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Commentary

THE history of electrical communications between this country and America could, until recently, have been divided into two well defined phases.

The first was the one in which the telegraph cable was dominant and it opens in 1858 with the laying of the first transatlantic telegraph cable by H.M.S. *Agamemnon* and the U.S. Frigate *Niagara* in the summer of that year. This cable functioned only for a few months. It failed completely in the autumn and there was a lapse of nearly ten years before the *Great Eastern* set out on her historic voyage.

Thereafter for some sixty years the telegraph cable reigned supreme and in that period a vast network of cables, spreading to the furthermost corners of the earth, had been buried beneath the oceans.

The supremacy of the telegraph cable was not to remain unchallenged and the second phase may be said to have started in 1927 when a radio telephone link was installed between Britain and America allowing the spoken word to be exchanged across the Atlantic for the first time.

What may be looked upon as a third phase in this history was ushered in on September 26 of this year, for it was on this day that the final splice in the first transatlantic telephone cable was made on board H.M.T.S. *Monarch*.

It was on December 1 1953 that the Postmaster General of the day announced that an agreement had just been signed between the British Post Office, the American Telephone and Telegraph Company, the Canadian Overseas Telecommunication Corporation and the Eastern Telephone and Telegraph Company for the construction of a transatlantic telephone cable to link the United Kingdom on the one hand with the United States and Canada on the other.

At the cost of some £15 million the scheme will provide 29 high grade and reliable telephone circuits to the United States and six to Canada with an additional circuit for telegraph channels to improve the telegraph service to Canada and, as necessary, to Australia and New Zealand via Vancouver.

In the spring of this year, the *Monarch*—the largest cable ship in the world—laid out from Clarenville, Newfoundland, the first section of the transatlantic telephone cable consisting of some 200 miles in length. In August, after loading of more cable at Erith, Kent, *Monarch* picked up the buoyed end of the first section and then laid in one length 1 500 nautical miles of cable with repeaters across the deepest part of the Atlantic, buoying the end at Rockall Bank, some 500 miles South West of Oban. After loading again at Erith she proceeded to Rockall Bank to pick up the buoyed end and laid the last 500 nautical miles to Oban. Meanwhile, the *Iris*, a smaller cable ship, had laid out a short length from the terminal station at Oban and on September 26 aboard the *Monarch* the final splicing took place of the 2 000 miles of cable stretching back to Newfoundland with the shore end and for the first time Newfoundland and Great Britain were linked by a submarine telephone cable. Thus was completed the first stage of this remarkable example of joint Anglo-American enterprise.

A similar cable carrying the east to west circuits will be laid in the spring of next year. The transatlantic section of the telephone cable system will consist of about 2 000 nautical miles of cables with repeaters at intervals of some 40 miles from Oban in Scotland to Clarenville in Newfoundland. At the Canadian end the cable will continue overland for some 60 miles where it will connect with another submarine cable of 280 nautical miles to Nova Scotia and thence overland by a 360 mile microwave radio link to the United States border where it will feed into the Bell System network. At an intermediate point a connexion will be made to the Canadian network. Meanwhile, the existing coaxial trunk cable system in this country is being extended to Oban to link up with the International Telephone Exchange in London.

When the first transatlantic cable was laid in 1858 the Society of Telegraph Engineers, as the Institution of Electrical Engineers was originally known, was not in being —it was not founded until 1871—but the opening of the radio telephone link in 1927, as some of the older members of the Institution will recall, was marked by a unique meeting.

On February 16 1928 there was held at the Institution a special meeting, over which Sir Archibald Page presided, simultaneously with a meeting of the American Institute of Electrical Engineers and greetings were exchanged between the two Institutions over the newly opened radio telephone link.

A paper[1] of joint Anglo-American authorship, in which the historical background and the main technical details of the cable and repeaters are given, has already been presented to the Institution of Electrical Engineers and the American Institute of Electrical Engineers. When this remarkable achievement is completed, towards the end of 1956, more detailed papers will, no doubt, be presented.

It is to be hoped, therefore, that the suggestion of Sir Archibald Gill[2] will be acted upon and that a further joint meeting of the Institute and the Institution will be conducted over this new service.

1. KELLY, M. J., RADLEY, G., GILMAN, G. W., HALSEY, R. J. A Transatlantic Telephone Cable. *Proc. Instn. Elect. Engrs.* *102*, Pt. B, 117 (1955).
2. GILL, A. Discussion on above. *Ibid* p. 131.

An Instrument for **Investigating Automobile Brake Usage** *under Practical Conditions*

By H. Rosemary Taylor*, M.Sc., A.Inst.P.

The instrument provided a detailed record of when and how the brakes on a car were used so that the effects of speed, driver, route, traffic, etc., could be investigated. A tachometer generator was driven by the propeller shaft and so generated a voltage proportional to the angular speed of the rear wheels. This signal was differentiated electrically to measure the angular deceleration of the wheels. A voltage proportional to the power dissipation at the brakes was derived from the product of the speed signal and the hydraulic pressure in the braking system which was measured with a potentiometer transducer. The temperature of the bulk of one brake drum was measured by a thermistor. These four signals were suitably amplified and applied to a four-pen recorder. Timing marks were provided so that the following quantities could be calculated: brake application time, mean deceleration, mean power dissipation and total energy dissipated during each application and during each journey. A relay circuit ensured that the pens recorded during brake applications only.

The speed unit was calibrated against a hand tachometer with the rear wheels jacked up. The deceleration and power dissipation units were calibrated by comparison with the speed/time record during controlled brake applications. The power calibration equation and the conditions under which it was valid are derived in the article. The temperature unit was calibrated against a thermocouple with the appropriate brake drum in an oven.

The article concludes with an assessment of the accuracy of each measurement and a few examples of the type of records produced in normal driving.

THIS instrument was designed to record, without attention from the driver, the significant factors of all brake applications during normal driving on public roads, so that the effects of average speed, driver, route, traffic, etc. on brake usage could be investigated Four variables were automatically plotted against time for each brake application. They were (1) angular speed of the brake drums, (2) angular deceleration of the brake drums, (3) power dissipation at the brakes, and (4) the bulk brake drum temperature. From the chart, the following were calculated: (5) the application time, (6) the mean deceleration, (7) the mean power dissipation, and (8) the total energy dissipated during the application. Each test consisted of a journey of about 50 miles over a given route, and the following were calculated for each test: (9) the number of applications, (10) the total braking time, (11) the total energy dissipated.

From this information, a semi-statistical survey of brake usage, test by test, and a detailed examination, application by application, are being made. Many results have already been obtained and details of the automobile engineering aspects will be published elsewhere.

Measuring Techniques

The four signals were recorded simultaneously, together with timing marks, on a Kelvin and Hughes pen recorder which was chosen for its suitability for use in a moving vehicle. Typical records are shown in Fig. 1.

The investigation was concerned with brakes, and it was therefore considered preferable to measure the angular speed of the brake drums, which was proportional to the forward speed of the car, except for generally small dis-

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crepancies due to tyre slip. This was done by driving a tacho-generator from the propeller shaft. The output of the generator was sufficient to operate the recorder without amplification.

Similarly, the deceleration to be measured was that of the brake drums and so the deceleration signal was derived from the speed signal by electrical differentiating circuits. A further reason for adopting this technique in place of the more usual spring-mass or pendulum decelerometer was that both the latter are subject to errors from road gradient and the tilt of the car on its springs. Being concerned with practical conditions, the investigation included work in very hilly country as well as flat country and towns, and road gradient could not be ignored. When measuring the maximum deceleration it was necessary to set an arbitrary upper frequency limit because the motion of the wheels, drums and propeller shaft included vibrational accelerations and decelerations far in excess of the deceleration produced by the brakes. The amplitudes and frequencies of these vibrations depended on the speed of the car and the nature of the road surface, but there was an important component at the natural frequency of the car springs which was about 2c/s. Smoothing circuits were added and adjusted until these vibrations were just tolerable on the roughest roads in the test routes. The response time was then approximately the same as that of the car suspension. Hence the record produced by the instrument was almost the same as would be produced by any decelerometer placed in the car body except that such an instrument would also be subject to errors due to car tilt and road gradient.

The power dissipation at each brake is given by the torque multiplied by the angular speed and, with the tacho-generator available, the problem reduced to measuring the total effective torque of the four brakes. Direct methods were not considered because of the drastic alterations to the car which would be required, and although a deceleration measurement was available, it could not be considered proportional to the torque because of the effects of rolling resistance, engine drag and road gradients. However. the torque was proportional to the hydraulic pressure in the braking system provided that the shoe factor* remained constant and that allowance was made for the pressure required to overcome the pull-off springs and move the linings into contact with the drums.

A pressure transducer was fitted in the hydraulic pipeline and could be energized by the tacho-generator to record power dissipation or by a constant voltage to record the pressure. Measurements of pipe-line pressure were not needed in the main investigation but extended the usefulness of the instrument.

A temperature measurement was required that would enable sections of tests to be compared, although it was not necessary to know the distribution of temperature round the drum or the surface temperature. It is well known that the surface temperature of a drum is very much higher than the bulk temperature and may be very uneven[1,2,3,4]. The output of a thermocouple would be too small to operate the recorder with only one stage of voltage amplification, and would introduce difficulties with spurious thermo-electric e.m.f.'s and contact resistance at the slip-rings, and a radiation detector would require cooling. A thermistor was therefore chosen as the transducer and it

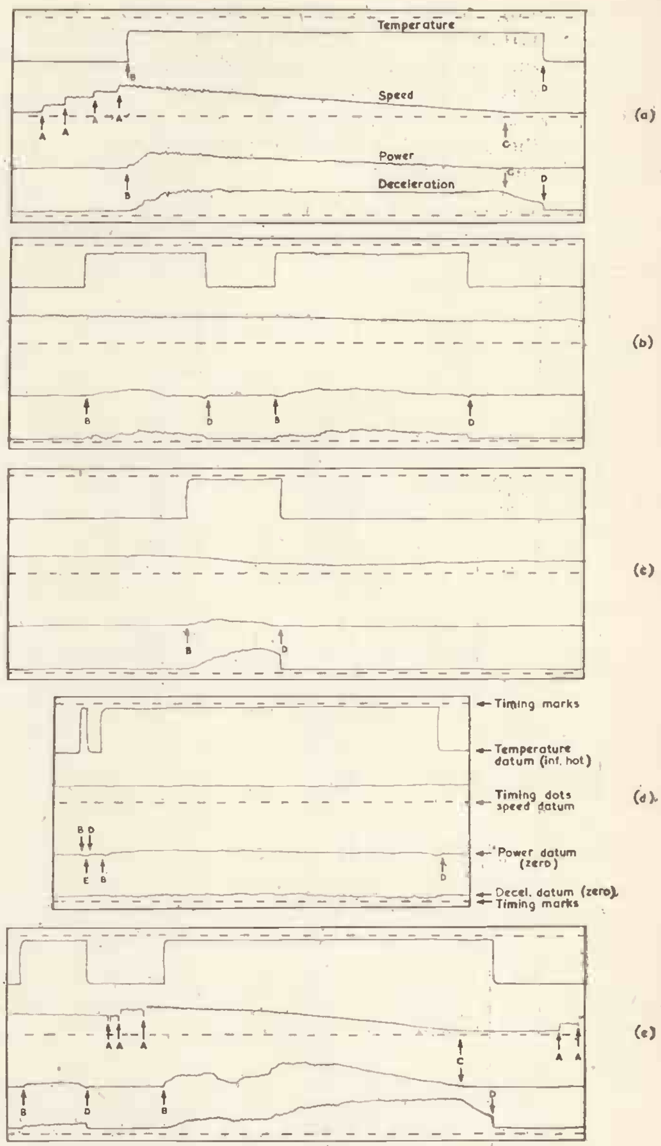

Fig. 1. Typical records

(a) Constant deceleration application for calibration calculations. Intermittent recording. (b) Braking for a corner on a downhill gradient. Continuous recording. (c) Braking for a corner on an uphill gradient. Continuous recording. (d) Keeping speed constant on a downhill gradient. Intermittent recording. (e) Braking to standstill at a road junction. Intermittent recording.

was soldered inside one brake drum as closely as possible to the braking path.

A system of relays ensured that the temperature, power and deceleration pens recorded datum lines before and after each application. Speed and timing marks could be recorded during applications only or throughout the test as required.

Transducers and Circuits

A block diagram is shown in Fig. 2 and a complete circuit diagram in Fig. 3. The apparatus was in three separate cases linked by multi-pin plugs. The "power unit" contained two vibrator packs, two rotary convertors, fuses and part of the control circuit. The "amplifier unit" contained three amplifiers, most of the control circuit and the speed unit. The "recorder" was a Kelvin and Hughes four-pen recorder Mark IV, which included an oscillator to

* *Shoe factor: The frictional force at the braking surface divided by the force applied to the toe of the brake shoe. It is a function of the coefficient of friction and of the geometry of the brake.*

provide timing marks. This apparatus was a semi-permanent fixture in a Bristol 2-litre saloon car. The back seat had been replaced by a wooden shelf to which the power unit was fastened directly and the amplifier and the recorder by anti-vibration mountings. The batteries were screwed to the shelf behind the apparatus and were independent of the car battery.

Control Circuit

The power supply for all the apparatus was controlled by

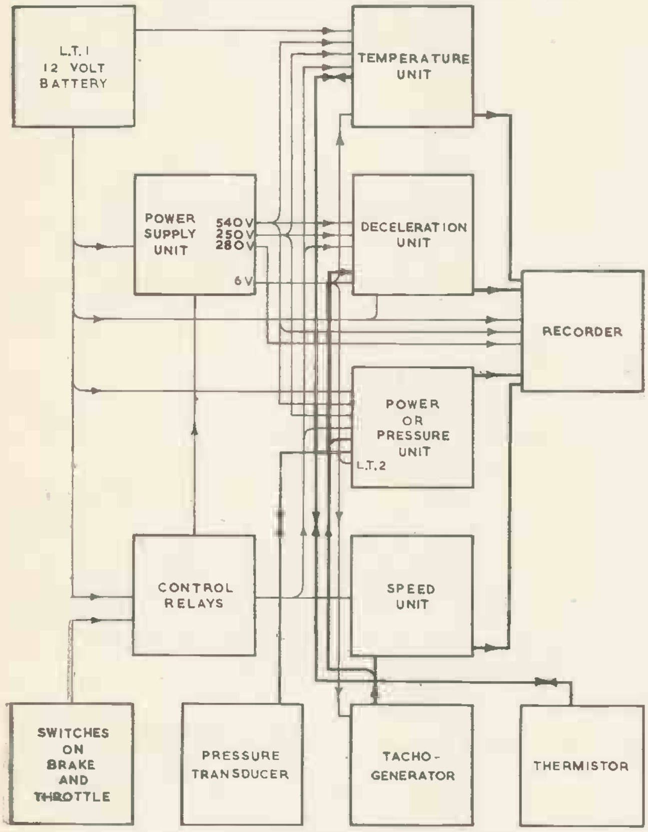

Fig. 2. General arrangement of the apparatus

the main switch S_5, which allowed current to reach the valve heaters and the barretters. The Switch S_6, which was not closed until immediately before a test, connected the l.t. supplies to the remainder of the apparatus, excepting the chart motor. When a record was required throughout a test as illustrated in Fig. 1 (b and c) the switch S_7 was closed which controlled the chart motor via relay contact E_1 and the timing marks and pen writing voltage via E_2. The deceleration, power dissipation and temperature pens were held in their datum positions by relay contacts $B_{1,2,3,4}$, until a slight depression of the brake pedal closed a micro-switch S_{10} and energized relay B. This is shown at B in Fig. 1 (a, b, c, d and e) and the opening of S_{10} is shown at D. When intermittent recording was required as illustrated in Fig. 1 (a, d and e), it was necessary to anticipate each brake application so that the chart motor accelerated and datum lines were recorded by each pen (except speed before the brakes were applied). A micro-switch S_{11} was mounted on the engine so that it was depressed when the throttle was closed and released as soon as the throttle began to open. When the driver was about to brake and took his foot off the accelerator pedal, this switch was operated and the capacitor C_9 discharged through relay A. During each brake application, relay B was energized as described above, and contact B_5 kept relay A energized and recharged C_9 so that after the brakes were released, relay A remained closed for a further three seconds or until the accelerator was pressed again, whichever occurred first. If the brakes were not applied within three seconds of the accelerator being released, relay A opened and only a few inches of chart were wasted. On the very rare occasions when the brakes were then applied, a normal record was produced except that the chart motor had to accelerate during the stop and the stopping time measurement was not reliable. Contact A_1 controlled the chart motor, pen writing voltages and timing marks.

Power Supply Unit

All the power was supplied by 12V batteries which had a total capacity of 115 ampère-hours. The negative terminal was connected to the chassis and cases.

The field winding of the tacho-generator was energized through two barretters, and the voltage across it was used as a 6V stabilized supply for the pressure transducer and the thermistor. Two vibrator packs (Masteradio VPG.554) provided h.t. at 250V for the output stages of the amplifiers. Rotary convertors supplied h.t. at 280V for the timing oscillator output stage and at 540V for the first stages of the amplifiers and the timing oscillator. The negative side of each h.t. supply was connected to the chassis, and the vibrator packs and rotary convertors ran throughout each test. An attempt was made to switch them on only when recording, but the extra current drawn from the batteries as the convertors accelerated caused a temporary fall in l.t. voltage which affected the amplifiers, the barretters being too slow to compensate for it.

Recorder

The recorder, which was modified to suit this investigation, produced the four records and three rows of timing marks on a Teledeltos chart 3in wide moving at about 1cm/sec. Four moving-coil pen units were mounted in one permanent magnet. Each coil was centre-tapped and had a resistance of about 4 200Ω and a sensitivity of about 95V for full-scale deflexion. A resistance-capacitance oscillator controlled the frequency of the timing marks at 2c/s. The output of the oscillator was applied to the inner grid of a gas filled tetrode in the cathode circuit of which was an air-cored transformer. When the grid was sufficiently negative the tetrode oscillated and the high secondary voltage from the transformer was applied to fixed pens at each side and in the centre of the chart and a short line was produced for each cycle of the oscillator.

Speed Unit

A tacho-generator was required with a d.c. output and sufficiently free from commutator ripple to produce a smooth deceleration record. A rotary convertor with its low voltage brushes removed was more suitable than commercial tacho-generators. It was mounted on a small platform attached to the differential case of the car and driven from the propeller shaft by a V-belt. Ripple in the output was reduced by the capacitor C_1. The output voltage was applied to one pen-coil through resistors adjusted to give sensitivities of 40 mile/h for full-scale deflexion for town driving (R_1, R_4 and R_5) and 80 mile/h for full-scale deflexion for country driving (R_2, R_6 and R_7). The centre-tap of the pen-coil was connected to the chassis so that a push-pull input was available for the deceleration unit. When the apparatus was not recording R_3 was connected

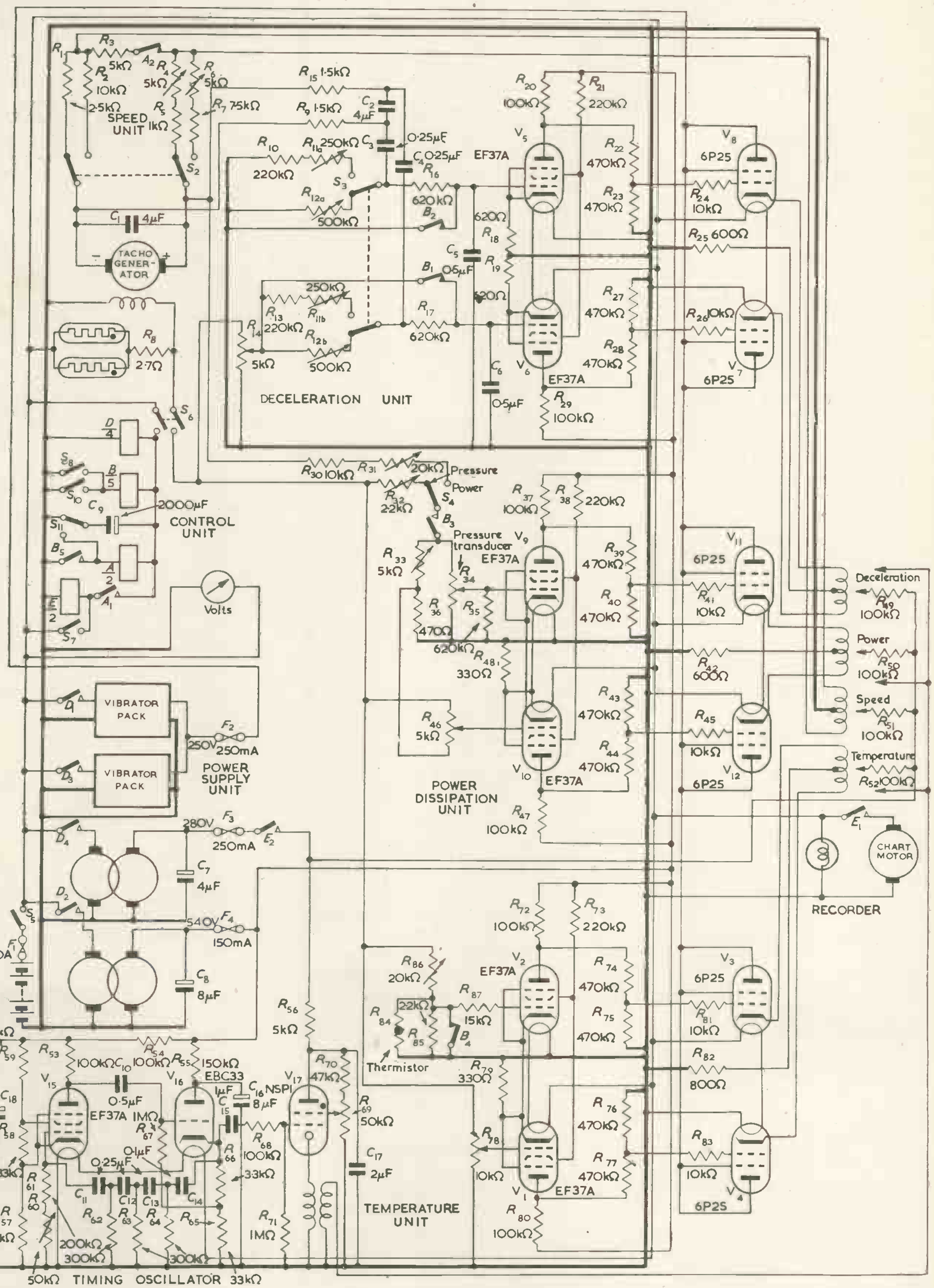

Fig. 3. The complete circuit

in parallel with the pen-coil to prevent the pen being overloaded by high speeds on the journey to or from a test. The pen was mechanically set to side zero.

Deceleration Unit

The input from the tacho-generator was partially smoothed by R_9, R_{15} and C_2. It was then differentiated by C_3, C_4, R_{10}, $R_{11a,b}$ and R_{13} or by C_3, C_4 and $R_{12a,b}$. The voltage representing deceleration was further smoothed by R_{16}, R_{17} and C_5, before being amplified. Screened wire was used for this part of the circuit to prevent spurious deceleration readings due to leakage to the grid connexions from the wire which carried the generator voltage. The time-constant of the actual differentiating circuit was chosen to give the required sensitivities, final adjustment being made by the double preset potentiometers $R_{11a,b}$ and $R_{12a,b}$. The values of the smoothing circuit components were then found by trial on the road.

The recorder pen coil required a symmetrical signal and the centre-tap had to be within 50V of the potential of the chassis. To give good linearity the amplifier was made push-pull throughout and consisted of a pentode voltage amplifying stage followed by a cathode-follower output stage. The voltage dividers R_{22}, R_{23} and R_{27}, R_{28} were required to provide suitable working voltages for the valves.

The pen was mechanically central and was set to side zero position by a voltage derived from the 6V stabilized supply which was added to the signal by series connexion (R_{14}). Relay contacts B_1 and B_2 short-circuited the signal except when the brakes were applied so that a datum line was recorded before and after each stop and so that accelerations did not force the pen backwards.

Power Dissipation and Pipe-line Pressure Unit

The pressure transducer was a Sangamo-Weston, type S122-4-28. It consisted of a Bourdon tube moving the sliding contact of a small potentiometer which was energized by the tacho-generator for power dissipation measurements and by the 6V stabilized supply for pressure measurements. When the pipe-line pressure was just sufficient to overcome the pull-off springs, a small fraction of the energizing voltage reached the grid of V_9 from the transducer but as no power was then being dissipated, it was necessary to apply a compensating voltage to the grid of V_{10}. This was derived through the resistance chain R_{33}, R_{36} and added to the zero-setting voltage produced across R_{46} and R_{36}. The compensation was not affected by the sensitivity adjustment (R_{31} or R_{32}) and the datum line which was recorded before and after each application represented either zero power dissipation or the pipe-line pressure necessary to overcome the pull-off springs. The correct value of this compensation was found by careful experiment.

The amplifier was similar to the deceleration amplifier except that the input was applied to one grid only and the two valves were coupled by the common cathode resistor R_{48}.

Temperature Unit

A small blind hole was drilled in the inner surface of one brake drum as closely to the braking path as possible. Into it was soldered a Standard Telephones & Cables type 'M' thermistor which consisted of a bead of semi-conductor enclosed in glass and mounted on a copper disk 5/32in in diameter. Connexions were made to two slip rings on the flat part of the drum and two brushes per slip ring were mounted on the back plate.

The resistance of the thermistor was 100kΩ at 20°C, falling to 100Ω at 300°C; the rate of change of resistance being greatest at low temperatures. The range required with the thermistor in the position described was 50°C to 250°C, and so a fixed resistor R_{85} was connected in parallel with the thermistor to produce an S-shaped calibration graph. (Fig. 4(d)). The datum line before and after an application was produced by contact B_4 short-circuiting the thermistor and therefore represented an infinitely high temperature.

The amplifier was identical with the one in the power dissipation unit.

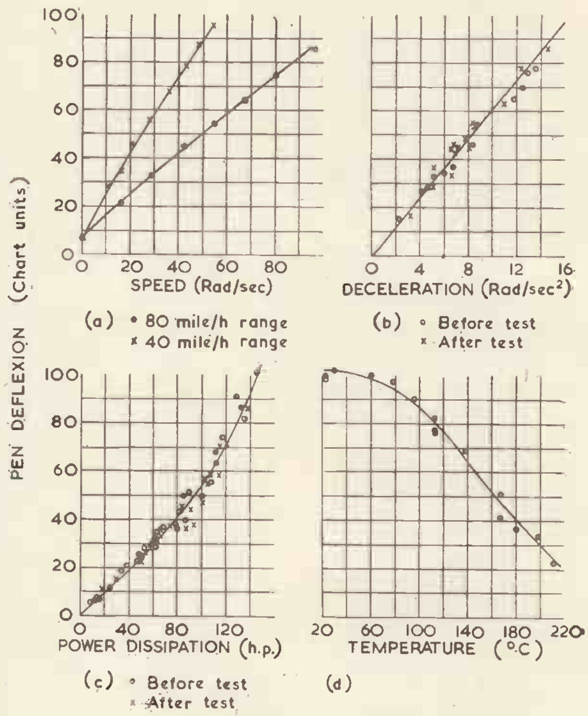

Fig. 4. Calibration graphs
(a) Speed. (b) Deceleration. (c) Power dissipation. (d) Temperature.

Calibration

Speed

The rear wheels of the car were jacked up and driven at various constant speeds. The surface speed of the V-belt was measured with a hand tachometer and the drum speed calculated from a knowledge of the back axle gear ratio and the radius of the V-belt surface over the propeller shaft pulley. Since there was no datum line on the chart, measurements were taken from the central row of timing dots, and a graph was plotted showing pen deflexion against drum speed in radians/sec. (Fig. 4(a)).

Deceleration

The car was driven at about 50 mile/h, put into neutral gear and the brakes applied, the driver keeping the deceleration as constant as possible. This was repeated at different decelerations and a typical record is shown in Fig. 1(a). When the chart was examined, a portion of each stop was selected in which the deceleration was constant. The initial and final speeds were determined from the chart and the interval measured by counting the timing marks. The mean deceleration in radians/sec² was thus calculated and plotted against the deceleration pen deflexion measured from the datum line at the beginning and end of each stop. (Fig. 4(b)).

POWER DISSIPATION

It is shown in the Appendix that the power dissipation *at the brakes* under certain conditions is:

$$(W/g)(1 + n)r^2(-\dot{\omega}' + d)\,\omega\text{ft.lb/sec} \quad \ldots\ldots (1)$$

$$= (W/550g)(1 + n)\,r^2\omega(\omega' + d)\text{ horse-power}$$

where W is the weight of the car (lb)

n is the kinetic energy of rotation of all revolving parts divided by the kinetic energy of translation of the whole vehicle at the same speed.

ω is the angular velocity of the wheels and drums (radians/sec).

$(-\omega')$ is the angular deceleration of the wheels and drums (radians/sec^2).

r is the effective radius of the wheels (ft).

$(-d)$ is the deceleration due to rolling and wind resistances at speed ω (radians/sec^2).

The conditions are:

(1) The road shall be flat so that there is no change in potential energy.

(2) Tyre slip shall be small and deceleration constant over those parts of the record which are measured.

(3) n and d must be determined with the car in the same gear as during the calibration. To keep these corrections small it is desirable that this should be neutral.

It can thus be seen that the unit can be calibrated by a comparison of the power dissipation record with the speed and deceleration records during controlled brake applications, although in normal driving the power dissipation cannot be calculated from the deceleration and speed records because of variations in road gradient, gear and tyre slip. The deceleration and power dissipation units were calibrated together from applications such as the one shown in Fig. 1(a) and the power dissipation was calculated from equation (1) for several speeds in each application. A typical calibration graph is shown in Fig. 4(c).

PIPE-LINE PRESSURE

The hydraulic braking system was connected to a dead-weight Bourdon-gauge tester, and the brake pedal tied down so that oil should not be forced back into the master cylinder. The pen deflexion was recorded on a few inches of chart at each of various pressures both in increasing and decreasing order.

TEMPERATURE

The drum with the thermistor attached was removed from the car, a calibrated thermocouple was clamped to it near the thermistor, and the whole was put in a thermostatically controlled oven with the thermistor well away from the walls. The calibration graph is shown in Fig. 4(d).

TIME

The frequency of the timing dots was measured before and after each test by counting the flashes of the neon tube V_{17} for one minute.

Accuracy

The apparatus was calibrated and used under similar conditions and the calibrations covered the full variation of battery voltage, brake drum temperature etc., and were repeated at frequent intervals, where possible before and after each test. Thus systematic errors were eliminated and the scatter of the points of the calibration curves was the random error of measurements on a single brake application. However, each test included scores of applications and so the accuracy of measurements referring to a whole test was considerably higher.

RECORDER

The response time of each unit was limited by the transducer and not by the recorder pen which was capable of recording frequencies up to 80c/s with a sensitivity within 6 per cent of the d.c. value. The deflexion/voltage characteristics were repeatable to within ± 1 per cent of full scale deflexion, i.e. reading accuracy, except that occasionally faults developed in the pen coils which altered the sensitivity without preventing them working altogether. Frequent calibration checks of the whole apparatus were therefore necessary. The recorder was checked with no electrical signals applied to the pen units and no spurious deflexions were produced by vibrations or accelerations during driving.

The full-scale deflexion of each pen was only 15mm and thousands of measurements were needed, so to improve accuracy and reduce eye strain, a chart reader was constructed. The chart was held stationary and a magnifying cursor 8in long was moved across it. A counter was coupled to the movement of the cursor and scaled in "chart units" which have been used throughout. 100 chart units = full-scale deflexion = 15·9mm. The long cursor was extremely useful for measuring mean deflexions, and it could be set to within half a unit on a trace which had vibrations of several units amplitude superimposed on the true deflexion. In general the reading accuracy was $\pm\frac{1}{2}$ chart unit.

There was also a movable non-magnifying cursor and a fixed scale for measurements along the length of the chart.

BATTERY VOLTAGE

The average decrease in battery voltage during a test was 3 per cent and a decrease of 16 per cent has been shown to cause a decrease of 5 per cent in the gain of an amplifier and 2 per cent in the 6V stabilized supply.

TIME

The frequency of the timing dots was checked before and after each test and was almost invariably within 1 per cent of its nominal value.

SPEED

The slip in the V-belt driving the tacho-generator was checked and found to be negligible. The terminal-potential difference/angular-speed graph was checked experimentally with the generator on load and found to be linear. Non-linearity would lead to errors in the deceleration and power dissipation measurements, although not in the speed measurement itself. Two calibration graphs, obtained as described in the previous section five months apart, were identical within 1 chart unit and the scatter of the points can be judged from Fig. 4(a).

DECELERATION

The response time was 0·6sec measured from the approximately exponential decay recorded when the deceleration ceased suddenly as the car came to rest. (See Fig. 1(a) and (e) C to D). The calibration accuracy depended on the accuracy of the speed unit and the timing marks and it was limited by the difficulty of maintaining a high constant deceleration for a measurable period, and by the presence of vibrational decelerations. The random errors may be assessed from the graph shown in Fig. 4(b), and it may also be seen that there was no measurable change in the sensitivity during the test.

PIPE-LINE PRESSURE

The calibration was linear and no hysteresis could be detected. Almost all the calibration points lay within 1·5 chart units of the line. The response time was less than 0·08sec.

POWER DISSIPATION

This measurement was necessarily less accurate than the speed and deceleration measurements since they were involved in its calibration, and a perfectly flat road surface was not obtainable. Errors also arose if the average shoe factor of the brakes varied, because of the assumption that braking torque was proportional to hydraulic pressure. It was therefore necessary to calibrate this unit before and after each test, and the principal use of the temperature unit was to ensure that during calibrations the brake drums reached temperatures as high as they had reached during the test. The effects of these errors may be seen in the calibration curve in Fig. 4(d). It may be noted that minor undulations in the road caused errors in the calculated values of power dissipation, but not in the pen deflexions, and they therefore increased the random scatter of the calibration points, but did not affect the instrument in use. The brake linings were specially selected for the constancy of their coefficient of friction.

Results

Some portions of typical records are shown in Fig. 1. Fig. 1(a) is a brake application for calibration of the deceleration and power dissipation units. The car was initially at rest and the four steps on the speed record marked A were produced by the release of the accelerator each time the gear was changed. The brakes were applied at B and the deceleration soon reached the constant value at which it remained until the car came to rest at C. The brakes were not released until D and the chart continued moving for three seconds more.

Fig. 1(b) and (c) are two extracts from one test in which the chart was moving continuously. (b) shows two applications in rapid succession on a downhill gradient and (c) shows one on an uphill gradient. The relative values of the power dissipation and the deceleration in the two records will be noted and in (c) the deceleration continued after the brakes were removed. Fig. 1(d) shows the brakes used to control the speed on a downhill gradient. There is a small but measurable power dissipation, and a deceleration followed by an acceleration with no overall change in speed. In hilly country much energy is dissipated in brake applications such as this because, although the power dissipation is small, the application time may be very long—three minutes has been recorded. At E the brake pedal has been pressed sufficiently to operate S_{10}, but too lightly to apply the brakes. The second application in Fig. 1(e) shows an entirely different use of brakes, in which the car was brought to a halt to wait for traffic at a road junction.

This article is concerned only with the apparatus. Results of the investigation into brake usage will be published in an automobile engineering journal. A preliminary account has already been published[5].

Acknowledgments

The author wishes to thank all those who have co-operated in the development of this apparatus and in particular Dr. N. Carpenter, Engineering Research Manager, Ferodo Limited. She is also grateful to Mr. D. K. Mackenzie, Manager of the Physics Laboratory, Ferodo Limited, for discussions, and to The Directors, Ferodo Limited, for permission to publish the article.

APPENDIX

THE POWER DISSIPATION AT THE BRAKES

Let the forward velocity of the car be $v$ft/sec and let the other symbols have the meanings previously listed.

There may be some slip between the tyres and road, in which case energy will be dissipated there when the car is braked. To a first approximation the slip speed may be assumed proportional to the tangential force between the tyres and road and hence to the deceleration. This approximation is justified because, provided none of the wheels is locked, the slip speed is only a small fraction of the wheel speed.

Then:

$$v = \omega r + kv' \quad \ldots\ldots\ldots\ldots (2)$$

where k is a constant.

Therefore, differentiating:

$$v' = \omega' r \quad \ldots\ldots\ldots\ldots (3)$$

provided that v' is constant.

The total kinetic energy of the car is:

$$E = (W/2g)\,(1 + n)\,v^2 \text{ft.lb} \quad \ldots\ldots\ldots (4)$$

Hence, provided that there is no change in potential energy, the total power dissipation is:

$$P = (-dE/dt) = (-W/g)\,(1 + n)\,v'v \text{ft.lb sec}^{-1} \quad \ldots (5)$$

This is equal to the total retarding force $\dfrac{W(1 + n)(-v')}{g}$ multiplied by the forward velocity v.

The retarding force may be separated into the braking force plus the wind and rolling resistances.

Thus:

$$-v' = r[(-\omega' + d) - d]$$

Substituting in (5):

$$P = (W/g)\,(1 + n)\,r\,[(-\omega' + d) - d]\,v \quad \ldots\ldots (6)$$

The forward velocity may be separated into the velocity of the tyre surfaces plus the slip speed. This is expressed in equation (2). Substituting in equation (6):

$$\begin{aligned} P &= (W/g)\,(1 + n)\,r\,[(-\omega' + d) - d]\,[\omega r + kv] \\ &= (W/g)\,(1 + n)\,r\,[(-\omega' + d)\omega r - d\,(\omega r + kv') + (-\omega' + d)\,kv'] \end{aligned}$$

The first term in the square brackets represents the power dissipated by the brakes, the second represents the power dissipated against wind and rolling resistances and the last term represents power dissipated at the tyres because of slip caused by braking.

Thus the power dissipated at the brakes is:

$$(W/g)\,(1 + n)\,r^2\,(-\omega' + d)\omega$$

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A Flutter Computer with Low Gain Amplifiers

By K. E. Wood* and I. V. Hansford†, B.Sc.

An analogue computer is described which solves aircraft flutter problems in two degrees of freedom. Low gain d.c. amplifiers are used rather than conventional high gain amplifiers, the principle of the computer being to shift all the poles of the solution to the left in the complex p plane. This method is especially useful in low-stiffness problems, with little feedback, where drift would make a solution difficult to obtain without the use of elaborate drift corrected amplifiers.

THE flutter equation for an aircraft structure, subject to aerodynamic and mechanical loading, for one degree of freedom, may be written in the form:

$$ax'' + (bv + d)\,x' + (cv^2 + e)\,x = 0$$

where a is the mechanical inertia coefficient;

b and d are the aerodynamic and mechanical damping coefficients;

c and e are the aerodynamic and mechanical stiffness coefficients;

v is the aircraft speed in suitable units;

x is the displacement.

In considering two or more degrees of freedom the flutter equations are not independent, due to both aerodynamic and mechanical coupling.

The equations to be solved by this particular computer may then be written as:

$$a_{11}x_1'' + (b_{11}v + d_{11})x_1' + (c_{11}v^2 + e_{11})x_1 = a_{12}x_2'' + (b_{12}v + d_{12})x_2' + (c_{12}v^2 + e_{12})x_2$$

$$a_{21}x_1'' + (b_{21}v + d_{21})x_1') + (c_{21}v^2 + e_{21})x_1 = a_{22}x_2'' + (b_{22}v + d_{22})x_2' + (c_{22}v^2 + e_{22})x_2$$

Rewriting the first row for example gives:

$$x_1'' = a_{12}/a_{11}\,x_2'' + (b_{12}/a_{11}\,v + d_{12}/a_{11})\,x_2' + (c_{12}/a_{11}\,v^2 + e_{12}/a_{11})\,x_2 - (b_{11}/a_{11}\,v + d_{11}/a_{11})\,x_1' - (c_{11}/a_{11}\,v^2 + e_{11}/a_{11})\,x_1 \quad \ldots\ldots\ldots (1)$$

$$x_1'' = \Sigma\,(\text{terms in } x_2'', x_2', x_2, x_1', x_1).$$

A block diagram of the computer to solve equation (1) is shown in Fig. 1.

The block diagram consists of an adding circuit into which are fed the above terms in x_2'', x_2', x_2, x_1' x_1, each multiplied by its corresponding coefficient ratio, which has been scaled to be less than unity. The output of this adding circuit is then x_1'', and if this is integrated twice outputs are obtained for x_1' and x_1. If a similar arrangement is carried out for the second row, solving for the principal term x_2'', then outputs are also obtained for x_2' and x_2. By completing feedback paths from the six available outputs, via phase invertors where necessary to give the correct sign, and coefficient potentiometers to the appropriate adding circuits, the complete loop will be closed, and the computer will simulate the above equations.

The Pole Shifting Method of Solution[1]

Most conventional analogue computers use very high gain d.c. amplifiers, which may or not be drift corrected, and consequently may be subject to serious drift troubles and parasitic oscillations. If the drift in an amplifier becomes appreciable then an error voltage is added to the coefficient quantities, causing a loss in accuracy. Generally in problems involving a large amount of feedback this is of little consequence, but where the feedback is small, in low-stiffness problems, the pole shifting method of solution shows a distinct advantage.

The computer, which is a closed loop, may be considered as a network with a transfer function $G(p)$. For a step function input the output at any point $v_o(p)$ may be expressed in general as:

$$v_o(p) = \sum \frac{A_n}{p + \alpha_n}$$

where A_n and α_n are constants.

Multiple poles may be included by letting:

$$\alpha_n \rightarrow \alpha_{n+1} \rightarrow \alpha_{n+2} \ldots$$

$$A_n \rightarrow A_{n+1} \rightarrow A_{n+2} \ldots$$

$$\therefore v_o(t) = \sum A_n \exp(-\alpha_n t)$$

Now suppose that throughout the whole system p is replaced by $(p + a)$, including modification of the input function, then:

$$v_o(p) = \sum \frac{A_n}{p + a + \alpha_n}$$

$$\therefore v_o(t) = \sum A_n \exp(-\alpha_n t)\exp(-at)$$

i.e. the original solution is obtained modified by the exponential term e^{-at}.

This result has been obtained by replacing p by $(p + a)$ in all the transfer functions, which serves to move all the poles of the solution a distance a further to the left in the complex p plane.

The transfer function of the integrating circuit shown in Fig. 2 is $1/pCR_1$, provided that the inner loop gain A of the amplifier is sufficiently high. If a resistor R_2 is put in parallel with the feedback capacitor C the transfer function is given by $1/CR_1 \,.\, 1/(p + (1/CR_2))$, which is of the form $k/(p + a)$, where $a = (1/CR_2)$ and $k = (1/CR_1)$.

If then all the integrators used, and the input function, have transfer functions of the form $k/(p + a)$, then the oscillatory solution obtained from the computer will be that given by using perfect integrators, and a step function input, multiplied by e^{-at}.

By having R_2 in parallel with C the d.c. gain is reduced to R_2/R_1, and the drift error will also be reduced in proportion.

The pole shifting method of solution has another advantage in that it may give a measure of the stability or instability of a problem. If a problem has a stable solution $F(p)$ then all the poles of the solution must be situated in the left-hand side of the complex p plane. If now in the

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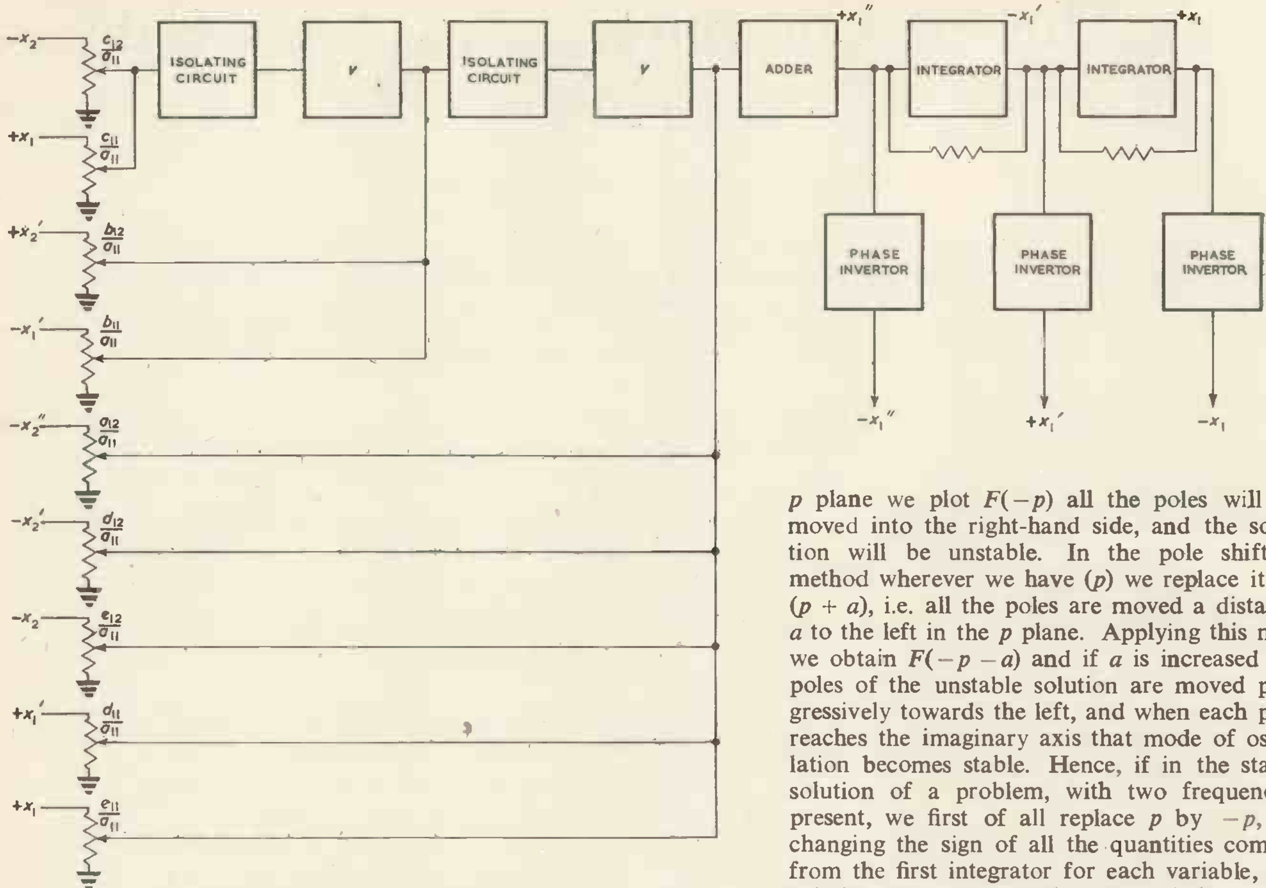


Fig. 1. Arrangement of computer to solve equation (1)

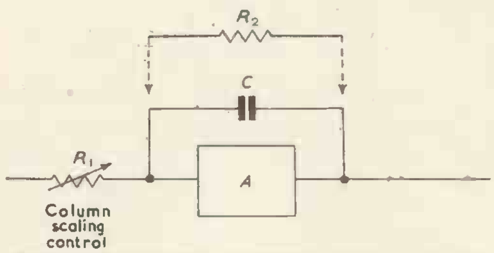
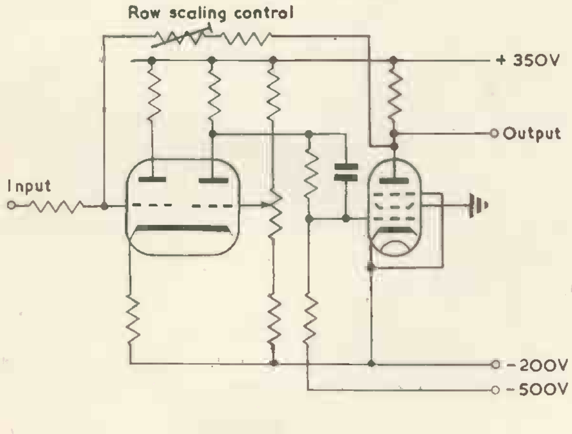


Fig. 2. Integrating circuit

Fig. 3. Adding amplifier



p plane we plot $F(-p)$ all the poles will be moved into the right-hand side, and the solution will be unstable. In the pole shifting method wherever we have (p) we replace it by ($p + a$), i.e. all the poles are moved a distance a to the left in the p plane. Applying this now we obtain $F(-p-a)$ and if a is increased the poles of the unstable solution are moved progressively towards the left, and when each pole reaches the imaginary axis that mode of oscillation becomes stable. Hence, if in the stable solution of a problem, with two frequencies present, we first of all replace p by $-p$, by changing the sign of all the quantities coming from the first integrator for each variable, the solution becomes unstable. Now a is increased, moving the poles to the left, and for a certain value of a one mode becomes stable and only one frequency is left. a is increased further till this mode also becomes stable. The two values of a then give a measure of the stability of the solution.

Similarly, if a problem is initially unstable, then on increasing a till stability is reached, a measure is obtained of the instability of the problem.

Description of the Units

The adding amplifier consists of a conventional d.c. amplifier with an inner loop gain of about 1 100. The output voltage is initially adjusted to be at earth potential by the balancing control. The circuit is shown in Fig. 3. The main adding amplifiers have 6 inputs each. A forcing function can be applied to either, corresponding to a disturbance being applied to either equation. To provide the row scaling facility the overall gain for each adding amplifier may be independently varied from $\frac{1}{2}$ to 3 in steps of $\frac{1}{2}$ by switching in resistors in the feedback path.

The phase invertor is a d.c. amplifier made up of a single pentode stage with a cathode-follower output. It has an inner loop gain of about 150, with the outer loop gain set to unity.

An isolating amplifier, with unity gain has to be inserted between the velocity coefficient potentiometers to prevent loading. The circuit is similar to that of the phase invertor except that a small indicator neon is used to provide a d.c. shift, thus increasing the inner loop gain.

As previously explained, it is necessary to use integrator circuits with transfer functions of the form $k/(p + a)$ where the d.c. gain is k/a. The transfer functions were chosen to be $120/(p + (1/5))$, i.e. the d.c. gain is 600. This is obtained by having a 5MΩ resistor in parallel with a 1μF feedback capacitor, and having an 8·2kΩ input resistor. The inner loop gain of the d.c. amplifiers used in the integrator circuits is 30 000.

The transfer function of the input function must be changed from that of a pure step $1/p$, to $1/(p + a)$, i.e. to that of a step with an exponential decay. A 1μF capacitor is charged from a potentiometer chain across the h.t. supply, serving as an input amplitude control, and discharged through a 5MΩ resistor into one of the adder circuits. To prevent loading of this resistor by the low input impedance of the adder circuit a cathode-follower is inserted in between, and this allows the input function to be displayed on a c.r.o. The slight loss of gain in the cathode-follower is of no consequence, as in this computer no meaning is attached to the amplitude of the input function.

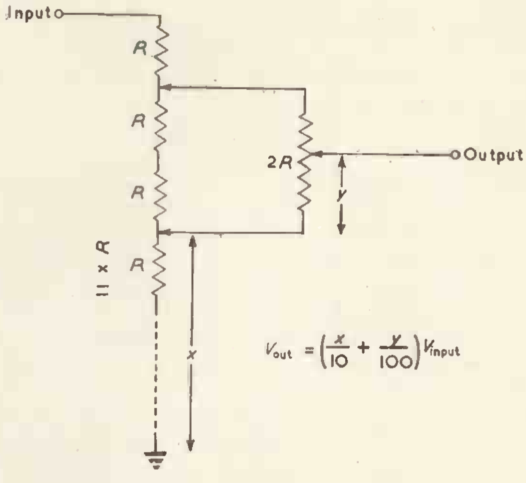


Fig. 4. Method of setting ratios

The Coefficient and Velocity Potentiometers

In order to set up a problem on the computer all the coefficients of each row are initially divided by the coefficient of the principal term, and then various operations may be performed to make these ratios less than unity. It was decided that it should be possible to set up the ratios so obtained accurately to three decimal places. The method used is based on the Kelvin-Varley slide-wire[2], and is shown in Fig. 4. One of the main advantages of the system is that it presents a constant input resistance.

For the coefficient potentiometers, the system consists of one decade switch followed by a wirewound potentiometer, together with a switch for selecting the correct sign of the coefficient. Provided the potentiometer can be set to 1 part in 100, the value of the coefficient can be set to 1 part in 1 000. To obtain the desired accuracy a bridge method was devised. For this an accurate potentiometer was built, consisting of three decade switches working on the above principle, and the coefficient value set up on this. The corresponding coefficient potentiometer system is then balanced against the accurate potentiometer by feeding the same voltage into both systems, and plugging a wander plug into a socket corresponding to the coefficient to be set up. In this way each coefficient can be easily and quickly set up accurately to three decimal places.

The method could be extended using further decade switches, instead of the potentiometer, to obtain any desired accuracy, but this would be very expensive if a large computer were involved. The fact that the setting up time by this method may be longer than by using three decade switches for each coefficient is of minor significance in comparison with the saving in cost, the coefficient control resistances becoming therefore less accurate and cheaper.

A decade system is used to obtain the value of the velocity coefficient v, except that a further 10-way wafer switch is used instead of the potentiometer. Hence v may be accurately varied from 0 to 1 in steps of 0·01. The two sets of switches for v are mechanically ganged together. If the scaling for v is 1 000ft/sec then the speed of the aircraft may be changed in steps of 10ft/sec. For certain problems these steps were found to be too coarse and a further subdivision into steps of 0·001 is contemplated.

The Method of Setting up the Problem on the Computer

Often the problem to be solved is not necessarily in the best state for solution on the machine. To achieve this scaling is carried out, and it must be such that the maximum value of any coefficient ratios is never greater than unity.

Scaling of the Whole Problem in Time

Consider the equation $x'' + x' + \omega^2 x = 0$, where the natural frequency of oscillation is ω.

In order to obtain a satisfactory display of the solution, i.e. to view a sufficient number of cycles to determine the true behaviour of the solution, it may be necessary to speed up the whole problem in time. This is done by having the transfer functions of the integrators to be of the form $k/(p + a)$. With these transfer functions the above equation may be written as:

$$x'' + kx' + k^2\omega^2 x = 0$$

i.e. the frequency of oscillation is now $k\omega$.

Having transfer functions of $k/(p + a)$ in all the integrators speeds up the whole problem k times, and does not necessitate any alteration in the coefficient values of the variables. The frequency cannot be increased indiscriminately due to stray capacitances. Any slight errors in the value for k will effect the flutter frequency more than the flutter speed.

Row Scaling

Referring to equation (1), in each row we solve for the principal term ax'', and so initially all the coefficients in that row are divided by a. Often a will be the largest coefficient in that row, and all the ratios so obtained will then be less than unity. However, this may not always be the case, and other means must be found for setting up these ratios on the computer.

In row scaling all the coefficients in a row, except the principal one, are divided by any suitable number which will reduce all the coefficient ratios to less than unity. This necessitates placing that amount of gain in the corresponding adder circuit. With the adders used only gains from $\frac{1}{2}$ to 3, in steps of $\frac{1}{2}$, may be used.

The above method for reducing the coefficient ratios may not be sufficient if the ratios are large, and then column scaling must be used.

Column Scaling

For the equation $x'' + x' + \omega x^2 = 0$ the natural frequency of oscillation is ω radians per second.

Suppose now the coefficient of x'' is divided by K^2,
that of x' is divided by K,
and that of x is divided by unity,

then the equation becomes:

$$(1/K^2)x'' + (1/K)x' + \omega^2 x = 0$$

where the frequency of oscillation is now $K\omega$, i.e. the time scale has been modified $1/K$ times natural time.

In the case of this computer, let the two expressions associated with one variable, i.e. a column, be scaled in the above manner by a constant K_1, and let the second column be scaled by K_2, where $K_1 \neq K_2$. Since the two time scales must be equal the transfer functions of the integrators will have to be altered accordingly to allow for $K_1 \neq K_2$.

The transfer function of the integrator circuits is chosen to be $120/(p + (1/5))$, where the d.c. gain is 600, and so the column scaled by the factor K_1 will have a time scale of $120K_1$ times natural time, and the column scaled by K_2 a time scale of $120K_2$ times natural time. The two time scales must be made equal, and this is done by reducing the d.c. gain of the integrators corresponding to the variable scaled by K_1 by the factor K_1, and similarly the d.c. gains of the other integrators corresponding to the variable scaled by K_2 are reduced by the factor K_2.

This reduction in the d.c. gain of the integrators is carried out on the computer by increasing the series input resistor R_1 of the circuit. The input resistors to each of the two integrators corresponding to one variable are ganged together, and consist of five 8·2kΩ resistors which may be switched in series. In this way scaling factors from 1 to 5 may be used.

Presentation of the Solution

An analogue divider[3] was developed to divide the output from the computer by e^{-at} to obtain the true solution, operating on a pulse time basis. A stable pulse repetition frequency is generated and triggers a Miller valve. This p.r.f. must have a stability better than the required divider accuracy, as the unit operates on a pulse duty cycle basis. The grid returning potential of the Miller valve is obtained from the waveform $V = V_0e^{-at}$, where V is the potential by which the solution is to be divided. This produces a pulse at the screen, the width of which is inversely proportional to this exponential waveform. The pulse amplitude is then limited to the instantaneous amplitude of the dividend. Hence an average reading instrument will give an output which is proportional to the product of the pulse duty cycle and the pulse amplitude. As the duty cycle is proportional to e^{+at}, the pulse amplitude, which is proportional to the computer output, is modified, and the true solution is obtained.

Since it is desired to view the corrected output for as long a time as possible, a wide range of multiplication is necessary. Thus for $a = (1/5)$, for 10sec viewing time a multiplying range of 7 is required, and for 20sec viewing time the range must be 54. For a 1 per cent accuracy in the run-down time the returning potential must have a minimum value of 100V. Hence for a multiplying range of 30 a 3kV supply is necessary. With a 3kV supply the limitation in accuracy now arises in the derivation of the waveform e^{-at}. This is because its source can no longer be isolated from its point of application, and loading on the source will occur. Assuming, however, that the potential at the grid of the run-down valve does not alter by more than 5V from earth, a minimum returning potential of 500V will ensure an accuracy better than 1 per cent. This will give a multiplying range of 6, and consequently a viewing time of 9sec. By using a higher gain than that provided by a single stage the input grid excursions may be reduced, with a consequent increase in accuracy in the waveform $V = V_0e^{-at}$ and in the linearity of the run-down.

A c.r.o. multiplier was also investigated in which the e.h.t. supply followed the law e^{-at}. Since on a c.r.t. the deflexion sensitivity is inversely proportional to the e.h.t. voltage on the final anode, by having an e.h.t. voltage following this law the waveform being displayed is effectively multiplied by e^{at}. Due to the variation in deflexion sensitivity as the e.h.t. supply changes a linear time-base cannot be used directly, but must be modified to be of the form te^{-at}. A maximum multiplication range of only 3, giving a viewing time of about 5sec may be obtained by this method.

The solution of an aircraft flutter problem has a natural frequency lying approximately between 1 and 50c/s. For this particular type of equation if the two degrees of freedom have similar flutter frequencies beats will be set up, which may be of a very low frequency, and a considerable period of time may be required to understand the true behaviour of the solution. It was found that neither the analogue divider, nor the c.r.o. multiplier, gave a viewing time which would be sufficient to interpret the behaviour of the solution near the flutter speed.

A simple alternative to the above methods of presentation was found by using a standard double beam Cossor c.r.o. with a long persistence tube. On one trace the exponentially decaying input function is displayed, and on the other the solution as given by the computer, which decays at an unknown rate. When the rate of decay of the solution is the same as that of the reference exponential, the true solution will just be maintaining oscillation. This method has the advantage that a complicated e.h.t. voltage and time-base are unnecessary, and that the viewing time may be considerably extended.

The initial amplitudes of the two display beams should be adjusted to be equal for this method of comparing the rates of decay to be valid. However, if this is not carried out the error introduced is not large, as initially the solution will contain transients, which are of no interest, and it is only after a certain time when the amplitudes will be more or less equal that the rates of decay need be compared.

Operation of the Computer

The output voltage level of all the various units in the computer must initially be set at earth potential. This figure is chosen so that the scaled values of the coefficients are those actually given on the coefficient potentiometers with no additional d.c. term involved. Switches allow the inputs to the units to be earthed in turn, and the outputs are then set at earth potential, the output voltage being shown on a centre reading instrument. It was found that the d.c. levels in the computer need only be set up every few weeks. Provision is made for injecting the input function into either equation, and any one of the six available outputs x_1'', x_1', x_1, x_2'', x_2', x_2, may be shown on the display unit. A decade counting system, which is gated for 10sec counts the number of cycles of the computer frequency during that time, and on applying a scaling factor the frequency of the original problem may be obtained. It is also possible to vary $1/a$ from $\frac{1}{2}$ to 5, in nine equal steps, by having the resistors R_2, which are in parallel with the feedback capacitors on the integrators, on a ganged switch. R_2 may be removed altogether, when the computer is operated as if with perfect integrators and a step function input.

Conclusion

This computer was built as a prototype model for a larger computer to prove the principle of the pole shifting method for the solution of linear differential equations, and could be extended for the solutions of non-linear differential equations. The method of presentation enabled the true behaviour of the solution from the computer to be interpreted to an accuracy of about 2 per cent.

The larger computer will handle a flutter problem involving nine degrees of freedom, containing 500 separate coefficients. The computer may be divided into two positions solving six and three degrees of freedom. Altitude may be simulated, and inertia terms containing both mechanical and aerodynamic quantities inserted. A variety of double input forcing functions may be simultaneously applied as separately calibrated quantities into each degree of freedom. High gain drift corrected amplifiers will be used throughout, with the pole shifting method available for use in low-stiffness problems.

Acknowledgments

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A Radio Controlled Tractor

A radio-controlled tractor has recently been demonstrated by the Tractor Division of the Ford Motor Company Ltd. The tractor used for the demonstration was a standard Fordson Major. Although intended only as a novel method of sales presentation and demonstration the Ford Company consider that it may have a future in mechanized farming and possibly also in the transport of dangerous materials, as for example, fissionable products at atomic plants.

The radio control is effected by means of a simple transmitter working at a frequency of 27·12Mc/s and provides for six separate but not simultaneous channels by means of audio-frequency modulation of the carrier, the receiver having tuned reed output relays for reception of the individual signals which in turn operate secondary relays to provide excitation from the 12V tractor battery for the solenoids connected to the tractor controls.

The transmitter is battery operated and has alternative aerial arrangements, one on the case itself and means for operating from an alternative fixed to a vehicle or platform. The receiver is also battery operated, enclosed in a robust case fixed to the side of the tractor in close proximity to the telescopic whip aerial. The operator's control box is separate from, but attached to, the transmitter by a flexible cable up to 8ft long. These controls are arranged to provide the following movements:—

1. Steering left. 2. Steering right. 3. Clutch out. 4. Implements raise. 5. Implements lower. 6. Engine stop.

Mounted on the tractor and associated with the receiver is a bank of coloured lights arranged to indicate the electrical functions 1 to 6 and provide a continuous check on the operation of the radio transmitter and receiving gear. It should be noted in this respect that function 6 is associated also with the patented safety device which comes into operation in the event of any failure occurring either in the transmitter or the receiver and the associated electrical gear, including batteries.

The radio equipment was supplied by Radio and Electronic Products of Richmond.

The radio-controlled tractor, the transmitter and control box

COLOUR TELEVISION TESTS

The BBC have recently begun a series of experimental transmissions in colour from its transmitting station at Alexandra Palace. These transmissions will in no sense be a public service, and are not an indication of any early start with colour television. They will consist mainly of still patterns, demonstration films in colour, and simple studio shots, designed to provide technical information. The transmissions will have no interest as programmes; they will take place outside normal programme hours and are likely to continue for some months. They are being carried out in agreement with the Television Advisory Committee, who have been asked by the Postmaster General to report on the whole field of colour television.

Preliminary work on a colour television system which might prove suitable for a public service in this country has been in progress in the BBC's Research Laboratories for some years. This work has been done in close collaboration with the Radio Industry, and it has now reached the stage where it is desirable to study the practical problems of transmission and reception such as would be encountered in a public service.

The system to be used in these first experiments is based on the N.T.S.C. system adopted for public service in the U.S.A., it has been adapted to make it suitable for British standards. Other systems may be tested later on. One of the problems is whether a truly compatible system should be the final objective, but it is first necessary to establish whether such a system is possible using the adapted N.T.S.C. standard. An important feature of the experiments will be to examine whether, if this system is used, there is any appreciable degradation of the picture as received on present-day black and white receivers.

The studio colour camera and the associated equipment to be used in these experiments has been supplied by Marconi's Wireless Telegraph Company Limited. The film and slide scanning equipment has been developed and built by the BBC Research Department.

Signal Analysis and Audio Characteristics of Pulse-Slope Modulation

By Jajneswar Das*

The Fourier analysis of different processes of p.s.m. and the experimental determination of its audio-frequency characteristics have been made. The maximum modulation possible is seen to be of the order of 80 to 90 per cent for general cases. The process of demodulation requires the use of a differentiator and a "boxcar" pulse-lengthener circuit. Experimental verification with a p.s.m. system of p.r.f. of 10kc/s and pulse duration of 10μsec shows that the modulation is linear over an input audio volume range of 35dB. The audio gain of the complete system is fairly constant up to 3kc/s with a variation of ±5dB only. The average distortion is within 5 per cent for a volume variation of 30dB at the input.

PULSE slope modulation[1,2], a new method of modulating video pulses, was reported sometime ago. It is known that on integration of a step function, the slope of the ideally-integrated function (known as Ramp function) is dependent on the amplitude of the step function and the constants of the integrating circuit. Further, on differentiation of a linear ramp function, the magnitude of the step function, thus generated, is dependent on the slope of the original function. On the basis of this, slope modulation has been realized by using a rectangular pulse as a gating pulse to a constant current charging circuit. The modulating signal voltage varies the charging current in the circuit and hence the slope of the output pulse. To produce a linear rise of the leading edge of the output pulse, the value of RC of the charging circuit has to be great compared to the duration of the gating pulse. Fig. 1(a) shows the block schematic of the slope-modulated pulse transmitter incorporating the sampling-pulse generator and the modulator. Demodulation is carried out by differentiating the slope-modulated pulses, thus giving rise to amplitude-modulated sharp pulses. These are passed through a "Boxcar"[3] pulse lengthener circuit and the necessary audio filter and amplifier. Fig. 1(b) shows the block schematic of a p.s.m. receiver circuit. Typical waveforms at different stages are also shown in the figure.

It is considered to be of interest to analyse the different processes involved in complete transmission and reception of p.s.m. signals, and to show how the overall characteristics compare with other similar pulse modulation systems. The general method of Fourier analysis[4] of waves derived by sampling a signal at discrete intervals has been followed. Spectrum analyses of the modulated pulses, the differentiated pulses and the complete demodulated output have been made. The overall amplitude characteristic, the linearity of modulation and the percentage distortion have been experimentally determined for a laboratory model of the p.s.m. system. The characteristics were found to compare very favourably with other similar pulse modulation systems.

Modulation

The Fourier series expansion of the trapazoidal pulses can be written as (Fig. 2):

$$f(t) = (a_o/2) + \sum_{n=1}^{\infty} (a_n \cos \omega nt + b_n \sin \omega nt)$$

where:

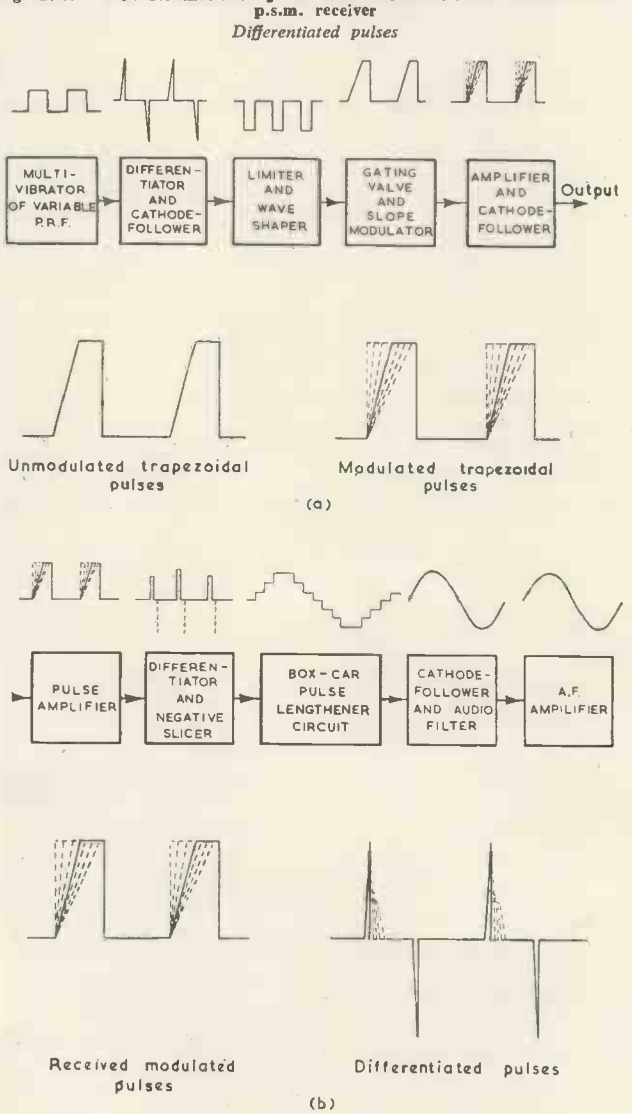
$$a_o/2 = (A/T_o)\ (d - (t_r/2))$$

$$= (A\omega/2\pi)\ (d - (t_r/2))$$

$$a_n = (2A/T_o) \int f(t) \cos \omega nt\ .\ dt$$

$$= (2A/T_o)\int_0^{t_r} (t/t_r)\ .\ \cos \omega nt\ dt + (2A/T_o)\int_{t_r}^{d} \cos \omega nt\ dt$$

Fig. 1(a). Block schematic of p.s.m. transmitter. (b) Block schematic of p.s.m. receiver



* *Indian Institute of Technology, Kharagpur.*

$$= (2A/T_o)\left(\frac{\cos \omega n t_r}{t_r\omega^2 n^2} - \frac{1}{t_r\omega^2 n^2} + \frac{\sin \omega n d}{\omega n}\right)$$

$$b_n = (2A/T_o)\int f(t)\sin \omega nt \,.\, dt$$

$$= (2A/T_o)\int_0^{t_r} (t/t_r)\,.\,\sin \omega nt\,.\,dt + (2A/T_o)\int_{t_r}^{d} \sin \omega nt\,.\,dt$$

$$= (2A/T_o)\left(\frac{\sin \omega n t_r}{t_r\omega^2 n^2} - \frac{\cos \omega n d}{\omega n}\right)$$

Hence:

$$f(t) = (A/T_o)(d - t_r/2)) + (2A/T_o)$$

$$\sum_{n=1}^{\infty}\left[\frac{\sin \omega n(d-t)}{\omega n} + \frac{\cos \omega n(t - t_r)}{t_r\omega^2 n^2} - \frac{\cos \omega nt}{t_r\omega^2 n^2}\right] \quad \ldots\ (1)$$

It has been shown[1] that on modulation by a voltage of the form $E \sin \phi$, the slope of the leading edge of the pulses vary according as:

$$\tan\theta = B + KE\sin\phi$$
$$= B(1 + KE/B)\ \sin\phi$$
$$= (A/t_r) \quad \ldots\ (2)$$

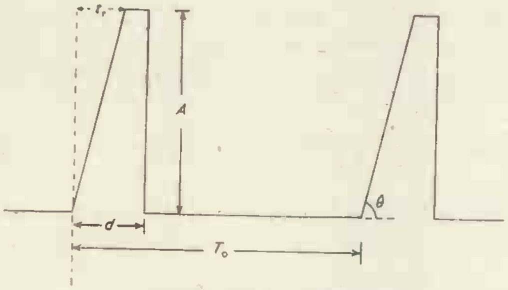


Fig. 2. Nature of trapezoidal pulses

t_r = rise time, d = pulse width, T_o = time period, tan θ = slope

where B and K are constants and E is the amplitude of the modulating voltage.

If $\tan\theta_{max}$ and $\tan\theta_{min}$ give the maximum and minimum slopes of the leading edge of the pulses during modulation process, the modulation-index can be defined as:

$$m = \frac{\tan\theta_{max} - \tan\theta_{min}}{\tan\theta_{max} + \tan\theta_{min}}$$

$$= \frac{1/t_{r(min)} - 1/t_{r(max)}}{1/t_{r(min)} + 1/t_{r(max)}}$$

$$= \frac{t_{r(max)} - t_{r(min)}}{t_{r(max)} + t_{r(min)}} \quad \ldots\ (3)$$

Further:

$$m = \frac{B + KE - (B - KE)}{B + KE + B - KE}$$

$$= KE/B \quad \ldots\ (4)$$

If t_{ro} corresponds to $\tan\theta_{(mean)}$, the slope without modution, then:

$$t_{ro} = (A/\tan\theta_{(mean)})$$

$$= \frac{A}{1/2\ (B + KE + B - KE)}$$

$$= A/B \quad \ldots\ (5)$$

We can now express the modulation in terms of changes in the risetime of the pulses, which is given by:

$$t_r = (A/\tan\theta)$$

$$= \frac{A}{B(1 + m\sin\phi)}$$

$$= \frac{t_{ro}}{(1 + m\sin\phi)} \quad \ldots\ (6)$$

Putting this new value of t_r in terms of the mean rise-time and the modulating voltage in equation (1), we have the Fourier series—expansion of the slope-modulated pulses given by:

$$F(t) = (Ad/T_o) + \frac{At_{ro}}{2T_o(1 + m\sin\phi)} + \frac{2A}{T_o}\sum_{n=1}^{\infty}\left[\frac{\sin \omega n(d-t)}{\omega n}\right]$$

$$+ \frac{2A}{T_o}\sum_{n=1}^{\infty}\frac{1 + m\sin\phi}{t_{ro}\omega^2 n^2}\left[\cos \omega n\left\{t - \frac{t_{ro}}{(1 + m\sin\phi)}\right\} - \cos\omega nt\right] \quad \ldots\ (7)$$

Power Relations

The power relations in p.s.m. are slightly involved due to the slope of the transmitted pulses and do not give the simple duty ratio as in the case of rectangular pulses.

In Fig. 3 the equation of the sloping edge of the pulse, AB, can be written as:

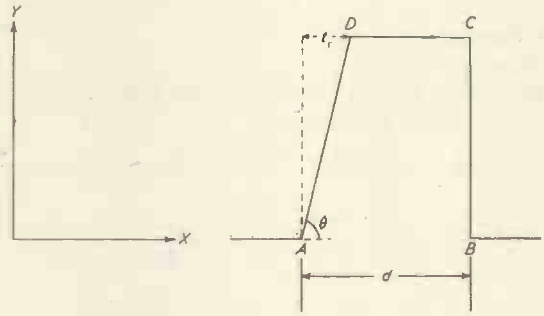


Fig. 3. A trapezoidal pulse

$$y = x\tan\theta$$
$$= xA/t_r$$

And the power at each point of AB is equal to y^2/R, where R is the load resistance. Average power in AB is equal to:

$$1/Rt_r\int_0^{t_r} y^2 dx = (A^2/3R)$$

Hence the average power in the pulse $ABCD$ (i.e. from $x = 0$ to $x = d$) is given by:

$$P_{av} = (1/d)\left[\frac{A^2 t_r}{3R} + \frac{A^2(d - t_r)}{R}\right]$$

$$= (A^2/Rd)(d - (2t_r/3)) \quad \ldots\ (8)$$

Total energy in one pulse:

$$P_{to} = (A^2/R)\ (d - (2t_r/3))$$

If the repetition frequency of the pulses is f_o, then the power transmitted per second:

$$P_o = (A^2 f_o/R)\ (d - (2t_r/3)) \quad \ldots\ (9)$$

Since the duty ratio in a pulse communication system may be defined as the ratio of the average power to the peak power and in case of non-rectangular pulses, this is computed from the area under the pulse, the duty-ratio in this case is equal to:

$$(A^2/T_oR).(d - (2t_r/3))$$

Equation (9) gives the power in unmodulated pulses. If the pulses are slope modulated, the value of t_r will be varying as:

$$t_r = \frac{t_{ro}}{(1 + m\sin 2\pi f_m t)}$$

(where f_m = modulating frequency).

but t_r will have discrete values depending on the phase relationship between the sampling pulses and the modulating voltage. If $t_{r1}, t_{r2}, t_{r3} \ldots$ etc. are the discrete values of the rise time at the points of sampling, then the transmitted power per second of the modulated pulses is equal to:

$$\frac{A^2 f_o d}{R} - (2/3)\cdot\frac{A^2 f_m}{R}\sum_{n=1}^{n=f_o/f_m}(t_{r1} + t_{r2} + t_{r3} + \ldots + t_{rn})$$

$$= \frac{A^2 f_o d}{R} - \frac{2A^2 f_m}{3R}\sum_{n=1}^{f_o/f_m}\left[\frac{t_{ro}}{1 + m\sin(2\pi f_m n/f_o)}\right] \ldots \quad (10)$$

$$\simeq \frac{A^2 f_o d}{R} - \frac{2A^2 f_m}{3R}\int_{n=1}^{f_o/f_m}\left[\frac{t_{ro}}{1 + m\sin(2\pi f_m n/f_o)}\right]\cdot dn;$$

if $f_o \gg f_m$.

Demodulation

(a) DIFFERENTIATION

The Fourier series expansion of the differentiated signal is given by (from equation (7)):

$$F'(t) = (d/dt)\ F(t) = (2A/T_o)\sum_{n=1}^{\infty} -\cos\omega n(t-d)$$

$$+ (2A/T_o)\sum_{n=1}^{\infty}\frac{(1+m\sin\phi)}{t_{ro}\omega n}\left[\sin\omega n\left\{\frac{t_{ro}}{(1+m\sin\phi)} - t\right\} + \sin\omega nt\right] \ldots\ldots \quad (11)$$

In evaluating the above equation, the assumption has been made that $t_r = t_{ro}/(1 + m\sin\phi)$ remains virtually constant during the time of sampling, i.e. for the duration of the differentiated pulse. Although the modulating voltage $E\sin\phi$ is a time-function of the nature $e^{j\omega t}$, the assumption is justified since the time period of the modulating signal is very large compared to the pulse-duration and there is no appreciable change in the modulating voltage during the duration of a signal pulse.

Adding the d.c. components in the two parts of equation (11), the equation can be rewritten as:

$$F'(t) = F(x) + F(y)$$

where:

$$F(x) = -(A/T_o) - (2A/T_o)\sum_{n=1}^{\infty}\cos\omega n(t - d) \quad \ldots\ (12)$$

$$F(y) = (A/T_o) + (2A/T_o)\sum_{n=1}^{\infty}\frac{(1+m\sin\phi)}{t_{ro}\omega n}\left[\sin\omega n\left(\frac{t_{ro}}{1+m\sin\phi} - t\right) + \sin\omega nt\right]$$

$$= \frac{(1+m\sin\phi)}{t_{ro}}\left\{\frac{At_{ro}}{T_o(1+m\sin\phi)} + \frac{2A}{T_o\omega n}\sum_{n=1}^{\infty}\left[\sin\omega n\left(\frac{t_{ro}}{1+m\sin\phi} - t\right) + \sin\omega nt\right]\right\} \ldots\ (13)$$

Equation (12), giving $F(x)$, is the Fourier series expansion of very thin negative pulses $[S(w) = 1)]$ occurring regularly at time $t = (d + nT_o)$, i.e. at the vertical trailing edges of the slope-modulated pulses (the product of its amplitude and pulse length is contained in the term A). These are removed by a diode and hence do not contribute any output.

Equation (13), giving $F(y)$, is the Fourier series expansion of repetitive pulses of length t_{ro}, simultaneously amplitude modulated by $E\sin\phi$ and inversely length modulated[4] by the same $E\sin\phi$, the length modulation being on the trailing edges of the pulses. It is thus seen that the highest pulse will have minimum length and vice-versa and this fact has been experimentally verified. Since the differentiated pulses have both amplitude modulation and length modulation incorporated in them, simple memory circuits such as half-wave diode peak detectors[5,6], cathode-follower detectors[5] and low-pass filters cannot be used to give distortionless output. If we expand equation (13) further with the help of Bessel function identity, it is seen that an infinite number of sideband frequencies are produced due to the inverse length-modulation and a certain amount of undesired sideband products will add to the audio output given by amplitude modulation, thus producing considerable distortion.

(b) "BOXCAR" PULSE-LENGTHENER CIRCUIT

It was, therefore, necessary to pass the differentiated signal through a "Boxcar"[3] pulse-lengthener circuit before finally filtering out and feeding the modulation signal to the audio amplifier. The effect of the pulse-lengthener circuit is to eliminate all length modulation from the differentiated pulses and produce pulses of equal duration only. The function $F(y)$ now becomes equal to:

$$\frac{(1+m\sin\phi)}{t_{ro}}\left\{\frac{At_{ro}}{T_o} + \frac{2A}{T_o\omega n}\sum_{n=1}^{\infty}\left[\sin\omega n(t_{ro} - t) + \sin\omega nt\right]\right\}$$

Wang and Uhlenbeck[3] have shown that the Fourier series expansion of an ideal "Boxcar" pulse is given by:

$$F_s(t) = 1 + \frac{\epsilon S_o f_o}{\pi f_m}\cdot\sin\frac{\pi f_m}{f_o}\left[\sin\left\{2\pi f_m t - \frac{\pi f_m}{f_o} + 2\pi f_m\delta\right\}\right.$$

$$\left. + \sum_{n=1}^{\infty}\frac{f_m}{nf_o \pm f_m}\cdot\sin\left\{2\pi(nf_o \pm f_m)t \mp \frac{\pi f_m}{f_o} \pm 2\pi f_m\delta\right\}\right]. \quad (14)$$

where ϵ = Peak amplitude of the modulating wave f_m

S_o = Mean amplitude of the differentiated pulses.

f_o = p.r.f.

δ = the phase difference between the modulating wave and the pulses.

It is assumed in the above equation that the length of the pulses is equal to the time period of the pulses and the memory circuit capacitance is held charged to the maximum amplitude of a pulse, until the next pulse comes and changes the potential of the capacitance to the new peak value of the pulse amplitude.

If we now use a filter with a cut-off frequency equal to f_m and if $f_m \leqslant f_o/2$, then the output of the audio amplifier is given by:

$$\frac{\epsilon S_o f_o}{\pi f_m}\cdot\sin\frac{\pi f_m}{f_o}\left[\sin\left\{2\pi f_m t - \frac{\pi f_m}{f_o} + 2\pi f_m\delta\right\}\right] \ldots\ (15)$$

The interesting point about equation (14) is that the peak amplitude of the audio output decreases with the increase of the modulating frequency due to the factor $[(f_o/\pi f_m)\sin(\pi f_m/f_o)]$ and falls to the value of $2\epsilon S_o/\pi$

for $f_m = (f_o/2)$ from the value of ϵs_o for $f_m \ll f_o$. Since the audio amplitude modulation in equation (13) is contained in the term $(1 + m \sin \phi)/t_{ro} \times (At_{ro}/T_o)$ and the length of the demodulated pulses is equal to T_o, the value of ϵS_o is related to the modulation index of p.s.m. as:

$$\epsilon S_o = \frac{C \,.\, m \,.\, A}{t_{ro}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (16)$$

where C is a constant and depends on the constants of the differentiator.

To compensate for the high-frequency loss in the pulse-lengthener circuit, an adequate equalizer[6] with inverse characteristics has to be used along with the audio amplifier to give a constant gain in the audio band.

Distortion

Since the "Boxcar" pulse lengthener circuit eliminates all length modulation from the signal pulses, the filtered audio output contains only the audio-modulating frequencies and no harmonics of audio frequencies are generated, so long as $f_m \leqslant (f_o/2)$. The system, therefore, should have no harmonic distortion, at least theoretically, if the process of slope modulation and the differentiation are linear.

If an ordinary memory circuit is used for demodulation, it is seen that the inverse length modulation contained in the terms:

$$\frac{At_{ro}}{T_o(1 + m \sin \phi)} + (2A/\omega n) \sum \left[\sin \omega n \left(\frac{t_{ro}}{1 + m \sin \phi} - t \right) + \sin \omega n t \right]$$

of equation (13) will contribute some audio output. If we expand these terms further, we get:

$$(At_{ro}/T_o)(1 - m \sin \phi + m^2 \sin^2 \phi - m^3 \sin^3 \phi + \ldots)$$
$$+ (2A/\omega n) \sum \left[\sin \omega n \left\{ t_{ro}(1 - m \sin \phi + m^2 \sin^2 \phi - \ldots) - t \right\} + \sin \omega n t \right] \quad \ldots \quad (17)$$

The higher powers of $m \sin \phi$ will give the harmonic distortion of audio frequencies and the latter part of the equation will produce an infinite number of sidebands due to the frequency modulation of the p.r.f. harmonics, adding, thereby, more distortion to the audio output. During initial experimentation, it has been found that the distortion in the audio output, using a cathode-follower-type demodulator[5], was always more than 8 per cent.

If it is desired to extract only the output due to the length modulation, the amplitude modulation of the pulses may be nullified by adequate limiters and the constant amplitude pulses filtered by suitable audio filters. But from the above reasonings, the harmonic distortion due to the higher powers of $m \sin \phi$ will persist in the output. The "Boxcar" pulse-lengthener circuit is therefore an essential requirement for distortionless output.

Circuits

The slope modulator (Fig. 4) in the p.s.m. transmitter (Fig. 1) consisted of a gating valve and a charging circuit with a constant current pentode. A negative gating pulse was applied to the switching valve, which, on being cut off, allowed the capacitor to be charged through the constant current pentode. A cathode degenerative resistance was used to make the charging more linear and to control the effective resistance of the charging circuit. The audio signal was fed between the cathode and ground, and the charging current through the pentode was varied due to the variation of the grid-cathode potential. The cathode-follower after the pulse amplifier was used to obtain proper matching with low impedance loads.

The "Boxcar" pulse-lengthener circuit (Fig. 5) in the p.s.m. receiver (Fig. 1) was of the type given by Schlesinger[7]. The gating pulse was obtained from the pulse-transmitter itself, but it can be generated in the

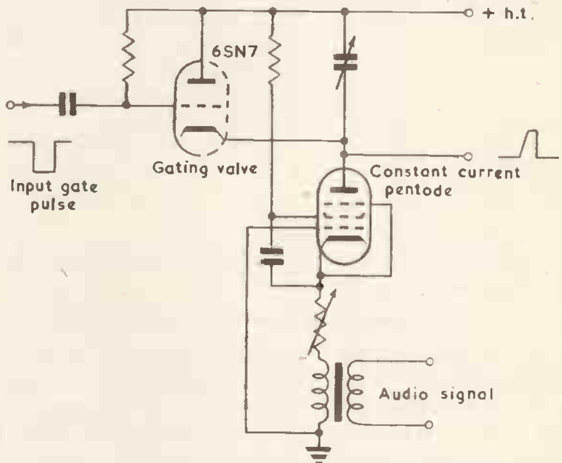


Fig. 4. Slope modulator

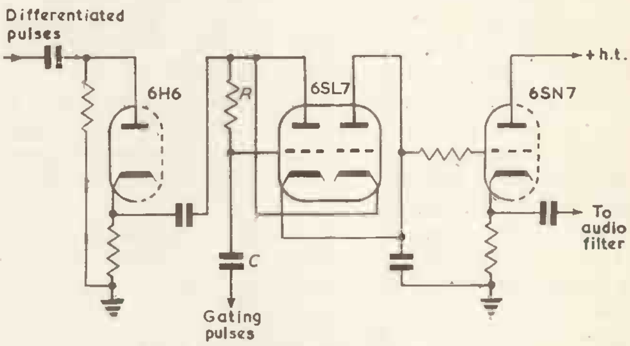


Fig. 5. "Box-car" pulse lengthener circuit

receiver by amplifying and limiting the received pulses. Since the gating pulses and the amplitude-modulated pulses have to be simultaneously applied to the pulse-lengthener circuit, it is not required to generate special gating pulses and there is no time delay necessary between the two pulses as required in the case of recycling detectors[8].

The cathode-follower used is essential, as otherwise the low input impedance of the amplifier stage allows the pulse peaks to fall exponentially and very rapidly during the quiescent period. This distorts the ideal staircase type of demodulated wave and requires more efficient filtering to eliminate the p.r.f. from the audio output.

The audio filter used had a nominal cut-off at 3·5kc/s and was of simple π type with additional losses introduced at 5kc/s and 8kc/s. Fig. 6 shows the gain characteristic of the audio amplifier with the audio filter in the circuit. It is seen that the minimum loss outside the pass-band is 25dB.

Experimental Results

The experimental pulse transmitter had the following

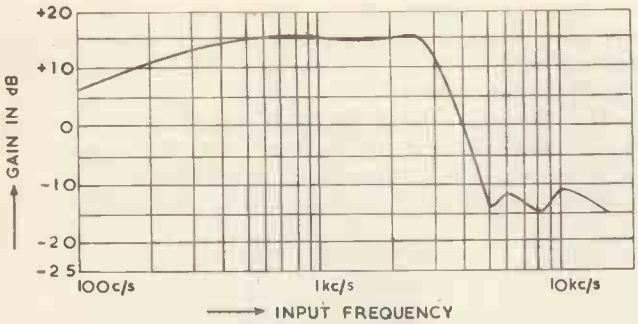


Fig. 6. Audio amplifier and filter characteristic
Input level = −15dB (reference 0dB = 1mW)

characteristics:

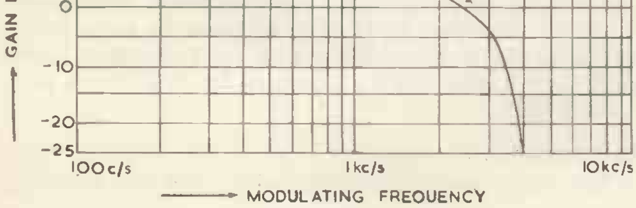
P.R.F. = 10kc/s.
Pulse duration = 10μsec.
Mean rise-time of the unmodulated pulses = 1μsec.
Bandwidth = 1·0Mc/s.
Audio frequency pass-band = 0 to 3 000c/s.

Curve A in Fig. 7 shows the audio frequency response of the pulse receiver with varying modulating frequencies. According to equation (14), curve B in the same figure shows the ideal response. The rise in gain at 500c/s is due to the non-ideal characteristics of the transformer-coupled amplifier stage and the filter and, of course, the sharp fall in gain after 3kc/s is due to the desired filter characteristic. A difference of 20dB in gain is obtained between 2·5kc/s and 4kc/s. The audio output characteristic follows that of the audio amplifier, which is shown in Fig. 5, in reasonably faithful manner.

The linearity of modulation is given by the curve in Fig. 8. It is seen that the audio output is proportional to the input modulating voltage over a range of about 35dB, which is sufficient to accommodate the usual volume range in speech transmission.

The distortion characteristic of the p.s.m. system is given in Fig. 9. The percentage distortion was measured with a bridge type r.m.s. distortion meter. The average distortion is within 5 per cent for a volume variation of 30dB approximately. The higher distortion at lower modulation index is due mainly to high hum and other noise level in the audio amplifier. Moreover, the pulse-lengthener circuit produces a certain amount of distortion at low input levels. The average distortion in the amplifier itself is 1·5 per cent. The higher distortion at higher modulation index is due to the non-linearity in the modulator and later pulse amplifiers in the transmitter. The non-ideal characteristic of the audio filter has also contributed to the distortion as the loss of 20dB only beyond cut-off frequency allows for 1·5 per cent distortion in the output. The higher distortion at 3kc/s input is due to lesser gain at this frequency and the presence of a nearer interfering sideband frequency (7kc/s). However, the overall distortion characteristic is satisfactory for the usual speech transmission circuits.

Fig. 7. Audio output characteristic of p.s.m. system without compensation
Input level = +10dB (reference 0dB = 1mW)



Conclusions

From equation (3), it is seen that the maximum modulation possible in p.s.m. is dependent on the bandwidth and the pulse duration of the system. Maximum value of m, that can be obtained with 1Mc/s bandwidth and 10μsec pulse-width, is of the order of 90 per cent.

In a usual pulse-amplitude modulation or pulse-length modulation system, using only a filter as the demodulator,

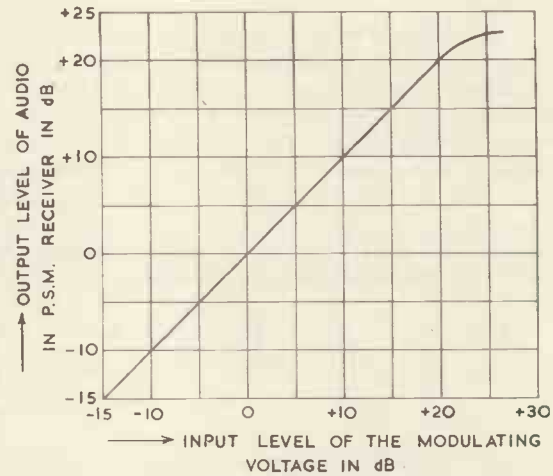


Fig. 8. Curve showing linearity of modulation in p.s.m.
Modulating frequency = 1kc/s (reference 0dB = 1mW)

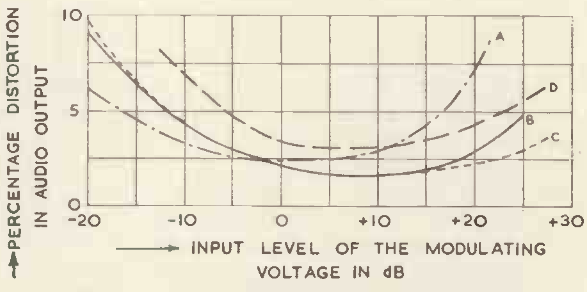


Fig. 9. Distortion characteristic of p.s.m. system
(reference 0dB = 1mW)
Curve A—for 500c/s modulating frequency
Curve B—for 1 000c/s modulating frequency
Curve C—for 2 000c/s modulating frequency
Curve D—for 3 000c/s modulating frequency

the amplitude of the audio output is equal to mAd/T_0, whereas in p.s.m. the amplitude of the audio output is equal to ($S_0 \epsilon f_0/\pi f_m$. sin $\pi f_m/f_0$). The audio voltage output is increased manifold in the ratio of T_0/d by the "Boxcar" pulse-lengthener circuit.

Kretzmer[4] has shown that even with an ideal low-pass filter and audio amplifier, the theoretical distortion in a p.l.m. system is of the order of 4 per cent with $f_m = (f_0/2)$. Levy[9] has reported a p.p.m. system, where the overall distortion in audio characteristics was of the order of 5 per cent. In comparison to these, the theoretical harmonic distortion in p.s.m. is zero with $f_m = (f_0/2)$, if the differentiating circuit and filter characteristics are ideal, and the amplitude distortion is of the order of -3dB at the highest modulating frequency. If a proper equalizer, ideal audio filter and amplifier can be used, the distortion in p.s.m. can be minimized to zero.

Acknowledgments

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Radio Network for Scottish Border Police

An extensive v.h.f. radio control scheme has recently been inaugurated by the Police Forces of Berwick, Roxburgh and Selkirk.

This radio network has been supplied and installed by Marconi's Wireless Telegraph Co. Ltd. Its range extends over some 1 250 square miles of the three Border Counties, and is an important addition to existing police radio installations, which already cover a large proportion of the country.

The new network consists of a master control room at the Police Headquarters at Jedburgh, with repeater stations at Hardens Hill and Meigle Hill. A further sub-control station is located at Hawick, the latter feeding into the main network via Meigle Hill. From these points two-way radiotelephonic communication is established with the radio-equipped police patrol cars. Owing to the hilly nature of the Eastern part of the area, very intensive technical tests were carried out to select the best sites for the project.

The equipment supplied includes four 10W fixed station v.h.f. installations type H16H, eleven 10W mobile v.h.f. equipment type H16 and one 5W mobile v.h.f. equipment type HP10.

An interesting feature is the principle of reversed frequency triggering, which was introduced into the Scottish Police communications systems by Chief Inspector W. N. Bruce of Edinburgh, and has proved highly successful. This system can effect considerable economies in capital cost, and operates as follows.

Contrary to standard practice, the master control room does not communicate with the patrol cars directly, but via one of the repeater stations. As a practical example, master control may be transmitting on a frequency of 80Mc/s and receiving on 90Mc/s. When a message is transmitted from master control it is received by the repeater station, the receiver of which is tuned to 80Mc/s. This receiver is connected directly to the input of the repeater station's transmitter which retransmits the message, this time on a frequency of 90Mc/s.

The receivers in the patrol cars are tuned to this frequency and therefore receive the re-transmitted message in the normal way. Conversely, the mobile transmitter in the patrol car transmits on 80Mc/s, to be picked up by the repeater station receiver and re-transmitted on 90Mc/s, the frequency to which, it will be recalled, the master control receiver is tuned.

(It is emphasized that the frequency figures quoted are hypothetical and given only as an example.)

Hitherto, master control has had to be sited for an extensive coverage, a requirement which is often inconvenient and difficult in a town centre. By the use of the reversed frequency method the master control can be at the most convenient point (usually Police Headquarters) even though it may be low-lying. The only requirement is that the repeater station is in line-of-sight with master control—usually a convenient hill-top in the rural area surrounding the town is easily found.

A possible alternative would, of course, be to site master control on the hill top and control it remotely by landline from Police HQ., but as this would demand the use of expensive equipment such as coaxial cable; the radio system outlined above is much less expensive to install and maintain.

A New V.H.F. Mobile Equipment

There are now more than 10 000 mobile radio-telephones in Britain, and these are operating on a frequency band which is being drastically reduced in width in order to make available additional channels for television purposes. The ever-increasing demand for mobile radio is therefore faced with a decreasing number of channels available for development.

In order to alleviate these difficult circumstances, Pye Telecommunications have developed a new dashboard mounting mobile radiotelephone for 25kc/s channelling known as the " Ranger ".

The " Ranger" is intended to work in either double or single frequency simplex on a press-to-talk basis, on channel spacings as close as 25kc/s over its whole frequency range (25-174Mc/s), which will enable the more efficient use of the available frequency bands, and permit the allocation of many more frequency channels. The power output at these frequencies is from 4 to 6W.

The specification to which the equipment works is tighter technically than the new specifications envisaged for use in the U.S.A., even though the development of two-way radio in Britain is still far behind the United States in numbers, where half a million radio-telephones are in use. At present mobile radio channels are spaced at 100kc/s in Britain and 60kc/s in the U.S. The U.S. plan to move to 30kc/s. It is claimed therefore that the " Ranger " equipment is the first in the world to offer 25kc/s channel spacing.

The receiver is a high performance double superheterodyne with efficient noise limiting characteristics. Attention has been given to the circuit design so that cross modulation and inter-modulation interference shall be at the lowest possible level. Two crystals and eleven valves are employed, comprising: 2 low noise r.f. amplifiers, 1st crystal oscillator and multiplier, low noise 1st mixer, 2nd mixer and 2nd oscillator, 3 i.f. amplifiers, 2nd detector-a.v.c. rectifier-a.f. amplifier, noise limiter, a.f. output.

The transmitter consists of a crystal oscillator, followed by 3 multipliers and a power amplifier, the modulator comprising a phase splitter and push-pull modulator.

The equipment is designed for operation on either 6 or 12V d.c. supply.

The fixed station transmitting and receiving equipment is of similar design, although existing installations can be modified to work satisfactorily with the new mobile equipment.

In order to provide accurate frequency checking of either the fixed or mobile equipment a portable sub-standard with internal temperature controlled crystals is available.

Colour Television in the U.S.A.

A Review of Recent Progress

By C. G. Mayer*

A colour television outside broadcast van.

IT was in December 1953 that the Federal Communications Commission adopted the NTSC colour television signal specification based on the RCA system. Since then great strides have been made in broadcasting colour to a mass audience. The major problem has been to produce a colour receiver which is as easy to operate and as reliable as a black-and-white receiver, and which can be sold at a price sufficiently low to be attractive to a large number of viewers[1].

Tri-colour Kinescope (c.r.t.)

Immense effort has been put into the engineering of colour tubes and the first commercially available tube was produced by RCA in 1953. This was a 15in glass tube of the now familiar shadow mask type. Further development resulted in production by November 1954 of a much improved 21in metal envelope tube which is used in sets today.

The RCA type 21AXP22, as it is called, has its phosphor-dot screen deposited directly on the curved face plate, and by so doing gives maximum picture area for a given screen diameter. The thin metal shadow mask, which has 300 000 holes, is formed to match the face plate and mounted directly behind it on frictionless spring supports. This method of mounting has proved successful in eliminating errors of mechanical registration, and is an important contribution to mass production[2]. One advantage of the metal shell construction is that it provides a substantial degree of shielding from the earth's magnetic field. The effect of what remains can be readily corrected by means of a magnetic field equalizer consisting of a system of small magnets placed around the front end of the tube.

Compared with a regular 21in black-and-white kinescope having the same deflexion angle of 70°, the colour tube is two inches longer and four pounds heavier. The difference in length is necessitated, not by the gun length which is about equal in the two tubes, but by the space requirements for the yoke and other components located on the neck of the colour tube. The 21AXP22 gives greater brilliance than earlier colour tubes, and for an average scene will produce a colour picture of 260 square inches area with highlights of 25 foot-lamberts. This brightness combined with a filter glass faceplate gives good contrast pictures under average ambient lighting conditions.

Various other forms of tri-colour tubes have been proposed involving different principles such as those using focus grids or sensing elements. Some have operated experimentally but have still to be developed into practical devices, and even the most vocal proponents of some of these other forms have adopted the shadow mask type for their colour receivers. RCA is convinced that the 21AXP22 is the only type of tube available now or in the near future which can be produced in the quantities required and at an acceptable price. Production techniques have been so well mastered that the reject rate is now approaching that for black-and-white tube manufacture. The tube is being produced at the RCA Lancaster plant at a rate of 5 000 per month, and to date more than 30 000 tubes have been delivered. Early this year the price was reduced from 175 dollars to 100 dollars and the design has excellent potential for further cost reduction. Confidence in the future is so great that facilities are being expanded for production of more than 30 000 tubes per month.

The Modern Colour Receiver

Meanwhile, work has been going on to simplify receiver circuits and to improve performance and reliability. The new RCA 21in colour receiver now in production and on the market contains only 26 valves including the kinescope, plus two crystal diodes and two selenium rectifiers. This is a considerable reduction from the 36 valves used in the 1954 21in set.

In the evolution of colour receivers emphasis has been placed on attaining the same high degree of stability which the public has come to expect in the operation of black-and-white sets. In earlier colour receivers the colour signal was demodulated at low level, matrixed and then amplified by three separate amplifiers. This method of operation required that the relative gain of the demodulators did not vary, and that the gain of the three amplifiers also remained constant in order to maintain colour balance. The new circuits demodulate at high level, using valves as switches, so that their characteristics do not affect the performance of the circuit. Colour balance is determined by the number of turns on a transformer, thus giving a very high order of stability with simple straightforward circuits.

With the NTSC specification demodulation can be accomplished in the colour receiver with unequal bandwidths as in the IQ system, or with equal bandwidths. Tests have shown that there is little difference to be seen in the picture received by the two methods. Advantage has been taken of this in the new colour receiver by producing the colour difference signals directly, with the

* *RCA European Technical Representative.*

advantage of improved stability and reliability at reduced cost.

Stability of colour phase is maintained by the use of colour synchronizing circuits in which almost all tuned circuits are common to both the chroma and colour synchronizing circuits, so that frequency drift has no effect.

Ease of operation is ensured by the provision of simple circuits to give stable automatic chroma control. In the same way that automatic gain control maintains the luminance output independent of variations in picture carrier signal, so automatic chroma control maintains the chrominance output constant for variations in the colour subcarrier signal.

As a result of these simplifications the only operating knobs exposed on the modern colour receiver are those of a normal black-and-white set—i.e. station selector, volume and brightness. The controls for colour saturation and colour phase or hue are available as pre-set controls under a hinged panel along with the normal raster controls.

Network and Programming

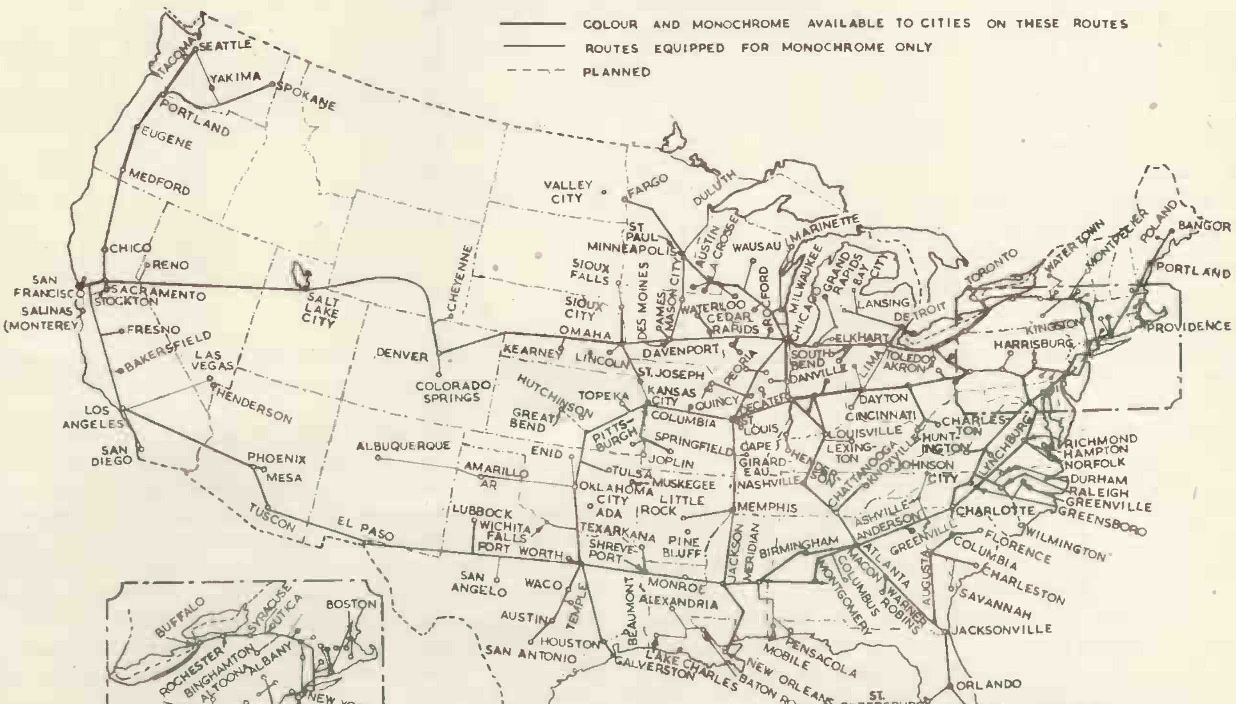
It is useless to expect colour television to make progress unless attractive programmes are on the air, and colour signals brought within reach of sets in peoples homes. As with black-and-white the main source of programmes is from the major studios, brought to local stations via microwave and coaxial cable networks. These networks have been engineered to provide adequate control of amplitude and delay characteristics over the bandwidth required by the monochrome signal. The colour signal has the same bandwidth but requires closer control of these characteristics, and especially of phase relationships in the region of the colour subcarrier. This means that more attention must be given to the high frequency end of the transmission characteristic. Differential phase must be controlled to a few degrees to avoid changes in colour hues, and variations in the amplitude characteristic kept to a minimum to reduce errors of colour saturation[3].

With the wideband microwave relays which predominate today there has been little difficulty. The older type of coaxial cable system, with its 3Mc/s transmission band, was a more troublesome problem. The solution adopted was to frequency-change the colour subcarrier and the chrominance components to a subcarrier frequency of 2·6Mc/s. This is done automatically at the input to the cable, and the colour signal is restored to its correct frequency at the receiving terminal. In this process, signal components between 2Mc/s and 3·3 Mc/s are discarded, resulting in some loss of definition in the picture. The newer sections of the cable system are provided with wideband coaxial circuits for which no special measures are required for satisfactory transmission of colour signals.

The Bell System now has about 47 000 channel miles capable of transmitting colour signals, and this represents about 70 per cent of available network facilities (Fig. 1). Colour service is available to about 250 stations in 150 cities, and out of a total of 440 television stations nearly one-half are equipped to transmit colour, so that about 90 per cent of American homes are within reach of a colour signal.

The number of stations capable of originating their own live colour programmes is also growing steadily. The NBC network includes 16 such stations, and another 35 can transmit colour films and slides. These numbers are increasing rapidly as apparatus is installed to meet existing orders. NBC has two major colour studios in New York and recently added "Colour City" on the West Coast. Additional studios are being built in New York to meet the increased demand for colour facilities. Both the Brooklyn and Hollywood studios are equipped

Fig. 1. The Bell System television network



with large-screen colour projectors so that studio audiences can watch the actual television production and also see in colour on a 20 by 15ft screen the picture on the air.

The increased cost of colour production compared with monochrome varies from 5 per cent to 40 per cent depending on the programme. Production experts agree, however, that costs will settle down to a point where, for example, a major variety show may be expected to cost 10 to 15 per cent more in colour than in black and white.

The National Broadcasting Company has recently announced a five-fold increase in colour programmes for this autumn. From October onwards approximately 40 hours per month will be devoted to live studio colour programmes. In addition NBC mobile colour units are being used to cover popular national sporting events. The Columbia Broadcasting System has also been broadcasting regular colour programmes and are also adding to their programmes.

It may be noted in passing that Canada is taking an active interest and will soon be broadcasting experimental colour programmes.

Transmitter and Studio Equipment

The broadcasting of colour usually involves the following steps:

(1) Modification of the black-and-white transmitter equipment to take the colour programmes from network sources.

(2) Addition of equipment for origination of pictures from colour films and slides.

(3) Camera equipment for origination of live studio pictures in colour.

Special consideration has to be given to certain characteristics of the transmitter[4], but usually one which performs well for black and white will be satisfactory for colour with only minor modifications and adjustments. Adequate frequency response over the entire bandwidth is an essential requirement: other important factors are linearity, which determines correct colour saturation and brightness; and, differential phase and envelope delay, which affect the quality of colour edges.

High quality colour broadcasting requires close adherence to carefully specified transmission standards, and much new and unique test equipment has been developed and is available for the purpose. One of these, known as the colour stripe generator, provides a test signal which enables servicemen to check colour receiver installations when only black-and-white programmes are on the air. This unit, when coupled to the video line feeding the transmitter, generates two colour bursts, one at the beginning and one at the end of each horizontal scanning line. If the transmission path is passing the colour burst, a vertical yellow-green stripe can be observed on colour sets by adjusting the horizontal frequency control until the colour circuits of the receiver are activated by the burst at the beginning of each line. The stripe is imperceptible on black-and-white receivers.

Under certain conditions of multi-path reception or faulty transmission lines it is possible to pick up a satisfactory black-and-white picture, but to have partial cancellation of the colour subcarrier. The stripe generator

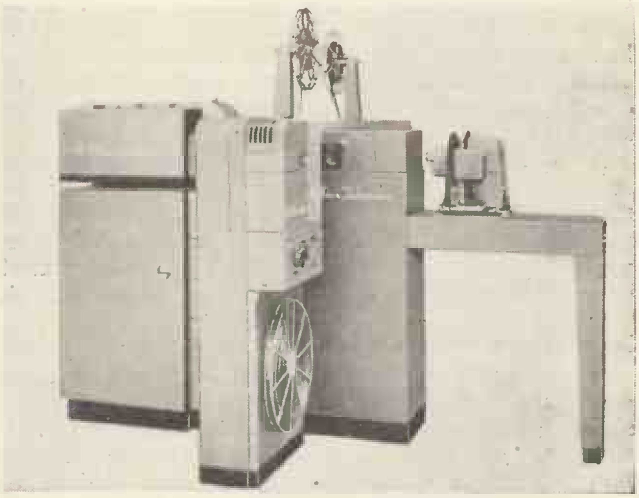
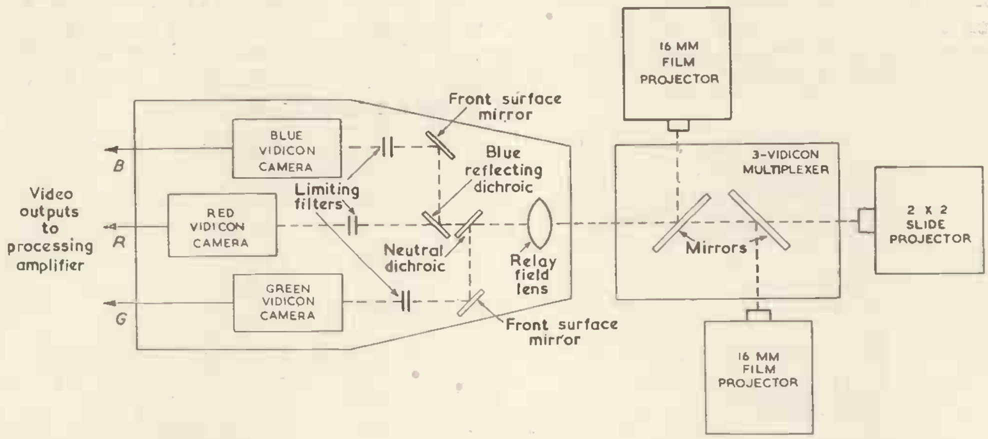


Fig. 2. Equipment for transmission of colour slides and films

Fig. 3. Arrangement of slide and film equipment



helps to avoid such situations by providing information to the serviceman without interfering with regular monochrome operations.

A convenient system for transmission of film and slide material is shown in Fig. 2. It consists of two 16mm projectors and a slide projector together with a hinged mirror arrangement to enable the desired picture to be directed on to a 3-Vidicon tube camera. The schematic diagram (Fig. 7) shows the multiplex unit, which can be controlled remotely; also the dichroic mirrors and colour filters which divide the light source into the red, blue and green components. Registration presents no difficulty and the system is very stable in operation, as well as giving excellent fidelity of colour reproduction.

Eventually stations will also need facilities for live pick-up in the studio and for local events. Colour cameras have been reduced in size and complexity so that the latest

Fig. 4. An RCA colour television camera

model is only a little larger and heavier than the regular monochrome camera. The camera shown in Fig. 4 contains three image orthicon tubes and the usual dichroic mirror arrangement.

The replacement of the three tubes in a colour camera by a single tube capable of generating simultaneously all the primary colours has been a major goal of television research. The RCA Laboratories recently demonstrated a developmental tube, known as the Tri-colour Vidicon[5]. This tube combines all the colour pick-up functions in a single tube, ensuring precise optical and electrical registration. It will be seen from Fig. 5 that the tube is no larger than the regular 3in image orthicon, and its use will permit greater simplicity and compactness in colour camera design. The tri-colour Vidicon has a unique and intricate colour-sensitive target applied to its faceplate by an evaporation technique. The target, a rectangle of 1½in diagonal, consists of nearly 900 fine vertical strips of alternate red, green and blue colour filters, covered by three sets of semi-transparent conducting signal strips. The signal strips corresponding to a given colour are all connected to a common output terminal, and insulated from the strips of the other colours.

As the scanning beam sweeps horizontally through all the strips, the tube generates directly the three simultaneous primary colour signals that form the composite transmitted colour signal. Current research is concentrating on the development of more sensitive photo-conductive materials that will permit a tube of this new type to operate as efficiently under all lighting conditions as present day studio-type camera tubes.

Fig. 5. The Tri-colour Vidicon

Video Tape Recording

A development which created widespread interest when it was first demonstrated by RCA in 1953 was the recording of television signals on magnetic tape[6]. Improved equipment has been installed for field test in the NBC studios in New York. A reduction in tape speed to 20ft per second and a new thinner tape enables a 15-minute colour programme to be recorded and wound on one 19in diameter spool. Closed circuit demonstrations of tape-recorded colour programmes have been given over commercial network facilities, and although some problems still remain to be solved, tape will soon be in experimental use in broadcast programmes. Kinescope film recording is already widely used to take care of time differences across the American continent, but for colour it is expensive, difficult and inconvenient. Tape provides the complete solution at a fraction of the cost.

Conclusion

Colour programming of about 15 hours per week will undoubtedly create and stimulate public interest in colour television. Colour receivers are available to meet the expected demand. Estimates of sets in use vary widely, but industry leaders have recently estimated fifty to one hundred thousand sets by the end of this year and about half a million in 1956. This belief in colour is based on the following factors:

Emergence of a satisfactory and low-priced colour set with 21in picture size.

Experience and technical facilities are now available to do a good job of colour broadcasting on a national basis.

Sponsors are willing to support colour and feel the extra cost is worthwhile.

Public awareness of the excitement of colour brought about by magazines, films, amateur photography and now by television in the home.

There are some, of course, who feel that sales will be slow at a price of 700 to 800 dollars for a set. However, this figure is proportionately less than the price at which black-and-white receivers were first sold—and now there are more than 35 million in America.

Another aspect of the price question is to consider it in relation to earning power. One finds that for the United States the current price for a 21in colour set represents about two months' earnings of an average worker. This same price ratio has not prevented widespread sales of television receivers in Europe.

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Maximum Power Transfer in Transitor Amplifiers

By G. L. Fougere*

Considerable simplification can be made to the analysis of transistor amplifiers by restricting observation to the condition of maximum power transfer. The power gain is then independent of generator and load impedances and can be related solely to the parameters of the transistor. Analysis yields a figure of merit in terms of these parameters and it is possible to make rapid assessment of the available power gain in a particular configuration. The relevant information is presented graphically for the three configurations which are most frequently encountered

THE algebraic complexity of transistor analysis is partly due to the unavoidable dependence of output impedance upon input impedance. All valve amplifiers, with due exception to the grounded grid amplifier, have a high output/input isolation and large changes can be made in one circuit without producing any effect in the other. The transistor, however, has input and output in series and the analytical difficulty which arises can be overcome by considering only the condition of maximum power transfer. This is fortunately an important condition in many applications and one which permits a simplified approach. It is assumed that the transistor parameters r_e, r_b, r_c and αr_c in Fig. 1 are ohmic and that r_e and r_b are comparable and both much less than r_c. It is also assumed that the transistor is used well within its specified frequency rating and that the phase shift is consequently negligible.

With such provisions, maximum power transfer will occur when the generator and load are matched to the corresponding image impedance of the four terminal network which represents the transistor amplifier. These image impedances are equal to the geometric means of the open- and short-circuit impedances of the network.

Grounded Emitter

Thus for the grounded emitter circuit the input impedance has been shown[1] to be:

$$Z_i = r_b + r_e \cdot \frac{r_c + Z_L}{r_c(1-\alpha) + r_e + Z_L}$$

Reference should be made to Fig. 1 which is a diagram of the equivalent circuit and contains definitions of r_c, r_b, r_e, α, Z_L and Z_G.

Thus for open-circuit when the load impedance $Z_L = \infty$

$$Z_{io} = r_b + r_e \simeq 2r_b$$

and for short-circuit when $Z_L = 0$.

$$Z_{is} = r_b + \frac{r_b \cdot r_c}{r_c(1-\alpha) + r_b}$$

Thus Z_1 the input image impedance is given by:

$$Z_1 = \sqrt{\left[2r_e\left(r_b + \frac{r_b r_c}{r_c(1-\alpha) + r_b}\right)\right]}$$

If $r_e = r_b$ and $r_c \gg r_b$ and since $r_c(2-\alpha) \gg r_e$

$$Z_1 = r_b\sqrt{\left[\frac{2r_c(2-\alpha)}{r_c(1-\alpha) + r_b}\right]}$$

Similarly:

$$Z_o = r_b + r_c(1-\alpha)\frac{r_b + Z_G}{r_b + Z_G + r_c}$$

Thus for open-circuit, $Z_G = \infty$

$$Z_{oo} = r_b + r_c(1-\alpha)$$

and for short-circuit $Z_G = 0$

$$Z_{os} = r_b + r_c(1-\alpha)\frac{r_b}{r_b + r_c}$$

If $r_e = r_b$ and $r_c \gg r_b$ then:

$$Z_{os} \simeq r_b + r_b(1-\alpha) = r_b(2-\alpha)$$

and Z_2 the output image impedance is given by:

$$Z_2 = \sqrt{\left[r_b(2-\alpha)\left\{r_b + r_c(1-\alpha)\right\}\right]}$$

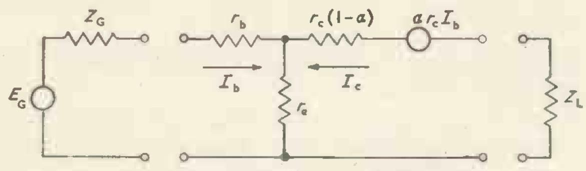
Thus the maximum power transfer:

$$\eta_{max} = \frac{Z_2\alpha^2}{Z_1(1-\alpha)^2}$$

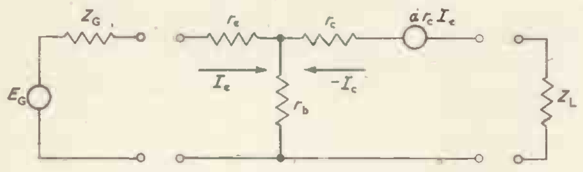
* *Royal Aircraft Establishment, Ministry of Supply.*

Fig. 1. Equivalent circuits

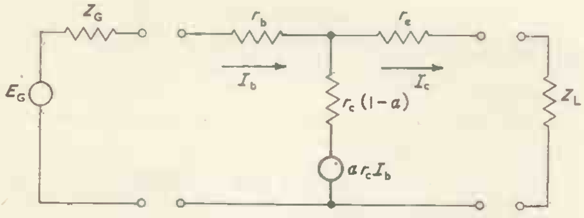
r_e = emitter resistance; r_b = base resistance; r_c = collector resistance; $\alpha = |\partial I_c/\partial I_e|_{V_c}$



Grounded emitter equivalent circuit

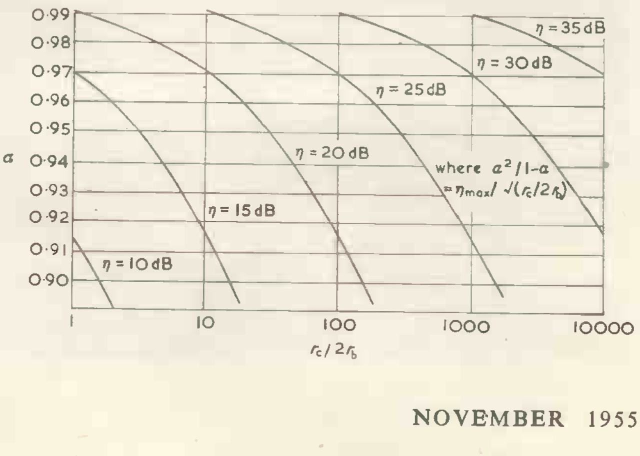


Grounded base equivalent circuit



Grounded collector equivalent circuit

Fig. 2. Grounded emitter power gain



$$= \frac{\alpha^2[r_b + r_c(1-\alpha)]\sqrt{r_b}}{(1-\alpha)^2\, r_b 2r_c}$$

$$\simeq \frac{\alpha^2\sqrt{(r_c/2r_b)}}{1-\alpha}$$

For this configuration the ideal transistor is one possessing a high r_c to r_e ratio and a current gain close to unity.

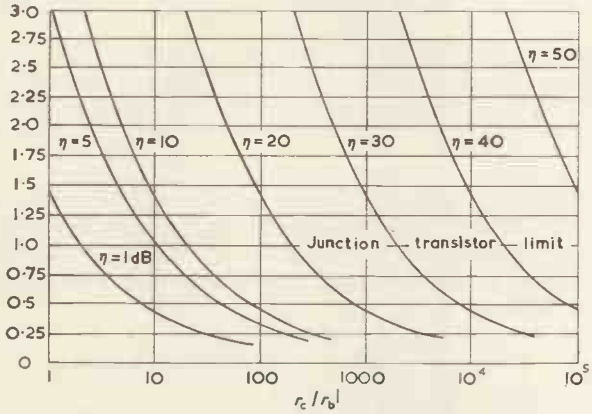


Fig. 3. Relation between maximum power gain, current gain and transistor constants for grounded base

In Fig. 2, η is plotted for various power levels and the transistor constants for a given level may be readily seen.

GROUNDED BASE

By a similar approach it can be shown that the grounded base configuration requires similar properties in the transistor, namely:

$$\eta_{max} = \alpha^2 r_c/2r_b$$

and this function is shown graphically in Fig. 3.

GROUNDED COLLECTOR

For the grounded collector amplifier, however, a quite different figure of merit emerges.

This configuration has a voltage gain approximating to unity and the power gain available may be regarded as the ratio of the input impedances.

The input impedance has been shown[3] to be:

$$Z_1 = r_b + r_c \cdot \frac{r_e + Z_L}{r_c(1-\alpha) + r_e + Z_L}$$

thus for open-circuit, when $Z_L = \infty$:

$$Z_{io} = r_b + r_c$$

and for short-circuit, when $Z_L = 0$:

$$Z_{is} = r_b + \frac{r_c r_e}{r_c(1-\alpha) + r_e}$$

If $r_e = r_b$ and $r_b \ll r_c$:

$$Z_{is} = \frac{r_b \cdot r_c(2-\alpha)}{r_b + r_c(1-\alpha)}$$

and therefore:

$$Z_1' = \sqrt{\left[\frac{r_b \cdot r_c(2-\alpha)}{r_b + r_c(1-\alpha)}\right][r_b + r_c]}$$

$$\simeq r_c\sqrt{\left[\frac{(2-\alpha)}{1 + (r_c/r_b)(1-\alpha)}\right]}$$

Let:

$$\rho = \frac{r_b}{r_c(1-\alpha)}$$

then:

$$Z_1' = r_c\sqrt{\left[(2-\alpha)\left(\frac{\rho}{\rho+1}\right)\right]}$$

$$\simeq r_c\sqrt{(2-\alpha)} \text{ when } \rho \gg 1$$

$$= r_c\sqrt{\left(\frac{2-\alpha}{2}\right)} \text{ when } \rho = 1$$

where ρ is likely to have any value between 1 and 10.

Similarly[2]:

$$Z_o = r_e + r_c(1-\alpha) \cdot \frac{Z_G + r_b}{Z_G + r_b + r_c}$$

thus for open-circuit, when $Z_G = \infty$:

$$Z_{oo} = r_e + r_c(1-\alpha)$$

and for short-circuit, when $Z_G = 0$:

$$Z_{os} = r_e + r_c(1-\alpha) \cdot \frac{r_b}{r_b + r_c}$$

Again, if $r_e = r_b$ and $r_b \ll r_c$ then:

$$Z_2' = \sqrt{\left[r_b + r_c(1-\alpha)\right]\left[r_b\left(\frac{r_b + r_c(2-\alpha)}{r_b + r_c}\right)\right]}$$

and since $\alpha \leqslant 1$:

$$Z_2' = r_b \cdot \sqrt{(2-\alpha)} \cdot \sqrt{[1 + (r_c/r_b)(1-\alpha)]}$$

$$= r_b \cdot \sqrt{(2-\alpha)} \cdot \sqrt{[(\rho+1)/\rho]}$$

$$\simeq r_b \cdot \sqrt{(2-\alpha)} \text{ when } \rho \gg 1$$

$$= r_b \cdot \sqrt{(2-\alpha)} \cdot \sqrt{2} \text{ when } \rho = 1.$$

Hence:

$$\eta_{max\;\rho\gg} = \frac{I_e^2 \cdot Z_2'}{I_b^2 \cdot Z_1'}$$

$$= \frac{r_b}{r_c(1-\alpha)^2}$$

and:

$$\eta_{max\;\rho=1} = \frac{2r_b}{r_c(1-\alpha)^2}$$

From this it can be seen that the figure of merit is quite different from that obtained in the other configurations, and that a useful power gain can be obtained only as $\alpha \rightarrow 1$. In Fig. 4, η is plotted against the transistor constants and the available gain can be readily obtained for various conditions.

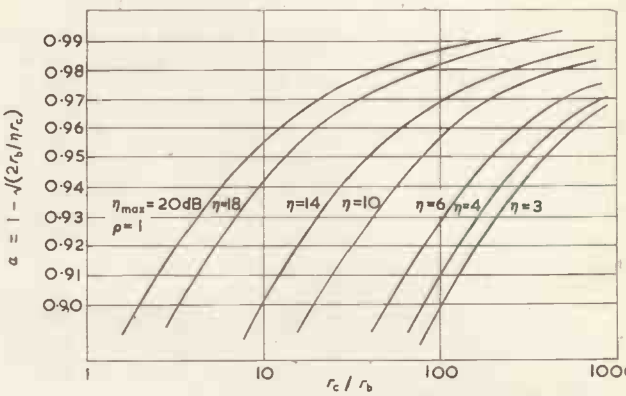


Fig. 4. Relation between maximum power gain, current gain and transistor constants for grounded collector

In general, this means that transistors should be designed or chosen for a particular purpose and that in certain circumstances it can be uneconomical to use a transistor in a configuration for which it is characteristically unsuited.

Acknowledgments

Acknowledgment is made to the Chief Scientist, Ministry of Supply, for permission to publish this article.

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A Q Meter Method of Measuring Very Low Reactance at High Frequencies

J. P. Newsome*, M.Sc., A.M.I.E.E.

The test impedance is connected into the Q meter measuring circuit through a mutual inductance or "step-up" transformer; a technique is developed which allows very low values of capacitive and inductive reactance to be measured. Details are given of a circuit designed to measure down to ·003μH at 1Mc/s.

FIG. 1 shows in basic form a Q meter circuit suitable for use as a direct reading Q meter up to a frequency of about 10Mc/s. The circuit consists of an oscillator working into a very low resistive impedance, thus providing a source of low impedance and constant e.m.f. The measuring circuit consists of an inductor in series with a tuning capacitor across which is connected a valve-voltmeter, which is used to indicate the circuit Q factor. The behaviour of this circuit is examined by the author in a previous paper[1].

Low values of reactance can be measured with this circuit by a direct series substitution method. Consider a circuit at resonance (defined as tuning capacitor voltage maximum) with a tuning capacitance C_{T1}; an inductive reactance ωl_x is inserted and resonance now obtained with a tuning capacitance C_{T2}. If the circuit Q factor, $Q_T \gg 1$, then l_x is given by:

$$l_x = 1/\omega^2 \cdot \frac{(C_{T1} - C_{T2})}{C_{T1}C_{T2}} \quad \ldots\ldots\ldots\ldots (1)$$

The lower limit of l_x is governed by the ability to measure $(C_{T1} - C_{T2})$ to an adequate degree of accuracy; this involves the use of an accurate incremental capacitor and an ability to fix precisely the circuit resonant condition. Consider a measurement at 1Mc/s; let $C_{T1(max)} = 900$pF, $(C_{T1} - C_{T2})_{(min)} = 10$pF, then $l_{x(min)} = 0{\cdot}32\mu$H.

By the introduction of mutual inductors (or measuring transformers) between the test impedance and the measuring circuit, a method has been developed which is capable of lowering the value of $l_{x(min)}$ by two or three decades. A measuring transformer has a frequency range of approximately 3:1 and is readily constructed. The method is characterized by rapid operation once the properties of a transformer are known. An accuracy of measurement of 5 per cent can be obtained with normal laboratory equipment over a wide range of test reactance values.

Very low reactance values are frequently of necessity calculated where the current path is well defined; when this is not the case, a measurement becomes necessary and the method presented here is considered of significance as no ready alternative exists.

The method was first developed by Lafferty[2]; it is here presented in a more precise and analytical form.

The Behaviour of a Mutual Inductor at High Frequencies

At high frequencies it is necessary to take into account the distributed capacitance system which exists in a screened mutual inductor. Where capacitance effects are limited, it is possible to represent them by lumped values. Fig. 2, (a) and (b), shows the equivalent circuits of a screened mutual inductor with the secondary circuit (a) isolated and (b) having one side connected to screen; in each case one side of the primary is connected to screen. It is assumed that losses associated with the capacitors may be ignored and that the inductors are air-cored. Winding resistances are eliminated from the diagrams for purposes of clarity.

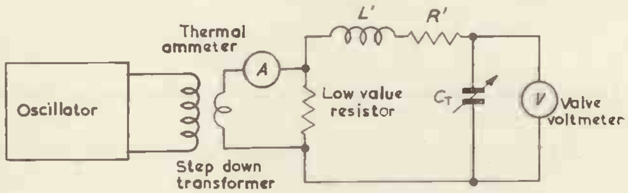


Fig. 1. Basic circuit of low impedance source Q meter

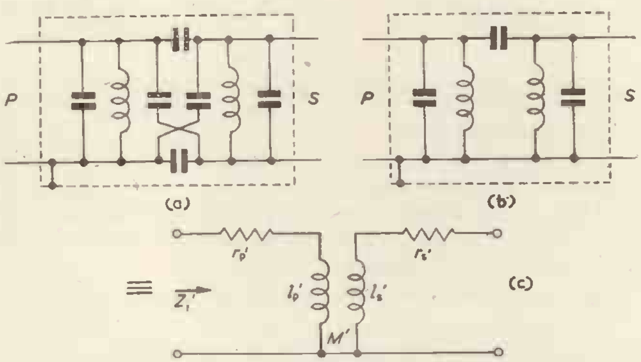


Fig. 2. High frequency equivalent circuits of a mutual inductor

Both these circuits can be reduced to the simplified equivalent circuit shown in Fig. 2(c), where the capacitance system has been suppressed and where all values may vary significantly with frequency. This circuit is an exact equivalent of Fig. 2(b), but is an equivalent of Fig. 2(a) providing the use is limited to a consideration of voltages applied to the windings and not across the windings. A method of converting Fig. 2(b) to Fig. 2(c) is due to Butterworth[3]. The effective mutual inductance M' is a complex quantity; it may be taken as a real number when the Q factors of the windings are large compared with unity.

The simplified equivalent circuit may be realized in the following way. Let:

$$Z_1' = R' + j\omega L'$$
$$\text{and } Q' = \omega L'/R'$$

Two measurements are undertaken from the primary terminals.

Measurement (1). Secondary Open-Circuit

$$Z_1' = r_p' + j\omega l_p' \quad \ldots\ldots\ldots\ldots (2)$$

* *University of Nottingham.*

so that:

$$L_1' = l_p' \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

$$Q_1' = \omega l_p'/r_p' \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

Measurement (2). *Secondary Short Circuit*

$$Z_i' = r_p' + r_s'\left\{\frac{\omega^2 M'^2}{r_s'^2 + \omega^2 l_s'^2}\right\} + j\omega\left\{l_p' - l_s' \cdot \left[\frac{\omega^2 M'^2}{r_s'^2 + \omega^2 l_s'^2}\right]\right\} \quad (5)$$

so that:

$$L_2' = l_p' - l_s' \cdot \frac{\omega^2 M'^2}{r_s'^2 + \omega^2 l_s'^2} \quad \ldots\ldots\ldots\ldots \quad (6)$$

$$Q_2' = \frac{\omega L_2'}{r_p' + r_s' \cdot \dfrac{\omega^2 M'^2}{r_s'^2 + \omega^2 l_s'^2}} \quad \ldots\ldots\ldots\ldots \quad (7)$$

Writing $Q_s' = \omega l_s'/r_s'$, equation (6) may be rewritten:

$$l_s' \cdot \frac{\omega^2 M'^2}{r_s'^2 + \omega^2 l_s'^2} = (L_1' - L_2')$$

or:

$$M'^2 = l_s'(L_1' - L_2')\left\{1 + 1/Q_s'^2\right\} \quad \ldots\ldots \quad (8)$$

Usually $Q_s' \gg 1$, then:

$$M'^2 = l_s'(L_1' - L_2')$$

List of Symbols Used

- C_p = effective primary capacitance of meas. trans.
- C_T = tuning capacitance
- C_x = effective series capacitance of test impedance
- L, L' = inductive component of Z_i, Z_i
- l_l = inductance of short-circuiting link
- l_p, l_s = primary, secondary inductance of meas. trans. (low freq. eq. cct.)
- l_p', l_s' = effective primary, secondary inductance of meas. trans. (high freq. eq. cct.)
- l_s'' = $l_s - l_l$
- l_x = effective series inductance of test impedance
- l_x' = $l_x + l_s''$
- L_T = total effective series inductance of Q meter meas. cct.
- M = mutual inductance of meas. trans. (low freq. eq. cct.)
- M' = effective mutual inductance of meas. trans. (high freq. eq. cct.)
- Q, Q' = ratio of $\dfrac{\text{reactive component of } Z_i, Z_i'}{\text{resistive component of } Z_i, Z_i'}$
- Q_I = Q meter reading (i.e. valve-voltmeter indication)
- Q_s, Q_s' = $\omega l_s/r_s, \omega l_s'/r_s'$
- Q_T = $\omega L_T/R_T$
- Q_x' = $\omega l_x'/r_x'$
- R, R' = resistive component of Z_i, Z_i'
- r_p, r_s = primary, secondary resistance of meas. trans. (low freq. eq. cct.)
- r_p', r_s' = effective primary, secondary resistance of meas. trans. (high freq. eq. cct.)
- R_T = total effective series resistance of Q meter meas. cct.
- r_x = effective series resistance of test impedance
- r_x' = $r_x + r_s$
- Z_i, Z_i' = input impedance to meas. trans., excluding C_p including C_p

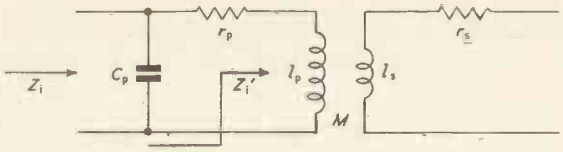


Fig. 3. Equivalent circuit of a measuring transformer

Likewise equation (7) may be rewritten:

$$r_s'/\omega^2 l_s'^2 \cdot \frac{\omega^2 M'^2}{(1 + 1/Q_s'^2)} = (\omega L_2'/Q_2') - (\omega L_1'/Q_1')$$

or:

$$Q_s' = \frac{Q_1'(L_1'/L_2' - 1)}{(Q_1'/Q_2' - L_1'/L_2')} \quad \ldots\ldots\ldots\ldots \quad (9)$$

Assuming l_s' is measured from the secondary terminals, the quantities r_p', l_p', M' and r_s' may be measured from the primary terminals and calculated using equations (3), (4), (8) and (9).

The measuring transformers required for this work employ secondary windings which are always smaller than the primary windings in physical dimensions and inductance value. It is permissible to ignore all residual capacitances except that shunting the primary winding; it is then possible to use the low frequency parameters of the mutual inductor and deal with the remaining residual capacitance in the following way. With reference to Fig. 3, it is taken that Z_i' is measured and that Z_i is required assuming that C_p is known.

If $Z_i' = R' + j\omega L_1'$, $Q' = \omega L'/R'$ and $Z_i = R + j\omega L$, $Q = \omega L/R$; when $Q' \gg 1$, it may be shown that:

$$R = \frac{R'}{(1 + \omega^2 L' C_p)^2} \quad ; \quad L = \frac{L'}{(1 + \omega^2 L' C_p)^2}$$

If C_T is the tuning capacitance required to resonate L', viz.:

$$\omega^2 = (1/L'C_T)$$

then:

$$R = \frac{R'}{(1 + C_p/C_T)^2} \quad ; \quad L = \frac{L'}{(1 + C_p/C_T)} \quad ; \quad Q = Q'(1 + C_p/C_T)$$

More complex relationships are required when Q' is not much greater than 1 and these are discussed in Appendix 2.

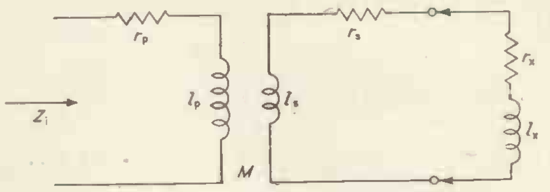
Theory of the Measurement

The test impedance is treated as a resistance r_x in series with an inductance l_x. If the unknown is effectively capacitive, it is measured as a negative inductance of numerical value:

$$(1/\omega^2 C_x)$$

The method assumes that C_p and l_s are known and that Z_i' is measured with the Q meter, Z_i being calculated as shown above. With reference to Fig. 4 (from which C_p has been withdrawn), three measurements are undertaken at

Fig. 4. Modified equivalent circuit of measuring transformer



the primary terminals as follows.

Measurement (1). *Secondary Open-Circuit*

$$Z_i = r_p + j\omega l_p \quad \dots\dots\dots\dots\dots \quad (10)$$

$$L_1 = l_p \quad \dots\dots\dots\dots\dots \quad (11)$$

$$Q_1 = (\omega L_1 / r_p) \quad \dots\dots\dots\dots\dots \quad (12)$$

Measurement (2). *Secondary Short-Circuit*

$$Z_i = r_p + r_s \cdot \frac{\omega^2 M^2}{r_s^2 + \omega^2 l_s^2} + j\omega \left\{ l_p - l_s \left[\frac{\omega^2 M^2}{r_s^2 + \omega^2 l_s^2} \right] \right\} \quad (13)$$

$$L_2 = l_p - l_s \cdot \frac{\omega^2 M^2}{r_s^2 + \omega^2 l_s^2} \quad \dots\dots\dots\dots \quad (14)$$

$$Q_2 = \frac{\omega L_2}{r_p + r_s \cdot \dfrac{\omega^2 M^2}{r_s^2 + \omega^2 l_s^2}} \quad \dots\dots\dots \quad (15)$$

Measurement (3). *Secondary closed with Test Impedance* $Z_x = r_x + j\omega l_x$

Writing $r_x' = r_x + r_s$, $l_x' = l_x + l_s''$, where $l_s'' = l_s - l_l$

$$Z_i = r_p + r_x' \cdot \frac{\omega^2 M^2}{r_x'^2 + \omega^2 l_x'^2} + j\omega \left\{ l_p - l_x' \left[\frac{\omega^2 M^2}{r_x'^2 + \omega^2 l_x'^2} \right] \right\} \quad \dots\dots\dots\dots \quad (16)$$

$$L_3 = l_p - l_x' \cdot \frac{\omega^2 M^2}{r_x'^2 + \omega^2 l_x'^2} \quad \dots\dots\dots \quad (17)$$

$$Q_3 = \frac{\omega L_3}{r_p + r_x' \cdot \dfrac{\omega^2 M^2}{r_x'^2 + \omega^2 l_x'^2}} \quad \dots\dots\dots \quad (18)$$

The quantities L_1, L_2, L_3 and Q_1, Q_2, Q_3 are determined from Q meter measurements. Assuming the transformer is designed to have $Q_s \gg 1$, M is given by:

$$M^2 = l_s(L_1 - L_2) \quad \dots\dots\dots\dots \quad (19)$$

Putting $Q_x' = (\omega l_x' / r_x')$, equations (17) and (18) may be rewritten:

$$\omega l_x' / \omega l_s \cdot Q_x'(1 + 1/Q_x'^2) = a = \frac{Q_1(1 - L_2/L_1)}{(L_3 Q_1 / L_1 Q_3 - 1)} \quad (20)$$

$$(\omega l_x' / \omega l_s)\ (1 + 1/Q_x^2) = b = \frac{(1 - L_2/L_1)}{(1 - L_3/L_1)} \quad \dots \quad (21)$$

Dividing these equations, which contain measured quantities only on the right-hand side:

$$Q_x' = a/b$$

Substituting for Q_x' in equation (21):

$$\omega l_x' = \omega l_s \cdot \frac{b}{(1 + b^2/a^2)} = \frac{\omega l_s}{(1 + b^2/a^2)} \cdot \frac{(1 - L_2/L_1)}{(1 - L_3/L_1)}$$

As $l_x' \to l_s$, so $L_3 \to L_2$ (when $Q_x' \gg 1$) and the above equation is more appropriately written:

$$l_x' = l_s \cdot \frac{1}{(1 + b^2/a^2)} \left\{ 1 + \frac{L_{32}}{L_1 - L_3} \right\} \quad \dots\dots \quad (22)$$

where $L_{32} = L_3 - L_2$.

l_s is the inductance of the secondary circuit when closed by the short-circuiting link. Assume that there is negligible mutual inductance between the secondary circuit and the link or test impedance. If l_l is the inductance of the link, then:

$$l_x' = l_x + l_s - l_l$$

or:

$$l_x = l_x' - l_s + l_l$$

$$= \frac{l_s}{(1 + b^2/a^2)} \cdot \left\{ \frac{L_{32}}{L_1 - L_3} - b^2/a^2 \right\} + l_l \quad \dots\dots \quad (23)$$

Where a low loss component is being tested, it is frequently found that $b^2/a^2 \to 0$ and the expression simplifies to:

$$l_x = l_s \left\{ \frac{L_{32}}{L_1 - L_3} \right\} + l_l$$

This expression is independent of ω and may be written in terms of the appropriate tuning capacitance values for each inductance.

The expression for r_x is not pursued since (1) it is not the object of the measurement, and (2) it may be highly dependent upon the contact resistance between the test impedance and the secondary circuit.

Experimental Procedure

The quantities L_1, L_2, L_3 and Q_1, Q_2, Q_3 are determined from the Q meter readings of tuning capacitance and indicated Q factor, Q_I. Assuming $Q_I \gg 1$, L_1 (etc.) is given by the expression:

$$1 = \frac{1}{\omega^2 L_1' C_T} = \frac{1}{\omega^2 L_1 (C_T + C_p)}$$

To obtain Q_1 (etc.) from Q_{I_1} (etc.), the residual impedances in the Q meter measuring circuit have to be taken into account. This conversion may be rapidly undertaken by using a procedure based on Q-type correction factors developed by the author[1].

It is necessary to provide a means of determining precisely the resonant condition. This may be accomplished by using a sensitive differential (or "backed-off") indicating instrument in the valve-voltmeter indicating instrument circuit.

It is important to proceed very carefully from measurement (2) to measurement (3). The difference between L_3 and L_2 may be extremely small and it is necessary to measure $(C_{T_2} - C_{T_3})$ with an incremental capacitor, the main tuning capacitor and frequency adjustment being untouched during this period.

To use equation (23), l_s and l_l must be known. l_s may be obtained by (1) a direct series substitution measurement in the Q meter at a frequency well removed from the self resonance of the primary circuit; (2) carrying out a measurement with a known value of l_x.

l_l is of the order of a correction factor and may be approximately calculated; it should be as small as possible.

Restrictions in the Measurement of Capacitive Reactance

When a capacitive reactance is under measurement, equation (17) may be written:

$$\omega L_3 = \omega l_p - (\omega l_s'' - 1/\omega C_x) \cdot \frac{\omega^2 M^2}{r_x'^2 + (\omega l_s'' - 1/\omega C_x)^2}$$

A certain range of values of C_x give rise to internal resonances in the transformer; under these conditions, L_3 may have a very high or low value and is negative over a certain range of values. The value of Q_3 under such conditions is always very low and these circumstances should always be avoided. For any transformer, the "forbidden range" of C_x values should be estimated; it is useful to plot L_3 against C_x at the upper and lower operating frequencies of the measuring transformer. Critical frequencies are as follows. (1) L_3 has positive and negative maxima at values of C_x given by:

$$1/\omega C_x = \omega l_s'' \pm r_x'$$

(2) L_3 is zero at values of C_x given by:

$$1/\omega C_x = \omega l_s'' \left\{ 1 - \frac{M^2}{2 l_s'' l_p} \left[1 \pm \sqrt{1 - \frac{4 l_p^2 r_x'^2}{\omega^2 M^4}} \right] \right\}$$

Where r_x' is small, the forbidden range is approximately given by putting r_x' to zero; the range then lies between:

$$C_x = (1/\omega^2 l_s'')$$

and:

$$C_x = \frac{1}{\omega^2(l_s'' - M^2/l_p)}$$

Procedure for the Design of a Measuring Transformer

For design purposes, it is useful to assume that $\omega l_x' \gg r_x'$. Writing:

$$M^2 = k^2 l_p l_s$$

where k is the coefficient of coupling, equations (14) and (7) simplify to:

$$L_2 = L_1 - M^2/l_s$$

$$L_3 = L_1 - M^2/(l_s + l_x)$$

Whence:

$$l_x = \frac{l_s}{\left\{\dfrac{k^2 l_p}{L_3 - L_2}\right\} - 1}$$

For l_x to be a minimum:

(1) l_s should be a minimum; the lower limit of l_s is fixed by the fact that it should be measurable by direct series substitution at a frequency much less than the self-resonant frequency of the primary circuit.

(2) k should be high, but not so high that Q_2, Q_3 become excessively low.

(3) l_p should be as high as possible.

(4) $(L_3 - L_2)$ should be as small as possible; this infers that C_{T2} should be as large as possible and $(C_{T2} - C_{T3})$ as small as possible.

Details of an Experimental Transformer

Details are to be given of a unit, which has been constructed and successfully operated by the author to the following specification. The measuring system to be capable of measuring $l_{x(min)} = \cdot 003\mu$H at 1Mc/s (viz. 1/100th of the value measurable by direct series substitution); the maximum tuning capacitance available is 900pF and the minimum value of incremental tuning capacitance permissible is 10pF.

Design Details

(1) A suitable minimum value for l_s is approximately $0{\cdot}5\mu$H.

(2) The smallest inductance to be measured in a test is L_2 (assuming l_x to be positive). Taking maximum tuning capacitance to be used, then $L_2 = 28{\cdot}1\mu$H.

(3) Taking $k = 0{\cdot}7$ gives:

$$l_p = \frac{28{\cdot}1}{1 - (0{\cdot}7)^2} = 55{\cdot}1\mu\text{H}$$

(4) If $C_p \simeq 6$pF, then $C_{T(min)} \simeq 60$pF is appropriate, since it is undesirable to operate the primary circuit too close to self-resonance. The largest inductance to be measured in a test is $L_1 = l_p$; at the highest test frequency $C_T = C_{T(min)}$, so that the maximum operating frequency is given by:

$$\frac{1}{2\pi\sqrt{(55{\cdot}1 \times 10^{-6} \times 60 \times 10^{-12})}} = 2{\cdot}6\text{Mc/s}$$

Construction

The Q meter used by the author is a Marconi Instruments Ltd. type TF329G and the measuring transformer in question was built into the case of an inductor type TM1438 (a range manufactured for use with this instrument). The unit is shown in Fig. 5 with the screening can removed. The primary winding consists of 100 turns of 3/3/3/3/46 s.w.g. enamelled copper wire wound in five sections on a 1¼in diameter polystyrene centre column. The secondary

Fig. 5. Measuring transformer with screening can removed

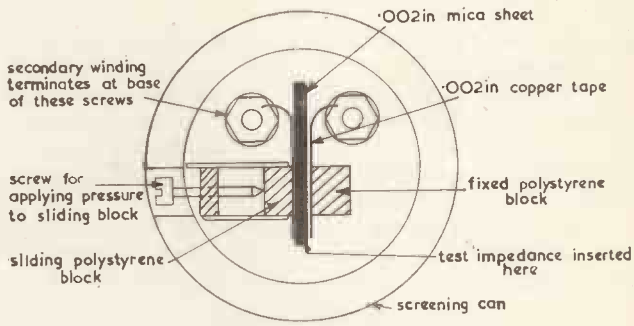


Fig. 6. Plan view of measuring transformer

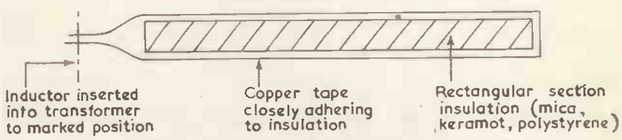


Fig. 7. Construction of "reference inductors"

winding of 4 turns of 18 s.w.g. enamelled copper wire is concentric with, but makes no contact with the primary winding. The terminating arrangements of the secondary winding are shown in Fig. 6. The short-circuiting link of copper tape was U-shaped and held by a small piece of polystyrene. It was inserted into the secondary winding termination in a standardized manner and was estimated to introduce an inductance of $160\mu\mu$H.

Testing

The measuring system was checked against "reference inductors", which were constructed as shown in Fig. 7.

The features of these units are:

(1) A low value of the conductor thickness/insulation thickness ratio; in this way skin and proximity effects have little effect on the inductance value, which may be calculated from formulæ based on uniform current distribution.

(2) A low value for the insulation of the width/length ratio; in this way the inductance contribution of the ends is made small.

The inductance of these units may be estimated from established formulæ, such as those given by Grover[4]. Units constructed by the author ranged from ·005 to ·05μH and used 0·002in by $\frac{1}{8}$in copper tape as the conductor.

Conclusion

This method of very low reactance measurement is considered significant as no ready alternative exists. The transformers used are simple in construction and it is therefore quite reasonable to make up a unit for a particular measurement.

Measuring transformers may also be applied to the measurement of medium value resistance with the Q meter. It is well known that a gap exists between low values of resistance measurable by series substitution and high values measurable by shunt substitution. This gap may be closed by using a measuring transformer.

Acknowledgments

The author wishes to thank Professor J. E. Parton, Ph.D., M.I.E.E., for facilities afforded him in connexion with this work.

Appendix

(1) Behaviour of the Q Meter when Q_T is not much greater than 1

It is shown in reference 1 that when Q_T is not much greater than 1, it is important to consider the conditions existing at resonance with care. Many commercial Q meters do not measure values of $Q_T < 25$ to any significant accuracy, they may be adapted to do so.

With reference to Fig. 1, assuming that R' is the major loss resistance in the circuit and that resonance is approached by variation of C_T, then at resonance:

$$\omega^2 = (1/L_T C_T) \cdot \frac{1}{1 + 1/Q_T{}^2}$$

and the valve-voltmeter indication, Q_I, is given by:

$$Q_I = \frac{\text{Input voltage to valve-voltmeter}}{\text{e.m.f. injected into measuring circuit}} = \sqrt{(1 + Q_T{}^2)}$$

where $Q_T = \omega L_T / R_T$.

L_T and R_T are respectively the total effective series inductance and resistance in the measuring circuit. Q_T is converted to Q', where:

$$Q' = \omega L'/R'$$

by eliminating the residual circuit impedances.

(2) Effect of Self-capacitance on the Behaviour of a Low Q Factor Inductor

Section 2 of the paper[1] deals with the evaluation of Z_l from Z_l' and C_p for the case when $Q' \gg 1$. When Q' is not much greater than 1 (due to the prominence of R and not C_p) a more complex relationship is required.

With reference to Fig. 8, equating the impedances of the two circuits:

$$R' = \frac{R}{(1 - \omega^2 L C_p)^2 + \omega^2 C_p{}^2 R^2}$$

Fig. 8. Effect of self-capacitance

$$L' = \frac{L(1 - \omega^2 L C_p) + C_p R^2}{(1 - \omega^2 L C_p)^2 + \omega^2 C_p{}^2 R^2}$$

Dealing with the conversion $L' \to L$: when C_p is small, then:

$$\omega^2 C_p{}^2 R^2 \ll (1 - \omega^2 L C_p)^2$$

so that:

$$L' = \frac{L}{(1 - \omega^2 L C_p)^2} + \frac{C_p \cdot R^2}{(1 - \omega^2 L C_p)^2}$$

The quantities:

$$\left(\frac{1}{1 - \omega^2 L C_p}\right) \quad ; \quad \frac{C_p \cdot R^2}{(1 - \omega^2 L C_p)^2}$$

are correction factors and may be approximated as follows:

$$\frac{1}{1 - \omega^2 L C_p} = (1 + \omega^2 L' C_p) \quad \text{(very nearly)}$$

$$\frac{C_p R^2}{(1 - \omega^2 L C_p)^2} = \frac{\omega^2 L'^2 C_p}{Q^2} \quad \text{(very nearly)}$$

Thus:

$$L' = L(1 + \omega^2 L' C_p) + \frac{\omega^2 L'^2 C_p}{Q^2}$$

or:

$$L = \frac{1}{(1 + \omega^2 L' C_p)}\left\{L' - \frac{\omega^2 L'^2 C_p}{Q^2}\right\}$$

Dealing with the conversion $Q' \to Q$.

$$Q' = \omega L'/R' = \frac{\omega L(1 - \omega^2 L C_p) + \omega C_p R^2}{R}$$

$$= Q(1 - \omega^2 L C_p) + \frac{\omega^2 L C_p}{Q}$$

or:

$$0 = Q^2 - Q \cdot \frac{Q'}{(1 - \omega^2 L C_p)} + \frac{\omega^2 L C_p}{(1 - \omega^2 L C_p)}$$

giving:

$$Q = \frac{Q'}{(1 - \omega^2 L C_p)}\left[1 - \frac{\omega^2 L C_p (1 - \omega^2 L C_p)}{Q'^2}\right] \text{(very nearly)}$$

Writing:

$$(1 - \omega^2 L C_p) = \frac{1}{(1 + \omega^2 L' C_p)}$$

and:

$$\omega^2 L C_p = \frac{\omega^2 L' C_p}{1 + \omega^2 L' C_p}$$

gives finally:

$$Q = Q'(1 + \omega^2 L' C_p)\left\{1 - \frac{\omega^2 L' C_p}{(1 + \omega^2 L' C_p)\, Q'^2}\right\}$$

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The Response Functions and Vector Loci of First and Second Order Systems

By David Morris*, D.Sc., A.M.I.E.E.

(Part 3)

An Asymmetrically-Resonant Response and a Cordiform Locus

In this article certain passive and active second order systems are considered which have a finite response at zero frequency and an anti-phase zero-limit at infinite frequency. The vector locus for such systems has a characteristic shape, and its properties are analysed.

Fig. 12 illustrates two passive second-order systems, which have a similar form of response. In Fig. 12(a), if the input X is the displacement given to the point A, and the output Y is the resulting displacement of the mass M, we may write for sinusoidal conditions:

$$F = (X - Y)K = (j\omega B + j^2\omega^2 M)Y \quad \ldots\ldots (23)$$

where F is the force in the spring K. Hence:

$$\text{Response} = (Y/X) = \frac{K}{K + j\omega B + j^2\omega^2 M} \quad \ldots (24)$$

Set $T_0 = 1/\omega_0 = \sqrt{(M/K)}$, and $Q = (1/B)\sqrt{(MK)} = K/\omega_0 B$, and also let γ = relative frequency $= \omega/\omega_0 = \omega T_0$. Then:

$$Y/X = \frac{1}{1 + j\gamma/Q + j^2\gamma^2} = \frac{Q}{(1-\gamma^2)Q + j\gamma} \quad \ldots (25)$$

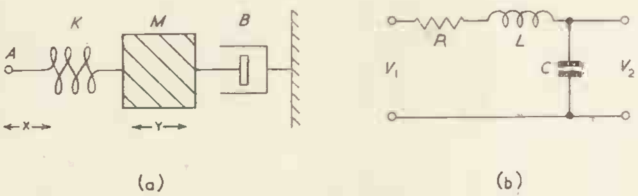

Fig. 12. Two passive second-order systems

The same expression serves as the transfer function V_2/V_1 for the circuit of Fig. 12(b), if we set $T_0 = \sqrt{(LC)}$ and $Q = (1/R)\sqrt{(L/C)}$. Further examples of systems having the same responses are the two active systems of Fig. 13. A servo-mechanism damped by velocity feedback and/or viscous friction, as illustrated in Fig. 13(a), has a response as given by equation (25), provided that the time-lag in the amplifier is negligible. A similar expression describes the behaviour of any d.c. amplifier having two stages of lag, when used with or without simple feedback. An example is provided by the current gain I_3/I_1 of the dynamo-electric amplifier of Fig. 13(b), provided that non-linearities and parasitic brush-currents have negligible effect. For this system, the parameters of equation (25) are[5]:

$$T_0 = \sqrt{\frac{T_2 T_3}{1 + K_{23}K_{32}}} \quad \ldots\ldots\ldots\ldots\ldots (26(a))$$

and:

$$Q = \frac{\sqrt{(T_2 T_3)}}{T_2 + T_3}\sqrt{(1 + K_{23}K_{32})} \quad \ldots\ldots (26(b))$$

where T_2 and T_3 are the time-constants of the armature circuits and $K_{23}K_{32}$ is the feedback factor.

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If the input displacement in the system of Fig. 12(a) is varied in such a manner as to maintain the output displacement constant, the necessary input in terms of the output is given by the function:

$$X/Y = 1 + j\gamma/Q + j^2\gamma^2 = (1 - \gamma^2) + j\gamma/Q \quad \ldots\ldots (27)$$

The frequency-dependent functions (25) and (27) are dimensionless, and have unity value at zero frequency.

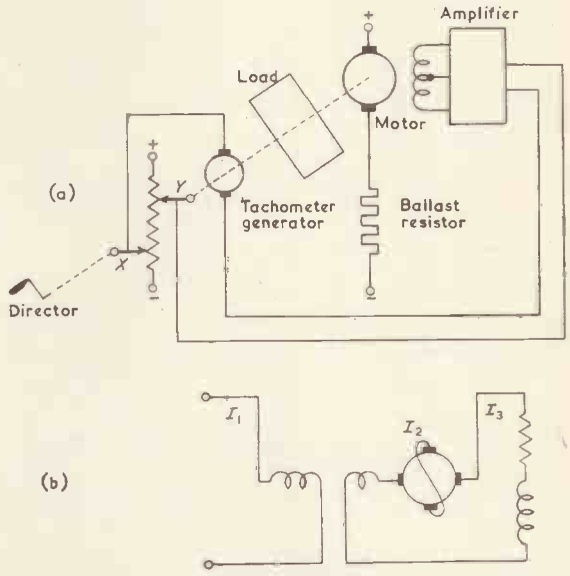


Fig. 13. Two active systems having second order responses

The Nature of the Resonance

From equation (25) we may write:

$$|Y/X|^2 = \frac{1}{(1-\gamma^2)^2 + (\gamma/Q)^2} = \frac{1}{\gamma^4 - \dfrac{2Q^2 - 1}{Q^2}\gamma^2 + 1} \quad \ldots\ldots\ldots\ldots (28)$$

Differentiating with respect to γ^2 we find that this has a maximum value of:

$$|Y/X|^2_{max} = \frac{4Q^4}{4Q^2 - 1} \quad \ldots\ldots\ldots\ldots (29(a))$$

when:

$$\gamma^2 = \frac{2Q^2 - 1}{2Q^2} = \gamma_0^2 \quad \ldots\ldots\ldots\ldots (29(b))$$

From equation (29(b)) it is seen that there is no peak in the response magnitude unless $Q > \frac{1}{2}\sqrt{2}$. Moreover, since the magnitude tends to zero for very large values of γ, but does not tend to zero for very small values of γ, the curve of response magnitude is not symmetrical when plotted against $\log\gamma$.

If we set $\Gamma = \gamma/\gamma_0$ it is easily shown that:

$$|Y/Y_{\max}|^2 = \frac{1}{1 + \dfrac{(2Q^2 - 1)^2}{4Q^2 - 1}[\Gamma^2 - 1]^2} \quad \ldots\ldots\ (30)$$

The curve of response magnitude is thus symmetrical when plotted against an arithmetic scale Γ^2, γ^2 or ω^2. A comparison of function (28) plotted against γ and against γ^2 is shown in Fig. 14. A point of interest is that the initial slope ($\gamma = 0$) is zero in the first case, but not in the second.

If an attempt is made to measure Q by the bandwidth method, the frequencies to give a response magnitude $\frac{1}{2}\sqrt{2}$ of the peak value can be found from equation (30), and are given by the expression:

$$\Gamma^2 = 1 \pm \frac{\sqrt{(4Q^2 - 1)}}{2Q^2 - 1} \quad \ldots\ldots\ (31)$$

If Γ_1 and Γ_2 are the values derived from equation (31), the apparent value of Q is:

$$Q' = \frac{1}{\Gamma_2 - \Gamma_1} \quad \ldots\ldots\ (32)$$

By substitution of numerical values in equations (31) and

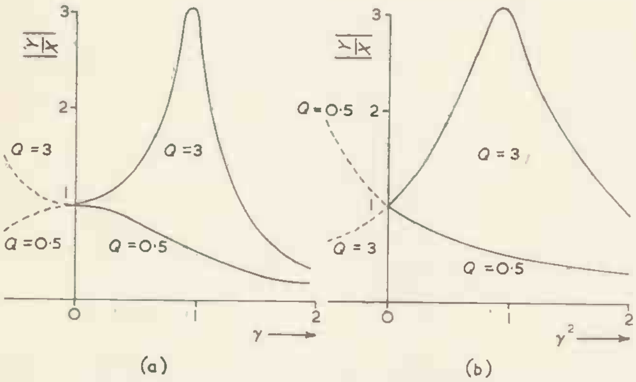


Fig. 14. Two methods of plotting the resonance curve

(32) we can relate actual and apparent values of Q as in Table 1.

TABLE 1

Q	1·308	3	5	10
Q'	0·707	2·83	4·72	9·96
Q'/Q	0·54	0·95	0·985	0·996

For values of $Q < 1·308$ the measurement cannot be made since Γ_1 is imaginary. The table shows that even though the resonance is not symmetrical, the bandwidth method for determining Q can be applied within certain limits of approximation.

The Vector Loci of the Asymmetrically-Resonant System

The behaviour of the function X/Y as ω varies from zero to infinity may be represented by a vector plot of $(1 - \gamma^2) + j\gamma/Q$ as γ varies from zero to infinity. As the variation of the real part is proportional to the square of the imaginary part the locus is a parabola, the axis of which is the axis of reals. The double intercept on the imaginary axis is $2/Q$, and the vertex is at unit distance from the origin. When $Q = \frac{1}{2}$, these two distances are in the ratio 4:1, and the origin is then the focus of the parabola.

The behaviour of the function Y/X may be represented by a vector plot of the function $\dfrac{1}{(1 - \gamma^2) + j\gamma/Q}$ as γ varies from zero to infinity. The loci are shown in Fig. 15 for values of Q equal to 0, $\frac{1}{2}$, 1, 2 and 3. Loci of constant-γ

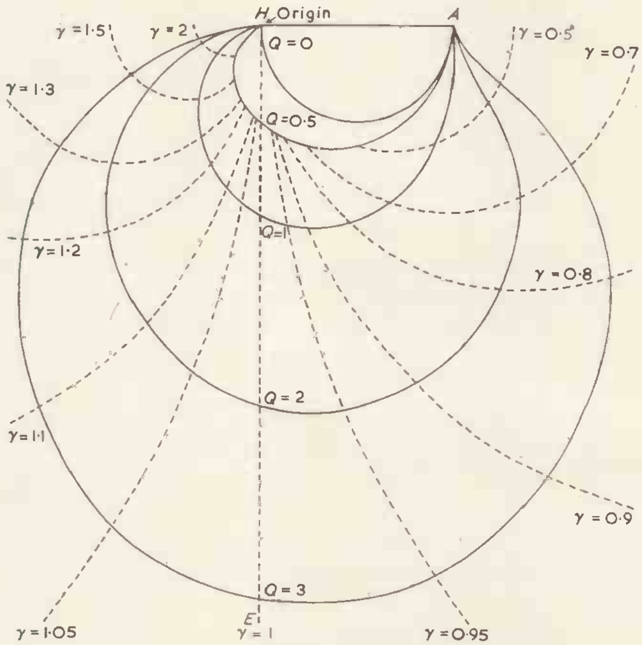


Fig. 15. The cordiform loci for various values of Q

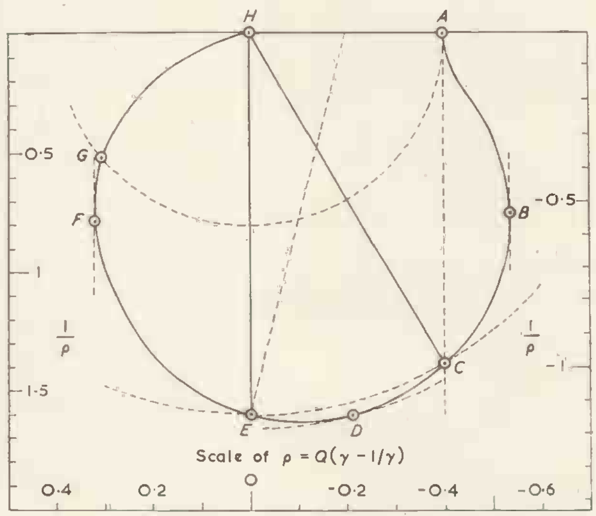


Fig. 16. Special points on the cordiform locus

are shown by dotted lines. Inspection of the above function shows that if Q varies while γ is held constant, a circle of radius $\dfrac{1}{2(1 - \gamma^2)}$ is obtained, the zero-Q point being at the origin. The Q value of any constant-Q locus is very easily determined by inspection because when the response has zero real part (i.e. $\gamma = 1$), the quadrature response is Q times the zero-frequency response.

The loci are the curves obtained by inverting the above-mentioned parabola about a point on its axis. There appears to be no name to characterize this class of curves.

TABLE 2

Details of Special Points on the locus $\dfrac{1}{1 + j\,\gamma/Q + j^2\,\gamma^2}$

POINT	γ^2	DEFINITION OF POINT	MAGNITUDE[2]	$-\tan\theta$
A	0	Response magnitude unity In-phase response unity	1	0
B	$1 - \dfrac{1}{Q}$	In-phase response at positive maximum	$\dfrac{Q^3}{2Q-1}$	$\sqrt{\left[1-\dfrac{1}{Q}\right]}$
C	$1 - \dfrac{1}{Q^2}$	Response magnitude Q In-phase response unity	Q^2	$\sqrt{(Q^2-1)}$
D	$1 - \dfrac{1}{2Q^2}$	Response magnitude maximum	$\dfrac{4Q^4}{4Q^2-1}$	$\sqrt{(4Q^2-2)}$
E	1	Response magnitude Q In-phase response zero	Q^2	∞
F	$1 + \dfrac{1}{Q}$	In-phase response at negative maximum	$\dfrac{Q^3}{2Q+1}$	$-\sqrt{\left[1+\dfrac{1}{Q}\right]}$
G	$2 - \dfrac{1}{Q^2}$	Response magnitude unity	1	$\dfrac{\sqrt{(2Q^2-1)}}{1-Q^2}$
H	∞	Infinite frequency point	0	0

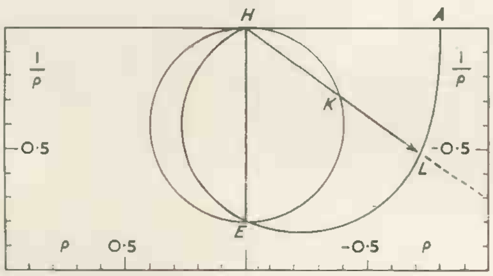
In the more general case, the inverse of a parabola about any point in its plane is a "bicircular quartic", which degenerates to a "bicircular cubic" when the centre of inversion lies on the curve. In the more particular cases, the inverse of a parabola about its vertex is a "cissoid", while the inverse about its focus is a "cardioid". Half of the latter curve is illustrated by the case $Q = \frac{1}{2}$ in Fig. 15. The complete curve, including the section representing negative values of γ, has the conventional heart shape from which it derives its name. For the purposes of these articles, the constant-Q curves will be referred to as the "cordiform loci" (i.e. heart-shaped loci), although for high values of Q the heart must be regarded as being in a somewhat unhealthy condition.

The loci of constant-γ can be calibrated in Q by means of linear scales of Q and $1/Q$, analogous to the linear scales of γ and $1/\gamma$ described in connexion with the first-order systems in Part 1. Further properties of the cordiform locus are illustrated in Fig. 16, which is drawn for the particular case $Q = 2$. The locus can be quickly sketched if it is remembered that there are two points E and C for which the response magnitude is Q, the in-phase component of response being zero and unity respectively. The direction of the locus at E is also given by the fact that the normal to the curve at this point passes through the point $+\frac{1}{2}$ on the real axis. The in-phase component of the response is at a positive and a negative maximum at the points B and F, for which $\gamma = \sqrt{(1 \mp (1/Q))}$. At these points the response vector is:

$$Y/X = \frac{\pm Q^2}{2Q \mp 1} - j\,\frac{Q^{3/2}\sqrt{(Q \mp 1)}}{2Q \mp 1} \quad \ldots\ldots (33)$$

When $Q > 1$, there is a point of inflexion between points A and B, for which $\gamma^2 = -1 + \sqrt{[(4Q^2 - 1)/3Q^2)]}$. Further details of special points are given in Table 2.

Fig. 17. Derivation of cordiform locus from circular locus



The Frequency Calibration of the Locus

Equation (25) may be re-written as:

$$Y/X = -j\,(Q/\gamma)\,.\,\frac{j\gamma/Q}{1 + j\gamma/Q + j^2\gamma^2} \quad \ldots\ldots (34)$$

Apart from an additional constant lag of 90°, the phase characteristics of this response are identical with those of the symmetrically-resonant system that has been examined in Part 2. Scales of $p = \frac{1}{2}(\gamma - 1/\gamma)$ and $1/p$ can therefore be inserted as shown in Fig. 16. Moreover, if ω_1, ω_0 and ω_2 are the frequencies that give 45°, 90° and 135° lag respectively, then $Q = (\omega_0/(\omega_2 - \omega_1))$ precisely.

Furthermore, if a circle of diameter $Q = HE$ is drawn, the cordiform locus can be derived from the circle by making $HL = HK/\gamma$. For large values of Q a major portion of the circle HKE is traversed for a very small variation in γ. The correction KL is therefore small except at extreme values of frequency, and the real half of the cordiform locus therefore takes on an almost circular appearance, as seen in Fig. 17.

Other responses of second order systems may be represented by other cordiform loci, and the relationships among these will be considered in Part 4.

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Effects of Heater-Voltage Variations of Valves in Certain Stabilizer Circuits

By F. A. Benson*, M.Eng., Ph.D., A.M.I.E.E., M.I.R.E., and G. V. G. Lusher*, B.Eng., Ph.D.

Mathematical analyses have been made to determine the effects of varying heater voltage of valves on the stability of several series-parallel valve stabilizer circuits in which a resistor is placed in the cathode lead of the amplifier valve. It is evident from the calculations that analyses assuming constant heater voltage are not perfectly adequate, but the effects of heater-voltage variations are not as serious as might be expected, because the two valves produce changes in output voltage of opposite sign.

CALCULATIONS have previously been carried out by the authors[1] on several voltage-stabilizer circuits for operating microwave oscillators to find the effects of heater-voltage variations of the valves on stability. Additional calculations have now been made on a number of series-parallel valve circuits in which a resistor is included in the cathode lead of the amplifier valve. In such cases the effective grid base of the amplifier valve is increased. In this way the range over which stabilization is obtained may be extended. Some useful circuits of this type, which do not appear to be well-known, have been investigated by Oxbrow[2] but, as is common, he neglects the effects of heater-voltage variations of the valves in his calculations to determine the performance of the circuits. It is with these stabilizers that the present article is concerned and the formulæ developed should prove of considerable value.

The same method of calculation is used as in the previous analyses[1]. It has been shown[3] that the anode voltage V_a, the grid voltage V_g and the anode current I_a of a triode are related by the expression:

$$V_a = r_a I_a - \mu V_g - c r_a \quad \ldots\ldots\ldots\ldots (1)$$

where μ is the amplification factor of the valve,
r_a is the anode resistance of the valve,
and c is a constant which for many triodes is normally nearly zero.

Thus, for each circuit examined an expression can be obtained for the output voltage in terms of several known values and certain quantities (c's) which are functions of heater voltages. Therefore, providing it is known how the various c's depend on heater voltage (which can easily be determined experimentally for the valves used) the ratio of the percentage change of output voltage to the percentage change of heater voltage can be found for a circuit.

Analysis of Fig. 1

Let the currents and voltages be as shown and neglect current I_2 in calculations for simplification.

With linear valve characteristics for the parallel valve:

$$V_{ap} = I_1 r_{ap} - \mu_p V_{gp} - c_p r_{ap} \quad \ldots\ldots\ldots\ldots (2)$$

Also:

$$V_{EF} = V_i - R_1 (I_1 + I_L) \quad \ldots\ldots\ldots\ldots (3)$$

$$V_{gp} = \frac{R_3}{(R_2 + R_3)} V_{EF} - v - I_1 R_6 \quad \ldots\ldots\ldots (4)$$

and:

$$V_{ap} = V_{EF} - I_1(R_5 + R_4) - I_1 R_6 - v \quad \ldots\ldots (5)$$

With linear valve characteristics for the series valve:

$$V_{as} = I_L r_{as} - \mu_s V_{gs} - c_s r_{as} \quad \ldots\ldots\ldots\ldots (6)$$

Also:

$$V_{gs} = V_{as} - R_4 I_1 \quad \ldots\ldots\ldots\ldots (7)$$

and:

$$V_{as} = V_{EF} - I_L R_L \quad \ldots\ldots\ldots\ldots (8)$$

From equations (2) and (5) eliminating V_{ap}:

$$I_1 r_{ap} - \mu_p V_{gp} - c_p r_{ap} = V_{EF} - I_1(R_4 + R_5) - I_1 R_6 - v \quad \ldots (9)$$

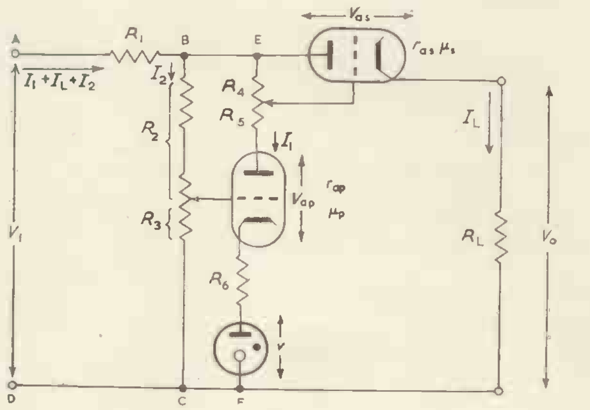

Fig. 1.

From equations (9) and (4):

$$I_1 r_{ap} + \mu_p v + \mu_p I_1 R_6 - c_p r_{ap} = V_{EF}\left(1 + \frac{\mu_p R_3}{R_2 + R_3}\right) - I_1(R_4 + R_5 + R_6) - v \quad \ldots\ldots\ldots\ldots (10)$$

From equations (10) and (3):

$$I_1 = \frac{(V_i - R_1 I_L)\left(1 + \dfrac{\mu_p R_3}{(R_2 + R_3)}\right) - v(1 + \mu_p) + c_p r_{ap}}{r_{ap} + \mu_p R_6 + R_4 + R_5 + R_6 + R_1\left(1 + \dfrac{\mu_p R_3}{(R_2 + R_3)}\right)} \quad (11)$$

From equations (6), (7) and (8):

$$V_{EF} - I_L R_L = I_L r_{as} - \mu_s(V_{EF} - I_L R_L - R_4 I_1) - c_s r_{as} \quad \ldots (12)$$

Using equation (3):

$$I_1 = \frac{V_i(1 + \mu_s) + c_s r_{as} - I_L(R_1 + R_L + r_{as} + R_1\mu_s + R_L\mu_s)}{R_1 + \mu_s R_1 + \mu_s R_4} \quad (13)$$

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From equations (11) and (13) eliminating I_1:

$$\frac{(V_i - R_1 I_L)\left(1+\frac{\mu_p R_3}{R_2+R_3}\right) - v(1+\mu_p) + c_p r_{ap}}{r_{ap}+\mu_p R_6 + R_4 + R_5 + R_6 + R_1\left(1+\frac{\mu_p R_3}{R_2+R_3}\right)} = \frac{V_i(1+\mu_s) + c_s r_{as} - I_L(R_1 + R_L + r_{as} + R_1\mu_s + R_L\mu_s)}{R_1 + \mu_s R_1 + \mu_s R_4} \quad \ldots\ldots (14)$$

Therefore:

$$I_L\left[R_1\left(1+\frac{\mu_p R_3}{R_2+R_3}\right)(R_1+\mu_s R_1+\mu_s R_4) - K(R_1+R_L+r_{as}+\mu_s R_1+\mu_s R_L)\right]$$

$$= V_i\left(1+\frac{\mu_p R_3}{R_2+R_3}\right)(R_1+\mu_s R_1+\mu_s R_4) - v(1+\mu_p)(R_1+\mu_s R_1+\mu_s R_4) + c_p r_{ap}(R_1+\mu_s R_1+\mu_s R_4) - \left\{V_i(1+\mu_s)+c_s r_{as}\right\} K \quad \ldots\ldots\ldots (15)$$

where:

$$K = r_{ap}+\mu_p R_6 + R_4 + R_5 + R_6 + R_1\left(1+\frac{\mu_p R_3}{R_2+R_3}\right) \quad \ldots\ldots (16)$$

Also:

$$V_o \text{ the output voltage } = I_L R_L \quad \ldots\ldots (17)$$

Therefore:

$$\frac{dV_o}{dV_h} = \frac{R_L\left\{r_{ap}(R_1+\mu_s R_1+\mu_s R_4)\frac{dc_p}{dV_h} - K r_{as}\frac{dc_s}{dV_h}\right\}}{R_1\left(1+\frac{\mu_p R_3}{R_2+R_3}\right)(R_1+\mu_s R_1+\mu_s R_4) - K(R_1+R_L+r_{as}+\mu_s R_1+\mu_s R_L)} \quad \ldots\ldots (18)$$

and:

$$\frac{dV_o}{V_o}\cdot\frac{V_h}{dV_h} = \frac{\left\{r_{ap}(R_1+\mu_s R_1+\mu_s R_4)\frac{dc_p}{dV_h} - K r_{as}\frac{dc_s}{dV_h}\right\} V_h}{V_i\left\{\left(1+\frac{\mu_p R_3}{R_2+R_3}\right)(R_1+\mu_s R_1+\mu_s R_4) - K(1+\mu_s)\right\} - v(1+\mu_p)(R_1+\mu_s R_1+\mu_s R_4)} \quad \ldots (19)$$

This assumes that c_p and c_s are each zero.

Analyses of Figs. 2, 3, 4, 5 and 6

Since the analyses of these circuits are basically similar to that for Fig. 1, only the resulting expressions for the quantity $dV_o/V_o \,.\, V_h/dV_h$ will be given.

FIG. 2

$$\frac{dV_o}{V_o}\cdot\frac{V_h}{dV_h} = \frac{\left[r_{as}\left(r_{ap}+\frac{R_1 R_3 \mu_p}{R_2+R_3}+\mu_p R_6 + R_4 + R_5 + R_6\right)\frac{dc_s}{dV_h} - (\mu_s R_4 + R_1 + r_{as}) r_{ap}\frac{dc_p}{dV_h}\right] V_h}{\left[r_{ap}+\frac{R_1 R_3 \mu_p}{R_2+R_3}\left\{1-\frac{(\mu_s R_4 + R_1 + r_{as})}{R_1}\right\}+\mu_p R_6 + R_4 + R_5 + R_6\right] V_i + v(1+\mu_p)(R_4\mu_s + R_1 + r_{as})\Big)\Big]} \quad \ldots (20)$$

when $c_s = c_p = 0$.

FIG. 3

$$\frac{dV_o}{V_o}\cdot\frac{V_h}{dV_h} = \frac{\left[r_{as}k\frac{dc_s}{dV_h} - r_{ap}(\mu_s R_3 + \mu_s R_1 + R_1)\frac{dc_p}{dV_h}\right] V_h}{V_i(1+\mu_s)k - (\mu_s R_3 + \mu_s R_1 + R_1)[(1+\mu_p)V_i - \mu_p v]} \quad \ldots\ldots (21)$$

when $c_s = c_p = 0$ and where $k = r_{ap} + \mu_p R_5 + R_3 + R_4 + R_5 + R_1 + R_1\mu_p$.

FIG. 4

$$\frac{dV_o}{V_o}\cdot\frac{V_h}{dV_h} = \frac{\left[\left\{\mu_s R_5 + R_1(1+\mu_s)\right\} r_{ap}\frac{dc_p}{dV_h} - r_{as}(r_{ap}+\mu_p R_6 + R_1 + R_4 + R_5 + R_6)\frac{dc_s}{dV_h}\right] V_h}{\left\{V_i - v - \mu_p v\right\}\left\{\mu_s R_5 + R_1(1+\mu_s)\right\} - (r_{ap}+\mu_p R_6 + R_1 + R_4 + R_5 + R_6)\left\{V_i(1+\mu_s)\right\}} \quad \ldots (22)$$

when $c_s = c_p = 0$.

FIG. 5

$$\frac{dV_o}{V_o}\cdot\frac{V_h}{dV_h}=\frac{\left[\frac{dc_s}{dV_h}\, r_{as}(r_{ap}+\mu_p R_6+R_4+R_5+R_6)-r_{ap}\frac{dc_p}{dV_h}(r_{as}+\mu_s R_4+R_1)\right]V_h}{V_i(r_{ap}+\mu_p R_6+R_4+R_5+R_6)+(r_{as}+\mu_s R_4+R_1)(v+\mu_p v)} \qquad (23)$$

when $c_s = c_p = 0$.

FIG. 6

$$\frac{dV_o}{V_o}\cdot\frac{V_h}{dV_h}=\frac{\left[K_3 r_{as}\,\frac{dc_s}{dV_h}-K_2 r_{ap}\cdot\frac{dc_p}{dV_h}\right]V_h}{\left(V_i+\frac{(R_1+r_{as})\;v}{R_5}\right)K_3+K_2\mu_p v} \qquad (24)$$

when $c_s = c_p = 0$.

where:

$$K_1=R_1+r_{as}+R_L+\frac{(R_1+r_{as})}{R_2+R_3}R_L+\frac{(R_1+r_{as})}{R_5}R_L \qquad (25)$$

$$K_2=r_{as}+\mu_s R_4+R_1-\frac{(R_1+r_{as})}{R_2+R_3}R_3 \qquad (26)$$

and:

$$K_3=r_{ap}+\mu_p R_3+R_4+R_3-R_3{}^2\frac{(1+\mu_p)}{R_2+R_3} \qquad (27)$$

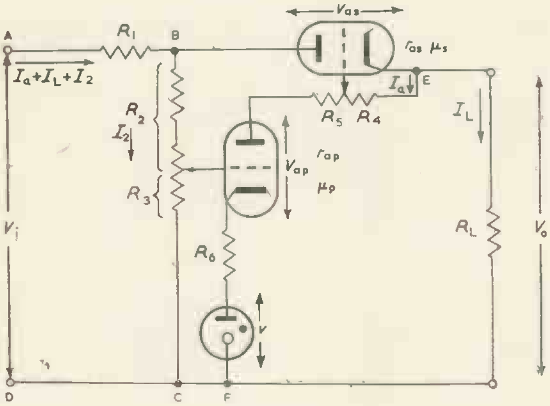
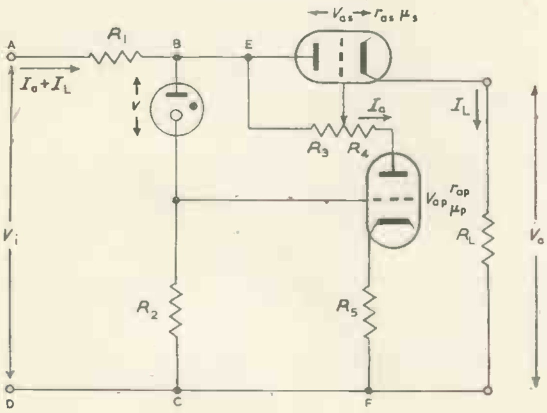


Fig. 2.

Fig. 3.



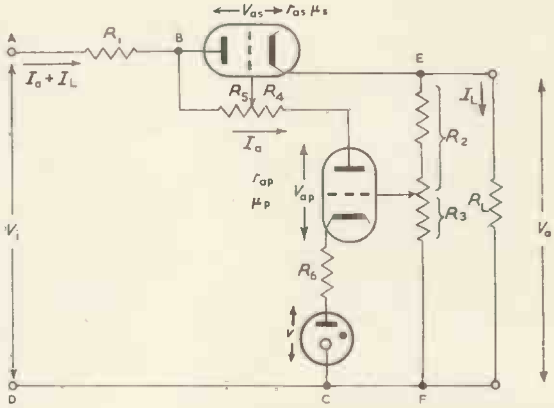


Fig. 4.

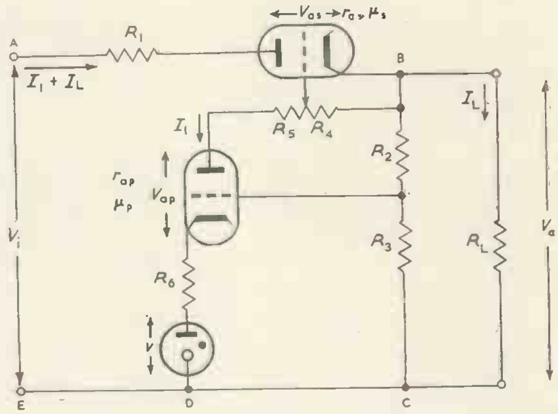


Fig. 5.

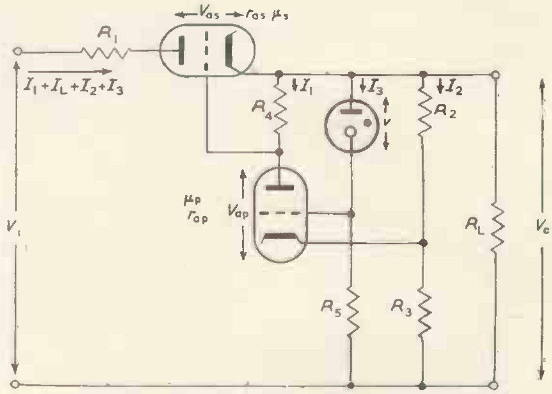


Fig. 6.

Conclusions

The formulæ developed enable the change in output voltage of each of the various stabilizers to be calculated for a given change in the valve heater voltages. The expressions are relatively complex, but when such calculations are made for a given circuit it is found that an analysis which assumes constant heater voltages of valves is not adequate. Generally the heaters of the valves are supplied from windings of transformers fed directly by the same mains voltage that supplies a rectifier unit providing the voltage requiring stabilization. Thus, as the input voltage to the stabilizer increases, so do the valve heater voltages.

The effects of heater-voltage variations on the stability of series-parallel valve arrangements are not as serious as have sometimes been stated or as might be expected. This is so because the two valves produce changes of output voltage, with variations of their heater voltages, of opposite sign, which is evident from a cursory inspection of the expressions developed. It is interesting to note that for many circuits the stability obtained when both input voltage and heater voltages change simultaneously is better than when the input voltage alone varies.

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Combined Lifting Rope and Cable for Underwater Television Cameras

SUCCESSFUL trials were recently held on the lake at Zurich, Switzerland, using an underwater television camera fitted with a specially designed multi-core signal cable which also acts as a lifting rope. The trials were carried out by Pye Ltd.

The tests showed that by using this type of cable the time taken to sink and lift an underwater camera is very much reduced. For instance, lowering a camera to a depth of 400 to 500 feet usually takes about 45 minutes whereas the new cable enables this to be done in two minutes. This great reduction is due to the absence of the usual wire hawser which makes a slow descent imperative because of the danger of its fouling the electric cable.

The new combined lifting rope and cable used was manufactured in one length of 1 000 feet by British Insulated Callender's Cables Ltd. It is a double plastic sheathed version of a standard polythene insulated multi-unit television camera cable with a loom woven textile finish of hemp as a final covering. This special interlocking weave provides extremely high tensile strength combined with flexibility and resistance to abrasion. It cannot fray or run loose, and owing to the individual threads being parallel with the axis of the cable there is no undue strain on the cable cores. The loom braid is clamped to the camera coupler and thus provides the necessary strength for lifting the camera.

For the purposes of the trials the underwater camera was connected to the cable by means of an underwater coupler developed jointly by Pye Ltd and B.I.C.C. Ltd., the cable passing over a pulley of approximately 1 foot diameter and from there to a capstan of the same diameter. The cable was then flaked on to the deck of the boat in a figure-8 formation to avoid twisting, and connected to the viewing unit and ancillary equipment. The camera was lowered by allowing the cable to slip round the capstan at the rate of approximately 250 feet a minute for 2 minutes and this was repeated many times. This was a severe test for the cable as the diameters of both capstan and pulleys were smaller than that normally recommended for this size of cable. The rubbing of the hemp finish on the capstan was also a severe test of the durability of the cable. At the same rate of lowering the camera into the water many depths around 150/250 feet were explored.

At all depths the cable resisted any twisting action and allowed the camera to remain in a fixed position indefinitely. While some wear was noticeable after three days of trials this was not considered unsatisfactory as the woven hemp finish had no protective covering. No electrical failures occurred and even after winding at high speed over small diameter pulleys the circuits were all intact.

The cable supported the weight of the camera (approximately 2½cwt) for long intervals above the surface of the water without showing signs of stretching and while the woven hemp finish tightened up on the p.v.c. sheath when in the water, no damage to the sheath occurred.

Additional tests included lifting the camera twenty feet in the air and allowing it to drop near the surface of the water, the whole strain being taken by the cable. No electrical failures occurred in the cable nor did the textile finish show signs of stretching or fracture.

At one time the camera was towed 100 to 200 feet behind the boat at a speed of approximately 12 knots and again the cable stood the test well.

The underwater camera suspended from the new cable

It is understood that Pye Ltd. are ordering a number of lengths of this type of B.I.C.C. cable which has greatly simplified the handling of underwater cameras, although even further improvements have been suggested as a result of the trials.

Notes from North America

IRE Annual Awards

The Institute of Radio Engineers has named John V. L. Hogan, President of Hogan Laboratories, the recipient of the IRE Medal of Honour, the highest American technical award in the radio engineering profession. The award, which was given "For his contributions to the electronic field as a founder and builder of The Institute of Radio Engineers, for the long sequence of his inventions, and for his continuing activity in the development of devices and systems useful in the communications art," will be presented during the IRE National Convention in New York City next March.

Mr. Hogan began his long career in radio in 1906 as a laboratory assistant to Lee De Forest. In 1912 he helped found the Institute of Radio Engineers, which has since grown to be one of the largest engineering societies in the world with an international membership of over 43,000. He served as Vice-President of the IRE from 1916 to 1919 and President in 1920. He has also served frequently as a member of the Board of Directors and on many IRE committees.

The Morris Liebmann Memorial Prize, awarded annually to an IRE member who has made a recent important contribution to the radio engineering art, was given to Kenneth Bullington, Bell Telephone Laboratories, New York, N.Y., "for his contributions to the knowledge of tropospheric transmission beyond the horizon, and to the application of the principles of such transmission to practical communications systems".

Mr. Bullington joined Bell Telephone Laboratories in 1937 and since that time has been engaged almost entirely in studies relating to the propagation of radio waves.

Wilbur S. Hinman, Jr., Director of the Diamond Ordnance Fuze Laboratories, Washington, D.C., received the Harry Diamond Memorial Award, which is given to persons in government service for outstanding work in radio and electronics. The award was presented "for his contributions to the electronic art in the fields of meteorology and proximity fuzes".

Scholarships in Librarianship

The Scholarship and Student Loan Fund Committee of Special Libraries Association announces two $500 scholarships to be granted for the academic year 1956-1957 for graduate study in librarianship, leading to a degree at an accredited library school. Applicants must be college graduates of high academic achievement who need financial assistance in obtaining the professional education necessary for work in the special library field.

Application forms and details of eligibility for the scholarship award may be obtained from the Executive Secretary, Special Libraries Association, 31 East Tenth Street, New York 3, N.Y. Applications must be received not later than March 1, 1956. The awards will be announced at the annual convention of the Association in Pittsburgh, Pennsylvania, June 1956.

The Special Libraries Association is an international organization of librarians working in libraries and other information centres concerned with special subjects and serving business, industry, science, social welfare, government and the arts. The Association has over five thousand members in the United States and Canada, as well as members in other countries.

World Symposium on Applied Solar Energy

A number of foreign scientists are expected to attend the Applied Solar Energy Conference, details of which were announced in ELECTRONIC ENGINEERING, October issue (p. 460).

Among those expected are: Prof. V. A. Baum, head of the Heliotechnical Laboratory, G.M. Krzhizhanovsky Power Institute, Moscow, U.S.S.R.; Prof. Felix Trombe, Director, Laboratoire de l'Energie Solaire, France; Dr. Harold Heywood, Imperial College of Science and Technology, England; Dr. H. Tabor, Director, National Physical Laboratory, Israel; Dr. M. L. Khanna, National Physical Laboratory of India; Dr. Hiroshi Tamiya, Director, Tokugawa Institute, Japan.

Their participation has been made possible by financial support from the National Academy of Science, the National Science Foundation, The Ford Foundation, The Rockefeller Foundation, Office of Naval Research and U.S. Air Force and UNESCO.

Miniature Volume Controls

The Stackpole Carbon Company has developed a small and efficient variable composition-resistor—the Type F Volume Control.

The important mechanical specifications for Type F include:

control diameter 0·637in

shaft diameter 0·124in
(available with knurled, slotted or flatted shaft)

bushing diameter 0·250in, $\frac{1}{4}$-32 thread.

Notably quiet and smooth in operation, the type F volume control gives stable, long-life performance even under wide humidity variations. It is rated for 0·3W up to 10kΩ; for 0·2W above 10kΩ.

Single and double pole switches for type F controls will be available within a few months.

Further specifications and particulars are available from Ad. Auriema, Inc. 89 Broad Street, New York 4, N.Y.

The miniature control compared with one of standard size.

Short News Items

The United Nations are to publish the proceedings of the International Conference on the Peaceful Uses of Atomic Energy, held in Geneva in August this year. The Proceedings, which will be published in sixteen volumes of approximately 500 pages each, will constitute the complete, unabridged record of the Conference, and comprise all papers, whether presented orally or in written form at the Conference, together with a record of the discussions concerning each paper. The volumes will be published in several languages; the English edition will be available in the beginning of 1956, the others at a later date still to be determined. A special pre-publication price of £39 for the full series of sixteen volumes has been established. This will be protected for all advance orders received up to 31 December 1955. Orders for the full set or for individual volumes (the prices of individual volumes will be announced later) may be placed with any Sales Agent for United Nations publications, including H.M. Stationery Office, P.O. Box 569, London, S.E.1, and N.V. Martinus Nijhoff, Lange Voorhout 9, The Hague, Holland.

Eurovision is now established on a permanent basis. Programmes will be seen simultaneously by people in Britain, France, Italy, Switzerland, Western Germany and Berlin, the Netherlands and Belgium. Austria will join later this year, and Denmark perhaps next year. The BBC will take programmes from the Continent on an average of once a week. Eurovision is organized by the European Broadcasting Union, which has its administrative headquarters at Geneva and technical headquarters in Brussels. Its President for this year is Sir Ian Jacob, Director-General of the BBC.

The Radio and Electronic Component Manufacturers' Federation occupied space of 1 300 square feet at the recent British Exhibition in Copenhagen. The exhibits covered almost the whole range of components for the radio and electronic industries.

Marconi's Wireless Telegraph Co Ltd announce that work on the 450ft tower for the Independent Television Authority's Midland transmitting station at Lichfield is proceeding according to schedule, and has progressed beyond the 90ft level. Marconi's are also responsible for the design and construction of the aerial array and feeder system.

Marconi's have built and supplied a 20kW h.f. broadcasting transmitter for the Tanganyika Broadcasting Service. This transmitter has now been installed, with the assistance of Marconi engineers, and is at present undergoing tests. It has been installed in a new broadcasting station situated off the Pugu Road in Dar es Salaam. In addition to the transmitter, Marconi's are supplying a studio control console, programme lines equipment, programme input equipment, and the complete aerial system.

Marconi Instruments Ltd, in conjunction with Svenska Radioaktiebogalet, have been successful in obtaining a £23 000 contract for signal generators from the Royal Swedish Air Board.

A joint Conference of the Physical Society and the Royal Meteorological Society will be held in the Department of Meteorology, Imperial College, Exhibition Road, London, S.W.7. on Wednesday and Thursday, 4 and 5 January, 1956. The subject of the Conference, which is being organized by Professor P. A. Sheppard, will be "Cloud Physics". Inquiries should be addressed to the Physical Society, 1, Lowther Gardens, Prince Consort Road, London, S.W.7, marked for the attention of Miss Miles.

The Technical College, Bradford, has issued a prospectus of part-time day and evening courses for the 1955-56 session. The Department of Electrical Engineering provides courses in Radio and Television Servicing, Electrical Engineering, Telecommunications Engineering, Electrical Installation, etc. Particulars may be obtained from Dr. G. N. Patchett, Head of Department of Electrical Engineering, Technical College, Bradford 7.

Dr. D. G. Tucker has been appointed to the Chair of Electrical Engineering at the University of Birmingham. Among many publications, he will be remembered for his various contributions to ELECTRONIC ENGINEERING, in particular the series of articles on the Synchrodyne Receiver which appeared during 1947 and 1948. For the past five years Dr. Tucker has been working with the Royal Naval Scientific Service on underwater detection, research and development.

Sir Patrick Branigan, Q.C., has recently joined the board of Suflex Ltd as Chairman of the company. Mr. B. H. Hornung remains as Managing Director and Mr. H. S. Payman has been appointed General Manager. These new appointments have been part of a general expansion of Suflex Ltd, who are now manufacturing polystyrene capacitors in addition to their sleeving, wire and braided products.

The Ministry of Supply announces that Dr. W. B. Littler has been promoted to Chief Scientific Officer and appointed Principal Director of Scientific Research (Defence) in succession to Dr. Cawood, who is to become Principal Director of Scientific Research (Air) on 1 November.

Mr. F. S. Barton has been appointed Adviser, Defence Supplies (Ministry of Supply) to the United Kingdom High Commissioner in Canada. Dr. D. H. Black has been appointed Principal Director of Electronics Research and Development in succession to Mr. Barton, and Dr. W. H. Wheeler will succeed Dr. Black as Head of the United Kingdom Ministry of Supply Staff, Australia, and Scientific Adviser to the United Kingdom High Commissioner in Australia.

The BBC announces that it has been decided, after consultation with the GPO, the radio industry and the trade, that the new television station at the Crystal Palace, which is to be opened early in 1956, will use the same method of transmission of the vision signals as is used at all the post-war BBC television stations, the upper sideband being partially suppressed. In this respect, the new station will differ from the existing station at Alexandra Palace, which transmits both sidebands equally. The new station will use the same frequencies and polarization as Alexandra Palace. The actual date when it will come into service will be announced as soon as practicable.

Sir Walter Monckton, Q.C. M.P., Minister of Labour and National Service, is to be the guest of honour at the annual dinner of the Radio Industry Council at the Dorchester Hotel, London, on 23 November.

International Aeradio Ltd are to provide air traffic control, communications and fire and crash rescue services at Lympne Airport. They offer a complete service to airport owners and administrations responsible for the provision of aerodromes, air routes, air traffic services, communications and radio aids to navigation. A licence for Lympne Airport was recently issued to Skyways Ltd.

The British Standards Institution has now opened a Sales Office in the Birmingham Chamber of Commerce, 95, New Street. This development is in line with the B.S.I.'s policy of making British Standards readily available at important industrial centres.

Addendum. In the article "The Design and Performance of a Simple V.L.F. Oscillator" on page 380 of the September issue V_2 and V_3 (Fig. 4) are Mullard type 85A2.

LETTERS TO THE EDITOR

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

Regulated d.c. Power Supplies at Low Voltages

DEAR SIR,—The article in the July 1955 issue appears to show the "hard way" to accomplish the intended purpose. The circuit configuration titled conventional in Fig. 5 results in impractical tube operating conditions for low voltage operation.

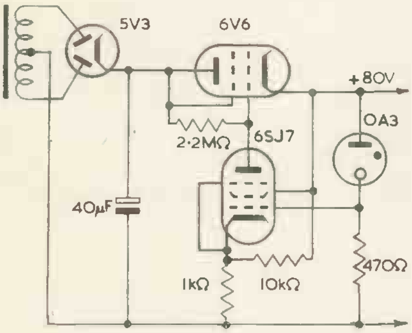


Fig. A. Regulated supply

Fig. A shows a simple regulator circuit configuration that gives normal tube operating conditions and provides an output potential a few volts higher than the regulator tube drop. The impedance is a couple of ohms and the ripple a few millivolts. A simple modification of this circuit, Fig. B, provides an adjustable output down to a few volts with little additional complication. Both

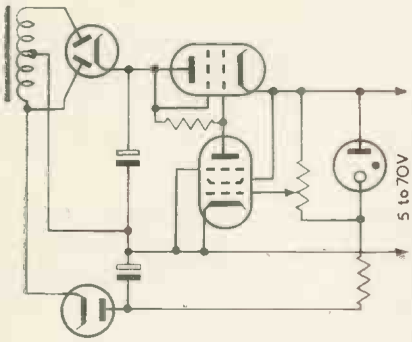


Fig. B. Alternate method

methods use but one voltage standard. The original circuit has been used commercially for years. The output may be shunted with a capacitor to maintain low impedance at high frequency beyond cut-off of the amplifier stages.

I would suggest that your readers compare the circuit complexity, count the electrical components, appraise their cost, reliability, size, weight, and heat dissipation, then compare the performance and draw their own conclusions. The merits of the tubes or glow lamps used are equally applicable to either design. The 6AU6 is a good choice of voltage amplifier for the alternative method circuit as it has extremely high gain as a "starved current" pentode, that is with screen potentials of 10V or lower.

The Hunt-Hickman article on "Electronic Voltage Stabilizers" which appeared in the January, 1939, issue of *The Review of Scientific Instruments*, covers most of the features of the circuits under discussion and may be of interest to your readers.

Yours faithfully,
SERGE L. KRAUSS,
Electronic Product Design,
C. G. Conn Ltd,
Elkhart, Indiana.

The authors reply:

DEAR SIR,—We agree with Mr. Krauss that the circuit of Fig. 5 is unsuitable for low voltage operation and, indeed, we have explained why this is so. There are a number of objections to the circuit of Fig. A. The characteristics of the regulator tube OA3 including ageing effects are: striking voltage 105V max, stabilizing voltage 68V min. and 80V max. The effect of a tube with the maximum striking voltage is that the output rises to 105V before the stabilizer tube strikes, and the circuit starts stabilizing. On switching on, there is thus a spike of some 35V which is particularly objectionable when feeding electronic switching circuits. The wide tolerance in stabilizing voltage gives an ambiguity of 12V at the output.

The circuit of Fig. B has improved operating conditions for the stabilizer which now runs at a more nearly constant current and the switching-on voltage spike is absent. The upper limit of the stabilized voltage is again not designable. As regards the lower limit of output voltage the amplifier valve (assumed to be 6SJ7) will not be effective with a 5V screen supply and the lower limit is about 30V. The use of an additional negative supply was mentioned in our article but it is more efficient to use a stabilizer tube and a valve running at low voltage and current, than to use a rectifier and an electrolytic capacitor, and moreover the former arrangement is less likely to go faulty. The choice of a valve with a somewhat high resistance gives better regulation than a low resistance valve, but at the cost of reduced efficiency.

If our circuit is compared with that of Mr. Krauss, with both circuits operating from 250-0-250V transformer, and giving an output of 50V, our circuit gives 80mA per series valve whereas the circuit of Fig. B gives only 45mA, allowing for supply voltage variations in both cases. Simple but inefficient circuits ease the problem of the designer, at the expense of the user who, on building the power supply into his own equipment, has to make provision for dissipating the heat unnecessarily generated. The second circuit of Mr. Krauss is well suited to laboratory bench work where the first cost may be important and heat dissipation and maintenance present no problem.

Yours faithfully,
R. K. HAYWARD,
Post Office Engineering Dept.,
Research Station,
London, N.W.2.

An Inexpensive Dekatron Scaler

DEAR SIR,—Dr. Kerkut's article in the September issue, includes a statement that the component values recommended in the manufacturer's literature are unsatisfactory. In all fairness we would point out that the network referred to was designed as a combined coupling and quenching circuit for use with the GTE175M cold cathode gas filled trigger tube, and it is not surprising that different component values are needed when used in the anode circuit of a hot cathode high vacuum tube such as the 6SN7.

Yours faithfully,
L. C. BURNETT,
Cold Cathode Tubes
Application Laboratory,
Ericsson Telephones Ltd.

The author replies:

DEAR SIR,—Mr. Burnett's criticism of the wording of my article is perfectly fair. However, the fact remains that when I used a cheap and easily available double triode for coupling, the circuit required modification.

Yours faithfully,
G. A. KERKUT,
Sub-Department of Physiology,
The University of Southampton.

Isocline Diagrams for Transistor Circuits

DEAR SIR,—I note that Mr. Oakes agrees that certain errors occurred in his derivation of the expression for the radius of curvature of the circular arcs used in the construction of the isoclines. We apparently differ, however, in our interpretation of the magnitude of this error applied to the complete isocline.

We are both agreed that the radius of curvature error is represented by the term

$u/k^2\,(v_c' - v_t)\,dv_t/du$ in the expression

$$\frac{-k}{1 - u/k^2\,(v_c' - v_t)\,dv_t/du}$$

This error term, referring to his Fig. 4 can be rewritten

$$QV.PV/(QP)^2 \quad dv_t/du$$

Considering one particular point V on the curve $v_t = f(u)$ the expression

$$QV.PV/(QP)^2$$

will vary between the limits

$$0 \rightarrow 0{\cdot}5 \rightarrow 0 \rightarrow -0{\cdot}5 \rightarrow 0$$

as PV varies from $+\infty$ through 0 to $-\infty$.

Also, dv_t/du will, for different points V on the curve $v_t = f(u)$, vary between $\infty \to 0 \to -\infty$.

Thus the error term $QV.PV/(QP)^2\ dv_t/du$ may be of considerable magnitude compared with unity, and, what is more to the point, may be of negative sign.

When negative, of course, the resultant error may cause the radius of curvature to change sign and the isocline will bend the wrong way. Also, when it is small, i.e. the curvature large, it is further evident that errors in the degree of curvature will have an unfortunately severe effect on the final destination of the isocline.

I will now heartily agree with Mr. Oakes that this error "restricts the useful length of the circular arcs". I will further agree that if the arcs are made short enough, and their number correspondingly large, errors in the calculation of the curvature, even of the magnitude I have outlined above, will be rendered impotent in causing any deviation from the true isocline. What I would like to point out, however, is that if the arcs are going to be so small as to render large errors in radius of curvature unimportant, then firstly, there was no point in estimating the radius of curvature at all, since the curve could be drawn equally well by constructing it of a series of short sections of the tangent, and secondly, a construction of this nature, which could not place reliance on the radius of curvature to keep the sections, and thus the total constructions, down to reasonable number, would be so laborious as to be impractical.

Yours faithfully,

A. J. Buxton,
Highcliffe, Hants.

Three Input Binary Adder and Subtractor Making Use of Crystal Diode Gates

Dear Sir,—There is much to be said, in digital computers making use of crystal diodes, for keeping the number of electronic valves down to a minimum. Gating operations are easily performed with crystal diodes and resistors, but the operation of negation needs at least one electronic valve (assuming that it is not convenient to use pulse transformers or transistors), and for this reason a circuit capable of adding or subtracting while making use of only one negating circuit, is likely to be of value. The selection of addition or subtraction can be by means of a waveform applied to one or other of two control lines.

The circuit described here makes use of the logical arrangement used in the binary adder described in "High-Speed Computing Devices" (Engineering Research Associates, Inc.) in Fig. 13-8, p. 277 (First Edition). A slight modification is incorporated in the logic of the addition, in order to facilitate simple conversion to a subtractor. This type of adder was adopted as a basis because it makes use of only one negating circuit.

The table shows the outputs required to form $A + B + C''$ for addition, or $A - (B + C'')$ for subtraction, for any combination of A, B and C''. C'' = third input, usually carry from previous digit. S = sum or difference, and C = carry to next digit. It will be seen that the output for S is the same in both cases, so that the conversion from "add" to "subtract" may be done by merely altering the C output.

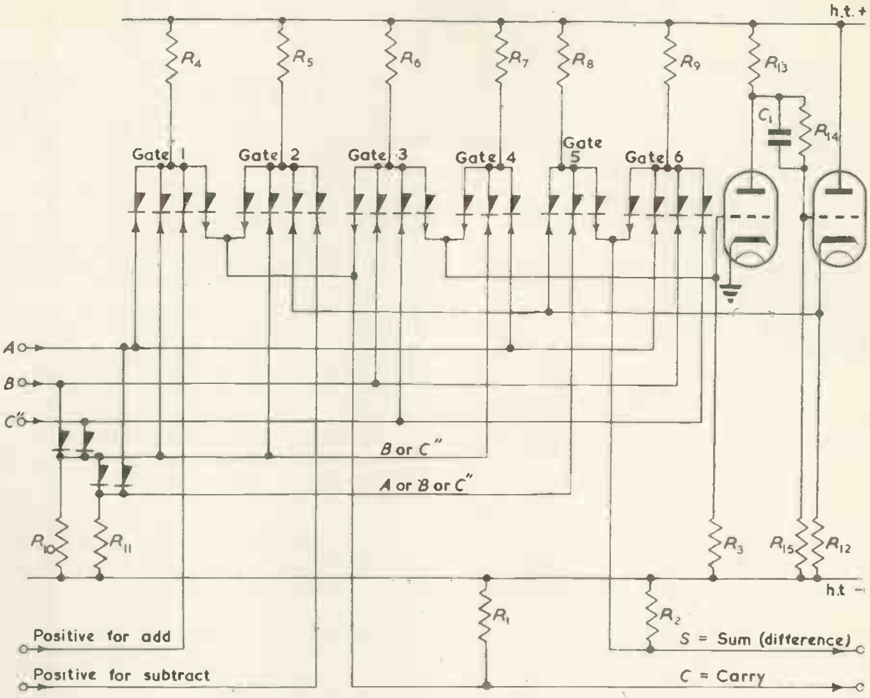


Fig. 1. Binary adder-subtractor

INPUTS			OUTPUTS				
			ADD		SUBTRACT		
A	B	C''	S	C	S	C	
0	0	0	0	0	0	0	1
0	0	1	1	0	1	1	2
0	1	0	1	0	1	1	3
0	1	1	0	1	0	1	4
1	0	0	1	0	1	0	5
1	0	1	0	1	0	0	6
1	1	0	0	1	0	0	7
1	1	1	1	1	1	1	8

The circuit shown in Fig. 1 gives outputs of positive polarity from the gates along the lines with arrows pointing downwards, output from a gate being obtained when all the input lines to that gate (with arrows pointing upwards) are pulsed positively. The valve circuits have the property of negating the polarity of its input. Outputs from the gates are given for the following conditions, neglecting the "add" and "subtract" lines:—

Gate 1: A and (B or C'')

Gate 2: B or C'' and not [(A and B) or (B and C'') or (A and C'')]

Gate 3: B and C''

Gate 4: A and (B or C'')

Gate 5: A or B or C'' and not [(A and B) or (B and C'') or (A and C'')]

Gate 6: A and B and C''.

It will thus be seen that gates 5 and 6 mixed give outputs for exactly the conditions needed to form S (see Table). Gates 1 and 3 mixed give outputs for the conditions needed to form C (add) and gates 2 and 3 mixed for the conditions needed for C (subtract). Thus by mixing gates 5 and 6 for the S outputs, and by mixing gates 1, 2 and 3 and inhibiting gate 2 for the C (add) output and gate 1 for the C (subtract) output, the required operation is derived.

It is assumed that positive pulses are used, of height V volts, extending from voltage $-V$ to earth. The two valves shown are an invertor and a cathode-follower, the resistors R_{13} to R_{15} and capacitor C_1 being adjusted so as to give an adequately large and correctly shaped pulse, at the correct d.c. potential, from the output. Resistors R_1 to R_3 should be capable of discharging any capacitance connected to them, over voltage V, in time t, where t is the desired response time. Resistors R_4 to R_9 should be capable of charging any capacitance attached to them over V in time t, of holding any of resistors R_1 to R_3 attached to them to earth potential, and of supplying the reverse current of any appropriate crystals. The values of R_{10} to R_{12} may be similarly calculated. Given these resistor values, the output impedance of the circuits needed to supply A, B and C'' may be calculated in a manner similar to the above.

Yours faithfully,

B. R. Taylor,
The British Tabulating Machine Co. Ltd,
Electronics Research Laboratories.

BOOK REVIEWS

Electronic Circuits

By T. L. Martin, Jr. 707 pp. 464 figs. Demy 8vo. Prentice-Hall, Inc., New York. Bailey Bros. and Swinfen, Ltd., London. 1955. Price 99s.

IT is difficult to understand how it comes about that electrical engineering students in this country appear to be so ill-informed in the subject of electric circuit analysis when one surveys the wealth of text-book information which has arrived from the United States during recent years. These books bear evidence of the tremendous advances that have been made in the United States in the teaching of the basic technological principles of electrical engineering. Considerable effort has been directed towards the integration of circuit theory fundamentals with the view to realizing a unity of approach and rigour of thought which contrasts with the *ad hoc* and incoherent collection of outdated notions that constitute most electronic circuits courses in this country.

This new book by Professor T. L. Martin, Jr., of the University of Arizona, is a good representation of current practice in electronic engineering education in the United States. It contains nothing that the student will have to unlearn if he subsequently extends his studies into the realm of advanced circuit engineering. The book is divided into three parts: Part 1 is introductory, Part 2 deals with class A circuits and Part 3 with circuits operating in the switching mode.

The first part opens with an explanation of the author's particular approach to the analytical treatment of circuits incorporating electronic devices. This is by way of what he calls the principle of linear approximation according to which every such system has an equivalent linear circuit representation containing a selector switch. The continuous characteristic curve of any non-linear device is approximated by a discontinuous series of straight line segments and the switch in the equivalent circuit permits the selection of that part of the circuit which has as its characteristic the appropriate straight line segment. The author defines class A operation to be that for which the switch remains in one position only and switching mode operation to be that for which the switch moves from one position to another. Switched equivalent circuits are then developed to represent the thermionic valve and the transistor. In this way the author is able to bring these devices within the compass of ordinary linear circuit theory.

The second chapter outlines the principles of linear circuit theory. The subject is presented along modern lines and begins with some general remarks concerning the relation between excitation function, system function, and response function. Kirchhoff's laws are distinguished not as the first and second (Kirchhoff himself was inconsistent here) but less ambiguously as the node and loop laws. An explication of modern loop-nodal analysis follows and leads naturally to the principle of duality. These important topics are dismissed somewhat hastily but at least the student is in contact with up-to-date technique. No undue significance is attached to sinewave excitation; the reader is led at once to the method of cisoidal oscillations and hence to the Laplace transform as a general instrument of linear system analysis. The graphical method of determining both time and frequency response functions on the complex frequency plane is clearly described and the chapter concludes with a short reference to physical realizability and minimum phase circuits.

Part 2 begins with Chapter III which comprises a thorough-going account of the principles of thermionic valve amplifier theory. The section on the calculation of amplifier gain avoids the current abuse of the use of decibels by following the recommendations of J. W. Horton who suggested the new but as yet non-standard unit the *logit*. The time and frequency response characteristics of aperiodic, tuned, and distributed amplifiers are the subjects of the next three chapters. The discussion is extensive and full use is made of the circuit theory principles explained in Part 1.

Chapter VII deals with feedback amplifiers. There appears to be two schools of thought concerning the interpretation of A in the formula $A' = A/(1 - \beta A)$ for the voltage amplification of a feedback amplifier. Some writers take it to represent the closed-loop voltage amplification of the system with respect to the input terminals of the A circuit while others take it to mean the voltage amplification of the open-loop system with the β circuit absent. Unfortunately, the latter interpretation is the one adopted by the author. The two viewpoints lead to conflicting results in specific cases. For example, in the case of simple cathode feedback, the co-efficient β according to author's viewpoint comes out to be a function of the valve parameters. Consequently the author is compelled to infer that the feedback voltage is not equal to the voltage across the cathode impedance. The chapter concludes with a discussion of the Nyquist and Routh-Hurwitz stability criteria and the remaining four chapters of Part 2 deal with such topics as transistor amplifiers, noise, and oscillators. The treatment of oscillators is thorough and includes an explanation of the method of isoclines and the use of cyclograms as aids to the study of negative resistance oscillators.

Part 3 occupies the final seven chapters and begins with a general exploration of the factors involved in the operation of thermionic valves as power amplifiers in the switching mode. The technique is based upon a Fourier analysis of the anode current waveform resulting from the application of a sine-wave input voltage of sufficient amplitude to operate the hypothetical selector switch in the equivalent circuit. Amplitude, angle, frequency, and phase modulation are expounded at length in Chapter XIII which is followed by two chapters concerned with rectification and detection. Chapter XVI introduces the fundamentals of magnetic amplifier theory and the last two chapters give an account of wave shaping, computing circuits, trigger circuits, and relaxation oscillators in the light of the principle of linear approximation.

The author's approach is essentially analytical but the book contains a great deal of useful circuit design information. Emphasis throughout is on the formulation and solution of electronic circuit problems. Inevitably there are points of view expressed on which agreement will not be unanimous but the general spirit of the book is compatible with the modern philosophy of engineering science. In the hands of the more enlightened of our technical college teachers it would make an excellent course book. Problem lists append every chapter but no answers are provided. The book is one of the Prentice-Hall Electrical Engineering Series; it is remarkably well printed and handsomely bound but the price will place it beyond the reach of most students.

S. R. DEARDS.

Linear Feedback Analysis

By J. G. Thomason. 355 pp. 237 figs. Demy 8vo. Pergamon Press Ltd. 1955. Price 55s.

THIS is the sixth volume of the "Electronics and Waves Series" edited by Dr. Fry, of Harwell, and published by the Pergamon Press, Ltd. The book, which is intended for the graduate reader, gives an introduction to modern network analysis. The initial chapters deal with the mathematical techniques while the later chapters give examples of the application of these techniques to specific circuit problems.

The first chapter dealing with steady state analysis using mesh and nodal methods, together with the next two chapters discussing the Laplace transform, cover ground which is not necessarily unfamiliar to the present-day graduate. This, however, does not detract from their usefulness. Chapter IV gives an elementary introduction to feedback circuits and Chapter V describes the design of amplifying stages. The whole of the discussion in this chapter is restricted to one particular pentode valve (Services Type CV138) and it is stated that the pentode is the "accepted choice" for the voltage amplifier; this is not generally true as triodes are also used extensively both in this country and the U.S.A. Chapters VI and VII deal with the three different ways in which the stability of a feedback system can be studied. Here the emphasis is on the gain/phase analysis of H. W. Bode and this method is mainly used in the examples given in subsequent chapters.

The author was formerly at A.E.R.E., Harwell, and is now at R.R.E., Malvern, and the practical examples given in the later chapters refer to topics of particular interest to these establishments. The examples given include audio and video amplifiers, feedback-integrators and stabilized power supplies. The only feedback system having mechanical constants that is discussed is the velodyne; this section could well have been much extended so as to increase the utility of the book to the servo-mechanisms engineer.

A list of thirty-one references is given at the end of the book. This is inadequate for a book of this nature and, furthermore, over half of the references cited are incomplete.

V. H. ATTREE.

Precision Electrical Measurements

300 pp. 80 figs. Demy 8vo. Her Majesty's Stationery Office. 1955. Price 27s. 6d.

THIS book contains the proceedings of aa four-day international symposium held at the National Physical Laboratory in November last.

Twenty-six papers were presented during the symposium, which was divided into five sessions, each devoted to a particular branch of the subject, i.e., Capacitance and Dielectrics, Inductance and Magnetics, Electrotechnics, High Voltage Impulse Testing Techniques.

The publication gives the papers in each session in full and summarizes the discussions with which the sessions ended.

Calibration of Temperature Measuring Instruments

47 pp. 27 figs. Demy 8vo. Her Majesty's Stationery Office. 1955. Price 2s.

THIS twelfth publication in the series Notes on Applied Science prepared by the National Physical Laboratory describes the basic scales of temperature and the instruments which are available for making temperature measurements. The methods and apparatus which are described are mainly those which are used at the National Physical Laboratory for routine checking of accuracy in temperature equipment and in the maintenance of standards of temperature.

Photo-Electric Handbook

By G. A. G. Ive, 152 pp. 108 figs. Demy 8vo. George Newnes Ltd. 1955. Price 17s. 6d.

THIS book has been specially prepared to give engineers the particular information they require regarding the special problems of photo-electric cells and their installation.

The first five chapters of the book deal in detail with the various parts of the equipment, i.e. light sensitive cells, thermionic valves and circuits, relays and auxiliary equipment, projectors and photocell receivers, and control units. Chapter VI gives information on the wiring and voltage supply, and Chapter VII deals with setting up and checking installations. Some useful fault finding charts form the subject of Chapter VIII, and the book concludes with some tables and an appendix on the measurement of light.

ELECTRONIC EQUIPMENT

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

E.H.T. Power Unit

(Illustrated below)

THIS power unit (type 105) provides an e.h.t. supply suitable for most types of radiation detector, as well as supplies for a probe unit such as the E.T.L. type 109, or type 110, or the A.E.R.E. type 1014A.

The e.h.t. supplies are continuously variable from 300V to 3·3kV in three ranges. Maximum current on the 300V to 1·2kV and 1kV to 2kV ranges is 2mA and on the 1·8kV to 3·3kV range it is 0·5mA. The output voltage is indicated on a 4in moving-coil meter. Maximum ripple is 20mV peak-to-peak and is much less than this on small output loadings.

A shunt type stabilizer employing a high gain amplifier gives a stability ratio (percentage change in input voltage percentage change in output voltage) greater than 100 to 1 for mains changes between

− 10 per cent and + 5 per cent of nominal.

The stabilized supplies available for a probe unit are + 300V, 10mA; − 150V, 7mA; and − 100V, 1·5mA, in addition to an unstabilized 6·3V, 2·5A a.c.

Ericsson Telephones Ltd,
Beeston,
Nottingham.

New Valves

(Illustrated above right)

THE QV20-P18 is a radial-beam tetrode for use in the pulse modulators of medium power radar transmitters. It gives power outputs up to 300kW.

This valve is the first of a range of hard valve modulators which Mullard are introducing for all forms of ground and airborne radar equipment. Advances in radar techniques such as pulse switching have created a demand for these valves as their use eliminates the need for the complex switching necessary when other means are employed.

The QV20-P18 has peak anode voltage and current ratings of 20kV and 18A. It is convection-cooled, with a rated anode dissipation of 60W. The heater ratings are 26V, 2·25A. The valve is of compact and sturdy construction. It is an exact equivalent of U.S. type 4PR60A, and is also a plug-in replacement for types 5D21 and 715C, though the lower ratings of these do not permit complete interchangeability.

The power triode TY4-350, which is a direct replacement for the U.S. type 883A is now available. The valve, which is widely used in r.f. heating equipment, has a maximum rated anode dissipation of 300W, and operates with h.t. voltages up to 3kV. It has a thoriated tungsten filament which operates at 10V, 10A.

Mullard Ltd,
Century House,
Shaftesbury Avenue,
London, W.C.2.

Lead-Acid Cell

(Illustrated below)

WITH a nominal capacity of 20Ah, this new cell, type ZVP19, is primarily for use with the vibrator power packs of portable radio equipment for the Armed Services. It can also be used for other similar applications and where high rate discharges with relatively high terminal voltages are required.

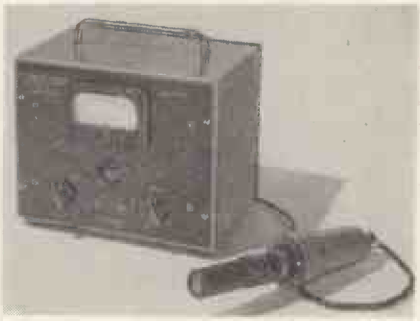
An exceptionally stringent specification ensures maximum reliability and efficiency under arduous climatic or operating conditions. The cell is assembled in a robust container moulded in translucent, high-impact polystyrene. The lid, an intricate one-piece moulding in the same material, is cemented in position by

a special technique which produces a reliable and leak-proof joint.

One of the most notable features is the combined gas vent and acid trap moulded integrally with the lid. The vent outlet is so designed that the possibility of its becoming blocked by dirt or by some object resting upon it is practically eliminated. The moulding of the acid trap and vent in the lid itself rather than in the filling plug enables a far more efficient design of trap to be employed. A separate filling plug is moulded with a broad slot at the top so that it can be screwed up or unscrewed with a coin.

The plates are separated by sheets of Norvic and resin-bonded glass wool.

Chloride Batteries, Ltd.
Clifton Junction,
Swinton,
Manchester.



Radioactivity Meter

(Illustrated above)

DESIGNED in collaboration with A.E.R.E., Harwell, and the Sub-Committee of the British Science Masters' Association, as a rugged, inexpensive, yet precision instrument for carrying out an illustrative range of experiments with radioactive isotopes, the radioactivity meter type 552 is intended primarily for educational purposes, particularly in schools. It is also suitable as a simple, easy to use, general-purpose rate measuring instrument, giving a useful accuracy for quantitative work.

It consists of a beta-gamma geiger counter probe and ratemeter circuit, employing a single cold-cathode trigger tube. The counter is of a robust pattern with window thickness 3·5 to 4m/cm². Indication of the low count rate is shown by a flash for each pulse, in the trigger tube. The built-in e.h.t. supply is continuously variable from 300 to 600V, as indicated on the meter.

Isotope Developments Ltd,
Beenham Grange,
Aldermaston Wharf,
Reading,
Berkshire.

Electronic Timing Units

(Illustrated above right)

THESE timing units may be used individually, or a number may be connected together to form a sequence or cyclic timer. The timing sequence being

initiated by the momentary closing of a contact, e.g. a push-button.

Units can be supplied for timing between the range of 0·1sec aand 5min, the repetitive accuracy being ± 2 per cent at mid-scale. Individual units are fitted in a case measuring 5 by 4 by 4in.

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

P.V.C. Decals

THESE p.v.c. decals are intended for identification marking of p.v.c. cables and are supplied in the form of ordinary (reverse) transfers. They are applied to the cable by means of a special solvent, the main constituent of which is cyclohexanone. This liquid dissolves the p.v.c. of both lettering and cable to some extent. When the two are applied together a firm bond results. Cables are ready for immediate use after the decal is applied, although maximum hardness is not attained for 24 hours.

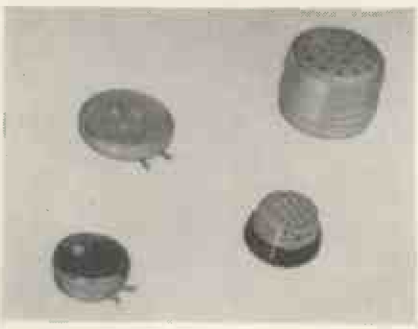
The Murray-Hill Co.,
Link Hill, Sandhurst,
Kent.

Microphone Inserts

(Illustrated below)

AMPLIVOX have recently introduced a range of miniature magnetic diaphragm microphones and microphone inserts; four types are illustrated in the accompanying photograph. The two flat insert type units have been specially designed for use in transistor hearing aids while the other two, being fully protected against moisture, are suitable for close speaking as in radio communication applications.

The microphones are intended primarily for high intelligibility speech reproduction and they all have a controlled response substantially free from resonances within the speech band. In spite of the necessarily large amount of damping, sensitivity is high, being maintained within ± 5dB at − 75dB ref 1V dyne/cm² into 300Ω, i.e. close speech produces peak outputs of the order of 10mV.



The largest of the four microphones illustrated is of the differential noise-cancelling pattern, consisting of an anti-phase back to back pair.

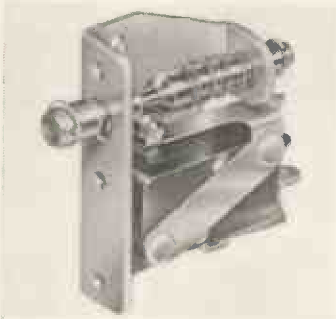
Amplivox Ltd,
2, Bentinck Street,
London, W.1.

Safety Switch

(Illustrated below)

THIS new door safety switch provides the usual protection against accidents to operators of cubicled high-voltage equipment but in addition, it affords a simple and quick method of maintaining the equipment in its operative condition while the service engineers require it open.

Full protection is automatically restored as soon as the door is closed



without the engineer touching the switch.

Burgess Products Co. Ltd,
Dukes Way, Team Valley,
Gateshead.

250kW Magnetrons

MULLARD Limited are now producing some additional types of "packaged" magnetrons for the 3-centimeter band. The JP9-250 operates in the frequency range 9 345 to 9 405Mc/s and the JP9-250A in the range 9 003 to 9 138Mc/s. The peak output power is 250kW. Maximum anode voltage and current are 23kV and 27·5A. Pulse durations can be up to 6μsec. The valves are equivalent to U.S. types 4J50 and 4J78.

Mullard Ltd,
Century House,
Shaftesbury Avenue,
London, W.C.2.

Miniature Micro-Switch

(Illustrated above right)

THE Burgess type 4A micro - switch has approximate overall dimensions of ¾ by ½ by ¼in. It will make and break 5A at 250V a.c. or 2A at 30V d.c. It is a single-pole changeover switch with a positive snap action. Con-

tact separation is 0·010in minimum.

The switch is operated by depressing a small bakelite plunger, a force of 7oz (max) being required. For applications where it is not possible to apply the operating force at 90 degrees to the alignment of the switch or when the actuating medium is a cam or sliding dog, auxiliary actuators with either plain levels or roller levers are available.

Burgess Products Co Ltd,
Dukes Way, Team Valley,
Gateshead 11.

Heavy Duty Balancing Machine

(Illustrated below)

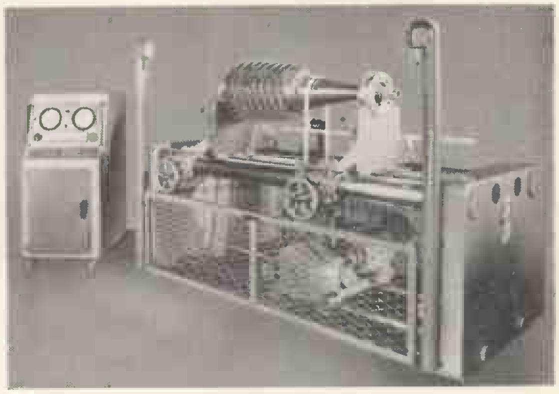
THIS newest and largest of the range of dynamic balancing machines produced by E.M.I. is particularly suitable for balancing gas turbine rotors, impellers and other rotating parts weighing from 75 to 1 500lb, and with lengths and diameters up to 60in and 36in respectively.

The sensitivity is such that forces displacing the centre of gravity by amounts between ·00002 and 0·06in (0·5 to 1 500 microns) can be measured quickly and precisely.

The rotor under test, mounted if desired in its own bearings (up to 10in outer diameter), is belt driven at a convenient speed between 250 and 1 200 rev/min (determined by the weight of the test-piece). The polar position and amount of any unbalance in the two predetermined correction planes are then immediately and directly indicated on large dials on the control console. The operator does not have to make any adjustments or identify points of resonance.

The machine is fitted with a safety guard, and a system of interlocks prevents accidental misuse.

E.M.I. Ltd,
Hayes, Middlesex.



MEETINGS THIS MONTH

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 30 November. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, W.C.1.
Lecture: High Fidelity Loudspeakers: The Performance of Moving-Coil and Electrostatic Transducers.
By: H. J. Leak.

North-Western Section

Date: 3 November. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester.
Lecture: Ground Controlled Approach and the Instrument Landing System.
By: R. H. James and N. MacKinnon.
To be preceded by the film "In on the Beam".

North-Eastern Section

Date: 9 November. Time: 6 p.m.
Held at: Neville Hall, Westgate Road, Newcastle upon Tyne.
Lecture: Turret Tuners for Multi-Channel Television Reception.
By: R. Holland.

Merseyside Section

Date: 9 November. Time: 7 p.m.
Held at: The Council Room, Chamber of Commerce, 1 Old Hall Street, Liverpool 3.
Lecture: The Dekatron Tube in a Digital Transmission System.
By: G. Shand.

West Midlands Section

Date: 9 November. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: Closed Circuit Industrial Television.
By. H. A. McGhee.

Scottish Section

Date: 10 November. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: Some Applications of Electronics to Marine Echo-Sounding.
By: F. Baillie.

THE BRITISH KINEMATOGRAPH SOCIETY

Date: 23 November. Time: 7.15 p.m.
Held at: The Gaumont-British Theatre, Film House, Wardour Street, London, W.1.
Lecture: The Display of Television Pictures in the Home.
By: E. P. Wethey.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.
Date: 3 November.
Lecture. The New High-Frequency Transmitting Station at Rugby.
By: C. F. Booth.
Date: 10 November.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.

Radio and Telecommunication Section

Date: 9 November.
Lecture: A Transistor Digital Fast Multiplier with Magneto-Strictive Storage.
By: G. B. B. Chaplain, R. E. Hayes and A. R. Owens.
Date: 21 November.
Discussion: The Reception of Band I and Band III Television Programmes.
Opened by: E. P. Wethey.

Measurement and Radio Sections

Date: 15 November.
Lectures: An Electrolytic-Tank Equipment for the Determination of Electron Trajectories, Potential and Gradient.
By: D. L. Hollway.
A Method of Tracing Electron Trajectories in Crossed Electric and Magnetic Fields.
By: J. H. Westcott.

Mersey and North Wales Centre

Date: 21 November. Time: 6.30 p.m.
Held at: The Town Hall, Chester.
Lecture: Control of d.c. Machines.
By: E. P. Hill and A. A. L. Bentall.

North-Eastern Centre

Date: 4 November. Time: 6.15 p.m.
Held at: The Neville Hall, Newcastle upon Tyne.
Lecture: Transatlantic Telephone Cable.
By: M. J. Kelly, Sir Gordon Radley, G. W. Gilman and R. J. Halsey.

North-Eastern Radio and Measurements Group

Date: 7 November. Time: 6.15 p.m.
Held at: King's College, Newcastle upon Tyne.
Lecture: An 8-MeV Linear Accelerator for X-Ray Therapy.
By: C. W. Miller.
Date: 21 November. (Time and place as above).
Lecture: Electrical and Magnetic Measurements in an Electrical Engineering Factory.
By: D. Edmundson.

North Midland Centre

Date: 1 November. Time: 6.30 p.m.
Held at: The Central Electricity Authority, Yorkshire Division, 1 Whitehall Road, Leeds.
Lecture: A Transatlantic Telephone Cable.
By: M. J. Kelly, Sir Gordon Radley, G. W. Gilman and R. J. Halsey.

North-Western Radio and Telecommunication Group

Date: 9 November. Time: 6.45 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: High-Speed Electronic-Analogue Computing Techniques.
By: D. M. MacKay.

Northern Ireland Centre

Date: 8 November. Time: 6.30 p.m.
Held at: Lecture Room A, Engineering Department, Queen's University, Belfast.
Lecture: Maintenance Principles for Automatic Telephone Exchange Plant.
By: R. W. Palmer.

North Scotland Sub-Centre

Date: 8 November. Time: 7.30 p.m.
Held at: The Caledonian Hotel, Aberdeen.
Lecture: Thermionic Valves of Improved Quality for Government and Industrial Purposes.
By: E. G. Rowe, P. Welch and W. W. Wright.
Date: 9 November. Time: 7 p.m.
Held at: Electrical Engineering Department, Queen's College, Dundee.
Lecture: Thermionic Valves of Improved Quality for Government and Industrial Purposes.
By: E. G. Rowe, P. Welch and W. W. Wright.

South Midland Radio and Telecommunication Group

Date: 28 November. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: A Radio Position Fixing System for Ships and Aircraft.
By: C. Powell.

Southern Centre

Date: 30 November. Time: 6.30 p.m.
Held at: Brighton Technical College.
Lecture: The Breaking of a.c. Circuits and the Short Circuit testing of a.c. Switches.
By: E. W. Goodman.

Maidstone District

Date: 1 November. Time: 7 p.m.
Held at: The Prince of Wales Hotel, Railway Street, Chatham.
Lecture: The Possibilities of a Cross-Channel Power Link between the British and French Supply Systems.
By: D. P. Sayers, M. E. Laborde and F. J. Lane.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 8 November. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Lecture: Economic Principles of Telecom's Plant Provision.
By: N. V. Knight.

RADIO SOCIETY OF GREAT BRITAIN

Date: 11 November. Time: 6.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place. London, W.C.2.
Lecture: Compressed Beams.
By: G. A. Bird.

SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 29 November. Time: 7 p.m.
Held at: Manson House, Portland Place, London, W.1.
Lecture: Electronic Computing Methods.
By: A. St. Johnston.

THE TELEVISION SOCIETY

Date: 11 November. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: The Application of Semi-Conductor Diodes to Television Circuits.
By: J. I. Missen.
Date: 24 November. (Time and place as above).
Lecture: Interference with Television Reception: Its causes and cures.
By: R. A. Dilworth.

PUBLICATIONS RECEIVED

PROCEEDINGS OF THE EASTERN JOINT COMPUTER CONFERENCE gives the papers and discussions presented at The Joint Computer Conference held in Philadelphia in December, 1954. American Institute of Electrical Engineers, 33 West 39th Street, New York 18, N.Y. Price $3.00.

NICKEL ALLOYS IN ELECTRONIC VALVES describes and illustrates some of the uses of wrought nickel and nickel alloys in electronic valves. Copies of the booklet may be obtained, free of charge, from The Publications Department, Henry Wiggin & Co Ltd, Thames House, Millbank, London, S.W.1.

THE SUPPRESSED FRAME SYSTEM OF TELERECORDING is the first of a series of monographs, which the BBC has decided to publish, on technical developments related to sound and television broadcasting. Each monograph will describe work that has been done by the Engineering Division of the BBC and will include, where appropriate, a survey of earlier work on the same subject. It is proposed to publish about six monographs each year. They will cost 5s. each post free; an annual subscription of £1 will cover all the issues in one year, post free. Orders should be placed with BBC Publications, 35 Marylebone High Street, London, W.1.

TECHNIQUE ET APPLICATIONS DES TRANSISTORS, by H. Schreiber, is a well presented and illustrated book by an author who has had considerable experience in the field of transistors. Societe des Editions Radio, 9 Rue Jacob, Paris 6e. Price Fr.720.

BASIC ELECTRONICS in four volumes gives the texts of the entire basic electricity and basic electronics courses, as currently taught at U.S. Navy speciality schools, the information having recently been released for civilian use. John F. Rider Publisher, Inc, 480 Canal Street, New York, 13. Price $2 per volume.