same form as the corresponding curves b and c of Fig. 5; the same symbols have been used as in Fig. 5, and the explanation of the details of these wave-forms is the same as that for the corresponding wave-forms in Fig. 5.

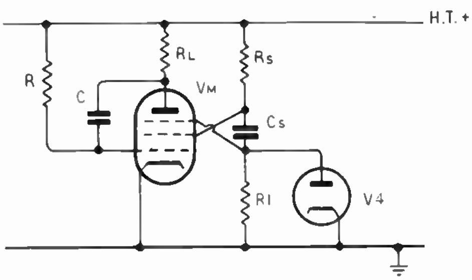


Fig. 7.—The "Miller-Transitron" circuit.

The switching action is performed by the "transitron" circuit constituted by the screen grid and the suppressor grid and the associated components. At the start of the run-down, the suppressor grid is near earth potential, and the bulk of the valve current goes to the anode. The screen grid is therefore at a potential near H.T.+. As the run-down proceeds, the control grid runs positive, so that the screen-grid current increases. The screen grid therefore runs negative as shown at AB in Fig. 8c. The coupling capacitor C_s (Fig. 7) carries the suppressor grid negative at the same rate as the screen grid, as shown at CD in Fig. 8d. At the end of the run-down the anode "bottoms," and the screen-grid current suddenly increases. This drives the screen grid negative, and with it the suppressor grid. As the suppressor grid goes negative it increases the tendency of the current to go to the screen grid rather than to the anode. This cumulative action produces a negative transient BE on the screen grid, resulting in an equal negative transient DG on the suppressor grid. The constants of the circuit are so chosen that the suppressor grid is driven more negative than the critical potential E_c (Fig. 8d) at which most of the current is cut off from the anode. The valve is thus switched to the condition to produce the fly-back. The peak P on the control-grid curve produces a negative over-shoot E_r on the screen grid, but once the fly-back is complete the screen grid remains at a constant potential shown at FJ. The return EF of the screen-grid over-shoot is reflected at GH on the suppressor-grid curve. Thereafter the suppressor grid returns, exponentially, towards earth potential due to discharge of the capacitor C_s through the resistance Rl, this exponential curve being shown at HQ in Fig. 8d. At the point M, however, the suppressor grid reaches the critical potential at which an appreciable proportion of the current begins to go to the anode. When this occurs, the screen-grid current is correspondingly reduced. This allows the screen grid to run positive, carrying the suppressor grid positive through the coupling capacitor C_s. This in turn allows more current to pass to the anode, producing another cumulative action resulting in positive transients

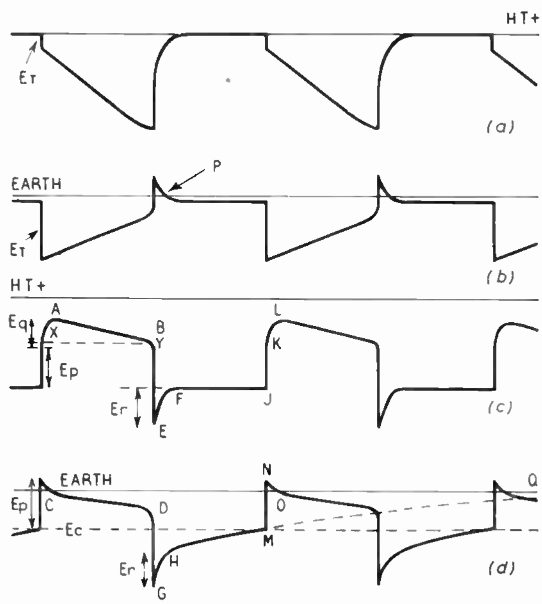


Fig. 8.—Wave-forms of the potentials at the anode (a), control grid (b), screen grid (c), and suppressor grid (d), of the "Miller" valve in the circuit of Fig. 7.

on the screen grid and the suppressor grid. The suppressor grid is held at earth potential by the diode V4 (Fig. 7); the transient MN at the suppressor grid ($= E_p$, say) is therefore less than that at the screen grid, which can be divided into two portions, JK = E_p and KL = E_q. The portion JK is produced by the cumulative "transitron" action of the screen grid and suppressor grid. The portion KL is produced by discharge of the capacitor C_S through the resistor R_S; the rate of change of

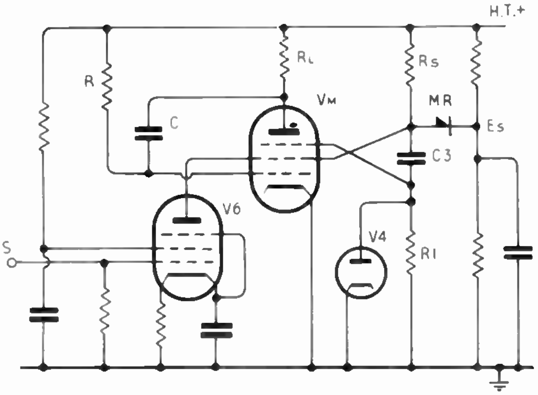


Fig. 9.—Improved "Miller-Transitron" circuit.

potential in the portion KL is therefore less than that in the portion JK. The positive run KL of the screen grid exerts a positive drive through the capacitor on to the suppressor grid, drawing the diode V4 into conductivity, and producing the small overshoot NO on the suppressor-grid curve. At the points L and O on the screen-grid and suppressor-grid curves, the potentials are the same as at the points A and C, and the cycle of operations then repeats itself, thus producing the required switching action on the suppressor grid of the valve.

The operation of the circuit of Fig. 7 is not perfect. The negative-running portions AB and CD of the screen-grid and suppressor-grid curves produce a corresponding decrease in the mutual conductance of the valve. The linearizing action of the "Miller" circuit depends on the mutual conductance of the valve remaining substantially constant. Consequently the linearity obtained from the circuit of Fig. 7 is not particularly good.

The circuit shown in Fig. 9 is an improved form* of that shown in Fig. 7. A rectifier MR is connected between the screen grid and a constant supply potential E_s, the latter being so chosen as to hold the screen grid always more negative than the potential at B in the curve of Fig. 8c. The effect of this is to cut off the sloping top of the screen-grid curve along the dotted line XY. As the screen grid does not now run negative, neither does the suppressor grid, so that the linearizing action of the "Miller" circuit is restored.

Synchronization of the time-base wave-forms generated by the circuits of Figs. 7 and 9 to the signal under investigation can be effected by applying a portion of the signal to the screen grid or to the suppressor grid of the "Miller" valve. This superposes this signal upon the suppressor-grid wave-form during the fly-back, as shown in Fig. 10, causing the suppressor-grid wave-form to cross the critical potential line E_c in synchronism with a positive-going portion of the signal.

It is necessary to exclude the signal from the screen grid and suppressor grid of the "Miller" valve during the run-down, otherwise the run-down is modulated by the signal. This is automatically achieved by the circuit of Fig. 9, in which the signal is applied at S to the control grid of a valve V6, the anode of which is connected directly to the screen grid of the "Miller" valve V_M. As the screen grid is held by the rectifier MR to the constant potential E_s during the run-down, the signal does not appear at the screen grid during this period. During the flyback, however, the rectifier MR does not

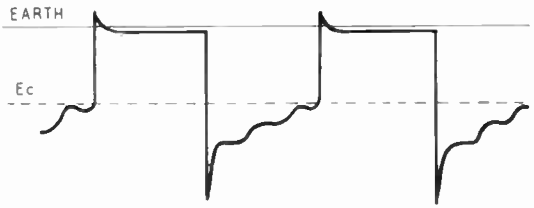


Fig. 10.—Suppressor-grid wave-form of the circuit of Fig. 9.

conduct. The signal therefore appears at the screen grid and is fed by the capacitor C3 on to the suppressor grid.

While the circuit of Fig. 9 is especially well suited to the generation of a "continuous-running" time base, it can be adapted to the generation of a triggered time base. For this purpose the resistor R1 is returned not to earth but to a negative supply line, the potential of

* British Patent No. 679,294.

which is more negative than the critical potential E_c in Fig. 10. Then, in the absence of a triggering signal, the circuit remains quiescent with the suppressor grid at the potential of the negative supply line, thus cutting off the current to the anode, which therefore assumes the potential of the positive supply line H.T.+. The circuit is triggered by a negative-going signal at S, producing a positive-going signal on the screen grid of the "Miller" valve, which is communicated via the capacitor C3 to the suppressor grid, thus initiating the run-down. The fly-back follows automatically, but the circuit then again becomes quiescent until another triggering signal appears at S.

3.3. *Particular Problems arising in Time-Base Design*

In applying the circuit of Fig. 9, the first problem is to select a suitable valve for the "Miller" valve. In the "Miller-Transitron" circuit, the driving electrode is the screen grid of the valve, and the switching electrode is the suppressor grid. These two facts largely determine the choice of a suitable valve. Clearly the valve must be such that the current to the anode can readily be cut off by biasing the suppressor grid to a negative potential. This excludes virtually all valves designated power pentodes or output pentodes, for in these the suppressor grid consists only of a few turns of wire, with the result that negative bias on the suppressor grid is ineffective in cutting off the current to the anode. Clearly, also, the screen grid must be capable of developing the necessary switching potential. This is not difficult in low-speed work, for then the screen resistance R_s can be made large; but for high-speed work the screen impedance is limited by stray capacitance, and here one of the major limitations on the maximum speed obtainable from the time base is set by the current available from the screen grid, which in turn is determined by the maximum rated power dissipation of the screen grid. One of the most satisfactory valves for the purpose, owing to its relatively high permissible power dissipation at the screen grid, is the Mullard EF50, but this valve has now, unfortunately, been transferred to the maintenance list. There is, however, a very useful special-purpose valve in the Mazda 6F33. This valve has a very close-mesh suppressor grid, resulting in an exceptionally sharp cut-off when the suppressor grid is biased negatively. At first sight this valve would appear to be at a disadvantage with respect to the EF50, since the maximum screen dissipation is little more than one-third that of the EF50, but this is more than counterbalanced by the very sharp suppressor-grid cut-off, which reduces the switching current required from the screen grid. The working conditions of the 6F33 require to be chosen with more exactitude than those of the EF50 on account of its great sensitivity to changes in suppressor-grid potential, but, properly handled, it is a very useful valve.

The circuit of Fig. 9, as shown, provides a time base of only one writing speed. For a general-purpose oscilloscope a wide range of writing speeds must be provided, preferably continuously variable. Steps in writing speed at convenient intervals may be provided by switching capacitors of different values into circuit at C, the writing speed being inversely proportional to the capacitance C. A continuously variable fine control is conveniently provided by returning the resistor R not to H.T. + but to a potentiometer connected between H.T. + and earth. The writing speed is directly proportional to the potential to which the resistor R is returned. The capacitor C3 must also be switched so that the capacitance C_3 changes in proportion to the capacitance C.

The lower limit to the speed of run-down at which the time base can be conveniently operated is set by the size of the capacitors which it is convenient to employ. They begin to become inconveniently large when the time of the run-down exceeds about a second. The upper limit to the speed of run-down is set by stray capacitances in the "Miller" circuit. In the conditions normally prevailing in an oscilloscope, these stray capacitances become serious when the time of run-down is of the order of a microsecond. The effect of stray capacitance at the screen grid is usually the most serious, as it limits the switching power of the transitron circuit. Stray capacitance at the grid of the "Miller" valve is also troublesome if it is comparable with the capacitance C. The effect is to produce a capacitance potential divider in the anode–grid feed-back circuit, so that, in order to produce a transient E_T at the grid (see Fig. 8b), a transient larger than E_T must appear at the anode. Furthermore, due to stray capacitance at the anode, this "transient" is not a true transient but a change taking place in a time which is a significant part of the time of run-down. The effect is to produce a non-linear run-down having a fast start. To reduce this effect it is helpful to make R as small as practicable, thus increasing the value of C required for a particular speed of run-down, so that the stray capacitance present is relatively smaller compared with C.

The circuits described above produce only a potential "run-down," synchronized to or triggered by the external signal under investigation. For most cathode-ray tubes, if defocusing of the trace is not to be produced at the edges of the scan, it is necessary to apply equal and opposite deflecting potentials to each pair of deflection plates. This means that a potential "run-up" must be produced for the second X-deflection plate equal and opposite to the run-down applied to the first plate. For this purpose a paraphase amplifier having a gain of −1 is normally employed.

A very convenient form of direct-coupled paraphase amplifier is the so-called "anode-follower" circuit, the basic form of which is shown in Fig. 11. The gain of the anode-follower stage is substantially equal to −1 if $R_2 = R_1$. If the time base is required to operate at frequencies where stray capacitances are comparable in impedance to the resistances R_1 and R_2, capacitors must be connected in parallel with R1 and R2 such that the impedance ratio of the total effective capacitances present is equal to the ratio of the resistances.

It is necessary also to provide an X-shift for the oscilloscope. This is very conveniently done by applying d.c. bias to the grid of the anode follower, for example by taking the resistor R3 to a variable potential instead of to the negative line. It is, however, desirable that the X-shift should also be applied as an equal and opposite potential to the X-deflection plates of the cathode-ray tube, otherwise some defocusing of the trace will result together with a change of deflection sensitivity with X-shift. To this end a second paraphase amplifier can be driven from the anode of the first paraphase amplifier V, the X-deflection plates being directly connected to the anodes of the two paraphase amplifiers. The X-shift is then produced by applying a d.c. bias at the grid of the first paraphase amplifier.

The anode-follower circuit is very easy to operate at low frequencies. The limitations at high frequencies are set principally by the stray capacitances at the anodes of the valves. The problem presents a somewhat different aspect according to whether the output of the valve is positive-going or negative-going. For the negative-going valve, the problem is for the valve to produce sufficient current to charge the stray capacitance. Any current required to produce the necessary potential drop across the anode load is wasted as far as the problem of charging the stray capacitance is concerned.

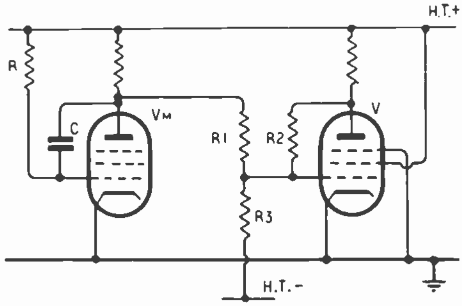


Fig. 11. The basic "anode-follower" circuit.

Consequently the anode-load resistance should be relatively high so that the current required is small, although it cannot be made very high, for that would produce an unduly long fly-back time since, during the fly-back, the stray capacitance has to be discharged through the anode-load resistance. For the positive-going valve, the problem is to get the stray capacitance discharged through the anode load. Consequently it is advantageous to keep the anode-load resistance as low as possible, the lower limit being set by the maximum current which it is permissible to draw from the valve, which must develop the required output potential across the anode-load resistance. Inductance in series with the anode load, which is largely immaterial in association with the negative-going valve, is detrimental with the positive-going valve, since it increases the impedance through which the stray capacitance must discharge.

3.4. *The Need for Direct Coupling in the Time Base*

For a "continuous-running" time base of relatively high sweep-speed it is not important whether or not direct coupling is employed. At lower speeds the time-base sweep may be distorted if capacitive coupling is employed due to partial discharge of the coupling capacitors during the sweep. This distortion does not arise if direct coupling is employed. In triggered operation, however, if the time base is not direct coupled, the starting potential at the plates of the cathode-ray tube will depend on the recurrence frequency of the triggering signal, due to charging up of the coupling capacitors during each sweep. Consequently, with a triggering signal occurring at irregular intervals, the starting point of the time base will change with the recurrence time of the triggering signal, the reproduction of the signal on the screen of the cathode-ray tube thus being undefined in position. This difficulty does not arise when the time base is direct coupled.

4. Amplifiers and the significance of Direct Coupling

The term "Direct-Coupled Amplifier" suggests that a direct connection exists between the output of each stage and the input of the following stage. This is true of many d.c. amplifiers, which may be called "straight" d.c. amplifiers; but "Carrier Frequency D.C. Amplifiers" also exist, where the coupling between certain stages may be through capacitors or h.f. transformers. All amplifiers in which the output signal produced is proportional to the input signal applied, the proportionality factor being independent of the frequency of the input signal for all frequencies below a given limit, may be considered "Direct-Coupled Amplifiers," by whatever means the result is achieved.

The advantages of employing a direct-coupled amplifier are largely self-evident, such, that it enables d.c. as well as a.c. potentials to be amplified, that it allows any d.c. component in an a.c. signal to be investigated, and that low frequency cut-off does not occur. There are, however, certain other advantages which follow from these. For example, it is possible to provide a calibrated d.c. potential which may be applied to the input of the amplifier, and so produce a shift on the screen of the cathode-ray tube which may be compared with the amplitude of a signal under investigation, thus providing direct measurement of the amplitude of the signal. An important aspect of the presence of the d.c. component of an a.c. signal appears in the investigation of a signal, such as a television picture signal, which contains a fixed reference d.c. level, but has a varying amplitude of signal, or contains pulses of varying mark-space ratio. When such a signal is passed through a reactance-coupled amplifier the reference d.c. level is lost, but this is retained when the signal is amplified by a direct-coupled amplifier. But further, a direct-coupled amplifier can handle a short duration pulse of nearly double the amplitude of the largest pulse which can be handled by an equivalent reactance-coupled amplifier. This is because in a reactance-coupled oscilloscope amplifier, each stage, when in its quiescent state, must be biased to the centre of its useful working range so that it may be able to handle a received pulse of either polarity without cut-off. However, when a direct-coupled amplifier is employed to investigate a pulse of particular polarity, the amplifier may be biased by a d.c. potential applied to the first stage so that each stage is biased towards the end of its working range opposite to that towards which it will be driven by the pulse. Thus, whereas in a reactance-coupled amplifier no stage can be driven by the pulse through more than half its working range, in a direct-coupled amplifier the full working range is available in the stage which receives the greatest drive.

4.1. *"Straight" D.C. Amplifiers, their advantages and limitations*

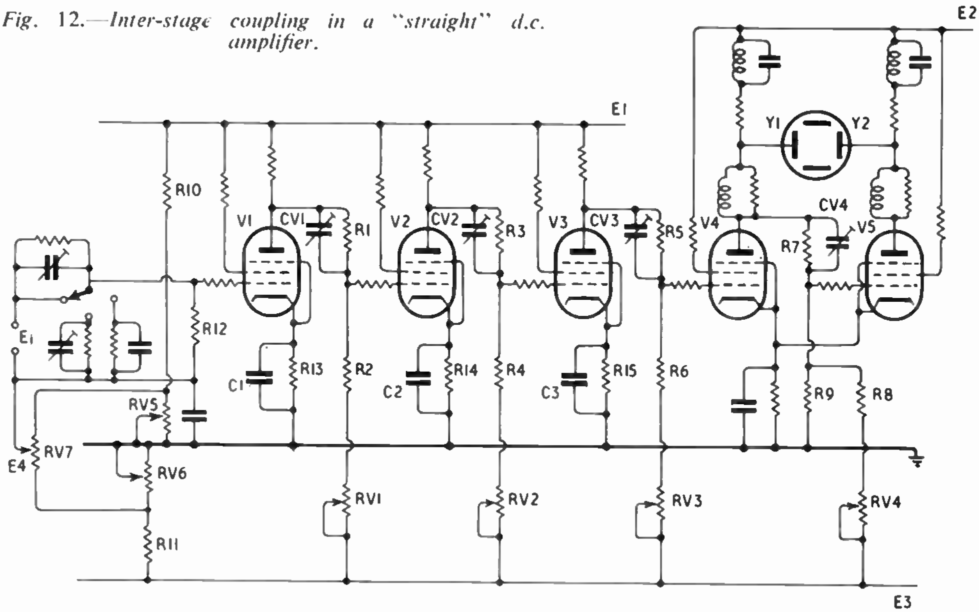
"Straight" d.c. amplifiers have the fundamental advantage of simplicity. Provided that the bandwidth required from the amplifier is small, a single pentode valve with a high anode-load resistance is capable of giving an amplification of thousands. When a wide-band amplifier is required, however, it is necessary to reduce the anode-load resistance on account of stray capacitance present, and then a number of stages is required if high gain is needed. This involves various complications.

If such a multi-stage amplifier is mains driven, then any mains voltage change at the anode of the first stage will be fed directly to the input of the second stage, and so will appear at the output, amplified by all stages except the first. For this reason it is necessary that the earlier stages should be fed from a stabilized supply.

Furthermore, if the anode of the valve of one stage is connected directly to the grid of the valve of the next stage, then the grid of the valve of each stage is at the potential of the anode of the valve of the preceding stage. This means that the potential supply for an *n*-stage amplifier requires to be *n* times that needed for each stage, which can be very inconvenient if more than two or three stages are required. One way of obviating this is employed in the "straight" d.c. amplifier shown in Fig. 12. The anode of the valve V1 is coupled to the grid of the following valve V2 through the resistor R1, with a second resistor R2 taken to a negative supply line E_3. The variable resistor RV1 provides an adjustment of the d.c. level applied to the control grid of the valve V2. It is necessary that both the positive line E_1 and the negative line E_3 supplying the earlier stages should be stabilized, and it is also necessary to connect a capacitor CV1 in parallel with the resistor R1 to compensate at high frequencies for the stray capacitance to earth of the grid of the valve V2 and its associated connections. This stray capacitance is largely constituted by the capacitance to earth of the grid of the valve, which may vary with change of heater current or bias conditions in the valve. Consequently the adjustment of the capacitor CV1 is not correct for all working conditions of the valve, so that the circuit is not very satisfactory. This is less serious if the negative supply line E_3 has a negative potential large compared with the positive potential E_1, but that is inconvenient. Alternatively the resistor R1 may be replaced by a voltage-stabilizing device, such as a neon stabilizer, which has a low effective d.c. resistance. Stabilization is not essential for positive line E_2 supplying the valves V4 and V5 which drive the plates Y_1 and Y_2 of the cathode-ray tube.

The amplifier shown in Fig. 12 incorporates a calibrated d.c. potential E_4. This is derived from the potentiometer RV7 connected to the resistor chains R10, RV5 and R11, RV6, which are supplied from earth and the stabilized potentials E_1 and E_3. The potentiometers RV5 and RV6 provide pre set adjustments for calibrating RV7. The potential E_4 is connected in series with the input signal E_i to the control grid of the valve V1 via a capacitance-corrected resistance attenuator, the grid resistor R12 of the valve V1 being returned to the calibrated potential E_4. This calibrated potential may thus be compared with the input signal, and a scale

Fig. 12.—Inter-stage coupling in a "straight" d.c. amplifier.



can be attached to the potentiometer RV7 to enable the amplitude of the input signal to be measured directly in volts.

This amplifier is suitable for providing a band-width of the order of 10 Mc/s and a gain of the order of 250. Compensating capacitors C1, C2 and C3 are connected in parallel with the cathode resistors R13, R14 and R15 of the valves for these stages. Valves V4 and V5 are driven in push-pull with a common cathode load. Cathode correction cannot, therefore, be applied to these stages, and so inductive correction is employed in the anode loads. The variable resistors RV1, RV2, RV3 and RV4 are adjusted so that the mean working potential of the control grid of each of the valves V2, V3, V4 and V5 is earth potential. This facilitates the lining up of the amplifier, particularly the application of a signal to the control grid of any stage to check the frequency response.

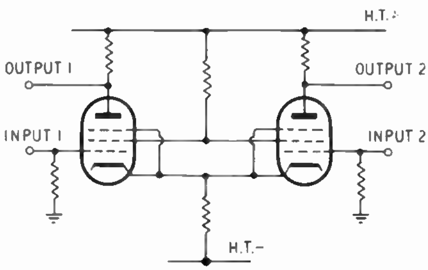


Fig. 13.—The "long-tailed pair."

The process of reducing the anode loads of each stage of the amplifier in order to increase the band-width cannot be continued indefinitely, otherwise the stage ceases to amplify. Connecting two or more valves in parallel allows the anode loads to be reduced somewhat further. The useful limit to this process is, however, quickly reached, because the addition of each extra valve also adds the anode capacitance of that valve to the output capacitance of the stage. Further increase in band-width can, however, be realized by using the technique of the distributed amplifier.[5,6]

The circuit shown in Fig. 13, known as the "long-tailed pair," is very useful in "straight" d.c. amplifiers. It has the advantage that any supply potential change produces nearly equal effects on the two valves of the pair. Consequently the effects of supply potential changes are largely cancelled, so that the need for employing stabilized supply potentials is much reduced. The two valves are operated with the input signal to the stage from the output of the preceding stage applied in push-pull between the control grids of the valves and the output signal taken in push-pull from the two anodes. This circuit is also very useful as the input stage where a push-pull input signal is to be used. When a single-sided input is employed, the grid of one valve of the pair can be earthed.

4.2. *Carrier Frequency D.C. Amplifiers and their applications*

The basic principle of the carrier frequency d.c. amplifier is that the input signal is caused to modulate a carrier signal produced by an oscillator, the modulated carrier then being amplified by a carrier amplifier. After amplification the signal is demodulated, when it may be fed direct to the output of the amplifier, or may be further amplified by a "straight" amplifier of relatively low gain. The modulator is so arranged that, when the input signal is zero, the carrier fed to the carrier amplifier is about half the peak available maximum amplitude.

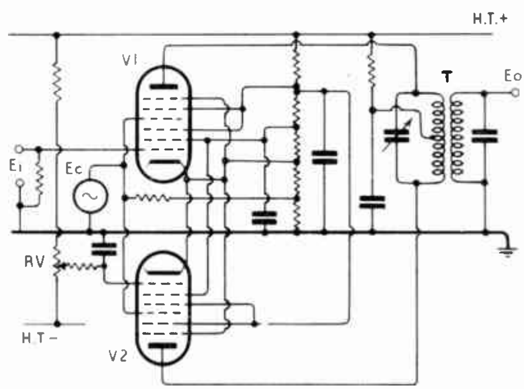


Fig. 14.—Basic circuit of modulator for carrier frequency d.c. amplifier using octode valves.

An input signal of one polarity increases the carrier amplitude, while an input signal of the other polarity reduces the carrier amplitude. It is this fact which enables the information relating to a d.c. input signal to be transmitted by the carrier amplifier although this is in itself essentially a reactance-coupled amplifier. At zero input signal the demodulator is also biased to the middle of its characteristic, thus enabling the d.c. component of the signal to be reproduced. The advantage of this system is that the carrier amplifier is insensitive to supply potential changes, so that, if the amplification after demodulation is relatively small, the only place where special safeguards have to be taken against supply potential changes is in the modulator. This type of amplifier has most advantage where a moderately high gain and a moderately large band-width are required simultaneously, for example in the region of 20,000 gain and 2 Mc/s band-width.

Various forms of modulator circuit may be employed. Fig. 14 shows a basic modulator circuit employing octode valves. The carrier signal E_c generated by the oscillator is fed to the fourth grid of each of a cathode-coupled pair of octode valves V1 and V2. The input signal E_i is fed to the first grid of one valve, V1. The first grid of the other valve V2, is fed from a potentiometer RV, which is so adjusted that, when the input signal is zero, the carrier output is about half the peak available output. The valve V2 in fact produces a constant carrier signal at its anode, say, $a \cos 2\pi ft$, where f is the frequency of the carrier. The valve V1 produces a signal at its anode consisting of the carrier signal modulated by the input signal. If the input signal be represented as $\phi(t)$, then the signal at the anode of the valve V1 may be represented by $b \cos 2\pi ft + \phi(t)b' \cos 2\pi ft$. The signal appearing at the output of the transformer T is proportional to the difference between the signals produced at the anodes of the two valves, and hence may be represented by:

$$k\left[(b-a)\cos 2\pi ft + \phi(t)b'\cos 2\pi ft\right]$$

This is a modulated carrier proportional to the input signal $\phi(t)$ superposed on a constant carrier, representing an arbitrary d.c. level which can be varied by varying a; but a can be controlled by varying the setting of the potentiometer RV, which can therefore be used as the Y shift control of the oscilloscope.

A valve mixer stage of this type has a low conversion efficiency. In fact such a modulator may well produce a loss of 20-fold, so that the gain of the amplifier excluding the modulator must have a gain 20 times higher than the total overall gain required. Thus any noise arising at the input of the carrier amplifier will be amplified 20 times more than would be necessary if there were no loss in the modulator. As noise constitutes a major limitation on the maximum practicable gain of a wide-band d.c. amplifier, it follows that a more efficient modulator is very desirable.

Figure 15 shows the basic circuit of a very useful modulator employing four rectifiers, crystal rectifiers being the most convenient.

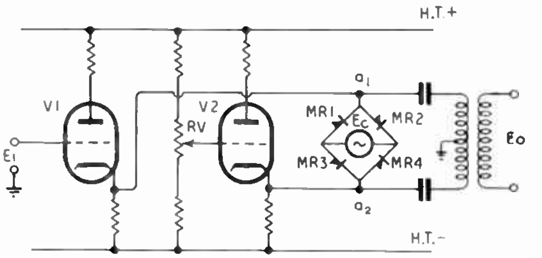


Fig. 15.—Basic circuit of modulator for carrier frequency d.c. amplifier employing crystal rectifiers.

The rectifiers MR1, MR2, MR3 and MR4 are connected in the form of a square, with the carrier signal applied across one diagonal of the square, and the rectifiers so directed that they all conduct during one half-cycle of the carrier signal and are all non-conductive during the other half-cycle. The input signal E_i is fed via the cathode-follower valve V1 on to one end a_1 of the other diagonal of the square, while the potential derived from the potentiometer RV is fed via the cathode-follower valve V2 on to the opposite end a_2 of that diagonal. The output signal E_o at the output transformer is of the same form as that obtained from the modulator of Fig. 14. The crystal modulator of Fig. 15 can, however, be made much more efficient than the valve modulator of Fig. 14.

The potentiometer RV of Fig. 15 may also be used to produce the Y-shift for the oscilloscope. Alternatively, the input signal may be applied in push-pull between the control grids of the valves V1 and V2. The Y shift is then supplied by other means. For example, the input signal may be fed through two cathode followers in cascade on to the crystal modulator,

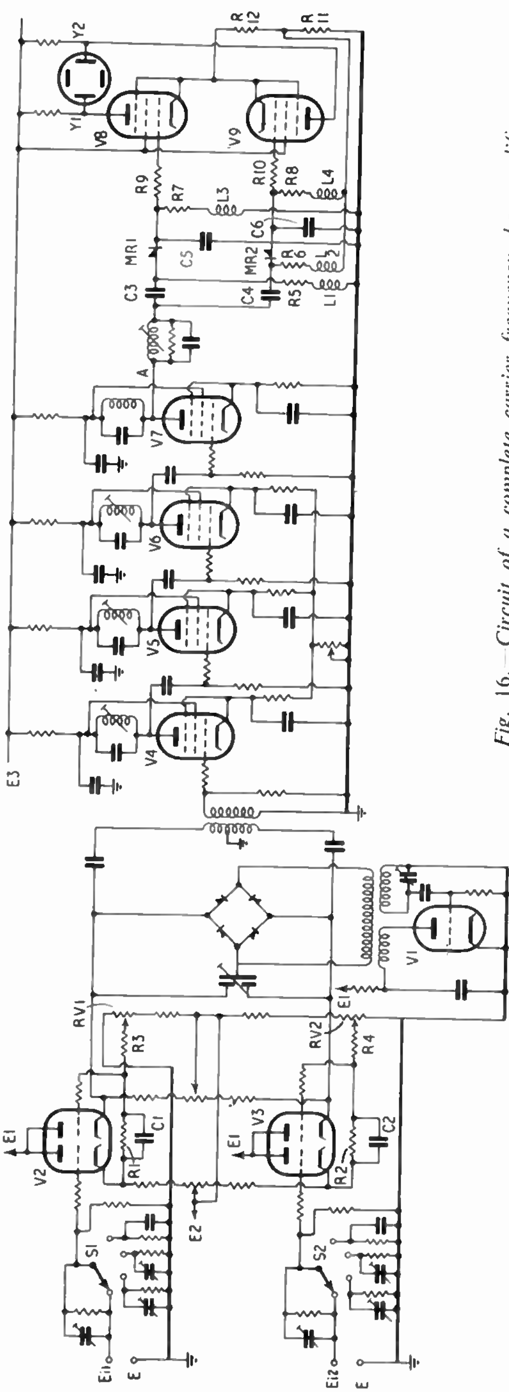


Fig. 16.—Circuit of a complete carrier frequency d.c. amplifier.

the Y shift being applied between the two cathode followers in one or both inputs.

Figure 16 is the circuit diagram of a complete amplifier in which this type of input is employed. The input signal may be applied between earth E and either E_{i1} or E_{i2}, or in push-pull between E_{i1} and E_{i2}. The switches S1 and S2 are ganged, and operate capacitance-corrected resistance attenuators through which the input signal is fed on to one control grid of each of the twin triodes V2 and V3. The signals appearing at the corresponding cathodes are fed through the potential dividers R1, R3 and R2, R4 on to the control grids of the other halves of the twin triodes, the capacitors C1 and C2 providing high frequency compensation for stray capacitance to earth at the control grids. The Y shift is provided by the potentiometers RV1 and RV2, to which the resistors R3 and R4 are returned. These potentiometers are ganged, and may be calibrated directly in volts as described with reference to the corresponding circuit in Fig. 12. The valve V1 is an oscillator supplying the carrier frequency to the modulator.

The output from the modulator is fed to a conventional carrier amplifier comprising the valves V4, V5, V6 and V7. The modulated carrier from the output of the carrier amplifier at A is fed through capacitors C3, C4 to crystal rectifiers MR1, MR2 connected in opposite sense. The rectified signal is developed across the load impedance constituted by the resistor R7 in series with the coil L3 and the resistor R8 in series with the coil L4. The amplifier output is applied to the plates Y1, Y2 of the cathode-ray tube. The channel of the demodulator fed by the capacitor C3 is biased to earth. That fed by the capacitor C4 is biased to a more positive point provided by the potential divider R11, R12 in the cathode circuit of the output valves. Thus at zero carrier level the output valves are biased asymmetrically. The constants of the circuit are so chosen that, at the carrier level corresponding to zero input to the modulator, the output valves are brought to the centres of the useful parts of their characteristics.

The positive supply potential E_1 and the negative supply potential E_2 feeding the modulator should be stabilized. Stabilisation of the supply potential E_3 feeding the carrier amplifier and the output stage is not so important.

4.3. *Drift in D.C. Amplifiers*

One of the main limiting factors on the maximum gain which can usefully be obtained from a d.c. amplifier is drift. Various methods of reducing drift have been tried, but they have severe limitations when applied to oscilloscope amplifiers. One method is to apply negative feedback, that is to say a fraction of the output of the amplifier is fed back to the input in opposite phase to the input signal. This reduces the gain of the amplifier to a value given approximately by the ratio of the impedances determining the feedback. Since this does not depend on the valve characteristics or the supply potentials of the circuit, the gain is thus stabilized. This method has, however, two serious limitations. The first is that the output of the amplifier is only strictly in phase with or in antiphase to the input signal at zero frequency. As the frequency of the applied signal is increased, so the phase difference between input and output increases, the rate of increase depending *inter alia* on the number of stages in the amplifier. The negative feedback therefore decreases with increase of frequency, and, if the band-width of the amplifier is sufficient, eventually becomes positive feedback, which sets the amplifier in oscillation. Thus this method of reducing drift is only applicable to amplifiers of restricted band-width or few stages. The second limitation of this method is that effects of changes in cathode emission of the input valve do not come within the feedback loop, and so are not reduced by the negative feedback. As the input valve is the most sensitive part of the amplifier, this limitation largely off-sets the advantages of the method.

Negative feedback may be applied to each stage of the amplifier instead of from output to input. This greatly reduces the effects of phase change with increase of frequency, but the effect of changes in cathode emission in *each* stage is then excluded from the feedback loop, with corresponding decrease in the effectiveness of the method to reduce drift.

A balanced input of the type shown in Fig. 13 is an effective method of reducing drift.

An important means of reducing drift is the stabilization of the supply potentials. The h.t. supplies, at least to the earlier stages of a high gain d.c. amplifier, should be stabilized. The heater potential should also be stabilized at least for the input valves. Mechanical movement of the cathode relative to the heaters may, however, also cause changes of cathode emission. The valves, especially the input valves, should therefore be of a type so constructed that the electrodes are not liable to change position under the influence of vibration or temperature change. Changes of ambient temperature may cause drift, but it is not practicable to house an oscilloscope amplifier in a thermostatically controlled chamber.

In carrier frequency d.c. amplifiers, the most serious source of drift is at the cathode of the input valves. With a modulator of the type shown in Fig. 15, this source of drift may be avoided by omitting the input cathode followers and feeding the input signal through a resistance on to the modulator at a_1. This, however, results in an input resistance of a few thousand ohms only, which is insufficient for many purposes.

When a very high input impedance is required, drift may occur due to grid current at the input valves. In such cases, therefore, it is necessary to choose input valves which draw little grid current.

5. Associated Equipment

The range of useful applications of an oscilloscope can be considerably extended with the aid of appropriate auxiliary equipment.

5.1. *Probes*

It is very important that connecting the oscilloscope at any point should not affect the signal under investigation. When signals of high frequency are being investigated, the input capacitance of the amplifier of the oscilloscope may easily influence the circuit at the point at which it is connected. The cathode-follower circuit provides a useful means of reducing the input capacitance of the amplifier. If, however, it is necessary to use a long lead to the oscilloscope, the capacitance to earth of the lead may affect the signal. To overcome this, a cathode-follower stage may be mounted in a probe which can readily be applied to the point to be examined, the lead to the oscilloscope being taken from the cathode of the cathode follower. When pulses with short rise-time are being investigated, however, the cathode follower can produce serious distortion if the pulse amplitude exceeds the grid base of the cathode-follower valve. If a positive-going transient of infinitesimally short rise-time is applied to the grid of the cathode-follower valve, the cathode charges positively following in exponential curve of time-constant $\dfrac{C}{g_m + \dfrac{1}{R}}$, where C is the capacitance to earth at the cathode of the valve, g_m is the mutual conductance, and R is the cathode-load resistance. The same applies to a negative-going transient if the amplitude of the transient is less than the grid base of the valve. Normally $1/R$ is small compared with g_m, so that the time constant reduces approximately to C/g_m. If, however, the negative-going transient exceeds in amplitude the grid base of the valve, then the grid is driven negative beyond the cut-off point of the valve current, the cathode being held instantaneously to its original potential by the capacitance to earth at the cathode. When the valve is thus driven beyond cut-off, g_m becomes very small, and the time constant of the exponential curve reduces approximately to CR. Thus a pulse having positive- and negative-going transients of infinitesimally short rise-time is reproduced as a pulse with a positive-going transient following an exponential curve of time constant C/g_m and a negative-going transient following an exponential curve of time constant CR, the "fall" therefore being much slower than the "rise."

If a probe is required for the investigation of pulses of larger amplitude than the grid base of any valve suitable for use in the cathode-follower probe, several ways of overcoming the problem are available. When all the voltages which it is required to investigate are large, it is convenient to employ a simple resistance-capacitance probe providing a fixed attenuation ratio, using a circuit such as that shown in Fig. 17. If $R2$ represents the combined resistance between the amplifier and earth provided by the amplifier input resistance and the resistor R3, then the attenuation ratio is $\dfrac{R_2}{R_1 + R_2}$. The variable capacitor CV is adjusted so that, if C_2 represents the combined capacitance to earth provided by the input capacitance of the amplifier, the capacitance of the co-axial cable C and the capacitor CV, then $C_2R_2 = C_1R_1$. The input capacitance of the probe is equal to the capacitance C_2 which would be present without the probe, reduced by the attenuation ratio.

When the probe is required for use with signals having a range of amplitudes, so that a constant attenuation ratio is not permissible, the problem becomes more difficult. Perhaps

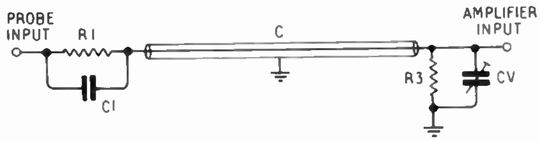


Fig. 17.—Simple resistance-capacitance probe.

the most satisfactory solution is to provide two probes—a cathode-follower probe for small signals and a resistance-capacitance probe for larger signals. A cathode-follower probe alone will suffice if it has an attenuator at the input, but this is liable to make it rather cumbersome. Alternatively it is possible to employ a corrected cathode-follower probe in which the distortion arising with large pulses in a simple cathode follower is compensated. A circuit of such a probe is shown in Fig. 18. The valve V1 is the cathode follower, the output being taken from a potential divider R2, R3 connected to the cathode and so chosen that the d.c. output is zero for zero input. The valve V2 is the correcting valve, and is connected to form the cathode load of the cathode follower. The cathode follower is provided with an anode load R1, and the anode is coupled to the grid of the correcting valve by a

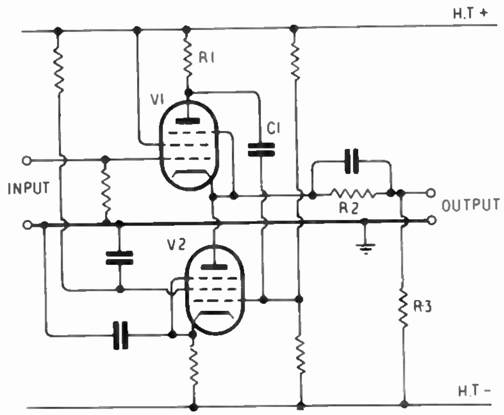


Fig. 18.—Corrected cathode-follower probe.

capacitor C1. When the cathode follower is functioning normally, so that the cathode "follows" the grid, the current change in the valve V1 is very small. Consequently the signal produced at the anode is very small, and the correcting valve therefore receives only a very small signal through the capacitor C_1. If, however, the grid of the cathode follower is driven suddenly negative to such an extent that the cathode current is cut off, then the anode receives a positive pulse which is conveyed through the capacitor C1 on to the grid of the correcting valve. This produces a sudden increase in current in the correcting valve which serves to drive the cathode of the cathode follower negative, thus forcing it to "follow" the grid.

5.2. *High Discrimination Input Units*

The term "discrimination" is applied to amplifiers having a push-pull input, to indicate the ratio of the push-pull output provided per unit push-pull input signal to the push-pull output resulting from unit in-phase, or "push-push," signal. High discrimination is often very important when the cathode-ray oscilloscope is employed to investigate electrical signals arising in living organisms. Such investigations are being carried out more and more extensively in biological and medical work. The body of an animal or person under investigation normally has a high resistance to earth, and a large capacitance to surrounding objects. This situation is one where unwanted signals from surrounding objects, such as mains hum, are very liable to be picked up. Furthermore, the actual signals under investigation, the signals arising in heart, muscles, nerves or brain, are always small, and so are easily swamped by signals picked up from the surroundings. If, however, the difference of potential is investigated between two points in the body of the subject, then the signals picked up from the surroundings at the two points are usually very nearly equal. Consequently, if the amplifier has a high discrimination, the push-pull output produced by the signals picked up from the surroundings will be small, and any "push-push" output is not reproduced on the cathode-ray tube.

In practice, the "long-tailed pair" circuit shown in Fig. 13 provides quite a high discrimination. This can be further increased by employing a pentode valve in place of the common cathode resistor. It is possible to provide a preset adjustment for the discrimination by feeding the screen grids of the input valves to variable supply points; a potentiometer connected between the screen grids with the slider taken to the supply potential is convenient. A suitable circuit is shown in Fig. 19.

5.3. *High Resistance Input Units*[7]

In some kinds of work it is required to investigate very small currents, or potential variations arising at points of very high impedance, for example, in investigating phenomena in ionized solutions, for certain investigations in biological work and the like. For such purposes a very high input resistance at the amplifier is necessary, and to obtain this a special input unit is used employing an electrometer valve. This is essentially a thermionic valve run at low heater current and supply potentials, with the object of minimizing grid current. Using a good quality electrometer valve, an input resistance of 10^{10} ohms can be employed without undue difficulty. In order to avoid mains hum being developed across the very high input resistance, it is preferable that the electrometer valve should be battery operated.

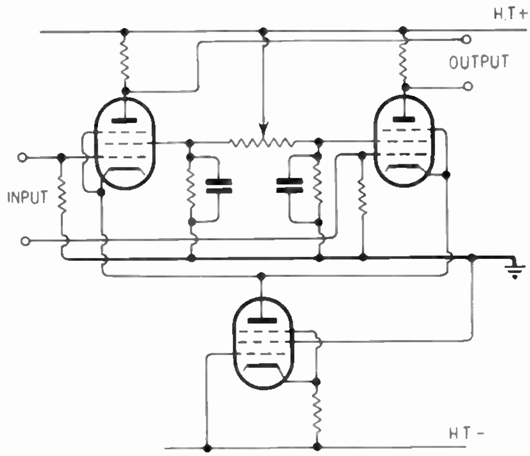


Fig. 19.—High discrimination input unit.

5.4. *Beam-Blanking Units*

Usually the beam of the cathode-ray tube is switched on during the scan of the time base and suppressed during the fly-back by a brightening pulse derived from the time base generator. For some purposes, however, it is desirable to be able to brighten or suppress the beam by a manually or mechanically controlled switch. For example, a moving film is sometimes used to record phenomena occurring over a longer time than the period of one scan of the time base at the time-base speed required, or phenomena of such low frequency that it is difficult to produce electronically a time base slow enough for satisfactory recording of the phenomena. In such cases the moving film itself constitutes the time base and the time-base generator of the oscilloscope is not used. It is then convenient that a brightening pulse should be actuated mechanically from the drum of the moving film camera when the film commences

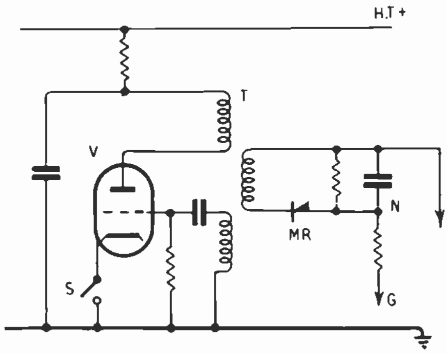


Fig. 20.— Beam-blanking unit.

to move and should be suppressed after one rotation, or after a specified number of rotations of the camera drum. For this purpose a beam-blanking unit is provided, controlled by a switch operated from the drum of the camera. Since the duration of the brightening pulse may well be required to be relatively long, at least of the order of seconds, capacitive coupling to the grid of the cathode-ray tube is unsuitable, and direct coupling is needed. This is difficult since the grid is usually at about 2 kV or more negative to earth. A convenient solution is provided by use of the principle of the carrier frequency d.c. amplifier. Fig. 20 shows a suitable circuit. The valve V with its associated components constitutes an oscillator producing the carrier frequency. This is coupled through a transformer T to a rectifier MR so that, when the oscillator is working, a negative d.c. potential is produced at the point N relative to the point B. The point B is connected to the intensity-control potentiometer of the cathode-ray oscilloscope, while the point G is taken to the grid of the cathode-ray tube. The oscillator can be switched on or off by the switch S, which is operated by the camera drum or other means by which it is desired to control the beam intensity. The beam is adjusted for the desired intensity with the switch S open. Closing the switch S then suppresses the beam. When the switch is subsequently opened the beam is then switched on at the correct intensity and remains on until the switch is again closed.

While it is common practice when using the time-base generator to couple the brightening pulse to the grid of the cathode-ray tube through a capacitor, direct coupling has inherent advantages. At low repetition rates capacitive coupling presents the usual difficulties of providing a sufficiently large capacitance to avoid discharge during the cycle of operations, and this is aggravated by the fact that the coupling capacitor requires to withstand the large voltage between the time-base generator and the grid of the cathode-ray tube. There is, however, a further disadvantage in capacitive coupling. The cathode of the cathode-ray tube is tied to the negative supply line of the cathode-ray tube supply unit, and so receives directly any fluctuations in supply potential. The grid, however, is tied through the coupling capacitor to the time-base generator, which is relatively near earth potential. Consequently the grid bias potential fluctuates with fluctuations in supply potential, causing changes in the intensity of the beam or even extinguishing it completely for a short time. This effect is very troublesome if the oscilloscope is fed from a mains supply which is subject to appreciable surges. It can be completely eliminated by using a beam-brightening system similar to that shown in Fig. 20. In this case the switch S of Fig. 20 must be replaced by an electronic switch operated by the time-base generator, which may either function to turn the valve V on and off, or be interposed between the valve V and the transformer T.

5.5. *Delay Lines*

A special problem arises where it is required to observe a pulse of very short duration, due to the finite time which inevitably elapses between the arrival of a triggering pulse at the trigger input terminal of the time-base generator and the actual start of the time-base sweep. If the pulse is simultaneously applied to the Y deflection plates of the cathode-ray tube, either directly or through an amplifier, a substantial portion, or even the whole, of the pulse may be lost to view due to the fact that this part of the pulse has already occurred before the time-base sweep has commenced. This difficulty may be overcome by delaying the arrival of the pulse at the Y deflection plates of the cathode-ray tube by passing the pulse through a delay line. A compact delay line is provided by a network consisting of a number of equal coils connected in series, with each junction connected through a capacitor to earth, all the capacitors also being of equal value. It is essential that the delay line should be terminated by a resistance equal to the characteristic impedance of the line, otherwise reflections will occur, distorting the pulse. If n is the number of coils employed, L the inductance of each coil and C the capacitance of each capacitor, then the delay produced is $n\sqrt{(LC)}$, and the characteristic impedance of the line is $\sqrt{(L/C)}$.

6. Acknowledgments

Many of the circuits described in this paper have been developed in the research laboratories of Messrs. Nagard, Ltd., of Belmont, Surrey. The author is indebted to the company for the loan of apparatus for demonstration at the oral presentation of the paper.

The author would also like to express his thanks to his colleagues in the company for helpful discussions, drawing of diagrams and reading of proofs, and in particular to Mr. M. J. G. Hinton for working the apparatus at the oral presentation of the paper.

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EXAMINATION PRIZE WINNERS—1953

THE PRESIDENT'S PRIZE

Meir Weger, who receives the President's Prize as the most outstanding candidate taking all parts of the Graduateship examination at one sitting, was born in Germany in 1932 and went to Israel in 1936. He continued his secondary education by evening study while apprenticed as a radio technician, and after matriculating in 1950 completed a year's study at the Jerusalem University before joining the Israel Air Force for two years' service. He has now returned to the university to study for a physics degree. Mr. Weger registered as a Student of the Institution in 1953 and was transferred to Graduate in May, 1954, following his success in the November, 1953, examination.

THE S. R. WALKER PRIZE

S. Janakiraman, recipient of the S. R. Walker Prize, was born at Virudhunagar, South India, in 1928. He holds various City and Guilds of London Institute certificates in Radio, and since 1952 has been employed at Nagpur as a radio technician in the Directorate of Co-ordination of the Government of India, where his duties are concerned with the maintenance of police radio equipment. Mr. Janakiraman registered as a Student of the Institution in 1952, and in May, 1953, sat for the Graduateship examination, being placed second among candidates passing all parts of the examination at one sitting. He was transferred to Graduateship in November, 1953.

ELECTRONIC MEASUREMENTS PRIZE

Srinivasa Ramabhadran, who was born at Terizhandur in 1918, has been awarded the Electronic Measurements Prize. He was educated at St. Joseph's College, Trichinopoly, obtaining a B.Sc. degree in 1940; he received an M.A. degree in Physics from Madras University in 1946, and a Diploma in Electrical Communication Engineering of the Indian Institute of Science, Bangalore, in 1947. Mr. Ramabhadran is at present a lecturer in communications at the College of Military Engineering at West Kirkee, and has held similar appointments at the University of Roorkee and the Indian Army School of Signals, Mhow. He registered as a Student in 1949 and, on completing the Graduateship examination, was transferred to Associate Membership in January, 1954.

AUDIO FREQUENCY ENGINEERING PRIZE

Peter Gareth Lovell is the recipient of the Audio Frequency Engineering Prize. He was born in Aberystwyth in 1926 and was educated at Lytham St. Annes. Following part-time study at the Regent Street Polytechnic and Acton Technical College he obtained a B.Sc. degree in Physics and Higher National Certificate with endorsements in Radio and Electro-acoustics. From 1945 to 1952 Mr. Lovell was a technical assistant, first with British Relay Wireless Co. and later with Philips Electrical, Ltd., and he is now a development engineer with the Plessey Co. Ltd. He registered as a Student in March, 1953, and was transferred to Graduate six months later, following his completion of the examination requirements.

UNBALANCE EFFECTS IN MODULATORS*

by

D. G. Tucker, D.Sc., Ph.D. (*Member*)†

SUMMARY

The various groups of output frequency-components which can be produced in rectifier modulators of the shunt (or "Cowan") and ring types are separated into those which are inherent in the operation of the modulator and those which are absent when the modulator is perfectly balanced. It is shown that in the shunt modulator only one balance control potentiometer is needed, and only one unbalance component can, in general, be brought to a real minimum by a particular adjustment; nevertheless, all the unbalance components tend to vary (and balance) together. In the ring modulator, two independent balance controls can be provided, and two (though not any two) unbalance components can be simultaneously brought to a minimum. The effect of the signal voltage on the magnitude of unbalance output-components is examined, and shown to be zero in a modulator which is perfectly balanced in the absence of signal, and to be noticeable only near the overload point even when an initial unbalance exists.

LIST OF SYMBOLS

f_c = carrier frequency.

f_s = signal frequency.

m may be zero or any positive integer.

r may be zero or any positive integer.

n may be any positive integer.

$r_f(v)$ = forward resistance of rectifiers as a function of voltage.

E = signal e.m.f. = $E_0 \cos 2\pi f_s t$.

V_c = carrier voltage across each rectifier.

R = resistance of external circuit (see Figs. 1 and 2).

V_{cL} = carrier-leak voltage.

$V_{cL}(t)$ = carrier-leak voltage as a time-function.

α = coefficient of square term in rectifier resistance/voltage law.

β = coefficient of cube term in rectifier resistance voltage law.

v_s = signal voltage.

v_0 = displacement of rectifier characteristic along voltage axis, causing unbalance in modulator.

a = $\frac{1}{2}\times$ carrier current in Appendix 1, and equals carrier voltage in Appendix 2.

b = $\frac{1}{2}\times$ signal current in Appendix 1, and equals $\frac{1}{2}\times$signal voltage in Appendix 2.

c_n = coefficient of nth power term in power-series expansion of rectifier voltage in terms of current.

d_n = coefficient of nth power term in power-series expansion of rectifier current in terms of voltage.

sign = "signature."
(sign $x = +1$ when $x>0$
and -1 when $x<0$.)

1. Introduction

Balanced modulators are in common use in a variety of applications, and no doubt reliance is usually put on the partial suppression of certain output components as a result of the nominal balance of the circuit. However, the number of output components which are produced in an unbalanced modulator in addition to those produced in a perfectly balanced one is very large, and the conditions for attaining a high degree of balance in a practical modulator are usually rather complicated. It is therefore rather surprising that so little has been published on the subject of unbalance effects. References 1 to 5 cover the existing literature known to the author. There is a particular lack of discussion of any aspect of the subject other than that of "carrier-leak," i.e., the direct leakage of the local, or "carrier" or "switching," oscillation. In practice, however, other unbalance effects

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are often important, e.g. even-order intermodulation and harmonic distortion, input-signal leak, the worsening of the carrier-leak as the input signal voltage is raised, etc. The discussion below attempts to clarify the overall picture of the subject by showing how the various groups of unbalance output components are produced and how they can be balanced out by simple circuit adjustments in the well-known shunt (or "Cowan") and ring rectifier modulators. The influence of signal voltage on carrier-leak is included. The discussion is illustrated by three simple mathematical appendixes and by experimental results.

2. Balancing the Shunt Modulator

The shunt modulator, shown in Fig. 1, is the simpler one to deal with, as only one balance control can be provided. The control potentiometer can be fitted anywhere in the bridge, but always affects the same group of output components. Ideally, i.e. with perfect-switching

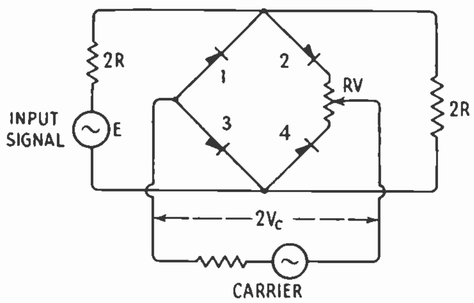

Fig. 1.—Shunt (or "Cowan") modulator.

(or "linear") identical rectifiers and a square-wave carrier, and with the signal level below overload point, the output contains only the input signal (frequency f_s) and the sidebands of odd orders of modulation, $(2m + 1)f_c \pm f_s$, where m may be zero or any positive integer. If a sine-wave carrier is used (or any non-square waveform), but still with perfect-switching rectifiers, then non-linear distortion[6] occurs due to the interference of the input signal with the switching of the rectifiers, which should be, ideally, entirely controlled by the carrier. These distortion products are of the type $(2m + 1)f_c \pm (2n + 1)f_s$, where n takes all integral values from unity upwards. When a sine-wave carrier is used with real rectifiers, with their gradual transition from high to low resistance and vice versa, even orders of modulation are introduced since the modulating function is no longer symmetrical about its mean value.[1] Thus output components of the type $2mf_c \pm (2n - 1)f_s$ appear, generally at a fairly low level compared with the others. These components are not due to unbalance, and cannot therefore be eliminated by adjustment of resistances.

If the four real rectifiers are not identical, other output components of the following types appear due to unbalance:

(*a*) Carrier-leak and its harmonics, nf_c.

(*b*) Even-order distortion products of even-order modulation, $2mf_c \pm 2nf_s$, which are produced in each individual rectifier even when these are perfect switches, but ideally cancel out at the output terminals.

(*c*) Even-order distortion products of odd-order modulation, $(2m + 1)f_c \pm 2nf_s$, which are not produced at all in perfect-switch rectifiers, and are probably always negligible except with large signal or low carrier voltages.

(*d*) A d.c. component.

All these components can be grouped into one main class:—$mf_c \pm 2rf_s$, where r, like m, may be zero or any positive integer. All members of this class can be reduced to a minimum by adjustment of the potentiometer RV. Generally, at low frequencies where capacitive unbalances are insignificant, any one component may be brought near to zero, but the optimum balance point is usually different for the different components. Appendix 1 gives a more detailed explanation of how all these various frequency components arise and how balancing controls them. Some experimental results on a shunt modulator, using four germanium rectifiers (not specially selected) at audio-frequencies, are shown in Fig. 2, which is a correlation diagram in which the output of $2f_c$ has been arbitrarily chosen as abscissa, and where the various points were obtained by varying the setting of RV. It will be seen that the products f_c, $2f_c$ and $2f_c - 2f_s$ have balance points fairly close together and vary together. When the potentiometer was placed at the junction of rectifiers 1 and 2 instead of as in Fig. 1, rather worse results were obtained; the balance points agreed much less closely, and the lines in the graph did not approximate to straight lines. In these experiments, the output of $2f_c \pm f_s$ was 1·15 mV and did not vary by more than $\pm$ 0·5 db over the range of potentiometer settings involved in Fig. 2.

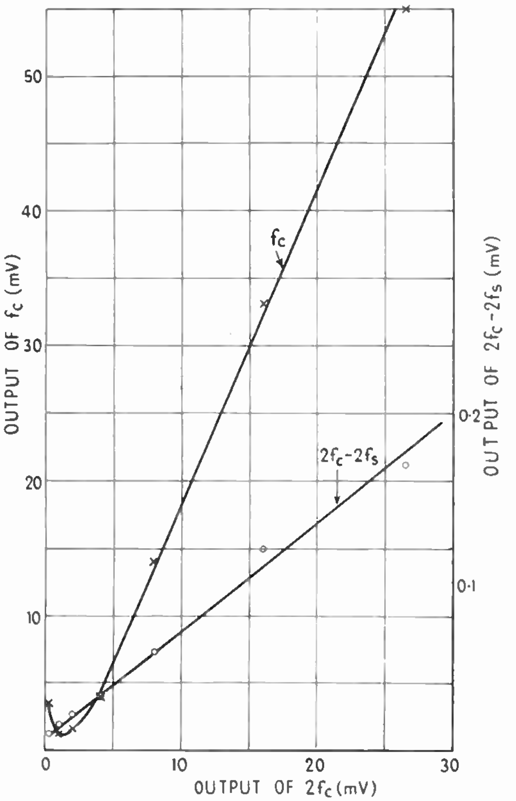

Fig. 2.—Shunt modulator. Effect of balancing potentiometer in position shown in Fig. 1.

Experimental results:
germanium rectifiers; $R = 1300\Omega$; $f_c = 3$ kc/s;
$f_s = 300$ c/s; $V_c = 0·6$ V approx.;
output of $f_c \pm f_s = 70$ mV.

It should be noted that a d.c. component is produced by unbalance, and the fact that the balance adjustment for this usually gives a fairly good (though not optimum) balance for the other components leads to a simple means of adjusting and maintaining carrier-leak by a d.c. meter.[2] This is discussed in detail, together with an account of the phenomena associated with capacitive unbalance, in a forthcoming paper by Connon.[7]

3. Balancing the Ring Modulator

The ring modulator, shown in Fig. 3, is much more complex than the shunt modulator, as the four positions of balancing potentiometers shown in the diagram all have different effects. With identical linear rectifiers and square-wave carrier, the only output components are $(2m + 1)f_c \pm f_s$; and with identical linear rectifiers and sine-wave carrier, the non-linear distortion products

$$(2m + 1)f_c \pm (2n + 1)f_s$$

are added. Identical real rectifiers would not alter this situation, since the modulating function would be symmetrical about its mean level (of zero). All other components which can be introduced are due to unbalance.

Thus, the unbalance output components comprise:

(*a*) Carrier-leak, nf_c.

(*b*) Input signal leak, nf_s (although amplitudes for $n > 1$ are usually negligible).

(*c*) Sidebands of even-order modulation, $2mf_c \pm f_s$.

(*d*) Even-order distortion products of even-order modulation, $2mf_c \pm 2nf_s$.

(*e*) Odd-order distortion products of even-order modulation, $2mf_c \pm (2n + 1)f_s$.

(*f*) Even-order distortion products of odd-order modulation, $(2m + 1)f_c \pm 2nf_s$.

(*g*) A d.c. component, although this is lost unless a variant of the ring modulator which has no output transformer[3] is used.

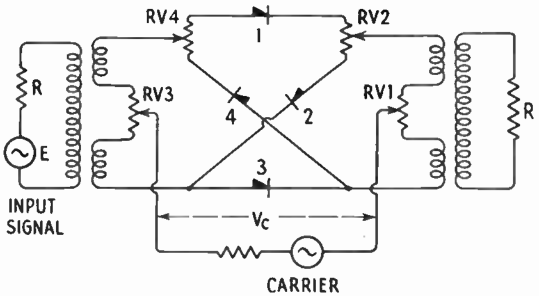

Fig. 3.—Ring modulator.

Of these, (*e*) and (*f*) are thought to be insignificant in practice, except with large signal or low carrier voltage; they do not occur with perfect-switch rectifiers.

These products, (*a*) to (*g*), can be grouped in three main classes:

$$(2m + 1) f_c \pm 2rf_s \quad \ldots \quad (A)$$

$$2mf_c \pm 2rf_s \quad \ldots \quad (B)$$

$$2mf_c \pm (2r + 1) f_s \quad \ldots \quad (C)$$

where m, as before, takes all integral values from zero upwards, and r does the same. These three classes have different balancing properties.

For components of class (A), the fundamental carrier-leak ($m = 0$, $r = 0$) may be taken as typical, and this is produced when the carrier current divides unequally into the two halves of the primary of the output transformer, in such a way that there is a resultant flux which is different in magnitude and/or polarity on the positive and negative carrier half-cycles.

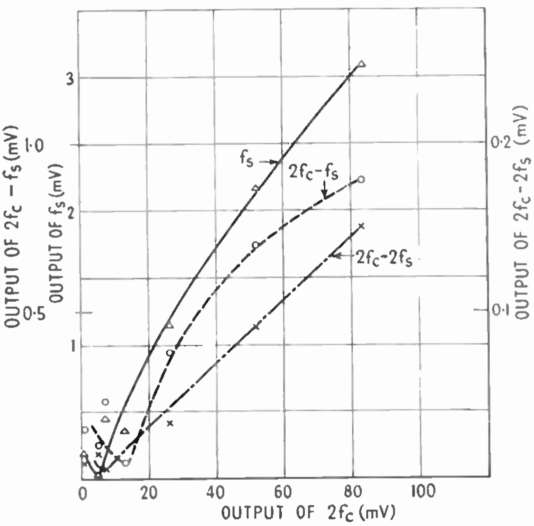

Fig. 4.—Ring modulator. Effect of balancing potentiometer RV2.

Experimental results: germanium rectifiers; $R = 5\text{k}\Omega$; $f_c = 3$ kc/s; $f_s = 300$ c/s; $E/V_c = 0{\cdot}3$ approx; $V_c = 0{\cdot}5$ V peak; output of $f_c \pm f_s = 47$ mV.

This unbalance can therefore be corrected by adjustment of either RV1 or RV4.

For components of class (B), the second harmonic of carrier-leak ($m = 1$, $r = 0$) may be taken as typical, and this is produced when the carrier current divides in the output transformer in such a way that there is a resultant flux which is not equal and opposite on both positive and negative carrier half-cycles. This unbalance can therefore be corrected by adjustment of either RV2 or RV3.

For components of class (C), the input-signal leak ($m = 0$, $r = 0$) may be taken as typical, and this is produced when the signal current flowing through the output transformer is not equal and opposite on both positive and negative carrier half-cycles. This unbalance can therefore be corrected by adjustment of either RV2 or RV4.

All the higher-order components included in classes (A) to (C) flow in the individual rectifiers, and the argument regarding balancing is the same as for the typical components chosen. A detailed illustration of the various effects discussed above is given in Appendix 2.

Components of all classes can thus be simultaneously adjusted by the use of *either* RV1 and RV2, *or* RV3 and RV4.* It will be exceptional, of course, for the optimum settings of each potentiometer to be identical for all components controlled by it, since this would only occur if the system were non-reactive and the only difference between one rectifier and another were a fixed resistance which did not vary with the voltage across the rectifiers. But usually, in practice, one pair of settings gives a reasonable suppression of all unbalance output components, as is illustrated by measured results shown in Fig. 4, made on a typical ring modulator using germanium rectifiers chosen at random from a large batch. It will be seen that four important components controlled by RV2 give optimum balances fairly close together, and tend to vary together. The various points on the graph were obtained by adjusting RV2.

In practice it will usually prove convenient to fit a fixed resistor in the connection to the bottom side of the primary of the output transformer in order to compensate for the resistance of RV2 in the other side—if this is not done, RV1 will need to be larger than otherwise necessary in order to cope with the longitudinal unbalance introduced. A similar argument applies to RV4 and RV3.

*It should be noted that adjustment of RV1, by slightly redistributing the voltages across the rectifiers, does slightly change their resistances and so has a second-order effect on the adjustment of RV2. Similarly RV2 reacts slightly on RV1, and RV3 on RV4 and vice versa. These interactions show up in practice only when a high degree of balance is sought.

4. The Influence of the Signal Voltage on Unbalance

There appears to have been no published discussion of this matter, although it is probably well known that carrier-leak, which is usually adjusted in the absence of signal, can be altered by the application of the input signal.

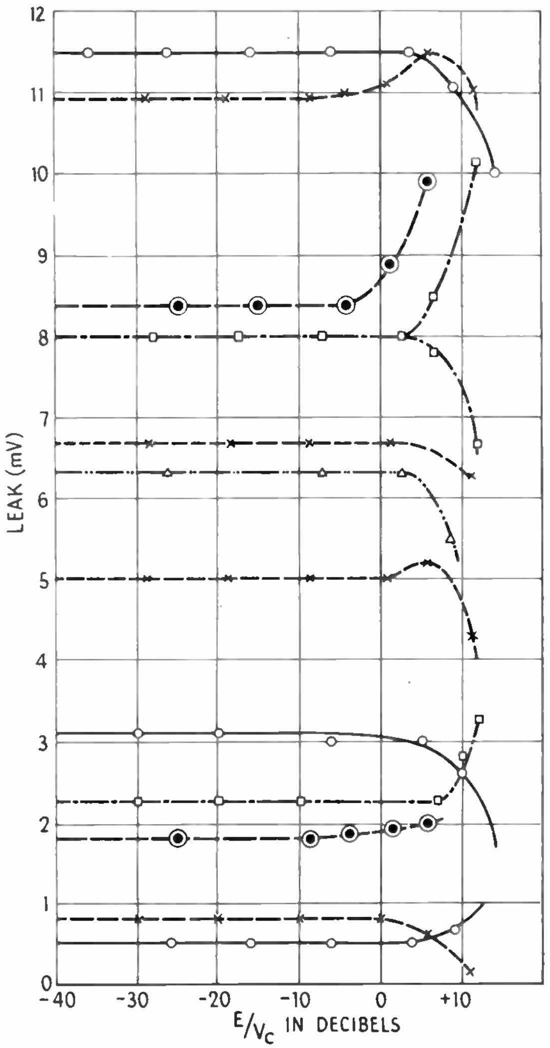

Fig. 5.—Effect of Signal Level.

Experimental results:

—o—o— = shunt, germanium, R = 2·5kΩ, square carrier
—x— x — = shunt, germanium, R = 2·5kΩ, sine carrier
—□—·—□— = shunt, copper oxide, R = 2·5kΩ, sine carrier
} f_c = 3 kc/s, f_s = 200 c/s

—Δ- ·· -Δ— = shunt, copper oxide, R = 2·5kΩ, sine carrier, f_c = 15 kc/s

—⊙— —⊙— = ring, germanium, R = 5kΩ, sine carrier, f_c = 3 kc/s

If a modulator is perfectly balanced (i.e. has all rectifiers identical) in the absence of a signal, then the application of a signal of perfectly symmetrical waveform (i.e. positive and negative excursions identical) cannot cause unbalance products to be formed whatever the shape of the rectifier characteristics. This conclusion, although perhaps an intuitive one, is probably readily acceptable as "obvious." However, proof for two very different cases is available. Firstly, the ideal ring modulator with perfect-switching rectifiers but with a sinusoidal carrier voltage, discussed in Reference 6, has a total output given by equation (11) of the paper cited; this equation includes the effect of signal voltages on the switching of the rectifiers, but can easily be shown to contain no components at the carrier frequency or its harmonics, nor at the other frequencies which appear due to unbalance. Secondly, Appendix 3 shows that a simple shunt modulator with rectifiers very different from perfect switches has no carrier-leak if the rectifiers are identical, irrespective of the signal voltage. No general treatment of the effect of signal voltages on the unbalance of a modulator which is already unbalanced in the absence of the signal can be attempted here, but Appendix 3 shows that in a simple case which is easily analysed the effect definitely exists but is small, only becoming noticeable as the overload point is approached. Theory such as this cannot be expected to be adequate for prediction of practical carrier-leak magnitudes, and serves primarily to elucidate the principles. Nevertheless, it is clear that the effect is very dependent on the actual shape of the rectifier characteristics and on the nature of the initial unbalance, and will, in practice, be very hard to predict.

Figure 5 shows the results of numerous measurements of the effect of signal voltage on the fundamental component of carrier-leak in shunt modulators using germanium and copper-oxide rectifiers, with square-wave and with sine-wave carriers, and, in the case of copper-oxide rectifiers, with a higher carrier frequency so that reactive effects might be included with a low backward resistance. Measurements on a ring modulator are also included. In all cases the effect is of the nature predicted above, and the order of magnitude does not conflict too seriously with that suggested by equations (6) and (7) of Appendix 3. Equation (7) can account only for an increase in leak, but since increases and decreases occur in practice, it is clear that equation (6) may be more realistic; in this equation the effect of the signal depends, both in magnitude and sign, on the relative magnitudes and sign of the coefficients a and b of the rectifier resistance/voltage law.

One particularly interesting feature in Fig. 5 is the result using the shunt modulator with copper-oxide rectifiers and $f_c = 3$ kc/s, the initial leak being 8 mV. Here the balancing potentiometer was adjusted first to one side of optimum balance, and then to the other side, the leak being set to 8 mV in each case. The effect of the signal voltage was opposite, however, as one side caused a decrease due to signal and the other side caused an increase. Although this type of result was often obtained, it was by no means the rule.

5. Conclusions

The effects of unbalance in rectifier modulators of the shunt (or "Cowan") and ring types have been discussed in relation to all output components which can be produced, and it has been shown that the use of balancing potentiometers can lead to substantial suppression of the following types of product:

Shunt Modulator: $mf_c \pm 2rf_s$

Ring Modulator: $(2m + 1)f_c \pm 2rf_s$

$2mf_c \pm 2rf_s$

$2mf_c \pm (2r + 1)f_s$

where m and r take any positive integral value or may be zero. In the shunt modulator only one balance control is possible, and thus, in general, only one of the products can be brought to a real minimum by a particular adjustment; nevertheless, all the unbalance products tend to vary (and balance) together. In the ring modulator, two independent balance controls can be provided, and one product from each of two of the three classes shown above can be simultaneously balanced to a true minimum, and a fair overall balance of all three classes of product can usually be achieved.

The types of output product which cannot be balanced out at all are:

Shunt Modulator: $mf_c \pm (2r + 1)f_s$

Ring Modulator: $(2m + 1)f_c \pm (2r + 1)f_s$

The effect of the signal voltage on the magnitude of the carrier-leak (i.e., $m = 1$, $r = 0$ in the class $mf_c \pm 2rf_s$ or $m = 0$, $r = 0$ in the class $(2m + 1)f_c \pm 2rf_s$) has been examined, and shown to be zero in a modulator which is perfectly balanced in the absence of signal, and to be noticeable only near the overload point even when an initial unbalance exists.

6. Acknowledgments

This paper is published by permission of the Admiralty.

7. References

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8. Appendix 1: Types of Output Product in the Shunt Modulator

It may not be immediately clear why the output products of the shunt modulator with real rectifiers and non-square carrier waveform fall into the two classes:

$mf_c \pm 2rf_s$ which can be balanced out, and

$mf_c \pm (2r + 1)f_s$ which cannot be balanced out.

A brief discussion of a simple case should readily clarify this matter.

Consider a non-reactive shunt modulator which has "constant-current" signal and carrier sources, and takes its output as the voltage developed across the rectifier bridge. The voltage across any individual rectifier can be related to the current through it by a power series of the type

$$v = \sum_{n=0}^{\infty} c_n i^n \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

and it is assumed that the rectifier resistance does not become infinite since the "constant-current" drive must always be maintained. Assume for simplicity at this stage that coefficients c_{n1} apply to rectifiers 1 and 4, and c_{n2} to rectifiers 2 and 3. This justifies the assumption of equal division of current between the two branches of the carrier circuit and between the two branches of the signal circuit. Then if $I_c \cos 2\pi f_c t = 2a$ is the carrier current and $I_s \cos 2\pi f_s t = 2b$ is the signal current, then the current in rectifiers 1 and 4 is $(a - b)$ and that in rectifiers 2 and 3 is $(a + b)$, taking the direction of best conduction of the rectifiers as the positive flow. The output voltage of the modulator is therefore

$$\sum c_{n2}(a+b)^n - \sum c_{n1}(a-b)^n \quad \ldots . (2)$$

$$= \tfrac{1}{2}\sum (c_{n2} + c_{n1})\left[(a+b)^n - (a-b)^n\right]$$
$$+ \tfrac{1}{2}\sum (c_{n2} - c_{n1})\left[(a+b)^n + (a-b)^n\right] \quad \ldots\ldots\ldots\ldots (3)$$

When the modulator is perfectly balanced, $c_{n2} - c_{n1}$ is zero for all values of n, and only the first term in (3) remains. Thus the output products which cannot be balanced out are those contained in

$$\sum (a+b)^n - \sum (a-b)^n \quad \ldots\ldots (4)$$

while the products which can be eliminated by balancing are those contained in

$$\sum (a+b)^n + \sum (a-b)^n \quad \ldots\ldots (5)$$

The expansion of (4) and (5) for $n = 4$ and 5 will sufficiently demonstrate the effect. Thus

$$(a+b)^4 - (a-b)^4 = 8a^3b + 8ab^3$$
$$(a+b)^5 - (a-b)^5 = 10a^4b + 20a^2b^3 + 2b^5$$

giving rise to the frequencies f_s, $f_c \pm f_s$, $2f_c \pm f_s$, $3f_c \pm f_s$, $4f_c \pm f_s$, $3f_s$, $f_c \pm 3f_s$, $2f_c \pm 3f_s$, $5f_s$, i.e. the group $mf_c \pm (2r + 1)f_s$. These cannot be balanced out.

Also

$$(a+b)^4 + (a-b)^4 = 2a^4 + 12a^2b^2 + 2b^4$$
$$(a+b)^5 + (a-b)^5 = 2a^5 + 20a^3b^2 + 10ab^4$$

giving rise to the frequencies 0, f_c, $f_c \pm 2f_s$, $f_c \pm 4f_s$, $2f_c$, $2f_c \pm 2f_s$, $3f_c$, $3f_c \pm 2f_s$, $4f_c$, $5f_c$, $2f_s$, $4f_s$, i.e. the group $mf_c \pm 2rf_s$. These can be balanced out. The condition for *simultaneous* balancing-out of all such products by means of a resistance adjustment is evidently

$$c_{n1} - c_{n2} = \text{constant for all values of } n,$$

where it is assumed that c_{n1} applies to rectifiers 1 and 4 and c_{n2} to rectifiers 2 and 3. If all four rectifiers are different the condition becomes

$$c_{n1} - c_{n2} - c_{n3} + c_{n4} = \text{constant}.$$

These conditions are unlikely to be met in practice.

9. Appendix 2: Types of Output Product in the Ring Modulator

Following the lines of Appendix 1, the working out of a theoretically simple case of the ring modulator for real rectifiers and non-square carrier waveform should help the understanding of the general statements of Section 3.

Consider a non-reactive ring modulator which has "constant-voltage" (or zero-resistance) signal and carrier sources, and takes its output as the current through a terminating resistance of zero. If $V_c \cos 2\pi f_c t = a$ is the carrier voltage and $V_s \cos 2\pi f_s t = 2b$ is the signal voltage, then the voltages across the four rectifiers are $(a + b)$, $-(a - b)$, $(a - b)$ and $-(a + b)$ for rectifiers 1 to 4 respectively. As in Appendix 1, polarity is taken with respect to the direction of best conduction of the rectifiers. Let the current through a rectifier be related to the voltage across it by the power series

$$i = \sum_{n=0}^{\infty} d_n v^n \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

In general the four rectifiers have different sets of coefficients, d_{n1}, d_{n2}, d_{n3} and d_{n4}, and in these circumstances the assumption made above that half the signal voltage appears across each rectifier is not strictly justified. But it greatly simplifies the calculation and leads to an error only in the amplitude of unbalance components; it does not affect the nature of the result nor the conditions for balance, discussed later. The assumption is completely justified if balancing condition (9a)—see below—applies, whether the other conditions (7) and (11) are met or not.

The output current is $i_1 - i_2 - i_3 + i_4$

$$= \sum d_{n1}(a+b)^n - \sum d_{n2}(-1)^n(a-b)^n - \sum d_{n3}(a-b)^n + \sum d_{n4}(-1)^n(a+b)^n \quad \ldots\ldots(2)$$

$$= \sum [d_{n1} + (-1)^n d_{n4}](a+b)^n - \sum [(-1)^n d_{n2} + d_{n3}](a-b)^n \quad \ldots\ldots(3)$$

If all d_n are equal, clearly all terms in which n is odd cancel out, and the even terms leave only

$$2\sum d_n(a+b)^n - 2\sum d_n(a-b)^n, \; n \text{ even} \quad \ldots(4)$$

which is therefore the output which cannot be balanced out. By expanding a few terms as in Appendix 1, it is quickly seen that this contains only frequencies of the type $(2m+1)f_c \pm (2r+1)f_s$. The remaining terms when the d_n are unequal are those which can be balanced out, and can similarly be seen to fall in the three classes

$$(2m+1)f_c \pm 2rf_s \quad \ldots\ldots(A)$$

$$2mf_c \pm 2rf_s \quad \ldots\ldots(B)$$

$$\text{and } 2mf_c \pm (2r+1)f_s \quad \ldots\ldots(C)$$

The output in these classes can be obtained by writing (3) as

$$\tfrac{1}{2}\sum [d_{n1} + (-1)^n d_{n4} + (-1)^n d_{n2} + d_{n3}] [(a+b)^n - (a-b)^n] + \tfrac{1}{2}\sum [d_{n1} + (-1)^n d_{n4} - (-1)^n d_{n2} - d_{n3}] [(a+b)^n + (a-b)^n] \quad \ldots\ldots(5)$$

From this it can be seen that the output in class (A) is

$$\tfrac{1}{2}\sum (d_{n1} - d_{n4} + d_{n2} - d_{n3}) [(a+b)^n + (a-b)^n], \; n \text{ odd} \quad \ldots\ldots(6)$$

and this can be balanced out by making

$$d_{n1} + d_{n2} = d_{n3} + d_{n4} \quad \ldots\ldots(7a)$$

by means of potentiometer RV1, or by making

$$d_{n1} - d_{n4} = d_{n3} - d_{n2} \quad \ldots\ldots(7b)$$

by means of potentiometer RV4.

The output in class (B) is

$$\tfrac{1}{2}\sum (d_{n1} + d_{n4} - d_{n2} - d_{n3}) [(a+b)^n + (a-b)^n], \; n \text{ even} \quad \ldots\ldots(8)$$

and this can be balanced out by making

$$d_{n1} + d_{n4} = d_{n3} + d_{n2} \quad \ldots\ldots(9a)$$

by means of potentiometer RV3, or by making

$$d_{n1} - d_{n2} = d_{n3} - d_{n4} \quad \ldots\ldots(9b)$$

by means of potentiometer RV2.

The output in class (C) is

$$\tfrac{1}{2}\sum (d_{n1} - d_{n4} - d_{n2} + d_{n3}) [(a+b)^n - (a-b)^n], \; n \text{ odd} \quad \ldots\ldots(10)$$

and this can be balanced out by making

$$d_{n1} - d_{n4} = d_{n2} - d_{n3} \quad \ldots\ldots(11a)$$

by means of potentiometer RV4, or by making

$$d_{n1} - d_{n2} = d_{n4} - d_{n3} \quad \ldots\ldots(11b)$$

by means of potentiometer RV2.

10. Appendix 3: The Influence of Signal Voltage on Unbalance—An Example

Consider a shunt (or "Cowan") modulator, as shown in Fig. 1, which has a square-wave carrier voltage of amplitude $2V_c$ with zero source impedance, and a circuit resistance R which is very large compared with the forward resistance, $r_f(v)$, of the rectifiers, which is considered as a function of voltage. The signal e.m.f. is $E = E_0 \cos 2\pi f_s t$. Consider only the forward half-cycle of carrier; we shall assume that there is no conduction on the backward half-cycle.

The signal voltage developed across each rectifier is

$$v_s = E\, r_f(v)/4R \quad \ldots\ldots(1)$$

and we assume $E < V_c$ so that switching is not affected by the signal.

Rectifiers 1 and 4 have resistance $r_f(V_c - v_s)$ and 2 and 3 have resistance $r_f(V_c + v_s)$, if all four rectifiers are identical in the absence of signal. But, for generality, assume that the rectifiers are initially unbalanced by having their characteristics displaced in opposite directions along the voltage axis, so that 1 and 4 have resistance $r_f(V_c - v_0 - v_s)$ and 2 and 3 have $r_f(V_c + v_0 + v_s)$.

It is easily shown that the carrier-leak voltage is

$$V_{cL} \simeq V_c \left[\frac{r_f(V_c - v_0 - v_s) - r_f(V_c + v_0 + v_s)}{r_f(V_c)} \right] \quad \ldots\ldots(2)$$

on the assumption of a relatively small unbalance.

Now the resistance/voltage characteristic of r_f can be represented approximately and arbitrarily by a law which gives a falling curve thus

$$r_f(v) = r_0 + \alpha(2V_c - v)^2 + \beta(2V_c - v)^3 \quad \ldots(3)$$

On substituting in (2) we obtain

$$V_{CL} = [V_c/r_f(V_c)]\,[4\alpha V_c(v_0 + v_s) + 2\beta\{(v_0^3 + 3V_c^2 v_0) + (v_s^3 + 3V_c^2 v_s + 3v_0^2 v_s) + 3v_0 v_s^2\}] \quad \ldots\ldots(4)$$

This is the value of the leak at any instant during the forward half-cycle of carrier. To obtain the full time-function of the leak we must introduce the switching function, i.e., if f_c is the fundamental frequency of the carrier,

$$V_{CL}(t) = V_{CL}[\tfrac{1}{2} + \tfrac{1}{2}\,\mathrm{sign}\,(\cos 2\pi f_c t)] \quad \ldots(5)$$

and it is immediately clear that since v_s is alternating, the only contribution to leak at the carrier frequency and its harmonics comes from the "d.c." component of (4), i.e., on putting $E = E_0 \cos 2\pi f_s t$ and using equation (1), the carrier-leak is

$$V_{CL}(t) = [V_c/r_f(V_c)]\,[\tfrac{1}{2} + \tfrac{1}{2}\,\mathrm{sign}\,(\cos 2\pi f_c t)] \times \left\{4\alpha V_c v_0 + 2\beta(v_0^3 + 3V_c^2 v_0) + 3\beta v_0 \frac{E_0^2[r_f(V_c)]^2}{16R^2}\right\} \quad \ldots\ldots(6)$$

If the initial unbalance is due to rectifiers 1 and 4 having a coefficient β_1 in equation (3) and 2 and 3 having a different coefficient β_2, instead of being due to differences in contact potential, then the carrier-leak is easily shown to be

$$V_{CL}(t) = [V_c/r_f(V_c)]\,[\tfrac{1}{2} + \tfrac{1}{2}\,\mathrm{sign}\,(\cos 2\pi f_c t)] \times (\beta_1 - \beta_2)\left\{V_c^3 + \frac{3}{2}V_c E_0^2\,[r_f(V_c)]^2/16R^2\right\} \quad \ldots(7)$$

which corresponds very closely to equation (6).

From this work it is clear that

(*a*) the leak is zero, irrespective of the value of E_0, if $v_0 = 0$ and $\beta_1 = \beta_2$, i.e. if the initial balance is perfect.

(*b*) the effect of the signal voltage is in any case quite small—arising only from the cube term in (3) — since $E_0 < V_c$ and $r_f(V_c) \ll R$.

It appears, too, that these conclusions are unaffected if the carrier is sinusoidal instead of square-wave.

GRADUATESHIP EXAMINATION—NOVEMBER 1954

FINAL PASS LIST

This list contains the results of the remaining oversea candidates not included in the lists published in the February and March issues of the *Journal*. A total of 553 candidates entered for the Examination

The following candidates have completed the requirements of the Graduateship Examination and are eligible for transfer or election to graduateship or higher grade of membership

DUTTA, Asim Kumar. (S) *Barrackpore.*
GOUDAS, Nicolaos. (S) *Alexandropolis.*
HAIRETAKIS, Emmanuel. (S) *Piraeus.*
MATHIOUDIS, Miltiades. (S) *Athens.*
SERVETAS, Evangelos. (S) *Aegaleo.*
VASSILIOU, Athena. (S) *Athens.*

The following candidates passed the parts indicated against their names

BALASUBRAMANIAN, Venkatram. (S) *Delhi.* (II).
CARLIS, Georges. (S) *Athens.* (IIIa).
CHATTERJEE, Bhabatosh. *Calcutta.* (I).
CHOPRA, Jamak Kumar. (S) *Delhi.* (IIIb).
CHRISTODOULOU, Christos. (S) *Athens.* (II).
DIMAS, Dimitrios. (S) *Athens.* (II, IIIa).
FATSIS, Nicholas. (S) *Piraeus.* (II, IIIa).
GEORGIOU, Gregory. (S) *Athens.* (II).
HARCHARAN SINGH. (S) *Mhow.* (IIIb).
KANWAR, Randhir Singh. (S) *Delhi.* (II).
NEGREPONTIS, Eleutherios. (S) *Athens.* (II, IIIa).
PANDAZIS, Georgios. (S) *Athens.* (II).
PAPADOPOULOS, Emmanuel. (S) *Athens.* (II, IIIa).
PITSINIGOS, Savas. (S) *Athens.* (II).
POLITIS, Athanasius. (S) *Athens.* (II).
POLYMENEAS, George. (S) *Piraeus.* (IIIa).
ROMANIDIS, Andrew. (S) *Athens.* (IIIa, IIIb).
SODHI, Pashori Lal. (S) *Simla.* (II).
SPYROPOULOS, Nicholas. (S) *Athens.* (II).
ZAFIROPOULOS, Peter. (S) *Athens.* (IIIb).

(S) denotes a Registered Student

APPLICANTS FOR MEMBERSHIP

New proposals were considered by the Membership Committee at a meeting held on March 31st, 1955, as follows: 23 proposals for direct election to Graduateship or higher grade of membership and 26 proposals for transfer to Graduateship or higher grade of membership. In addition, 46 applications for Studentship registration were considered. This list also contains the names of six applicants who have subsequently agreed to accept lower grades than those for which they originally applied.

The following are the names of those who have been properly proposed and appear qualified. In accordance with a resolution of Council and in the absence of any objections being lodged, these elections will be confirmed 14 days from the date of the circulation of this list. Any objections received will be submitted to the next meeting of the Council with whom the final decision rests.

Direct Election to Member

KNOX, Commander George Frederick Edmund, R.A.N. *London, W.C.2.*

Transfer from Associate Member to Member

THWAITES, George Percy, B.Sc. *Petts Wood, Kent.*

Direct Election to Associate Member

BENNETT, Sqdn. Ldr. Herbert Ernest, R.A.F. *Singapore.*
BRENNAND, Robert. *Chessington, Surrey.*
GRAY, Capt. John Edward, R. Sigs. *Watchfield, Wiltshire.*
RADFORD, Norman Arthur William, B.Sc.(Eng.). *Sarnia, Natal South Africa.*
RICHARDSON, Arnold, B.Sc. *Shrivenham, Wiltshire.*
TOOMBS, Capt. Jack, R.E.M.E.. *Arborfield, Berkshire.*
WHITE, Flt. Lt. Clarence Bent, R.A.F. *Felixstowe.*

Transfer from Associate to Associate Member

BUNNER, Flt. Lt. Alfred John, R.A.F. *Chippenham.*
FORD, Edmund Alfred. *Ashford, Middlesex.*
MAINE, Arthur Edward. *St. Albans.*
MURRAY, John Walter. *Lagos.*
SHAPLAND, Albert John. *Swansea.*
WHITWORTH, Francis William. *Hatch End, Middlesex.*

Transfer from Graduate to Associate Member

COTTRELL, John Gilmour. *Bognor Regis.*
KIRYLUK, Wlodzimierz. *Bromley, Kent.*
REID, John Michael. *New Malden.*
WILKINSON, William Dinsdale. *Baltimore, U.S.A.*

Transfer from Student to Associate Member

STOKLE, Flt. Lt. Norman, R.A.F. *Manchester.*

Direct Election to Companion

BROWNE, Rupert Pollard, O.B.E., B.Sc. *Twickenham.*

Direct Election to Associate

CLEGG, William Arthur. *Farnborough, Hants.*
DALZELL, Thomas Derek. *Wilmslow.*
EVANS, Ronald Harold. *London, N.22.*
FOREMAN, Jack. *London, W.9.*
KEEN, George Frederick Sidney. *Nairobi.**
MORRIS, Lionel Alfred Dodsworth. *Pontypridd.*
PRINGLE, Arthur Graham. *Welgedacht, South Africa.*
TEMPLE, John Richard. *Wembley.*
VALE, Gordon Thomas. *London, S.E.2.*

Transfer from Student to Associate

BRADLEY, Wilfred. *Welwyn Garden City.*
COLLINGBOURNE, William Edwin. *Loughton.*
MATHEWS, Vincent Victor. *Singapore.*

Direct Election to Graduate

ARUNACHALAM, M.P., B.A.(Hons.). *Madras.*
BALL, Douglas Cedric, B.Sc. *Wellington, New Zealand.*
COLLINS, John Gilbert. *Banstead.*
DAY, Lieut. Geoffrey, R.N. *Amesbury.*
PAGE, Roger Francis Scott. *Maidenhead.*

Transfer from Student to Graduate

BANNOCK, Keith. *Hounslow West, Middlesex.*
DAVIES, Mervyn William. *Johore, Malaya.*
FEAR, Peter William. *Gravesend.*
HARKNETT, Maurice Richard. *Portsmouth.*
NKELE, Aluma. *Bende, Nigeria.*
RHODES, Alan Temple. *Havant.*

* Reinstatement

STUDENTSHIP REGISTRATIONS

ACKROYD, Sydney. *Henlow.*
BELCHER, John Charles. *Whitley Bay.*
BELFORD, Bertram Colin. *Coventry.*
BENNETT, Arthur John. *Brentford.*
BLEACH, Frederick Roy. *Helston.*
BLOOMFIELD, Oldman. *London, E.1.*
BRAITHWAITE, Clive. *Hull.*
BRIJLALL, Arora. *Ambala City.*
BROUGHTON, Flg. Off. George Herbert, B.Sc., R.A.F. *West Hartlepool.*
CLARKE, Arthur Philip Blake. *St. Davids.*
CLARKE, David Kelvin Jenner. *Sutton.*
CLEMENTS, John Henry. *Bristol.*
CORKETT, Raymond John. *Luqa, Malta.*
DE RUYTER, Albertus Hermanus Maria. *Eindhoven.*
DOYLE, Denis. *Limerick.*
GUHA ROY, Jayanta K. *Bangalore.*
GUILDFORD, Leslie Henry. *Brighton.*
GUPTA, Sudhaker, B.E. *Bihar.*
HADFIELD, William Norman Walter. *Kingsbridge.*
HALE, Peter Stacey. *Grimsby.*
HALPIN, Robert Joseph. *Ilford.*
HARDIE, Kenneth. *Glasgow.*
HIND, Derrick Duncan. *Burlington, Ontario.*
HOLMES, John. *Zomba, Nyasaland.*
JACKSON, Eric. *London, E.11.*
JACKSON, Michael Clifford. *Bristol.*
JAEGER, Eric. *Farnborough, Hants.*
KLOSOSZCZYK, Leopold. *Wollongong, New South Wales.*
KRISHNA VIR SINGH, B.Sc.(Hons.). *Bangalore.*
LARGE, Douglas Blake. *Great Yarmouth.*
LAWSON, Oswald Edward. *Polruan.*
LOCK, Robin Ian. *Helston.*
MARSHALL, Anthony Robert John. *Glasgow.*
NG WAI CHUNG. *Hong Kong.*
OWEN-JONES, Edward. *London, S.W.4.*
PODLASKI, Jan. *Manchester.*
RAJAPAKSE, Chandrapala, B.Sc. *Colombo.*
RAMAKRISHNAN, Krishna Sivarama, B.Sc., *Hyderabad.*
SHARMA, Rishi Prakash. *Rewa, India.*
SIRCAR, Pratap Kumar. *London, N.W.6.*
STEYN, Marthimus Jacobus. *Goodwood, South Africa.*
SUITER, John Ross, B.Sc. *London, W.3.*
TAYLOR, Robert Thomas. *Basingstoke.*
TWISS, Leonard Arthur. *London, N.22.*
VIRINDER SINGH. *Bangalore.*
WILLIAMS, Frederick. *Liverpool.*

PHOTO-ELECTRIC SMOKE DETECTION AND TEST TECHNIQUES USING BARRIER-LAYER CELLS*

by

R. W. J. Cockram†

A paper presented during the Industrial Electronics Convention held in Oxford in July 1954.

SUMMARY

Two automatic smoke detection equipments for use in aircraft, which depend on obscuration and reflection respectively, are described, and the techniques evolved for testing are given. The methods advanced are developed from a common basis of arranging for calibration of a photo-cell either for each test or from a sub-standard illumination.

1. Introduction

The combustion of most materials is accompanied by the generation of smoke which often occurs before actual ignition takes place. For this reason a means of automatically detecting the presence of smoke can form an important part of fire protection systems.

The Smoke Detector described here was originally developed for detecting smoke generated in baggage and freight compartments of aircraft, and has since found application on Marine and Industrial Applications. The operation principle involves the use of a photo-electric cell of the barrier-layer type, so arranged that the presence of smoke in a beam of light changes the measured illumination.

During the development of the Detector attention was focused on to various calibrating and measuring problems associated with such equipment.

The test techniques described here are based on the use of the simple barrier-layer cell to measure light flux values received from a beam of light by direct illumination, by light scatter from suspended particles in the beam or by light reflection from plane surfaces. In each case the substantially linear response characteristic of the cell is employed for comparison purposes and methods are described to ensure that the margin of error introduced is restricted to practical limits.

With developments in the barrier-layer cell and its resultant stability over substantially long time intervals, the test techniques described here are made possible and considerable advantages are available to many fields of Industry by the use of these cells on a Laboratory and Test House substandard basis. By these methods test results do not depend for accuracy on such variables as "standard" candles, the differing visual powers of different observers, the dependence of final results upon linear measurements related by the inverse square law, etc. All observations are taken by direct reading indicating electrical instruments wherein any errors may be evaluated and compensation made to give reasonable consistency of results.

Each of the techniques are basically of a very simple nature but investigation shows a wide field of application in each case.

2. Automatic Smoke Detection Equipment

Smoke detector designs fall into two main groups depending upon the manner in which a beam of light is arranged with respect to a pair of photo-cells which are arranged in a balanced bridge network with a sensitive current-measuring device. In each case one of the photo-cells, the Detector Cell, is arranged to respond to any variation in light intensity due to the presence of smoke whilst the Balancer Cell is substantially unaffected.

2.1. *Obscuration-Type Detector*

In this type of detector the beam of light is focused to fall directly upon the detector cell face through a long tube and a secondary beam from the projector lamp illuminates the balancer cell. Upon smoke entering the tube the intensity of illumination on the detector cell is reduced and the balanced condition of the bridge is disturbed so that the sensitive relay initiates the alarm. (*See Fig.* 1.)

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When smoke or other polluting media enter a beam of light, the reduction in transmitted light flux is not so much due to losses caused by absorption of the light flux as due to the deflection of the flux away from its otherwise normal path. By designing a detector, therefore, to operate on the flux reflected by the smoke particles, it is possible to reduce the circulating current required in the network to very low values, obtain a high standard of stability from the device and, at the same time, produce a highly sensitive detector which immediately responds to any serious pollution of the air traversed by the light beam.

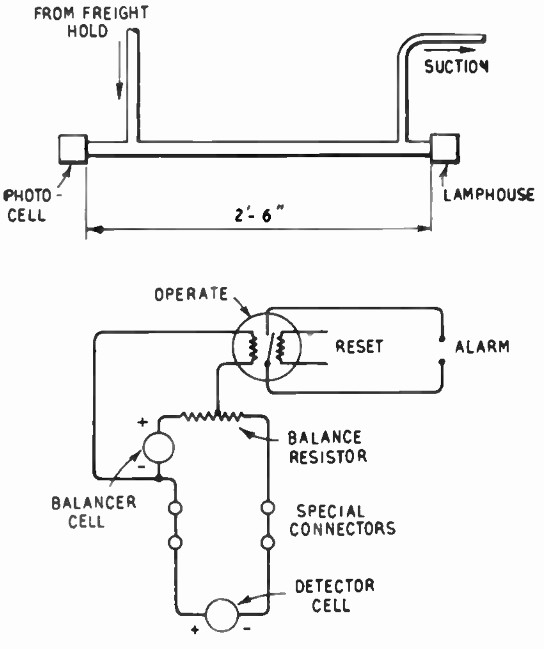

Fig. 1.—Obscuration-Type Smoke Detector, and simplified circuit diagram.

2.2. *The Reflection-Type Smoke Detector*

The smoke detector shown in Fig. 2 operates on the reflection principle. The unit employs a short light beam 6 in. in length and if the beam is terminated at a light-absorbing screen of high efficiency the circulating current through the photo-cell is about 8 microamperes. A schematic of the detector arrangements has been

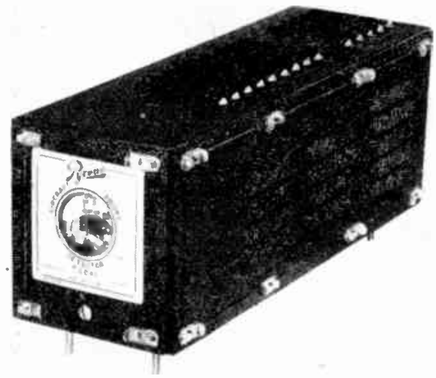

Fig. 2.—External view of Reflection-Type Smoke Detector.

given elsewhere[3] in a discussion contribution by the present author.

The detector comprises a rectangular metal box which is divided into three compartments. That at the front contains the sensitive relay which has an operating current of 2 microamperes only, and uses contacts formed from permanent magnets so that the low torque value is only used to move the needle against the gravity-control weights and the resistance of the damping mechanism. The rear compartment contains a 75-watt 45-volt projector lamp which is run at reduced voltage by series connection with a 2·5-ohm resistor, so that the lamp voltage approximates to 24 volts d.c. for a supply voltage of 29 volts d.c. The central compartment houses the photo-electric cells, the light absorbing screen, test lamp, signalling relay, and is fitted with light-tight louvres. A 28 dioptre bi-convex lens gives a substantially parallel beam of light from the projector lamp, and this passes through the detection compartment in line with the sensitive face of the detector cell. A small window located beneath the lens allows a secondary beam of light to illuminate the balancer cell which is fitted with a variable position shutter adjustable from the front of the detector. With the projector lamp on and all covers in position it is therefore possible to so adjust the position of the shutter of the balancer cell that this produces sufficient output to equate to the stray-light output of the detector cell. Upon smoke or other polluting media entering the light beam the output of the detector cell is increased by the additional intensity of its illumination, but the output of the balancer cell will remain substantially constant. The out-of-balance current can then only pass through the operating coil of the sensitive relay, and move the needle to the contact-making position. In this way the alarm is raised via the power relay which, for convenience, is housed in the detection compartment.

For test facilities, a small lamp is positioned in the detection compartment which, upon operation, is energized by the volt drop across the projector lamp resistor, thereby simulating the presence of smoke in the detector. By using the lamp resistor in this manner, the test signal can be raised if the cells are in working order, and the projector lamp circuit is complete.

As a means of temporarily increasing the sensitivity of the detector during balancing operations, a diverting switch is fitted whereby the projector-lamp resistor is cut out, thereby energizing the lamp at supply voltage. The increased light flux obtained is sufficient to ensure that the detector may be balanced even if the supply falls to 18 volts and that stability of balance is maintained over an 18-29 volts d.c. voltage range.

3. Smoke Detector Calibration Technique

Having developed a device which is ready for service trials, the next problem is usually one of determining that the further models required will have the same sensitivity characteristics as that tested under laboratory conditions. Furthermore, such a device must be checked from time to time to determine any loss in operating efficiency. The obvious problem here was to evolve some unit of calibration against which detectors could be set and which unit was readily reproducible without recourse to intricate apparatus.

To meet these requirements it was noted that the obscuration effect of smoke could be measured by its effect on the electrical output of the photo-cell receiving a beam of light flux. Since, within the limits of practicability, the electrical output of the photo-cell follows a straight line law characteristic, a unit of smoke density was evolved dependent upon the percentage obscuration in microamperes as exhibited by a photo-cell receiving light flux under given conditions.

The unit of smoke density may therefore be defined as that density which reduces the electrical output of a "standard" photo-cell with an external impedance of 290 ohms, from 100 to 99 microamperes when receiving light of a beam length of 4 feet.

Although this relationship does not hold good for very high intensities of illumination, the law gives a reasonably equally divided scale between 0 and 100 microamperes for a cell of surface area 1·0 in^2, providing the external resistance does not exceed 300 ohms.

3.1. *Sensitivity Test Equipment*

The test equipment shown in Figs. 3 and 4 comprises a substantially airtight wooden case with provision inside for the support of the smoke detector test model. An external lamphouse with lens and projector lamp is arranged to throw a beam of light across the lower portion of interior where it is reflected by a mirror so that it twice traverses the cabinet before falling on to the sensitive face of the "standard" cell which is protected from stray light by means of a hood.

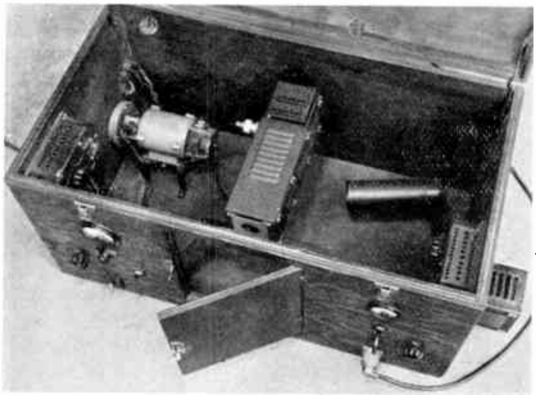

Fig. 3.—Sensitivity Test Equipment.

Means are provided for energizing the detector over a wide voltage range by "coarse" and "fine" rheostats. The voltage supply to the calibrating projector lamp may also be controlled.

The smoke detector may be operated for test, and reset conditions from the control panel and a suction fan is used to draw smoke and other polluting media into the cabinet, and circulate this by internal swirling action.

The test procedure entails positioning the detector on its supports in the test cabinet, connecting it to the supply, and adjusting the supply voltage. The calibrating projector lamp is then energized, and its voltage adjusted to give 100 microamperes output from the photocell with clean air in the test cabinet.

The suction fan is then started and smoke carefully introduced at the collecting hood so that the microammeter slowly reduces its deflection. Upon the detector responding to the smoke in the test cabinet, as is indicated by the warning lamp, the microammeter reading is noted. The difference between the original and this reading of the microammeter is a measure of the sensitivity of the detector, and is usually expressed as x per cent. smoke obscuration.

Tests show that the detector is, as would be expected, subject to variation in sensitivity with voltage variation. A detector set to operate at 10 per cent. obscuration at 29 volts requires approximately 33 per cent. obscuration before it operates when energized at 21 volts.

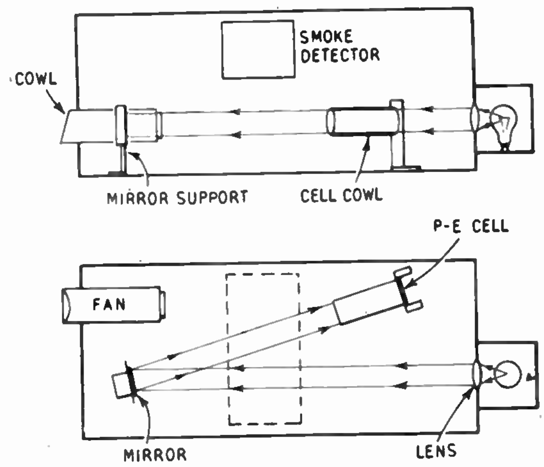

Fig. 4.—Sectional and plan views of the Sensitivity Test Equipment.

This cabinet is used regularly on test and inspection work under production conditions. All detectors are recorded for smoke density operation before despatch, and the value adjusted to be between 8 and 15 per cent. smoke obscuration. It is of interest to note that a detector despatched calibrated to operate at 12 per cent. smoke obscuration, was returned for inspection after 1,000 flying hours. Upon checking the obscuration value required, it was found to have risen to 30 per cent. Subsequent cleaning of the apparently clean lamp, lens and cell windows, however, restored the detector to its original sensitivity of 12 per cent. It was then possible to advise the user that the recommended cleaning instructions had not been carried out.

3.1.1. Comparison of effects of dusts

Whilst the detector was designed to operate on smoke, it was of course expected that there must be some degree of sensitivity to dust-laden atmospheres. In order to determine the characteristics of dusts as compared with smoke, a series of tests were instituted using mixtures of finely powdered white china clay and black charcoal.

It was found, as shown by the graph of Fig. 5, that even with white dust the sensitivity of the detector was less than with smoke. In addition by plotting the results obtained against the variation of dust mixtures at both 29 and 21 volts supply two, curves were obtained. Consideration of these curves show :—

(1) The repeatability characteristic of the test technique.

(2) The reflection of the photo-cell output characteristic.

(3) The degree of reliability which may be obtained from the technique described.

3.1.2. Fog tests

In order to simulate fog under laboratory conditions the intense white mist obtained when liquified carbon dioxide gas is discharged was used.

With the detector energized, a CO_2 fire extinguisher was carefully discharged into the test cabinet. Although the obscuration value was raised to 100 per cent. no signal was registered. This can only be attributed to the elevated temperature of the detection compartment due to the heating effect of the detector projector lamp.

4. Photo-Cell Calibration Technique

The calibration of photo-cells is a comparatively simple matter providing one can determine the intensity of illumination at which to measure the cell output.

The method employed in this technique is to arrange a light-tight cabinet, so that the illumination intensity of a ground glass screen can be varied at will by means of a simple lamp and rheostat. An example of this design is shown in Fig. 6.

In operation the diffusion screen illumination is adjusted to the test illumination using a specially calibrated photo-cell with a microammeter of the same internal resistance as the calibration instrument. The standard cell is then replaced by the test cell and the deflection recorded.

Standardized photo-cells are found to experience little change in calibration value over several years of time.

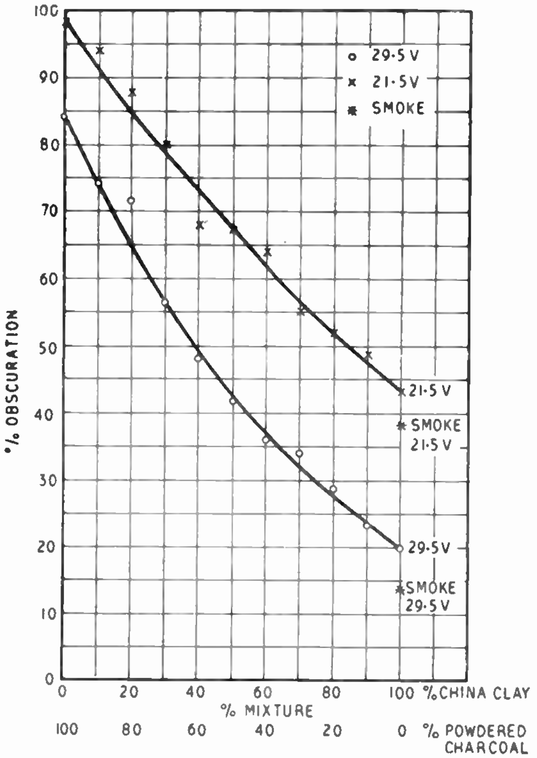

Fig. 5.—Graph of dust tests on reflection-type smoke detector.

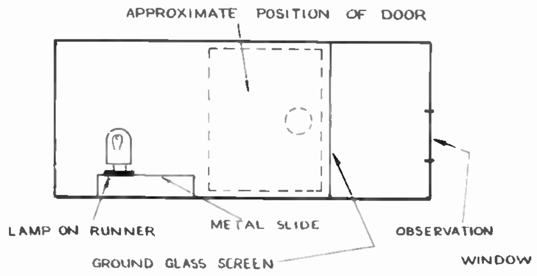

Fig. 6.—Photo-cell Calibration Cabinet.

4.1. *Sensitive Instrument Calibration*

When calibrating sensitive instruments of the microampere range a most useful source of supply is a photo-cell illuminated at controllable intensity. For such work the cell test cabinet is most advantageous.

The sensitive relay with magnetic contacts used on the smoke detector has a full scale deflection of 2 microamperes only. Testing this instrument can present problems which are avoided by connecting the test instrument in series with a substandard microammeter and a photo-cell. Upon illuminating the cell to low intensity it is found that the current in the relay slowly rises and "dips" temporarily at the point of contact-making. The peak current registered by the substandard meter is therefore a convenient method of determining the operating current of the test instrument.

5. Calibration of Light-Absorbing Screens

An important component of the smoke detector is the light-absorbing screen. If this is not of high efficiency, then the sensitivity and stability of the detector will be impaired, as a high value of light scatter in the detection compartment entails a high value of circulating current between the two photo-cells.

The test equipment shown in Fig. 7 was developed as a means of comparing various types of light-absorbing screens. The apparatus, arranges for a photo-electric cell to measure the intensity of a beam of light flux specularly reflected from an inclined plane bearing the test sample, and an additional photo-cell is positioned directly above the sample to measure also the intensity of the scattered light flux in accordance with standard methods of investigating the quality of finish.

Early tests showed that this simple apparatus was capable of detecting differences in the light-absorbing qualities of different materials as indicated by the different photo-cell readings at the same intensity of illumination. This justified the original reason for making up the experimental equipment in that it provided a ready means of determining the relative merits of the various materials available from which the smoke detector light-absorbing screens could be made.

From these tests it was determined that stiff pile-carpeting materials of the type used for

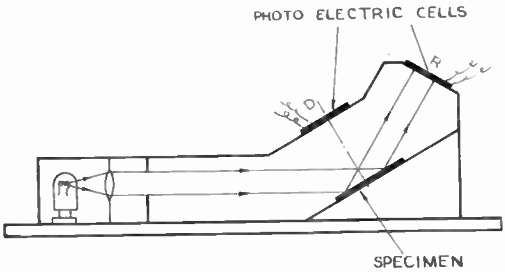

Fig. 7.—Light Scatter Comparator.

the floor covering of modern automobiles was a most efficient form of light-screen material in a form of ready availability.

6. Investigation of Surface Characteristics

Some experimental work using flat anodized aluminium sheets instead of light-absorbing materials showed that sheets of similar appearance recorded widely different dispersed reflections. Whereas those showing high dispersion values gave "woolly" reflections of bright objects, those of low dispersion values gave "sharp" reflections of bright objects. A quantitative means of discriminating between these characteristics is therefore available by determining numerically the dispersed light values obtained at constant illumination.

6.1. *Investigation of Surface Finish*

In this case both dispersed and specular reflections are measured by the same photo-cell and ammeter, the photo-cell being changed from the dispersed to the specular reflection position for each intensity of illumination.

Upon conducting this test using a sheet of matt black-paint finished aluminium it was found that for a wide range of intensity of illumination the ratio of dispersed to reflected light ($D/100R=K$) is equal to a constant throughout the test. Upon changing the test sample to another sheet of material of apparently the same appearance, the K value was found to be considerably altered.

Records of the original values obtained from this test are shown in Table 1, wherein D shows the microammeter readings for dispersed light, R shows readings for specular reflected light and $K=D/100R$.

Table 1

Sample I			Sample II		
D	R	$K=D/100R$	D	R	$K=D/100R$
12.5	16	78.1	11.6	18.5	62.6
10.8	13.7	78.7	10.7	17.2	62.5
9.7	12.3	78.8	9.9	16.0	61.8
8.5	11.0	77.2	9.1	14.8	61.5
7.4	9.4	78.7	8.6	14.0	61.5
6.0	7.6	79	7.8	12.5	62.4
4.6	6.0	76.6	7.0	11.3	61.8
3.7	4.8	77.2	6.7	10.7	62.6
2.9	3.7	78.4	5.9	9.3	63.4
2.0	2.6	76.9	5.4	8.6	62.8
1.4	1.8	77.8	5.0	8.0	62.5
Average $K=78$			Average $K=62$		

The matt black-painted aluminium was then replaced by a sheet of writing notepaper to obtain another value of K. Upon replacing the test sample by another portion from the same sheet the K value remained unchanged.

The readings taken in this case are shown in Table 2.

Table 2

Sample I			Sample II		
D	R	K	D	R	K
46.2	21.0	218	47.1	21.6	213
39.0	17.6	222	34.3	15.6	220
28.3	12.7	223	23.6	10.7	221
17.3	7.8	223	14.2	6.5	218
Average $K=221$			Average $K=218$		

In all these tests it was observed that the K value remained constant in the linear portion of the response curve of the photo-cell. Inaccuracies were only introduced upon obtaining readings of less than 1 microampere and here the "reading" accuracy of the meter would account for the errors observed.

The technique established was therefore explored in other fields.

Two sample steel plates, Nos. 4 and 6, known to have been treated by dissimilar processes, yet both having the same appearance with respect to colour and surface finish were checked for *K* value as shown in Table 3. This table also includes results from testing sample 2 (of different appearance) and sample 7 which was of untreated steel. It will be noted that a wide difference exists between the readings of the apparently similar samples 4 and 6, whilst the obviously different sample (2) falls between these values. Sample 7, presenting a smooth polished surface of relatively high reflectivity, has a *K* value widely removed from the other non-reflective plates.

Further tests conducted using groups of treated plates wherein each group had been treated in precisely the same manner gave results shown in Table 4.

In each case it will be noted that at least one sample has a different *K* value to the remainder of the group. At present the precise reason for this difference is indeterminate as in each case the sample differing from the remainder of its group can only be identified using this method of test.

Whilst no clearly practical application is yet apparent for the *K* value discussed here, the technique may have possibilities in connection with the grading of apparently similar surface finishes.

It will be noted that this technique has advantages over more usual methods in that instead of obtaining values for total and specular reflection from the surface investigated, a direct comparison is made between measures of the diffused and specular reflection. Whilst therefore, the technique may not be of such a precise nature the *K* value described here is more directly linked with the surface characteristics on a more empirical basis.

7. Acknowledgments

Acknowledgments are due to The Pyrene Company Limited for permission to publish the information contained in this paper, and for the use of illustrations and diagrams contained therein. The author would also wish to express thanks to Dr. G. A. Veszi, of Megatron Limited for data received with respect to photo-cells and permission to publish the typical output curves included in this paper.

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Table 3

Sample 4 (Black)			Sample 6 (Black)			Sample 2 (Mid-Grey)		
D	*R*	*K*	*D*	*R*	*K*	*D*	*R*	*K*
1.3	1.3	104	19.4	11.8	165	37.2	22.2	147
2.1	2.1	100	11.5	6.9	167	23.4	15.9	147
4.6	4.6	100	14.3	8.5	168	15.4	10.4	146
6.8	6.75	102	9.0	5.5	163	9.5	6.5	146
10.5	10.4	102	4.9	2.8	175	6.7	4.6	146
17.0	16.8	102	2.8	1.6	175	2.8	2.0	140
Average $K=101$			Average $K=169$			Average $K=145$		

Sample 7 $D=0.5$ $R=32.6$ $K=1.39$

Table 4

GROUP I		GROUP II		GROUP III		GROUP IV		GROUP V	
Sample	*K*	Sample	*K*	Sample	*K*	Sample	*K*	Sample	*K*
P1	175	22	140	40	140	202	78	232	116
P2	177	23	130	41	123	203	82	233	116
P3	174	24	146	42	143	204	82	234	110
P4	160	—	—	—	—	—	—	—	—

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9. Appendix 1: Photo-electric Cell Circuits

All circuits described here assume that no extraneous light is allowed to affect the photo-cells.

9.1. *Simple Cell and Light Beam*

Where the cell is used for direct measurement of intensity of illumination, the circuit shown in Fig. 8 is employed.

The photo-cell output is controlled by:
(1) The surface area of the cell.
(2) The intensity of illumination.
(3) The impedance of the external circuit.

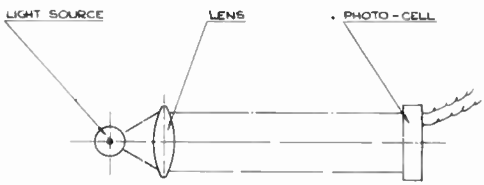


Fig. 8.—Simple cell and light beam.

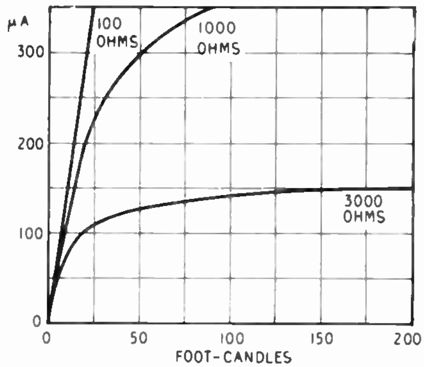
The current output of a photo-cell is directly proportional to the surface area, and is usually found to range between 0·5 and 0·65 microamperes per candle foot per square centimetre, providing the external resistance is maintained small compared with the internal resistance of the cell.

Since, however, the internal resistance varies inversely with both the cell area and the intensity of illumination, this relationship will not hold good for high intensity of illumination on large cells unless the external resistance is reduced to very low values.

Reference to Fig. 9 shows the comparative characteristics of two cells, one of 67 mm diameter, the other of 25 mm diameter with external resistance values of 100, 1,000 and 3,000 ohms.

It will be noted that the 25-mm diameter cell will give approximately a straight line response curve between 0 and 100 microamperes with an external resistance of 1,000 ohms, whilst with an external resistance of 100 ohms the straight line characteristic is maintained to above 350 microamperes.

(*a*)



(*b*)

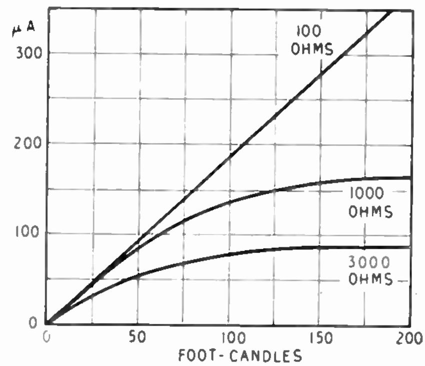


Fig. 9.—Output curves for (a) 67-mm and (b) 25-mm dia. Cells.

9.2. *Balanced Bridge Circuit for Smoke Obscuration*

In this circuit, shown in Fig. 1, two cells are employed. The detector cell is directly illuminated as in Fig. 8, and a balancer cell is also illuminated either from the same or a separate source.

For smoke detection applications the circuit has the following disadvantages:—

(1) Sensitivity is proportional to circulating current between the two cells.

(2) Deposits of dust on the detector cell can cause false warnings in service.
(3) A comparatively long tube of light flux is required to provide reasonable sensitivity.
(4) The equipment must usually be piped to exhaust ducting for efficient operation.

9.3. *Balanced Bridge Circuit for Smoke Reflection*

When smoke enters the path of a light beam, much of the obscuration effect is caused by light deflected from its normal path. It has been shown how this arrangement is employed to produce a smoke detector of more advanced design wherein :

(1) Sensitivity is independent of cell circulating current.
(2) Dust deposits on the cell face only serve to reduce the sensitivity of the unit.
(3) A short tube of light flux may be employed with an efficient light absorbing screen.
(4) No connection to exhaust ducting is required.
(5) False operation can be generated by open circuiting the balancer cell.

10. Appendix 2: Smoke Detector Sensitivity to other than Smoke Media

Mention has been made of the relative sensitivity of the reflected light form of smoke detector to other media than smoke. (Sections 3.1 and 3.2).

Further tests in this series revealed that whilst the equipment is comparatively insensitive to minute particles of water as under fog conditions, mists formed by apparently more persistent liquids, such as a solution of vegetable oil in a solvent, are quickly detected.

When the detector was enclosed in a metallic box fitted with 100-mesh gauze (10^4 holes per in.^{2}) and a vegetable oil solution sprayed into a further outer enclosure, the signal was raised within one minute of commencing the test. The spray was immediately stopped and balance conditions could not be restored to the cell circuit for a further 2 minutes. This condition was considered due to either the persistance of the spray or due to temporary condensation on the projection lens.

The former condition raised the possibility of other liquids causing reactions from the photo-cell circuit, so a number of chemical liquids were tested using for simplicity the diffusion reflection position (D) of the surface comparator (Fig. 7). Test specimens were introduced by means of a watch-glass placed immediately below the projection lens.

The projector lamp was set to give a dispersed reflection value of 3·6 microamperes and readings were obtained for various chemical fluids as shown in Table 5.

Table 5

Chemical	Photo-cell Current	Signal Current
Acetic Acid	3.6	0
Ammonium Sulphide	5.7	2.1
Ammonia	6.5	2.9
Hydrochloric Acid	6.2	2.6
Nitric Acid	5.0	1.4
Benzol Chloride	3.6	0

It is seen that acetic acid and benzol chloride give no increase in reflected light, whilst the presence of the remaining liquids may be detected by the reflected light method. In each of the examples used no fumes were visible under normal lighting conditions but, where a positive result was obtained, fumes were clearly seen on visually inspecting the light beam. For all tests the ambient temperature within the test cabinet was 21·5°C. for an external ambient of 21°C.

It should be pointed out that no claims can be made for the suitability of the detector for working with these liquids, many of which are of a highly corrosive nature with respect to selenium. Were it required to produce a detector to meet such working requirements, then considerable care in design would be necessary to ensure complete and hermetic sealing of the photo-cells so that there would be no possibility of these chemicals gaining access to the cell plates used.

11. Appendix 3: Contact Techniques

Contact engineering for power circuits is a highly complicated subject for which little assistance may be derived on a precise basis. Much of the design work carried out has, of necessity, to be based upon trial and error and whereas certain metallic combinations are recognized as most suitable for specific applications, special cases usually entail a complete restart involving many years of experimental work before definite conclusions can be reached.

With the exceedingly low power of the photo-cell circuit much of the work entailed is

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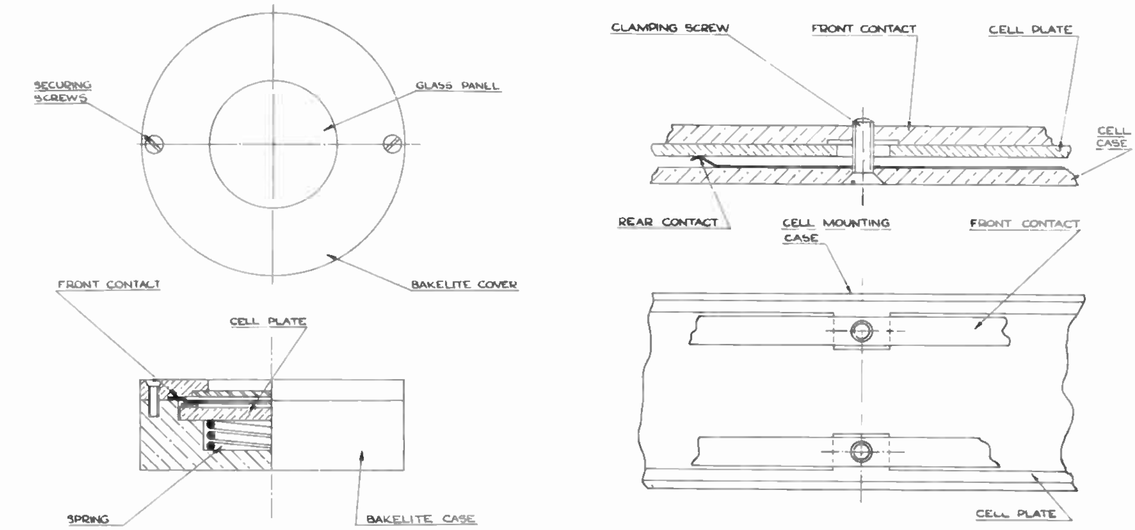

Fig. 10.—Cell contact design (a) with spring contact. (b) without spring contact.

directed to the maintenance of good contact surfaces under conditions of continuous "make." Here the problems of contact resistance heating effects, as experienced with power circuits is avoided but, where photo-cell circuits have to be broken and remade at will, the problem becomes more difficult than the power case, in that there is no real e.m.f. value available to break down the contact resistance film.

11.1. *Permanent Contacts*

Experience has shown that where permanent contacts in the cell are required to maintain efficiency over long time intervals, both the contact surface area and contact pressure should be as high as is practicable. High pressure small area combinations usually result in cutting through the sprayed metal contact on the cell face resulting in short-circuit conditions. An example of this design is shown in Fig. 10a, where it will be noted that the contact is formed by a metal ring having notches cut for resilience.

Figure 10b shows an improved form of contact incorporating no spring action, but using the cell casing as one portion of a clamp. The other clamp member is a solid flat strip of heavy section metal which is forced against the sprayed metal by means of the clamping screws. A slight recess close to each clamping screw ensures that the pressure point is removed from the edge of the cell.

11.2. *Switching Contacts*

Normal switching techniques find little successful application on photo-cell circuits. This would appear to be almost entirely due to insufficient contact making pressure on modern point to point or point to butt type contacts. Furthermore, and what seems to be more important, the degree of contact wiping available on modern switches appears to be wholly insufficient to ensure that the mating contacts are cleaned to the necessary degree during the making operation.

The ideal form of switching contact is probably one in which the two contacts faces are literally "rung" together as carried out, for example, in the assembly of the inspector's slip gauges.

It would appear that the development of a reliable photo-cell switch would provide further applications for photo-cells in that this would remove the present necessity of maintaining these on permanently closed circuits where efficiency and reliability are essential qualities required of the apparatus on which they are used.

NEW PARTICLE ACCELERATORS FOR RADIOTHERAPY

The principal methods used in radiotherapy for the treatment of cancer are by radioactive isotopes and high-voltage x-rays. The development of the latter technique using the well-known type of x-ray tube was continuous up to about the beginning of the last war but it became apparent that for various technical reasons, chiefly the difficulty of insulation, the limit had been reached with the design of orthodox equipment operating at 1 or 2 MeV.

As frequently happens in technical developments a fresh approach to the problem was required and this came about through the applications of techniques evolved for radar to the acceleration of particles, generally electrons, to extremely high energies for nuclear research. Work in this field has been carried out over the past years in a number of countries using different types of particle accelerators, and two different approaches were recently demonstrated at the Christie Hospital and Holt Radium Institute, Manchester, one of the leading centres for the treatment of cancer. These were a travelling wave linear accelerator and a betatron, producing 4 million volts and 20 million volts x-rays respectively, and developed by Metropolitan-Vickers Electrical Company Limited.

The use of both these high-energy accelerators as radiotherapeutic equipment provides a very useful and flexible means of investigating what is to a certain extent an experimental problem. Although the effects of x-rays generated by the normal type of apparatus are very well known, the increased energy available presents both advantages and fresh problems.

The advantages may be summarized as follows:—

(*a*) there is greater penetration of the radiation making it easier to deliver an adequate dose at a depth;

(*b*) the maximum dose is produced not on the surface as at 250 kV but some distance below, thus reducing undesirable skin effects. (For 4 million volts x-rays, the maximum is at about 1 cm below the surface, while for 20 million volts, it is 4 cm deep.);

(*c*) the amount of scattered radiation, and hence the dose received outside the useful beam, is greatly reduced;

(*d*) below 1 million volts, bones absorb x-rays much more than does normal tissue (which is why bones are revealed on x-ray pictures), whereas above 1 million volts bone and soft tissue absorb x-rays nearly equally, a very real advantage in radiotherapy.

One of the complications which arise is that the greater penetration achieved may in some circumstances lead to an increased exit dose being given to the skin at the point where the beam leaves the patient's body.

Although both the new equipments deal with the same problems, their routine use presents different difficulties, and for some time they will be employed in rather different ways. The linear accelerator operating at 4 MeV presents few additional problems in dose measurement and treatment planning. The advantage to be gained from its use can, therefore, be predicted with some degree of assurance, and the equipment has, in fact, been used for routine treatment for some time.

The betatron on the other hand is the only clinical adaptation of this machine in this country. Although some data are available from the use of betatrons both in Europe and America and a little experience of very high-energy radiation has been obtained in this country using a synchrotron, at very low output, a considerable amount of experimental work will be required before the betatron can be used for treatment, otherwise serious damage might result.

The mode of operation of the linear accelerator has already been dealt with in papers in the *Journal* describing some of its industrial applications.* Briefly, electrons are accelerated by an axial electric field associated with waves propagated inside a corrugated waveguide. The electrons "ride" on the wave in much the same way as a surf-rider may ride on a sea wave. Metallic disks (or irises) along the guide control the phase velocity of the wave and so arrange that the wave and the electrons accelerated by it travel at the same speed. By variation of the iris

*C. W. Miller. "Industrial radiography and the linear accelerator," *J.Brit.I.R.E.*, **14**, p. 361, August 1954; also, "Applications of high-energy electrons to the sterilization of pharmaceuticals and the irradiation of plastics," *loc. cit.*, p. 637, December 1954.

NEW PARTICLE ACCELERATORS FOR RADIOTHERAPY—(*contd.*)

and guide dimensions, wave velocity and accelerating field can be controlled, and the wave velocity can be increased progressively along the length of the guide towards light velocity and the electrons collected by the wave thus accelerated.

The guide is 1 m long and in addition to the axial accelerating field there is also a steady axial magnetic field for focusing, provided by a solenoidal winding surrounding the accelerator guide.

To obtain the high-electron energies required, high fields in the corrugated waveguide (of the order of 30 to 40 kV/cm) and hence very large radio power fluxes of the order of megawatts are necessary. Such considerations enforce pulse operation of the equipment, and, since the free-space wavelength may conveniently be 10 cm, operation becomes similar to that of pulsed radar equipment, a 2-MW magnetron being used.

The x-ray beam is defined by means of adjustable diaphragms in the x-ray head which provide for continuous adjustment of field size from 4×4 cm up to 25×30 cm. At the maximum repetition rate of 500 pulses per second having 2-μsec duration, an unfiltered x-ray beam output of 330 roentgens per minute at 1 m is obtained.

The mounting is arranged so that the machine can be moved to any angle required on a 120° arc lying in a vertical plane. The axis of rotation coincides with the treatment point 100 cm from the target, a convenient arrangement for setting up a 100-cm target-skin or target-tumour distance as required.

The 20-MeV betatron consists of a large alternating current magnet, whose central limb has specially shaped pole pieces between which is inserted a continuously evacuated glass toroid or "doughnut." The betatron magnet is excited from a 150-c/s tripling transformer fed from the three-phase mains supply. 60-keV electrons are injected into the toroid by means of a small electron gun, which is pulsed to this voltage for 4 μsec when the magnetic field between the poles is at an appropriate very small value and thus sends a stream of electrons circling round the doughnut. As the magnetic field increases sinusoidally the electrons circulate round the doughnut on an orbit of 19 cm radius, and for each revolution gain 80 electron-volts of energy by magnetic induction. The electrons reach their maximum energy in 1/600th of a second, during which period they travel nearly 300 miles. At this stage the orbit is expanded so that the electrons spiral outwards and hit a small platinum target attached to the end of the gun. The x-rays are produced in the forward direction in a relatively narrow cone and occur as 2-μsec pulses at a repetition rate of 150/sec.

In order to reduce the power consumption, a resonant circuit is employed in which the magnet is paralleled with a 1,600-kVA capacitor bank so that it is only necessary to supply the losses of the magnet, which amount to about 20 kW.

The magnet, which weighs 7½ tons, can be rotated through 120° so that the x-ray beam can be set at any direction from 15° above the horizontal down to 15° beyond the vertical. The framework in which the magnet is held can be raised or lowered through a distance of 7 ft. As in the case of the linear accelerator, radiation from the machine is applied to a patient through a beam-defining device or collimator, into which different treatment cones are fitted according to the field shape required. The maximum size of field that can be irradiated is 15 cm diameter at 1 m from the target, and the x-ray yield at 1 m from the target is 100 roentgens per minute.

The installation at the Christie Hospital is supplemented by a 300-kV x-ray equipment which is intended to serve as a control. This equipment is of the resonant transformer type in which a multi-section x-ray tube is mounted coaxially within the transformer. The Director of the Radiotherapy Department, Dr. Ralston Paterson, stressed that this lower-voltage equipment is not to be regarded as outmoded by the new accelerators since in many cases completely satisfactory results are obtained with lower energies. Although spectacular results should not be expected, higher-energy x-rays provide the possibility of treatment in more difficult cases. However, as already indicated, much careful research work will be required.

MULTI-ELECTRODE COUNTING TUBES*

by

K. Kandiah, M.A., B.Sc.† and D. W. Chambers†

A Paper presented during the Industrial Electronics Convention held in Oxford in July 1954

SUMMARY

The paper considers possible uses of hot and cold cathode decimal counters in systems other than straightforward counting. It deals with their uses for adding and subtracting, for counting a predetermined number, and for dividing by numbers less than ten. The principles of a pulse-amplitude analyser using trochotrons and dekatrons in a matrix system are outlined. References are also made to other possible uses where they are expected to be more reliable than more conventional methods.

1. Introduction

Multi-electrode counting tubes such as the Ericsson Telephones GC10B, the Mullard E1T, and the L.M. Ericsson RYG10, have made it possible to count in decimal figures in a manner which is considerably simpler than in those systems which use a multiplicity of thermionic or cold cathode tubes. It is now possible to use these new tubes in equipment for counting nuclear events and for various counting applications in industrial process control, with the advantage of direct indication of the number stored, high counting speeds, and simple zero resets. The reliability of these tubes has encouraged their use in further applications. It is the purpose of this paper to assess features of these tubes which make them suitable for some unusual applications.

The characteristics of these tubes have been described in detail in the literature.[1, 2, 3, 4, 5]

2. Counting in Decimal Figures

Most of the tubes available at present are designed to count in decimal figures and they are all suitable for connection in a cascade system providing a store of as many digits as desired. The complexity of such a complete unit depends largely on the type of pulses required to drive the tubes. These are single pulses, or pairs of pulses, having fairly wide permissible tolerances for the gas-filled tubes, and pulses of defined amplitude and shape for the E1T and RYG10-type tubes.

All tubes available at present in this country provide an output pulse which is of smaller energy content than that required to drive a succeeding counting tube. It is therefore necessary to provide a driving stage, using at least one thermionic valve, or cold cathode trigger tube between the stages of a cascade. A gas-filled counting tube has been described[6] in which the necessary coupling element in the form of a trigger tube is built into the envelope which contains the counting tube.

The Ericsson Telephones GC10B has the advantage of being completely symmetrical, and can therefore be driven in either direction. It is only necessary to reverse the connections to the transfer guides in order to reverse the direction in which the glow transfers. If, however, it is necessary to make the tube add or subtract when an input pulse is applied to one of two alternative points, it is necessary to have independent driving circuits which apply transfer pulses in the correct sequence. When a cascade of GC10B stages is required to add or subtract, the problem becomes more involved, and one method of providing this facility has already been described.[7] A disadvantage of this method is that ambiguity in reading the digits may sometimes be experienced, and it will then be necessary to determine the digit stored by reference to the position of the glow in the previous tube.

An advantage of the method[8] of driving a cascade of GC10B stages by utilizing a combination of a common pulse generator and gating tubes as coupling elements is that it readily lends itself to a reliable method of addition and subtraction. In its simplest form, the circuit of one stage of a cascade is as shown

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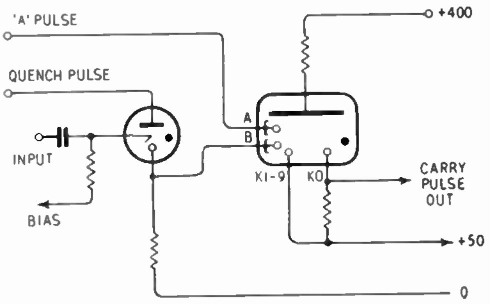

Fig. 1.—One stage of a cascade.

in Fig. 1. In this system, the discharge rests on a "B" cathode of the counting tube and the "A" and "K" cathodes are used as transfer guides. All the stages in the cascade are identical and "A" pulses and "quench" pulses from a common pulse generator are applied to all the stages of the cascade whenever an input pulse is received into the first stage. The waveforms in the first two stages in such a system for addition are shown in Fig. 2. Although pulses are applied to all the "A" transfer guides, the glow will advance by one digit in only those stages which have received a pulse into the trigger tube.

In principle, a cascade of stages using the circuit of Fig. 1 can be made to subtract if the common pulse generator delivers the pulses shown in Fig. 3 which also shows the appropriate waveforms in the first two stages when they count backwards from 10. It should be noted that in this method the first trigger tube receives a pulse which is delayed by about 200 μsec on the input pulse and that all "A" transfer guides receive a negative pulse from the common pulse generator during this delay.

The realization of a practical unit which will add or subtract, working on the principles outlined, will require some changes to the circuit shown in Fig. 1 and to the waveforms from the common pulse generators.

During the period of the negative pulse to all "A" transfer guides, the discharge in the counting tubes will move from the "B" guide to an adjacent "A." This will result in the lowering of the cathode potential of the trigger tube associated with each stage. The tube may therefore fire even in the absence of an input pulse to the trigger electrode and the unit will miscount. It is therefore necessary to modify the cathode circuit of the trigger tubes so that they will fire only when an input pulse is applied to the trigger electrode and not when the associated counting tube momentarily passes current to a guide other than the "B" guide.

Another difficulty arises from the delay between the positive-going back edge of the "A" pulse to a GC10B—which results in that

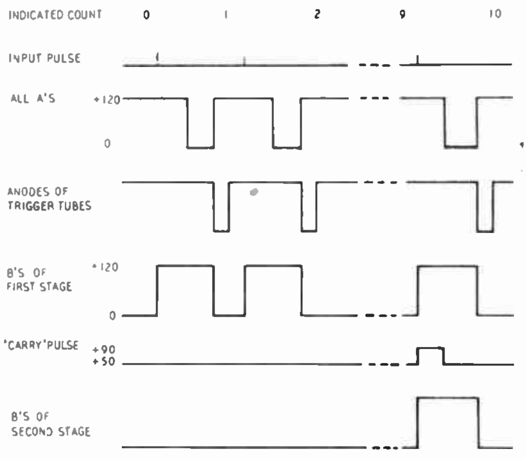

Fig. 2.—Waveforms for addition.

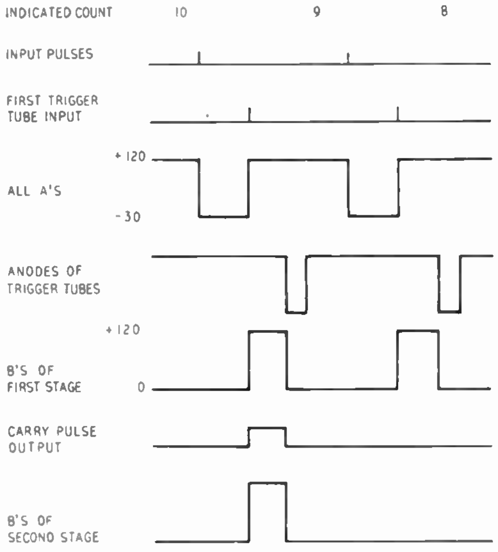

Fig. 3.—Waveforms for subtraction.

particular stage counting to 10, and the firing of the subsequent trigger tube. This delay is partly due to the trigger tube, and in the case of a fast trigger tube, mainly due to the rate of rise of the GC10B anode voltage and the rate of transfer of glow to the output guide Ko. It is possible to keep this delay down to about 10 μsec per stage of the cascade. In a 5-stage unit the delay may therefore be 40 μsec. The resulting waveforms on the 1st and 5th stages during addition and subtraction are shown in Fig. 4. In the case of addition the effect of the delay can be completely overcome by increasing the delay between the input pulse and the negative-going front edge of the "A" pulse. For subtraction, however, it is necessary to ensure that the back edge of the "A" pulse is returned to + 120V line only when the trigger tube of that particular stage has fired. If the trigger tube does not receive an input pulse, the "A" guides are returned to + 120V when the quench pulse is applied to all the trigger tubes.

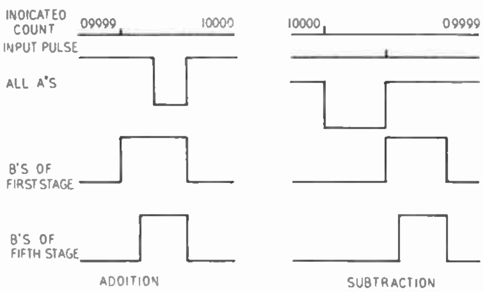

Fig. 4.—Effect of cumulative delay in five stages of a cascade.

The circuit diagram of a typical stage of a cascade capable of adding and subtracting, with the modifications necessitated by the above considerations, is given in Fig. 5. The waveforms for addition and subtraction are given in Fig. 6. Let us first consider the quiescent state of the typical stage shown in Fig. 5. The trigger tube V1 is not conducting and the anode voltage on this tube is at the recommended value for the tube. The cathode of V1 is held at earth potential by the rectifier MR1, there being a small current through R1 to − 20V. The GC10B is passing current to a guide "B" which is held at the potential of the cathode of V1 by rectifier MR2, the current through R2 being slightly greater than the current through the GC10B. Guides "A" and "K" of the GC10B are at + 60V.

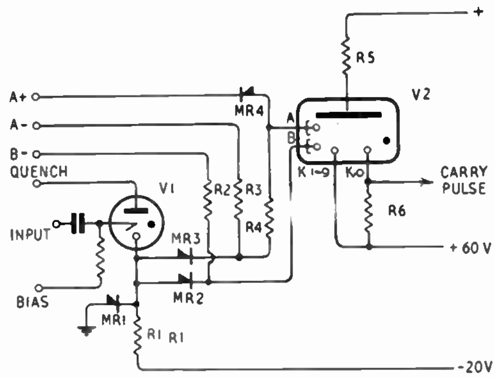

Fig. 5.—Typical stage of cascade for addition or subtraction.

When the unit is required to add, a trigger pulse is applied to V1 and after about 200 μsec, the common "A" pulse is applied to the A + line as shown in Fig. 6. At the end of this pulse, the quench pulse is applied to the anodes of all the trigger tubes. The GC10B then transfers forward via a guide "K" and a guide "A" to the following guide "B." The waveforms in the 5th stage of the cascade are shown in Fig. 6, where it is seen that the positive-going front edges of the pulses on guides "A" and "B" of this stage have the cumulative delay through the four preceding stages. This does not result in erratic counting provided that the delay between the positive-going edge of the "B" pulse and the negative-going edge of the "A" pulse in this stage is greater than 80 μsec. It should be noted that, when V1 is fired, current flows through R2 and R3 to the B − and A − lines respectively. Since the number of trigger tubes that are fired by a pulse at the input to the cascade can vary from 1 to 5, it is important to ensure that the A − and B − lines do not change their potentials by more than about 5V as a result of this, since it may lead to erratic counting. It is also necessary to ensure that the pulse applied to the A + line from the common pulse generator is capable of taking this line to 0V notwithstanding the current that will flow through the resistors R4 of each stage in which V1 has fired.

When the unit is required to subtract, pulses from the common pulse generator are applied to the A —, B —, and quench lines as shown in Fig. 6. Also, the pulse that is applied to the first trigger tube is delayed by about 100 μsec on the input pulse. The sequence of events in the various stages will now be described. On receipt of the input pulse, the glow is made to transfer to the guides "A" immediately preceding the guides "B" that were glowing. When the delayed input pulse fires the first trigger tube, the guides "A" and "B" are taken to + 120V with the result that the guide "K" immediately preceding will now glow. When the quench pulse is applied, the glow will rest on guide "B" which is one digit earlier than it was at the beginning of the cycle. If the glow passed to a Ko guide during the above sequence a "carry" pulse will have been produced which will have fired the succeeding trigger tube, and the associated GC10B will have moved back one digit. In the case of those stages where a "carry" pulse is not received from the previous stage, the glow will return to the original guide "B" at the end of the cycle. It can be seen from Fig. 6 that the effect of the time delays due to the trigger tube and the rate of rise of the output pulse from the GC10B tube will result in a shortening of the duration of the glow on a guide "K" in the later stages of the cascade. The duration of the common A — and B — pulses and hence the delay to the front edge of the quench pulse from the input pulse must be made long enough to avoid erratic counting. It should be noted that, as in the case of addition, the common pulse generator should be capable of maintaining the correct potential at the A —, B — and A + lines irrespective of the number of trigger tubes in the cascade that happens to fire for an input pulse.

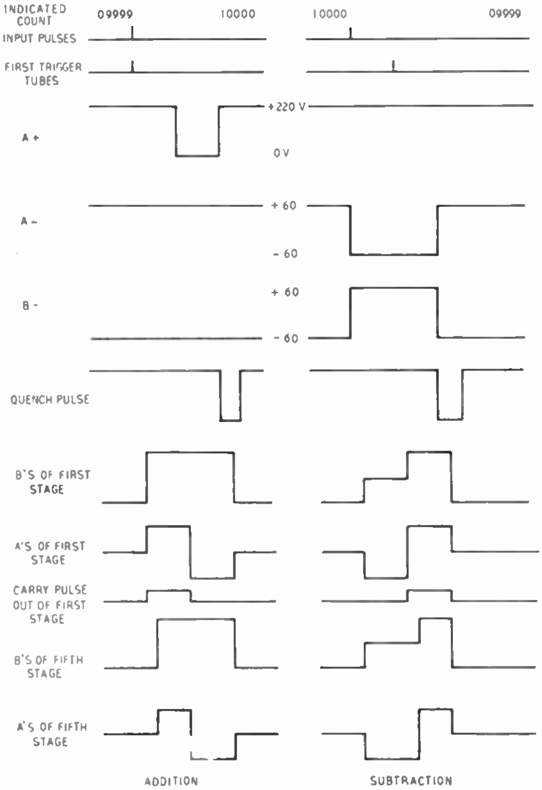

Fig. 6.—Waveforms for addition and subtraction using circuit of Fig. 5.

3. Frequency Division and Counting a Predetermined Number

The use of the new counting tubes for dividing the repetition frequency of a train of pulses by factors of 10 is a further application of their normal scaling circuits. Some important industrial uses are: the accurate measurement of an unknown frequency using standard timing pulses as a reference; measuring the time between two events; and the provision of frequency signals which are sub-multiples of a standard frequency.[9, 10] In general, the total number of components for each divider using one of the new counting tubes will be greater than that needed by one of the monostable trigger circuits which are normally used where the pulse spacing is fixed. However, one important advantage of the circuits using the new tube is that the number of close tolerance components in each dividing stage can be smaller, and further, the dividing factor is completely independent of pulse spacing. It is expected that the circuits using these tubes will give long periods of trouble-free operation.

One advantage of the monostable trigger circuits is the ease with which the dividing factor can be changed from 1 up to about 10. Provided that the input repetition frequency does not vary by more than say 5 per cent., this method of frequency division is probably the simplest. Where, however, the dividing factor has to be independent of the input frequency, it is necessary to use a counting circuit. It is possible to alter the dividing factor in a circuit using the E1T counting tube from 1 to 10 in a relatively simple way. It is seen from the characteristics of the E1T shown in Fig. 7 that the 10 stable positions of the beam correspond to 10 voltage levels at the right-hand

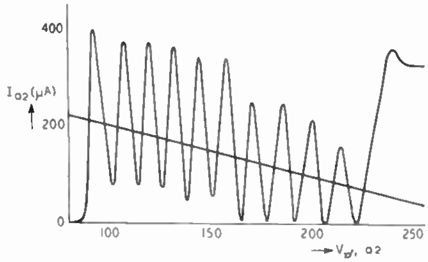

Fig. 7.—Typical characteristic of the E1T.

deflector, with a mean separation of about 14V between two adjacent points. The "0" position corresponds to about 240V and the "9" position about 100V. When pulses are applied to the left-hand deflector, the beam moves from one stable position to the next, and after passing position "9," the beam is normally reset to the "0" position by momentarily cutting off the beam current. This makes the right-hand deflector return to the highest potential which gives a stable beam position. If the right-hand deflector is caught at an intermediate potential by a diode, then the beam can be made to reset to any one of the other stable positions. This reduces the number of stable positions available and consequently allows division by any number up to 10. The use of the E1T in a counting unit to provide an output at the end of a predetermined number of counts, N follows from the above method of setting the beam to any one of the 10 stable positions. If a cascade of say five stages is used, then the stages are initially set up to the number $100{,}000 - N$ by momentarily arranging to "catch" the right-hand deflectors at suitable potentials whilst cutting off the beam currents. If, now, N pulses are applied at the input, the unit counts in the normal manner and the last stage will give an output pulse at the Nth input pulse. It can then be again set to the original $100{,}000 - N$ position by resetting as above.

It is also possible to arrange for the trochotron RYG10 to divide by any number less than 10. Let us first consider the normal operation of the trochotron in the arrangement which allows division by 10. The basic circuit diagram is given in Fig. 8, and the waveforms in Fig. 9. Normally the spade resistors R1, R2 – R10 are all equal, and so are all the spade capacitors C1, C2 – C10. A negative pulse whose duration is approximately $0{\cdot}8\ CV/I$ (where C is the total capacitance at the spade, V is the spade voltage, and I is the cathode current) is applied to the anode in order to shift the beam from one stable position to the next. If the negative input pulse on the anode moves the beam from spade one to spade two, the voltage on spade one rises with the time constant C1R1 whereas the voltage on spade two falls much more rapidly since the current flowing to this spade during the input pulse is the full cathode current of about 10 mA. The quiescent current on the spade which passes current is usually less than 1 mA. The waveforms for this condition are as in Fig. 9a. If the duration of the negative pulse

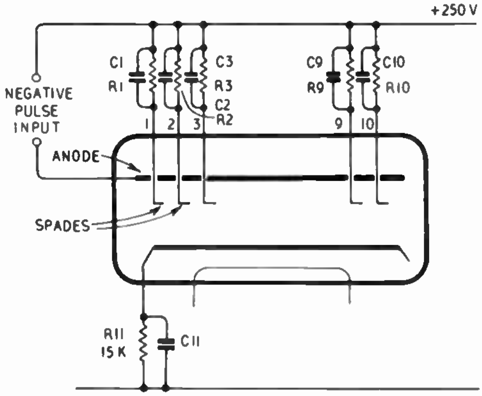

Fig. 8.—Basic circuit for trochotron RYG10.

on the anode is increased, the beam will move past spade two after the potential of this spade has fallen by about 80V. By increasing the duration of the driving pulse to $2 \times CV/I$ it is possible to make the beam miss every other spade and therefore divide by five. Suppose it is necessary to divide by nine. If one of the spades, say spade two, has a larger load resistance and a much smaller capacitor the beam will always miss spade two, and the waveforms will be as shown in Fig. 9b. If this technique is used on many spades, the value of the cathode

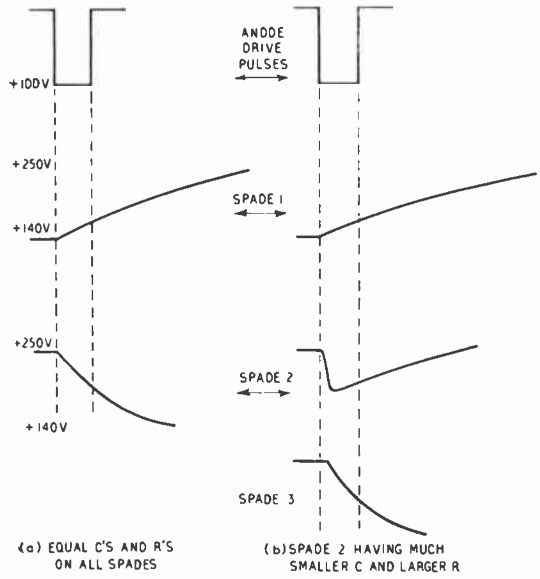

Fig. 9.—Waveforms in trochotrons.

resistor R11 will become critical at high repetition frequencies. An alternative method of making the trochotron divide by a number less than 10 is to join the requisite number of spades together, and connect them to a common load resistor and capacitor. If more than three spades are joined together for this application the anode and spade supply voltages will require re-adjustment.

In the gas-filled counting tubes, the digit stored in the tube can be identified by the guide to which current is passing. In order to be able to set the tubes to any required number, it is necessary to bring the 10 guides corresponding to the numbers to separate connections outside the tube. By then applying suitable initial potentials to these guides, these tubes can also be used in a predetermined counter. These tubes require a large negative pulse to a guide in order to bring the glow to that guide. In the case of the GC10B, a negative rectangular pulse of at least 200V amplitude with a duration of a few milliseconds is necessary to reset the glow to an arbitrary guide. A pulse with an exponential decay is also effective provided that the time constant is more than 10 milliseconds. The faster gas-filled tubes can be reset with slightly shorter pulses, but it is generally true that the time necessary to reset in this manner is much greater than the resolving time of the tubes.

It will be seen from the preceding statements that the use of tubes other than the E1T or the RYG10 for frequency division by a factor other than 10 will result in more complicated circuits associated with each tube. There are two exceptions corresponding to division by 2 or 5. If we connect together K0, K2, K4, K6 and K8 of a GS10B, in which all cathodes K are brought out to separate leads, and derive the "carry" pulse from this common point, it will be seen that this stage will divide by 2. Similarly, if we take the "carry" pulse from K0 and K5 connected together then we can divide by 5. In general, by connecting together any number N of cathodes K in a 10-position tube, we can divide by the number $10/N$, but the output pulses from such a stage will not be equally spaced except in the cases of division by 2, 5 or 10. Where regular pulse spacing is unnecessary, for instance, when calibrating ratemeters, then the circuit of Fig. 10, giving a variable dividing factor from 1 to 10, is very economical in components.

In certain cases of frequency division such as in the generation of standard time signals from a quartz crystal oscillator,[11] it is necessary to maintain the least possible time delay between the output pulse and the associated input pulse. In the case of the trochotron RYG10 and the E1T this time delay can be made less than 1 μsec per stage, using the circuits recommended for those tubes. With slight modifications to the circuits, this delay can be made less than 0·3 μsec, but this will in general lead to closer tolerance for some of the component values. Using negative-driving pulses, the time delay between input and output pulses for the GC10D and the G10/241E will be 20 μsec or more, and for the GC10B it will be more than 160 μsec per stage. By driving the gas-filled tubes with positive pulses as in the system using a common pulse generator, this time delay can be made less than 2 μsec per stage for the GC10D and 10 μsec per stage for the GC10B.

4. Switching Applications

All of the counting tubes discussed earlier with the exception of the E1T are capable of being used as 10-way switches. A modified version of a tube similar to the E1T, and other switching tubes of the trochotron type have been described[10] recently. One possible application

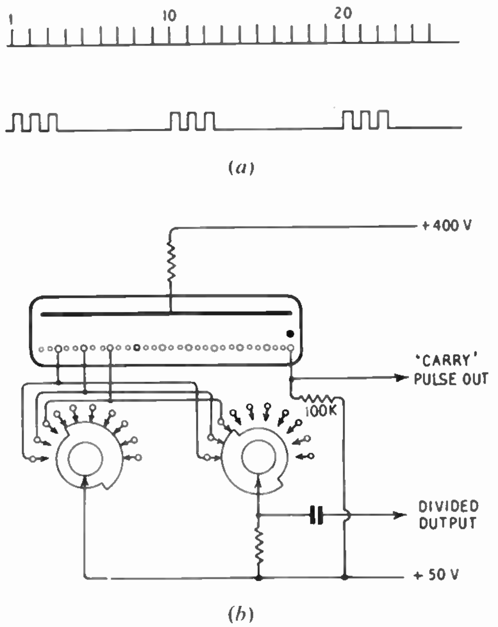

Fig. 10.—Method of obtaining calibrating frequencies.
(a) Waveforms with K1, K2, K3 connected together.
(b) Switching arrangement.

of switching tubes is in communications where normal audio or radio frequency signals are used. A prime requirement for such applications is that the switch should not introduce spurious signals or distort or attenuate the signal being transmitted through the switch. The gas-filled tubes have the disadvantages of being able to switch only at relatively low speeds and transmitting satisfactory signals only at audio frequencies. If the signal is fed into the anode of a gas-filled tube and taken out of one of the guides the voltage and current gains are equal to 1. The signal can therefore be transmitted through a number of tubes without appreciable attenuation. The noise in the tubes is generally not greater than a few millivolts r.m.s. between anode and guide, and in the case of good tubes may be as small as a few hundred microvolts. The maximum signal that can be transmitted through the tubes is about 10V r.m.s. with the current being limited to about 100 μA r.m.s.

At the other extreme of switching speeds the trochotron of the RYG10 type can change over from one spade to the next in about 0·3 μsec. Signals applied to the cathode can be switched to one of the 10 spades. The voltage gain of the tube is slightly less than 1, and the current gain is less than 0·05 for the RYG10. In the case of the tubes with separate collectors[10] such as the TR-SR-11, the current gain is about 0·2 and the voltage gain can exceed unity. It is seen that these tubes attenuate the signal considerably and the use of tubes in cascade for switching purposes involves amplification of the signal and will result in additional valve circuits in series with each signal. The greatest disadvantage of these tubes lies in the high noise level which can be as high as 10 μA r.m.s. for a mean spade current of 1 mA. It should be observed that the complete shot noise given by

$$\bar{I}^2 = 2eI\,\delta F$$

for a current of 1 mA is 0·018 μA assuming δf to be 10^6 c/s. In tubes such as the TR-SR-11 where the collectors are not in any way connected to the spades, the noise current is expected to be a smaller fraction of the mean collector current although it will still be considerably higher than the true shot noise.

The use of various counting tubes as coders and decoders in pulse code modulation systems and as decimal-to-binary converters is expected to lead to a reduction of the total number of valve elements in such systems. An interesting immediate application is in a system where information is converted from an analogue to a digital system. One particular case is the pulse-amplitude analyser. In this instrument the pulses in the amplitude range $0 - E$ volts are counted in n channels each of which accepts pulses in the range 0 to $\frac{E}{n}$, $\frac{E}{n}$ to $2\frac{E}{n}$, $2\frac{E}{n}$ to $3\frac{E}{n}$, . . . and $(n-1)\frac{E}{n}$ to E respectively. In one method, a capacitor is charged to the peak voltage of each signal pulse as it arrives, and at the same time an accurate linearly increasing voltage and a pulse train of fixed spacing is generated. The first pulse of the train opens the first channel, the second pulse closes the first channel and opens the second channel, the third pulse closes the second channel and opens the third channel and so on. When the linearly increasing voltage reaches the voltage of the capacitor, which remains constant at the peak signal pulse voltage, a pulse is applied to all the channels and only the channel which is open at that instant will register this pulse. Some form of digital counting system is used in each channel and the number of pulses of a particular amplitude range is therefore counted in one channel. As soon as the pulse is counted in a channel the capacitor is discharged and the system is ready to accept another pulse.

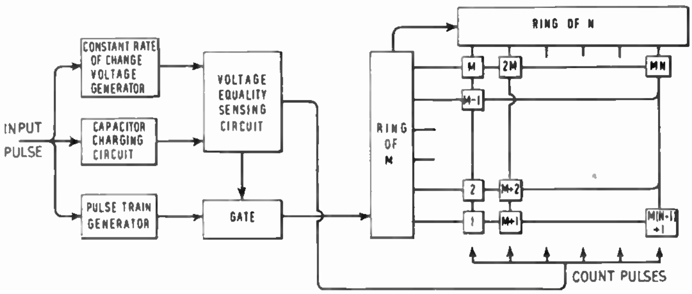

Fig. 11.—Block schematic of analyser.

A novel method[11] of reducing the total number of valves in a multi-channel pulse analyser of the above type is to use two counting circuits consisting of a ring of m and a ring of n respectively and to connect them in a matrix as in Fig. 11. Every complete circulation in the ring of m moves the ring of n by one stage. There are mn combinations of states in the two ring circuits in which one stage of each ring circuit is in an odd state. In order to use this in a pulse-amplitude analyser of the type just described, the pulse train is applied to the input of the ring of m. Each channel of the analyser consists of a triple coincidence arrangement which receives an input from one stage of the ring of m, an input from one stage of the ring of n, and a counting pulse, which is common to all channels. The pulse train, which starts when the capacitor is being charged to the peak input pulse, is counted in the ring circuits until the linearly increasing voltage reaches the voltage on the capacitor. When this state is reached the pulse train is interrupted and the counting pulse is applied to all channels. Only that channel which is connected to the odd stage in the ring of m and the odd stage in the ring of n will register the pulse. This is therefore a pulse-amplitude analyser with mn channels, where the mnth channel counts all pulses above a certain level.

An interesting system has been worked out on the above general principles but using RYG10-type tubes in place of the ring counters and using GC10B tubes to perform the combined functions of the coincidence unit and the digital counter of each channel. In the proposed system there are no components in each channel other than one GC10B, the anode load of the GC10B and the output cathode resistor. The pulses out of the GC10B in each channel can be counted in further counting stages of any desired type.

The block schematic of the new 100-channel pulse analyser is shown in Fig. 12 and one of the two ring circuits and gates is shown in Fig. 13. The ring circuits are identical to each other. The main differences in function are that ring A receives its drive pulses directly from the pulse-train generator, and the pulses through the gating valves V2, V4 . . . V20 are fed to the "A" guides of the GC10B tubes; whereas the ring B receives its drive pulses from the "carry" pulses out of ring A, and the pulses through the gating valves are connected to the "B" guides of the GC10B tubes. The RYG10 has the usual cathode resistor R4 and spade resistors R5, R8 . . . R32. The nine spades that are not passing current are caught by the corresponding diodes V1, V3 . . . V19 at + 250V. The spade that is conducting will be at about 10V below the potential of the cathode of the RYG10. The grids of the left-hand sections of the gating valves V2, V4 . . . V20 are connected to spades 1, 2 . . . 10 respectively of the RYG10. The grids of the right-hand sections of the valves are connected together and returned to the junction of R2 and R3 which is about 20V above the potential of the cathode of the RYG10. The anodes of the right-hand sections are connected to the GC10B's, those from ring A being connected to the "A" guides and those from ring B to the "B" guides as shown in Fig. 12.

In Fig. 13, the valves V22 and V24 are normally cut-off. Hence the cathodes of V1, V3 . . . V19 are held by diode V23 at + 250V and the cathodes of the gating valves V2, V4 . . . V20 are at about + 275V, there being no current through any of the gating valves. One of the gating valves, corresponding to the spade of the RYG10 that is conducting, has its left-hand

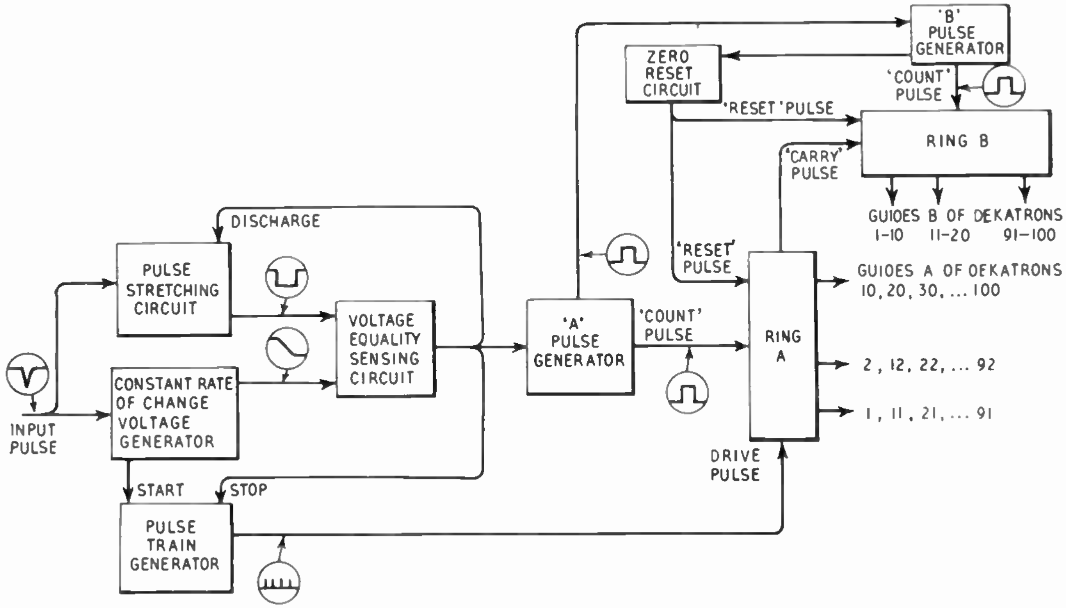

Fig. 12.—*Block schematic of pulse analyser using trochotrons and dekatrons.*

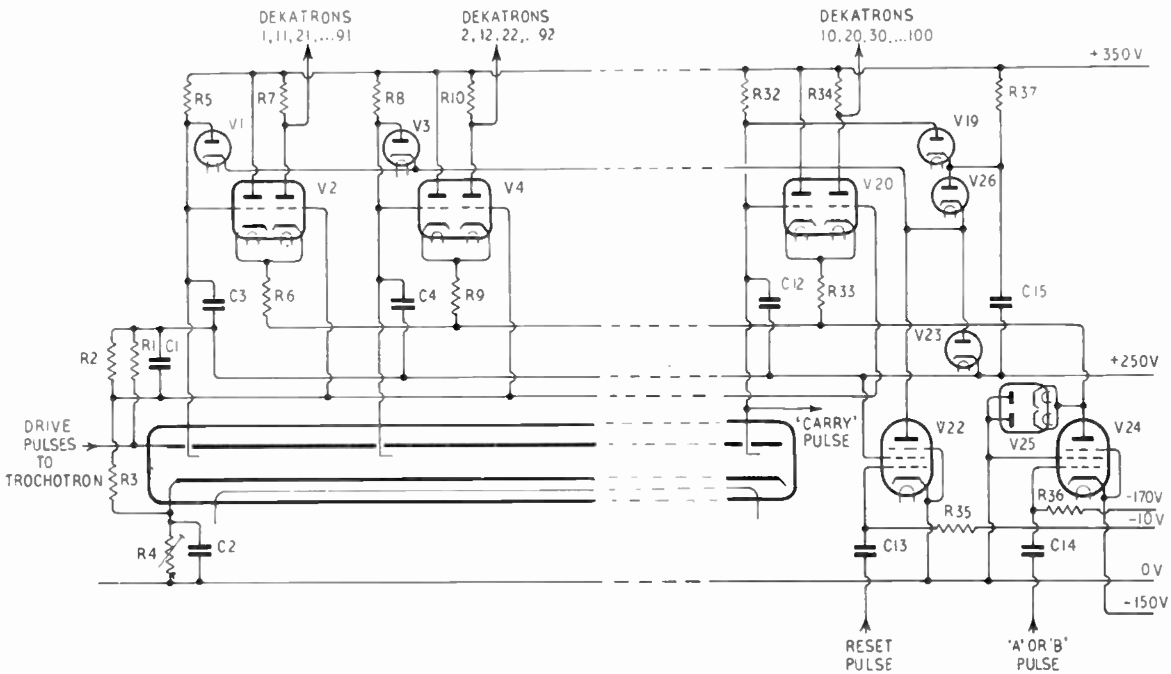

Fig. 13.—*Ring circuit and gates for pulse analyser using trochotron RYG10.*

grid at about + 120V, whereas the left-hand grids of all the remaining gating valves are at + 250V. The right-hand grids of all valves are at about + 150V. If the valve V24 is now brought into conduction there will be a negative pulse out of the right-hand anode of that gating valve which is connected to the conducting spade.

The operation of the complete pulse analyser can now be described with reference to Fig. 12 and the waveforms as shown in Fig. 14. The waveforms refer to a signal pulse of an amplitude corresponding to channel 25. The signal pulse is fed through a diode to a capacitor which is therefore charged to the peak value of the signal. The capacitor remains charged to this voltage until the information corresponding to this signal is stored in the appropriate channel. When the capacitor is charged, the constant rate voltage and pulse-train generators are started simultaneously. The pulse train which can have a spacing of about 2 μsec is fed to ring A and the carry pulse from ring A is fed to ring B. When the constant rate voltage reaches the voltage stored on the capacitor the level-sensing circuit stops the pulse train. Immediately after this a positive rectangular pulse of about 100 μsec duration is fed to V24 of ring A which therefore applies negative transfer pulses of about 120V to the "A" guides of the GC10B's 5, 15 . . . 95 in the case of this particular pulse amplitude since ring A will be stopped on position 5. Immediately after the "A" pulse, a similar positive "B" pulse is applied to the corresponding point in ring B which will therefore apply a negative transfer pulse to the "B" guides of GC10B's 21 to 30. Since the only GC10B which received both an "A" and a "B" pulse is tube 25, this tube and no other will transfer by one digit. At the end of the "B" pulse a positive reset pulse is applied to the grids of the resetting valves V22 in the two-ring circuits. This causes the cathodes of the diodes V1, V3 . . . V19 to come down to nearly earth potential and recover slowly to the normal potential of + 250. Spades 1 to 9 will recover faster than spade 10, as shown in Fig. 14, since the recovery of spade 10 will be determined by R37 and C15, which is made longer than the recovery of the normal spade circuits.

In a complete analyser it is necessary to have certain ancillary circuits in order to avoid errors. It will be appreciated that in the system described above, any pulse larger than E volts in amplitude will be counted in one of the channels as if it was a pulse of amplitude less than E volts. It is therefore necessary to have a circuit which

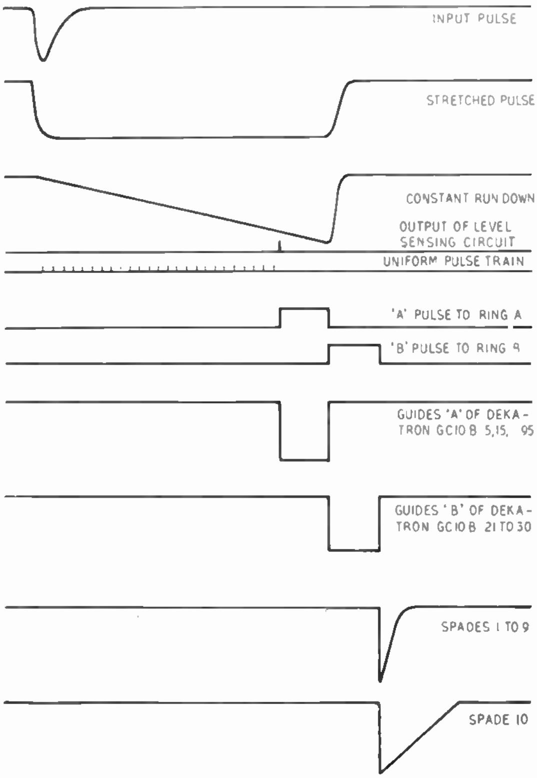

Fig. 14.—The waveforms in the pulse analyser of Fig. 12.

ensures that each gate shall be open once only during each cycle. It is also necessary to introduce a gate which prevents pulses entering the stretching circuits while a previous pulse is being analysed by the unit.

The simplicity of the above system can be appreciated by comparing the total number of components in the circuits associated with the sorting of the pulses into the various channels in different types of analysers. In one of the earlier types of analyser[13] there are approximately 80 valves and 400 components in the sorting circuits for a 30-channel unit. For the purposes of this estimate, double valves in one envelope have been counted as two valves and the valves and components which perform functions common to many units are not included. The scaling and counting circuits in the individual channels are not included in the above estimate. On the same basis the analyser described in this paper uses 72 valves and 100 components in the two ring circuits and 100 GC10B dekatrons and 200 resistors in the 100 channels. It will accept signal pulses with a resolving time of about 500 μsec, whereas the 30-channel analyser referred to will resolve signal pulses which are 150 μsec apart. It can be said that the new analyser provides 100 channels with fewer valves and much fewer components than the 30-channel analyser.

Another comparison can be made with the slow analyser[14] working on similar principles. In this unit, which will resolve signal pulses 100 msec apart, there are approximately 160 valves and over 200 components. It should be noted, however, that no further components are necessary to store the number of pulses in each channel since mechanical registers which can store four digits are used. In the unit described in this paper it will be necessary to use 100 valves in order to work the registers following the dekatrons, and there will be five components associated with each valve. The main improvement that has been obtained is the resolving time of 500 μsec as opposed to 100 msec in the slow analyser.

Another important instrument for the study of the interaction of neutrons with nuclei is the time-of-flight spectrometer.[12] In this instrument a pulse of neutrons of short duration but of varying energies is allowed to travel a known distance, and the time delay in the arrival of the neutrons at the detector is measured in order to give information on the energy of the neutron.

The number of neutrons arriving with time delays of 0 to t, t to $2t$, $2t$ to $3t$ . . . $99t$ to $100t$ are counted in 100 separate channels. It is seen that the ring circuits and associated GC10B's in the analyser described above, can be used directly for this purpose. It is only necessary to use the pulse-train generator which is started when the neutrons leave the source, and stopped when a neutron is received at the detector. This system will only permit the recording of one neutron during each complete cycle; but in those instances where the neutron source transmits one neutron in the direction of the detector for every few pulses at the source this does not lead to serious error in the measurement.

The above instrument can be used for any application in which a delayed signal is generated from some main event and where the exact time delay is not known and where there is a fluctuation in the time delay. A simple decimal counting unit with two decades is sufficient to measure a time delay of up to 100 units only where there is one time delay involved.

5. Conclusions

The counting tubes discussed have shown great promise not only in simple circuits where the total number of events of one type is to be counted, but also in complex equipment such as pulse-amplitude analysers and time-of-flight spectrometers. In every application where information in a digit form has to be analysed or stored, these tubes provide a means of obtaining accurate information of a complex nature in relatively small and simple equipment. The life of these tubes has been found to be comparable to normal valves and since one of these tubes performs the functions of many valves the ultimate reliability of equipment is expected to be improved considerably. This should lead to the use of these tubes in industrial measurements and process control of complex types in which combinations of valve circuits have lacked the necessary simplicity and reliability.

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DISCUSSION*

N. Armitage: I am interested in protection problems on power lines and in these circumstances reliability is most important. We have begun experiments with cold cathode tubes for various switching relay circuits. We have no operating experience of the life of these tubes and I would therefore like to ask if Dr. Taylor can give us actual figures of their life? We have been unable to get any information from the manufacturers apart from "several thousand hours."

I would also like to ask a question of Mr. Kandiah. We may use cold cathode tubes for counting fault incidence. When the tube comes to and end of its useful life, is that end indicated rapidly; or does the valve gradually become unreliable, and, if so, in what manner?

Dr. D. Taylor (*in reply*): I do give one figure in the paper† for the life of cold cathode tubes. Replying to the question further, the type of trigger tube used in the monitor described by Stephens‡ has a failure rate of the order of ½ per cent. per annum. We have used the multi-electrode tube that Mr. Kandiah describes in one of our digital computers, and the life figures are of the same general order as the better-type thermionic valves.

K. Kandiah (*in reply*): I would like to make one further comment in addition to what Dr. Taylor has said about cold cathode switching tubes. I feel that there has been some conflict in the general opinion on the life of cold cathode trigger tubes. When it is stated that reliability is poor, I am certain that this is due to lack of appreciation of the limitations of these tubes, which are of quite a different character to thermionic valves. Taking those limitations into mind there is no doubt that the life of these tubes is very long.

Multi-electrode tubes and switching tubes do not fail catastrophically. If a particular characteristic is measured, it is extremely unlikely that this will have changed appreciably in the next few days except in a few special cases. The few special cases are those in which the characteristic is a function of the nature of the surface inside the tube, such as in some types of activated tubes or when the tube contains hydrogen or similar gases. Some of those tubes can change their characteristics in a few hours, but in general it is a very slow process and this sort of defect can very easily be overcome by marginal test before the tubes are put into use.

W. Nock: We have been using switching tubes of the type described by Mr. Kandiah since their introduction in 1950; our primary use is in time measurements, although in recent years we have used them for batching and control purposes. We have had very reliable results except for the first version of the GC10D (20 kc/s tube) which became "sticky" after a short time in use. We have recently obtained a modified version of this tube and this is giving very promising results. We have had a timer using one in use for four months.

Mr. Kandiah mentions the use of selector tubes with ten cathodes, GS10B, to give fixed counts of 2 and 5. Using the GS12B with 12 cathodes, it is also possible to produce stages with counts of 2, 3, 4, 6 and 12. In addition we have used a GS10B with forced reset to give stages counting in any number up to 10 at speeds up to 2 kc/s.

*Part of this Discussion has already been published in the November 1954 issue of the *Journal*, following the paper by Dr. Taylor. It is repeated here for the sake of completeness.

†*J.Brit.I.R.E.*, 14, p. 568, November, 1954.

‡*J.Brit.I.R.E.*, 14, p. 377, August, 1954.