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Commentary

IT has not been an easy matter in the past to judge the rate of scientific progress in the Soviet Union and it has been mistakenly assumed that the lack of news from behind the Iron Curtain is an indication that the Russians are a long way behind the rest of the world.

There have, of course, been exchange visits in recent years between British scientists and their Russian counterparts, but the reports which have been published by our scientists on their return do not always give a complete view of the Russian scene and with the present state of international relationships it is perhaps too much to hope that they would.

However, they do show, as one would expect, that in a country which entered the scientific and industrial era at a late stage, there are deficiencies here and there, but where effort has been concentrated in a particular direction the results have been quite outstanding for all the world to see, and if any emphasis were needed in this respect the latest Russian successes in their space programme including the photographing of the remote side of the moon have provided all that is required.

The exchange visits have been sufficiently frequent and varied in character to show that the Russian effort has not been concentrated on space research alone to the exclusion of all other fields and the signs are that considerable attention is now being devoted to developing the country's industrial strength.

Some further light has been thrown on this subject by a two-week visit to the Soviet Union by a team of six British engineers, led by Professor A. Tustin of Imperial College, to study progress in automatic control applied to industrial production and their report has just been published by the Department of Scientific and Industrial Research.

The visit was made in May of this year following an earlier visit to this country by six Soviet engineers mainly from the Institute of Automatics and Telemechanics in Moscow and led by its Deputy Director, Professor B. S. Sotskov.

This visit to the Soviet Union, which was the first of its kind in the field of automatic control, was exploratory in nature and was intended to do no more than to open up channels of communication and to establish personal relationships. Nonetheless the team were able in the two weeks available to visit a number of factories and research and teaching institutes in Moscow, Leningrad and Kiev. They visited the Institute of Automatics and Telemechanics where the highlight was the development of an automatic control system for a four-stand tandem mill which the team found interesting as an advanced theoretical study, but saw no evidence of its practical use. Included in the itinerary was a visit to the Institute of Automation of the Ukraine at Kiev which concerns itself with problems of automation and instruments for automation in heavy industry including chemistry, metallurgy and mines as well as electric power systems.

Their conclusions were that work in progress on automatic control in the Soviet Laboratories is being applied on an increasing scale to all industrial applications and extensive use is being made of both analogue and digital computers although the work on the whole was not so advanced as it is in this country and chemical plant process control is cited as a case in point.

On the other hand because of the particular needs of the Soviet Union there is considerable activity going on in control systems for the steel industry and integrated process control simulators are being developed to assist in the design of multi-variable control systems and later to operate as a part of more ambitious systems.

In telemetry and remote control effort is not lagging and these principles have already been applied to the control of long gas transmission pipelines and to electrical power networks.

In short, automatic control and automation in industry is being given considerable attention in the Soviet Union and to this end large and well equipped research and teaching centres have been set up from which a vast number of trained engineers will eventually emerge. These centres have been in existence only for the past two or three years so that the impact they have made is not yet very great and there is a vast leeway to make up. Nevertheless the general impression gained by the team is that engineering education is both massive and effective, and they go so far as to say that in a few years the application of automatic control in Soviet industry will have overtaken that in this country. The stage will then be reached when further advance will depend on breaking new ground but the team found insufficient evidence on which to judge how good the Soviet engineers would prove to be at this further stage although their present level of achievement in basic theoretical work is high and in other fields initiative has not been lacking.

The conclusions which may be drawn on reading this report are plain for us to see. It has been known for a long time that the Soviet Union is sparing no effort to establish itself as one of the world's leading industrial powers and the challenge to this country must not be allowed to pass.

The history of our industrial growth has been long and exciting with many "firsts" but we can no longer rely on our past achievements. We are at the beginning of another industrial revolution and we are witnessing the rise in industrial power of large nations hitherto regarded as backward and under developed. We must not be lulled into a sense of false security by the words in the report that "it is easier to catch up than to surpass."

A PROGRAMMED TEST SET

By D. W. Bradfield*, B.Sc., A. M. East*, and H. F. Rourke*, B.Sc., A.R.C.S.

This article describes how, by programming the test sequence with uniselectors, complex electronic equipment may be checked in a fraction of the time required to carry out manual tests. The test results are indicated by means of a 'go-no go' display. The particular programme was evolved to test the power system of a guided weapon, but the principles are generally applicable to a wide range of equipment.

(Voir page 774 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 779)

As a result of the increasing complexity of present day electronic equipment, the task of testing is becoming more difficult. As a consequence, it is desirable to check such equipments by means of automatic test gear. The test set described was specifically designed to test a missile power supply in accordance with a predetermined programme. The tests which are carried out on the various units of the power supply are as tabulated below:

Unit	Details of Test
(1) *Turbo-Alternator*	(1) Run-up time
	(2) The output voltage under both steady state conditions and transient conditions when the turbo-alternator is subjected to a load change.
	(3) The frequency under both steady state conditions and transient conditions when the turbo-alternator is subjected to a load change.
(2) *Rotary Convertor*	(1) The synchronism between the turbo-alternator and the rotary convertor.
(3) *Power Pack*	(1) The output voltages both alternating and direct.
	(2) The ripple level on certain of the lines.
(4) *Programming Switch*	(1) Static tests on the initial conditions.
	(2) A timing check on the switch operation.
	(3) A test to ensure that the switch has reset correctly.
(5) *Gain Controlling Potentiometers*	(1) The total resistance value.
	(2) The slider position.

The operation of a single switch starts a complete test sequence which controls the system under test as well as measuring the various parameters as required. The display consists of 'pass' and 'fail' lamps to indicate the state of individual tests in addition to the main 'Pass' and 'Fail' lamps.

The programme is divided into two parts:

(a) Static tests which are carried out with the turbo-alternator stationary, and

(b) Dynamic tests which are carried out with the turbo-alternator running.

The complete sequence is, therefore, as follows:

(1) A warming-up period for the power pack and rotary convertor during which time they are fed from an external supply simulating the turbo-alternator output.

(2) Static tests on the programming switch initial conditions.

(3) Static tests on the gain controlling potentiometers for total resistance value and slider position.

(4) Automatic start of the turbo-alternator and a subsequent check of its run up time, frequency and voltage and, in addition, a measurement of the latter two parameters when the turbo-alternator is subjected to a load change.

(5) A timing check on the programming switch operation.

(6) Automatic transfer of the power pack and rotary convertor from the warming-up supply to the turbo-alternator.

(7) A check of the output voltage of all the lines of the power pack and, in some cases, a measurement of the ripple level.

(8) A check to ensure that the turbo-alternator and rotary convertor are synchronized.

(9) Shutdown of the turbo-alternator under specified load conditions.

(10) A test to ensure that the programming switch has reset correctly.

The Principle of the Test Set

The principle of the test set is demonstrated in Fig. 1.

The test programme is originated by means of a 'sequence generator' which initiates the commands to operate the 'control mechanism', the 'measurement blanking mechanism' and the 'measurement selector mechanism.' In addition, 'pass' and 'fail' lamps are lit via the sequence generator, the lighting of a particular 'pass' lamp indicating that all tests, up to and including that indicated by the lamp have been passed.

The control mechanism sends the appropriate control signals to the 'stimuli generators' to feed the necessary inputs, including test signals, into the equipment under test. For example, it switches the external supply on and off the power pack and rotary convertor and activates the primary source of power for the turbo-alternator. In addition, the control mechanism is used to cycle the 'dummy load' for the purpose of carrying out transient tests on the turbo-alternator.

The measurement selector mechanism connects the appropriate measuring unit to the parameter to be measured and selects the correct reference level and limits. It should be noted that the measurement is normally inhibited for a finite period, by means of a 'measurement blanking mechanism,' after switching in a measuring unit in order to allow any transients which may have been set up to subside.

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The 'measurement section' contains the necessary circuits for converting the parameters to be measured into suitable form to be checked against a reference by means of various comparators. The references are direct voltages except in the case of measurements involving time intervals when the reference consists of a master timing pattern generated by the sequence generator. In all cases, except one, the pass or fail of a measurement is recorded by the state of a thyratron which, in turn, determines the condition of a relay. In the event of a particular parameter being recorded as being within limits by the measurement section, a 'pass' lamp is lit via the sequence generator on the completion of the test and the programme is allowed to proceed. In the event of a fail being recorded, information is sent back to the sequence generator to stop the test sequence. The missile power system is then switched off via the control mechanism and the relevant 'fail' and the main 'fail' lamps are lit.

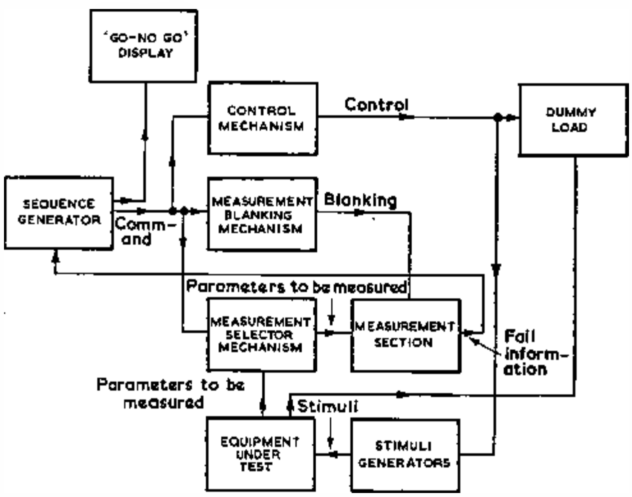

Fig. 1. Simplified diagram of an automatic test set

The Controller

In any automatic device, the controller is the heart of the system. It consists of a programme generator, control mechanism and the measurement selector and blanking mechanisms.

The programme pattern is derived from two six bank, twenty five way uniselectors in series. The operation of the 'start' switch while the 'auto-manual' switch is in the 'auto' position initiates the necessary mechanism to step the first of the uniselectors forward at either 10 or 5 positions per second. This uniselector will be referred to as the fast uniselector and it drives the second uniselector forward one position every 2·5sec. The second uniselector will be called the slow uniselector. The fast uniselector is stepped forward by a relay closing the 50V return on to its energizing coil, this relay in turn being operated by the output of a Wien bridge oscillator and squarer which runs at 10c/s during slow uniselector positions 1-22 inclusive and 5c/s for slow uniselector positions 23 and 24. The mechanism to terminate the run is initiated on position 25 of the slow uniselector when the fast uniselector reaches position 1. Thus are created 601 possible test positions explored in 65sec.

Within the framework of this general pattern, control is exercised by the operation or release of relays by the uniselectors or associated slave relays. The contacts of these relays then provide paths to pass the necessary stimuli to the equipment under test, to route the information received from the equipment under test together with the appropriate references to the measurement section concerned, to remove the blanking at the time of test and to indicate the progress of the test run by lighting pass lamps as each test is successfully passed and appropriate fail lamps should any test fail. This procedure is considerably simplified by dividing the tests into two groups, viz.,

(a) Those which, due to their time and duration of test, can be associated with a particular unique position of the slow uniselector. These are termed 'instantaneous tests' and

(b) Those which extend over longer periods of time and are termed 'continuous tests'.

Instantaneous Tests

Because of their association with particular slow uniselector positions and the fact that a fail in any test immediately stops the test run, the control and recording of these

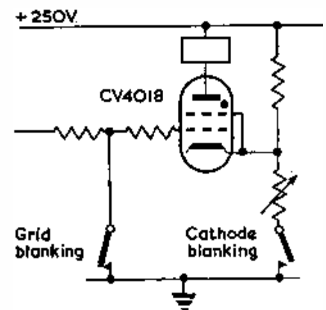

Fig. 2. Fail thyratron blanking

tests is organized by both uniselectors as follows:

(a) The arrival of the slow uniselector at a particular position lights the pass lamp for the previously conducted instantaneous test and operates a relay which:

(1) Passes the parameter under test and an appropriate reference (if required) to the relevant measurement section.

(2) Partially removes the blanking from the fail thyratron for the particular measurement by connecting the lower end of the cathode potential divider chain to earth (see Fig. 2).

(b) The arrival of the fast uniselector at position 18 operates a relay which removes the remaining blanking by disconnecting the grid of the fail thyratron from earth. Thus a delay of 1·8sec is achieved after the measurement unit is switched in before any test result is recorded and this ensures that any transients which might have been set up have completely subsided.

(c) The arrival of the fast uniselector at position 23 releases the relay described under (b) thus restoring the grid blanking and terminating the measurement.

(d) Should the measurement be outside the permitted limits, the fail thyratron is fired which, in turn, operates a fail relay in its anode. A fail signal is returned from the measuring unit to the controller where the 'instantaneous test fail relay' is energized. This relay then lights the appropriate fail lamp via one of the banks on the slow uniselector and energizes the 'main fail relays' which in turn initiate the stop and shut-down actions. The shut-down action reimposes the blanking on all measurements to ensure that no additional fails are recorded when the system under test is closed down.

Continuous Tests

The problem of controlling the continuous tests is considerably simplified by the facts that they are few in number and that no two tests are of the same type. Consequently it is arranged that the relevant parameter is fed to the appropriate measurement section continuously and the relevant test periods are determined by the blanking mechanism. In general the blanking mechanism is similar to that for instantaneous tests except that the relay initially energized by the uniselectors remains energized until released by the operation of another relay (i.e. it is not simply energized for one slow uniselector position). In certain cases, however, because of the need for flexibility, or for a time of commencement and/or duration not associated with uniselector positions, the blanking is organized by Miller delay circuits triggered by an appropriate reference. These apply particularly to the master timing pattern generated for the programming switch measurement and to the blanking and limit changes on the frequency measurements under transient conditions. As in the case of instantaneous tests, should the measurement fall outside limits a fail signal is returned to the controller but, in this case, energizes the 'main fail relays', the appropriate fail lamp being lit directly by an additional signal from the measurement section. The passing of the continuous tests is indicated by the lighting of the appropriate pass lamps just prior to the termination of the test sequence, i.e., after the continuous tests have been completed.

Due to the nature of the equipment under test, i.e., a system in which a failure of certain parameters may be catastrophic if allowed to continue indefinitely and in which few parameters are adjustable on site to convert a fail to pass, the philosophy adopted with this equipment was that a fail of any test, besides stopping the test, would also switch off the system under test. For the purposes of test and calibration of the test set itself, however, 'single shot', 'clear fail' and 'override' facilities have been included. When the 'auto-manual' switch is set to 'manual' the fast uniselector coil is connected to a 'single shot' switch which allows the fast uniselector to move round one position every time the 'single shot' switch is released. The 'clear fail' switch breaks the earth return of the fail thyratrons thus extinguishing any that have fired and also removes the 50V from the holding contacts of the instantaneous test fail relay and other fail relays which energize it. The 'override' switch releases or prevents the energizing of the main fail relays which initiate the stop action. Thus, for a more general type of equipment in which adjustment is both possible and permissible these facilities could play an important part in producing fully adjusted systems.

The successful completion of a test run is indicated by the lighting of a prominent green main 'pass' lamp in addition to the pass lamp of the final test and a lamp indicating that the test run has been terminated. A fail during the sequence is indicated by the lighting of a prominent red main 'fail' lamp in addition to the appropriate fail lamp indicating the unsuccessful parameter and the 'test run terminated' lamp.

At the termination of the test run, whether successful or not, the equipment is reset by operating the 'reset' switch which initiates the necessary mechanism to connect the uniselector coils to their homing banks and removes the energizing supply from the other relays, thus restoring the equipment to its initial condition.

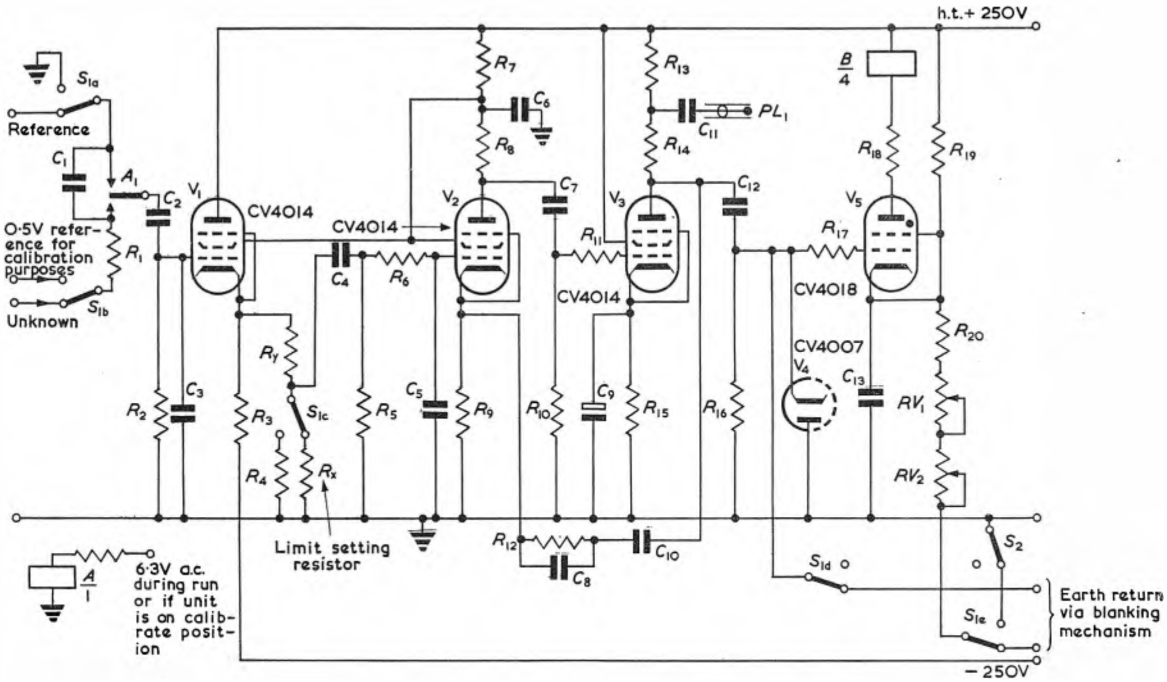

Fig. 3. Simplified circuit of voltage measuring unit

S_1 *run calibrate*
S_2 *clear thyratron*
RV_1 *set fail level—coarse*
RV_2 *set fail level—fine*

Voltage Measurement

The d.c. voltages from the missile power pack are checked by comparing them against precision references and using the 'error signal' to fire a 'fail' thyratron if the voltage under test is outside its permitted limit.

A simplified circuit diagram of the voltage measuring unit is shown in Fig. 3.

The voltage to be tested is routed on to the bottom plate of C_1 via R_1 and its corresponding reference to the top plate by the measurement selector mechanism. Consequently a voltage equal to the error is built up across this capacitor which is subsequently converted into a square wave by means of a chopper relay, A, and amplified in order to obtain a signal whose amplitude is sufficient to ensure reliable operation of the thyratron (V_5). It will be appreciated that the thyratron is set to fire at a discrete voltage level and, consequently, it is necessary to ensure that the square wave is at this value when the voltage under test is at its limit. This is ensured by means of a potential divider (R_yR_x) interposed between the cathode-follower (V_1) and the two stage feedback amplifier (V_2, V_3) in which R_x is chosen by the measurement selector mechanism to determine the limit of the voltage under test. Since the thyratron is set to fire with an effective 'error signal' of 0·25V, it can be shown that for a limit of L volts the value of the limit setting resistor is:

$$R_x = \frac{R_y}{4L - 1} \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(1)}$$

In the event of the fail thyratron being fired, information is fed to the sequence generator by contacts on relay B to stop the test sequence and light the appropriate 'fail' lamp.

It should be noted that R_1 and C_1 form a low-pass filter to prevent any ripple that might be on the line under test causing any appreciable error in the measurement. The d.c. restorer (V_4) is used to double effectively the signal that is fed into the thyratron and to prevent any changes in the mark-to-space ratio of the chopper from modifying the fail level due to a change in the average value of the signal.

The steady state output of the turbo-alternator and the l.t. lines of the power pack are measured by converting the a.c. to d.c. by means of two detectors whose output is proportional to the average value of the input, and using the same measuring unit as is used for the d.c. lines.

The necessary reference voltages are derived from two divider chains built up of 0·1 per cent precision wire wound resistors across the positive and negative 300V lines of a power pack which has a drift of 0·1 per cent per 1 000 hours.

The voltage measuring unit is calibrated by feeding a 0·5V reference voltage into the unknown terminal with the reference terminal connected to earth and adjusting the 'set fail level' controls until the thyratron fires, this being indicated by a pilot lamp lit via a pair of contacts on relay B. The average reading detectors are calibrated by feeding in a signal, checked by means of a dynamometer type meter, equal to the nominal output of the line in question and nulling the output of the detector against its reference. The positive and negative 300V lines of the reference power pack are set up by means of a precision grade 1 moving-coil meter.

In the event of the limit not being symmetrical, the reference is chosen to be equal to the average of the upper and lower voltage level corresponding to the limit values and the limit is made equal to the average of the upper and lower limit.

Measurement of Resistance

The value of an unknown resistance can be checked by means of the voltage measuring unit by measuring the voltage at the junction of a potential divider made up of two resistors with a 0·1 per cent tolerance with the lower limb shunted by the unknown resistor as shown in Fig. 4. The Thévenin equivalent circuit of this is shown in Fig. 5 and hence it will be seen that the test voltage is given by:

$$V_T = \frac{V_{oc}R_x}{R_x + R_{in}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(2)}$$

Now if R_x has a fractional tolerance of x, the voltage at the junction when the resistance under test is at its lowest limit is:

$$V_{T(min)} = \frac{V_{oc}(1 - x)R_x}{R_{in} + R_x(1 - x)} \quad \ldots\ldots\ldots \text{(3)}$$

and when the resistance is at its upper limit:

$$V_{T(max)} = \frac{V_{oc}(1 + x)R_x}{R_{in} + R_x(1 + x)} \quad \ldots\ldots\ldots \text{(4)}$$

and hence the reference against which the junction voltage must be compared is chosen to be equal to the arithmetic mean of equations (3) and (4) whereas the limit voltage is equal to:

$$V_{(limit)} = \frac{V_{T(max)} - V_{T(min)}}{2} \quad \ldots\ldots\ldots \text{(5)}$$

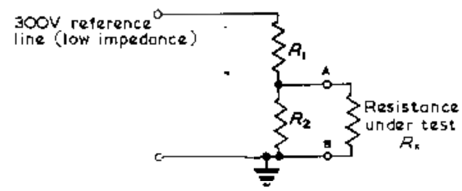

Fig. 4. Test circuit for resistance

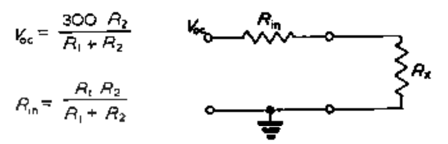

Fig. 5. Thévenin equivalent of test circuit for resistance

R_1 and R_2 are chosen to obtain adequate sensitivity and it should be noted that R_2 provides a path to charge the 'error signal' capacitor in the voltage measuring unit in the event of the resistance under test being open-circuited.

Voltage Transient Measurements on Turbo-Alternator

The transient response of the turbo-alternator voltage output is checked by instantaneously increasing the load and checking that the voltage recovers to the required values during each of two discrete time intervals after the application of the increased load. In addition, a check is conducted to ensure that the overshoot is not excessive.

A simplified circuit diagram of the transient measuring unit is shown in Fig. 6. When the load is increased by the control mechanism, a signal is simultaneously fed on to the grid of a thyratron (V_4) and causes this to fire, thus generating a positive going step function at its cathode. This step function is then differentiated to provide a triggering pulse to activate the cathode coupled monostable multivibrators (V_3) in three similar channels. These multivibrators produce gate periods corresponding to the measurement period to feed on to the screen of a thyratron coincidence circuit. These thyratrons can only strike, however, if a positive signal is simultaneously applied on to their grids. The grid input signal is derived by clipping the negative peak to the turbo-alternator output voltage by means of a clipping circuit (MR_1, MR_2) whose clipping level is set to correspond to the test level. Thus a signal is only obtained from the clipper if the turbo-alternator output exceeds the clipping level. The clipped peak is then amplified and inverted by V_1 and will fire the thyratron coincidence circuit providing the gate is still open. It will, therefore, be appreciated that the thyratrons in the 'speed of recovery' channels are fired by a 'pass' signal, while in the case of the 'overshoot' channel the thyratron is fired by a 'fail' signal.

In the anode of the thyratron in each 'speed of recovery' channel there is a relay with normally closed contacts which are opened by the thyratron firing, whereas in the anode of the thyratron in the 'overshoot' channel there is a relay with normally open contacts which can only be closed by the thyratron firing.

The contact position of these relays are sampled at the completion of the test period and in the event of any one being in the closed position a stop action is initiated.

In order to calibrate the unit an astable multivibrator is switched on which provides recurrent square waves which can be differentiated to provide a train of trigger-ing pulses to drive the gate generators. The gate lengths can then either be set up by means of an oscilloscope with a precision time-base, or a counter. The clipper levels can be set up by feeding a signal corresponding to the required level into the unit and adjusting the clipper level until the relevant thyratron fires, this being indicated by a pilot lamp operated by a pair of contacts on the relay in the anode circuit.

relays A, B, C and D are those controlled by the screen currents of the Miller delays and are energized except during the period of the delay, and relay G is a slave relay of the contact under test which is energized when the contact makes and released when the contact breaks.

The sequence of events in checking the make time of a contact then becomes as follows:

(a) The delays controlling relays A and B are triggered by the reference event and contacts A_1 and B_1 assume the position shown.

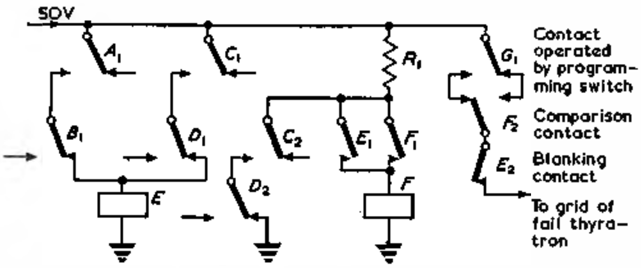

Fig. 7. Typical test circuit for the programming switch

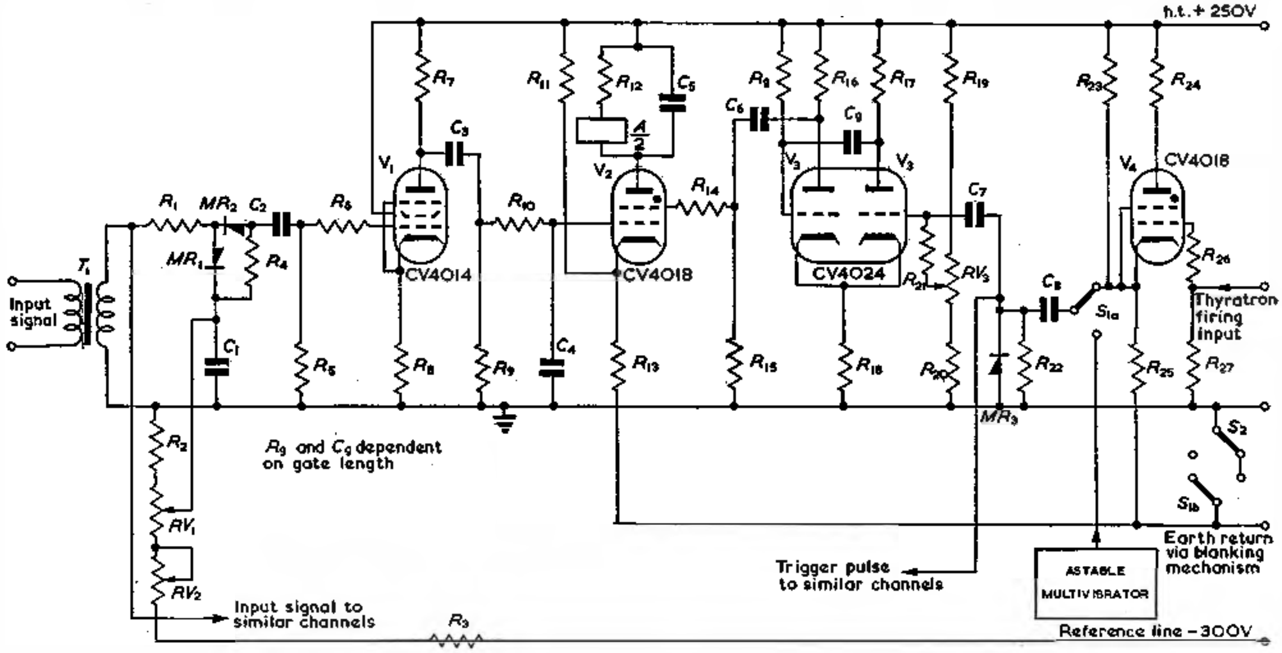

Fig. 6. Simplified circuit of transient measuring unit

S_1 run calibrate

S_2 clear thyratron

RV_1 set clipper level—coarse

RV_2 set clipper level—fine

RV_3 set gate length

Programming Switch Measurement

In order to check the timing of the programming switch it is necessary to ensure that the various contacts operate within the appropriate time intervals. Since the contacts operate relative to some other event either in the test set or the equipment under test, the technique adopted is to blank the measurement between the appropriate time limits and to sample the contact position by means of relay operated comparison contacts at other times. The time intervals when the comparison contacts operate is con-trolled by Miller delays operating relays, the delays being triggered by the appropriate reference event or suitable time-associated event. The circuit arrangement for checking the make and subsequent break times of a representative programming switch contact is as shown in Fig. 7, where

(b) Delay A terminates: relay A re-energizes; relay E becomes energized by 50V via contacts A_1 and B_1; the blanking is applied by contact E_2; relay F becomes energized by 50V via R_1 and contact E_1 and is subsequently held by contact F_1.

(c) Delay B terminates; relay B re-energizes; relay E is released by the breaking of contact B_1; the blanking is removed by contact E_2.

Thus, provided that contact G_1 does not move across before contact E_2 opens and has moved across before con-tacts E_2 closes again (contact F_2 having moved across in the interim) no path is formed for 50V to the grid of the fail thyratron and no fail is recorded.

A similar sequence of events checks the breaking of the contact, the blanking period being set by contacts C_1 and

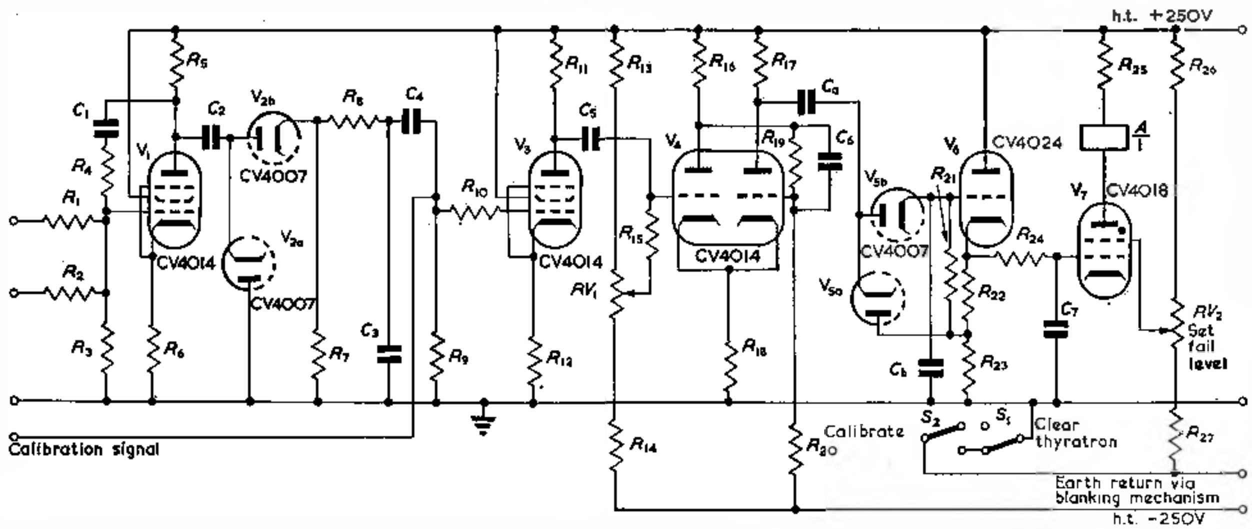

Fig. 8. Simplified circuit of synchronism measuring circuit

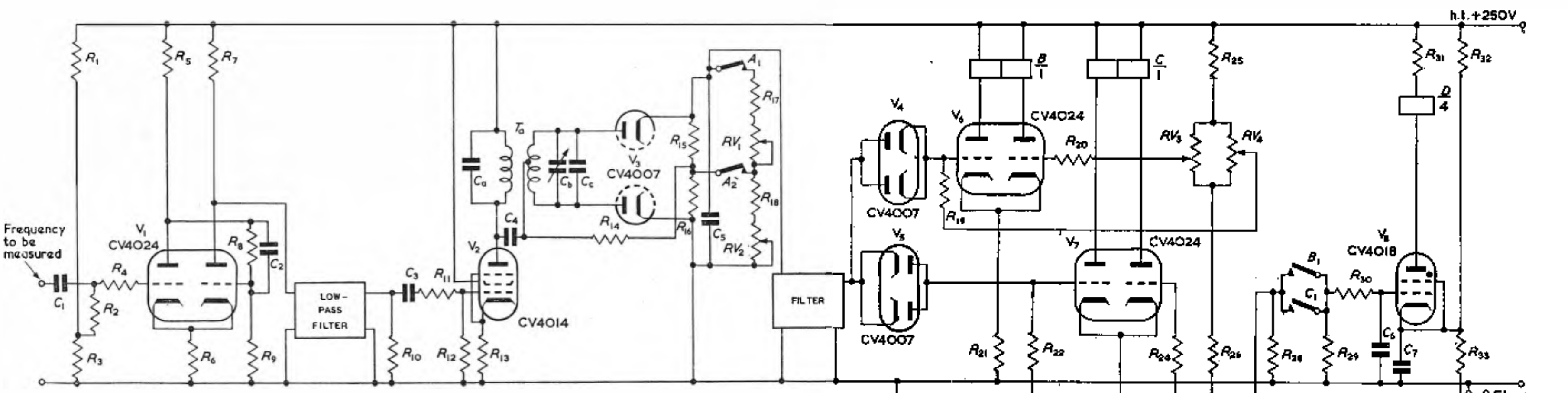

Fig. 9. Simplified circuit for frequency measuring unit

T_a, C_a, C_b, C_c — *dependent upon frequency*

RV_1 *adjust discriminator damping*
RV_2 *adjust discriminator damping*
RV_3 *adjust upper limit reference level*
RV_4 *adjust upper limit fail level*
RV_5 *adjust lower limit reference level*
RV_6 *adjust lower limit fail level*

D_1 and relay F being restored to its initial position by contacts C_2 and D_2. In this case, however, since the operation of relay F is not directly associated with that of relay E, it is necessary to choose relays to fulfil the following requirements:

(a) That the make time of relay E is less than the short-circuit break time of relay F which ensures that contact E_2 opens and applies the blanking before contact F_2 moves across, and

(b) That the make time of relay F is longer than the open-circuit break time of relay E which ensures that, when relay D re-energizes, relay E is released before relay F can re-energize via contact E_1.

The apparent measurement inaccuracies due to relay make and break times are eliminated by calibration techniques which measure the time lapse between the application of 50V to the coil of the slave relay (i.e., relay G) and the operation of the blanking relay (i.e., relay E).

Synchronism Measurement

The synchronism between the turbo-alternator and rotary convertor is checked by ensuring that the difference frequency between the fundamental of the turbo-alternator and the appropriate harmonic of the rotary convertor output corresponding to the turbo-alternator fundamental does not exceed its permitted limit.

A simplified circuit diagram of the synchronism measuring unit is shown in Fig. 8 from whence it will be seen that the output of the turbo-alternator and the harmonic of the rotary convertor output are added by means of a feedback summing amplifier (V_1) whose output is fed into a detector (V_2) to obtain the difference frequency. The resultant signal is then fed via a filter and amplifier (V_3) on to the grid of a bistable cathode coupled multivibrator (V_4) from whose anode is obtained a train of square waves whose recurrence frequency corresponds to the difference frequency. The square wave train is then fed into a ratemeter[1] (V_5, V_6) whose d.c. output is proportional to the frequency of the input. This d.c. output is then compared against a d.c. reference by means of the thyratron comparator (V_7) and in the event of the difference frequency being outside its permitted limit the thyratron is fired which in turn energizes relay A. The fail information is then fed via contacts on this relay to stop the test sequence and light the 'sync' fail lamp. It should be noted that the diode (V_{5a}) of the ratemeter is 'boot-strapped' to the output of the cathode-follower (V_6) in order to improve the linearity of the circuit and the sensitivity in the fail region[2].

The necessary harmonic of the rotary convertor is obtained by feeding the output into a half wave rectifier and extracting the required harmonic by means of a band-pass filter. A half wave rectifier is used because in the equipment being described the relevant harmonic required was an even one, and since a half wave rectified sine wave does not contain any odd harmonics the problem of filtering was simplified. The harmonic so obtained is then fed via a two stage feedback amplifier into the summing amplifier previously mentioned.

The unit is calibrated by feeding a signal at the difference frequency into the amplifier (V_3) and adjusting the 'set fail level' control (RV_2) until the thyratron fires, this being indicated by the lighting of a pilot lamp via a pair of contacts on relay A.

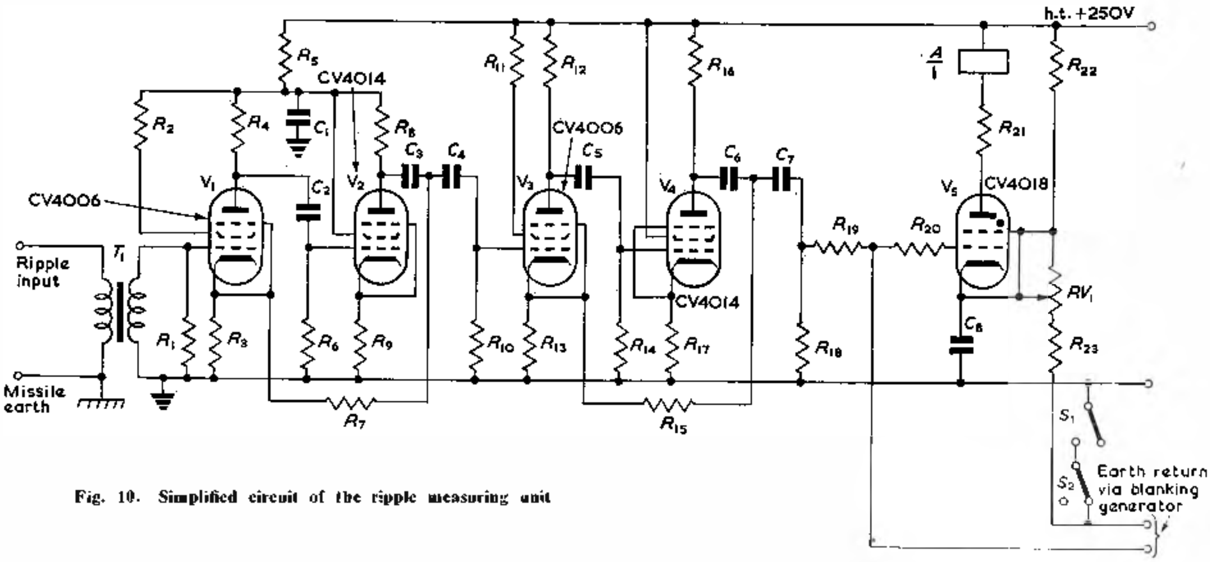

Fig. 10. Simplified circuit of the ripple measuring unit

Frequency Measurement

The frequency of the turbo-alternator is measured by means of a discriminator centred on the nominal frequency of the turbo-alternator, which converts any error in frequency to a d.c. output which can be compared against d.c. voltages representing the upper and lower limits.

As can be seen from the simplified circuit diagram of the frequency measuring unit (Fig. 9), the output of the turbo-alternator is first fed into a bistable cathode coupled multivibrator (V_1) whose output is then fed into a low-pass filter to obtain a sine wave at the turbo-alternator output frequency whose amplitude is independent of the signal level of the input to the multivibrator. This limiter has been included because of the considerable difference in level of the turbo-alternator output and the calibration signal. The signal so obtained is then fed into a Foster-Seeley discriminator[3] (V_2, V_3), whose d.c. output is fed into two long-tailed pair comparators (V_6, V_7) via a filter and the diode gates (V_4, V_5), the purpose of the latter being to prevent the comparator grid returns from damping the discriminator output.

In the event of the frequency falling outside limits, the relevant long tailed pair comparator changes the state of a high speed relay in its anode circuit. The fail thyratron (V_8) is then fired by having 50V fed on to its grid by the contacts of the comparator relay. As a consequence, relay D is energized and the relevant fail information is fed back into the sequence generator to stop the test run and light the 'frequency' fail lamp.

In addition to the steady state checks, certain transient tests are carried out on the frequency when the turbo-alternator has an instantaneous load change applied. These transient tests are normally carried out by allowing a relaxed limit for a certain duration after the load change and checking that the frequency recovers to within its normal limits during this time. It will, therefore, be appreciated that for a given reference level it is necessary to reduce the slope of the discriminator for the period that the relaxed limit is permitted. This is done by de-energizing relay *A* which then inserts damping resistors across the discriminator.

The discriminator is adjusted on to its centre frequency by inserting the output of a crystal oscillator into the frequency measuring unit and adjusting the discriminator tuning capacitance (C_a) for zero output as indicated on a

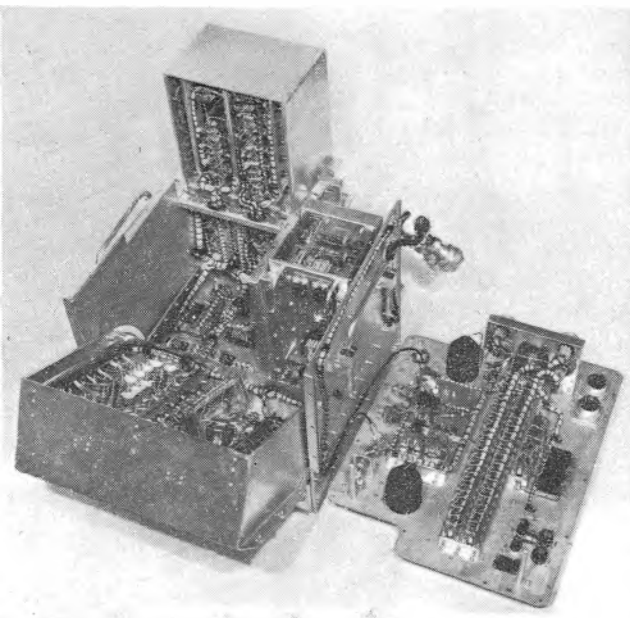

Fig. 11. Mechanical construction of the controller

valve-voltmeter placed across the discriminator load. The comparators are set by noting the output of the discriminator when frequencies corresponding to the nominal fail values are fed into the unit and setting the 'reference level' and 'fail level' controls ¼V above and below this level respectively.

Ripple Measurement

The ripple is measured by feeding it through an amplifier and using the amplified signal to fire a 'fail' thyratron in the event of the permitted fail level being reached.

As can be seen from the simplified circuit diagram (Fig. 10) the ripple to be measured is fed through two stages of feedback pairs (V_1, V_2 and V_3, V_4) via an input transformer (T_1) whose purpose is to isolate the power supply earth and measuring unit earth in order to minimize errors due to earth return currents. The feedback amplifiers[4,5] have been designed to be flat over the frequency band of interest and it should be noted that the input stage (V_1) uses a reliable low noise, low microphony pentode to minimize errors due to these causes, the error due to microphony when the equipment is vibrated having been found to be particularly severe with normal valves. In the event of the ripple exceeding the permitted limit, the thyratron V_5 is fired and the closure of contacts on relay *B* transmits the necessary fail information to the sequence generator.

It will be appreciated that the fail level is not identical on all lines and potential dividers are provided prior to the transformer to reduce all fail levels to that set up on calibration. In addition, the ripple that can be tolerated is dependent upon the frequency and hence it is necessary to include weighting filters in the form of low-pass and high-pass filters in the noise measuring channels. Where the weighting characteristic is common to all lines it has been included in the amplifier but, in the cases where different characteristics are required for the various lines, the filters have been included prior to the transformer and are chosen by the selector mechanism.

In order to prevent any breakthrough from the chopper in the voltage measuring unit introducing an error on the ripple measurement, the ripple on any particular line is measured one position ahead of its corresponding voltage measurement. The lamp indicating which line has failed, however, is operated from the same mechanism as for a voltage fail. Hence the fail information is stored until the

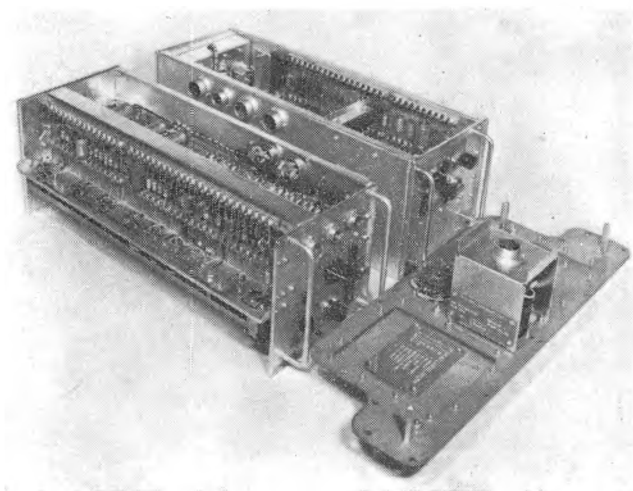

Fig. 12. Mechanical construction of a typical measuring unit

line in question is prepared for voltage measurement when the fail mechanism is immediately operated. In this case, however, the 'ripple' fail lamp is lit in addition to that indicating which line has not met its specification.

In order to calibrate the unit a signal corresponding to the normal fail level is fed into the unit and the 'set fail level' control adjusted until the thyratron fires as indicated by a pilot lamp being lit by a pair of contacts on relay *A*.

Conclusions

The article has demonstrated the basic principle of an automatic test set which is used to check the power supply of a guided weapon. The capabilities and potentialities of this type of equipment are shown by the fact that thirty-seven tests are conducted in sixty-five seconds with, at times, as many as five tests being conducted simultaneously. This automation has resulted in reduced test time, lower level of skill to carry out the tests and more consistent results.

Acknowledgments

The authors wish to thank The Plessey Company Limited for permission to publish this article.

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Automatic Control of Distortion in Wideband Frequency Modulated Microwave Links

By J. Tolman,* B.Sc.(Eng.), A.C.G.I., A.M.I.E.E.

This article discusses the problem of distortion generated in broad band frequency modulated microwave links by non-uniform group-delay characteristics in r.f. amplifiers. An experimental automatic equalizer correcting group delay slope is described and results obtained on a 600 channel circuit are given. Methods of correcting higher orders of distortion are indicated.

(Voir page 774 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 779)

IN the development of broad-band microwave systems carrying many hundreds of telephony channels by means of an f.m. carrier, it is found that one of the main sources of distortion is that due to a non-uniform group delay characteristic in any amplifier transmitting the f.m. signal. The distortion may arise in various parts of the system such as i.f. amplifiers, microwave filters, travelling-wave tubes, aerial feeders or as a product of disturbed propagation conditions. In general, suitable fixed equalizers can be inserted at appropriate points to correct for most of the delay distortion. At i.f., for instance, a combination of one or more all-pass delay equalizers can provide the necessary characteristic.

With extensive systems however, the overall group-delay characteristic will change with time due to temperature, mis-alignment etc. leading to a degradation of performance. It is desirable, therefore, to provide a means of 'mopping-up' the distortion automatically if possible to avoid undue maintenance and testing.

Before describing the means of achieving this automatic control it will be useful to determine the order of the correction needed.

Group-delay Requirements

Any broad-band multi-channel system designed to conform with C.C.I.R. standards must not introduce more than a certain amount of noise into the circuit. For the purpose of specification the C.C.I.R. has introduced the fictitious reference circuit of 2 500km in length and carrying 120 channels or more. The reference circuit can normally be divided into nine sections of 280km, each being terminated in a radio modulator or demodulator and associated channelling equipment (Fig. 1).

It is laid down[1] that at a zero reference point in a channel (300 to 3 400c/s) at the receiving terminal the noise power must not exceed 7 500pW (pshophometric) for more than 1 per cent of the busiest hour. The frequency translating equipment is allocated 2 500pW and thus 5 000pW can be assigned to the radio equipment. The maximum power which can be introduced by one 280km section is 5 000/9 = 556pW or 1 110pW if the noise is measured in a 0 to 4kc/s band with an unweighted quadratic meter.

Taking an average radio path length of about 56km (35 miles) there will be four repeaters in the standard section. The noise power can be further sub-divided into intermodulation and basic (thermal) noise in the various amplifiers. If 250pW, say, is allowed for basic noise for the whole 280km section, this leaves 860pW due to

(a) Modem (modulator-demodulator) transfer non-linearity.

(b) Group delay distortion in modem and five transmitter pairs.

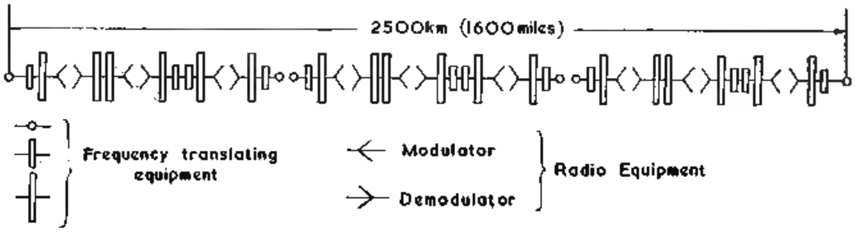


Fig. 1. General arrangement of the C.C.I.R. reference circuit for radio relay systems

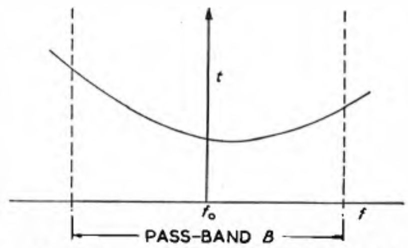


Fig. 2. Typical group-delay characteristic of a radio link

Allowing 70pW for (a) and 30pW for echo distortion in each feeder (10 in all), then 490pW remains for the r.f. and i.f. circuits. It is necessary to separate the contribution due to feeder echoes because as explained later these cannot be simply corrected.

The allocation of power in the system is, of course, a matter for the designer, but the preceding treatment has been based on performances attainable with practical systems of up to 960 channels capacity.

Distortion as a Function of Group-Delay

The group-delay characteristic of a radio link (Fig. 2) will naturally be quite complicated, but, neglecting feeder echoes, the variation due to amplifier and filters etc. can normally be considered to be a combination of a linear slope together with a quadratic variation in the pass-band.

$$\text{i.e. } t = t_1 + t_2 \frac{f - f_0}{B/2} + t_3 \frac{f - f_0}{(B/2)^2}$$

* *Marconi's Wireless Telegraph Co. Ltd.*

t_1 is the absolute value of the delay at the band centre and is of no significance in the calculation of distortion as it merely adds a time delay to the whole signal.

$t_2/(B/2)$ is the linear slope normally expressed in mμsec/Mc/s

$t_3/(B/2)^2$ is the quadratic curvature normally expressed in mμsec/Mc/s²

B is the bandwidth of the system.

It should be noted that although t_2 may be positive or negative, the curvature coefficient t_3 is almost always positive, i.e., there is a minimum of group delay near the band

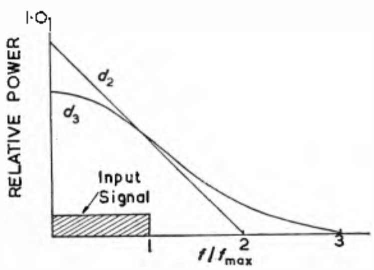


Fig. 3. Amplitude intermodulation spectra (2nd and 3rd order)

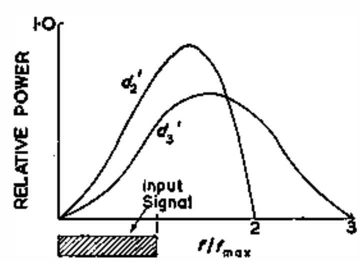


Fig. 4. Intermodulation spectra due to group-delay distortion (2nd and 3rd order)

centre. The pass-band B containing the region of interest depends on the capacity of the system, but has been estimated at 12Mc/s for a 600 channel system and 18Mc/s for 960 channels.

For the purpose of specification and testing of multi-channel telephony systems it is convenient to represent the baseband signal by a 'white noise' signal of suitable level and bandwidth. The method has been fully described elsewhere[2]. If the noise signal is applied to the input of a microwave system the output will consist of:

(a) The original white noise or uniform power-frequency distribution signal.

(b) Intermodulation noise due to non-linearity of modem transfer characteristic (amplitude distortion).

(c) Intermodulation noise due to non-uniformity of group delay in amplifiers and mismatched feeders.

(d) Thermal or basic noise introduced mainly in the early stages of repeaters and amplifiers.

The signal has a uniform power-frequency spectrum but the level of intermodulation and basic noise varies throughout the band. Fig. 3 shows the distribution of the amplitude distortion power plotted against f/f_{max} where f_{max} is the highest frequency of the baseband input signal (2·54Mc/s for 600 channels). Curves for the predominant 2nd and 3rd order distortion powers d_2 and d_3 in a channel centred on f are shown. It will be seen that intermodulation power is at a maximum at the lowest frequency of the baseband.

Fig. 4 shows similar distribution curves for noise due to group-delay distortion, d_2' being the second order distortion power in a channel arising from the slope coefficient t_2, and d_3' the third order distortion due to curvature coefficient t_3. These curves can be obtained by multiplying the ordinates of the amplitude distortion case by ω^2. Here it is noted that intermodulation power is greatest at the highest frequency in the baseband, in fact, the maxima are reached above the traffic band at about $1{\cdot}3f_{max}$ for d_2' and $1{\cdot}5f_{max}$ for d_3'.

In practice it is found that noise due to group-delay distortion and thermal noise usually exceeds that due to transfer non-linearity and thus the noisiest channel is the highest frequency one. It can be shown[3] that the second

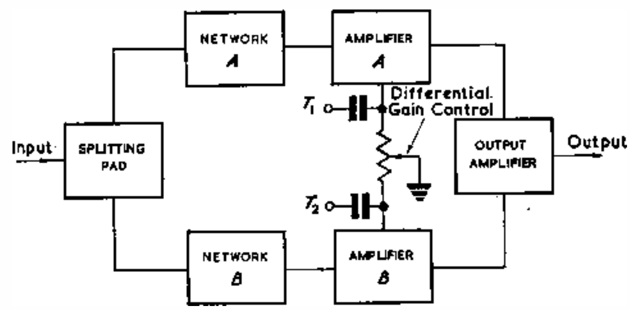


Fig. 5. General arrangement of the 70Mc/s variable group-delay equalizer

order intermodulation power d_2' in the highest frequency channel is

$$\frac{\omega^2 \Delta f^2 P}{8N}(t_2/(B/2))^2 \quad \ldots\ldots\ldots\ldots \quad (1)$$

where $\omega = 2\pi f_2$, Δf = peak deviation of r.f. carrier, P is the total baseband power of the fully loaded system, N the number of channels and $t_2/(B/2)$ the group delay slope.

Similarly the third order intermodulation power d_3' in the highest frequency channel is

$$\frac{\omega^2 \Delta f^4 P}{20N}[t_3/(B/2)^2]^2 \quad \ldots\ldots\ldots\ldots \quad (2)$$

where $t_3/(B/2)^2$ is the group delay curvature.

P as recommended by C.C.I.R. is $-15+10 \log N$, i.e. +13dBm at a point of zero level for 600 channels. The deviation for a channel test tone level of 0dBm is 200kc/s r.m.s.

It was found that the maximum noise power due to group delay distortion in a five hop section was 490pW and assuming $d_2' = d_3' = 245$pW, then equations (1) and (2) give a maximum permissible group delay slope of 0·40mμsec/Mc/s and curvature of 0·51mμsec/Mc/s². For 960 channels the figures are reduced to 0·20mμsec/Mc/s and 0·20mμsec/Mc/s². If it is desired to increase still further the channel capacity, tolerances become even more severe. Since $P = -15 + 10 \log N$, P/N is constant, $\Delta f \propto \sqrt{N}$ and $\omega \propto N$ (for N large). Hence from equations (1) and (2) for a fixed value of d_2' and d_3'

$$t_2/(B/2) \propto 1/N^{3/2} \quad \text{and} \quad t_3/(B/2)^2 \propto 1/N^2$$

Method of Equalization

From the previous discussion it can be seen that intermodulation power due to phase non-linearity in a system is at a maximum above the highest frequency of the baseband signal. Accordingly at every 280km in the reference circuit, this out of band signal is available as a measure of the distortion introduced by the preceding section. There is thus the possibility of utilizing the noise to control automatically the setting of an i.f. equalizer inserted immediately before the demodulator. A suitable continuously variable equalizer has been described by Hamer and Wilkinson[4].

As shown in Fig. 5 the 70Mc/s i.f. signal is divided equally by a splitting pad into two paths A and B, amplified and combined in an output amplifier. If network A is a positive slope fixed equalizer and B a similar negative slope equalizer, then providing certain precautions are observed, the overall group delay characteristic can be varied smoothly between the limits set by the networks by controlling the gains of the amplifiers differentially. This control is most simply achieved by varying the bias voltages to the amplifier valves by a potentiometer.

Fig. 6 shows the arrangement used to demonstrate an experimental automatic equalizer. The white noise signal simulating a fully loaded 600 channel system is applied through two band-stop filters to the modulator. The first filter inserts a loss of >80dB in a small band centred on the measuring channel of 2 466kc/s. The second filter inserts a loss of >50dB at 3·2Mc/s to remove any noise in the control channel which may remain due to the deficiencies of the low-pass filter in the white noise generator. The deviation at 70Mc/s is 875kc/s r.m.s. and no pre-emphasis is used.

The 70Mc/s carrier is applied to the demodulator through two similar variable slope equalizers. In the first or error-inserting equalizer the slope can be pre-set manually to any value between +1·0 and −1·0mμsec/Mc/s. In the second or control equalizer, a 50c/s search voltage is applied to terminals T_1T_2 (Fig. 5) in the differential gain control circuit; the group delay slope and hence the intermodulation noise due to the inserted delay distortion will vary at this rate. The noise power in a 50kc/s band centred on 3·2Mc/s is amplified, detected and the modulation envelope applied to one winding of a two phase servo-motor. The other winding of this motor is energized by a 50c/s voltage from the same source as the search voltage. The motor is mechanically coupled to the potentiometer which sets the mean slope of the equalizer. The 50c/s component of the error signal will rotate the motor, with correct phasing in a direction such as to reduce the intermodulation power and hence the equalizer is finally set in a position giving minimum intermodulation power.

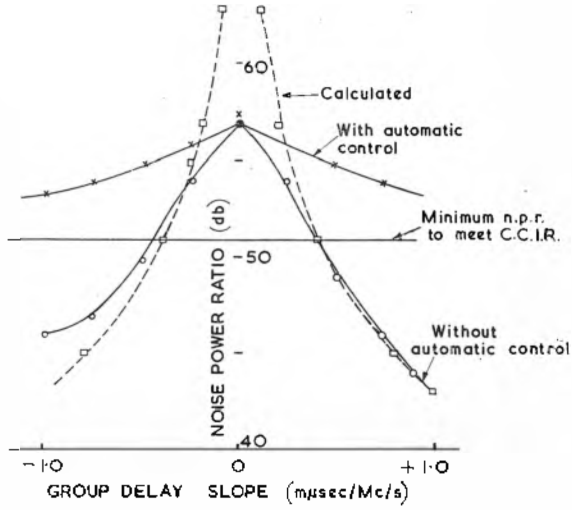
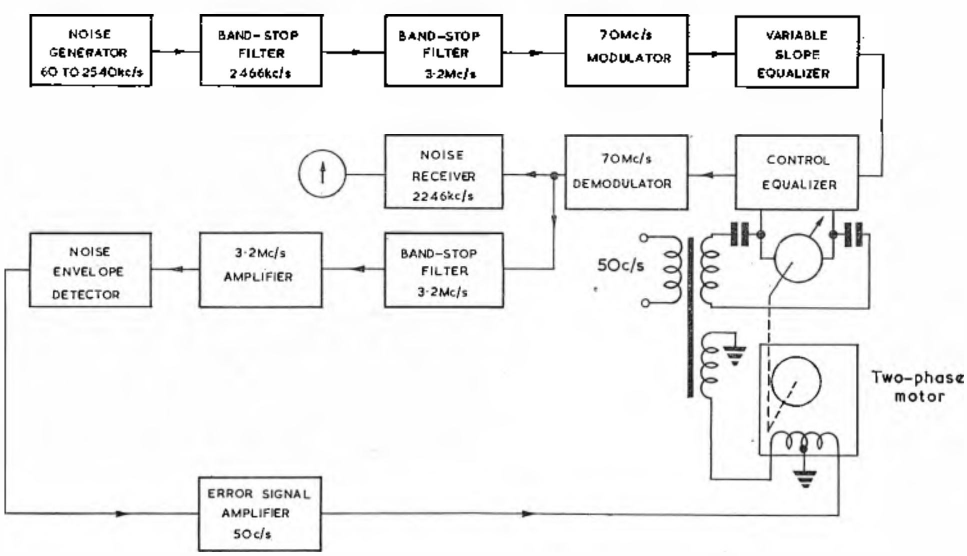


Fig. 7. Intermodulation power with and without the automatic equalizer

The advantage of using a motor control as against electronic is that should the 50c/s voltage be removed at any time the equalizer is left at the optimum setting. The 50c/s drive to the equalizer is adjusted until the intermodulation noise in the 2 466kc/s channel is increased just noticeably at zero-slope and thus there is a negligible degradation in performance.

Fig. 7 shows the intermodulation measured at 2 466kc/s with and without correction as the slope introduced by the first equalizer is varied between +1·0 and −1·0mμsec/Mc/s. The results are given as a noise power ratio (n.p.r.), i.e., the ratio of the noise powers in the measuring channel when the 2 466kc/s band stop filter is switched from the 'out' to 'in' position. The n.p.r. is the most useful figure in evaluating the performance of a multi-channel telephony system; the channel signal-to-noise ratio may be obtained by adding 15dB to the n.p.r. A considerable improvement is obtained with automatic correction and in particular if the equalizer is placed in the five hop section mentioned previously, the cumulative group delay slope can rise to at least 1·0mμsec/Mc/s before performance is degraded below that estimated to meet C.C.I.R.

Fig. 6. General arrangement of the experimental automatic group-delay equalizer



Reasonable agreement is shown with results calculated from equation (1) for large delay slopes. For small values, other sources of distortion, such as delay curvature, transfer non-linearity and thermal noise, are predominant.

Slope and Curvature Correction

System performance in general is most sensitive to errors in group delay slope, this being the first order effect of slight detuning of amplifiers and filters. To maintain highest standards, however, it is desirable to provide curvature correction as well.

The scheme[5] is outlined in Fig. 8. The general principle is the same as slope correction, but in this case it is necessary to generate two independent control signals.

The torque generated in a two phase motor is proportional to the product of the reference signal and the control signal. Thus, if the reference signal is $A \sin \omega t$ and the control signal $B \sin (\omega t + \phi)$, the torque T is proportional to

$$\begin{aligned} T &= AB \sin \omega t \cdot \sin (\omega t + \phi) \\ &= (AB/2) [\cos \phi + \cos (2\omega t + \phi)] \end{aligned}$$

and the non-pulsating component of the torque is $T_{dc} = (AB/2) \cos \phi$. The reference signal to servo motor (1) is $A \sin (\omega t + \pi/4)$ while the control signal to it contains two components in phase quadrature $B_1 \sin (\omega t + \pi/4)$ and $B_2 \sin (\omega t - \pi/4)$. Therefore the steady torque has two components:

$(AB_1/2) \cos((\pi/4) - (\pi/4))$ and $(AB_2/2) \cos((\pi/4) + (\pi/4)) = 0$

Thus each servo motor will respond to one control signal only, and the delay slope and curvature correction will take place independently. The curvature equalizer is similar to the slope, except that in Fig. 5 network A is replaced by a fixed equalizer of approximately parabolic shape with maximum group delay at band centre, and network B is a cable of uniform group delay. Variation of the differential gain control thus gives an equalizer in which the curvature can be set from zero up to a maximum determined by network A.

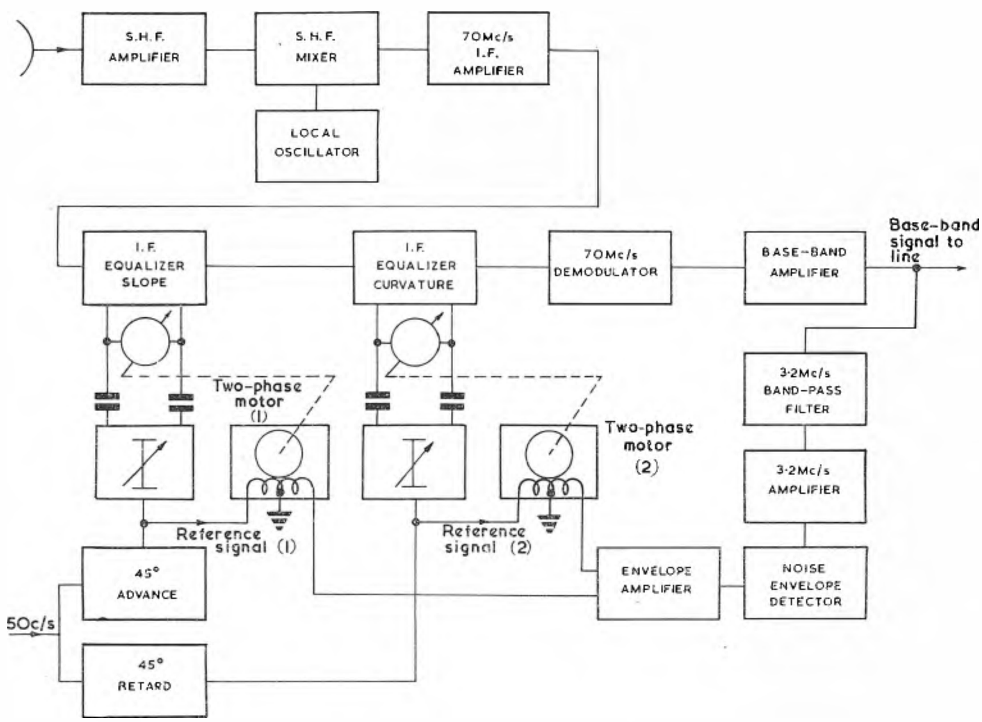


Fig. 8. Block schematic of proposed slope and curvature correction equalizer

Further Development

The preceding treatment has dealt only with the correction of relatively simple group delay characteristics. In particular the delay of a link with aerial feeder echoes will show rapid changes across the band. The form of the characteristic due to a single strong echo is that of a sinusoidal ripple, the frequency spacing between successive maxima being $\frac{1}{2}T$ where T is the feeder delay (one-way). For long feeders of say, 100ft, echo delays of 200mμsec are typical and thus the spacing is only 5Mc/s. If the system is to carry 960 channels, there will be more than three complete cycles across the working band. It is clear that a parabolic curvature equalizer will be quite unable to deal with this shape. Correction of higher terms of the delay function would be possible with a transversal equalizer[6] which can provide as many terms as needed in theory, but will be very complicated in practice. However, the introduction of ferrite isolators has mitigated at least in part the problem of echoes in long feeders.

Conclusions

The design and operation of a simple automatic group delay equalizer has been discussed and results show that such a device can give a worthwhile improvement in f.m. microwave systems. As systems of larger capacity carrying up to 1 800 channels and colour television signals come into use, the principle of automatic correction will undoubtedly become more attractive in order to maintain the extremely small tolerances in phase and transfer linearity required.

Acknowledgements

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An Equipment for Measurement of Tool Cutting Forces

By J. Goldberg*

This article describes equipment developed for the sequential measurement of components of cutting force on lathe tools during machining operations. The force sensing elements are wire resistance strain gauges used in conjunction with a two-component force dynamometer. The measuring circuits are formed from transformers with inductively coupled windings of 1:1 ratio supplying sets of gauges suitably connected. Each component of force is measured by a separate bridge circuit, the input and output circuit of each bridge being switched in the required sequence between a single a.c. driving source and a common detector. By this means the effect of switch contact resistance on the accuracy of the measuring loop is rendered insignificant.

Phase sensitive detection is employed for observing the change in the resistive component of unbalance of the bridges, as well as the sign of the strain.

The display system is a pen recorder which incorporates a device for raising and lowering the pen in order to share the recording time between the number of channels in use.

Full-scale readings corresponding to forces of 200lb, 500lb, and 1 000lb, are provided, and the equipment can be readily calibrated by static loading. Records of cutting force obtained with the equipment are shown.

(Voir page 774 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 779)

In the investigation of metal cutting processes, a knowledge of the force acting on the cutting tool during the actual machining operation is of basic importance. The magnitude and direction of the force when related to such factors as tool geometry, cutting speeds, workpiece materials, etc., give an insight into the mechanism of the machining operation.

A dynamometer is used to sense and measure the magnitude and direction of the force. It usually consists of a framework to which the cutting tool is fixed and which is capable of very small deflexions in appropriate directions under the influence of a cutting load.

Ideally, it is desired to measure the values of the various components of cutting force at the same instant of time. However, it is known that variations in cutting force during a cut are so slow that the forces can be regarded as quasi-static. The present equipment uses a single channel to sample the forces at the rate of three per second.

The Transducer

A typical dynamometer[1], as used in a lathe turning operation to measure the tangential and axial components of cutting force is shown in Fig. 1. It is in the form of an octagonal ring to which the tool is fastened and the complete arrangement is then clamped to the saddle of a lathe.

A theoretical investigation of the ring shows that when tangential load is applied there are four points on the ring which are not strained, namely the positions marked A; similarly, when axial load is applied there are four strain nodes marked T (see Fig. 1). Resistance strain gauges are fastened to these nodal positions and it is found that if axial and tangential loads are applied simultaneously to the tool they can be measured with only slight cross-interference (caused mechanically), by connecting the relevant gauges into two independent bridges.

The Measuring System

A schematic diagram of the measuring system as used with the two-component dynamometer is shown in Fig. 2. The two independent bridge circuits are sequentially switched between a single oscillator and a common detector, the forces and zeros of the apparatus being displayed by the pen recorder.

The forces to be measured can reach 1 000lb in both directions, a typical value being 200lb. Preliminary investigations showed that the strain sensitivity of the dynamometer is about the same in both directions and the measured force-strain relationship is approximately

$$F = 10^7\epsilon,$$

where ϵ is the strain and F is the force in pounds.

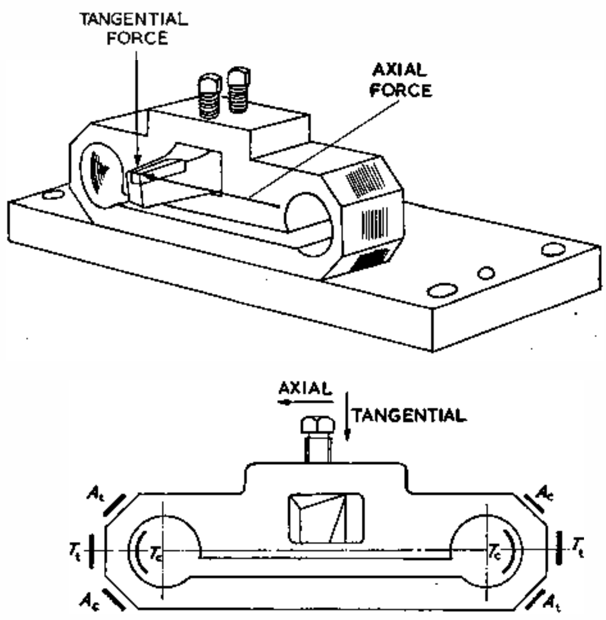

Fig. 1. Force measuring transducer

The recorder and dynamometer together impose a limit of resolution of about 5 per cent of full scale, 200lb, on the most sensitive range. This means that a force of approximately 10lb has to be read corresponding to a strain of 1μin/in.

With a gauge factor $g = (\delta R/R)/\epsilon = 2{\cdot}60$, a change of resistance of $\delta R = 3 \times 10^{-4}\Omega$ in a 120Ω strain gauge, has to be resolved.

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It has been found that contact resistances in selector switches of the telephone type having nickel-silver contacts are much larger than $3 \times 10^{-4}\Omega$ and are very variable. Switching the complete bridge system in the manner described renders contact resistance variations insignificant.

The Transformer Bridge Circuit

The gauges corresponding to each force component are connected in series pairs, one pair in compression and the other in tension, and are arranged with parallel balancing resistances across the secondary of a transformer with inductively coupled windings of 1 : 1 ratio, as shown in Fig. 3. These windings provide bridge ratio arms of high stability and, because of their low impedance and the three terminal connexion, minimize the effects of capacitances to earth.

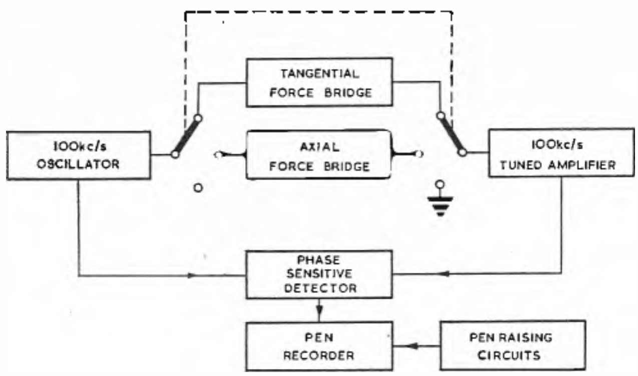


Fig. 2. Force measuring system

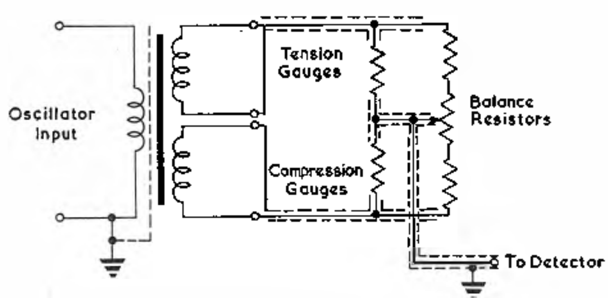


Fig. 3. Strain gauge bridge

The resistive balance controls have about 70kΩ in each arm and, while they provide sufficient adjustment for balancing the bridge, their effect on the gauge factor is negligible. This arrangement of strain gauges will yield the maximum sensitivity in the bridge and will provide some degree of temperature compensation when the gauges, surrounding bond and metal, have come to temperature equilibrium.

The transformer primary is energized at 100kc/s to provide a suitable carrier for vibration and chatter modulation frequencies which lie in the audio band. These, if desired, can be viewed on an oscilloscope. However, when the equipment is used with a pen recorder, the higher frequencies are not displayed.

Effects of Transformer Imperfections

The transformer bridge circuit may be represented by an equivalent form (Fig. 4) in which e_1 and e_2 are voltages induced in the secondary windings, L_1 and L_2 are leakage inductances of each half winding, r_1 and r_2 the combined effects of transformer winding and connecting cable resistances, C_1 and C_2 the distributed capacitances to earth of the screened connecting cable, and R_1 and R_2 are total strain gauge resistances.

The voltages e_1 and e_2 will differ from each other by a small amount since leakage flux from the primary may not link both halves of the secondary winding symmetrically. In addition, the leakage impedances on both sides will differ slightly. Thus, any reactive or resistive loading of the ratio windings will cause the two voltages applied to the strain gauges to differ in magnitude and phase by a small amount.

The errors caused by these effects are constant for a given transformer and associated leads to the transducer. The most serious effect is that a bridge unbalance signal is produced in quadrature with that caused by changes in gauge resistance and may overload the phase sensitive detector. Careful construction of the transformer to equalize the leakage impedances will reduce this effect to a safe value.

Bridge Details

The transformer is wound on a 3B3 Ferroxcube pot core in which the air gap has been almost eliminated by grinding and lapping.

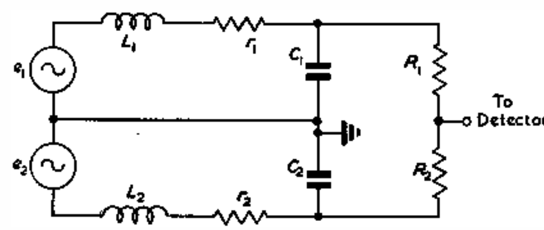


Fig. 4. Approximate equivalent circuit of bridge

Primary: 65 turns of 29 B.S. enamelled copper wire.
Secondary: 9 turns of 27/40 enamelled silk-covered litz wire.

The secondary is bifilar wound on top of the primary, the latter winding being electrostatically shielded from the secondary. The strain gauges are of the bonded type having a d.c. resistance of 120Ω and a gauge factor of 2·6. The bridge excitation voltage is approximately 0·65V per side at the secondary on the lowest range and thus the r.m.s. current in the gauges is less than 3mA in the arrangement adopted. This minimizes drift due to self-heating and gives satisfactory long-term stability.

Oscillator

The oscillator (Fig. 5) supplies an adjustable voltage for the bridges and a two-phase voltage for the phase sensitive detector.

A Wien bridge, shown in the dotted square, is driven by two separate amplifiers V_1, the output being applied via buffer cathode-followers V_2 across one diagonal of the Wien bridge. The bridge unbalance signal which appears across the opposite diagonal is applied to the amplifier input terminals.

At a frequency $f_r \simeq 1/2\pi RC$, where $R_2 = 2R_3 = R$, $C_2 = C_1/2 = C$, this bridge unbalance signal is real and in the same phase as the driving voltage. As the amplifier gain is high, the amount of unbalance required in the bridge to maintain oscillation is small. The thermistor element adjusts itself so that the above conditions are realized, e.g., if unbalance voltage in excess of that required to just maintain the system in oscillation is applied to the amplifier input terminals, increased voltage will appear at the amplifier output and be applied to the bridge. Increased dissipation in the thermistor will lower its resistance and will tend to restore the bridge unbalance to its former value. The oscillator output amplitude is largely independent of amplifier characteristics and depends mainly on the Wien bridge.

The values of resistance and capacitance in the frequency dependent arms of the bridge are chosen to make the impedances of these arms equal at 100kc/s.

Variations in oscillator output are mainly caused by the effect of ambient temperature on the thermistor. They are made small by enclosing the bridge circuit in a polished metal can and thermally insulating it from the chassis material. There is provision for metering the oscillator voltage.

The circuit consists of two pentodes V_4, V_5 in a symmetrical arrangement. The control grids of both valves receive the in-phase and quadrature components of the bridge unbalance signal. The suppressor grids are excited from the oscillator, the phase of the voltage at one suppressor being 180° out of phase with that at the other. The sine waves at each suppressor grid are d.c. restored below earth potential by the diodes. The gain of the detector under these conditions is proportional to $E_2 \cos \phi$, where E_2 is the voltage applied to either suppressor grid and ϕ is the phase-angle between E_2 and the input voltage to the control grids. Thus any voltage in quadrature with E_2 is rejected in the output.

The phase detector gain is approximately 100 and, at its output terminals, provides a d.c. voltage swing of ± 60V for full scale deflexion of the recorder, the output being applied to the recorder via two cathode-followers V_6 and V_7 which drive the low impedance pen movement and produce damping. An output meter is also provided.

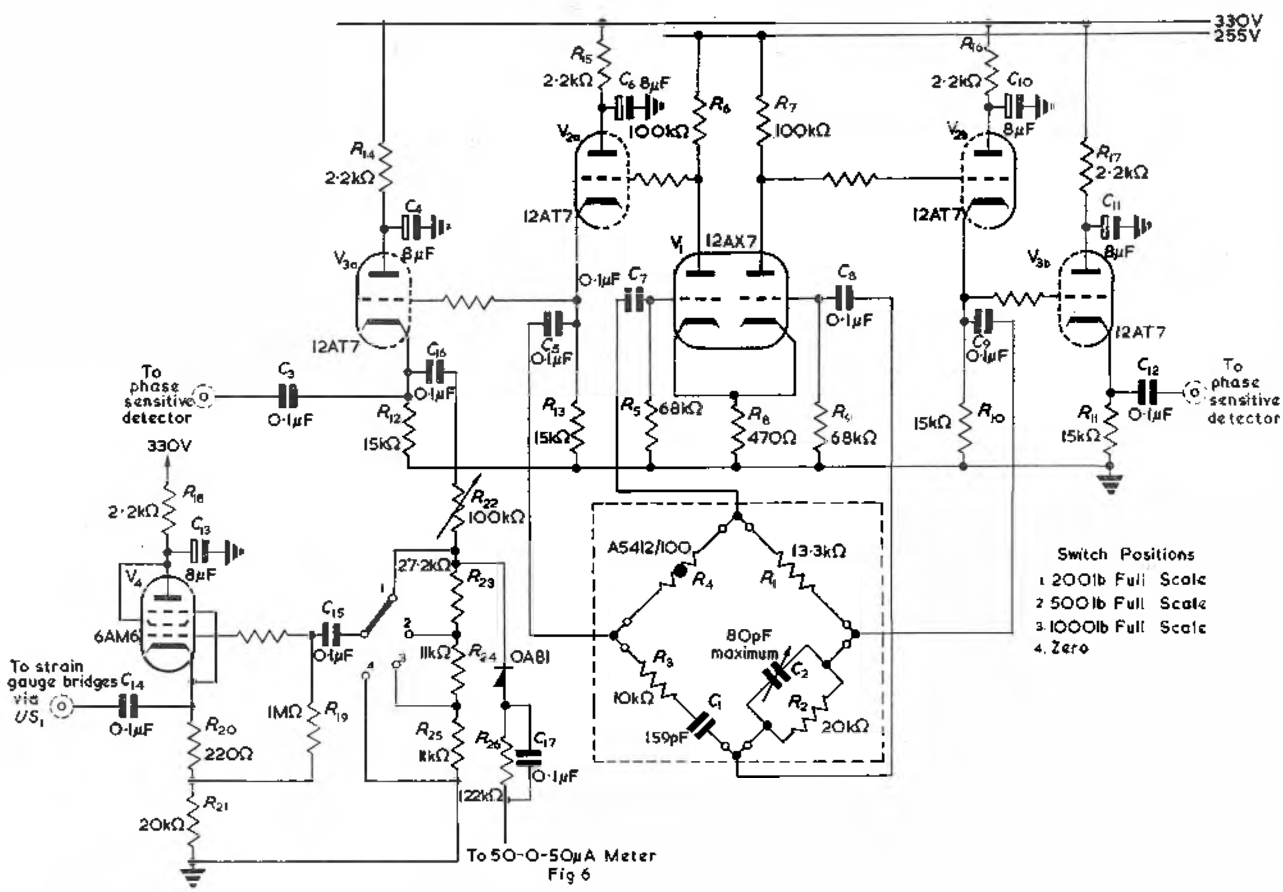


Fig. 5. 100kc/s oscillator

Amplifier and Phase Sensitive Detector

The amplifier (Fig. 6) is conventional, consisting of two stages V_1 and V_2, which employ single tuned circuits and have the input matched to the source impedance of the strain gauge bridges, which is approximately 100Ω. The overall gain is about 12 000 and the gain of each valve is to some extent stabilized by having the cathode resistors high in value and unby-passed.

The bandwidth is approximately 2kc/s and is quite adequate to pass the modulation frequencies of the carrier and allow for drift in the oscillator frequency. Output is taken via a cathode-follower V_3 and passed to a phase sensitive detector. This circuit, which is also shown in Fig. 6, has to sense the sign of the strains and to provide an adequate voltage swing to operate the pen recorder. In its output it must also reject the quadrature component of the bridge unbalance signal.

System Stability

The stability of the measuring system without the dynamometer was checked by disconnecting the strain gauges and deriving a small signal through an attenuator from the secondary of the ratio transformer and applying this to the amplifier input. The system output variation with time was recorded and was found to be ±0·2 per cent of full scale over a period of three hours and was thus insignificant.

Next, the stability of the complete system including the strain gauges was checked and was found to be ±0·4 per cent of full scale. This order of stability was obtained with strain gauges applied to the transducer with thermo-setting adhesive followed by several cycles of oven drying. To maintain this order of stability during a cutting test it was found necessary to remove the heat from the tool by circulating water through the tool shank (Fig. 11).

and the method used to display the force measurements will now be described.

The bridge unbalance signals are the result of quasi-static cutting forces. This allows a relatively slow sampling technique to be used.

The display system uses a pen recorder with the addition

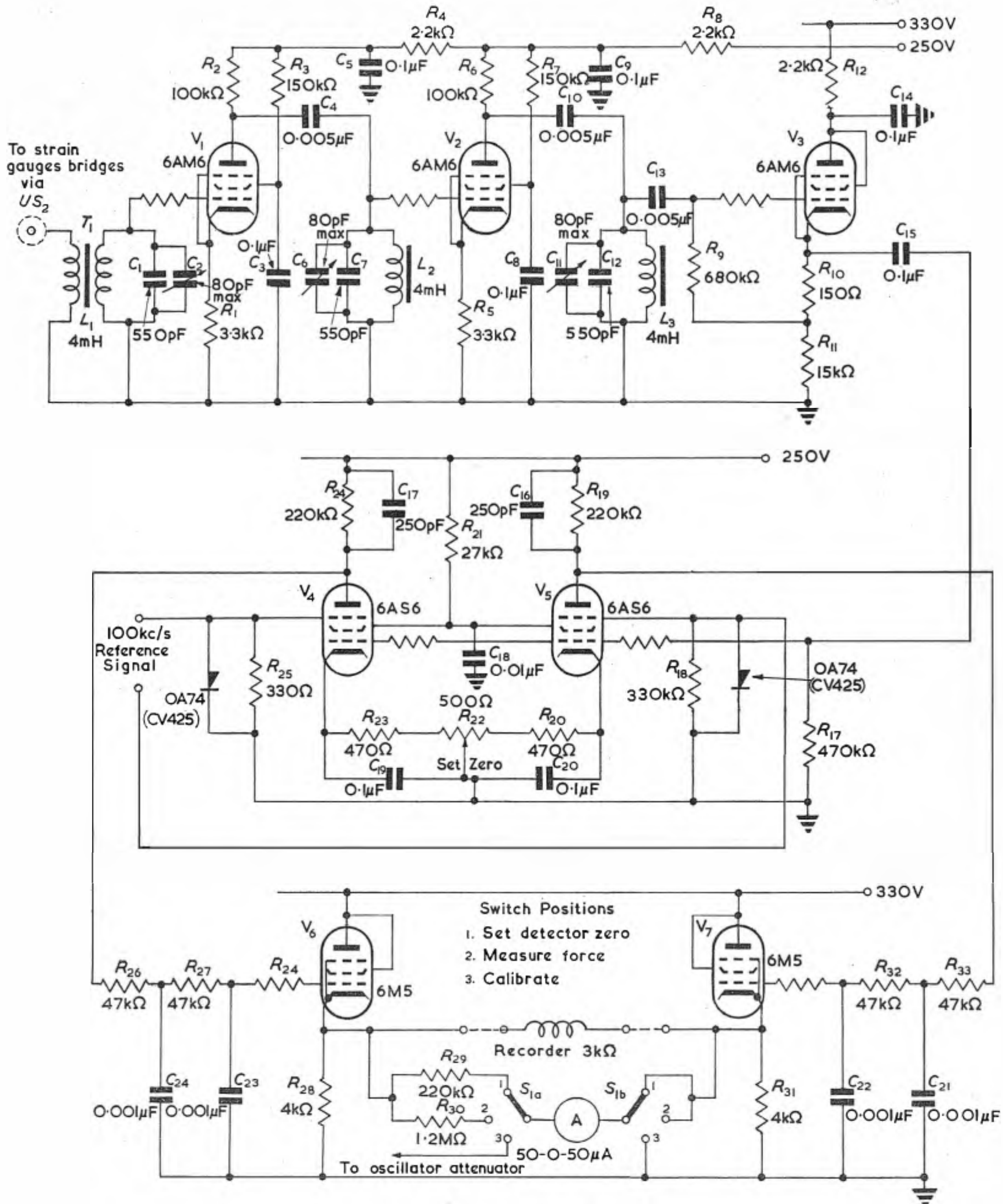


Fig. 6. 100kc/s amplifier and phase sensitive detector

Calibration

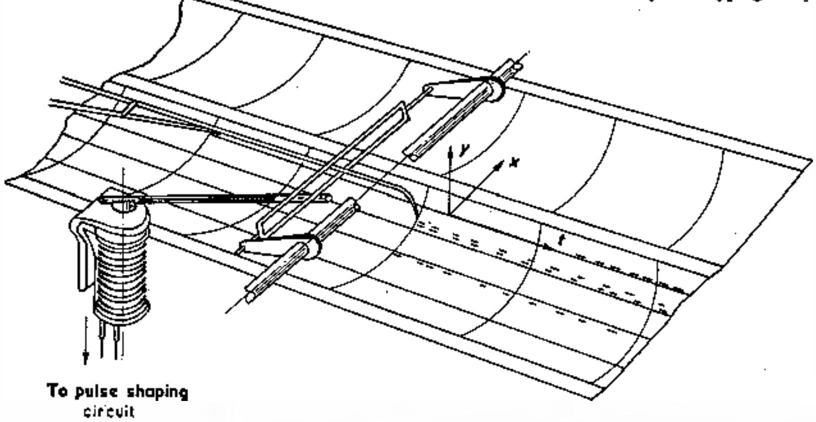
The equipment is calibrated by applying known static loads directly to the tool shank by a moment method. The calibration is thus made independent of the value of the gauge factor.

With a 200lb static load applied to the tool the full scale reading of the recorder is set by adjusting resistor R_{22}, shown in Fig. 5. The remaining ranges, 0 to 500lb and 0 to 1 000lb, are obtained by further attenuation.

Switching and Display System

As indicated in Fig. 2, each bridge circuit is sequentially switched between a single oscillator and a common detector of a relay to raise and lower the pen (Fig. 7). The chart moves in the t-direction and impulses in the relay coil move the pen vertically in the y-direction while the measuring system is being switched to the appropriate channel. This switching is performed by two synchronized uniselectors, and the vertical pen movement by a standard 3 000 type relay with armature slightly modified to carry a lever. This operates a chopper bar constructed to allow free movement of the pen in the x-direction as it is being raised. The chart record thus consists of a sequential series of small ink marks, as indicated in the figure. For the present purpose it was considered sufficient to switch channels at the rate of three per second.

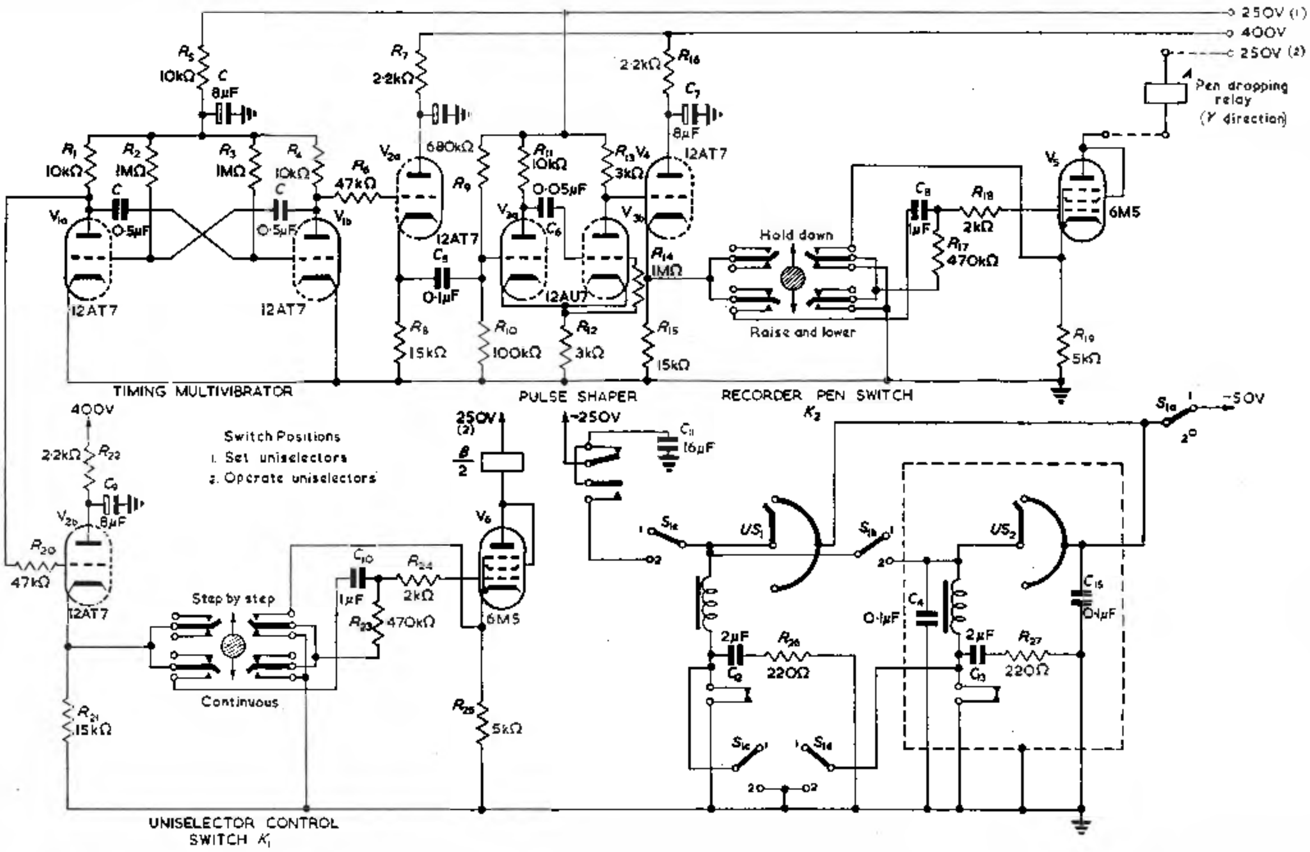
Fig. 7. Arrangement of recorder and pen dropping relay



Circuit Description

The circuit is shown in Fig. 8(a), and idealized waveforms in Fig. 8(b).

An astable multivibrator V_1 generates timing waveforms (a) and (b) at each anode at the channel switching frequency (3c/s). These are of equal mark-to-space ratio. The waveform (a) is passed via cathode-follower V_{2b} and switch K_1 to V_6. In the switch position marked 'continuous', the waveform turns the current in V_6 on and off, operating and releasing the relay B in the anode circuit. In the released position a capacitor C_{11} is charged to −250V and in the operate position this capacitor is discharged



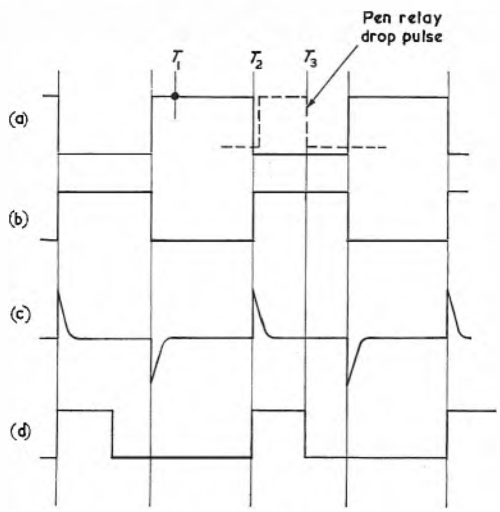


Fig. 8(a). Switching unit
(b). Idealized waveforms of switching unit

through the driving coils of uniselectors US_1, US_2, pulling in both armatures. At the cessation of the discharge the armatures release and the contacts are made for the next channel at time T_1 (Fig. 8(b)).

When the channel is connected by US_1 and US_2 the recorder pen moves and takes up a position ready to drop on the chart paper before switching to the next channel. It is desirable to delay the start of the pulse for operating the pen-dropping relay until the pen has overshot and returned approximately to its true record position. To perform this operation, the waveform (b) which is in opposite phase to (a) is differentiated as in (c) and the positive going pulses trigger a monostable multivibrator V_3, thus producing the positive going pulses for vertical movement of the pen relay.

For continuous operation the recorder pen switch K_2 is in the 'raise and lower' position, allowing the sequence of pulses to be passed to the valve V_5.

The delay between the making of channel contacts at time T_1, and the start of the relay drop pulse at time T_2 is about 100msec. The drop pulse width, $T_3 - T_2$ is chosen to give an optimum length of mark on the paper and is approximately 100msec. In conjunction with a chart speed of $\frac{1}{4}$in/sec a satisfactory record is obtainable.

Operation on a Single Channel

If a continuous record of a single channel is required, key switch K_1 is in the 'step by step' position and key switch K_2 in the 'hold down' position. For K_1 the normally earthed end of the resistor R_{23} is connected to V_6 cathode, thus changing the anode current from a small value to a value sufficient to operate the relay B. The continuous waveform is disconnected during this operation so that only one channel can be switched for each operation of K^1. The same method is used to hold the pen-dropping relay A down and to disconnect the continuous sequence of pulses from the valve V_5.

Arrangement of Selector Switches

The switching is carried out by two synchronized uniselectors US_1 and US_2, as shown in Figs. 8(a) and 9. Uniselector US_1 switches the 100kc/s carrier voltage to the primary side of the ratio transformers. The wiper arm is shown on contact number 1. In this position both transformers receive the carrier and 100kc/s current will flow through both sets of gauges. This is advisable in order to minimize warm up drift and make the equipment ready for instant use after an idle period.

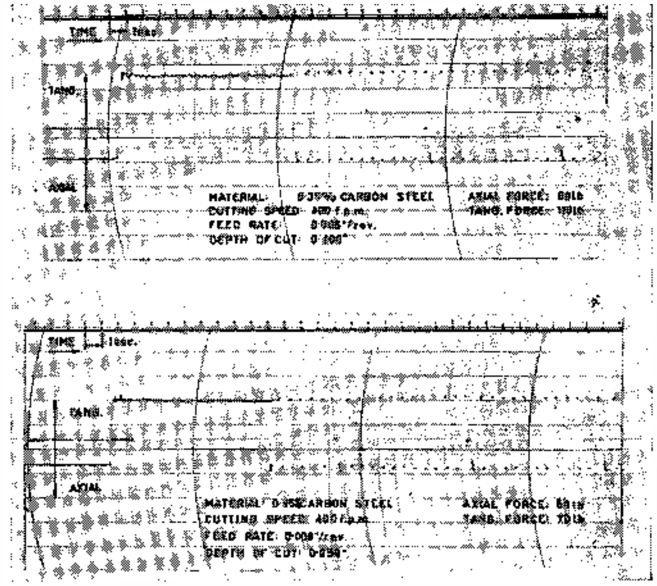


Fig. 10. Typical records of cutting force obtained with the equipment

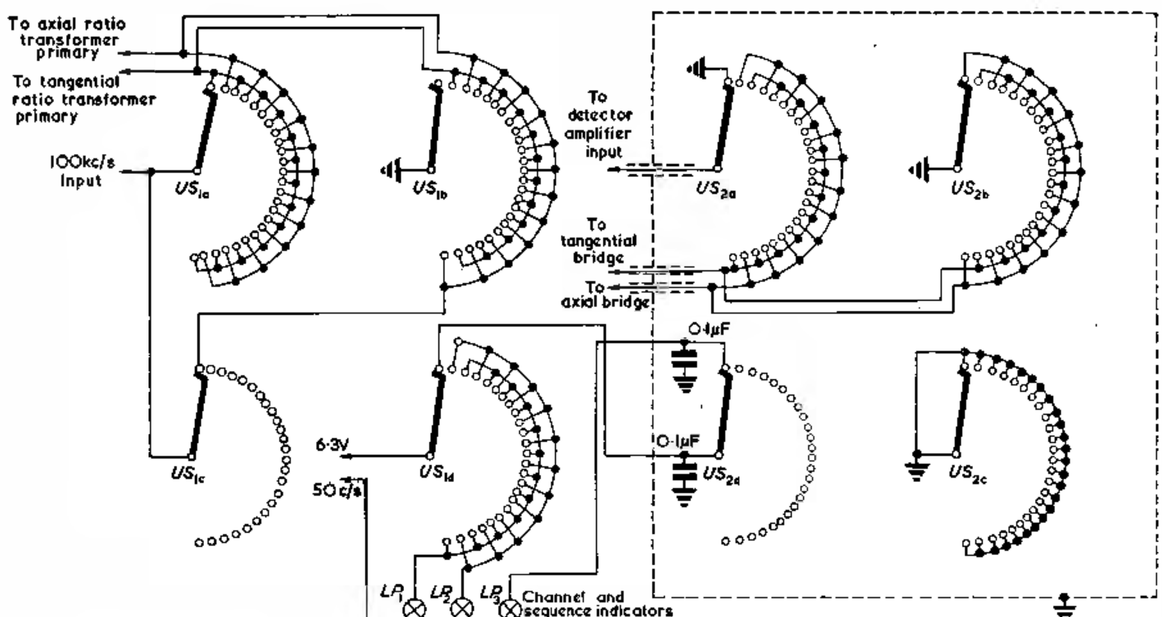


Fig. 9. Arrangement of selector switch contact banks

Uniselector US_2 is used to switch the detector arm of the bridges to the amplifier input. This switch is shielded from US_1 to prevent pick-up directly from the oscillator. On contact number 1 the amplifier input is earthed and a zero mark will therefore be registered on the chart. The remaining contacts (Fig. 9) switch the two bridges sequentially and provide visual indication, by pilot lamps, of the particular channel in use at a given instant or select a particular channel. It has been found advisable to short-circuit the primary of the ratio transformer not in use at a given instant to prevent pick-up from the adjacent channel.

To facilitate a rapid setting of both selectors to contact number 1 the homing arcs are used. In position 1, switch S_1 (Fig. 8(a)) applies -50V to both selectors through the interruptor contacts. In position 2 the selectors are ready for normal operation. This feature is important because the selectors tend to get out of step if operated with capacitor C_{11} not fully charged. The selectors may be brought into step rapidly by operation of S_1 and this state is shown by a pilot lamp LP_3.

Records of Cutting Forces

A typical pen record taken with the equipment during a lathe-turning operation on steel is shown in Fig. 10. Each channel was first recorded separately and then continuously sampled in sequence.

Power Supplies

The complete equipment uses an a.c. stabilizer on the mains input. The measuring circuits use a conventional rectifier and filter and a chain voltage regulator valves

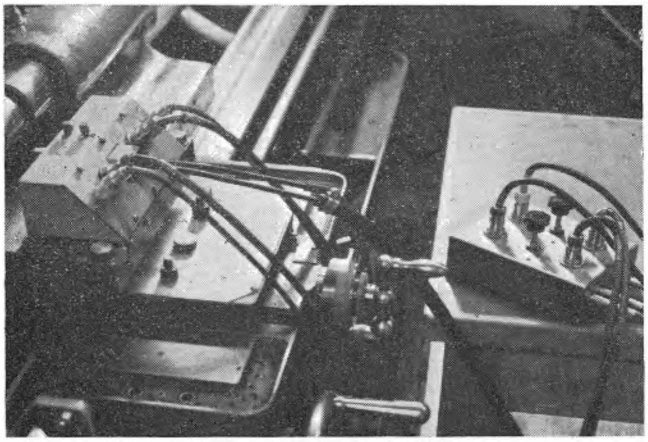


Fig. 11. Transducer and bridge balance controls

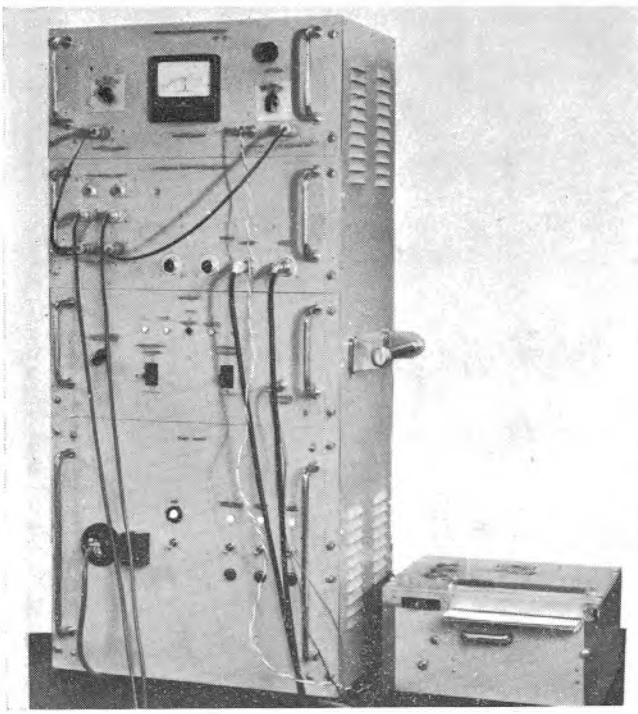


Fig. 12. Electronic equipment and pen recorder

to give 255V and 330V. The switching unit uses a separate supply with the 255V line regulated as above and an unregulated 400V line for the cathode-followers. Because of the pulsed nature of the load, the relay and uniselector circuit elements are worked from separate unregulated h.t. supplies, having the values shown in Fig. 8(a).

Constructional Details

Fig. 12 shows the electronic equipment and pen recorder. The units are arranged as follows commencing from the top.

(1) Oscillator and phase sensitive detector.

(2) Inductive ratio transformers, uniselectors and detector amplifier.

(3) Switching circuits.

(4) Power supplies.

There is provision for extension to four channel working.

All valves are mounted horizontally on chassis attached to the front panels. The top and bottom of the cabinet are covered with fretted steel, and with the bottom raised slightly above the bench the air circulation is adequate.

Fig. 11 shows the dynamometer attached to the lathe. The dynamometer housing is sealed with rubber to prevent ingress of moisture. The external balancing resistors are contained in a box a short distance away to keep them clear of the lathe saddle when it is in motion. The coaxial cables connecting these units are protected with heavy, braided copper sleeving to prevent damage by metal turnings.

Acknowledgments

The author would like to thank his colleague Mr. H. A. Ross, Senior Research Officer, Division of Metrology, for encouragement and helpful advice during the course of this project and Mr. A. M. Thompson, Principal Research Officer, Division of Electrotechnology, for helpful criticism. The equipment was developed to assist the research programme of the Applied Mechanics Section, Division of Metrology, C.S.I.R.O.

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A Car Radiotelephone System

A pilot car radiotelephone service in the South Lancashire area has been inaugurated by the Postmaster General. This service will enable users in the area whose vehicles are fitted with suitable radio equipment to make or receive calls from any telephone subscriber in the United Kingdom network.

The system comprises two base stations, one at Liverpool to serve that area and the other at Winter Hill, near Horwich, to serve the remainder of the South Lancashire area. The latter station is augmented by auxiliary radio receivers in Manchester to improve the service in the centre of the city.

F.M. is used with equipment which is suitable for channels having carrier-frequency separations of 50kc/s, and the same characteristics as have been agreed for the v.h.f. international maritime services have been adopted. These include a maximum deviation of $\pm$15kc/s, pre-emphasis and de-emphasis of 6dB/octave within the band 300 to 3 000c/s and frequency tolerances of $\pm$2 and $\pm$3kc/s for the base and mobile stations, respectively. The power of the base stations is about 35W (e.r.p.); for the mobile stations the power is restricted to 25W (e.r.p.) by licence but 15W is expected to be employed. Two-frequency channels are used in all cases; the base station operates on a duplex basis but the mobile subscriber has choice of simplex or duplex, the G.P.O. naturally preferring the latter.

The frequencies being used in the first instance are in the 159 to 164Mc/s band.

The four-wire radio circuits, i.e. the transmitting and receiving circuits are extended from the base stations over land lines to Peterloo Exchange, Manchester, where they are converted to a two-wire circuit via radio terminal equipment for connexion to the public telephone network. This radio terminal equipment comprises apparatus to ensure full modulation of the radio transmitter, irrespective of the speech volume received from the telephone subscriber within wide limits. It will also give protection against excessive re-radiation of noise or signals from the radio receiver.

In order to ensure a satisfactory signal to the telephone subscriber the radio receiver is set to operate on signals which have a signal-to-noise ratio of 25dB or better; it is not disconnected until the signal-to-noise ratio drops to 15dB. The auxiliary receivers at Manchester are set to operate at a somewhat higher signal-to-noise ratio and once operated they are connected to the telephone circuit in place of the Horwich base-station receivers.

The coverage is qualified as primary and secondary service areas. In the former, service should be generally good, but there may be some locations of no-service due to local screening. In the secondary area good service may be obtained on comparatively high ground and more generally in a stationary vehicle, perhaps at a selected site. The nominal service area is that which should obtain for 15W mobile station equipment in the mobile-to-base direction, in the reverse direction the range may be a little greater.

Wide-Band Analysis of Valve Phase-Splitting Circuits

By L. J. Giacoletto*

A phase-splitting circuit which is an electronic counterpart of a balanced output transformer has two outputs whose voltages are ideally equal in magnitude but 180° apart in phase. This ideal operation is realized at low frequencies. Interelectrode capacitances alter the output relations at higher frequencies, and it is this effect which is analysed herein. Accurate as well as approximate design relations are developed and compared with measured data for a typical circuit. It is found that near ideal operation to moderately high frequencies can be obtained by using suitable equal load impedances. Performance can be improved somewhat by using slightly unequal load impedances including a small peaking inductor in the anode impedance.

(Voir page 774 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 779)

It is frequently necessary to transform from an unbalanced voltage to a balanced voltage. The operation of a bridge measurement circuit is an example. Here, either the signal source or the detector must be balanced. A wide-band balanced detector has been previously analysed[1]. This article will analyse the wide-band operation of a phase-splitting circuit useful for obtaining a balanced signal from an unbalanced source. The circuit consists of a thermionic valve with load impedances in both the cathode and anode circuit as shown in Fig. 1. Since the circuit has two outputs, it is more difficult to analyse than the conventional single input-single output circuit; this is probably the reason why an analysis has not previously been published. Only small-signal (linear) operation will be considered. The effects of inter-electrode capacitances will be included.

The low-frequency version of the circuit of Fig. 1 with equal load resistors is well known and often used in audio amplifiers. Only a triode is considered practical for phase-splitting circuits as current division in other multi-electrode valves would make exact phase division difficult even at d.c.

Analysis

A variety of methods can be used for analysis. The technique employed here is to carry out equivalent circuit transformations as shown in Fig. 2. In Fig. 2(a) the circuit of Fig. 1 is shown with the conventional π equivalent circuit for the triode. By standard methods[2] the π equivalent circuit can be transformed to a T equivalent circuit as shown in Fig. 2(b). The results are:

$$\left.\begin{aligned} z_g &= \frac{g_{ac} + j\omega C_{ac}}{\Delta y} \\ z_c &= \frac{j\omega C_{ga}}{\Delta y} \\ z_a &= \frac{j\omega C_{gc}}{\Delta y} \\ z_m &= -(g_m/\Delta y) \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ (1)$$

where

$$\Delta y = -\omega^2(C_{ac}C_{gc} + C_{ac}C_{ga} + C_{ga}C_{gc}) + j\omega(g_mC_{ga} + g_{ac}C_{ga} + g_{ac}C_{gc}) \ \ldots\ldots\ (2)$$

Now, the cathode load impedance, Z_{cn}, can be combined with z_c to form a composite T equivalent circuit of the conventional type with a single input and single output. The various circuit operations can then be formulated[3]. In particular, the grid current and the anode-to-grid and cathode-to-grid voltage amplifications are:

$$I_g = \frac{z_a + z_c + Z_{cn} + Z_{an}}{D} V_{gn} \ \ldots\ldots\ldots\ (3)$$

$$V_{an}/V_{gn} = \frac{(z_m + z_c + Z_{cn})Z_{an}}{D} \ \ldots\ldots\ldots\ (4)$$

$$V_{cn}/V_{gn} = \frac{(-z_m + z_a + Z_{an})Z_{cn}}{D} \ \ldots\ldots\ldots\ (5)$$

where

$$D = z_g(z_a + z_c + Z_{cn} + Z_{an}) + (z_c + Z_{cn})(z_a - z_m + Z_{an}) \ \ldots\ (6)$$

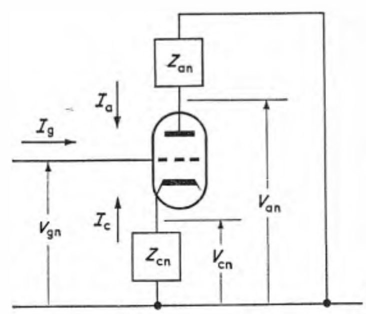

Fig. 1. Basic phase-splitting circuit

For typical triode parameters and at moderate frequencies, $\Delta y \simeq j\omega C_{ga}g_m$ with the result that equations (4) and (5) can be approximated as:

$$V_{an}/V_{gn} \simeq -\frac{[1 - j(1+g_mZ_{cn})(\omega C_{ga}/g_m)]\, g_mZ_{an}}{D'} \ \ldots\ (7)$$

$$V_{cn}/V_{gn} \simeq \frac{[1 + j((C_{gc}/C_{ga}) + g_mZ_{an})(\omega C_{ga}/g_m)]\, g_mZ_{cn}}{D'} \ (8)$$

where

$$D' = [(g_{ac}/g_m) + j(C_{ac}/C_{ga})(\omega C_{ga}/g_m)][1 + (C_{gc}/C_{ga}) + g_m(Z_{cn} + Z_{an})] + [1 + g_mZ_{cn}][1 + j((C_{gc}/C_{ga}) + g_mZ_{an})\,\omega C_{ga}/g_m] \ \ldots\ (9)$$

At d.c. the above equations can be further simplified to:

$$V_{an}/V_{gn} \simeq -\frac{g_mR_{an}}{D''} \ \ldots\ldots\ldots\ldots\ (10)$$

$$V_{cn}/V_{gn} \simeq \frac{g_mR_{cn}}{D''} \ \ldots\ldots\ldots\ldots\ (11)$$

where

$$D'' = g_{ac}/g_m\,(1 + (C_{gc}/C_{ga}) + g_mR_{cn} + g_mR_{an}) + (1 + g_mR_{cn}) \ \ldots\ldots\ldots\ (12)$$

These last equations are the usual results applied to the phase-splitting circuit of Fig. 1. It is seen that for d.c. exact phase splitting will be obtained if $R_{an} = R_{cn}$. At higher fre-

* *Ford Motor Company, U.S.A.*

quencies exact phase splitting will in general no longer be obtained.

For broad-band phase splitting, the question arises: what load impedances, Z_{an} and Z_{cn}, will give the best broad-band operation? Clearly from equations (4) and (5), exact phase splitting will be obtained if:

$$z_m - z_a - Z_{an} = (z_m + z_c + Z_{cn})\, Z_{an}/Z_{cn} \quad \ldots\ldots \ (13)$$

One method of satisfying this equation would be to make $Z_{an} = Z_{cn}$, $|Z_{an}| \ll z_m$, $|z_a| \ll |z_m|$, and $|z_c| \ll |z_m|$. These conditions could be generally satisfied by choosing $R_{an} = R_{cn} \ll 1/g_m$ provided a large voltage amplification were not important. Suppose, as a compromise, $R_{an} = R_{cn} = 1/g_m$ is chosen. Then the low-frequency voltage amplifications would be about $\frac{1}{2}$ (provided $g_{ac} \ll g_m$). The ratio of the anode to cathode voltage, $V_{an}/V_{cn} \simeq 1\ \angle(2\pi - \phi)$ (provided $C_{gc} - C_{ga}$) where ϕ (radians) $\simeq 4\omega C_{ga}/g_m$ [provided $\phi \ll 1$]. Thus, the first effect of higher frequency operation is to cause the phase-angle between V_{an} and V_{cn} to differ from 180°. The magnitudes of V_{an} and V_{cn} are still approximately equal although not the same as their low-frequency values. The source impedance at the cathode is much smaller than at the anode, and for this reason small unbalances between cathode and anode load impedances will produce much larger unbalances in output voltages.

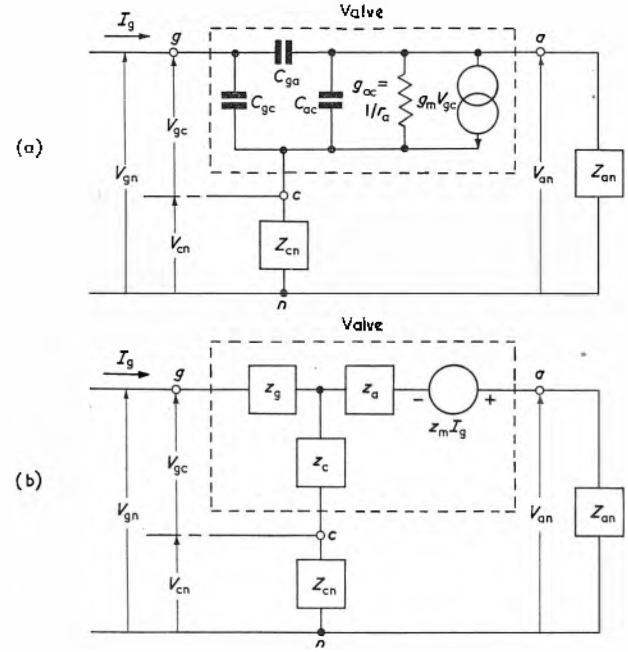

Fig. 2. Equivalent circuits

(a) Conventional π equivalent circuit

(b) T equivalent circuit

If large voltage amplification is important, $R_{an} = R_{cn} \gg 1/g_m$ is required. For this choice of Z_{an} and Z_{cn}, departures from exact phase splitting might be significant at relatively low frequencies. Some compensation for better frequency characteristics can be obtained by suitable design of Z_{an} and Z_{cn}. Thus, using the approximation, $\Delta y \simeq j\omega C_{ga} g_m$, equation (13) becomes:

$$\frac{-(C_{gc}/g_m C_{ga}) + j(1/\omega C_{ga})}{(1/g_m) + j(1/\omega C_{ga})} = Z_{an}/Z_{cn} + \frac{2Z_{an}}{(1/g_m) + j(1/\omega C_{ga})} \quad \ldots\ldots\ldots\ldots \ (14)$$

If, now the arbitrary choice is made that $1/Z_{cn} = G_{cn} + j\omega C_{cn}$ then:

$$Z_{an} = \frac{1/G_{cn}\,(1 + j(\omega C_{gc}/g_m))}{1 + \dfrac{\omega^2 C_{cn} C_{ga}}{G_{cn} g_m} - j(\omega C_{ga}/G_{cn})(2 + (G_{cn}/g_m) - (C_{cn}/C_{ga}))} \quad \ldots\ldots\ldots\ldots \ (15)$$

If the ω^2 term in the denominator is discarded as being negligible compared to unity, then Z_{an} can be synthesized by a series R_a and L_a and the combination shunted by C_{an} where:

$$R_a = (1/G_{cn})$$

$$L_a = (C_{gc}/G_{cn} g_m)$$

$$C_{an} = C_{ga}\,((C_{cn}/C_{ga}) - (G_{cn}/g_m) - 2) \quad \ldots\ldots \ (16)$$

The peaking inductor, L_a, will usually be very small; C_{cn} will have to be somewhat greater than $2C_{ga}$ in order for C_{an} to be positive. C_{an} and C_{cn} will be approximately equal if $C_{cn} \gg C_{ga}$.

Calculations

Calculations of the wide-band performance of the phase splitting circuit are straightforward but lengthy. The results of calculations for a circuit using half of a 12AU7 tube with equal cathode and anode load resistors of 5kΩ are shown in Fig. 3.

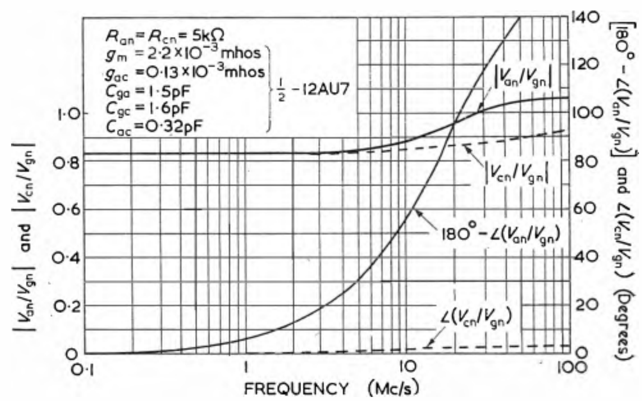

Fig. 3. Calculated performance with equal resistance loads

The valve is assumed to be operating with a floating heater supply and under standard conditions: Anode voltage = 250V; Anode current = 10·5mA; Grid voltage = −8·5V; and the valve parameters are as shown in Fig. 3. The low-frequency voltage amplification is 0·827; using equations (10) to (12) a voltage amplification of 0·820 is computed. The phase-angle of the anode-to-neutral voltage varies rather rapidly with frequency as compared to the cathode-to-neutral voltage. This result can be qualitatively explained by the fact that the quadrature current introduced through R_{an} by C_{ga} is about twice 'normal' value because V_{ga} and V_{an} are out-of-phase; in contrast the quadrature current introduced through R_{cn} by C_{gc} is quite small since V_{gc} and V_{cn} are in-phase. At higher frequencies both anode and cathode tend to be connected directly to the grid as the impedances of C_{ga} and C_{gc} become small compared to R_{an} and R_{cn}. This explains why although $|V_{an}/V_{gn}|$ does not vary much its phase-angle varies rapidly and approaches zero at the highest frequencies. Since $|V_{cn}/V_{gn}|$ is at the same time approaching unity, the normal $g_m V_{gc}$ generator of the valve becomes inoperative.

Figs. 4 and 5 show similar results for calculations carried out when R_{an} and R_{cn} are both shunted by 3pF capacitors and 15pF capacitors. With 3pF shunting capacitors $|V_{cn}/V_{gn}|$ increases but slightly at higher frequencies and then decreases. The phase-angle of $|V_{cn}/V_{gn}|$ is now negative in contrast with a slightly positive value in Fig. 3. With 15pF shunting capacitors (Fig. 5), the amplitude variations, $|V_{an}/V_{gn}|$ and $|V_{cn}/V_{gn}|$, are more nearly equal although differences in amplitudes occur at lower frequencies than in Fig. 4 and 3. The 3pF shunting capacitors (Fig. 4) represents about the minimum shunting effect that can be achieved when operating into a similar cathode-follower stage with the same value—1·5pF for grid-to-anode capacitance, $1{\cdot}6\,(1 - 0{\cdot}83) = 0{\cdot}3$pF for effective grid-to-cathode capacitance, and 1·2pF for stray capacitance.

The circuit shown in Fig. 6 was used to obtain operating data. The summing measurement circuit was used to give $A = |(V_{an}+V_{cn})/V_{gn}|$ at its output. Measured results are shown in Fig. 7 for two load conditions of the phase-splitting circuit. For the uncompensated load circuit ($L_a = 0$, $C_{an} = C_{cn} \simeq 3$pF) the anode and cathode shunting capacitors were removed and the 3pF shunting capacitance arose from the summing measurement circuit and stray capacitance. The compensated load circuit ($L_a = 3{\cdot}6\mu$H,

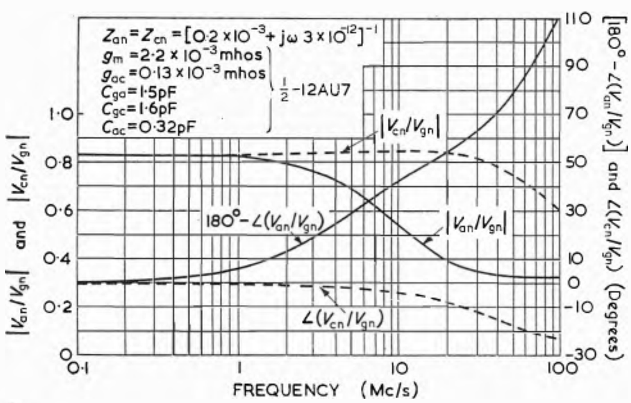

Fig. 4. Calculated performance with equal resistance loads and small shunting capacitance

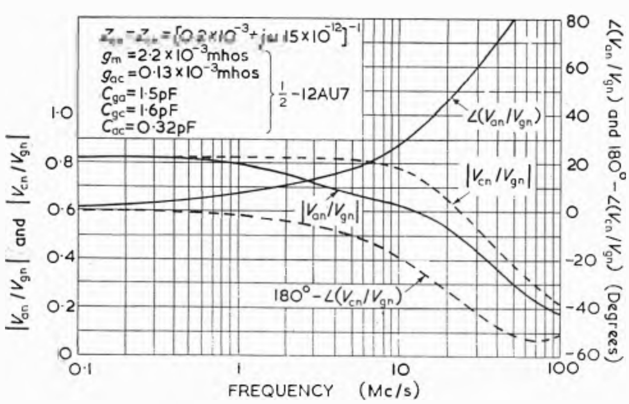

Fig. 5. Calculated performance with equal resistance loads and moderate shunting capacitance

C_{an} = 10·5pF, and C_{cn} = 13·7pF) was designed in accordance with equation (16) and achieved by 'padding out' the anode and cathode loads. The measured data indicates that A for the compensated circuit was about half that for the uncompensated circuit. However, both sets of measurements were somewhat less than calculated values of A also shown in Fig. 7. Possible explanations for the discrepancy between calculation and measurement are:

(1) Accurate measurements are difficult to carry out particularly at higher frequencies. The summing circuit was carefully balanced for zero output at lower frequencies where an accurately balanced input signal was available. This balance was then assumed to hold at higher frequencies. The summing circuit operates ideally only when exactly balanced. Any unbalance introduces degeneration and therefore likely accounts for the smaller measured amplitudes.

(2) Calculations were not carried out for the compensated circuit.

(3) At higher frequencies the equivalent circuit should be modified to include lead inductance and transit-time effects.

No measurements were made of the linearity of the phase-splitting circuit. Linearity should be reasonably good since the circuit is moderately degenerative.

Conclusions

If large voltage amplification is not important a simple phase-splitting circuit can be made effective over a wide range of frequencies by choosing load resistors which are

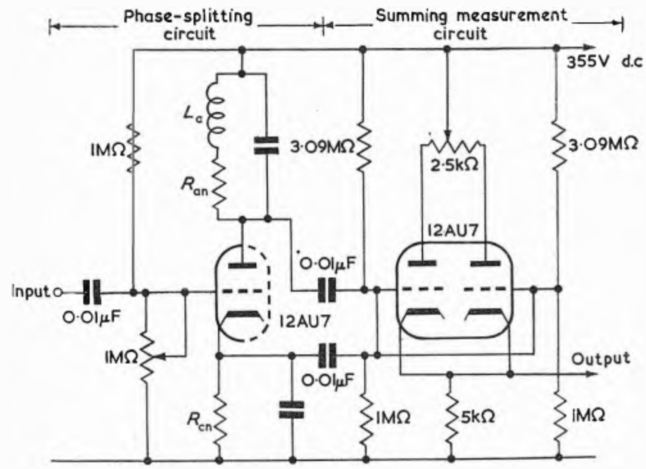

Fig. 6. Test phase-splitting circuit and associated summing measurement circuit

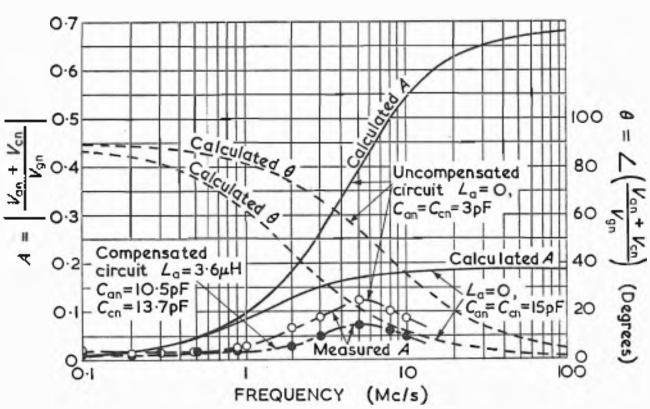

Fig. 7. Calculated and measured performance of combined cathode and anode outputs

much smaller than $1/g_m$ and assuring that the load impedances are equal. For best performance, the loading effect of subsequent circuits must be included into the cathode and anode loads of the phase-splitting circuit. Excellent use can then be made of a triode phase-splitting circuit as a balanced signal source in place of a balanced transformer for driving bridges and similar circuits.

Acknowledgments

The original analysis of the phase-splitting circuit was carried out by the author in 1952 while at the RCA Laboratories, Princeton, N.J. The calculations were programmed on an IBM 650 computer by Philip J. Rosen. The measurements were carried out by William Duane.

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AN AUTOMATIC CAMERA CONTROL

By H. R. A. Townsend*, M.B., B.Ch.

A miniature cold-cathode tube used to control a multi-contact relay forms a versatile unit with which to build a process controller. The camera control described incorporates a simple delay circuit, a synchronizing circuit and a remote firing circuit, as well as a binary counter also constructed with miniature cold-cathode tubes. It performs the relatively complex sequencing needed for fully automatic photography of traces displayed on a cathode-ray oscilloscope.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 780)

THE apparatus described is essentially a simple process controller, and is designed around miniature cold-cathode tubes and associated high impedance Post Office type 3 000 relays.

In taking a photograph from the screen of a c.r.t. a cycle of operations has to be performed in the order as listed below.

(1) Open shutter.

(2) Illuminate the face of an electromechanical counter for a predetermined period. (The counter is in the field of view of the camera and records a serial number on each frame.)

(3) Keep the shutter open for the exposure proper.

(4) Switch on the beam brightness during a flyback blanking period.

(5) Count a preset number of sweeps (1, 5, 10, 25, 50, 100) or alternatively if the camera is operating in the continuous feed condition, i.e. with a stationary spot and moving film, to count a preset number of frames passing the aperture.

(6) Switch off the beam brightness during a flyback blanking period.

(7) Close shutter.

(8) Advance electromechanical counter one digit.

(9) Wind on film one frame.

Three post office relays are used. Functions (1) and (2) are performed by the first relay; function (3) by the second relay; and functions (8) and (9) by the third relay. A pair of miniature high speed relays perform functions (4) and (6). Function (5) is performed by a seven-digit binary counter, consisting of a chain of bistable pairs of cold-cathode tubes.

A fourth Post Office relay is used to set the required number into the cold-cathode tube binary counter, and pulls in during the resting period between cycles; counting is performed backwards from the set number to zero, i.e. the state of the counter at any time shows the number of sweeps which have yet to be recorded.

Three remote controls are provided. A START button which initiates the cycle and is thereafter inoperative. A CANCEL button which is inoperative until a cycle has started but can cause closure of the shutter and a wind-on at any time during the cycle. A WAIT switch, which is a useful facility when adjustments have to be made in the middle of an exposure. In the waiting condition the beam brightness is turned off, and the binary counter ceases to operate, but holds the current count unchanged.

A MASTER switch sets the type of photograph which is to be taken. Not only does the master control the number of sweeps or frames to be counted, but motor connexions, c.r.t. plate connexions, and c.r.t. brilliance level are all adjusted appropriately. (The latter pair of functions via relays in the oscilloscope.)

Cold-Cathode Tubes

The chief advantage of using cold-cathode tubes to pull in the relays lies in the comparatively small power required to trigger the tubes. Two types of tube are used, both miniature and of essentially similar characteristics. (See Table 1.)

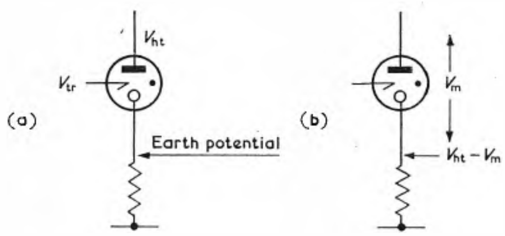

Fig. 1. (a) Untriggered tube; (b) Triggered tube

Referring to Fig. 1, the untriggered tube is subjected to the full h.t. potential (250 to 275V). Now if the trigger electrode, which in this state presents an input impedance of the order of megohms, exceeds a voltage of approximately +150V, with respect to cathode (which is at earth potential), then a discharge will form causing the whole tube to become conducting. The tube now drops a voltage

TABLE 1

Comparison of the two types of cold-cathode tube used in the apparatus

PARAMETER	TYPE 1 (RELAY ACTUATOR)	TYPE 2 (COUNTER)
Max. H.T. Voltage	310V	310V
Limits of V_m	121V 113V	140V 95V
Max. transient current	12mA	9mA
Recommended continuous current	3mA	4·5mA
Max. V_{tr} below which no tube will fire	137V	110V
Min. V_{tr} above which all tubes will fire	153V	170V

NOTE: *Type 1 tubes have priming cathode and will function in the dark. Type 2 tubes require a minimum of 5ft. candles incident illumination.*

* *The National Hospital for Nervous Diseases.*

V_m approximately 115 to 120V across it and the current is determined by the cathode load resistor, the cathode potential with respect to earth is now approximately +150V.

The Basic Relay Stage

This stage is shown in Fig. 2. The trigger tube simply acts as a switch pulling in the relay when it is turned on. Immediately the relay has pulled in it holds on, via the normally open contact, and disconnects the trigger tube, and all previously energized stages, via the normally closed contact. The slight complication of a 'hold' circuit is useful where a fairly heavy multi-contact relay is to be pulled in. Since the trigger tube only has to pass current for a short time a much smaller tube may be used than if it had to remain 'on' passing the relay current for long periods.

The use of a chain of contacts in the h.t. line might give rise to trouble if the number of stages were large, but it does have the advantage of being a simple and positive arrangement. Once a trigger tube has been switched on it can only be put out again by interrupting the h.t. supply. If the h.t. supply to a stage of this sort is restored at the same time as the 'hold on' circuit of the relay is broken then the large negative spike which will appear at the cathode due to the inductance of the relay is sufficient to re-trigger the tube even with the trigger electrode connected

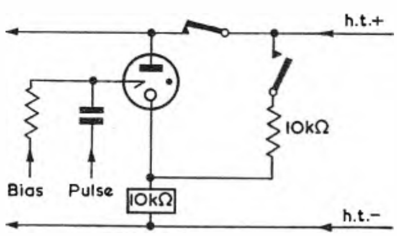

Fig. 2. Basic relay stage

to earth or to cathode, unless special arrangements such as short-circuiting the spike with an appropriate diode, are made. The use of a chain of contacts not only gets over these troubles but also provides an 'interlock' ensuring that a new cycle cannot be initiated until the previous one is complete.

Inputs to a Trigger Tube

A very useful input to a trigger tube is the pulse plus bias combination. A +100V pulse and a +100V d.c. bias are used, as indicated in Fig. 2. If the bias resistor is returned to earth and a pulse applied via the capacitor the tube will not be triggered, because V_{tr} will not have reached the critical value of >110V. Similarly the bias alone will not be effective but a pulse arriving when a bias is already present will produce a V_{tr} of the order of +200V and will be effective. This system is used for the input to the first stage of the counter. A variant of the same system is used for waveforms such as those produced at the output of counter stages, this waveform can be considered as supplying both pulse and bias through the same line and the input resistance and capacitor are simply placed in parallel. (Fig. 4(b)). Fig. 4(c) shows a method of firing a tube by means of a remote mechanical contact. Closing the contact causes the trigger electrode to be connected to a potential divider across the h.t. supply and so long as the resultant potential exceeds the critical value of V_{tr} then the tube will fire. The advantage of this arrangement is that, when the mechanical contact is open the capacitance of the line to earth is isolated from the trigger electrode by a series resistor which may be made of the order of megohms so long as the bias resistor is also large. Fig. 4(a) shows the circuit of a delay stage. The normally closed relay contact keeps the trigger potential close to earth until the contact is opened. The trigger potential then rises exponentially towards h.t.+ until the critical value of V_{tr} is exceeded when the tube will fire. The series resistor limits the trigger current to a safe value.

The Binary Counter

Fig. 3 shows a single stage of the binary counter. This consists of two tubes with a 0·1μF capacitor connected between their cathodes.

Suppose that initially both tubes are non-conducting. Now closing the contact marked 'set count' will fire the right-hand tube (Fig. 4(c)). The cathode potential of this tube will rise rapidly to +150V charging the 0·1μF capacitor to this voltage. The tapping point on the cathode

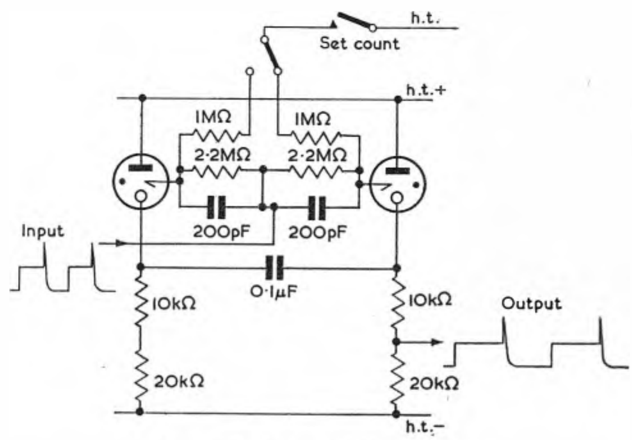

Fig. 3. Single stage of binary counter

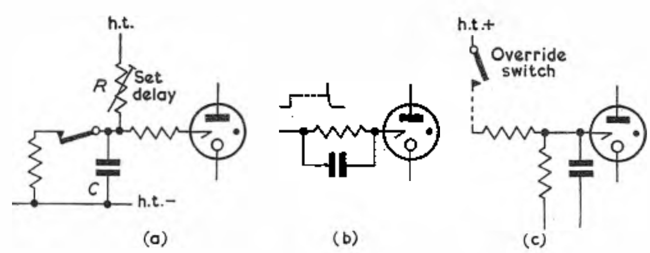

Fig. 4. Different methods of firing relay stage tubes

resistor chain will rise to +100V. The stage is stable in this condition. The set count contact is opened and the counter is ready to operate.

Assuming that the input has been at earth potential it now rises to +100V and the trigger of the non-conducting (left-hand) tube also rises to this voltage. The trigger of the right-hand tube (which is conducting) acts as a probe in the discharge and maintains itself at a voltage of about +170V but is effectively isolated by the large value series resistor.

Now when the input rises rapidly to +200V it will again carry the trigger of the non-conducting tube up to this potential, causing the non-conducting tube to fire. As the cathode of this tube (left-hand tube) rises to +150V the inter-cathode capacitor will cause the right-hand cathode to rise by a further 150V to +300V (almost). This is equivalent to removing the h.t. from the tube, which is extinguished. The voltage on the cathode of the right-hand tube thus produced will die away with a time-constant determined by the value of the inter-cathode capacitor and the cathode load resistance. This time-constant must be made of the order of 3msec in order to give the tube time to de-ionize.

The voltage at the tapping point will rise from +100V to +200V providing the output pulse. The stage acts as a binary counter because each +200V pulse at the input causes a change of state, but a +200V pulse at the output is only produced when the right-hand tube extinguishes.

The seven-digit binary counter is assembled on the tag-board at the top of the rack. Its h.t. supply is partially parallel stabilized and the counter which uses 20 per cent tolerance components works reliably in normal room lighting so long as the h.t. voltage is maintained within ±10V of the optimum value, determined by experiment as 240V.

The maximum speed of the counter has not been checked as the maximum required in this application is only 100 counts/sec.

B which holds on via contact B_1 and unlatches relay *A* via contact B_2. (A_8 is now open again and the START button has no further control until the cycle is complete.) Contact B_3 is also opened and the capacitor *C* charges up towards h.t.+ at a rate determined by the pre-set delay resistor. Other contacts on relay *B* (not shown) open the camera shutter and light up the face of an electromechanical counter so that a photographic record of the serial number of the frame is made. The delay period times this exposure and also gives the oscilloscope amplifier time to settle down after the transient caused by the operation of the shutter solenoid in the camera.

When the voltage across capacitor *C* exceeds the triggering voltage of V_1 this tube conducts and relay *C* is energized. Relays *A* and *B* are now disabled. Relay *C* carries a contact (not shown) which takes over from relay *B* before that relay falls out and keeps the camera shutter open in readiness for the exposure proper. Contact C_1 is again a hold contact: contact C_2 disconnects the previous stages. Contact C_3 applies h.t. voltage to V_2 and also a +100V bias to the first stage of the binary counter, provided that the 'run-wait' switch is not switched to wait. For most applications the binary counter is supplied with +100V pulses generated in the c.r.o. at each flyback and reference to Fig. 7 will explain the course of events.

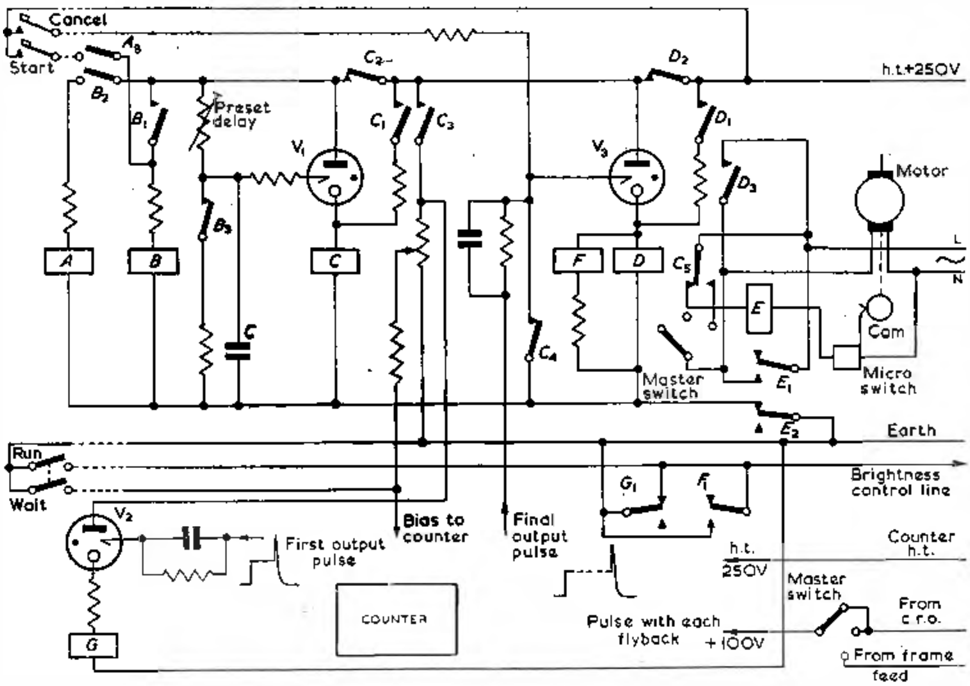

Fig. 5. Circuit diagram of complete controller
Note: Contacts on relays which are not essential to the sequencing of the controller are not drawn (e.g. contacts on relays 2 and 3 which energize the shutter solenoid in the camera)

The Camera Control

Turning now to the diagram of the apparatus as a whole (Fig. 5). With power off all relay contacts are in position as follows: B_2, C_2, B_3, D_2, and C_4 closed; C_5 to the left; G_1, F_1, E_1 and E_2 up. So that when first switched on relay *A* will be energized and pull in. All the power supplies come via the camera, and the apparatus will not switch on unless the camera is correctly set and loaded with film.

Contacts 1 to 7 on relay *A* are not shown on the diagram. They are the 'set count' contacts for the seven binary stages (see Fig. 3). The two way switch, determining whether the right- or left-hand tube will light, is replaced by a wafer on the master switch.

The master switch is the large rotary switch in the centre of the rack (seen end on in Fig. 6). It is an 11-way switch with 12 wafers. Seven of these wafers each control the setting of one stage of the binary counter.

Thus immediately the apparatus is switched on, a number, depending on the position of the master switch is set up in the counter, and contact A_8 is also closed.

The START button may now be pressed, energizing relay

It is assumed that five sweeps are to be superimposed on the film and the binary counter is initially set to the binary number 5 or 101. The first flyback pulse to arrive after the bias has been connected turns over the input binary stage and the pulse generated (in this instance at the cathode of the left-hand tube—ringed—) is arranged to light V_2. Relay *G* which is a small high speed relay in the cathode of this tube pulls in and switches on the oscilloscope brilliance, so that subsequent sweeps are photographed. Thus during the first sweep the counter displays 100 (binary 4). V_2 (the centre tube) forms a convenient 'brilliance on' indication. Note that the run-wait switch is in parallel with the contacts of relay *G*. In this oscilloscope the brilliance is turned off by earthing the brightness control line, so the run-wait switch both prevents the counter from counting flybacks and also prevents the uncounted sweeps from being photographed. This facility is very useful if adjustments have to be made during a long run of 50 or 100 sweeps.

As previously explained a carry occurs in the counter everytime that the right-hand tube of a pair is extinguished. Now during the 5th sweep the binary counter shows all zeros 0000. The flyback pulse occurring at the end of the 5th sweep causes a carry to run right along the line setting the stages to 1111 and extinguishing the right-hand tube of the last stage. This pulse—ringed—is the final output pulse which triggers V_3.

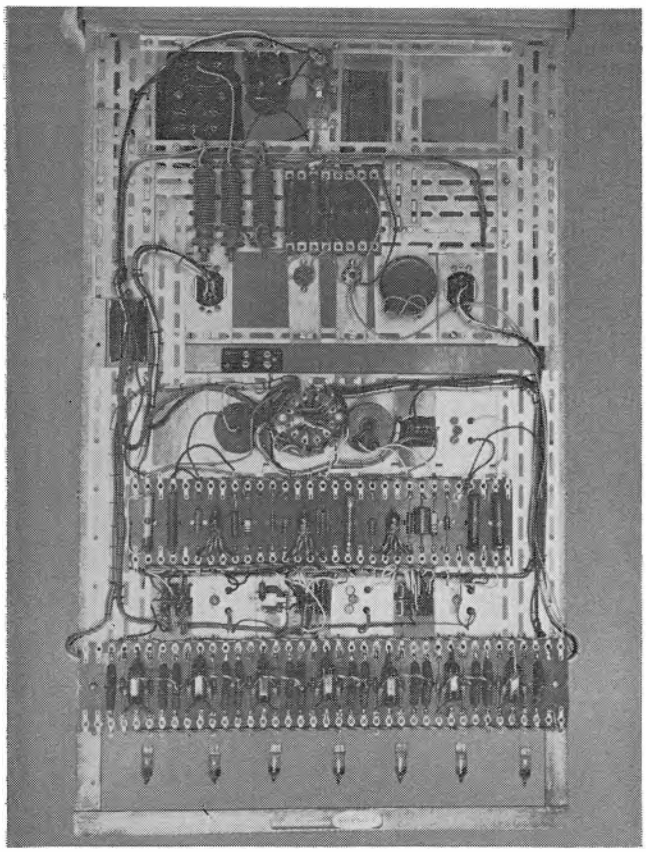

Fig. 6. A rear view of the camera controller

Contact C_4 is of course open at this time. This contact has the function of preventing inadvertent triggering of V_3 either by pressing the CANCEL button instead of the START button or by causing a general carry while altering the number set up in the counter. The CANCEL button is of course used in order to terminate a long count prematurely if desired.

The master switch is illustrated with three possible positions: The first (as drawn) corresponds to 'single shot', i.e. counter set to 0000001, so that V_3 is triggered after only a single sweep. The second position shown corresponds to 'superimpose'. In the present apparatus 5, 10, 25, 50, or 100 sweeps may be superimposed. The same brilliance level is used in all these cases, and corresponds to the minimum brilliance at which a trace will just register on the film. The brilliance is altered at these positions by a relay in the oscilloscope actuated by a spare wafer of the master switch.

The third position corresponds to 'continuous feed'. In this condition the oscilloscope time-base is disconnected leaving the spot stationary in the centre of the screen. This again is done by means of a relay in the oscilloscope. The film transport motor is made to run continuously while relay *C* is energized by the normally open contacts C_5 in series with the contacts on the master switch. The normally closed contacts C_5 open and prevent relay *E* (an a.c. relay in the camera, see below) from being operated in this condition. The counter is also switched by the master to count frames of film, instead of flybacks, via a small pulse transformer and settings are provided to count 8, 16 or 32 frames.

In all cases when the count is exhausted V_3 is triggered. Both relay *D* and the high speed relay *F* pull in together. A high speed relay is used to shut off the brightness because, in some instances if the flyback and waiting time are short, part of a further sweep may be recorded before relay *C* has unlatched. Contact D_1 is again a hold contact and

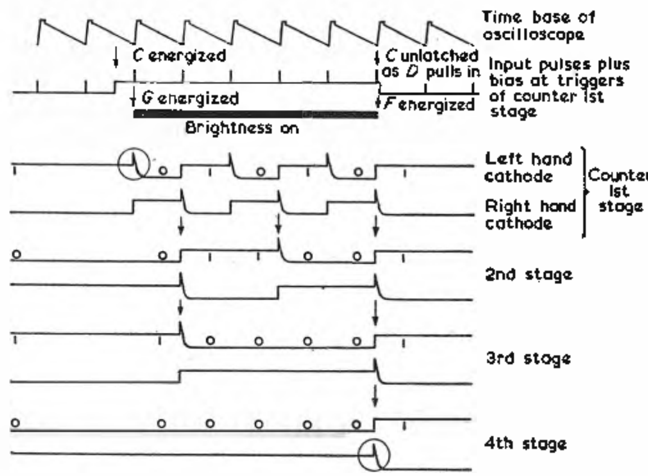

Fig. 7. Waveforms diagram to show relative timing

D_2 disconnects previous stages. Note, however, that the counter has its own h.t. supply and is unaffected. Contact D_3 causes the frame feed motor to start running. A further contact on relay *D* (not shown) causes the mechanical counter to advance one step during the wind on.

After about ½ turn of the frame feed shaft a microswitch closes, and energizes relay *E*—an a.c. relay in the camera—contacts on this relay E_1 close and keep the motor running until the microswitch opens again. Meanwhile, contacts E_2 open and break the main h.t. negative line, resetting the apparatus to its initial state.

Conclusion

The apparatus has been in routine use in the form shown in the photograph for the past twelve months. It has behaved reliably and required no maintenance.

Acknowledgments

The author wishes to thank the Department of Photography at the Institute of Neurology, Queen Square, who produced the illustration.

The author also desires to express his gratitude to the staff of the E.E.G. Department, particularly Mr. J. Plumb, who carried out the necessary alterations to the oscilloscope.

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The Electronic Simulation of the Load Applied to an Insect Muscle

By K. E. Machin*, M.A., Ph.D., A.M.I.E.E.

Certain insect muscles when suitably loaded contract rhythmically; the oscillation is largely controlled by the mechanical parameters of the load. An apparatus has been developed in which the mass, viscosity and stiffness of the load are simulated electronically, and can therefore readily be varied. The relation between force and length for an oscillating muscle can be displayed on a cathode-ray tube, while the frequency of oscillation and power output are presented on meters. The apparatus can also be used for direct measurement of the mechanical impedance of a muscle as a function of frequency.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 780)

NORMAL muscles when stimulated electrically give one transient contraction or 'twitch' for each electrical impulse; as the frequency of the stimulating impulses is increased these twitches fuse into a steady contraction or 'tetanus'. The wings of many insects are operated by muscles carrying out a series of twitches; in the motor nerve feeding the flight muscles there is therefore one electrical impulse for each wing beat. In certain insects, however, the motor nerve impulses are quite asynchronous with the wing-beat, and often of much lower frequency. The wing-beat frequency is then governed by the mechanical resonance of the wing system; the flight muscles when stimulated appear to be able to excite mechanical systems into oscillation over a wide range of frequency. In electrical analogy terms, when the 'circuit' (muscle) is 'switched on' (stimulated) its output impedance has a negative resistance term which can set into oscillation a 'tuned circuit' (mechanical resonant system) to which it is connected.

In the investigation of the properties of these 'fibrillar muscles' it was necessary to find how the amplitude, frequency etc. of the oscillation depended upon the parameters of the mechanical load, such as its mass, stiffness and viscous damping. Early experiments[1] were carried out using physical masses and springs, but it became clear that if adequate information were to be obtained during the time that the muscle preparation remained fresh (a few hours) some rapid method of varying the mechanical parameters of the load would be necessary. An apparatus was devised in which the mass, stiffness and viscosity of a load could be simulated electronically, and varied rapidly by potentiometers.

It was thought worthwhile in the present article to describe in some detail this apparatus and the ancillary units used with it, for the following reasons.

(1) The simulation of all three parameters of a mechanical load is believed to be novel.

(2) Certain unusual features in the use of mechano-electrical transducers are involved.

(3) The apparatus illustrates one definition of experimental zoology, as a subject in which '. . . the size of the apparatus exceeds by several orders of magnitude that of the animal.'

The Simulation of Mechanical Parameters

The muscle is attached to an electro-mechanical transducer which produces a force proportional to the current supplied to it (Fig. 1). To this is mechanically connected a mechano-electrical transducer giving an output proportional to displacement. This voltage is differentiated twice to produce 'velocity' and 'acceleration' voltages. Three potentiometers select fractions of the displacement, velocity and acceleration voltages. These fractions are added together and a current proportional to their sum is supplied to the electro-mechanical transducer.

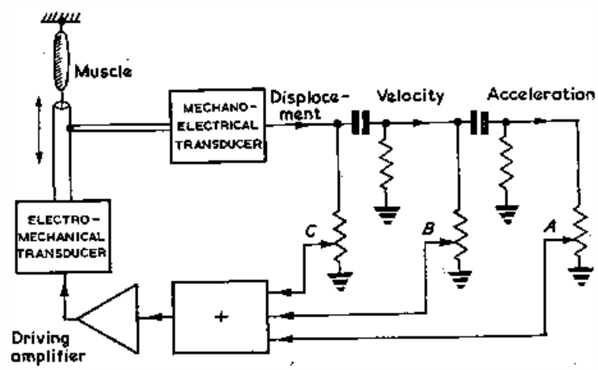

Fig. 1. Simulation of mechanical parameters

Hence the force which this transducer exerts can be expressed as:

$$F = A\,(d^2x/dt^2) + B\,(dx/dt) + Cx$$

where A, B and C are constants depending on the potentiometer settings. This equation is identical with that which describes a combination of mass A, viscous damping B and a spring of stiffness C; the shaft of the transducer will therefore behave in exactly the same way as such a combination of mechanical quantities.

It is also possible to make any of the constants A, B, C negative; the resulting negative values of apparent mass, viscosity and stiffness can cancel out the mass, viscosity and stiffness of the electro-mechanical transducer, or indeed of the muscle itself.

The apparent mass presented by the transducer shaft, while possessing normal inertial properties, is of course not subject to gravitational attraction.

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Transducer Head and Preparation Mounting

The electro-mechanical transducer is a moving-coil vibrator (Goodmans type V47, 300Ω). Attached to its output shaft is a photo-electric displacement transducer, described below. The mountings of these two transducers must be exceedingly massive and rigid as otherwise spurious oscillations may occur at one of the resonant frequencies of the mounting. The mounting shown in Fig. 2 was originally used, but has now been modified so that the transducers are clamped directly on to a 12kg block of iron.

The insect under study is impaled by two needles into a universally mounted holder. A piece of 0·0024in Nichrome wire is tied to the end of the muscle under investigation, and connects it to the vibrator shaft.

The preparation holder is mounted on a mechano-electrical transducer (described below) which measures the force applied to it, and hence to the muscle.

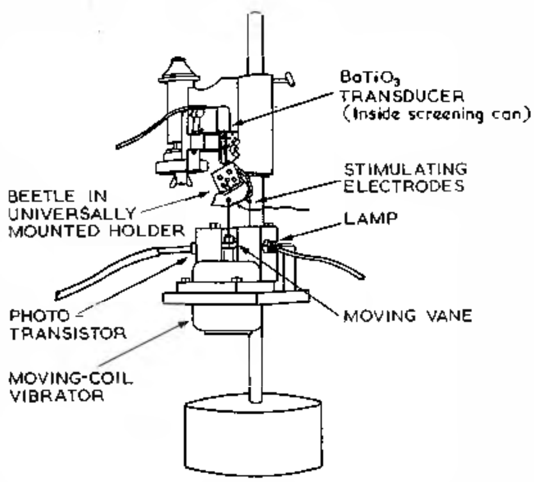


Fig. 2. Transducer head and preparation mounting

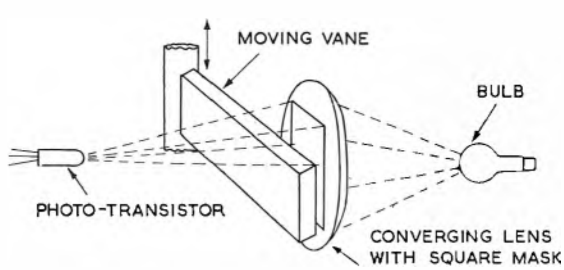


Fig. 3. Photo-electric displacement transducer

Photo-electric Transducer

The light from a 6V, 0·3A bulb is focused by a small lens on to the end of an OCP71 photo-transistor; the lens is masked to have a rectangular aperture. Immediately in front of this aperture moves a stiff vane made of electron metal, which is attached to the output shaft of the moving-coil vibrator. The movement of the vane thus causes a change in the current of the photo-transistor. The arrangement is shown diagrammatically in Fig. 3.

Early attempts to use this system were frustrated by non-linearity of response and by temperature drift of sensitivity and mean current. It was not at first clear whether these effects were occurring at the photo-junction or in the transistor amplification process; if the latter, then improvement might be obtained by applying negative feedback. Accordingly a calibration of the OCP71 with and without feedback was carried out with the aid of a point source of light, an optical bench and the inverse square law. Fig. 4 shows that with feedback both linearity and temperature stability are greatly improved, at the expense of a corresponding loss of sensitivity.

The circuit finally used (Fig. 5) incorporates negative feedback; by varying the degree of feedback the sensitivity of the transducer can be set to a standard value. Push-pull output is provided. Fig. 5 also shows the simple stabilizing circuit used to feed the lamp.

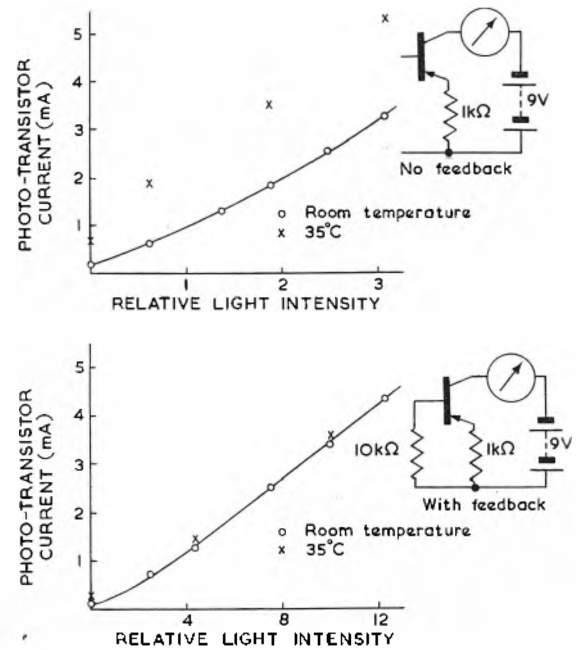


Fig. 4. Performance of photo-transistor with and without feedback

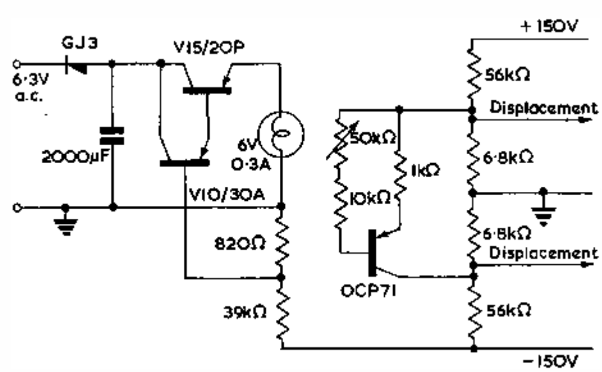


Fig. 5. Photo-transistor circuit and lamp stabilizer

Differentiating Circuits

The potentiometers which set 'mass' 'viscosity' and 'stiffness', together with the circuits between them and the driving amplifier, are shown in Fig. 6. V_1 acts as a feedback differentiator fed from the 'mass' control, and as a straightforward feedback amplifier fed from the 'viscosity' control. The output from V_1 is fed via C_1 into the very low (~300Ω) input impedance of the driving amplifier; a further differentiation thus occurs. Each differentiation is sufficiently accurate that less than 1° of spurious phase shift is introduced at all frequencies below 300c/s. The 'stiffness' control feeds through R_1 directly into the input of the driving amplifier; the two resistors associated with the control neutralize the d.c. component at the output of the photo-electric transducer. A manually variable current ('force') is also fed into the driving amplifier.

The ranges of the controls are as follows: mass, -4 to $+30$ grams; viscosity, -4000 to $+5000$ dyn/cm sec^{-1}; stiffness, 0 to 2×10^6dyn/cm.

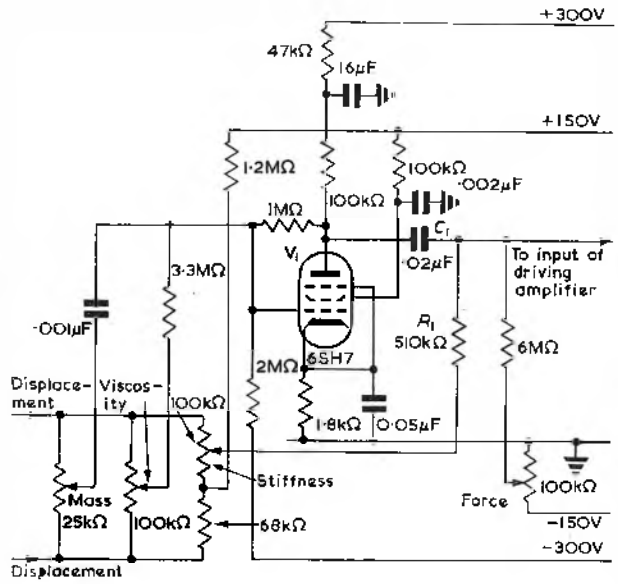


Fig. 6. Differentiating circuit

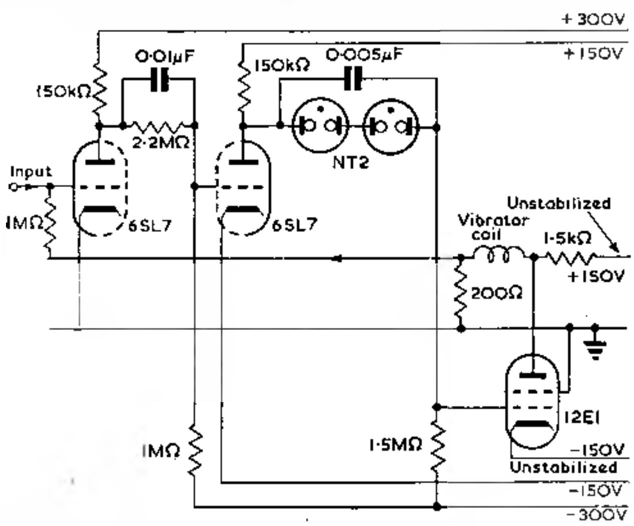


Fig. 7. Driving amplifier

Driving Amplifier

The amplifier which drives the moving-coil vibrator is shown in Fig. 7. It is designed to produce a current in the vibrator which is an amplified version of the current fed into the input terminal. The current gain is 5 000 over the frequency range 0 to 1kc/s; the phase shift below 200c/s is less than 1°. Due to the heavy negative feedback the input impedance is very low—1kΩ at zero frequency, falling to 300Ω at 100c/s.

Force Transducer Circuit

To avoid phase-shift in the force transducer system, its resonant frequency must be well above the working range; in addition its compliance must be very low, as otherwise unrecorded changes in the length of the muscle could occur. A barium titanate piezo-electric transducer was finally chosen; the unit used (G.E.C. type S.P.1100, force applied along long axis) had a compliance of a less than 10^{-10}cm/dyn. The combination of transducer and specimen holder had a resonant frequency of about 5kc/s.

However, a piezo-electric transducer has a frequency response which does not extend down to zero; the cut-off frequency depends on the resistive load on the piezo-element. It was essential in the present application to have a measure of mean muscle force as well as the oscillating component, and so response to zero-frequency force was necessary. This was achieved as follows.

At low frequencies the current in the vibrator is a true representation of the force which it exerts. Some of this force is used to overcome suspension stiffness; for small movements this 'lost force' is proportional to displace-

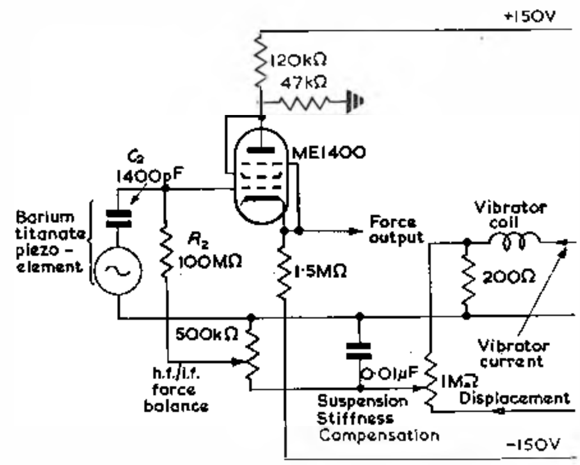


Fig. 8. Force transducer circuit

ment. Thus if a voltage proportional to vibrator current is suitably mixed with a voltage representing displacement, the resultant voltage is a measure of the force exerted by the vibrator on the external load. Such a voltage is only a true representation at low vibration frequencies; at higher frequencies some of the force is needed to overcome inertia and damping. On the other hand the output of the piezo-electric transducer represents force accurately only at high frequencies. By mixing the two voltages, an output truly representing force at all frequencies from zero to near the transducer resonance can be obtained.

It will be seen from Fig. 8 that the l.f. compensating voltage is fed into the bottom of the grid resistor of the valve following the piezo-element. The transducer output reaches the grid through the high-pass filter C_2R_2, while the compensating voltage is applied through the low-pass filter R_2C_2. The sum of the two voltages is thus a frequency-independent measure of force.

An alternative explanation of the operation of the circuit can be framed in the following terms. When a force is first applied to the transducer, a voltage appears across it. The charge tends to leak away via R_2, but before appreciable charge has dissipated the lower end of R_2 is brought to the same potential as its upper end. No further discharge current flows.

The 'crossover frequency" is arranged to be about 1c/s.

Cathode-ray Tube Display

The voltage from the 'force' output, suitably amplified, is fed to the Y plates of a c.r.t., and the 'displacement'

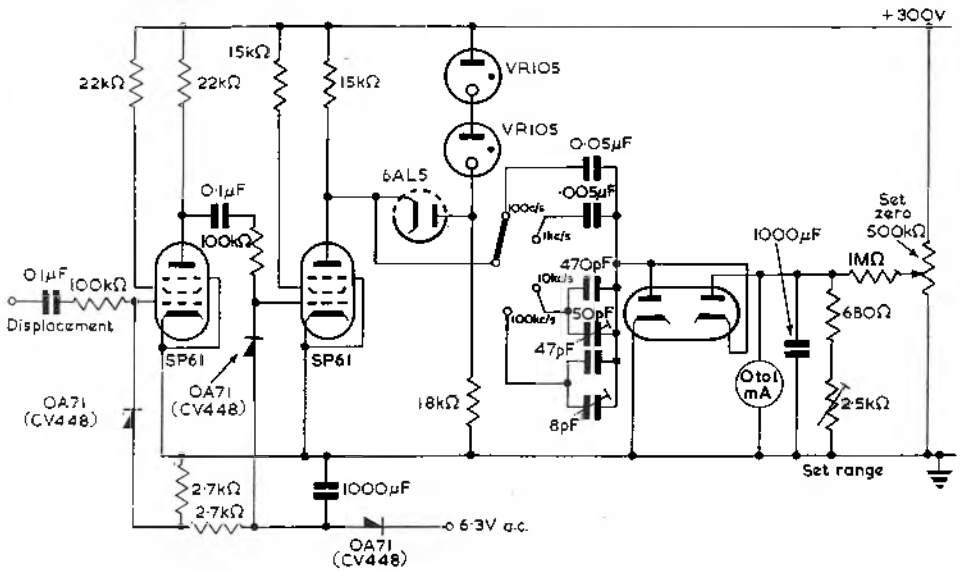


Fig. 9. Frequency meter

output is similarly fed to the X plates. The tube, which thus displays a plot of tension against length, is photographed. The relation between tension and length for the unstimulated and stimulated muscle can be photographed point by point by varying the 'force' control. When the muscle is oscillating, the spot on the screen traces out an approximately elliptical path once per cycle; the area within the ellipse represents the work done by the muscle per cycle.

Setting-up Procedure

After an initial warming-up period, the photo-transistor feedback control is adjusted, with the driving amplifier off and no mechanical load on the vibrator, so that the voltage at the 'displacement' output has a standard value. The sensitivity of the displacement transducer is thereby standardized. The driving amplifier is switched on and the 'suspension stiffness compensation' control adjusted until no movement in the 'force' direction is seen on the c.r.t. display as the vibrator is moved up and down with the 'force' control. A dummy muscle, such as a length of rubber, is now mounted on the 'h.f./l.f. force balance' control adjusted until sudden movement of the 'force' control produces a movement of the c.r.t. spot which is free from overshoot or undershoot.

Ancillary Units

In addition to the cathode-ray tube display two other output indicators are provided. Firstly, the frequency of oscillation is indicated on a meter. Secondly, a direct meter presentation of mean power output of the muscle is available; this is particularly useful in finding the optimum conditions for oscillation.

Frequency Meter

The design of the frequency meter (Fig. 9) is conventional. The input waveform is squared twice, the excursion of the second squarer anode being defined by a catching diode and two neons. The squared output feeds a diode pump with provision for backing-off the standing emission and for setting the instrument against a known frequency. The upper two ranges are not used in the present experiments.

Power Meter

This instrument (Fig. 10) displays the mean power output of the muscle when it is driving the resonant load. It is based on a twin-coil dynamometer movement, the coils being driven by amplifiers with heavy current feedback. One amplifier is fed with a voltage from the 'force' output, while the other receives the 'displacement' output after differentiation by C_3R_3. The meter thus reads the mean value of force times velocity, i.e. mean mechanical power output.

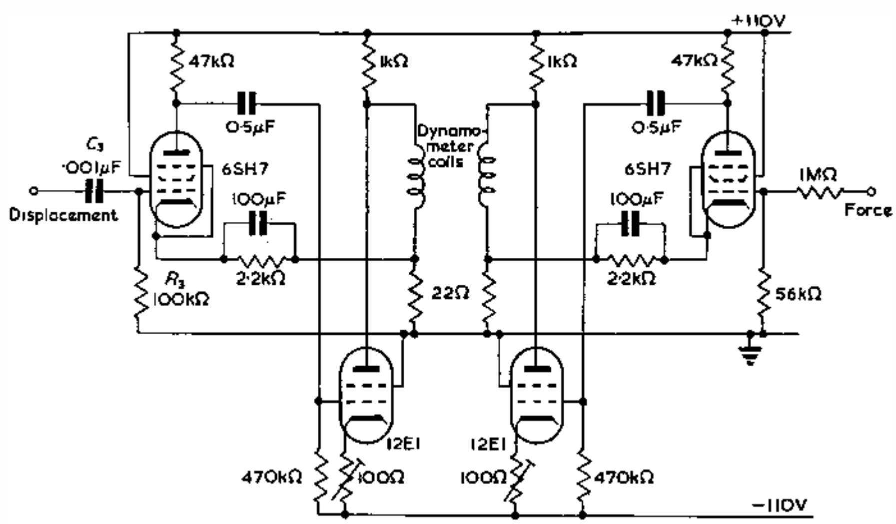
Direct Measurement of the Mechanical Impedance of the Muscle

It was clear from 'free oscillation' experiments[3] that not only was the power output of the muscle a function of the frequency, but that its apparent stiffness also varied with frequency. In other words, the mechanical impedance of the muscle had real and imaginary parts which were both frequency-dependent. Provision was therefore made for measuring directly the mechanical impedance of the muscle at various frequencies. This involved varying the length of the muscle sinusoidally and measuring the resulting force in both amplitude and phase.

Circuits for Driven Oscillation

The natural frequency of the vibrator was about 10c/s, making it impossible merely to drive it directly from an oscillator. Instead, displacement feedback was applied (Fig. 11) to the input of the driving amplifier, and a small alternating current injected from an oscillator. The feedback was equivalent to a very high setting of the 'stiffness'—about 100 times the full scale of the normal control. This

Fig. 10. Power meter



raised the vibrator resonant frequency by a factor of fifty to about 500c/s, and gave its shaft an apparent stiffness of about 10^8dyn/cm (100kg.wt/cm). The muscle could therefore be driven sinusoidally in length at any frequency up to 250c/s, the drive having a very high mechanical impedance.

The capacitor C_4 stabilizes the feedback loop, while C_5 is chosen to minimize the high-frequency phase shift between oscillator voltage and displacement. It is then possible to set the phase shift exactly to zero at, say, 250c/s with the 'mass' control, which provides variable second differential feedback.

The oscillator used is part of the Solartron 'transfer function analyser', the other unit of which is a phase-sensitive voltmeter displaying directly the components of its input which are in phase and in quadrature with the oscillator. Thus if this phase-sensitive voltmeter is fed from the 'force' output, it will display directly the elastic and viscous components of the force. The real and imaginary components of the mechanical impedance follow immediately.

Conclusion

The apparatus described was designed, or more accurately it evolved, as a tool for a particular research project. It is hoped that the multiplicity of pre-set controls, valve types and power supplies can therefore be excused. However, one point of sound design is claimed: every valve and transistor in the circuit is within at least one feedback loop.

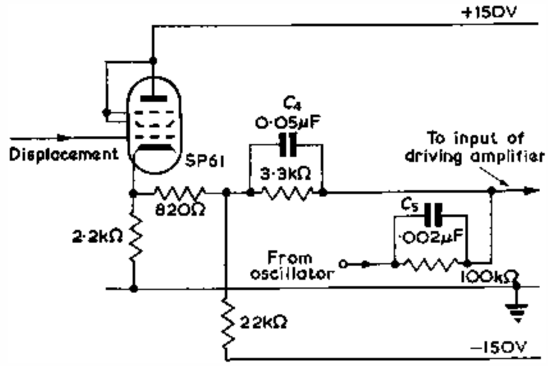


Fig. 11. Circuit arranged for driven oscillation

Acknowledgments

The author is grateful to Dr. J. W. S. Pringle, F.R.S., for much valuable discussion. The transfer function analyser was acquired through a grant from the Royal Society to Dr. Pringle.

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A Prototype Electronic Telephone Exchange

An experimental electronic telephone exchange is now undergoing exhaustive tests at the Post Office Research Station at Dollis Hill, London. When these tests have been completed, equipment for about 1 000 lines will probably be installed within two years for the public exchange at Highgate Wood, in North London.

The introduction of electronics to telephone exchange design is expected ultimately to reduce initial cost, provide smaller and less equipment which will need less space in buildings, and give greater reliability because of the virtual absence of moving parts.

Although this experimental model is not built as a public exchange, it does provide facilities for telephone calls within the Research Station and also into the London public telephone system. It will be used to prove the principles involved in designing electronic exchange systems.

As early as 1955 it was realized that a very big research and development effort would be required to apply electronic techniques to telephone switching and provide an effective answer within a reasonable time.

So in 1956 the five principal manufacturers of telephone exchange switching systems and the British Post Office decided to set up a Joint Electronic Research Committee (J.E.R.C.) under the chairmanship of Sir Lionel H. Harris, K.B.E., T.D., Engineer-in-Chief of the Post Office.

In pooling their research and development resources, the six organizations making up the Committee decided to undertake long-term studies on the best system to be employed and also to obtain practical experience in designing and operating exchanges using electronic equipment.

The experimental exchange at Dollis Hill will help in furthering both objects. Each member organization of J.E.R.C has undertaken detailed study of a portion of the system, and the individual parts are now working together in the experimental exchange.

The multiplex and speech circuit equipment have been designed jointly by the Post Office and the General Electric Co. Ltd; the supervisory equipment by Ericsson Telephones Ltd; the register equipment by Siemens Edison Swan Ltd, and the translation, metering, line information and other equipment jointly by Automatic Telephone & Electric Co. Ltd and Standard Telephones & Cables Ltd. Similar divisions of responsibility and manufacture will occur in the Highgate Wood public exchange installation of which the Dollis Hill model is the laboratory version.

Basically the exchange will employ the principle of time division multiplex and switched highways for the ultimate connexion, where each highway can carry up to 100 simultaneous conversations and sufficient highways are provided depending on the traffic of the exchange. The exchange lines are arranged in groups and each line in a group is connected by electronic gate circuits to the highway common to all lines in that group. Each highway in turn is able to be connected to any other, again by electronic gate circuits. These highways transmit speech and other signals to and from the lines in the form of amplitude modulated pulses. The number of lines in a group connected to a highway is several times the possible number of simultaneous conversations depending on the traffic originated by the lines.

Each connexion on a highway uses a pulse channel transmitting one of a series of interleaved cyclically recurring pulse trains, typically of 10kc/s repetition frequency. Lines are connected to channels by applying suitably timed pulses to the gate circuits. Magneto-strictive delay lines are used as temporary memories for storing the pulses which have been allocated to the subscribers' gates for connexion to the highway, and also for controlling the gates inter-connecting the highways. Magneto-strictive delay-lines are used as registers to receive information dialled by subscribers. They count and store the dialled pulses, obtain any translations which are necessary to route calls to other exchanges and transmit digit pulses when outgoing routes are established. Delay-lines also control the supervision of connexions after they have been established. They connect tones to the calling lines and ringing to the called lines at appropriate times, detect answering and clearing and arrange for the metering of successful calls.

A magnetic drum is used as a semi-permanent memory. This drum carries all the information relative to all subscribers' lines and junctions, acting in this respect as the exchange library. Another portion of the drum co-operates with register equipment by providing translation facilities for routing the call either to a local subscriber or to subscribers on other exchanges. Positions on the drum particular to each subscriber are used to record the accumulated total of unit fee registration debited against the subscriber and as such replace the individual electro-mechanical message registers provided per subscriber in existing systems. Such accumulated totals can be read from the drum as and when required for subsequent automatic accounting for the final subscribers' account.

A Simple Current Stabilizer for Electromagnets

By M. H. N. Potok*, M.A., B.Sc., Ph.,D., A.M.I.E.E., M.I.R.E.

For certain experiments it was necessary to provide a magnetic field which was free of ripple and of slow drift, the electromagnet concerned required about 10A at 100V. In this article the various methods of supplying the electromagnet are briefly reviewed and the solution adopted is described. The power is supplied by an Amplidyne. Transistors are used to reduce ripple and a d.c. amplifier is employed to overcome slow drift.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 780)

IN connexion with some experiments designed to measure the *g* value of electrons it has become necessary to provide a magnetic field which would be free of ripple as well as of slow drift. This problem has recently been receiving the attention of several workers[1,2,3] but it may be of interest to give some details of a comparatively simple solution. well at the lower load required. The commutator ripple was found to be of the order of 0·5V, and the attempt to remove it by an *LC* smoothing circuit led to instability when feedback was applied, because the Amplidyne is a two-lag device, and the smoothing circuit provided the third lag leading to self-oscillation at a few cycles per second. Eventually, it was decided to use only capacitance for rough smoothing and power transistors for final ripple reduction while a d.c. amplifier is used to correct for slow drift by comparing the voltage across a standard resistor against a dry battery, and feeding the error signal into the field coil of the Amplidyne. The Amplidyne response with sufficient resistance in the quadrature winding is of the order of 0·1sec, and the transistor smoothing circuit takes over below that.

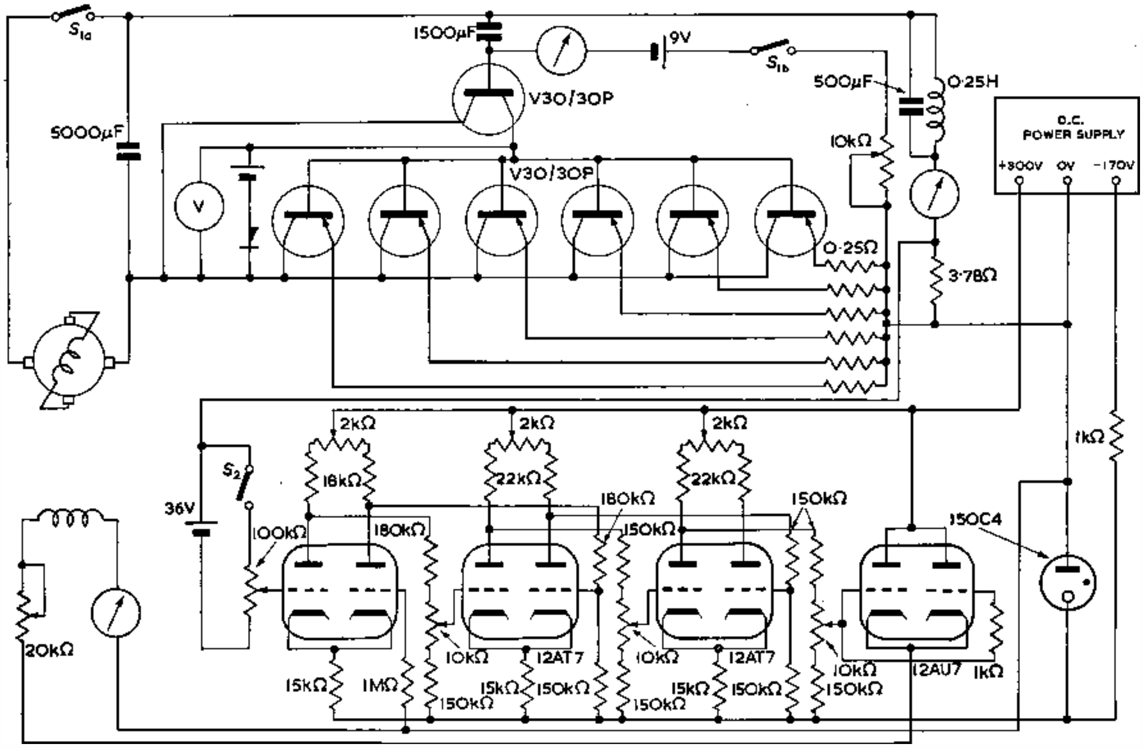

Fig. 1. The current stabilizer circuit

The air cored electromagnet as designed required about 10A at 100V. Various methods of stabilizing the field were considered, such as the proton resonance effect, the cyclotron effect and the Hall effect but in view of the comparatively weak field required (less than 300G), shortage of space inside the coil and the high stability aimed at, these were found to be unsatisfactory, and magnet current stabilization was chosen as the best approach. Here the choice was between a battery, a d.c. generator and a rectified a.c. supply. The battery would avoid the ripple problem but it was realized that that would not be as great a problem as the current drift due to battery voltage and magnet resistance changes. Also a battery to give 10A, 100V is bulky and would require frequent recharging. It was decided eventually to use an Amplidyne. The one available was rated at 25A, 220V but behaved reasonably

The circuit is as shown in Fig. 1.

The d.c. amplifier is a conventional one consisting of double triodes in long-tail pairs. The transfer conductance is of the order of 1A/V. The drift is of the order of 1mV between input grids. This is rather more than one would like, and to reduce its effect, and in view of the extra power available from the Amplidyne, a comparatively large standard resistance was chosen dropping 36V

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which is compared with a 36V dry battery. This should give current stability approaching 3 in 10^5. It is proposed eventually to place the resistor in a bath of alcohol, arranged so that the heat produced will just boil the alcohol, and so keep the temperature constant. The alcohol is, of course, condensed and returned to the bath. (This idea is due to Dr. McCutchen of this Laboratory, and has worked well in another connexion.) The transistor smoothing circuit gives a 400 to 1 reduction in ripple and reduces considerably such troublesome effects as the Amplidyne rumble. This is unexpectedly large and irregular and many attempts were made to reduce it, such as attention to the bearings, commutator and brushes but without significant improvements. The transistor circuit copes with it successfully except for very occasional spikes which are thought to be due to the carbon brushes. The principle of operation of this transistor smoothing circuit has been described by the author earlier[4]. The transistors used here are the V30/30P rated at 10W, 3A maximum. Four of these are required in parallel to allow a load of 10A but the current amplification at 2·5A is low, hence it has been decided to use six of them. This also provides a safety margin in case of sudden overload. The diode in the circuit is used to safeguard the transistors against overvoltage. and the resulting burn out. Current to be stabilized can be changed by varying either the reference voltage or the standard resistance. Since long term stability is not required, the reference voltage can be changed by tapping a resistor across the reference battery. A very smooth control of current is thus obtained. The resistor (100kΩ), of course, discharges the battery slowly thus causing a slow drift in current. The present stability is of the order of 5×10^{-5}, and it is proposed to use a chopper d.c. amplifier to obtain a greater stability if and when required.

The operation is comparatively simple although some care has to be exercised in bringing the circuit into operation if sudden overloads are to be avoided.

No attempt has been made to follow the fashion of using transistors only in place of the 'anachronistic' valves[3], since it is strongly held by the author that where a 'one-off' piece of equipment is required for a comparatively short-lived experiment, ingenuity in devising novel circuits where old-established ones will work, is totally misplaced both in terms of economy and effort.

Ackowledgments

The author wishes to acknowledge many useful discussions with Mr. H. V. Beck and Dr. McCutchen of this Laboratory.

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A New Acoustic Research Laboratory

A NEW acoustic research laboratory has recently been opened at Hemel Hempstead, Herts, by the Acoustical Investigation and Research Organization Ltd (AIRO) who are an independent, non-trading organization formed for the purpose of practical investigation and research into all aspects of noise problems.

The new laboratory is the first in this country to bring under one roof all the special equipment and techniques required for acoustic analysis and measurement and it has been designed in collaboration with the National Physical Laboratory and the Building Research Station with whom AIRO maintain close liaison.

The laboratory will undertake these principal types of work.

(1) Measurement of the acoustic characteristics of machines, materials and electro-acoustic transducers.

(2) Development projects involving acoustical problems. One of the first to be investigated by the laboratory will be the provision of effective sound insulation in lightweight structures such as office partitions and bulkheads between passenger cabins on ships.

(3) Fundamental research in the field of acoustic engineering and electro-acoustics.

The laboratory consists of four experimental chambers and an underground duct. All are connected to a control room, to which information is transmitted for recording and analysis of the data obtained.

Two of the experimental test rooms require maximum insulation from extraneous noise. These are the anechoic chamber and the reverberation chamber, where all surfaces reflect noise. To achieve the necessary degree of insulation, both these chambers are separately supported on high-compression rubber mountings. They are, therefore, completely independent of the structure of the building. In addition incoming ducts and leads are insulated and a mineral wool blanket is suspended in the cavity between these chambers and the main structure.

In the anechoic chamber there is a sound-absorbent lining on all walls floor and ceiling. This consists of 'Bondacous' in a specially devised formation up to a thickness of 30in. There is a retractable metal grid floor which is used for setting up experiments.

In the reverberation chamber all surfaces are reflective and adequate diffusion of sound is obtained by introducing reflector plates suspended from the ceiling. This technique has not been used previously in this country but is now internationally accepted.

In the ceiling of the reverberation chamber there is a pre-cast concrete panel which can be raised by hoist into an upper source chamber above. It can then be removed along a gantry and lowered to ground level for resurfacing with new materials which require to be tested for sound transmission properties.

No two walls in the reverberation chamber are parallel to one another the object being to prevent undesirable effects associated with rectangular rooms.

The same requirement applies to the source chamber which although not suspended independently, has highly-reflective surfaces and no two walls in parallel.

The underground duct which is adjacent to the building is 40ft long by 3ft square. It contains an anechoic wedge structure which can be blanked off by reflective plates when required.

The scope of the laboratory's analytical and measuring facilities can be indicated by a brief description of the type of work which will be undertaken.

Measurement of Sound Absorption

The sound absorption coefficients at random incidence of any natural or manufactured material over the audible frequency range 40c/s to 10kc/s can be accurately measured. In the case of frequencies at or above 100c/s, these tests will be carried out in the reverberation chamber according to British Standards and in conditions identical to those at the National Physical Laboratory, using samples of 100ft² placed on the floor of the chamber. Measurements of the chamber's reverberation time with and without the sample enables the absorption coefficient of the materials to be calculated. Measurements of the sound absorption coefficients of materials at frequencies below 100c/s will be carried out in the underground duct using the normal standing-wave technique. There are also facilities for measuring the sound absorption coefficients of materials at normal incidence in the frequency range 125c/s to 4kc/s using the same technique in the standing wave tube.

Sound Transmission and Sound Insulation

This facility is particularly useful in the designing of sound-reducing partitions and windows, either from the point of view of their construction or the materials used.

For example, in testing partition materials and constructions, the experimental sample is placed in the test aperture between the reverberation chamber and source chamber. An accurate measurement can then be taken of the airborne sound transmission loss characteristics of any type of partition at ⅓rd octave intervals in the frequency range 40c/s to 10kc/s according to B.S.2750:1956.

Similar measurements can be made for sound reducing floors and ceilings. In this case the test sample is placed in the aperture between the upper source chamber and the reverberation chamber. The efficiency of a floor or ceiling in reducing the transmission of impact sound, such as is caused by, for example, footsteps or machinery, can also be measured.

In testing window constructions under normal conditions the sample is placed in the aperture between the reverberation chamber and the anechoic chamber, since the anechoic chamber gives the effect of normal open air conditions outside a window.

Noise Characteristics of Machines

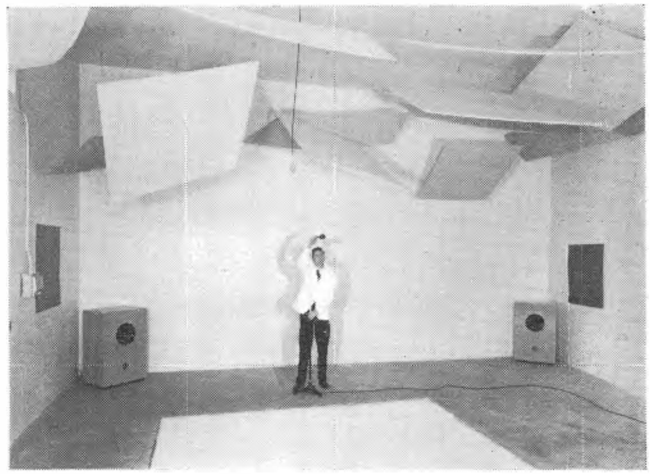
Measurement of the noise characteristics of all types of mechanical and electrical machines and components will be carried out in the laboratory. The reverberation chamber, the anechoic chamber and the underground duct will all be employed to calculate sound power levels and spectral analysis of machine noise in the frequency range 20 to 20 000c/s at frequency intervals of from ⅓rd octave to 1/40 octave as required. Directivity patterns and polar diagrams can also be produced.

In particular, the noise characteristics of all types of ventilation plant can be measured. Fan speeds can be varied within wide limits and measurements can be taken under given suction or discharge pressure conditions. Where overall sound power levels are to be determined the test fan will be mounted in the reverberation chamber. Alternately, the test plant can be mounted in the anechoic chamber to enable directivity characteristics to be measured in open air conditions. Ventilation plant may be tested either working into free air or with duct work installed at suction and/or discharge sides.

The laboratory will also investigate the problems of noise generation by all types of machines, and projects involving the reduction of noise in specific mechanical devices will be undertaken.

Vibration

The laboratory is fully equipped to carry out detailed measurements of all the parameters of mechanical vibrations, spectral analyses of vibration acceleration, and determination of velocity and displacement. The problems of vibration isolation can be fully investigated and resolved.



The reverberation chamber

Electro-Acoustics

A major activity at the laboratory will be the measurements and calibration of all types of electro-acoustic transducers, in particular, microphones and loudspeakers.

Measurements of the sensitivity of all types of microphones will be carried out in the anechoic chamber at any required range or incidence. Free field microphone frequency response measurements can be taken over the frequency range 20c/s to 20kc/s and the response can be measured either as a continuous function of frequency or as point-to-point measurements, at frequency intervals which can be as close as 1/40 octave apart. Free field measurements at frequencies above 100c/s will be carried out in the anechoic chamber and measurements at frequencies below 160c/s will be made in the underground duct. An alternation technique will be used.

Full facilities are also available for the determination of microphone directivity characteristics and sets of polar diagrams can be produced for all frequencies under free field conditions.

The laboratory standard reference microphones will be periodically calibrated at the National Physical Laboratory. In addition absolute reciprocity calibrations will be carried out frequently in the laboratory.

Comprehensive measurements on the performance of loudspeakers under free-field conditions can be undertaken in accordance with B.S. 2498:1954. The measurements include: —Sensitivity, Frequency responses, Mean spherical responses, Polar responses, Impedance measurements, Transient responses, Non-linear distortion, Efficiency.

These measurements can be carried out in the frequency range 20c/s to 20kc/s. Frequency response curves will be produced on the basis of either point-to-point measurements or continuous frequency responses whichever is required.

To simulate free field conditions the loudspeaker will be placed on the anechoic chamber. At frequencies below 160c/s the underground duct will be used.

The laboratory will also undertake research on the design of loudspeakers.

Measurement can be made of the performance of many other types of electro-acoustic transducers—such as artificial ears and telephones—and of audio-electronic instruments, such as tape recorders and reproducers, amplifiers, etc.

High Frequency Transistor Filter Synthesis

By L. M. Vallese*, D.Sc.

Methods of synthesis of high frequency transistor filters based on the use of the forward equivalent network transformation are developed. For purposes of simplification the transistor intrinsic feedback, which is very small in diffused base units, is neglected. Two canonical ladder structures, of hybrid-π and of hybrid-T type respectively, are considered for the synthesis of complex pole pairs: these are cascaded with (and in certain cases without) the interposition of buffer stages. Examples of video and band-pass filters are shown.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 780)

In recent years the problem of the synthesis of active networks has been receiving increasing attention as attested by the large number of technical papers on the subject. The attractiveness of these networks arises from their greater freedom of design and from their gain feature, characteristics which with proper use may offset the disadvantages of lesser reliability of the parameters.

The synthesis of transistor networks is usually complicated by the presence of intrinsic reactances and intrinsic feedback. However, modern technical advances have provided graded base structures with negligible intrinsic feedback, which permit a considerable simplification of the design. In the following, the problem of the realization of high frequency transistor filters is discussed taking into consideration the base-emitter capacitance and neglecting the collector-base capacitance and the base spread resistance. Methods of synthesis of low- and band-pass ladder structures are developed using a procedure based on the network transformation from the non-reciprocal to the reciprocal form, which was previously discussed [1,2]. The entire filter is built with canonical structures of the hybrid-π or of the hybrid-T type, which realize each one pair of complex poles; the hybrid-π structures are cascaded with the intermediary of buffer stages, while the hybrid-T structures are cascaded directly. Experimental verifications and examples of design are included.

Network Transformation

Given a two-port non-reciprocal network N with admittance matrix:

$$\begin{Vmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{Vmatrix} \quad \text{(1)}$$

and load admittance Y_L, the network of matrix

$$\begin{Vmatrix} Y_{11}+Y_{21}(Y_{21}-Y_{12})/(Y_{22}+Y_L) & Y_{21} \\ Y_{21} & Y_{22} \end{Vmatrix} \quad \text{(2)}$$

possesses the same forward transfer and driving immittance properties and the same stability as N, when closed on the load Y_L.

Similarly, in terms of the impedance matrix of N

$$\begin{Vmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{Vmatrix} \quad \text{(3)}$$

indicating with Z_L the load impedance, the corresponding forward equivalent matrix is

$$\begin{Vmatrix} Z_{11}+Z_{21}(Z_{21}-Z_{12})/(Z_{22}+Z_L) & Z_{21} \\ Z_{21} & Z_{22} \end{Vmatrix} \quad \text{(4)}$$

Finally in terms of the chain matrix (A B C D) of the network N the corresponding reciprocal forward equivalent matrix is

$$\begin{Vmatrix} A & B \\ C & D+(1-AD+BC)/(A+BY_L) \end{Vmatrix} \quad \text{(5)}$$

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The principle of the design procedure used in the following investigation consists of synthesizing the forward equivalent structure with appropriate restrictions, and then retransform the final network to the non-reciprocal form. For this purpose it is convenient to consider canonical structures of the type shown in Fig. 1(a) for the realization of

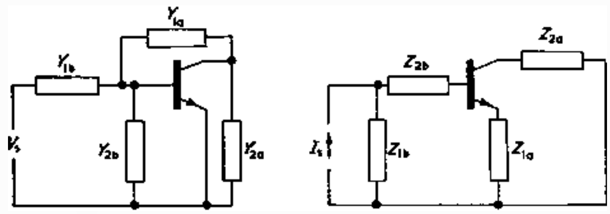

Fig. 1. Canonical structures for the realization of one complex pole pair
(a) Hybrid-π type
(b) Hybrid-T type

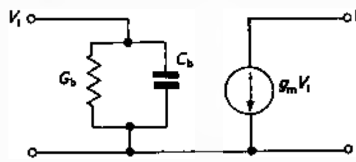

Fig. 2. Simplified common emitter equivalent circuit

transfer voltage or transfer admittance functions of type:

$$T_{21} = \frac{H(s)}{s^2 + bs + c} \quad \text{(6)}$$

Similarly the structure of Fig. 1(b) is used for the realization of transfer current or transfer impedance functions of type (6). Both these canonical structures may be cascaded with the interposition of unilateral buffer stages of common emitter or of common base type, using graded base type transistors, provided the input circuits of these stages are made part of the load elements Y_{2a} and Z_{2a} respectively. For the purpose of the synthesis the equivalent circuit of the transistor used is simplified as shown in Fig. 2, neglecting intrinsic feedback, base spread resistance and collector conductance, which are all presumed to be very small with respect to the extrinsic parameters.

As a result, in general transfer functions of the type:

$$T_{21} = \frac{H(s)}{(s^2 + b_1s + c_1) \ldots (s^2 + b_ns + c_n)} \quad \text{(7)}$$

may be realized within a frequency range depending upon the transistor used.

Synthesis of Canonical Structures

Starting point of the synthesis is a prescribed transfer function of type (6), In particular, if the latter is frequency normalized, the same normalization should be applied to the transistor equivalent circuit, multiplying the value of C_b by the normalization factor.

Considering first the network of Fig. 1(a), the following forward equivalent admittance matrix is obtained:

$$\begin{Vmatrix} Y_{1b} + Y_{2b} + Y_{1a} + Y_b + Y' & -Y_{1a} + g_m \\ -Y_{1a} + g_m & Y_{1a} + Y_{2a} \end{Vmatrix} \quad \ldots\ldots (8)$$

where $Y' = g_m(g_m - Y_{1a})/(Y_{1a} + Y_{2a})$. The corresponding ladder unit is shown in Fig. 3(a), where, indicating with M an impedance normalization factor, one has:

$$Y_{2a}' = (1/M)\,(Y_{2a} + g_m)$$
$$Y_{1a}' = (1/M)\,(Y_{1a} - g_m)$$
$$Y_{2b}' = (1/M)\,(Y_{2b} + G_b + sC_b + g_m Y_{2a}'/(Y_{1a}' + Y_{2a}'))$$
$$Y_{1b}' = (1/M)\,Y_{1b}$$

$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (9)$$

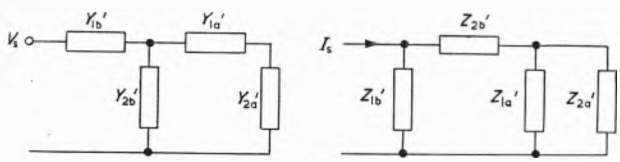

Fig. 3. Reciprocal forward equivalent forms of the canonical networks

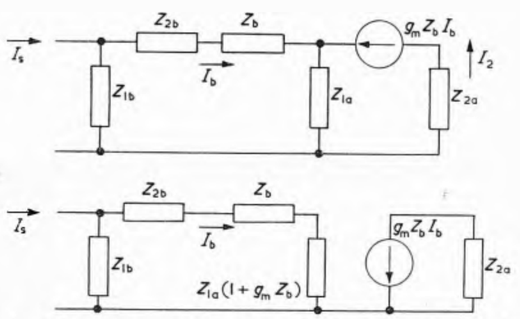

Fig. 4. Hybrid-T type circuit and its unilateral forward equivalent

Similarly, indicating with z_{ije} the common emitter impedance parameters, the following is the reciprocal forward equivalent matrix of the network of Fig. 1(b):

$$\begin{Vmatrix} Z_{1b} + Z_{2b} + z_{11e} + Z_{1a} + Z' \;, & Z_{1a} + z_{21e} \\ Z_{1a} + z_{21e} & Z_{1a} + Z_{2a} + z_{22e} \end{Vmatrix}$$

where $Z' = (z_{21e} - z_{12e})\,(z_{21e} + Z_{1a})/(Z_{1a} + Z_{2a} + z_{22e})$.

$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (10)$$

The corresponding network representation is shown in Fig. 3(b), where:

$$Z_{2a}' = (1/M)\,(Z_{2a} - z_{21e} + z_{22e})$$
$$Z_{1a}' = (1/M)\,(Z_{1a} + z_{21e})$$
$$Z_{2b}' = (1/M)\,(Z_{2b} + z_{11e} + z_{12e} + (z_{21e} - z_{12e})Z_{2a}'/(Z_{1a}' - Z_{2a}'))$$
$$Z_{1b}' = (1/M)\,Z_{1b}$$

$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (11)$$

However, with the assumed simplifications of the transistor parameters, the quantities z_{12e} and z_{22e} become infinite; therefore, equations (11), valid in general, do not apply in the specific case considered here.

Redrawing the circuit of Fig. 1(b) in Fig. 4(a), with consideration of the transistor equivalent circuit, one finds that the output current I_2 is unaffected by the impedance Z_{2a} and is given by the expression

$$I_2 = g_m Z_b Z_{1b} I_1/(Z_{1b} + Z_{2b} + Z_b + Z_{1a}\,(1 + g_m Z_b) \quad \ldots. (12)$$

Hence, as far as the current I_2 is concerned, this may be determined from the network of Fig. 4(b), which is the 'unilateral' forward equivalent of the network of Fig. 4(a).

Furthermore, it is noted that, because the circuit of Fig. 4(a) terminates in a current generator, it can be cascaded directly with similar circuits, without the interposition of buffer stages.

The method of synthesis of the above transformed networks is not unique. Thus, for RC networks with real zeros the Cauer ladder development may be applied[2,3,4]; for voltage transfer functions in general the matrix factorization method of Pantell[5,6] may be used. Still other procedures of synthesis may be followed.

As an example, consider the synthesis of a Butterworth or a Tchebyscheff filter. If the specifications call for a band-pass with prescribed bandwidth $\omega_c = \omega_a - \omega_b$, with centre frequency $\omega_o = \sqrt{(\omega_a \omega_b)}$ and prescribed off band attenuation (in addition to the value of the ripple in the Tchebyscheff case), one can easily derive the corresponding frequency normalized low-pass transfer function[7]. The latter is synthesized and then denormalized to the bandwidth ω_c by dividing inductances and capacitances by ω_c. Finally, applying the low-pass band-pass analogy, the network is converted to the band-pass form adding in series with each inductance L a capacitance $C = 1/\omega_o^2 L$ and in parallel with each capacitance C' an inductance $L' = 1/\omega_o^2 C'$.

Assume, therefore, that a prescribed voltage transfer function is to be realized by means of a ladder section of the type of Fig. 1(a). Considering the corresponding chain matrix (5) of the entire network, the parameter A, which remains unchanged under the forward equivalent network transformation, is the reciprocal of the voltage gain A_v. Following Pantell's procedure, A is written as the product of four functions of s, i.e.,

$$A = (k_1 f_1)\,(k_2 f_2)\,(k_3 f_3)\,(k_4 f_4) \quad \ldots\ldots\ldots (13)$$

where the $f_i\,(s)$ will be further specified later and the k's are positive or negative constants. Factoring the reciprocal forward equivalent matrix (5) in an appropriate way, it is shown that the following relations apply for the ladder elements:

$$Y_{2a}' = k_4 f_4$$
$$1/Y_{1a}' = k_3 f_3 - 1/k_4 f_4$$
$$Y_{2b}' = k_2 f_2 - 1/k_3 f_3$$
$$1/Y_{1b}' = k_1 f_1 - 1/k_2 f_2$$

$$\ldots\ldots\ldots\ldots (14)$$

Further simplifications may be obtained grouping together Y_{1b}' and Y_{2b}' hence using only three $f_i\,(s)$ functions. Using equations (9) and (14) the following relations are derived for the extrinsic circuit elements:

$$Y_{2a} = M\,k_4 f_4 - g_m$$
$$Y_{1a} = M/(k_3 f_3 - 1/k_4 f_4) + g_m$$
$$Y_{1b} + Y_{2b} = M(k_2 f_2 - 1/k_3 f_3) - G_b - sC_b - \cfrac{g_m k_4 f_4}{k_4 f_4 + \cfrac{1}{k_3 f_3 - 1/k_4 f_4}}$$

$$\ldots\ldots\ldots\ldots\ldots (15)$$

From the latter equations it is seen that the $f_i(s)$ may be p.r. (positive real) functions, with the k appropriate positive or negative constants.

For example, if a voltage transfer function:

$$A_v = 1/(s^2 + bs + c)$$

$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (16)$$

is to be synthesized, one may let:

$$A = \left(k_2\,\frac{s^2 + bs + c}{s + z}\right)\left(k_3\,\frac{s + z}{s + z'}\right)[k_4\,(s + z')] \quad \ldots.. (17)$$

with:

$$f_4 = s + z',\; f_3 = (s + z)/(s + z')$$
$$f_2 = (s^2 + bs + c)/(s + z) \quad \ldots\ldots\ldots. (18)$$

There follows:

$$Y_{2a}' = k_4 (s + z') \quad \ldots\ldots\ldots\ldots \quad (19)$$

where k_4 may be taken equal one. Furthermore, if gain larger than one is desired, Y_{1a}' should have a negative real part; letting $k_3 = -1/(z' - z)$, with $z' > z$, there follows that Y_{1a}' is negative real, i.e.,

$$Y_{1a}' = -(z' - z) \quad \ldots\ldots\ldots\ldots \quad (20)$$

Finally one has:

$$Y_{1b}' + Y_{2b}' = k_2 \frac{s^2 + bs + c}{s + z} + z' - z \quad \ldots\ldots \quad (21)$$

In order to determine z, z' and k_2 equations (15) are considered. These are written:

$$Y_{2a} = M (s + z') - g_m$$
$$Y_{1a} = -M (z' - z) + g_m \quad \ldots\ldots\ldots \quad (22)$$
$$Y_{1b} + Y_{2b} = Mk_2 \left(s + b - z + \frac{c - z (b - z)}{s + z} \right) + M (z' - z) +$$
$$+ M \frac{(z' - z)^2}{s + z} - G_b - sC_b - g_m - g_m \frac{(z' - z)}{s + z}$$

Hence, letting $n = z'/z$, the following conditions must be satisfied for $R - C$ realizability:

$$0 < z < b, \; n > 1$$
$$g_m/nz \leqslant M \leqslant g_m/(n - 1) z$$
$$M (k_2 (c - z (b - z)) + (n - 1)^2 z^2) = g_m (n - 1) z$$
$$M (k_2 (b - z) + (n - 1) z) \geqslant G_b + g_m$$
$$Mk_2 \geqslant C_b$$
$$\ldots\ldots\ldots\ldots\ldots\ldots \quad (23)$$

The value of the factor M is:

$$M = \frac{g_m (n - 1) z}{k_2 (c - z (b - z)) + (n - 1)^2 z^2} \quad \ldots\ldots \quad (24)$$

The constant k_2 must satisfy the inequality:

$$\frac{(n - 1)^2 z^2}{(b - z)(1 + nh_{fe}) z - c (1 + h_{fe})} \leqslant k_2 \leqslant \frac{(n - 1) z^2}{c - z (b - z)} \quad \ldots \quad (25)$$

where h_{fe} is the common emitter current amplification factor. If the load conductance is prescribed, the value of k_2 is:

$$k_2 = \frac{(n - 1) z^2}{c - z (b - z)} \left(1 - \frac{nG_{2a}}{G_{2a} + g_m} \right) \quad \ldots\ldots \quad (26)$$

provided the inequality (25) is satisfied. From equation (26) it is seen that the constant n falls in the range:

$$1 < n < 1 + g_m/G_{2a} \quad \ldots\ldots\ldots\ldots \quad (27)$$

The value of G_{2a} has been assumed to be large with respect to the collector shunt conductance.

Substituting equation (26) in equation (24) there follows:

$$M = (1/nz)(g_m + G_{2a}) \quad \ldots\ldots\ldots\ldots \quad (28)$$

Alternatively, the constant k_2 could be obtained in terms of a prescribed source conductance G_{1b}.

The zero frequency gain is:

$$A_{vo} = -\frac{(n - 1) z \, G_{1b}'}{k_2 c} \quad \ldots\ldots\ldots\ldots \quad (29)$$

This is maximum if G_{1b}' is made as large as possible. Assuming that G_{2b} is negligible, one has:

$$G_{1b}'_{(max)} = k_2 \frac{(b - z) nz - c}{(n - 1) z} - G_b/M \quad \ldots\ldots \quad (30)$$

$$A_{vo(max)} = -\frac{(b - z) nz}{c} + 1 + \frac{G_b (n - 1) z}{M k_2 c} \quad \ldots\ldots \quad (31)$$

In particular, if the last term in equation (31) is negligible, $A_{vo(max)}$ is largest when $z = b/2$ and n is as large as possible.

Recapitulating, the canonical network realizing the voltage ratio (16) is of the type shown in Fig. 5. The components of this network are obtained by first choosing appropriate values for z, n, k_2 and then computing M and the following expressions:

$$G_{2a} = Mnz - g_m$$
$$C_{2a} = M$$
$$G_{1a} = g_m - M(n - 1)z$$
$$G_{1b} + G_{2b} = M(k_2(b - z) + (n - 1)z) - G_b - g_m$$
$$C_{2b} = Mk_2 - C_b$$
$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (32)$$

The upper limit of realizable bandwidth with a single given transistor in the present configuration is determined by the value of C_b and corresponds to the condition $Mk_2 = C_b$.

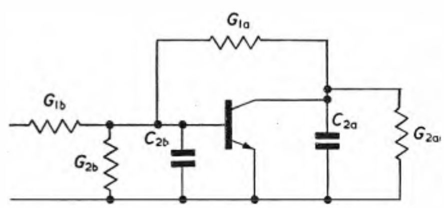

Fig. 5. Final form of the hybrid-π section

The reliability of the design is increased if:

$$G_{1b} + G_{2b} \gg G_b, \quad C_{2b} \gg C_b, \quad M(n - 1)z \ll g_m$$
$$\ldots\ldots\ldots\ldots\ldots\ldots \quad (33)$$

Consider now the case of the synthesis of a current ratio:

$$A_i = P(s)/Q(s) \quad \ldots\ldots\ldots\ldots \quad (34)$$

It has been shown that in the canonical structure of Fig. 4 one has:

$$A_i = g_m Z_b Z_{1b} / (Z_{1b} + Z_{2b} + Z_b + Z_{1a}(1 + g_m Z_b))$$
$$\ldots\ldots\ldots\ldots\ldots\ldots \quad (35)$$

Solving for Z_{1a} there follows:

$$Z_{1a} = \frac{1}{g_m + Y_b} \left(Z_{1b} \left(\frac{g_m Z_b Q(s)}{P(s)} - Y_b \right) - Z_{2b} Y_b - 1 \right.$$
$$\ldots\ldots\ldots\ldots \quad (36)$$

Hence the problem of the synthesis is reduced to determining Z_{1b} and Z_{2b} so that Z_{1a} is passive realizable.

For example; let:

$$P(s) = H(s + \alpha), \quad Q(s) = s^2 + bs + c$$
$$\ldots\ldots\ldots\ldots\ldots\ldots \quad (37)$$

There follows:

$$Z_{1a} =$$
$$\frac{1}{g_m + G_b + sC_b} \frac{Z_{1b}}{s + \alpha} \left[s^2(g_m/H - C_b) + s(g_m b/H - G_b - \alpha C_b) \right.$$
$$\left. + cg_m/H - \alpha G_b \right] - 1 - Z_{2b} (G_b + sC_b)$$
$$\ldots\ldots\ldots\ldots\ldots\ldots \quad (38)$$

From equation (38), letting $H = g_m/C_b$, the following necessary conditions are derived:

$$\alpha < b - G_b/C_b = b - \omega_{\alpha e}$$
$$\alpha < cC_b/G_b = c/\omega_{\alpha e}$$
$$\ldots\ldots\ldots\ldots\ldots \quad (39)$$

where $\omega_{\alpha e}$ is the 3dB down angular frequency of the common emitter current amplification factor.

Letting:

$$Z_{2b} = 0, \quad Z_{1b} = R_{1b} = 1/G_{1b}, \quad H = g_m/C_b$$

$$p = (g_m + G_b)/C_b,$$
$$\beta = [cC_b - \alpha(G_b + G_{1b})]/[(b-\alpha)C_b - G_b - G_{1b}] \quad \text{(40)}$$

(where p is the 3dB down angular frequency of the common base current amplification factor), equation (38) is written:

$$Z_{1a} = \frac{(b-\alpha)C_b - G_b - G_{1b}}{C_b G_{1b}} \cdot \frac{s+\beta}{(s+\alpha)(s+p)} \quad \text{(41)}$$

This is a p.r. function provided:

$$\alpha + p > \beta \quad \text{(42)}$$

In particular, if $\alpha \ll p$, this condition may be written:

$$\alpha < b - G_b/C_b - cC_b/(g_m + G_b) = b - \omega_{\alpha e} - c/p \quad \text{(43)}$$

In this case the conductance G_{1b} is chosen in accordance with the inequality:

$$G_{1b} \leqslant [b - \alpha - cC_b/(g_m + G_b)]\,C_b - G_b \quad \text{(44)}$$

As a particular case, when $\alpha < p$, G_{1b} may be chosen so that $\beta = p$, letting:

$$G_{1b} = [(b - G_b/C_b)p - c)C_b - \alpha g_m]/(p - \alpha) \quad \text{(45)}$$

The low frequency gain is:

$$A_{i0} = g_m\alpha/cC_b \quad \text{(46)}$$

which is proportional to α. Hence, if A_{i0} is to be larger than one, α must be larger than cC_b/g_m.

Recapitulating, the synthesis of the hybrid-T circuit for prescribed current ratio $A_i = (s+\alpha)/(s^2 + bs + c)$ is obtained as follows: Letting $b = \omega_0 b'$, $c = \omega_0^2$ and assuming that the transistor has a cut-off frequency $\omega_{\alpha e}$ and current amplification factor h_{fe}, one verifies that $\alpha \ll p$ and satisfies the following inequalities:

$$\frac{\omega_0}{\omega_{\alpha e} h_{fe}} \leqslant (\alpha/\omega_0) \leqslant (\omega_0/\omega_{\alpha e}) \quad \text{(39a)}$$

$$\frac{\omega_0}{\omega_{\alpha e} h_{fe}} \leqslant (\alpha/\omega_0) \leqslant b' - (\omega_{\alpha e}/\omega_0) - \frac{\omega_0}{\omega_{\alpha e}(1 + h_{fe})} \quad \text{(43a)}$$

A necessary condition for the solution of this problem is that:

$$b' > (\omega_{\alpha e}/\omega_0) + (\omega_0/\omega_{\alpha e})\frac{1 + 2h_{fe}}{h_{fe}(1 + h_{fe})}$$

i.e. for given b' the ratio $\omega_0/\omega_{\alpha e}$ should fall in the range:

$$(b'h_{fe}'/2) - \sqrt{[(b'h_{fe}'/2)^2 - h_{fe}']} \leqslant (\omega_0/\omega_{\alpha e}) \leqslant (b'h_{fe}'/2) + \sqrt{[(b'h_{fe}'/2)^2 - h_{fe}']}$$

where $h_{fe}' = (h_{fe} + 1)h_{fe}/(1 + 2h_{fe})$. In certain cases the latter condition may be satisfied by 'decreasing' artificially the value of $\omega_{\alpha e}$. If α is not negligible with respect to p, the general condition (42), rather than (43) should be satisfied.

Applications

Design applications of the previous synthesis procedure have been made using drift transistors of type 2N384 (RCA). At the bias point $I_e = -1{\cdot}5$mA, $V_{ce} = -6$V this possesses the typical parameter values $G_{ii} = 960\mu$mhos, $C_b = 90$pF, $g_m = 0{\cdot}0568$mhos, $r_{bb}' = 50\Omega$. Furthermore, $\omega_{\alpha e} = 10{\cdot}7 \times 10^6$, $p = 642 \times 10^6$, $f_{\alpha e} = 1{\cdot}58$Mc/s, $h_{fe} = 59$. Actual parameter values may be measured from the transistor performance as a simple resistance loaded amplifier, using the low frequency input conductance, the low frequency voltage gain and the high frequency cut-off.

Design (1)

Design a video amplifier having a fourth order Butterworth voltage gain response, a bandwidth of 100kc/s and the load conductance $G_{2a} = 10^{-3}$mhos.

The transfer voltage ratio is in normalized form:

$$A_v = 1/(s^2 + 0{\cdot}7654s + 1)(s^2 + 1{\cdot}8478s + 1) \quad \text{(47)}$$

where the normalization factor is $2\pi 10^5$. Correspondingly, the capacitance C_b will be normalized to $C_b 2\pi 10^5 = 56{\cdot}5 \times 10^{-6}$F. The entire filter will be obtained with three transistors, two in the configuration of Fig. 1(a) (or 5) and a third one as a common emitter buffer stage.

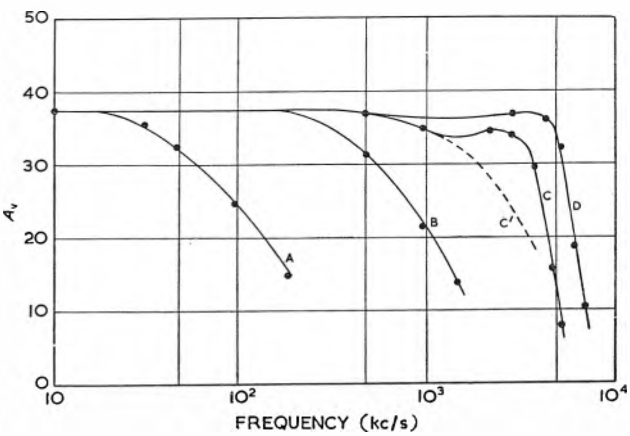

Fig. 6. Voltage gain response of a particular hybrid-π section

A $C_{2a} = 1910pF$, $C_{2b} = 76m\mu F$
B $C_{2a} = 151pF$, $C_{2b} = 7510pF$
C $C_{2a} = 19pF$, $C_{2b} = 70pF$
C' *Theoretical response*
D $C_{2a} = C_{2b} = 0$

Starting from the output side one synthesizes one of the complex pole pairs, for example $s^2 + 0{\cdot}7654s + 1$. Since G_{2a} is prescribed, the value of n must be chosen in the range:

$$1 < n < 1 + g_m/G_{2a} = 57{\cdot}8$$

Letting $z = b/2 = 0{\cdot}3827$ and $n = 50$ one has from equations (26) and (27) respectively:

$$k_2 = 1{\cdot}1346, \; M = 3{\cdot}0206 \times 10^{-3}$$

From these quantities the following normalized parameter values are computed:

$$G_{2a} = 10^{-3}\text{mhos}, \; C_{2a} = 3{\cdot}0206 \times 10^{-3}\text{F}$$
$$G_{1a} = 156 \times 10^{-6}\text{mhos}, \; G_{1b} + G_{2b} = 155{\cdot}3 \times 10^{-6}\text{mhos}$$
$$C_{2b} = 3370\mu\text{F}.$$

Finally, denormalizing to $f = 10^5$c/s, one has:

$$C_{2a} = 4809{\cdot}9\text{pF}, \; C_{2b} = 5367\text{pF}.$$

In addition one may let:

$$1/G_{1b} = 9{\cdot}5\text{k}\Omega, \; 1/G_{2b} = 20\text{k}\Omega$$

and furthermore use a portion of $1/G_{1b}$, for example 1kΩ, as the load of the preceding buffer stage.

The input admittance of the buffer stage is $G_b + sC_b$ and will be incorporated in the admittance Y_{2a} of the preceding section, which synthesizes $s^2 + 1{\cdot}8473s + 1$.

Letting again $G_{2a} = 10^{-3}$mhos and $n = 50$, and taking $z = b/2 = 0{\cdot}9239$, one finds $M = 1{\cdot}2512 \times 10^{-3}$ and obtains the following extrinsic normalized parameters

$$G_{2a} = 10^{-3}\text{mhos},\ C_{2a} = 1194{\cdot}5\mu\text{F},\ G_{1a} = 1{\cdot}7 \times 10^{-4}\text{mhos}$$

$$G_{1b} + G_{2b} = 43{\cdot}408 \times 10^{-3}\text{mhos},\quad C_{2b} = 481{\cdot}78 \times 10^{-4}\text{F}.$$

Denormalizing to $f = 10^5$c/s one has

$$C_{2a} = 1910\text{pF},\ C_{2b} = 76 \times 10^{-9}\text{F}$$

Finally the design may be completed letting

$$1/G_{1b} = 25\Omega,\ 1/G_{2b} = 300\Omega.$$

Assume now that the bandwidth of the above video amplifier is to be 1Mc/s or 10Mc/s, instead of 100kc/s. In this case a common base buffer stage might be used, to obtain a lower value of C_{2a}/G_{2a} ratio. The output section of the filter may remain unchanged, except for the frequency normalization factor, and the input section should be redesigned on the basis of values of $C_{2a} = M \geqslant C_b$ and $G_{2a} \geqslant G_b + g_m$.

In Fig. 6 are summarized the results of experimental investigations of the voltage ratio of the filter section (49). Good agreement was found with the theoretical response for cut-off frequencies of 100kc/s and 1Mc/s. At 10Mc/s the response follows only in part the theoretical curve and shows a peaking effect at 3Mc/s. In the same figure the response obtained removing completely the extrinsic capacitances is also shown.

Design (2)

Design a Butterworth second order filter of band-pass type, with centre frequency 100kc/s bandwidth 10kc/s, load conductance 10^{-3}mhos. For the low-pass analogue the expression of the voltage ratio is:

$$A_v = 1/(s^2 + \sqrt{2}s + 1)$$

with a frequency normalization factor of $2\pi 10^4$. The design proceeds substantially as before; however, an effort is made to obtain capacitors of reasonable value, since these must be resonated with parallel inductances at the centre frequency of the band-pass filter. Since the values of C_{2a} and C_{2b} are proportional to M, the latter should be made sufficiently small. From equation (28) it is seen that the minimum value of M occurs when n and z are both maxima, i.e., n is close to $1 + g_m/G_{2a}$ and z close to b. On the other hand, from equation (31) it is seen that the gain varies directly with $nz\,(b \quad z)$.

If one selects $n = 50$, $z = 1$, there follows

$$M = 1{\cdot}156 \times 10^{-3},\ k = 11{\cdot}2,\ A_{vo} = 15{\cdot}7$$

The normalized low pass parameters are

$$G_{2a} = 10^{-3}\text{mhos},\ C_{2a} = 1{\cdot}156 \times 10^{-3}\text{F},\ G_{1a} = 0{\cdot}0558\text{mhos}$$

$$G_{1b} + G_{2b} = 0{\cdot}0044\text{mhos},\ C_{2b} = 12{\cdot}95 \times 10^{-3}\text{F}$$

Denormalizing to $f = 10^4$c/s one has

$$C_{2a} = 18{\cdot}5 \times 10^{-9}\text{F},\ C_{2b} = 0{\cdot}206\mu\text{F}$$

Finally, returning to the band-pass analogue at the centre frequency of 100kc/s one has

$$L_{2a} = 135 \times 10^{-6}\text{H},\ L_{2b} = 12{\cdot}1 \times 10^{-6}\text{H}$$

Design (3)

Design a filter with a transfer impedance function of second order Butterworth, low-pass, bandwidth 1Mc/s, load conductance 10^{-3}mhos.

One has in normalized form

$$Z_{21} = V_2/I_2 = 1/(s^2 + \sqrt{2}s + 1)$$

with frequency normalization factor $2\pi 10^6$. Since $V_2/I_1 = Z_{2a} I_2/I_1$, one may let $Z_{2a} = 1/H(s + \alpha)$ and

$$I_2/I_1 = H(s + \alpha)/(s^2 + \sqrt{2}s + 1)$$

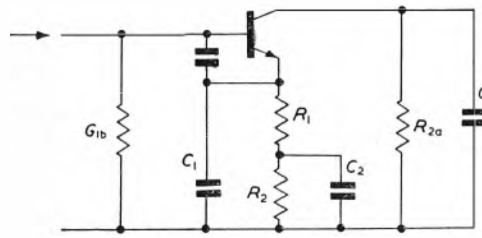

Fig. 7. Example of synthesis of hybrid-T section

In order to satisfy the conditions (39) and (43) with $\alpha < cC_b/g_m$ it is necessary to increase the value of C_b. For example, if C_b is doubled, the range of α is determined by the conditions

$$0{\cdot}0199 < \alpha < 1{\cdot}175,\ \alpha < 0{\cdot}5054$$

where $\alpha \ll p$. Choosing $\alpha = 0{\cdot}1$ and $H = 50$, equation (45) gives $G_{1b} = 4{\cdot}26 \times 10^{-4}$mhos; furthermore, from equation (44) one has

$$G_{1b} < 500 \times 10^{-6}\text{mhos}$$

For generality one selects $G_{1b} = 10^{-4}$mhos and obtains:

$$\beta = 2{\cdot}42,\ p = 51,\ Z_{1a} = 3750\,\frac{s + 2{\cdot}42}{(s + 0{\cdot}1)(s + 51)}$$

The synthesis of the latter impedance is shown in Fig. 7 where the normalized values of the components are

$$C_1 = 2{\cdot}67 \times 10^{-4}\text{F},\ R_1 = 77\Omega,\ C_2 = 5{\cdot}62 \times 10^{-3}\text{F},\ R_2 = 1700\Omega$$

The corresponding denormalized values of C_1 and C_2 are respectively $C_1 = 42{\cdot}6$pF, $C_2 = 895$pF.

Finally the load impedance is synthesized as

$$G_{2a} = 1/H'(s + 0{\cdot}1)$$

where $H' = G_{2a}/0{\cdot}1 = 10^{-2}$. Denormalizing to $f = 1$Mc/s one has $G_{2a} = 10^{-3}$mhos, $C_{2a} = 1590$pF.

Conclusions

Procedures of design for high frequency transistor filters have been developed, using the non-reciprocal-to-reciprocal and the non-reciprocal-to-unilateral forward equivalent network transformation. Two types of canonical structures, respectively a hybrid-π and a hybrid-T type have been considered for the synthesis of each complex pole pair. The hybrid-π networks are cascaded with the use of unilateral buffer stages, obtained using graded base transistors; the hybrid-T networks may be cascaded directly.

The investigation has not dealt in detail with the problems of reliability and of temperature stability, which constitute important but separate aspects of the active networks behaviour. In general the design affords a certain freedom of choice of the fundamental parameters, which may be used to satisfy particular reliability requirements of the networks.

Acknowledgment

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Variation of Input Conductance of a Grounded Base Junction Transistor

By S. Deb* and J. K. Sen*

This article investigates the problem of variation of input conductance of a grounded base junction transistor with the input d.c. current. Methods are suggested for maintaining this variation linear over a wide range of values of current and also for reducing the thermal drift. Performances of an amplitude modulator and an analogue multiplier using a transistor incorporating these features are described.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 780)

IT is known that to a first approximation the input resistance, r_{in} of a grounded base junction transistor amplifier varies inversely as the first power of the emitter bias current I_e. Several interesting uses may be made of this variation. Thus Chaplin and Owens[1] have proposed a novel type of d.c. amplifier using this phenomenon. These authors have also suggested its use for constructing the basic element of an analogue multiplier. The simple law of variation of r_{in} with I_e as stated above is, however, only valid over a limited range of operation, thus restricting somewhat the practicability of these and other suggestions. This limitation is discussed in the present article and it is shown that it can be overcome to a large extent by the use of a simple compensating element. The problem of thermal drift is also discussed and methods are suggested for its compensation. The utility of a transistor thus compensated is illustrated by an account of the experimental results obtained for two typical applications as (a) an amplitude modulator and (b) an analogue multiplier.

Variation of Input Conductance with Input Current

Basic Equations

The input conductance of a junction transistor in the grounded base mode is given by[2]:

$$g_{in} = (1/r_{in}) = \frac{1}{r_e + r_b(1-\alpha)} \quad \ldots\ldots\ldots (1)$$

where α is the short-circuit current amplification factor, r_b the base, and r_e the emitter resistance.

Now:

$$r_e = \frac{kT}{q(I_{in} + I_o)} \quad \ldots\ldots\ldots (2)$$

I_{in} being the input d.c. current and $I_o \ll I_{in}$. If $\alpha \sim 1$,

$$r_{in} = (kT/qI_{in}) \quad \ldots\ldots\ldots (3)$$

If now a small alternating voltage v_e is applied between the emitter and the base from a constant voltage source and the current I_{in} is parallel fed from a constant current source then the input driving current

$$i_e = (v_e/r_{in}) = Kv_eI_{in}$$

where $K = (q/kT)$ and the output a.c. current

$$i_o = K\alpha v_eI_{in} \quad \ldots\ldots\ldots (4)$$

Equation (4) is a very interesting relation which may be utilized for a variety of purposes. Its utility is, however, limited by the finite value of the term $r_b(1-\alpha)$ and the dependence on operating temperature and these will now be discussed.

Effect of Base Resistance

For high values of I_{in} the approximation $r_e \gg r_b(1-\alpha)$ is no longer admissible and the output tends to fall below that expected from equation (4). A linear $i_o - I_{in}$ relationship may, however, be obtained over a reasonably good range by deliberately decreasing the response for low values

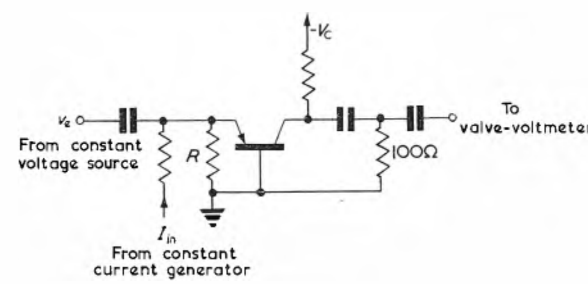

Fig. 1. Grounded base amplifier with resistive compensating element at the input

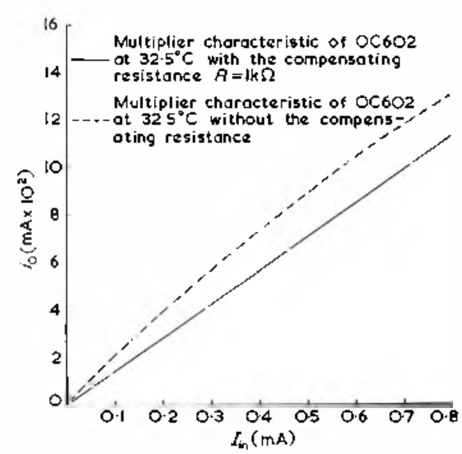

Fig. 2. Variation of the output signal current with the input d.c. current with and without a compensating resistance for transistor type OC602 at 32·5°C in the grounded base mode

of I_{in}. This is easily done by introducing a compensating resistance R at the input as shown in Fig. 1. For this case $I_{in} \neq I_e$ and:

$$i_o = Bv_eI_{in} \quad \ldots\ldots\ldots (5)$$

where

$$B = \frac{\alpha R}{[(kT/q) + r_bI_e(1-\alpha)]\,[(kT/qI_e) + r_b(1-\alpha) + R]} \quad \ldots (6)$$

I_{in} being the total current given by the sum of the junction current I_e and the current flowing through R. The problem is to maintain the value of B constant for the desired range of values of I_{in}. As both α and r_b tend to vary with I_e making B a complicated function of I_{in} a graphical method of solution of the problem is of advantage. This method

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showed that for an OC602 transistor the best value of R for constancy of B is about 1kΩ. In Fig. 2 one i_o/I_{in} characteristic has been drawn for this transistor with and without the compensating element R illustrating the improvement in the linear nature of the characteristic in the former case.

Effect of Ambient Temperature

Neglecting the term containing r_b and making use of equation (2) for r_e the output would be found to depend on temperature through the temperature sensitive factors K, I_o and α. With a compensating resistance R, as shown in Fig. 1, however:

$$I_e = \frac{I_{in}R}{R + \dfrac{kT}{q(I_e + I_o)}}$$

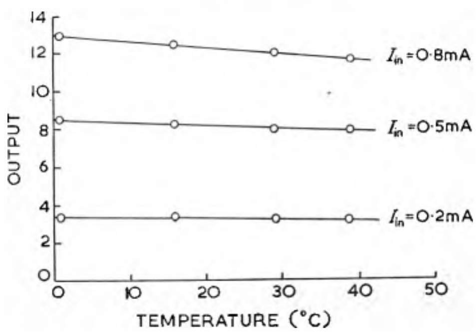

Fig. 3. Variation of the output with the temperature for different input currents without compensating elements for transistor type OC602 (v_e = 5mV)

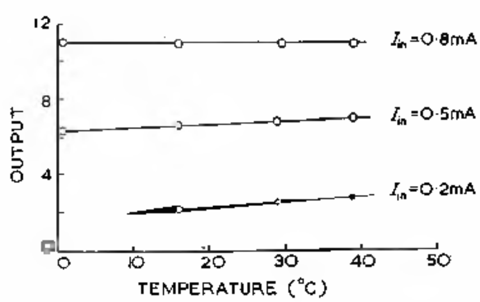

Fig. 4. Variation of output with temperature for different input current with the compensating element present for transistor type OC602 (v_e = 5mV)

whence:

$$I_e = \tfrac{1}{2}\left[(I_{in} - (kT/qR) - I_o) \pm \sqrt{\left\{(I_{in} - (kT/qR) - I_o)^2 + 4I_{in}I_o\right\}}\right] \quad \ldots\ldots\ldots\ldots (7)$$

so that

$$i_o = \frac{K\alpha v_e I_{in}}{2}\left[(1 - (kT/qRI_{in}) + (I_o/I_{in})) \pm \sqrt{\left\{(1 - (kT/qRI_{in}) - (I_o/I_{in}))^2 + 4(I_o/I_{in})\right\}}\right] \quad \ldots . (8)$$

With no compensating element, $R = \infty$ and:

$$i_o = v_e I_{in}\alpha\,(q/kT)\,[1 + (I_o/I_{in})] \quad \ldots\ldots\ldots . (9)$$

It follows from equation (9) that since α varies relatively slowly with temperature i_o should fall with rise in temperature when I_{in} is large. This is borne out by Fig. 3 in which the output has been plotted as a function of temperature for three values of I_{in} for transistor type OC602. When a compensating element is present one obtains from equation (8):

$$i_o = v_e I_{in}\alpha\,(q/kT)\,[1 - (kT/qRI_{in}) + (I_o/I_{in})] \quad \ldots . (10)$$

provided kT/qRI_{in} and I_o/I_{in} are both less than unity. Equation (10) shows that for I_{in} large the output would fall with rising temperature. For I_{in} low, however, the variation of I_o/I_{in} would be the controlling factor and the output should rise with rise in temperature. This also is borne out by Fig. 4 in which the output has been plotted as a function of temperature for three values of I_{in} for transistor OC602.

The simplest way to minimize the variation shown in Fig. 4 would be to connect a negative temperature coefficient resistor of the right value of temperature coefficient at the collector end and to take the potential drop across this as the output. For the OC602 transistor the compensation with this method was found to be very good within the range 22°C to 35°C for I_{in} values between 0·1 to 0·5mA. The drift may also be reduced by using a second grounded base amplifier stage following the first. As is shown from Fig. 3, for an uncompensated stage, the output tends to fall with increase in temperature; the rate of fall being a function

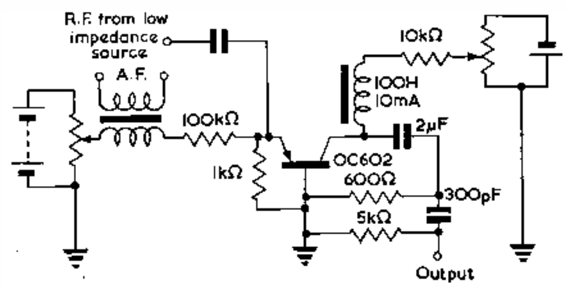

Fig. 5. Transistor amplitude modulator

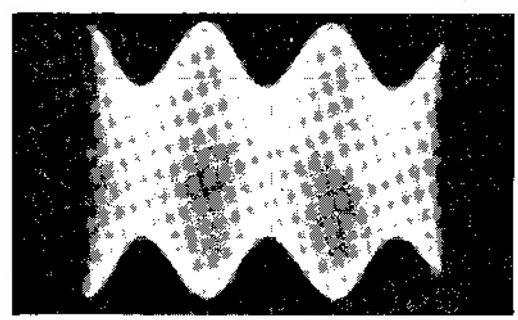

Fig. 6. Oscillogram of the modulated output obtained with the circuit shown in Fig. 5

of I_{in}. The rising tendency of the output from the first stage may thus be reduced by properly selecting the bias for this second stage.

Some Practical Applications

The fact that the a.c. output of a grounded base transistor amplifier may be made to vary linearly with the direct current input I_{in} and the a.c. signal input voltage v_e over a fairly wide range of operating conditions may be utilized to construct a number of useful devices. Its possible applications as a d.c. amplifier and an analogue multiplier have already been mentioned. Another possibility is its use as an amplitude modulator. Experiments carried out on the use as a modulator and as a multiplier gave very encouraging results and are briefly described below.

Amplitude Modulator

A simple modulator circuit using transistor type OC602 is shown in Fig. 5. The modulating voltage was connected in series with the source supplying I_{in} and the frequency of the signal v_e was set at the carrier frequency. A typical modulated wave obtained with a carrier frequency of 100kc/s and a modulating frequency of 400c/s, is shown in Fig. 6. Measurement with a distortion analyser gave the results as shown in Fig. 7, in which the measured values of percentage distortion are given as a function of percentage modulation for two values of I_{in}. The results indicate that the modulation is sensibly distortionless even for large percentages of modulation. The thermal drift could be reduced by connecting a negative temperature coefficient resistor at the output and confining the operation within the range $0{\cdot}1 < I_{in} < 0{\cdot}5$mA. It is to be noted that with transistor type OC602 operation at higher values of carrier frequency

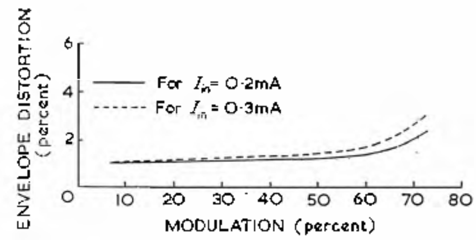

Fig. 7. Variation of percentage distortion of the envelope with percentage modulation for different input currents

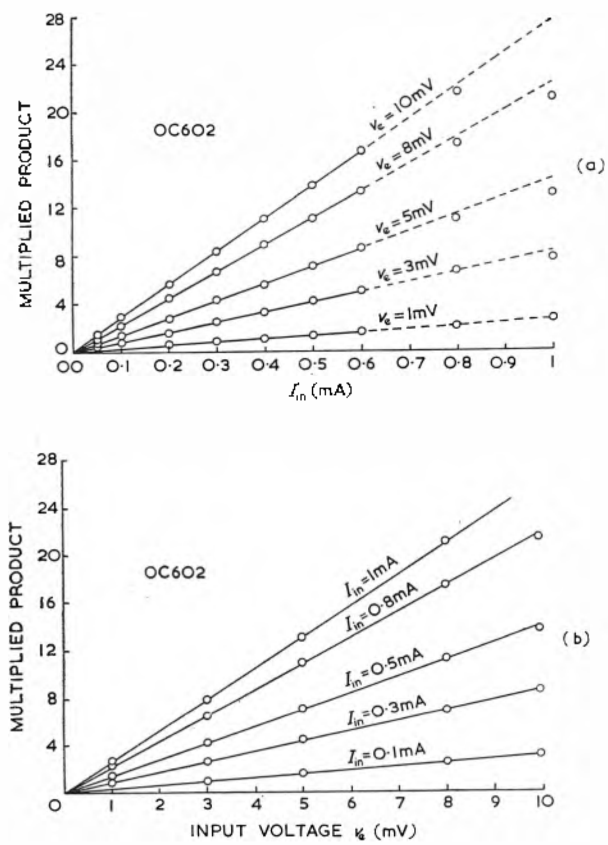

Fig. 8. Multiplier characteristics of OC602 compensated by 1kΩ at 32·5°C
(a) Output/input d.c. bias with the input a.c. voltage as parameter
(b) Output/input a.c. voltage with the input d.c. current as parameter

was not possible. Nevertheless, it establishes the possibility of the device and, with transistors having high cut-off frequencies, modulators employing this principle could be designed to operate at much higher frequencies.

Analogue Multiplier

The basic element of a multiplier circuit using type OC602 transistor is shown in Fig. 1. A set of characteristics giving the output as a function of I_{in} with v_e as a parameter is shown in Fig. 8(a). The characteristics giving the output as function of v_e with I_{in} as parameter are also shown in Fig. 8(b). The characteristics show that with resistive compensation the device gives satisfactory operation over a wide range of values of I_{in} and v_e. From a practical point of view, however, it is important that it also stands the test of thermal drift, frequency response and of long-term drift, if any. In testing these points an accuracy of response within 5 per cent was adopted as being tolerable.

With regard to drift due to variation in ambient temperature the arrangement consisting of a negative temperature coefficient resistor at the output was found to be quite satisfactory.

Curves giving the relative response as a function of frequency with I_{in} as parameter are shown in Fig. 9. These curves show that in respect of the variable v_e the response does not fall below 95 per cent of the low frequency value up to a frequency of 20kc/s even at high values of I_{in}. Experiments carried out with a modulator circuit as described above showed that in respect of the variable I_{in} the response is satisfactory up to about this value of frequency. The bandwidth of the multiplier element should, therefore, be about 20kc/s. The figure would obviously be higher if, as is usual, the bandwidth is measured up to the frequency at which the response is 3dB below the low frequency value. The position may be further improved by using transistors with higher values of cut-off frequency.

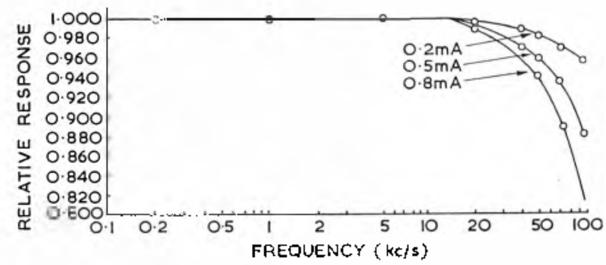

Fig. 9. Frequency response characteristics for different input currents of a multiplier using transistor type OC602 at 32·5°C

It is thus seen that with regard to bandwidth the device shown in Fig. 1 would be superior to many of the existing types of multiplier.

As regards long-term stability it is to be noted that observations lasting over 6 hours did not show any measurable variation of response.

It may be pointed out that a great advantage of the multiplier element described above is that the multiplied output appears as an a.c. signal which enables one to avoid the use of d.c. amplifiers for subsequent amplification.

Conclusion

The linear variation of the input conductance with the direct current input of a junction transistor amplifier in the grounded base mode of operation may be improved considerably by using a resistive compensating element at the input. The thermal drift can be minimized by simple methods. A transistor thus compensated may be utilized to construct an amplitude modulator giving faithful modulation envelope and an analogue multiplier having a very good frequency response.

Acknowledgments

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A High-Speed Ferrite Storage System

C. J. Quartly*, M.A.

A square loop ferrite storage system is described which overcomes the limitation on speed imposed by the core material when used in a conventional store using coincident drive selection. Each digit is stored as a difference in flux levels in two cores which are only partially switched during the writing operation. Measurements made on a pair of cores show how the output during reading varies with write pulse duration and amplitude and with digit pulse amplitude. The use of this principle with external word selection enables a read/write cycle time of 0·5μsec to be achieved in a small store.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 780)

AT the present time practically all high speed random access stores used in recently developed computers are matrix stores using small toroids of square loop ferrites as storage elements. With conventional methods of selection and driving[1,2] these have proved very successful from the point of view of reliability, ease of access and power requirements. However, these methods suffer from the limitation that the fastest cycle time achievable in a practical machine is about 5μsec depending on the system used and the size of the store. This speed has proved adequate for 1 to 3Mc/s serial logic using valves or transistors. Some problems require a computer working much faster than this and an increase in speed by a factor of, say 10, in both the logic and the store would be very desirable. The faster logic may now be achieved with the aid of newer types of transistor so that there is a requirement for a store with a cycle time of 0·5 to 1μsec. Various storage methods have been proposed for obtaining this speed including square loop ferrites[3], thin magnetic films[4,5,6], metal tape cores[7] and low temperature devices[8,9].

In view of the fact that square loop ferrite cores are readily available, whereas the materials needed for the other methods still require further development, the former are more likely to provide a working store in a short time, assuming a suitable system can be found. From preliminary investigations it appears that one particular system has many advantages. This is described below together with the results of measurements already carried out.

Basic System

The system described here is similar to one described by Rajchman[10] which was used for easing the output discrimination problem when using ferrite plates at normal speeds.

The present system differs in that short high current pulses are used which only partially switch the cores during the write process.

A word access system is used. For each digit stored two cores are arranged as in Fig. 1. They are both supplied by a digit pulse, at the same time as the write pulse. One core is arranged to experience the effect of the sum and the other core the difference of these two pulses. The digit pulse is of one polarity if a ' 1 ' is to be written and of the opposite polarity for a ' 0 '. The operation depends on the fact that the core in which the digit pulse adds to the write pulse receives a larger drive than the other, in which the two pulses are in opposition, and will consequently switch further during the writing period. When reading takes place both cores are returned to point P and the output of the core which has been switched further will be greater than the output from the other. Since the core outputs are in opposition on the output wire the polarity of the combined output indicates which digit has been stored.

The apparent advantages of this system are as follows:

(1) Heating

In this system self heating of the cores is not a serious problem with read-write cycle times down to 0·5μsec.

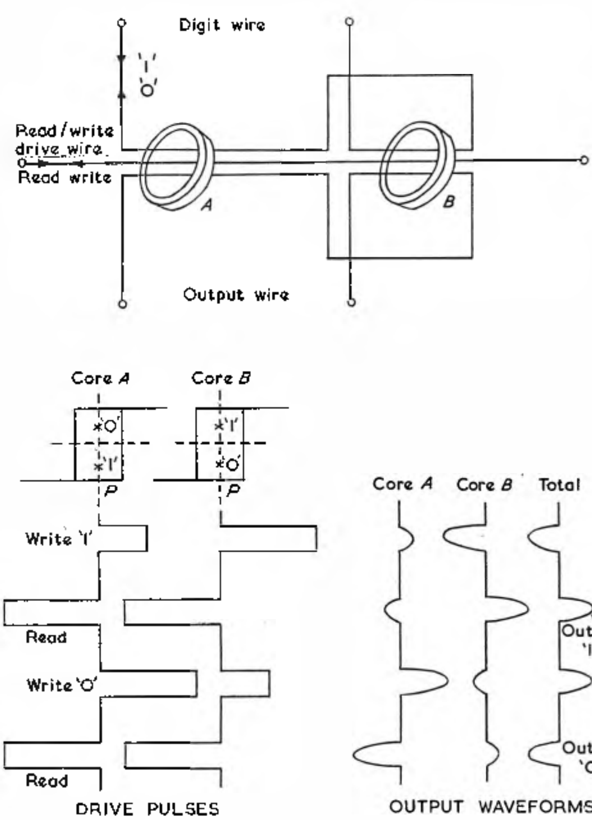

Fig. 1. Arrangement of cores and resultant waveforms

Firstly, since the cores are only partially switched, the amount of heat generated is less. Secondly, and more important, although the quantity of flux stored may change with temperature, the difference in flux between the two cores is used to determine the digit stored and this difference will not change sign due to temperature changes.

(2) Plane Wiring

The wiring of the storage planes is simple in principle, (Fig. 2) since it only requires co-ordinate wires and does not involve diagonal wires or small coupling loops. In practice it may be a problem to reduce the direct pick up between the drive and output wires to an acceptable level.

* Mullard Ltd.

(3) DISCRIMINATION

The discrimination between a '1' and an '0' should be much easier than in some systems since the outputs are of opposite polarities.

(4) DRIVE CURRENTS

None of the current drives is critical in amplitude. The read drive pulse must not be smaller than the sum of the write drive and digit pulses to ensure that both storage cores are always returned to saturation before writing. The '1' and '0' digit pulses should be similar in amplitude.

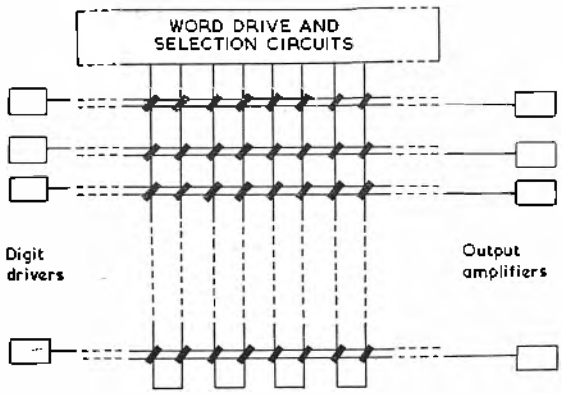

Fig. 2. Wiring of storage planes

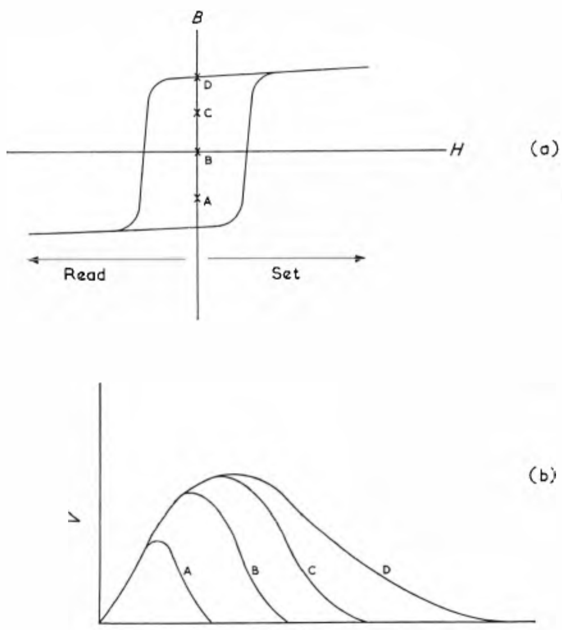

Fig. 3. Effect of remanance

(5) DRIVE PULSE GENERATION

The impedance presented by the cores to the drive pulse generators is the same regardless of the information stored. This could be an advantage when considering the design of these generators.

Core Properties

In this system two core effects are met which do not arise in conventional matrix stores used at lower speeds. Both of these are advantageous though not essential for the basic storage principle as described above.

The first concerns the switching time of cores when they are driven into saturation from various values of remanent magnetization. Fig. 3(b) shows the outputs obtained by applying a constant amplitude read pulse to a core which has been previously set to remanent states A to D in Fig. 3(a). It can be seen that considerably lower switching times are obtained from a partially switched core than from one which is switched from one saturation state to the other, although the exact switching time from a given state depends to some extent on the setting pulse used.

The second property is concerned with the viscosity effect which has also been found in metal tape cores[7]. Pulses very much larger in amplitude than the d.c. coercive force may be repeatedly applied to a saturated core without causing any appreciable irreversible change of flux, provided that the pulses are shorter than a certain length

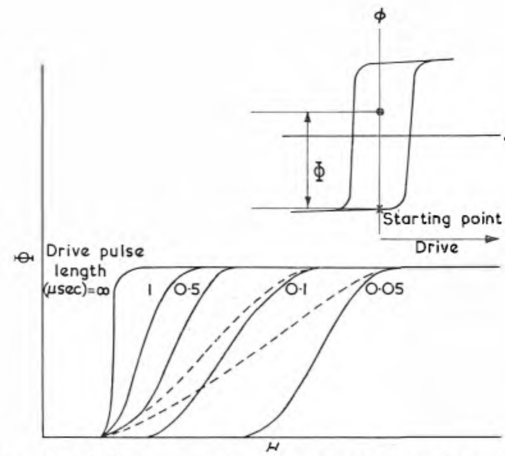

Fig. 4. Idealized curves of switched flux against driving field

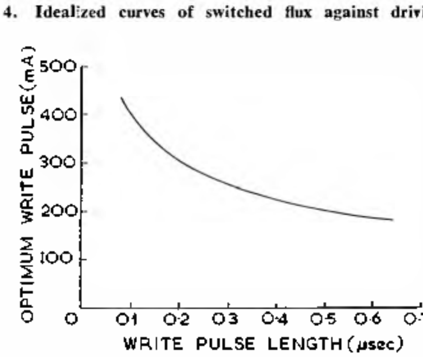

Fig. 5. Optimum write pulse amplitude against pulse length

which varies inversely as the amplitude. The idealized curves of Fig. 4 show the amount of switched flux plotted against driving field for various drive pulse lengths. It can be seen that the viscosity effect which is only significant for drive pulses shorter than 0·1μsec, leads to a higher rate of change of Φ with H than would result if it did not exist as indicated by the dotted lines. This means that even with a fast write pulse of large amplitude the two cores can be set to significantly different states of flux without using a large digit pulse.

Measurements

Initial measurements have been made on this system using a pair of 0·050in o.d. cores, type FX2140, which have a d.c. coercive force equivalent to a current drive of slightly less than 0·1At.

The variation of optimum write pulse amplitude, i.e. the amplitude which provides maximum read output, with write pulse length is shown in Fig. 5 and the effect of different values of digit current is shown in Fig. 6 for write pulses of length 0·1 and 0·5μsec. As expected, the output increases as the digit pulse increases but since the

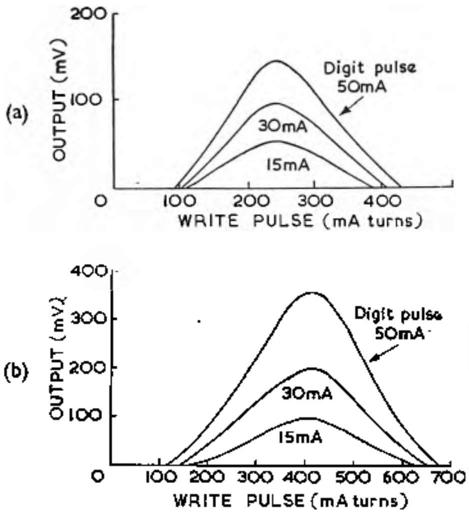

Fig. 6. Output against write and digit currents for write pulse lengths of 0·5 and 0·1μsec
(a) Read pulse 0·3At
(b) Read pulse 0·5At

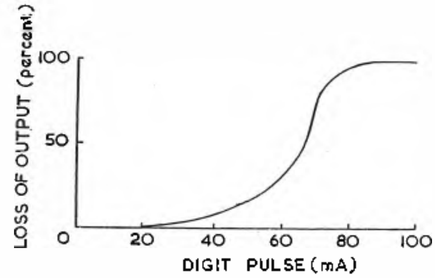

Fig. 7. Loss of output against disturbing pulse amplitude

digit pulse passes through other words its maximum amplitude is limited by the disturbing effect on the partially switched cores in these words. This disturbing effect is rather worse when the disturbing pulse is in the opposite direction to that used when the information was written and Fig. 7 shows the percentage loss of output after a large number of disturbing pulses plotted against the disturbing pulse amplitude.

It can be seen that digit pulses of 30 to 40mA can be tolerated and will provide an adequate output.

In all the above measurements a rectangular drive pulse with a fast rise-time was used but the system will work well with half sine wave pulses if required.

The drive pulses needed for a storage cycle time of 2μsecs are well within the capabilities of currently available transistors and there seems to be no reason why a store of this type should not be fully transistorized.

Conclusions

For initial measurements the system described above shows promise for achieving store cycle times from 0·5 to 2μsecs and may well form an economical solution for filling the gap between the speeds available from conventional ferrite stores and those using other principles such as super-conductivity.

Acknowledgments

The author wishes to thank the Directors of Mullard Research Limited for permission to publish this article.

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Semiconductor Production

The semiconductor factory of the General Electric Co. Ltd. at Hazel Grove, Stockport, Cheshire, is now in full production and is producing some 70 000 transistors and diodes per week while a further 50 000 diodes per week are being produced at the G.E.C. Radio Works, Coventry.

The factory at Hazel Grove first became available in 1956 and a feature of the production there is the high average yields that are obtained; for the factory as a whole the average yield is between 70 and 80 per cent, while for one device it is as high as 90 per cent. Unlike some semiconductor manufacturers, the G.E.C. has not considered it necessary, or desirable, to air-condition or maintain super-cleanliness throughout the factory: rather they have aimed at a general air of cleanliness throughout the factory with small super-clean areas and the fairly extensive use of glove boxes.

Although few of the actual manufacturing processes employed could be described as fully automatic, wherever possible jigs are employed and in a number of cases these are made by watchmaking techniques to the standards of that industry.

The very wide range of semiconductors produced includes low power audio frequency, radio frequency and power transistors; low power silicon junction diodes; and medium power silicon junction, diffused junction silicon, and germanium junction rectifiers—over 100 types in all—at Hazel Grove. In addition, about 30 types of radar mixers and point contact signal diodes are manufactured at Coventry.

Among the novel techniques which are employed is a glassing centrifuge which is used to ensure the maximum efficiency of the glass-to-metal sealing process. The centrifuge, which was developed at the Company's Research Laboratories at Wembley, consists of a rotating annular shaped boat and fitted lid which is enclosed in a firebrick enclosure. The boat, which is flushed with an inert gas, is heated by a bank of gas jets located in the enclosure. The method of operation, once the components have been assembled and jigged and placed radially in the boat, is completely automatic.

The cycle is started by setting the speed of rotation to the required level. The brick lid automatically descends to enclose the rotating boat and triggers the gas/air supply which heats the boat and contents to a preset temperature. The temperature is controlled for a given period after which the gas is turned off and forced air cooling under controlled conditions affects the annealing of the seals which are still in the inert atmosphere. A further period of uncontrolled cooling reduces the boat and contents to handling temperature.

The final hermetic sealing of the transistor envelopes is carried out by cold welding. The application of the process to transistors was a G.E.C. invention and has since been adopted by many other, including some American, companies. The surfaces of the metals to be joined must be cleaned before compression between dies to give a welded joint. The surface films remaining after cleaning are dispersed by the controlled lateral flow of the mating surfaces and contact under pressure is obtained between perfectly clean metals. In this way a true homogeneous weld is obtained without the application of heat.

In a cold welded joint the metal is truly continuous across the weld. The electrical and thermal conductivities of the joint are therefore equal to those of the parent metal. Cold welded joints are thus particularly suitable for the hermetic sealing of semiconductors, the absence of heat, solder, flux vapour, weld flash or gas flames, ensuring that there is no damage to, or contamination of, the device during the sealing process. This is a basic requirement for long life stability.

A new method of 'crystal dicing' has been developed. This is known as diamond scribing and it is claimed to be twelve times faster than the saw-cut technique it is replacing.

All-Metal Bakeable Taps for High Vacuum

By N. W. Robinson,* Ph.D., A.R.C.S.

In an increasing number of microwave and image intensifier tubes it is becoming necessary to create vacuums of the order of 10^{-10}mm Hg. This may be accomplished by a system described by Alpert: in this system a greaseless isolating tap is a necessary component. In this article a suitable all-metal bakeable tap is described and its performance enumerated.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 780)

THE problem of attaining vacuum in a system better than that obtained from diffusion pumps was shown by Alpert[1], to depend upon being able to isolate the system by means of a greaseless tap and then continue to pump with an ionization gauge. Pressures of $<10^{-10}$mm Hg have been obtained in this way and undoubtedly pressures of this order will become commonplace in pumping tubes in which the presence of traces of gas has a bad effect on noise (as in microwave and transmitter tubes) or on 'background', (as in image intensifier tubes).

Alpert described a simple metal tap in which the seal was made by ramming a hard conical plunger into a circular orifice in soft copper, the plunger thus making its own seating. The method has been widely copied with varying degrees of success, and this article describes taps which have been used satisfactorily in The Mullard Research Laboratories for many years.

Fig. 1 shows two taps almost identical with those described by Alpert. That on the right-hand side (a) is shown without the actuating screw and with a bridge piece to hold the tap open while baking. The actuating screw (b) is shown in front of the picture. The second tap (c) shows a production model with the actuating mechanism in position. An assembly drawing showing the construction of the tap is shown in Fig. 2.

Baking

All the materials used in the tap can be baked to approximately 600°C but in practice the baking temperature is limited to 450°C, which is the maximum temperature the glass will withstand. In some circumstances it is impossible to bake the tap along with the remainder of the glass work and an electric soldering iron element is therefore attached to the back of the support, as shown in Fig. 1. With this element it is possible to raise the temperature of the main body to 300°C. If the tap is lagged with heat insulating material the temperature will be substantially higher. During baking the tap must be kept open by screwing the plunger up to a bridge piece which spans the body of the tap. After baking, any scale formed on the screw threads must be removed by a wire brush to prevent the screws binding in use.

Tubulation

As shown in Fig. 1, the input and output tubulations, which are ¼in diameter, may be arranged at any position

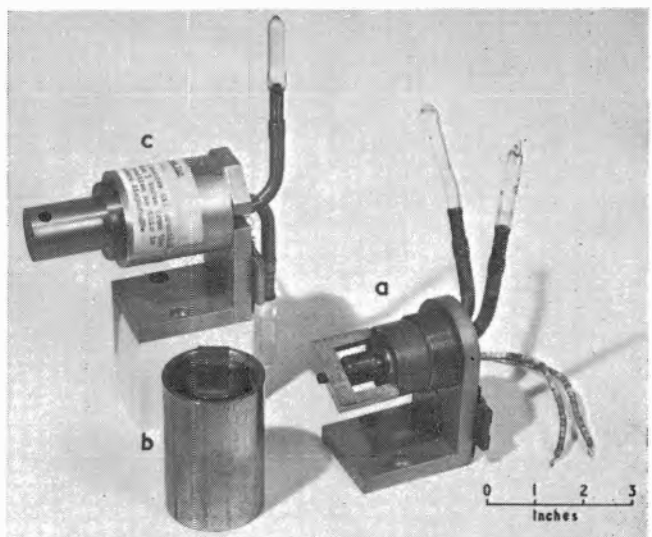

Fig. 1. Two taps similar to those described by Alpert
(a) Without actuating screw
(b) The actuating screw
(c) A production model with the actuating screw in position

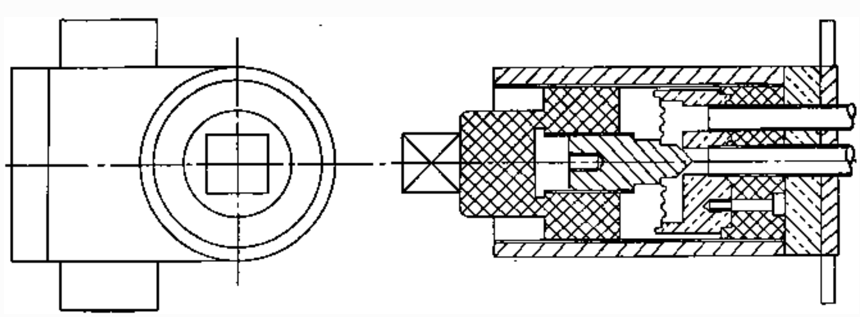

Fig. 2. The construction of the tap

relative to the main body, but that shown on the right-hand tap is considered the most generally useful. Since the tap is normally mounted rigidly on a bench, the glass-work attached must have sufficient flexibility to take up strains due to baking. This is ensured by incorporating loops or U bends in the connecting tubing. With the tubulations as shown it is almost impossible to incorporate the tap in a glass system without employing U bends.

* *Mullard Research Laboratories.*

Operation

The force on the piston is exerted by means of a differential screw operating between the plunger and the barrel. A torque of about 30lb ft is required, which can be applied by a man using a spanner about 8in long. Above a certain limit there is no advantage to be gained by applying increased torque.

The tap can also be operated by an electric motor coupled to the differential screw, so that it may be opened and closed by push-button control.

Tests

The taps have been tested using the arrangement shown in Fig. 3. One tap is used as a manifold tap to isolate the section to the left of the tap from the diffusion pump. The section between the test tap and the manifold tap encloses a known volume, including an ionization gauge. The ionization gauge is run at very small electron current to minimize pumping. The system is baked to 450°C for three hours with both taps open, both taps are then closed and the pressure in the enclosed volume is measured over several minutes until a steady value is reached. The pressure may show a slight fall over the interval of a few minutes due to ion pumping, depending on the electron current. This rate is determined by observations of pressure with time. The glass tip is then broken and measurements of pressure and time taken over a period of minutes or hours, depending upon the rate at which the pressure rises.

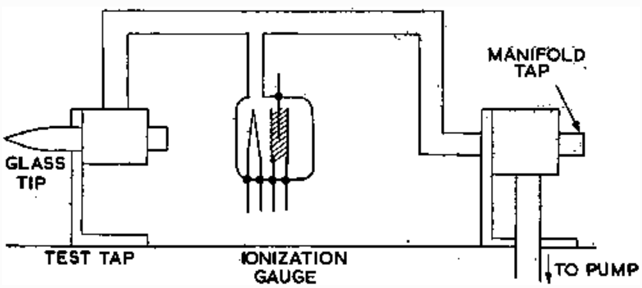

Fig. 3. Arrangement used for testing the taps

It is found that the pressure rises sharply almost immediately the tap is broken and thereafter may continue to rise at a much slower rate. This effect is due to gas molecules being driven from the inner surface of the diaphragms due to the impact of the atmospheric pressure on the exposed surface. The subsequent rise in pressure is determined by the leak rate.

There may be a leak rate in the manifold tap but the measurements of pressure versus time, taken before breaking the tip, give the combined effect of pumping by the ionization gauge and leakage through the manifold tap. The leak rate of the manifold tap is in any case very small since the pressures on each side of the seal are almost equal.

If $p_0$mm Hg is the initial pressure before breaking the tap, $p_1$mm Hg is the pressure after t seconds, V litres is the volume of the enclosed space and A is the atmospheric pressure in mm Hg, then

$$\text{Conductance} = \frac{(p_1 - p_0)V}{A \, . \, t} \quad \text{litres/second.}$$

The conductance measured in this way is higher than that obtained when pressures are so low that the mean-free path of the molecules is long, compared with the linear dimensions of the tap on both sides of the diaphragm, but for practical purposes the method is easy to do and gives directly the performance of the tap when used to shut-off a system temporarily against atmospheric pressure.

Results

The taps are found to have a continuously variable conductance from 0·3 litres/second when open to $<10^{-10}$ litres/second when closed by a torque of 30 to 35 lb ft. Intermediate values can be obtained and the taps may be used as uncalibrated leaks for mass spectrographic work.

These taps are widely used throughout the Mullard Research Laboratories and, although systematic life tests cannot be reported, many of them have been in constant use on normal pump benches for about two years.

A second tap, similar in construction to the first, but with 18mm bore tubulation and modified acutuating mechanism, is illustrated in Fig. 4. The tap may be mounted in several different positions on the stand and may be fitted with flanges for incorporation in an all-metal pump system. The sealing flanges and sealing rings are shown alongside the left-hand tap in the photograph. The sealing method is that described previously and the tap is fully bakeable without adjustment during and after bake[2]. The conductance is continuously variable from 3 litres/second to $< 10^{-10}$ litres/second at a torque of 35 lb ft. Some of these taps have been in use on pump benches for more than a year and are still in good condition.

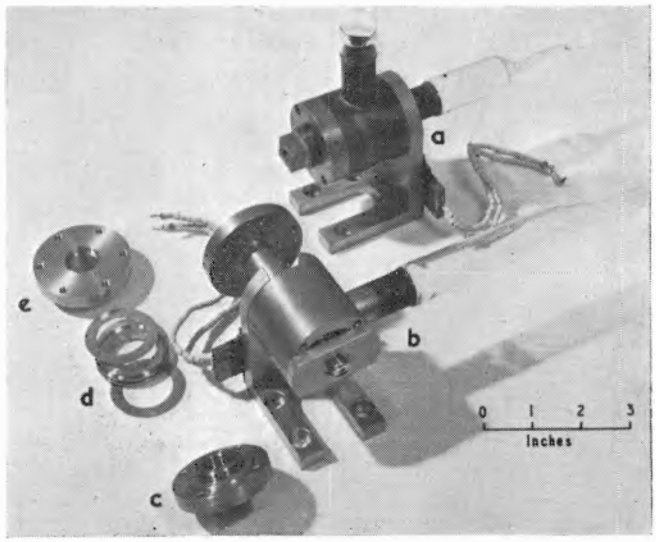

Fig. 4. A tap with 18mm tubulation and modified actuating mechanism

As in all types of metal taps, the chief cause of deterioration is contamination resulting from frequently letting them down to air while another tube is inserted in the manifold. This contamination cannot be cleared completely by baking to 450°C, with the result that the performance may deteriorate during life. Most of the taps have a closed conductance of considerably less than 10^{-10} litres/second when new—indeed the average is about 10^{-12} litres/second —while many have been made with zero conductances.

Acknowledgments

The author wishes to thank the directors of Mullard Research Laboratories and the directors of Mullard Ltd for permission to publish this article.

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LETTERS TO THE EDITOR

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

A Method of Measuring the Output Impedance of a Stabilized D.C. Power Supply

Dear Sir,—In a recent experiment on a design of a stabilized d.c. power supply, it was found desirable to be able to measure the output impedance of the supply while adjustments were made in the power supply circuit, and to view, at any rate qualitatively, the output impedance over a continuous range of load current 0 to 1A simultaneously. The power supply in question employed variable positive current feedback, which made a negative output impedance possible. The method of measurement used enabled this negative output impedance to be observed and measured readily. There is no danger of instability unless the load impedance is less than the numerical value of the output impedance, a condition which could not be achieved in the power supply in question.

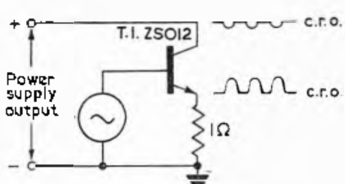


Fig. 1. Method of measurement

The method adopted is shown in Fig. 1. The base of an npn silicon power transistor was driven by an audio oscillator, so that the transistor conducted a half-wave sinusoidal current. A double-beam oscilloscope was used to examine the emitter current (which was taken as equal to the collector current) and the voltage waveform across the power supply output terminals. If the base of the transistor was driven by a sawtooth waveform, the output impedance at different current levels could be measured more easily, and the linearity of output impedance with current would be more easily observable.

It is felt that this is a simpler method of measuring low output impedance than the conventional 'backing off' method. Also, the output impedance can be measured at a variety of frequencies, up to 1Mc/s with the type of transistor used (T.I. 2S012).

Yours faithfully,

B. T. Denvir,
Texas Instruments Limited,
Bedford.

A Graphical Design Method for Non-Linear Resistive Devices

Dear Sir,—There is a tendency when dealing with non-linear elements such as Varistors to assume that the easiest design procedure is an intuitive 'cut and try' approach. A graphical approach however often proves the quickest and easiest method of analysis and design and certainly offers a clearer insight into the behaviour of the circuit.

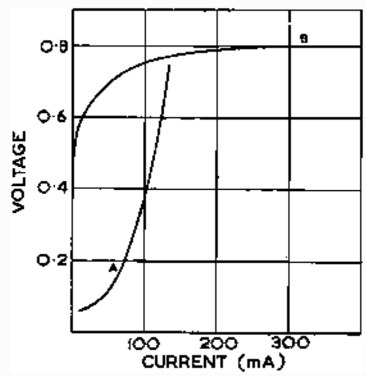
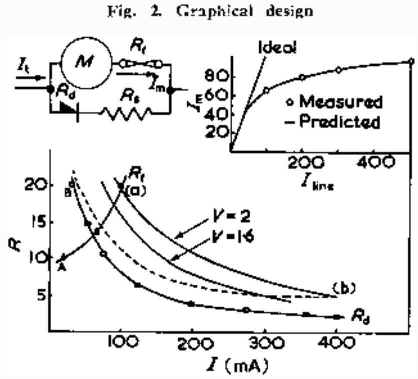


Fig. 1. Typical V-I characteristics

Non-linear devices generally have a V-I characteristic which is either of the type A or B in Fig. 1. Although the V-I plot is normally known or easily obtained for a non-linear element, this plane is not the most useful for graphical design. The resistance current characteristic of the device is of more practical use in design work. This plot has been suggested for analysing the time varying current effects also[1]. Resistance in this plane is defined as the ratio of voltage to current at any particular value of current.

A resistance-current plane carries as an implied auxiliary parameter a voltage which is equal to IR at any point. Constant voltage lines may therefore be plotted on the R-I plane giving a complete representation of R, V and I. Resistors in series, whether linear or not, may be added along constant current lines while one may move to parallel branches by traversing constant voltage lines. It is possible by this method to analyse the various currents and voltages in a non-linear circuit.

Fig. 2. Graphical design



To illustrate this procedure, a meter protection circuit of the type shown in Fig. 2 will be designed. The moving-coil meter has a f.s.d. of 100mA and an internal resistance of 0·6Ω. It is anticipated that line surges of 0·5A will occur and the meter current is not to exceed 100mA under these conditions. The non-linear elements have resistance-current characteristics as shown in Fig. 2. R_d is a silicon diode and R_f a 100mA fuse.

The steps in the design are as follows:

(1) Plot all resistance characteristics on the R-I plane. In this example the meter and fuse resistance have been combined in one line.

(2) Choose a meter current of 100mA (maximum conditions) and find intersection with R_f line (point (a)).

(3) Proceed down the constant voltage line to the 400mA line (point (b)). This is the desired current in the diode branch. It is necessary to add a resistance of R_sΩ to limit the current to 400mA.

(4) A new (R_d plus R_s) curve can now be plotted.

(5) The circuit response can now be predicted by assuming various meter currents and finding the corresponding diode current, i.e. choose say 80mA of meter current, find the intersection with R_f line. Proceed down constant voltage to (R_d plus R_s), read I_d. The meter current plus the diode current must be the total line current.

A plot of meter current against line current for predicted and measured values is shown in Fig. 2. It is possible from this plot to see the contribution to the protection curve of each element and it can quickly be seen what the effect of additional shunt or series non-linear resistors would be.

The example used as an illustration combines the effects of adding linear and non-linear elements in series and parallel. The important considerations are the same as in normal circuit analysis. The series elements must be added along constant current lines and to move to a parallel branch one must move via a constant voltage line. The method proposed is sufficiently general to permit the analysis of any circuit in which the d.c. voltage-current characteristics must be evaluated.

Yours faithfully,

B. A. Bowen,
University of Syracuse,
Syracuse, N.Y.

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Correction

In 'The Author Replies' by J. F. Young on page 623 of the October 1959 issue, the second sentence in paragraph one should read: "The effect with a particularly bad tube is shown in the photograph. The effect can be compensated by the use of potentiometers in the cathode circuits, though this is a complex solution."

Short News Items

Some of the results obtained with the Jodrell Bank Telescope were announced during the visit of H.R.H. the Duke of Edinburgh last month, and among these was the development of the technique by which radar echoes were obtained from the planet Venus at its close approach to the earth in September. The experiment was carried out by a group of scientists headed by Dr. J. V. Evans. The transmitter, part of which was mounted on the telescope and part on the ground, consisted of a klystron working on a frequency of 408Mc/s transmitting pulses of 1/30sec duration every second. This klystron was originally designed by Professor S. Devons, Langworthy Professor of Physics in the University, as part of a linear accelerator for research in nuclear physics. Some of the associated transmitter equipment was provided by the English Electric Co. of Stafford, and part of the expense of the installation was borne by a grant from the Electronics Directorate of the United States Air Force. On account of the great distance of the planet special arrangements were necessary to integrate the received signals over considerable periods of time. The preparations for this experiment took several years and the planet was already receding from the earth before the first tests were made in mid September. The distance was then 30 million miles, and the radio waves took 5min to travel from Jodrell Bank to Venus and back. The procedure consisted of setting the telescope in automatic motion to follow the planet across the sky; transmission of the radar pulses were then made for 5min, the transmitter was then switched off and the received power integrated for 5min. This process was repeated continuously and altogether about 60 hours' observations were made before the planet receded out of useful range at the end of September. The results indicate that the integrated echo occurred at the range suggested by the previous experiments carried out at Millstone Hill in America. The value of the solar parallax derived from this work is $8{\cdot}8020 \pm 0{\cdot}0005$ seconds of arc and can be regarded as a valuable clarification of the diverse values obtained by optical measurements over the past century. The intensity of the received echo was considerably less than that expected from theoretical calculations, and from that indicated on the basis of the Millstone results. The reason for this may either be connected with some peculiarities in the reflecting properties of the surface of Venus, or with the speed of rotation which is assumed to be about 23 days but cannot be determined by visual means because of the thick cloud cover of the planet. The resolution of these uncertainties would require the use of more powerful transmitting equipment.

The Institution of Electrical Engineers has arranged a Christmas Holiday Lecture to be given in the Lecture Theatre at Savoy Place, London W.C.2, at 3 p.m. on Wednesday, 30 December 1959 and to be repeated on the following afternoon, Thursday, 31 December at the same time. The subject of the lecture, which will be delivered by Mr. G. G. Gouriet will be 'Colour Television'. This is the fourteenth year in which The Institution has organized these Christmas Lectures, and an effort has been made to evolve a suitable technique for their delivery to schoolchildren, so that they may be of the maximum interest and benefit to them.

The Radio Trades Examination Board has recently announced the results of the 1959 Radio Servicing Examination and Television Servicing Certificate Examination. Of the 1 869 candidates who entered for the Radio Servicing Examination, 547 were successful, 291 were referred in the practical test and 1 099 failed. This represents the highest failure rate since the examination was first started in 1944. The main cause of failure was the second written paper although the overall results were also poor. The practical test was carried out on the "Trainer-Tester" system for the first time and proved to be a very satisfactory form of testing fault diagnosis. The Television Servicing Certificate results were very much better: 485 candidates were entered, of which 209 qualified for the award of the Certificate, 94 were referred in the practical test and 160 failed.

The Electronic Forum for Industry—E.F.F.I.—held its second Open Day in London last month under the chairmanship of Mr. C. Metcalfe, managing director of E.M.I. Electronics Ltd, when representatives of twelve 'user' associations were present. Three films were shown illustrating the role of electronics in industry and commerce and this was followed by an open forum at which representatives of the electronic industry answered questions posed by users. E.F.F.I. was formed in February of last year for the interchange of ideas between the rapidly expanding electronic industry and the growing number of user associations. Constituent members are, British Electrical and Allied Manufacturers' Association (BEAMA), British Radio Valve Manufacturers' Association (BRVMA), Electronic Engineering Association (EEA), Machine Tool Trades Association Inc. (MTTA), Office Appliances and Business Equipment Trades Association (OABETA), Radio and Electronic Component Manufacturers' Federation (RECMF), Scientific Instruments Manufacturers' Association (SIMA), and Telecommunication Engineering and Manufacturing Association (TEMA).

The Eighth Annual Convention of the Scientific Instrument Manufacturers Association was held at Brighton on 22 to 24 October under the presidency of Mr. L. A. Woodhead, Cossor Instruments Ltd. The theme of the Convention was 'Instruments of Destiny' and was discussed in four panels. The panels considered a number of subjects under the headings of 'Instruments of the Future', 'Future Materials and Production Techniques', 'Organizing for the Future' and 'The Individual and the Company'. The discussions covered the Nuclear Age, the Space Age, and the conditions, technology, instruments and components that would be needed in the future, including such purposes as ballistic missiles and satellites. The organizational, marketing and financial needs of the forward looking company were covered, and a number of subjects from the aspect of the individual in regard to such things as health, financial rewards, pensions, etc. were dealt with. Among a list of distinguished speakers were The Rt. Hon. Lord Piercy, C.B.E., chairman of the Industrial and Commercial Finance Corporation Ltd.; Professor H. S. W. Massey, F.R.S., Quain Professor of Physics, University College of London; Dr. H. M. Finniston, B.Sc., Ph.D., A.R.T.C., F.I.M., research manager of Nuclear Research Centre, C. A. Parsons & Co. Ltd; Dr. H. Beric Wright, M.B., F.R.C.S., Head of the Medical Research Unit, Institute of Directors.

The Microwave Research Institute of the Polytechnic Institute of Brooklyn announce that the tenth of its annual International Symposia will be held at the Engineering Societies Building, 33 West 39 Street, New York, from 19 to 21 April 1960. This symposium on "Active Networks and Feedback Systems" will be sponsored by the Department of Defence Research Agencies and the Institute of Radio Engineers as in the past. One of the aims of this symposium is to seek a unity of approach to the problems in the broad field of active networks, particularly with regard to linear systems. It will also present specific aspects of the general theory as they apply to electron devices and control systems. The papers given will include the following topics: Physical Realizability of Active Networks; Representation and Synthesis of Active Networks; Synthesis of Feedback Networks; Time Varying Systems; Applications of Active Network Theory to Electron Devices; Applications of Active Network Theory to Feedback Systems. The closing date for the submission of papers and/or 100 word abstracts is 15 January 1960. Correspondence should be addressed to Professor H. J. Carlin, Microwave Research Institute, 55 Johnson Street, Brooklyn 1, New York.

Norwood Technical College, Knight's Hill, West Norwood, London S.E.27, will celebrate its centenary on 16 and 17 December 1959 from 2.30 p.m. to 9 p.m., when the work of the various departments, including the Telecommunications, Physics and Chemistry Departments, will be displayed by films, demonstrations and models. Manufacturers of commercial, domestic and scientific electronic apparatus are also contributing a large number of exhibits. Admission is free.

Nash & Thompson Ltd have announced that they are now the exclusive selling agents in the United Kingdom and Commonwealth for KOVO, the foreign trade corporation for the Import and Export of precision engineering products made in Czechoslovakia. Among the instruments that will be imported into the U.K. for the first time are the Polaroscope and Polarographs designed at Professor Heyrovsky's Polarographic Research Institute in Prague. These are widely used in Eastern Europe and Baltic countries, a fact which was demonstrated by the papers read by Czech and Polish scientists at the International Polarographic Congress recently held at Cambridge. Possibly the most interesting of these instruments is the P.576 polaroscope, which is particularly useful for rapid polarographic determinations. An interesting feature is that the reversibility of the electrode reaction is immediately apparent. Other apparatus includes the Tesla B.S.242 intermediate electron microscope, telecommunication equipment, spectrophotometers and electrolytic analysis apparatus.

The British Electrical and Allied Manufacturers Association held their Fourth Publicity Conference in London on 3 and 4 November under the chairmanship of Mr. C. T. M. Bagnall, when some 180 delegates of the member firms attended. The principal topics for discussion were Industrial Photography, the Role of House Magazines in Industry, Editorial Information, Exhibitions and Publicity Overseas. The report of the Conference will be published in the near future and will contain the speakers' papers and the discussions.

Laurence, Scott & Electromotors Ltd would like to hear from ex-apprentices with whom they have lost touch. It is hoped to arrange for a get-together, possibly during the run of the Electrical Engineers Exhibition next year. Ex-apprentices are asked to give details of their present whereabouts, either to Mr. W. F. Symes at Gothic Works, Norwich, or by telephone to Mr. T. J. Barfield at Temple Bar 5223.

Standard Telephones & Cables Limited have planned, manufactured and installed Malaya's first main-line trunk microwave telephone network which was brought into service in September last. This company has been associated with the Pan-Malayan Telecommunications Department since 1931, when they installed Malaya's first three-channel carrier telephone systems. In recent years the major part of the trunk telephone traffic in Southern Malaya has been carried by v.h.f. radio links which provided 96 circuits between Kuala Lumpur, Malacca and Singapore. When increasing demands on the service made it necessary to provide additional circuits, the Pan-Malayan Department of Telecommunications anticipated future expansion by installing the new microwave system. Initially this system is equipped to provide less than its full capacity of telephone circuits and additions can be made when required.

The microwave network for Malaya comprises six radio links interconnected by multiplex channelling equipment to provide direct circuits between the various exchanges in the network.

The main route from Kuala Lumpur to Singapore and the spur from Tampin to Malacca consists of duplicated S.T.C. type RL4C radio equipments of 600 circuit capacity. This system operates in the 3 800 to 4 200Mc/s band, using the C.C.I.R. frequency allocation plan. It provides an output of 5W from a travelling wave amplifier which feeds a 10ft diameter parabolic antenna via waveguide feeders.

The spur routes from Tampin to Seremban and from Kluang to Johore Bahru will use the S.T.C. type RL7A equipment having a capacity of 240 telephone circuits. This work in the 7 300 to 7 500Mc/s band and will also be duplicated to provide interruption-free service. Eight-foot parabolic antennas are used with a transmitter power of ½W. Both systems are equipped with a supervisory system providing alarms in the case of an equipment or power supply failure. This also allows the control stations at Kuala Lumpur, Fort Canning and Malacca, to initiate certain remote control switching operations at the other stations which are designed for unattended operation. The supervisory and control equipment of the RL4C employs an independent s.h.f. vehicle channel to interconnect all stations on the route. The vehicle channel will use the C.C.I.R. recommended frequencies in the 3 800 to 4 200Mc/s band.

The systems are each equipped with multiplex channelling equipment of the type now usual for wideband radio or line telephone systems. This provides up to 600 C.C.I.T.T. carrier telephone circuits spaced 4kc/s apart in the band 60kc/s to 2·6Mc/s. Crystal filters enable a single stage of modulation to be employed in the formation of the basic 12-channel group in the frequency band of 60 to 108kc/s.

The equipment works from 240V single-phase a.c. mains. As is usual on a system of this type, motor alternator and engine alternator sets, specially designed for feeding telecommunications equipment are used. These are arranged in conjunction with standby power plant to provide a "no-break" power supply to the radio equipment.

One of the Royal Medals recommended by the Council of the Royal Society, has been awarded to Professor R. E. Peierls, C.B.E., F.R.S., professor of mathematical physics in the University of Birmingham, for his distinguished work on the theoretical foundations of high energy and nuclear physics.

Corrections

In the article 'An Electrical Analogue for Heat Flow Problems in Semiconductors' by N. L. Potter on page 454 of the August issue the following corrections are necessary:—

Line above equation (3) should read $J_y = -k\mathrm{A}\, d/dx\, (T + dT/dx\, \Delta\, x)$. The Δ is missing.

Six lines above equation (6) $\Delta V/x$ should be $\Delta V/\Delta x$

Line above equation (6) has the = sign missing.

Sixth line below Calibration of the Analogue has a small c_t which should be a large C_t.

Table on page 455

The symbols below R_t C_t R_θ and C_θ should be respectively

$$\frac{r}{kA},\ c\rho rA,\ -\frac{mzr}{kA} \text{ and } \frac{zc\rho rA}{m}$$

In the article 'Charts for Deriving Transistor r-Parameters from h-Parameters' by G. W. E. Stark on page 592 of the October issue the last sentence before the Acknowledgment should read:—

Finally, subtract this value of r_s from the known value of h_{ib} to give r_e.

BOOK REVIEWS

Design of Transistorized Circuits for Digital Computers

By A. I. Pressman. 324 pp. 219 figs. Medium 8vo. John F. Rider Publisher, Inc., New York. 1959. $9.95.

MORE than half of this book is concerned with the design of various arrangements of diode gates having transistor invertors at intervals, of similar arrangements in which resistors replace the gating diodes and of direct-coupled transistor logic. Although the basic circuits and principles involved here are few, well known and quite simple, these pages are well worth reading and constitute the major attraction of the book. For each of the arrangements considered, full details are given (including arithmetic values for a practical design using commercially available devices) of the procedure whereby tolerances of all components and power supplies are taken into account and desired switching speeds are obtained. The author has obviously been intimately concerned with such circuits and these pages teach by describing practical engineering steps with real numbers rather than by discussing disembodied theory. Another chapter similarly treats the design of multivibrators but this is perhaps not quite so successful.

The consideration of transistor transient response in all designs is a particularly good point and an early chapter on this is also well done. The now classical equations of Ebers and Moll are quoted and their meaning clearly illustrated by many practical values and graphs. Some confusion may be experienced here, however, owing to the lack of a clearly stated, or consistently used, sign convention.

A chapter on other logic circuits (storage and transducer circuits are completely omitted), attempts to correct for the otherwise somewhat restricted outlook, but is rather too brief. Moreover, figures 10.3, 10.4 and 10.7 of this chapter which illustrate the description of an alternative and important approach (ref. H. S. Yourke, Trans. I.R.E., Vol. CT-4, No. 3) show certain inductances wrongly connected and the effect of these is completely ignored in the treatment given of switching speed, although it may be extremely significant. The emphasis in most chapters on alloy junction transistors and the assumptions permittedly made for these may also tend to prevent a wider application of some of the ideas described.

There is a tendency for space to be wasted. Diagrams and graphs are often larger than necessary and sometimes not really justified at all. No doubt the early pages on logic and transistor fundamentals are helpful and make the book self-contained but one wonders whether certain elementary concepts, e.g. binary numbers and crystal lattices, need have been treated at such length.

T. W. SHEPPARD

Progress in Dielectrics, Vol. I

Edited by J. B. Birks and J. H. Schulman. 312 pp. 319 figs. Royal 8vo. Heywood & Company. 1959. Price 70s.

ANY doubts as to the need for this new serial publication in view of the annual American 'Digest of Literature on Dielectrics' are quickly dispelled when one finds that the book under review is a well produced and generously illustrated collection of reviews covering many years of publications in a skilfully chosen series of fields. These admittedly do not cover the wide range of topics (some of them marginal) to be found in the 'Digest', but any loss in this respect is amply compensated by the completeness, quality and readability of each chapter. The publishers would do their readers a favour, in any subsequent issues, by having section numbers printed at the top of each page and bibliography page numbers at the foot.

A panel of authors, drawn from Britain and America, consists of names familiar to workers in the field of dielectric materials. J. H. Mason (E.R.A., Leatherhead) deals with electrical breakdown in solids, starting with the thorny subject of 'intrinsic' breakdown and the degree to which theory and experiment agree, passing on to the more practical topic of pre-disposing causes of failure (internal discharges, etc.) and finishing with a short summary of insulation testing methods. The bibliography consists of 115 entries.

J. W. Davisson (U.S. Naval Research Laboratory) reviews the subject of directional effects in the breakdown of crystals, bringing out the close ties between the crystal structure and the complex patterns formed by partial-breakdown paths. The actual mechanism of breakdown, however, is still not free from doubt. Sixty-five references are given.

The chapter on high-field conductivity of dielectric liquids, written by T. J. Lewis of Queen Mary College, London, surveys the work between 1937 and 1958. Recent attempts to unify theories of breakdown and conduction have brought the former study into some prominence in this section, which concludes with a useful list of 99 references.

T. W. Liao and R. E. Plump (G.E. Co of America) contribute a chapter and 111 references on gaseous dielectrics. It is a pity that the authors' anxiety to avoid becoming involved in the intricacies of gas discharges should have led to so meagre a discussion of the conditions for breakdown. One is rewarded, however, in the experimental section where, as is to be expected, fluorine compounds figure prominently. The treatment is practical throughout.

In the chapter on ferro-electricity by Alan D. Franklin (National Bureau of Standards, U.S.A.) barium titanate is taken as a representative dielectric in this class, and the predictions of Devonshire's theory illustrated with its help. Some useful figures for more recently discovered ferro-electrics appear in an appendix. The bibliography of 64 entries contains only one British contribution apart from those by Devonshire. One wonders why.

P. Popper (British Ceramic Research Association) contributes a chapter, on non-oxide ceramic dielectrics, possessing the attractive feature of a substantial introductory section on solid-state theory. The electrical properties of a wide range of compounds are reviewed against this background, with the result that the chapter, which might have degenerated into a dull catalogue, is of absorbing interest. The bibliography contains 188 entries.

A similar revision course in fundamentals is included in the final chapter on electrophoretic deposition of insulating materials by one of the co-editors, J. B. Birks of Manchester University. The chapter covers every significant aspect of the subject, from the theoretical to the technological, with pleasing lucidity. References total 90.

The publishers are to be congratulated on this extremely timely, useful and well produced book. It will be welcomed by all those engaged in the field of dielectrics, the more enthusiastically because published papers in this field are scattered through the literature of engineering, physics and physical chemistry

K. A. MACFADYEN

Analysis of Linear Systems

By D. K. Cheng. 431 pp. 288 figs. Medium 8vo. Addison-Wesley Publishing Co. Inc., Massachusetts. Pergamon Press, Ltd, London. 1959. Price $8.50.

EVER since the publication of the first volume of Gardner and Barnes's *Transients in Linear Systems* there has been a continual flow of books from the United States on the use of the Laplace transformation in the study of linear systems. Most of them seem to have evolved from sets of lecture notes used by their authors at some time or other but none of them adds very much to what is already known of the techniques of linear system analysis. Nevertheless, this new book by Professor D. K. Cheng of Syracuse University is exceptionally good.

The book has evidently been modelled on that of Gardner and Barnes. After an introductory discussion of linear systems from both the physical and mathematical points of view, the classical methods of solving linear differential equations with constant coefficients are reviewed. This is followed by an excellent presentation of the fundamentals of linear electrical network analysis. The elements of an electrical network are defined, the loop and node methods of setting up the network integro-differential equations are explained, the principle of duality is introduced and Cauer's rule for obtaining the dual of a given planar network is described. A few typical network problems are solved and the discussion concludes with Thévenin's and Norton's theorems.

Following this is a similar presentation of linear mechanical network analysis. The mechanical network elements are defined and the analogy between mechanical networks, translational and rotational, and electrical networks is emphasized. Methods of transforming mechanical networks into electrical networks are discussed in detail. Both the force-voltage and the more natural force-current analogies are considered.

At this point the Fourier and Laplace transformations are introduced. The development of the Laplace transformation from the Fourier transformation is particularly well done. The use of the Laplace transformation in the solution of integro-differential equations and the determination of network response to a variety of excitation functions, including non-sinusoidal periodic functions, is treated at length. Also considered are the impulse function, the convolution and superposition integrals, and the inverse Laplace transformations of some irrational functions such as the gamma and error functions.

Linear systems incorporating feedback are dealt with next. Among the topics covered are the use of block-diagrams and signal flow graphs in the representation of feedback systems and the Routh-Hurwitz and Nyquist stability criteria. For the purpose of dealing with sampled-data feedback systems in which the actuating error is supplied intermittently at discrete intervals of time, the Z-transformation is introduced and its properties discussed. The Z-transformation is then applied to the solution of difference equations and to the study of open-and closed-loop systems. It is also shown how the modified Z-transformation is used to determine system response between sampling instants. The final chapter of the book is devoted to the application of the Laplace transformation to the analysis of systems with distributed parameters. Two appendices follow; the first gives an account of Graeffe's root-squaring method of finding the roots of polynomials and the second contains a list of Laplace transformation pairs.

The book is intended for advanced undergraduates or graduates. It is remarkably well planned and easy to read. Lists of problems append each chapter and answers are provided at the end of the book. A knowledge of the Laplace transformation calculus has become a necessity for the modern electro-mechanical engineer. This book can be thoroughly recommended to anyone looking for a lucid textbook on the analysis of linear systems.

S. R. DEARDS

Sampled Data Control Systems

By E. I. Jury. 453 pp. 120 figs. Demy 8vo. John Wiley & Sons, New York. Chapman & Hall Ltd., London. 1959. Price 128s.

RAPID advances in the design and application of digital computers have markedly increased the interest in the field of sample data systems. This is the second excellent volume to originate in Columbia University. The first by Ragazzini and Franklin formed an introduction to the subject, and the present volume which contains much original material covers more the application of the theory to system design. Obviously a volume of this nature is highly mathematical and is intended for the practising system designer, or qualified engineer. The work is based on the notes for a two-year post graduate course at Columbia University.

Application of the Z-transform to sample data systems forms an introduction to the subject following on from servomechanism theory. The use of digital computers as the compensating component of feedback systems requires the description of the output at all instants of time, this is achieved by a modified Z-transform method which is completely developed. A section on the investigation of various mathematical models is introduced at this stage. A chapter each is devoted to the application of the familiar root locus and frequency response methods to sample data systems. The author then discusses the synthesis problem and introduces a mathematical approach to discrete compensators. Continuous compensation is well discussed, in both the frequency and time domains.

Of invaluable use to the system designer will be a chapter on the physical realization of discrete compensators. Final chapters cover the application of the Z and modified Z-transforms to the approximate analysis of continuous systems and the exact analysis of sample data systems. An interesting section on the limitations of the Z-transform is included, also the application of the Z-transform to the solution of linear difference equations.

An extensive bibliography is given, also many references are quoted during the various chapters. Many examples, both solved and unsolved, are included, unfortunately many of them are hypothetical cases.

Worthy of particular mention are the many lengthy tables of transforms and various sample data system configurations. These tables will be extremely useful to the system designer.

E. F. SNARE

Radio Circuits

By W. E. Miller and revised by E. A. W. Spreadbury. 172 pp. 84 figs. Demy 8vo. 4th Edition. Iliffe & Sons Ltd. 1959. Price 15s.

EXPLAINED in this popular book, in simple language and in easy stages, are all the varieties of circuits found in radio receivers of the kind that are used for sound broadcasting. It does this by taking separately each stage of the receiver, piece by piece, and explaining it without the complication of all the associated circuits round it.

Since the last edition was published, several fundamentally new techniques have been developed, and these have given the new author considerable scope in revising the book for the fourth edition. The book now covers all types of receiver, including those using transistors, car radio, and frequency modulated receivers.

Conductance Design of Active Circuits

By K. A. Pullen. 323 pp. 166 figs. Medium 8vo. J. F. Rider Publisher Inc., New York. Chapman & Hall Ltd., London. 1959. Price 80s.

THIS book deals with the practical design of valve and transistor circuits from their characteristic curves. Each chapter gives a brief summary of the important equations and then works out a specific design based on a set of characteristics. The topics covered include audio amplifiers, sine-wave oscillators, mixers and various special circuits such as the cathode-coupled amplifier.

The book does not deal with the derivation of the various equations on which a design is based so that it is not suitable for use by students in university or technical college courses. The various active devices, both valve and transistor, are all of American types so that the text is not as useful to a British engineer as to his American counterpart. The book itself is almost devoid of references to original work, but a two-page bibliography appears at the end of the book; this bibliography is restricted almost wholly to the author's own publications.

V. H. ATTREE

Correction

In the book review of 'Solid State Magnetic and Dielectric Devices' by F. J. Hyde on page 620 of the October 1959 issue, "Electromechanical Applications" should appear instead of "Electrochemical Applications" in the second paragraph.

ELECTRONIC EQUIPMENT

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

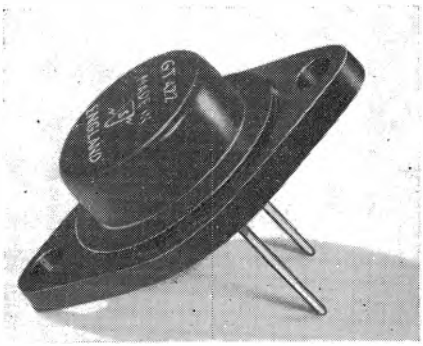
(Voir page 771 pour la traduction en français: Deutsche Übersetzung Seite 776)

GERMANIUM POWER TRANSISTORS

(Illustrated below)

Sylvania-Thorn Colour Television Laboratories Ltd, Great Cambridge Road, Enfield, Middlesex

A new series of npn 60V 6A germanium alloy transistors has been introduced by Sylvania-Thorn Colour Television Laboratories Ltd. Designed for power switching applications it is anticipated that they will give the designer of electronic equipment greater scope in the use of npn transistor technique at less cost than has been possible hitherto.



Among other applications the GT422 can be used in conjunction with a pnp type transistor as a complementary pair for transformerless radio or servo amplifier output stages, or as a series voltage regulator. Other types in the new series are the GT424 (60V 3A), GT425 (36V 6A) and GT426 (36V 3A). Maximum junction temperature is 85°C.

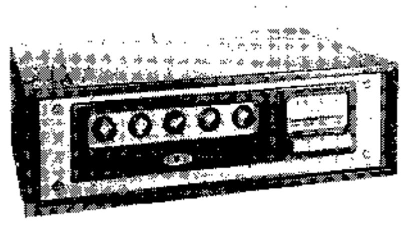
EE 16 751 *for further details*

NUCLEONIC COUNTING INSTRUMENTS

(Illustrated above right)

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

Isotope Developments Ltd have recently announced their latest range of equipment, the '1800' Series. This is a new series of 'budget' priced nucleonic counting instruments designed to give accuracy in operation, full flexibility of application and simplicity of maintenance and servicing. The instruments are supplied in attractively styled cabinets with tilt-up stand so that they can be used singly or stacked one above the other. They can also be converted readily to rack mounting instruments when required. Since all connectors are mounted at the rear and the cabinet is fitted from the front, access can be obtained to the interior without disconnecting cables.



Printed circuit plug-in boards are used extensively throughout the series and an identical power pack is fitted in all units to simplify spares requirements and servicing. Each instrument after 100 hours' test is despatched with a fully detailed service manual and packaged spares kits can be provided at the same time.

The first three members of the series now available for delivery are the scaler, type 1800, the high voltage supply, type 1820 and the amplifier/selector, type 1830.

Further instruments in the range will include a ratemeter, type 1810, a prescaler, type 1850, a potentiometer unit, type 1840, and other units to be announced later.

EE 16 752 *for further details*

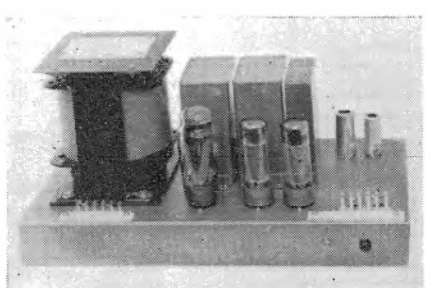
DEKATRON POWER UNIT

(Illustrated below)

Sandmar Electronic Products, 68 Nina Drive, New Moston, Manchester, 10, Lancashire

These power supply units, which are available in either sub-unit or 19in panel form, are designed specifically for supplying Dekatrons and their associated circuits.

Three stabilized outputs are provided: the first supplies +450V d.c. at a maximum current of 10mA with a ripple of less than 0·02 per cent; the second supplies +300V d.c. at 200mA with a ripple of less than 0·005 per cent and regulation better than 0·1 per cent for a change in load from 0 to 200mA; the third output provides −100V d.c. at 10mA. In addition an unstabilized a.c. output of 4 or 6·3V at 3A is provided. The unit is designed so that when switching on from cold the voltages will be applied in the correct sequence.



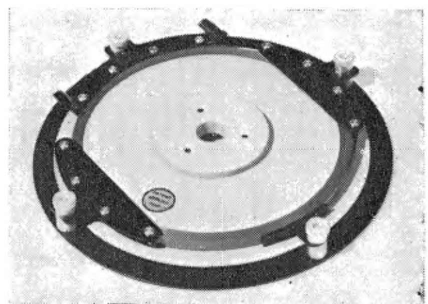
EE 16 753 *for further details*

ENDLESS-TAPE CASSETTE

(Illustrated below)

Guy's Calculating Machines Ltd, Truro Road, London, N.22

The 'Brittape' endless-tape cassette will convert any flat-topped 7in spool tape deck for continuous operation.



Mounted over one or other of the drive-spindles of the tape deck, the cassette provides up to 200ft of constantly circulating ¼in magnetic tape, either standard or long-play.

When in use, the cassette is driven by the tape being drawn in the normal way through the existing mechanism of the deck.

The 'Brittape' cassette enables one to circulate continuously a message or set of signals of up to 22 minutes duration, or to run a tape continuously in order to capture transient phenomena occurring at random intervals.

EE 16 754 *for further details*

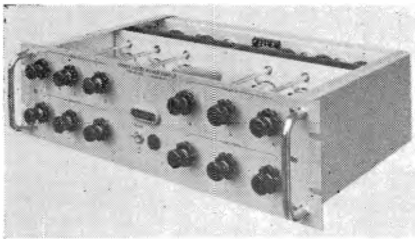
STRAIN GAUGE POWER SUPPLY

(Illustrated above right)

Solartron Electronic Group Ltd, Thames Ditton, Surrey

This fully transistorized six-channel strain gauge power supply unit, type AS958, has been introduced primarily for use with the Solartron pressure transducers.

When working from transducers into a high gain amplifier (such as the Solartron AA900 data amplifier), a normal stabilized power supply may produce a large amount of unwanted mains hum.



The AS958 is a new departure in that special transformers are employed to reduce such undesirable feed-through to a figure of less than 0·1 per cent of full scale output.

The unit has been designed to work into bridges of a nominal 350Ω resistance, but may also be used with resistance bridges of higher impedance. The unit is thus suitable for energizing most types of strain gauges, bonded or unbonded.

This mains-energized power supply incorporates six individual, 'fully-floating' stabilized power sub-units. Each sub-unit is capable of providing a nominal 5V of regulated d.c. at up to 20mA. A 'sensitivity' control allows variation of the output voltage by ±5 per cent.

Also incorporated is a 'balance' control capable of correcting initial bridge unbalance of up to 10 per cent of full-scale. 'Sensitivity' controls (for each channel) are brought out to the front panel and adjustments may be clamped by use of concentric control locks.

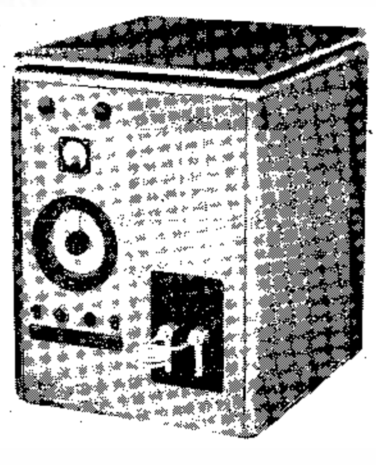
EE 16 755 *for further details*

R.F. INDUCTION HEATER

(Illustrated below)

Pye Ltd, R.F. Heating Division, 28 James Street, Cambridge

The second induction heater in the new Pye range is the R.F.2 which has medium impedance output and gives a continuous output of 1·5kW at 2Mc/s and has been designed for bench operation. It is similar in appearance to the first model being in an aluminium cubicle and fitted with safety switches on the removable sides. The overall dimensions of the unit being 24¾in high, 17½in wide, 21½in deep. The equipment is self contained with its own automatic resetting process timer, overload relay to protect the valves, and water pressure switch to protect the equipment. Provision is made for remote control and a pulse is available at the end of the heating cycle to initiate a quench or any other operation required. The equipment is designed to be operated from a single-phase supply, 180 to 250V, 50c/s, with a full load consumption of 3·1kW.



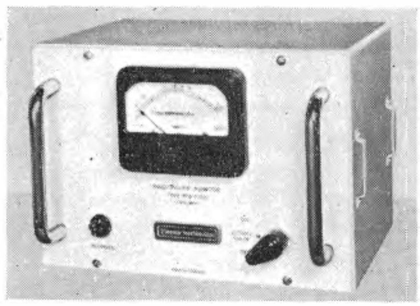
EE 16 756 *for further details*

PORTABLE BACKGROUND MONITOR

(Illustrated below)

Plessey Nucleonics Ltd, Weedon Road, Northampton

A new lightweight portable monitor has been designed and developed by Plessey Nucleonics Limited of Northampton for the measurement of background radiation.



This monitor, known as type PNI 1072, has been designed to fill the frequent need for an accurate means of measuring any background activity quickly and easily. Detection is accomplished by a built-in halogen-quenched Geiger Müller tube with measurement on a transistor ratemeter giving direct reading of dose rate on a 2½in scale moving-coil meter. The instrument has a range of from 0 to 3mr/h on a two-decade quasi-logarithmic scale, with a minimum scale calibration of 30μr/h. The unit is powered by a 9V dry battery and weighs 8 lb.

EE 16 757 *for further details*

BAND-PASS FILTERS

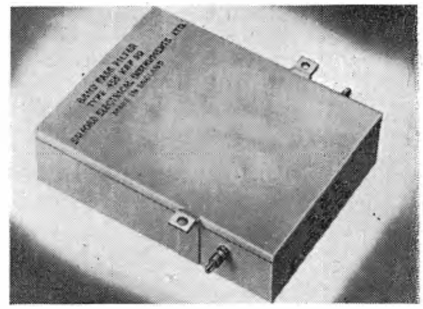
(Illustrated above right)

Salford Electrical Instruments Ltd, Times Mill, Heywood, Lancashire

Salford Electrical Instruments Ltd, a subsidiary of The General Electric Co. Ltd, has introduced two new standard filters designed to control the pass-band characteristics of v.h.f. communications receivers. Known as types 455KBP50 and 455KBP25, they provide the selectivity necessary to meet G.P.O. radio communication specifications at an intermediate frequency of 455kc/s, with either 25kc/s or 50kc/s channel spacing.

The filter is completely encapsulated and fitted in a metal box, input and output leads being brought out to two short stand-off pillars situated on either side of the box.

Fixing brackets are arranged so that the unit may be mounted through a cut-



out in the chassis in order to place the filter terminals close to the rest of the circuit.

EE 16 758 *for further details*

ELECTRONIC 'BUILDING BRICKS'

(Illustrated in next column)

Mullard Equipment Ltd, Mullard House, Torrington Place, London, W.C.1

Mullard Equipment Ltd has introduced a comprehensive series of electronic sub-assemblies designed as standard 'building bricks' for the construction of industrial control systems, digital computing elements, data processing equipment, measuring systems and other equipments based on digital circuit techniques.

These building bricks constitute complete, fully developed circuits guaranteed to work efficiently with one another in any required combination and stringently tested for reliable service under the severest operational conditions.

Two ranges of sub-assemblies are available: 'Norbits' and 'Combi-Elements'. Norbits have been developed primarily to allow industrial firms to build their own sequence control systems quickly, economically and without the services of trained electronic engineers. They can also be used as standard building bricks by electronics firms specializing in control systems.

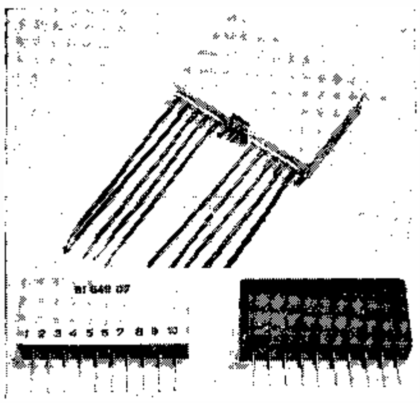
The range includes all the appropriate circuit elements for the construction of machine tool and transfer line controls, factory safety systems, automatic lift controls, sequential programming systems and so on.

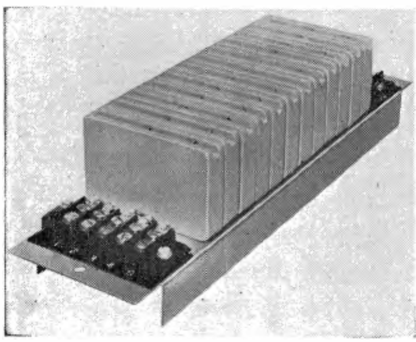
'Combi-Elements' are intended specifically to provide the manufacturer of electronic systems and equipment with a set of standard building bricks representing the most commonly used digital circuits. Besides cutting down development time and costs. Their use leads to substantial savings in space and affords greater flexibility in the building of larger systems and equipments.

To ensure a high standard of reliability both the Norbits and Combi-Elements are transistorized throughout. Transistors also make it possible for the various units to be housed in compact plastic cases of very small dimensions, reduce power consumption, simplify the problems of heat dissipation and avoid the need for high voltage sources.

'Norbit' building bricks are specifically designed so that the entire planning and construction of an industrial sequence control system can be carried out by the firm's plant engineer with no more than a very general knowledge of electronics.

The units have long flying leads so that inter-connexions may be made without intermediate tag-panels of additional wiring, and using solderless clinched joints. Channel units, equipped with terminating blocks for input and output connexions, are provided on which a number of Norbits may be mounted by single screw fixings.





The basic 'Norbit' unit is a transistor switching element which, in the interests of simplicity and flexibility, has been designed for use as an AND or an OR circuit, depending on the manner of connexion. On receipt of information from the machine's limit-switches or other input sources, this switching unit performs the necessary logic operations and delivers signals in the correct sequence to a suitable output unit, which in turn activates the transducers governing or monitoring the functions of the machine. The basic unit measures 2½in by 1 9/16in by ½in and twelve can be accommodated on a single channel.

A number of output units of various ratings up to 150W are available for operating contactors, electromagnetic clutches, solenoids, etc; other units include a counter, and a timer with a range of 0 to 60sec.

The range of Combi-Elements includes flip-flops and other multivibrator circuits, pulse shapers, gate circuits, amplifiers and emitter-follower circuits.

Each circuit of Combi-Element is hermetically sealed in a small plastic case measuring only 2·1in by 0·94in by 0·39in and designed for quick, easy mounting on a printed circuit board using the same techniques normally employed for mounting resistors and capacitors. The physical proportions of the units are designed so that the maximum possible number can be accommodated on a single board.

Assembly time and costs are thus minimized, and single elements or complete boards may be interchanged readily, giving greater flexibility in the composition of more elaborate systems and equipments and simplifying modifications.

The range of circuits has been based on the principle of d.c. gating and in general the signal transfer from element to element is at d.c. levels. The flip-flops and other multivibrators can be triggered by either a d.c. signal or a voltage step (a.c. signal), and depending on the application the units can be used for non-synchronous or synchronous logic, the latter using clock pulses.

EE 167 59 *for further details*

TEMPERATURE CONTROLLER

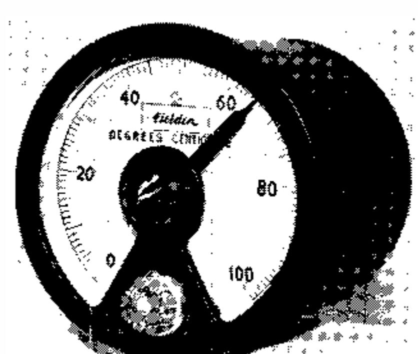
(Illustrated below)

Fielden Electronics Ltd, Wythenshawe, Manchester, Lancashire

The first of the new Fielden 'Bikini' range of transistorized instruments has just been released. This is a precision temperature controller with a differential of 0·5°C, high calibration accuracy and high long term stability.

The instrument is housed in a small meter type case and occupies a panel space of only 4½in diameter. The scale, which is 9in long, is calibrated directly in degrees Celsius or Fahrenheit and the pointer is set to the required control temperature by a small knob in the centre of the instrument. A red lamp visible from the sides as well as from the front indicates the control action. The meter type case is completely sealed and contains no delicate moving parts and the instrument can be mounted at any angle, for example, on a sloping control desk. Control can be at any temperature between −200°C and +500°C with a minimum temperature range span of 50°C and there are 12 standard calibrations.

A platinum resistance element is used at the temperature measuring point, this being housed in a robust stainless steel sheath of ¼in diameter. Ordinary connecting cable is used between the measuring point and the instrument and this can be up to 300ft in length without introducing errors.



The control action is on-off, either electrical or pneumatic. The electrical contacts give changeover switching at 5A 250V while the pneumatic valve requires 17 to 20 lb/in² input.

EE 16 760 *for further details*

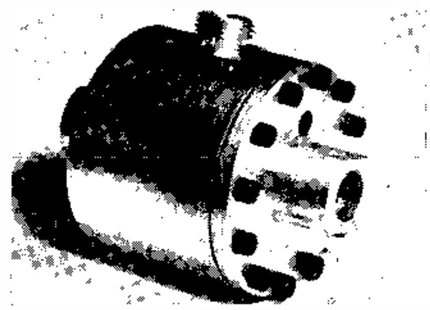
DIFFERENTIAL PRESSURE TRANSDUCER

(Illustrated below)

Southern Instruments Ltd, Frimley Road, Camberley, Surrey

Southern Instruments can now supply a differential pressure gauge type G255 which employs a variable inductance sensing element for use with their f.m. amplifier system.

Pressures can be applied to either, or both sides of a beryllium copper diaphragm: movement of this diaphragm is detected by an inductor cast into the body of the unit. The voltage output from the f.m. pre-amplifier is strictly proportional to the difference in pressure



on the side of the diaphragm. Linearity is better than 3 per cent of rated range. The range of pressures covered is 1-0-1 lb/in² to 1 000-0-1 000 lb/in².

EE 16 761 *for further details*

SILICON CONTROLLED RECTIFIERS

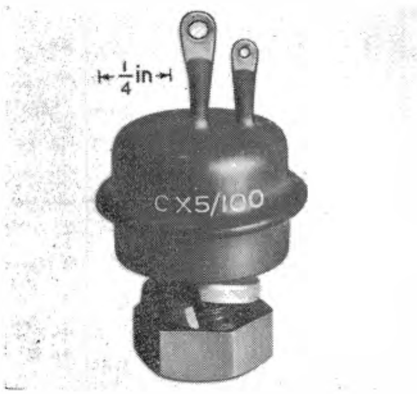
(Illustrated above right)

Associated Electrical Industries Ltd, Carholme Road, Lincoln

This new device, which is of the pnpn form, can be regarded as two transistors coupled by a common collector junction; it is the solid-state equivalent of the thyratron or the grid-controlled mercury-arc rectifier. It offers a high impedance (several megohms) in each direction, but can be triggered into a low impedance condition in one direction by the application of a small signal to the third electrode. It has the usual solid-state advantages of small size and absence of heater supplies, and has a high efficiency of operation.

Applications are similar to those of the conventional devices, such as the control of electric motors, firing of ignitrons, d.c. to a.c. conversion, frequency conversion, phase-controlled d.c. power supplies, static switching, etc.

The voltage/current characteristic of a controlled rectifier is such that in the reverse direction it is almost identical with that of a simple silicon rectifier. The forward direction is a mirror image of the reverse characteristics up to the



breakdown voltage. If a small current (10 to 100mA) is applied to the control electrode when the forward voltage is below breakdown, the rectifier will remain conducting until the voltage across it falls to almost zero. A common method employed is to control the output of the rectifier by changing the phase of the trigger current in relation to the a.c. voltage applied to the rectifier.

The present range covers ratings from 25 to 300V p.i.v. at forward currents of 5 and 10A.

EE 16 762 *for further details*

OSCILLATOR-AMPLIFIER

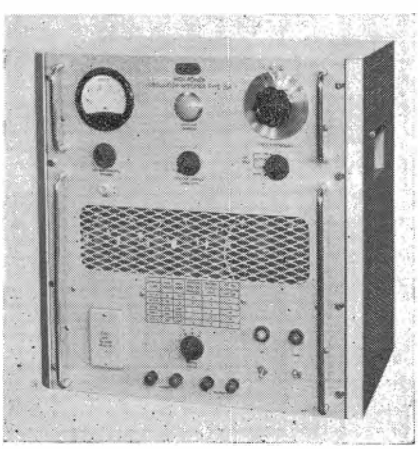
(Illustrated below)

Airmec Ltd, High Wycombe, Buckinghamshire

The oscillator-amplifier type 254 has been designed basically as a power source for small servometers, vibrators and 'odd-frequency' equipment.

The equipment consists of a power amplifier giving an output of 150W, with provision for being driven from an external signal or from an internal oscillator continuously variable over the range 30c/s to 30kc/s. If required, the amplifier may be isolated from the oscillator and used as a public address system or transmitter modulator.

The output stage consists of four EL34 valves in parallel push-pull working into an output transformer with a number of secondary windings, which may be connected in series-parallel arrangements to match a wide range of loads. A meter is fitted to monitor the output voltage. The valves are fully protected by fuses and spark gaps, and a warning lamp is provided to indicate overdriving or excessive anode current. In addition, the anodes of the valves may be viewed through a grille in the front panel of the equipment.



The oscillator employs a Wien bridge feedback network stabilized by a thermistor, and the frequency of oscillation is indicated on a large, clear dial to an accuracy of ±2 per cent. When driven from an external signal, the oscillator valve is switched to operate as a pre-amplifier, giving a sensitivity of 0·1V for full output.

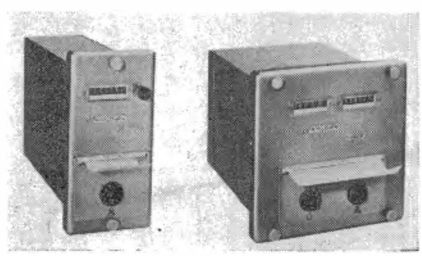
EE 16 763 *for further details*

PRINTING COUNTERS

(Illustrated below)

The Stonebridge Electrical Co. Ltd, 6 Queen Anne's Gate, London, S.W.1

This Company has introduced a new range of printing counters which print out on paper roll indications of count and time, and can be used for a wide variety of purposes in industry and research.



They are arranged as single or double units. The single units are normally six digit totalizing counters with electrical impulsing, printing and zero reset. Double units may be fitted with two such totalizing units or one totalizing and one unit giving date and time information. Impulsing rate is maximum 10 per second, printing is carried out in 100msec, and reset in 500msec.

Models are available for various voltages, both a.c. and d.c.

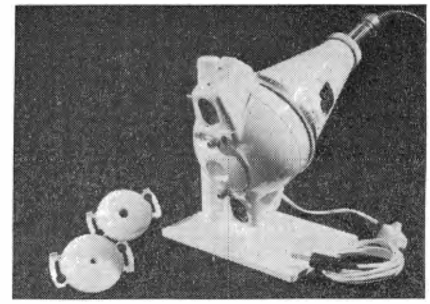
EE 16 764 *for further details*

MEDICAL SCINTILLATION COUNTER

(Illustrated above right)

Ekco Electronics Ltd, Southend-on-Sea, Essex

A recent addition to the range of nuclear equipment for medical use by Ekco Electronics Ltd is a lightweight scintillation counter, type N618, with a directional shield, type N651. The counter is constructed in a sealed stainless steel container enabling it to be cold-sterilized. The output from the photo-multiplier tube is fed via a cathode-follower stage to the permanently-attached cable from which connexions can be made to a suitable ratemeter or scaler. It can be connected directly to an Ekco type N600 ratemeter, but requires the use of a separate amplifier, such as the Ekco type N640, when connected to a typical ratemeter or scaler requiring a minimum input of 5V.



The directional shield retains the scintillation counter by means of a simple clamp ring and it can be fitted with any one of three standard collimators to provide alternative angles of acceptance. To facilitate accurate repositioning in relation to the patient, a continuously-adjustable coincident light-spot range-finder is incorporated.

EE 16 765 *for further details*

PORTABLE PRECISION METERS

(Illustrated below)

Metrix Instruments Ltd, 54 Victoria Road, Surbiton, Surrey

A complete range of Metrawatt portable precision ammeters, voltmeters and wattmeters is now being marketed by Metrix Instruments Ltd.

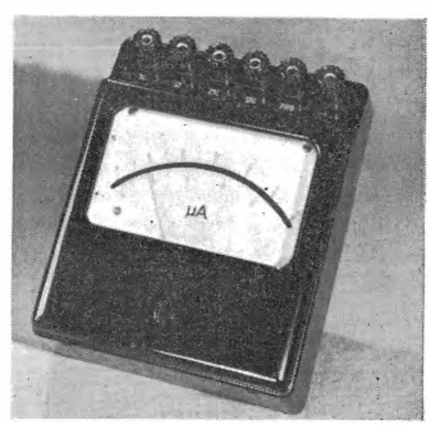
These meters which are of high quality construction, are designed to meet the need for sensitive instruments having an accuracy somewhat higher than that of general purpose bench models for test and laboratory measurements where it is inconvenient or uneconomical to use an expensive and often delicate standard instrument. The range is also suitable for use as transfer standards.

The range of moving-coil instruments have an accuracy of ±0·5 per cent and include a voltmeter covering 6 to 600V d.c. in five ranges. Four current measuring instruments cover 20μA to 2mA; 0·5 to 50mA; 0·2 to 20A, and 0 to 60mV for use with shunts up to 2 500A.

A similar range of precision moving-iron types having an accuracy of ±0·5 per cent a.c. ±1 per cent d.c. is also available.

These instruments are manufactured by Metrawatt A.G. Nürnberg.

EE 16 766 *for further details*



PUBLICATIONS RECEIVED

TRANSMISSION TESTING APPARATUS CATALOGUE contains a representative selection of the testing apparatus available, which has been developed mainly for the installation and maintenance of a wide variety of carrier telephone and other systems. Standard Telephones & Cables Ltd, Transmission Division, North Woolwich, London E.16.

UNILEVER RESEARCH is a booklet giving an impression of life in the Unilever research laboratories in England and will be of interest to those interested in careers in research. Unilever Limited, Unilever House, London E.C.4.

LOW FREQUENCY AMPLIFIER SYSTEMS is the 31st volume in the Electronic Technology Series edited by Dr. Schure. The content deals with low frequency amplifier systems with special emphasis on coupling methods suitable for this frequency range, phase inversion and inverse feedback, as well as the circuit design using thermionic valves and transistors. John F. Rider Publisher, Inc., 116 West 14th Street, New York 11.

ULTIMATE PARTICLES OF MATTER by D. T. Lewis is an attempt to present to the layman some of the more complex modern theories of chemistry and physics with regard to the structure of matter. There is a preface by Sir William Penney, and he refers to the phenomenal growth of the atomic energy industry and stresses the need for industrial staff who have a clear perception of the basic scientific ideas on which everything depends. Chantry Publications Limited, 63 Neal Street, Shaftesbury Avenue, London W.C.2. Price 15s.

TRAINING REGULATIONS has been published under the authority of The Council of The Institution of Electrical Engineers. Much of the content of this leaflet will not be new to those who have been concerned with the training of students and graduates, but they do indicate once again the importance which the Council attach to the ordered arrangements of this phase of a young engineer's preparation for his chosen profession. What is new is the recognition by the Council that some electrical organizations are unable to offer an organized type of training known as apprenticeship. The Regulations accept that, in such cases, effective training may be obtained by a junior engineer while holding a position on the staff of a chartered electrical engineer. Copies can be obtained from The Secretary, The Institution of Electrical Engineers, Savoy Place, London W.C.2.

PRINTED CIRCUITS by Morris Moses traces the development of printed circuits, tells how they are made commercially, how the reader can make them and how to repair, maintain and apply equipment, what valves, transistors, resistors, capacitors and other components are used with them and how to repair, maintain and apply them in home entertainment, industrial and test equipment. Gernsback Library Inc, 154 West 14th Street, New York 11. Price $2.90 for paper covered edition.

BRITISH WIEDEMANN TURRET PUNCH PRESSES is the latest comprehensive brochure issued by Dowding & Doll Ltd., 346 Kensington High Street, London W.14. The short production times, of which typical examples are given in this new brochure, should be of special interest to all concerned with the production of pierced work, such as chassis for electronic apparatus, aircraft, and electrical panels, where runs are comparatively short and modifications frequently made.

COPPER CABLES is a 62-page booklet devoted exclusively to information concerning the many types of insulated copper cables. There are chapters devoted to the properties of copper and some of its alloys, the manufacturers of copper wire, power, transmission and other types of insulated cables. Copper Development Association, 55 South Audley Street, London, W.1.

'WIRELESS WORLD' DIARY 1960 contains, in addition to the usual diary pages, an 80-page reference section which includes in tabloid form a wide variety of technical and general information. Published by T. J. & J. Smith Ltd., 12 Hanover Square, London W.1. Price: 6s. 3d. leather and 4s. 6d. rexine, postage 4d.

FIXED CARBON RESISTORS FOR USE IN TELECOMMUNICATION AND ALLIED ELECTRONIC EQUIPMENT PART I: GENERAL REQUIREMENTS AND TESTS (B.S. 2112:1959). This British Standard applies to fixed carbon resistors having a dissipation not exceeding 3W at 70°C and a rated resistance value of not less than 10Ω and not greater than 10MΩ. These resistors are intended for use over the ambient temperature range −40°C to +100°C in one of the humidity classes (H1, H2, H3 and H3A) of B.S. 2011, 'Basic climatic and durability tests for components for use in radio and allied electronic equipment'. British Standards Institution, Sales Branch, 2 Park Street, London W.1. Price 10s. (Postage extra to non-subscribers.)

STABILIZED TRANSISTOR POWER SUPPLIES, SOLARTRON CHOKELESS POWER SUPPLY SUB-UNITS and SOLARTRON STABILIZED POWER SUPPLY SUB-UNITS are all data sheets which have been issued by The Solartron Electronic Group Ltd, Thames Ditton, Surrey. They deal with the three ranges of Solartron Power Supply Sub-Units now available for direct incorporation into prototype or production electronic equipment.

ELLIOTT INSTRUMENTS IN MICROWAVE ANALYTICAL CIRCUITS FOR THE 8-12Gc/s BAND is a catalogue recently issued by Elliott Automation Limited, 34 Portland Place, London W.1, which gives illustrations of the scope and versatility of their X-Band range.

24 CHANNEL FREQUENCY MODULATED VF TELEGRAPH TERMINAL EQUIPMENT TYPE T24D is a leaflet describing equipment which is intended primarily for use over radio or carrier bearer circuits. The 24 channels lie within the speech-frequency band of 300 to 3 400c/s. At working speeds of 50 baud the average distortion is less than 5 per cent with receive level variations about the nominal of +10dB to −15dB. Telephone Manufacturing Company Limited, Technical Information Department, Hollingsworth Works, Martell Road, London S.E.21.

MOLYBDENUM DISULPHIDE IN ACTION has been issued by Moly-Paul, K. S. Paul (Molybdenum Disulphide) Ltd, Angel Lodge Laboratories and Works, Angel Road, London N.18. The reports in this booklet are mainly extracts of letters from users or reports of inspections of specified applications, classified under fourteen technical headings. Copies can be obtained free from the above Company.

DESIGNING FOR DIECASTING is one of a series of technical publications, and attempts to show how the design of pressure diecastings in particular, can be altered to give a combination of good properties, low cost and good design. This booklet is obtainable free on application to Fry's Diecastings Ltd, Merton Works, Prince Georges Road, Merton Abbey, London S.W.19.

STUDIES IN CRYSTAL PHYSICS by M. A. Jaswon is published by Butterworths Scientific Publications, 4 & 5 Bell Yard, London W.C.2. This booklet contains four articles on crystal physics which were published in Research in 1958. Each touches on a different topic, but all are concerned with physical effects related primarily to the crystal structure rather than to the specific kind of atom involved. A fifth article has been added which provides the thermodynamic background to the picture. Price 10s. 6d., postage 7d.

PRINTED WIRING & PRINTED CIRCUIT TECHNIQUES DESIGN & PRODUCTION CONSIDERATIONS is a document which has been prepared by the Electronic Engineering Association's Printed Wiring and Associated Techniques Committee. This is a 'first edition' and it is hoped as knowledge accumulates and techniques advance, to issue further editions. In order to guide the Committee's future efforts, appraisal of the document in the widest sense will be welcomed from all concerned, not only in respect of technical content, but also on layout, clarity and the need for expansion. Copies can be obtained free from The Electronic Engineering Association, 11 Green Street, London W.1.

Meetings this Month

THE BRITISH COMPUTER SOCIETY LIMITED

Date: 17 December. Time: 2.30 p.m.
Held at: Northampton College of Advanced Technology, St. John Street, London E.C.1.
Discussion: Conversion between Analogue and Digital Representation.

Leicester Branch

Date: 17 December. Time: 7 p.m.
Held at: Room 104, Leicester College of Technology.
Film Evening. Two films will be shown: Unilever's 'The Electronic Computer in Commerce' and I.C.T.'s 'Time to Think'.

South Wales and Monmouthshire Branch

Date: 17 December. Time: 6.30 p.m.
Held at: Small Shandon Lecture Theatre, University College, Cardiff.
Lecture: Mechanical Translation of Languages.
By: Dr. J. P. Cleve.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Medical Electronics Group Meeting

Date: 16 December. Time: 6.30 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London W.C.1.
Lecture: Measurements in the Presence of Noise.
By: Dr. D. A. Bell and Dr. G. D. Dawson.

North Eastern Section

Date: 9 December. Time: 6 p.m.
Held at: The Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: Micro Miniaturization.
By: H. G. Manfield.

Scottish Section

Date: 18 December. Time: 7 p.m.
Held at: Department of Natural Philosophy, the University, Drummond Street, Edinburgh.
Lecture: The Digital Voltmeter.
By: J. A. Irvine and D. A. Pucknell.

South Western Section

Date: 17 December. Time: 7 p.m.
Held at: School of Management Studies, Unity Street, Bristol.
Lecture: The Transistor and its Use in Communication and Control Equipment.
By: E. Wolfendale.

THE INSTITUTE OF PHYSICS

Electronics Group

Date: 15 December. Time: 5.30 p.m.
Held at: The Institute of Physics, 47 Belgrave Square, London S.W.1.
Lecture: Tubes for Colour Television.
By: K. G. Freeman.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings will be held at Savoy Place commencing at 5.30 p.m.

Electronics and Communications Section

Date: 16 December.
Lecture: The Laying of Submarine Cables.
By: Captain W. H. Leech.

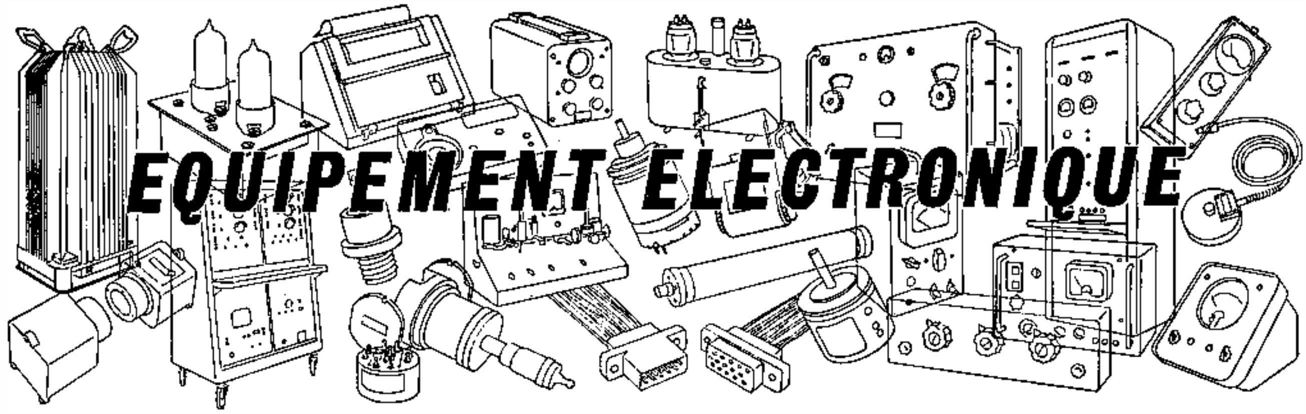
Measurement and Control Section

Date: 15 December.
Discussion: Data Handling Problems in Atomic Installations.
Opened by: D. Taylor.

THE RADAR AND ELECTRONICS ASSOCIATION

Date: 29 December. Time: 7.30 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London W.C.2.
Lecture: High Quality Sound Broadcasting.
By: K. R. Sturley.

(Pages 771 to 780 French and German Section)



EQUIPEMENT ELECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 766 à 769

TRANSISTORS DE PUISSANCE AU GERMANIUM

(*Illustration à la page 766*)

Sylvania-Thorn Colour Television Laboratories Ltd, Great Cambridge Road, Enfield, Middlesex

Une nouvelle gamme de transistors à alliage de germanium npn 60 V 6 A a été réalisée par Sylvania-Thorn Television Laboratories Ltd. Calculée pour les applications de commutation de puissance, elle fournira, prévoit-on, au constructeur d'équipements électroniques un rayon d'action plus étendu dans l'emploi de la technique des transistors npn et à moins de frais que cela n'a été possible jusqu'ici.

Entre autres applications, le GT422 peut être utilisé conjointement avec un transistor de type pnp, en tant que paire complémentaire pour radio sans transformateurs ou étages de sortie de servo-amplificateur, ou bien encore comme régulateur de tension série. Les autres modèles de la nouvelle série sont le GT424 (60 V 3 A), le GT425 (36 V 6 A) et le GT426 (36 V 3 A). La température de jonction maxima est de 85°C.

EE 16 751 *pour plus amples renseignements*

COMPTEURS NUCLÉAIRES

(*Illustration à la page 766*)

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

La société Isotope Developments Ltd vient d'annoncer la réalisation de sa toute dernière gamme d'appareils de comptage nucléaire. Il s'agit de la série 1800, à prix "budgétaires", calculée pour assurer un fonctionnement précis, une grande souplesse d'application, ainsi qu'un entretien et un dépannage faciles. Ces appareils sont livrés dans de séduisants coffrets avec supports inclinables, de façon à pouvoir être utilisés individuellement ou entassés l'un sur l'autre. Ils peuvent également être convertis, si nécessaire, en appareils montés sur crémaillère. Vu que tous les connecteurs sont montés à l'arrière et que le coffret est fixé par l'avant, on peut aisément vérifier l'intérieur sans débrancher les câbles.

Des plaquettes de circuits imprimés interchangeables sont utilisées sur une grande échelle dans toute cette série et tous les appareils sont munis de blocs d'alimentation identiques, afin de simplifier le dépannage et les besoins en pièces de rechanges. Après un essai de 100 heures, chaque appareil est livré avec un manuel d'entretien, et une trousse de pièces de rechange peut être fournie en même temps.

Les trois premiers appareils de la série, livrables dès maintenant, sont l'échelle, type 1800, le bloc d'alimentation haute tension, type 1820, et l'amplificateur/sélecteur, type 1830.

Les appareils complémentaires de la série comprendront un compteur, type 1810, une pré-échelle, type 1850, un élément potentiométrique, type 1840, ainsi que d'autres éléments qui seront annoncés plus tard.

EE 16 752 *pour plus amples renseignements*

BLOC D'ALIMENTATION POUR DÉKATRONS

(*Illustration à la page 766*)

Sandmar Electronic Products, 68 Nina Drive, New Moston, Manchester, 10, Lancashire

Ces blocs d'alimentation sont livrés soit sous forme de sous-ensembles, soit comme organes de 483 mm pour montage sur panneau. Ils ont été spécifiquement conçus pour l'alimentation des Dékatrons et des circuits qui s'y rapportent.

Trois sorties stabilisées sont prévues. La première fournit +450 V c.c. à un courant maximum de 10 mA, avec une ondulation de moins de 2,02%. La seconde fournit +300 V c.c. à 200 mA, avec une ondulation de moins de 0,005% et un réglage de plus de 0,1% pour un changement de charge de 0 à 200 mA. La troisième fournit—100 V c.c. à 10 mA. Il y a, en outre, une sortie non-stabilisée de courant alternatif de 4 ou 6,3 V à 3 A. Le bloc a été étudié de manière à ce que les tensions soient appliquées en succession exacte lorsqu'on effectue la mise en circuit à froid.

EE 16 753 *pour plus amples renseignements*

CHÂSSIS À BANDE MAGNÉTIQUE SANS FIN

(*Illustration à la page 766*)

Guy's Calculating Machines Ltd, Truro Road, London, N.22

Le châssis à bande magnétique sans fin "Brittape" peut convertir au fonctionnement continu n'importe quel pont plat à bobines de ruban magnétique de 178 mm. Monté sur l'une ou l'autre des broches d'entraînement du pont à bobines, il fournit jusqu'à 61 mètres de ruban magnétique de 6,35 mm à mouvement constant, soit du modèle courant soit à durée prolongée.

Durant le fonctionnement, le ruban est entraîné de façon normale par le mécanisme du pont sans modification aucune.

Le châssis "Brittape" permet de faire circuler continûment un message ou une série de signaux pendant une durée de temps pouvant atteindre 22 minutes. On peut aussi faire tourner un ruban continûment pour repérer des phénomènes transitoires survenant à intervalles irréguliers.

EE 16 754 *pour plus amples renseignements*

BLOC D'ALIMENTATION D'EXTENSOMÈTRE

(*Illustration à la page 767*)

Solartron Electronic Group Ltd, Thames Ditton, Surrey

Ce bloc d'alimentation entièrement transistorisé, à six canaux, type AS 958, pour extensomètres, a été prévu en premier lieu pour être utilisé avec les transducteurs de pression Solartron.

Lorsqu'on opère, à partir de transducteurs, dans un amplificateur à grand gain, tel que l'amplificateur de données Solartron AA 900, une alimentation stabilisée normale peut produire un niveau élevé de ronflement de secteur indésirable. Le bloc AS 958 constitue une véritable innovation, car des transformateurs sont utilisés pour réduire ce ronflement gênant à un chiffre de moins de 0,1% de la sortie totale.

L'appareil a été calculé pour des ponts d'une résistance nominale de 350 ohms, mais il peut, cependant, être utilisé également avec des ponts de résistances d'une impédance plus élevée. On peut donc, l'employer pour amorçer la plupart des types d'extensomètres.

L'ensemble d'alimentation amorçée par le secteur comprend six sous-blocs individuels d'alimentation stabilisée "entièrement flottants". Chacun de ces sous-blocs peut fournir un courant continu réglé de

5 V jusqu'à 20 mA. Une commande de sensibilité permet des variations de la tension de sortie de ±5%.

L'appareil comprend aussi une commande d'équilibre capable de corriger un déséquilibre de pont initial pouvant atteindre jusqu'à 10% de l'échelle totale. Les commandes de sensibilité (pour chaque canal) sont placées sur le panneau avant et les réglages peuvent être fixés au moyen de verrous de commande concentriques.

EE 16 755 *pour plus amples renseignements*

RÉCHAUFFEUR PAR INDUCTION HAUTE FRÉQUENCE

(Illustration à la page 767)

Pye Ltd, R.F. Heating Division, 28 James Street, Cambridge

Le second réchauffeur par induction de la nouvelle série Pye est l'appareil type R.F.2 avec sortie à impédance moyenne. La sortie continue de cet appareil est de 1,5 kW à 2 MHz et il a été réalisé pour l'emploi au banc. Il est semblable, en apparence, au premier modèle de la série, étant donc contenu dans un coffret d'aluminium avec interrupteurs de sûreté sur les côtés amovibles. Les dimensions hors tout de l'appareil sont: 616 mm de hauteur, 445 mm de largeur, 540 mm de profondeur. Il est complet par lui-même, étant muni de son propre compte-pose de réenclanchement automatique, d'un relais à maxima pour la protection des lampes et d'un interrupteur de pression d'eau pour la protection de l'ensemble de l'appareil. Ce dernier peut être également télécommandé, et une impulsion peut être émise à la fin du cycle de chauffage pour effectuer un étouffement ou toute autre opération voulue. L'appareil est prévu pour une source d'alimentation monophasée de 180 à 250 V, 50 Hz, avec consommation à pleine charge de 3,1 kW.

EE 16 756 *pour plus amples renseignements*

DÉTECTEUR DE RAYONNEMENT DE FOND PORTATIF

(Illustration à la page 767)

Plessey Nucleonics Ltd, Weedon Road, Northampton

Un nouveau détecteur portatif poids léger a été conçu par la Plessey Nucleonics Ltd, de Northampton, pour mesurer le rayonnement de fond.

Ce nouvel appareil, type PNI 1072, a été réalisé pour répondre à la demande fréquente d'un moyen de mesure précis, rapide et simple de tout rayonnement de fond. La détection s'effectue par tube de Geiger incorporé, à extinction au halogène, avec mesure sur un compteur à transistor donnant une lecture directe du dosage sur un dosimètre à cadre mobile avec échelle de 63,5 mm. L'appareil a une gamme de 0 à 30 milli-r/h sur une échelle quasi logarithmique à deux décades, avec un étalonnage d'échelle minimum de 30 micro-r/h. Il est actionné par pile sèche de 9 V et ne pèse que 3,6 kg.

EE 16 757 *pour plus amples renseignements*

FILTRES PASSE-BANDES

(Illustration à la page 767)

Salford Electrical Instruments Ltd, Times Mill, Heywood, Lancashire

La société Salford Electrical Instruments Ltd, qui est une filiale de la General Electric Co. Ltd, a réalisé deux nouveaux filtres pour contrôler les caractéristiques de bande passante des récepteurs pour hyperfréquences.

Appelés types 455KBP50 et 455KBP25, ces filtres assurent la sélectivité conforme aux spécifications des autorités postales concernant les radiocommunications, à une fréquence intermédiaire de 455 kHz, avec un espacement entre canaux de 25 kHz ou de 50 kHz.

Le filtre est entièrement blindé et logé dans une boîte métallique. Les conducteurs de sortie et d'entrée sont reliés à deux petites colonnes isolantes, placées sur les deux côtés de la boîte.

Les supports de fixage sont disposés de façon à ce que l'appareil puisse être monté au moyen d'une ouverture dans le châssis, en ayant les bornes du filtre près du reste du circuit.

EE 16 758 *pour plus amples renseignements*

"BRIQUES ÉLECTRONIQUES"

(Illustration à la page 768)

Mullard Equipment Ltd, Mullard House, Torrington Place, London, W.C.1

La société Mullard Equipment Ltd vient de réaliser une série étendue de sous-ensembles électroniques sous forme de "briques" pour la construction de systèmes de contrôle industriel, d'éléments de calculateurs numériques, d'appareils de dépouillement de données, d'équipements de mesure ainsi que d'autres matériels basés sur la technique des circuits numériques.

Ces "briques" constituent des circuits complets et entièrement mis au point, dont le fonctionnement efficace entre elles est garanti dans n'importe quelle combinaison. Elles ont été, en outre, soumises à des essais très stricts de sécurité de service, dans les conditions de marche les plus rigoureuses.

Ils existe deux gammes de sous-ensembles : NORBITS et COMBI-ÉLÉMENTS. Les premiers ont été étudiés principalement pour permettre aux sociétés industrielles de construire leurs propres systèmes de contrôle de cadencements, rapidement, économiquement et sans avoir besoin des services d'électroniciens expérimentés. Ils peuvent aussi être employés comme "briques de construction" normales par les compagnies d'électronique spécialisées dans les systèmes de contrôle. Cette gamme englobe tous les éléments de circuits convenant à la construction des commandes de machines-outils et de lignes de transfert, des systèmes de sécurité d'usines, des commandes automatiques de levage, des systèmes de programmation successive, etc.

Quant aux COMBI-ÉLÉMENTS, ils sont spécifiquement destinés à fournir au fabricant de matériel électronique un assortiment de "briques" courantes, représentant les circuits numériques les plus communément employés. Ils réduisent non seulement le temps et les frais de développement, mais leur emploi conduit aussi à une économie d'espace considérable et offre une plus grande marge de flexibilité dans la construction de systèmes et d'équipements plus importants.

Afin d'assurer un niveau élevé de sécurité de fonctionnement, tant les NORBITS que les COMBI-ÉLÉMENTS sont entièrement transistorisés. L'utilisation de transistors rend possible, de même, de loger les différents éléments dans des boîtiers compacts en matière plastique, aux dimensions fort réduites, ainsi que de diminuer la consommation électrique, de simplifier les problèmes de dissipation de chaleur et d'éviter de recourir à des sources de haute tension.

Les "briques" NORBIT ont été spécifiquement conçues pour que le projet et la construction d'un système de contrôle de cadencement industriel puisse être entièrement exécuté par l'ingénieur de l'usine, même si celui-ci n'a qu'une connaissance très générale de l'électronique.

Les éléments ont de longs conducteurs "volants", de sorte qu'on peut faire des inter-connexions sans panneaux à cosses intermédiaires avec câblage supplémentaire, et employer des joints rivés sans soudure. Des éléments à canaux, équipés de plaquettes de connexions d'entrée et de sortie, permettent de monter sur ces dernières un nombre de NORBITS par fixage à vis unique.

L'élément NORBIT de base est à commutation par transistor. Pour plus de simplicité et de flexibilité, il a été prévu pour être utilisé soit comme circuit "ET" soit comme circuit "OU", selon le genre de connexion. Au reçu des données des interrupteurs de limite de l'appareil ou d'autres sources d'entrée, cet élément de commutation exécute les opérations logiques nécessaires et délivre des signaux, en succession correcte, à un élément de sortie approprié lequel, à son tour, met en oeuvre les transducteurs contrôlant ou détectant les fonctions de l'appareil. L'élément de base mesure 63,5 × 39,7 × 12,7 mm et on peut en loger douze dans un seul canal.

Un nombre d'éléments de sortie, ayant des caractéristiques allant jusqu'à 150 W, sont livrés pour actionner des contacteurs, des embrayages électromagnétiques, des solénoïdes, etc. D'autres éléments comprennent un compteur et une minuterie avec gamme de 0 à 60 sec.

La série des COMBI-ÉLÉMENTS se compose de basculeurs monostables différentiels et autres circuits à multivibrateurs, de façonneurs d'impulsions, de circuits à porte, d'amplificateurs et de circuits à charge cathodique par transistor.

Chaque circuit de COMBI-ÉLÉMENT est hermétiquement scellé dans un petit coffret en matière plastique, ne mesurant que 53,3 × 23,9 × 9,9 mm, pouvant être monté facilement et rapidement sur une plaquette à circuit imprimé, par la même méthode que celle normalement employée pour monter des résistances et des condensateurs. Les dimensions physiques des éléments sont telles qu'on peut en loger le plus grand nombre possible sur une seule plaquette.

La durée de l'assemblage et les frais qui en découlent sont donc considérablement réduits. En outre, les éléments individuels ou des plaquettes complètes sont aisément interchangeables, donnant ainsi une plus grande souplesse à la composition de systèmes et d'équipements plus compliqués et simplifiant les modifications.

La gamme des circuits est basée sur le principe de déclenchement périodique c.c. et, en général, le transfert de signal d'élément à élément s'effectue aux niveaux c.c. Les flip-flops et autres multivibrateurs peuvent être déclenchés soit par signal c.c. soit par étage de tension (signal c.a.) et, selon l'application, les éléments peuvent être employés pour la logique synchrone ou non-synchrone, celle-là utilisant des impulsions de référence de temps.

EE 16 759 *pour plus amples renseignements*

CONTRÔLEUR DE TEMPÉRATURE

(*Illustration à la page 768*)

Fielden Electronics Ltd, Wythenshawe, Manchester, Lancashire

Le premier des instruments de la nouvelle gamme Fielden "Bikini" d'appareils transistorisés vient de sortir. C'est un contrôleur de température de précision, ayant un différentiel de 0,5°C, un étalonnage des plus exacts et une stabilité à long terme élevée.

Ce contrôleur est logé dans un petit coffret du type à compteurs et il n'occupe qu'un espace de 114 mm de diamètre sur le panneau. L'échelle, qui a 229 mm de long, est étalonnée directement en degrés centigrade ou Fahrenheit, l'aiguille étant réglée à la température de contrôle voulue par un petit bouton au centre de l'instrument. Une petite lampe rouge, visible tant des côtés que de l'avant, indique le processus de contrôle. Le coffret est complètement scellé et ne contient pas de parties mobiles délicates, de sorte que l'appareil peut être monté à n'importe quel angle, par exemple sur un pupitre de commande incliné. Le contrôle s'effectue à n'importe quelle température entre -200°C et $+500$°C, avec une gamme minima de températures de 50°C et 12 étalonnages normaux.

Un élément à résistance de platine est utilisé au point de mesure de la température. L'élément est logé dans un robuste fourreau en acier inoxydable de 6,35 mm de diamètre. Un câble de connexion ordinaire est employé entre le point de mesure et l'instrument. La longueur de ce câble peut atteindre 91 mètres sans, cependant, causer des erreurs.

Le mécanisme de contrôle est à "marche/arrêt", soit électrique soit pneumatique. Les contacts électriques donnent une commutation de 5 A 250 V tandis que le clapet pneumatique nécessite une entrée de 1,2 à 1,4 kg/cm².

EE 16 760 *pour plus amples renseignements*

TRANSDUCTEUR DE PRESSION DIFFÉRENTIEL

(*Illustration à la page 768*)

Southern Instruments Ltd, Frimley Road, Camberley, Surrey

La société Southern Instruments fournit maintenant un manomètre différentiel type G255 qui utilise un élément sensible à inductance variable destiné à l'emploi avec le système d'amplification à modulation de fréquence de cette société.

On peut appliquer des pressions à l'un ou l'autre des côtés, ou aux deux côtés d'un diaphragme en cuivre de beryllium. Le mouvement de ce diaphragme est décelé par un inducteur coulé dans le corps de l'appareil. La sortie de tension du préamplificateur à modulation de fréquence est rigoureusement proportionnelle à la différence de pression sur le côte du diaphragme. La linéarité est supérieure à 3% de la gamme nominale. La gamme de pressions s'étend de 0,5-0-0,5 kg/cm² à 70,3-0-70,3 kg/cm².

EE 16 761 *pour plus amples renseignements*

REDRESSEUR COMMANDÉ AU SILICIUM

(*Illustration à la page 769*)

Associated Electrical Industries Ltd, Carholme Road, Lincoln

Ce nouveau redresseur du genre pnpn peut être considéré comme étant formé de deux transistors accouplés par une jonction à collecteur commune. C'est en somme l'équivalent, sous forme solide, du thyratron ou du redresseur à vapeur de mercure et à grilles commandées. Il fournit une impédance élevée (plusieurs mégohms) dans chaque direction, mais il peut être déclenché à basse impédance dans une seule direction, par l'application d'un petit signal à la troisième électrode. Il possède les avantages habituels d'un encombrement réduit et de l'absence d'alimentation de chauffage. De plus, son fonctionnement est des plus efficaces.

Ses applications sont semblables à celles des redresseurs conventionnels, c'est à dire: la commande de moteurs électriques, l'amorçage d'ignitrons, la conversion de courant continu en courant alternatif, la conversion de fréquences, l'alimentation c.c. à commande de phase, la commutation statique, etc.

Les caractéristiques de tension/courant d'un redresseur commandé sont presques identiques dans le sens inverse à celles d'un simple redresseur au silicium. La direction avant est une image fidèle de la direction arrière, jusqu'à la tension de rupture. Si on applique un faible courant (de 10 à 100 mA) à l'électrode de commande, lorsque la tension avant est au-dessous de la rupture, le redresseur continuera à transmettre la tension qui le parcourt jusqu'à ce que celle-ci tombe presque à zéro. Une méthode communément employée consiste à contrôler la sortie du redresseur en changeant la phase du courant de déclenchement par rapport à la tension alternative appliquée au redresseur.

La gamme actuelle couvre des régimes de 25 à 300 V de tension inversée de crête, à des courants avant de 5 à 10 A.

EE 16 762 *pour plus amples renseignements*

OSCILLATEUR-AMPLIFICATEUR

(*Illustration à la page 769*)

Airmec Ltd, High Wycombe, Buckinghamshire

L'oscillateur-amplificateur type 254 a été étudié fondamentalement pour servir de source de puissance aux petits servo-moteurs, vibrateurs et appareils à fréquences non-courantes.

Il comprend un amplificateur de puissance fournissant une sortie de 150 W et pouvant être actionné par un signal extérieur ou par un oscillateur intérieur à variation continue, dans la gamme de 30 Hz à 30 kHz. Si nécessaire, l'amplificateur peut être isolé de l'oscillateur et employé comme système de sonorisation extérieure ou comme modulateur d'émission.

L'étage de sortie se compose de quatre lampes EL 34 fonctionnant en push-pull parallèle dans un transformateur de sortie avec un nombre d'enroulements secondaires qui peuvent être reliés en série-parallèle afin de s'adapter à une gamme étendue de charges. La tension de sortie est contrôlée par un compteur. Les lampes sont entièrement protégées par des fusibles et des éclateurs, tandis qu'une lampe d'alarme indique les surcharges ou les courants anodiques excessifs. De plus, les anodes des lampes peuvent être vues à travers une grille sur le panneau avant de l'appareil.

L'oscillateur utilise un réseau à réaction à pont de Wien, stabilisé par un thermistor, et la fréquence d'oscillation est indiquée sur un grand cadran d'une lecture facile, à un degré de précision de ±2%. Lorsque l'appareil est actionné par un signal extérieur, la lampe d'oscillateur est commutée pour fonctionner comme pré-amplificateur, donnant une sensibilité de 0.1 V pour une sortie totale.

EE 16 763 *pour plus amples renseignements*

APPAREILS IMPRIMEURS

(*Illustration à la page 769*)

The Stonebridge Electrical Co. Ltd, 6 Queen Anne's Gate, London, S.W.1

Cette société vient de présenter une nouvelle série d'appareils imprimant sur rouleaux de papier des indications de comptage et de temps et pouvant être utilisés pour une gamme étendue d'applications dans l'industrie et les travaux de recherches.

Ces appareils s'emploient comme ensembles uniques ou comme ensembles doubles. Les premiers sont généralement des totalisateurs à six numéros, à impulsions, impression et réenclanchement de zéro électriques. Quant aux ensembles doubles, ils peuvent comporter soit deux totalisateurs soit un totalisateur et un élément fournissant des données de date et de temps. Le taux maximum d'impulsions est de 10 par seconde, l'impression s'effectuant en 100 msec. et le réenclanchement en 500 msec.

Il existe des modèles pour différentes tensions, tant alternatives que continues.

EE 16 764 *pour plus amples renseignements*

COMPTEUR À SCINTILLATIONS MÉDICAL

(*Illustration à la page 769*)

Ekco Electronics Ltd, Southend-on-Sea, Essex

Le compteur à scintillations léger, type N618, vient de s'ajouter à la série des appareils nucléaires pour usages médicaux, construits par la société Ekco Electronics Ltd. Cet appareil, qui est muni d'un blindage directionnel type N651, est enfermé dans un coffret scellé en acier inoxydable permettant de le stériliser à froid. La sortie du tube photomultiplicateur est transmise, à travers un étage à charge cathodique, à un câble fixé de façon permanente, à partir duquel des connexions peuvent être effectuées avec un compteur ou une échelle appropriés. Il peut être relié directement au compteur EKCO type N600, mais un amplificateur à part, tel que l'amplificateur EKCO type N640, est nécessaire lorsqu'il est relié à un compteur ou à une échelle nécessitant une entrée minimum de 5 V.

Le blindage directionnel retient le compteur à scintillations par un simple collier de serrage et on peut fixer sur ce blindage n'importe lequel des trois collimateurs courants pour obtenir un choix d'angles d'admission. Afin de faciliter le repositionnement précis de l'appareil, par rapport au malade, il a également été doté d'un télémètre à tâche lumineuse coïncidente, à réglage continu.

EE 16 765 *pour plus amples renseignements*

APPAREILS DE MESURE PORTATIFS DE PRÉCISION

(Illustration à la page 769)

Metrix Instruments Ltd, 54 Victoria Road, Surbiton, Surrey

Une série complète d'ampèremètres, de voltmètres et de wattmètres portatifs de précision METRAWATT vient d'être lancée sur le marché par la société Metrix Instruments Ltd.

Cette série d'appareils, de construction particulièrement robuste, a été étudiée en vue de répondre à la demande d'instruments sensibles et d'une précision supérieure à celle des modèles universels d'établi, pour les mesures d'essai et de laboratoire où il est peu pratique ou économique d'employer un instrument ordinaire, coûteux et souvent délicat. Cette nouvelle série peut également être employée comme étalons de transfert.

Les instruments à cadre mobile ont une précision de $\pm 0{,}5\%$ et comprennent un voltmètre allant de 6 à 600 V c.c. en cinq gammes. Quatre instruments de mesure de courant comportent des gammes de 20 μA à 2 mA, de 0,5 à 50 mA, de 02 à 20 A et de 0 à 60 mV, pour des champs de shunt allant jusqu'à 2 500 A.

Il existe aussi une série analogue de modèles à fer mobile, d'une précision de $\pm 0{,}5\%$ c.a. et $\pm 1\%$ c.c.

Ces instruments sont fabriqués par la société Metrawatt A.G., Nuremberg.

EE 16 766 *pour plus amples renseignements*

Résumés des Principaux Articles

Un système d'essais par programme par D. W. Bradfield, A. M. East et H. F. Rourke

Résumé de l'article aux pages 714 à 721

Cet article décrit comment on peut vérifier des équipements électroniques complexes, en une fraction du temps qu'il faudrait pour exécuter des essais manuels, en établissant un programme de la succession d'essais au moyen d'unisélecteurs. Les résultats de ces essais sont indiqués par affichage "marche-arrêt". Le programme dont il est question ici fut établi pour vérifier le système de puissance d'un engin téléguidé, mais les principes sont généralement applicables à une étendue d'équipements.

Le contrôle automatique des distorsions dans les faisceaux hertziens à modulation de fréquence à large bande par J. Tolman

Résumé de l'article aux pages 722 à 725

L'auteur traite du problème de la distorsion produite dans les faisceaux hertziens à fréquence modulée à large bande, par les caractéristiques non-uniformes de freinage des faisceaux dans les amplificateurs haute fréquence. Il décrit un égalisateur automatique expérimental qui corrige ce freinage et il donne les résultats obtenus avec un circuit à 600 voies. Il indique, enfin, des méthodes pour corriger des distorsions d'ordres plus élevés.

Un appareil pour mesurer les forces de coupage d'outils par J. Goldberg

Résumé de l'article aux pages 726 à 732

L'appareil décrit dans cet article a été calculé pour la mesure successive de composantes de forces de coupage sur des outils de tour pendant les opérations d'usinage. Les éléments sensibles à la force sont des extensomètres à resistance bobinée, employés conjointement avec un dynamomètre d'effort, à deux composantes. Les circuits de mesure sont formés de transformateurs à enroulements couplés inductivement, d'un rapport 1 : 1, qui alimentent des séries d'extensomètres reliés de façon appropriée. Chaque composante de force est mesurée par un montage en pont séparé, le circuit d'entrée et de sortie de chaque pont étant branché dans la succession voulue, entre une source d'attaque à courant alternatif et un détecteur commun. Par ce moyen, l'effet de la résistance de contact de commutation est rendu insignifiant.

On emploie la détection sensible aux phases pour observer les changements de la composante résistive de déséquilibre des ponts, ainsi que les signes de contrainte.

Le système de présentation consiste en un enregistreur à plume contenant un dispositif pour lever et abaisser la plume, afin de partager la durée d'enregistrement entre le nombre de voies employées.

Des lectures totales, correspondant à des efforts de 100, 250 et 500 kgs. environ, sont fournies par l'appareil qui peut être étalonné directement par charge statique. L'article conclut pas des données de forces de coupage obtenues avec l'appareil.

Analyse à large bande de circuits de déphasage à tube électronique par L. J. Giacoletto

Résumé de l'article aux pages 733 à 735

Il s'agit ici d'un circuit de déphasage constituant la contrepartie électronique d'un transformateur à sortie équilibrée, ayant deux sorties dont les tensions sont idéalement égales en magnitude, mais à 180° de séparation en phase. Cette opération idéale est réalisée aux basses fréquences. Des capacités entre électrodes modifient les rapports de sortie aux fréquences supérieures et c'est précisément cet effet qu'analyse l'auteur. Après avoir examiné tant les rapports de dessin précis que les rapports approximatifs, il les compare avec les données mesurées d'un circuit-type. Il en résulte qu'un fonctionnement quasi idéal, à des fréquences assez élevées, peut être obtenu en employant des impédances appropriées de charge égale. On peut encore améliorer ce fonctionnement dans une certaine mesure en ajoutant un petit inducteur de pointe dans l'impédance anodique.

Une commande automatique de caméra par H. R. A. Townsend

Résumé de l'article aux pages 736 à 739

Un tube à cathode froide miniature pour commander un relais à plusieurs contacts constitue un élément fort souple au moyen duquel on peut construire un contrôleur de fonctionnement. La commande d'appareil dont il est question ici comporte un simple circuit à retard, un circuit de synchronisation, un circuit d'amorçage à distance, ainsi qu'un compteur binaire, également construit avec des tubes à cathode froide miniature. Cette commande exécute le cadencement relativement complexe nécessaire pour obtenir une photographie entièrement automatique des traces présentées sur un oscillographe à rayons cathodiques.

La simulation électronique d'une charge appliquée à un muscle d'insecte par K. E. Machin

Résumé de l'article aux pages 740 à 744

Certains muscles d'insectes se contractent rythmiquement lorsqu'ils sont chargés de manière appropriée. L'oscillation est commandée dans une grande mesure par les paramètres mécaniques de la charge. On vient de mettre au point un appareil qui simule électroniquement la masse, la viscosité et la raideur de la charge et qu'on peut donc faire varier facilement. Le rapport entre la force et la longueur d'un muscle oscillant peut être reproduit sur un tube cathodique tandis que la fréquence d'oscillation et la sortie de puissance sont présentés sur des compteurs. L'appareil peut être employé aussi pour mesurer directement l'impédance mécanique d'un muscle en fonction de la fréquence.

Un simple stabilisateur de courant pour électro-aimants par M. H. N. Potok

Résumé de l'article aux pages 745 à 746

Pour certaines expériences dont il est fait mention dans cet article, il a fallu fournir un champ magnétique exempt d'ondulations et de dérive lente, l'électro-aimant utilisé à cet effet nécessitant environ 10A à une tension de 100V. Les diverses méthodes d'alimentation de l'électro-aimant, ainsi que le mode de rotation employé, sont brièvement décrits. Le courant est fourni par une amplidyne, l'ondulation est réduite par des transistors et la dérive lente est évitée par l'emploi d'un amplificateur à courant continu.

La synthèse des filtres hautes fréquences à transistors par L. M. Vallese

Résumé de l'article aux pages 748 à 752

Cet article traite des méthodes de synthèse de filtres h.f. à transistors, basées sur l'utilisation de la transformation du réseau équivalant. Aux fins de simplification, la réaction intrinsèque des transistors, qui est fort peu importante dans les éléments à base diffuse, a été négligée. Deux structures à échelles réglementaires, du type hybride-π et hybride-T respectivement, sont étudiées pour la synthèse de paires à pôles composés; celles-ci sont mises en cascade par (et, en certains cas, sans) intercalation d'étages intermédiaires. L'auteur donne des exemples de filtre vidéo et passe-bande.

Variation de la conductance d'entrée d'un transistor à jonction, à base mise à la terre, et les applications pratiques de cette variation

par S. Deb et J. K. Sen

Résumé de l'article aux pages 753 à 755

Cet article traite du problème de la variation de la conductance d'entrée d'un transistor à jonction à base mise à la terre en fonction du courant direct d'entrée. Des moyens y sont suggérés en vue de maintenir cette variation linéaire dans une gamme étendue de valeurs de courant, ainsi que pour réduire le glissement thermique. Sur le plan pratique, une description est donnée des performances d'un modulateur d'amplitude et d'un multiplicateur analogique utilisant un transistor selon les méthodes indiquées.

Système à mémoire en ferrite à action rapide par C. J. Quartly

Résumé de l'article aux pages 756 à 758

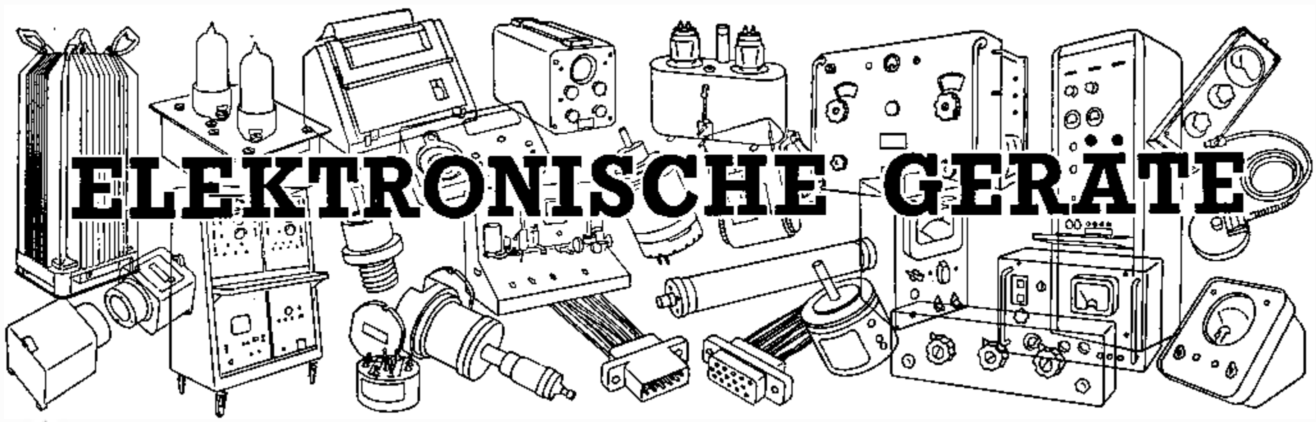
L'auteur décrit un système à mémoire en ferrite, à cadre, qui obvie à la vitesse limitée imposée par le matériau de noyau lorsqu'il est employé dans une mémoire conventionnelle avec choix d'entraînement coïncident. Chaque nombre est emmagasiné, sous forme de différence de niveaux de flux, dans deux noyaux qui ne sont que partiellement branchés pendant le processus d'écriture. Des mesures effectuées sur une paire de noyaux montrent comment la sortie varie pendant la lecture en fonction de la durée de l'impulsion d'écriture et de l'amplitude, ainsi qu'en fonction de l'amplitude de l'impulsion numérique. L'application de ce principe, avec choix de mots extérieur, permet de réaliser une durée cyclique de lecture/écriture de 0,5 microsecondes dans une petite mémoire.

Prises entièrement métalliques à l'épreuve du feu pour vides élevés par N. W. Robinson

Résumé de l'article aux pages 759 à 760

En raison du nombre croissant de tubes d'hyperfréquences et d'intensification d'images, il devient nécessaire de créer des vides de l'ordre de 10^{-10} mm. Hg. Ceci peut être réalisé selon un système décrit par Alpert, dont un élément essentiel est une prise d'isolement non-graisseuse. Cet article décrit une prise entièrement métallique à l'épreuve du feu, répondant à cette condition, dont le rendement est également donné.

ELEKTRONISCHE GERÄTE



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 766 bis 769

GERMANIUM-LEISTUNGSTRANSISTOR

(*Abbildung Seite 766*)

Sylvania-Thorn Colour Television Laboratories Ltd, Great Cambridge Road, Enfield, Middlesex

Sylvania-Thorn Colour Television Laboratories Ltd haben eine neue Serie legierter npn Ge-Transistoren für 60 V 6 A eingeführt. Sie wurden für Leistungsschalten entwickelt und werden voraussichtlich Konstrukteuren elektronischer Geräte für geringere Kosten als bisher mehr Spielraum für Anwendung der npn Transistortechnik geben.

Der GT422 kann, neben anderen Verwendungszwecken, mit einem pnp Transistor zusammen als ergänztes Paar für transformatorenlose Ausgangsstufen von Radio- oder Servoverstärkern, sowie in Serienspannungsreglern eingesetzt werden. Andere Typen dieser Serie sind die GT424 (60 V 3 A), GT425 (36 V 6 A) und GT426 (36 V 3 A). Die maximale Kristalltemperatur ist 85°C.

EE 16 751 *für weitere Einzelheiten*

KERNTECHNISCHE ZÄHLGERÄTE

(*Abbildung Seite 766*)

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

Mit der „1800“ Serie hat Isotope Developments Ltd. vor kurzem ihre neueste Geräteserie angekündigt. Es ist eine Serie preisgünstiger kerntechnischer Zählgeräte, die für Operationsgenauigkeit, volle Anpassungsfähigkeit, sowie einfache Wartung und Service entwickelt wurde. Die Geräte werden in formschönen Gehäusen mit Kippstand geliefert, sodass sie entweder einzeln oder aufeinander gestellt benutzt werden können. Auf Wunsch können sie auch leicht für Gestellmontage umgebaut werden. Da alle Zuleitungen von hinten angeschlossen werden und das Gehäuse von vorn montiert ist, kann das Innere des Gerätes ohne Lösung der Kabel zugänglich gemacht werden.

Einsteckplatinen mit gedruckten Schaltungen finden in der Serie weitgehende Verwendung, und identische Netzteile werden in alle Geräte eingebaut, um den Ersatzteilbedarf und den Service zu vereinfachen. Jedes Gerät wird nach einem 100 Stunden-Test mit vollen Bedienungsanweisungen und einer Ersatzteilausrüstung versandt.

Die ersten drei Geräte dieser Serie sind nunmehr lieferbar als Untersetzer Type 1800, Hochspannungsquelle Type 1820 und Verstärker mit Impulsselektion Type 1830.

Die folgenden Instrumente sind für die Serie vorgesehen: ein Ratemeter Type 1810, ein Vorsatz-Untersetzer Type 1850 und ein Potentiometergerät Type 1840. Weitere Geräte werden später angekündigt.

EE 16 752 *für weitere Einzelheiten*

DEKATRON NETZGERÄT

(*Abbildung Seite 766*)

Sandmar Electronic Products, 68 Nina Drive, New Moston, Manchester, 10, Lancashire

Die Netzgeräte wurden besonders zur Stromversorgung von Dekatrons und den zugehörigen Schaltungen entwickelt. Sie werden entweder als Chassisbaustein oder auf Frontplatte für 483 mm (19 Zoll) Gestelle geliefert.

Das Gerät liefert drei stabilisierte Ausgangsspannungen: die erste gibt bis zu 10 mA +450 V Gleichstrom mit weniger als 0.02% Restwelligkeit ab; die zweite +300 V Gleichstrom bis 200 mA mit weniger als 0,005% Restwelligkeit und reguliert innerhalb 0,1% für Belastungsschwankungen von 0 ... 200 mA; die dritte liefert −100 V Gleichstrom für 10 mA. Ausserdem sind unstabilisierte Wechselspannungen von 4 oder 6,3 V bei 3 A vorgesehen. Das Gerät ist so konstruiert, dass beim Einschalten die Spannungen folgerichtig abgegeben werden.

EE 16 753 *für weitere Einzelheiten*

KASSETTE FÜR ENDLOSES MAGNETBAND

(*Abbildung Seite 766*)

Guy's Calculating Machines Ltd, Truro Road, London, N.22

Die „Brittape“ Kassette für endloses Magnetband stellt jedes flache Laufwerk für 178 mm (7 Zoll) Spulen auf Dauerbetrieb um. Die Kassette wird auf eine der Laufwerk-Antriebsspindeln aufgesetzt und kann bis zu 61 m dauernd umlaufendes 6,35 mm (1/4 Zoll) Standard- oder Langspiel-Magnettonband aufnehmen.

In Betrieb wird die Kassette durch das Magnetband, das in üblicher Weise durch das bestehende Laufwerk gezogen wird, angetrieben.

Durch die „Brittape“ Kassette können Ansagen oder Signalzusammenstellungen bis zu 22 Minuten dauernd zirkulieren, oder das Band kann kontinuierlich laufen, um Einschwingvorgänge aufzuzeichnen, die in regellosen Intervallen auftreten.

EE 16 754 *für weitere Einzelheiten*

NETZGERÄT FÜR DEHNUNGSMESSER

(*Abbildung Seite 767*)

Solartron Electronic Group Ltd, Thames Ditton, Surrey

Dieses sechskanalige Netzgerät für Dehnungsmesser Type AS958 ist durchgehend mit Transistoren bestückt und wurde hauptsächlich für Verwendung mit Solartron Druckumwandlern eingeführt.

Ein übliches stabilisiertes Netzgerät kann viel unerwünschtes Brummen hervorrufen, wenn ein Umwandler direkt an einen Hochleistungsverstärker wie z.B. den Solartron AA900 Datenverstärker angeschlossen wird. Modell AS958 ist insofern eine Neuentwicklung, als ein Spezialtransformator die unerwünschte Übertragung zu weniger als 0,1% des Skalenvollanschlages verringert.

Das Gerät wurde zum Anschluss an Brücken mit einem Eingangswiderstand von 350Ω entwickelt, kann jedoch auch für Widerstandsbrücken höherer Impedanz Verwendung finden. Dadurch ist das Gerät zur Speisung fast aller verbundenen oder unverbundenen Dehnungsmessertypen geeignet.

Des Netzgerät hat sechs einzelne stabilisierte Stromversorgungs-Bausteine, deren Schaltungen keine gemeinsamen Punkte haben. Jeder Baustein kann 5 V regulierte Nenngleichspannung bis zu 20 A abgeben. Mit einem „Empfindlichkeits“-Regler kann die Ausgangsspannung ±5% geändert werden.

Ein eingebauter „Ausgleichs"-Regler kann die Nullstellung der Brücke bis zu 10% des Vollausschlages korrigieren. Die Empfindlichkeitsregelung kann für jeden Kanal auf der Frontplatte vorgenommen und die jeweilige Einstellung konzentrisch verriegelt werden.

EE 16 755 *für weitere Einzelheiten*

HF INDUKTIONS-ERWÄRMUNGSANLAGEN

(Abbildung Seite 767)

Pye Ltd, R.F. Heating Division, 28 James Street, Cambridge

Die zweite Induktions-Erwärmungsanlage in der neuen Pye-Serie ist das Tischmodell R.F.2 mit einer mittelgrossen Ausgangsimpedanz und einer Dauerleistung von 2 kW bei 2 MHz. Das Aluminiumgehäuse mit Sicherheitsschaltern und abnehmbaren Seitenwänden ist in Gestaltung dem ersten Modell ähnlich. Die Gesamtabmessungen des Gerätes sind 616 mm hoch, 445 mm breit und 540 mm tief. Die Anlage ist unabhängig, mit eingebauter automatischer Programmuhr mit Rückstellung, Schütze für die Röhren und Wasserdruckschalter zum Schutz der Anlage. Fernsteuerung ist vorgesehen, und am Ende der Erwärmungsperiode kann ein Signal die Abschreckung oder irgendeine andere Operation einleiten. Die Anlage arbeitet mit einer einphasigen Netzspannung von 180 ... 250 V 50 Hz und verbraucht bei Vollbelastung 3,1 kW.

EE 16 756 *für weitere Einzelheiten*

TRAGBARER UNTERGRUNDMONITOR

(Abbildung Seite 767)

Plessey Nucleonics Ltd, Weedon Road, Northampton

Plessey Nucleonics haben einen neuen, leichten, tragbaren Monitor zur Messung von Untergrundstrahlungen entworfen und konstruiert.

Dieser als Type PNI1702 eingeführte Monitor wurde entwickelt, um den Bedarf für ein genaues Instrument, das jede Untergrundstrahlung schnell und leicht misst zu decken. Der Nachweis erfolgt durch ein Geiger-Müller Halogenzählrohr, die Messung in einem Transistor-Ratemeter mit direkter Anzeige in Dosisrate auf der 63.5 mm Skala eines Drehspulinstrumentes. Das Messgerät hat einen Bereich von 0 ... 30 mr/h auf zwei quasilogarithmischen Skalen mit einer min. Skaleneichung von 30 μr/h. Das Gerät wird von einer 9 V Trockenbatterie gespeist und wiegt 3.6 kg.

EE 16 757 *für weitere Einzelheiten*

BANDFILTER

(Abbildung Seite 767)

Salford Electrical Instruments Ltd, Times Mill, Heywood, Lancashire

Salford Electrical Instruments Ltd, eine Tochtergesellschaft der General Electric Co. Ltd, hat zwei neue serienmässig hergestellte Filter eingeführt, die die Durchlasskennlinien für UKW-Fernmeldeempfänger bestimmen.

Die als 455KBP50 und 455KBP25 bekannten Typen entsprechen den für Selektion in den Fernfunknormen der britischen Postverwaltung für ZF festgelegten Kanalabständen von 25 kHz und 50 kHz.

Das Filter ist vollständig vergossen und in einen Metallbehälter eingebaut, wobei die Eingangs- und Ausgangsleitungen zu je zwei kurzen Säulen an jeder Seite des Behälters herausgeführt werden.

Die Befestigung ist so angeordnet, dass das Filter in einen Chassisausschnitt montiert werden kann, sodass die Filteranschlüsse dem Rest der Schaltung so nahe wie möglich liegen.

EE 16 758 *für weitere Einzelheiten*

ELEKTRONISCHE BAUSTEINE

(Abbildung Seite 768)

Mullard Equipment Ltd, Mullard House, Torrington Place, London, W.C.1

Mullard Equipment hat zur Konstruktion von elektronischen Steuerungssystemen, Digitalrechnerelementen, Datenverarbeitungsanlagen, Messystemen und anderen Geräten, die aus Digitalschaltungen aufgebaut sind, eine umfassende Serie elektronischer Standard-Bausteine eingeführt.

Diese Bausteine bestehen aus vollständigen, völlig entwickelten Schaltungen, die garantiert in jeder gewünschten Zusammenstellung bestens arbeiten und der schärfsten Prüfung für Betriebssicherheit unter den schwierigsten Betriebsbedingungen unterliegen.

Zwei Bausteinserien sind lieferbar: „Norbits" und „Combi-Elements". Norbits wurden hauptsächlich für Industriefirmen entwickelt, die ihre eigenen Steuerfolgesysteme schnell, wirtschaftlich und ohne ausgebildete elektronische Ingenieure bauen wollen. Sie können auch als Standard-Bausteine von Firmen verwandt werden, die sich auf Steuerungssysteme spezialisiert haben.

Die Serie umfasst alle zur Konstruktion von Werkzeugmaschinen-und Transferstrassensteuerung, Werkschutzsystemen, automatischer Liftsteuerung, Programmfolgesystemen usw. erforderlichen Schaltungen.

„Combi-Elements" sind besonders für Hersteller elektronischer Systeme und Anlagen gedacht und geben ihnen eine Serie von Standard-Bausteinen, die üblicherweise verwandte Digitalschaltungen darstellen. Abgesehen von der Verkürzung von Entwicklungszeit und -kosten, führt ihre Verwendung zu wesentlichen Raumersparnissen und gewährt grössere Anpassungsfähigkeit beim Bau grosser Systeme und Anlagen.

Zur Erreichung höchster Zuverlässigkeit sind sowohl Norbits als auch Combi-Elements voll transistorisiert. Transistoren ermöglichen auch den Einbau der verschiedenen Schaltungen in gedrungene Kunststoffbehälter kleinster Abmessungen, verringern den Energieverbrauch, vereinfachen Wärmeableitungsprobleme und vermeiden Hochspannungsquellen.

Norbit-Bausteine wurden besonders entwickelt, um dem Betriebsingenieur einer Firma, der nur eine sehr allgemeine Kenntnis der Elektronik hat, die ganze Planung und Konstruktion eines industriellen Programm-Steuerungssystems zu ermöglichen.

Die Bausteine haben lange Zuleitungen, sodass Verbindungen ohne Lötstützpunkte und zusätzliche Verdrahtung unter Verwendung geklemmter, nicht gelöteter Verbindungspunkte hergestellt werden können. Kanaleinheiten mit Klemmleisten für Eingangs- und Ausgangsverbindungen mit Einlochbefestigung für mehrere Norbits sind lieferbar.

Der grundsätzliche Norbit-Baustein ist ein Transistor-Schaltelement, das im Interesse der Einfachheit und Anpassungsfähigkeit, und entsprechend der Verdrahtung als UND- oder als ODER-Schaltung eingesetzt wird. Nach Erhalt der Information von den Machinengrenzschaltern oder von anderen Eingangsquellen führt der Baustein die erforderlichen logischen Operationen aus und liefert folgerichtige Signale an ein geeignetes Ausgangsgerät, das seinerseits Umwandler erregt, die die Maschinenfunktion bestimmen oder überwachen. Die Bausteinabmessungen sind 63,5 × 39,7 × 12,7 mm. Ein einziger Kanal kann 12 Norbits aufnehmen.

Mehrere Ausgangsgeräte mit verschiedenen Nennleistungen bis zu 150 W zur Betätigung von Schaltwerken, elektromagnetischen Kupplungen, Solenoids usw., sowie ein Zähler und eine Uhr von 0 ... 60 s und andere Bausteine sind lieferbar.

Die Combi-Elements-Serie umfasst Flip-Flop und Mehrfachvibratorschaltungen, Impulsformer, Torschaltungen, Verstärker und Emitterverstärkerschaltungen.

Jede Combi-Element-Schaltung ist luftdicht in einem kleinen Kunststoffbehälter von nur 53,3 × 23,9 × 9,9 mm verschlossen und für schnelle und leichte Montage auf Platinen gedruckter Schaltungen konstruiert, wobei dieselben Methoden wie für die Montage von Kondensatoren und Widerständen Anwendung finden. Die Bausteine sind raummässig so gestaltet, dass die grösstmöglichste Anzahl auf einer Platine untergebracht werden kann.

Dadurch werden Montagezeiten und -kosten reduziert, und einzelne Bausteine oder ganze Platinen können leicht ausgewechselt werden, wodurch eine grössere Anpassungsfähigkeit in der Zusammenstellung komplizierterer Systeme und Anlagen und eine Vereinfachung für Abänderungen erreicht wird.

Diese Schaltungsserie stützt sich auf das Gleichstromtor, und im allgemeinen werden Signale von Baustein zu Baustein durch Gleichspannungen übertragen. Flip-Flop und andere Mehrfachvibratoren können entweder durch ein

Gleichstromsignal oder eine Spannungsstufe (Wechselstromsignal) getriggert werden. Dem Einsatz entsprechend können die Bausteine für asynchrone oder synchrone Logik verwandt werden, wobei die letzere Zeitimpulse benutzt.

EE 16 759 *für weitere Einzelheiten*

TEMPERATURREGLER

(Abbildung Seite 768)

Fielden Electronics Ltd, Wythenshawe, Manchester, Lancashire

Fielden kündigt das erste Gerät in der neuen "Bikini"-Reihe transistorisierter Instrumente an. Es ist ein Präzisions-Temperaturregler mit einem Anzeigedifferenzial von 0,5°C, hoher Eichgenauigkeit und Dauerstabilität.

Das Instrument ist in ein kleines Messgerätgehäuse eingebaut und nimmt auf der Frontplatte nur einen Platz von 114 mm Durchmesser ein. Die 229 mm lange Skala ist direkt in Celsius oder Fahrenheit geeicht, und der Zeiger wird durch einen kleinen Knopf im Instrumentmittelpunkt auf die gewünschte Regeltemperatur eingestellt. Eine rote Lampe, die sowohl von der Seite als auch von vorn sichtbar ist, zeigt die Regeltätigkeit an. Das Messgerätgehäuse ist vollständig abgedichtet und enthält keine empfindlichen beweglichen Teile, sodass das Instrument in jeder Lage, z.B. auf einer geneigten Pultplatte, montiert werden kann. Die Regelung kann bei jeder Temperatur zwischen −200°C ... +500°C erfolgen. 12 Standard Eichungen sind vorgesehen.

Als Temperaturmesspunkt dient ein Platin-Widerstandselement, das in einer widerstandsfähigen Hülle aus rostfreiem Stahl von 6,35 mm Durchmesser untergebracht ist. Zwischen dem Messpunkt und dem Instrument kann gewöhnliches Kabel bis zu 90 m Länge benutzt werden, ohne dass Fehler auftreten.

Der Regler schaltet elektrisch oder pneumatisch ein - aus. Die elektrischen Kontakte schalten 5 A 250 V um, und das pneumatische Ventil benötigt einen Eingabedruck von 1,2 ... 1,4 kg/cm².

EE 16 760 *für weitere Einzelheiten*

DIFFERENTIAL-DRUCKWANDLER

(Abbildung Seite 768)

Southern Instruments Ltd, Frimley Road, Camberley, Surrey

Southern Instruments kann jetzt einen Differentialdruckwandler G255 liefern, dessen Fühlerelement veränderlicher Induktivität mit ihrem FM Verstärkersystem Verwendung findet.

Druck kann entweder auf jede einzelne Seite oder beide Seiten der Beryllumkupfermembran ausgeübt werden; die Membranbewegung wird von einer in den Wandler eingegossenen Spule entdeckt. Die Ausgangsspannung des FM Vorverstärkers ist der Druckdifferenz auf der Membranseite genau proportional. Die Geradlinigkeit über den Nennbereich ist besser als 3%. Der Wandler ist für Druckbereiche von 0,07-0-0,77 kg/cm² bis 70-0-70 kg/cm² lieferbar.

EE 16 761 *für weitere Einzelheiten*

GESTEUERTE SILIZIUMGLEICHRICHTER

(Abbildung Seite 769)

Associated Electrical Industries Ltd, Carholme Road, Lincoln

Dieses neue Element in pnpn Form kann als zwei Transistoren betrachtet werden, die durch eine gemeinsame Kollektorfläche gekoppelt sind; es ist das Festkörperequivalent des Thyratrons oder des gittergesteuerten Quecksilberdampfgleichrichters. Es hat in jeder Richtung eine hohe Impedanz von mehreren Megohm, kann jedoch durch ein kleines Signal zur dritten Elektrode in einen Zustand geringer Impedanz in einer Richtung getriggert werden. Es hat die üblichen Festkörpervorteile: kleine Grösse, keine Heizfadenversorgung und hohe Betriebsleistung.

Die Anwendungsgebiete sind denen herkömmlicher Vorrichtungen ähnlich, z.B. Motorenregelung, Zündung von Ignitrons, Gleichstrom/Wechselstrom-Umwandlung, Frequenzumwandlung, phasengesteuerte Gleichstromquellen, statisches Schalten usw.

In der Sperrichtung sind die Spannung/Strom-Kennlinien fast mit denen eines einfachen Silizium-Gleichrichters übereinstimmend. In der Durchlassrichtung sind die Kennlinien bis zur Durchschlagsspannung ein Spiegelbild der Sperrkennlinien. Wenn ein kleiner Strom von 10 ... 100 mA zur Steuerelektrode fliesst, während die Durchlassspannung niedriger als die Durchschlagsspannung ist, so wird der Gleichrichter leitend bleiben, bis die angelegte Spannung fast null ist. Der Gleichrichterausgang wird nach einer bekannten Methode durch Phasenverschiebung des Triggerstroms gegenüber der an den Gleichrichter gelegten Wechselspannung gesteuert.

Zur Zeit sind Gleichrichter im Bereich 25 ... 300 V Spitzensperrspannung mit Durchlassströmen von 5 und 10 A lieferbar.

EE 16 762 *für weitere Einzelheiten*

OSZILLATOR-VERSTÄRKER

(Abbildung Seite 769)

Airmec Ltd, High Wycombe, Buckinghamshire

Der Oszillator-Verstärker Type 254 wurde hauptsächlich als Antriebsquelle für kleine Servometer, Vibratoren und Geräte für ungenormte Frequenzen entwickelt.

Das Gerät besteht aus einem 150 W-Hochleistungsverstärker, der entweder durch ein Fremdsignal oder einen eingebauten Oszillator mit stufenloser Abstimmung von 30 Hz ... 30 kHz angetrieben wird. Der Verstärker kann auch vom Oszillator isoliert und als Rundspruchgerät oder Modulator eines Senders eingesetzt werden.

Die Ausgangsstufe besteht aus vier EL34 Röhren in Parallel-Gegentaktschaltung und einem Ausgangsumsetzer mit mehreren Sekundärwicklungen, die in Serien-Parallelschaltung einem weiten Belastungsbereich angepasst werden können. Die Ausgangsspannung wird durch ein eingebautes Messgerät überwacht. Die Röhren werden durch Sicherungen und Funkenstrecken voll geschützt. Eine Signallampe zeigt Übersteuerung oder unzulässigen Anodenstrom an. Ausserdem sind die Röhrenanoden durch Frontplattengitter sichtbar.

Die Oszillatorschaltung besteht aus einem thermistorstabilisierten Wienbrücken-Gegenkopplungsnetzwerk, dessen Frequenz auf einer klaren Grossskala mit ±2% Genauigkeit angezeigt wird. Für Fremdsignalantrieb wird die Oszillatorröhre als Vorverstärker geschaltet und hat dann bei voller Ausgangsleistung 0,1 V Empfindlichkeit.

EE 16 763 *für weitere Einzelheiten*

DRUCKENDE ZÄHLER

(Abbildung Seite 769)

The Stonebridge Electrical Co. Ltd, 6 Queen Anne's Gate, London, S.W.1

Die Firma hat druckende Zähler eingeführt, die auf Papierrollen Zählungs- und Zeitangaben drucken und in Industrie und Forschung weitgehende Verwendung finden können.

Sie sind als Einzel- oder Doppelgeräte lieferbar. Normalerweise sind die Einzelgeräte sechsstellige, summierende Zähler mit elektrischem Impuls, Druck und Löschung. Doppelgeräte können entweder mit zwei summierenden Zählern dieser Art oder mit einem summierenden und einem Gerät, das Datum und Zeit angibt, ausgerüstet werden. Die Impuls-Höchstgeschwindigkeit ist 10 pro Sekunde, Drucken wird in 100 ms und Löschen in 500 ms ausgeführt.

Modelle sind für die verschiedensten Gleich- und Wechselspannungen lieferbar.

EE 16 764 *für weitere Einzelheiten*

MEDIZINISCHER SZINTILLATIONSZÄHLER

(Abbildung Seite 769)

Ekco Electronics Ltd, Southend-on-Sea, Essex

Ekco Electronics Ltd haben vor kurzem ihr Kerngeräteprogramm für medizinische Zwecke durch Einführung des leichten Szintillationszählers N618 mit Richtungsschild N651 erweitert. Der Einbau des Zählers in einen abgedichteten Behälter aus rostfreiem Stahl lässt Kaltsterilisation zu. Der Ausgang einer Fotomultiplierröhre wird über eine Kathodenverstärkerstufe einem ständig angeschlossenen Kabel zugeleitet, mit Hilfe dessen ein Ratemeter oder Untersetzer angeschlossen werden kann. Der Zähler kann direkt an das Ekco Ratemeter N600 angelegt werden, benötigt jedoch einen getrennten Verstärker, wie z.B. Ekco N640, wenn er mit einem üblichen Ratemeter oder Untersetzer mit Mindesteingangsspannung von 5 V Verwendung finden soll.

Eine einfache Schelle befestigt den Szintillationszähler im Richtungsschild. Durch Einbau eines der drei Standard-Kollimatoren können verschiedene Aufnahmewinkel erzielt werden. Um den Zähler in Bezug auf den Patienten wieder in die genaue Lage bringen zu können, ist ein stufenlos einstellbarer Entfernungsmesser mit Koinzidenzlichtpunkten eingebaut.

EE 16 765 *für weitere Einzelheiten*

TRAGBARE PRÄZISIONS-MESSGERÄTE

(*Abbildung Seite 769*)

Metrix Instruments Ltd, 54 Victoria Road, Surbiton, Surrey

Metrix Instruments Ltd bringt neuerdings eine vollständige Auswahl tragbarer „Metrawatt" Präzisions-Strom-, Spannungs- und Leistungsmesser auf den Markt.

Diese sehr sorgfältig konstruierten Messgeräte wurden für Prüf- und Labormessungen entwickelt, die empfindliche Instrumente einer Genauigkeit erfordern, die über der der üblichen Universalinstrumente liegt, für die jedoch teure und oft empfindliche Normalinstrumente entweder unbequem oder unwirtschaftlich sind. Die Serie ist auch für Übertragungsnormen geeignet.

Die Drehspulinstrumente haben eine Genauigkeit von $\pm 0{,}5\,\%$. In dieser Serie ist auch ein Gleichspannungsmesser von 6 ... 600 V in fünf Bereichen lieferbar. Vier Strommesser für die Bereiche 20 µA ... 2 mA, 0,5 ... 50 mA, 0,2 ... 20 A und 0 ... 60 mV ermöglichen Messungen bis 2 500 A mit Nebenschlusswiderständen.

Eine ähnliche Serie von Weicheiseninstrumenten ist mit $\pm 0{,}5\,\%$ Wechselstrom- und $\pm 1\,\%$ Gleichstromgenauigkeit lieferbar.

Die Messgeräte werden von Metrawatt A.G. Nürnberg hergestellt.

EE 16 766 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein programmierter Tester von D. W. Bradfield, A. M. East und H. F. Rourke

Zusammenfassung des Beitrages auf Seite 714-721

Dieser Beitrag beschreibt, wie komplexe elektronische Geräte durch Programmierung der Prüffolge mittels eines nur in einer Richtung laufenden Wählers in einem Bruchteil der Zeit geprüft werden können, die manuelle Tests in Anspruch nehmen würden. Die Testresultate werden als "gut-fehlerhaft" angezeigt. Das beschriebene Programm wurde zur Prüfung des Stromversorgungssystems einer Fernlenkwaffe entwickelt, aber die Prinzipien sind im allgemeinen für eine grosse Auswahl von Geräten gültig.

Automatische Verzerrungsregelung in Mikrowellen-FM-Breitbandfunkstrecken von J. Tolman

Zusammenfassung des Beitrages auf Seite 722-725

Der Beitrag bespricht Verzerrungen, die in Mikrowellen-FM-Breitbandfunkstrecken durch ungleichförmige Gruppenverzögerungs-Kennlinien der HF Verstärker auftreten. Eine versuchsmässige automatische Korrektur der Gruppenverzögerungs-Steilheit wird beschrieben und die für eine 600-Kanal-Schaltung erzielten Resultate angegeben. Korrekturmethoden für Verzerrungen höherer Ordnung werden angedeutet.

Ein Gerät zur Messung der Werkzeug-Schneidkräfte von J. Goldberg

Zusammenfassung des Beitrages auf Seite 726-732

Dieser Beitrag beschreibt ein Gerät, das für die aufeinander folgenden Messungen der Schneidkraft-Komponenten eines Drehwerkzeuges während des Arbeitsablaufes entwickelt wurde. Als Kräftefühler werden Widerstand-Spannungsmesser mit einem Zweikraftdynamometer eingesetzt. In den Messschaltungen werden Übersetzer mit induktiver Kupplung und einem Verhältnis 1 : 1 benutzt, die Spannungsmessersysteme in geeigneter Weise speisen. Jede Kraftkomponente wird mittels einer getrennten Brückenschaltung gemessen, wobei die Eingangs- und Ausgangskreise der Brücken in der gewünschten Folge zwischen einer einzelnen Wechselstrom-Antriebsquelle und einem gemeinsamen Anzeiger umgeschaltet werden. Durch diese Anordnung wird der Einfluss des Kontaktwiderstandes auf die Genauigkeit der Messschleife unbedeutend.

Zur Beobachtung der Änderung der Widerstandskomponente der Brückenunsymmetrie, sowie des Vorzeichens der Spannung, wird eine phasenempfindliche Anzeige benutzt.

Zur Aufzeichnung wird ein Schreiber verwandt, in dem eine eingebaute Vorrichtung die Feder hebt oder senkt, damit die Aufzeichnungszeit zwischen mehreren in Gebrauch befindlichen Kanälen geteilt werden kann.

Für Kräfte von 90,7 kg, 223 kg und 453,5 kg sind Vollausschlagsanzeigen vorgesehen, und das Gerät kann durch statische Belastung direkt geeicht werden. Aufzeichnungen der mit dem Gerät gemessenen Schneidkräfte werden gezeigt.

Breitbandanalyse einer Phasenteilerschaltung mit Röhren von L. J. Giacoletto

Zusammenfassung des Beitrages auf Seite 733-735

Eine Phasenteilerschaltung ist das elektronische Gegenstück zu einem symmetrischen Ausgangsübertrager und hat zwei Ausgänge, deren Spannungen idealerweise gleicher Grösse, aber um 180° gegeneinander phasenverschoben sind. Diese idealen Bedingungen können für niedrige Frequenzen erreicht werden. Für höhere Frequenzen ändern Röhrenkapazitäten die Ausgangsbeziehungen, und dieser Einfluss wird analysiert. Genaue sowie annähernde Berechnungsbeziehungen werden entwickelt und mit Messwerten typischer Schaltungen verglichen. Es ergibt sich, dass bei Verwendung geeigneter gleicher Belastungsimpedanzen nahezu ideale Betriebsbedingungen bis zu mässig hohen Frequenzen erreicht werden können. Die Verwendung von Belastungsimpedanzen, die nicht ganz gleich sind, und Einschaltung einer kleinen Abstimmspule in die Anodenbelastung kann die Wirkungsweise etwas verbessern.

Eine automatische Kamerasteuerung von H. R. A. Townsend

Zusammenfassung des Beitrages auf Seite 736-739

Ein durch eine Kaltkathodenröhre gesteuertes Mehrfachkontaktrelais ist ein sehr vielseitiger Baustein für eine Programmsteuerung. Die beschriebene Kamerasteuerung enthält eine einfache Verzögerungsschaltung, eine Synchronisierschaltung, eine ferngesteuerte Zündschaltung, sowie einen auch mit Kaltkathodenröhren bestückten binären Zähler. Das Gerät führt die verhältnismässig schwierige Steuerfolge durch, die für die vollautomatische Fotografie der Elektronenstrahlspuren eines Oszillografen erforderlich ist.

Die elektronische Nachbildung der Muskelbeanspruchung eines Insektes von K. E. Machin

Zusammenfassung des Beitrages auf Seite 740-744

Wenn gewisse Muskeln von Insekten in bestimmter Weise beansprucht werden, ziehen sie sich rythmisch zusammen; die Oszillation hängt wesentlich von den mechanischen Parametern der Beanspruchung ab. In dem entwickelten Apparat werden Masse, Zähigkeit und Steifheit der Beanspruchung elektronisch nachgebildet und können daher leicht geändert werden. Für einen oszillierenden Muskel kann die Beziehung zwischen Kraft und Länge auf einer Bildröhre dargestellt werden, während die Oszillationsfrequenz und Leistungsabgabe auf Messgeräten angezeigt wird. Der Apparat kann auch zur direkten Messung der mechanischen Impedanz eines Muskels als Funktion der Frequenz eingesetzt werden.

Ein einfacher Stromstabilisator für Elektromagnete von M. H. N. Potok

Zusammenfassung des Beitrages auf Seite 745-746

Für bestimmte Versuche war ein magnetisches Feld erforderlich, das von Restwelligkeit und langsamer Drift frei war. Der in Frage stehende Elektromagnet benötigte ungefähr 10 A bei 100 V. Die verschiedenen Versorgungsmethoden werden in dem Beitrag kurz besprochen und die angenommene Lösung beschrieben. Als Stromquelle wird ein Amplidyne gewählt, Transistoren reduzieren die Restwelligkeit, und ein Gleichstromverstärker korrigiert die langsame Drift.

HF Transistor-Filter Synthese von L. M. Vallese

Zusammenfassung des Beitrages auf Seite 748-752

Synthesierungsmethoden für HF Transistor-Filter werden entwickelt, die sich auf die Transformation des äquivalenten Vorwärts-Netzwerkes stützen. Zur Vereinfachung wurde die transistoreigene Gegenkopplung vernachlässigt, da sie für Elemente mit diffundierter Basis sehr gering ist. Zwei kanonische Leiterstrukturen, und zwar des Zwitter-π-Types bzw. des Zwitter-T-Types, werden für die Synthese komplexer Polpaare in Betracht gezogen; sie werden unter (und in gewissen Fällen ohne) Zwischenschaltung von Pufferstufen in Kaskade arrangiert. Bild- und Bandfilter werden als Beispiel gezeigt.

Anderungen des Eingangsleitwertes eines Flächentransistors in Basisschaltung und ihre praktische Verwertung von S. Deb und J. K. Sen

Zusammenfassung des Beitrages auf Seite 753-755

Der Beitrag untersucht Anderungen des Eingangsleitwertes von Flächentransistoren in Basisschaltung in Abhängigkeit vom Eingangsgleichstrom. Methoden werden vorgeschlagen, die die Anderungen über einen breiten Stromwertbereich geradlinig halten und auch die thermische Drift verringern. Die Leistungen eines Amplitudenmodulators und eines Analogmultiplikators werden beschrieben, in denen ein Transistor unter diesen Bedingungen arbeitet.

Ein Ferrit-Schnellspeichersystem von C. J. Quartly

Zusammenfassung des Beitrages auf Seite 756-758

Ein Speichersystem wird beschrieben, das Ferritkerne mit rechteckiger Hystereseschleife benutzt und Geschwindigkeitsbeschränkungen überwindet, die dem Kernmaterial eigen sind, wenn es in den üblichen Speichern mit elektronischem Koordinatenschalter verwandt wird. Jede Zahl wird als Differenz des Flusses zweier Kerne gespeichert, die während des Einschreibens nur teilweise geschaltet werden. Messungen an einem Kernpaar zeigen, wie die Ausgangsleistung während des Lesens sich in Abhängigkeit von der Einschreibeimpulsdauer und -amplitude und der Zahlenimpulsamplitude ändert. Durch Anwendung dieses Prinzips mit Aussenwortabfrage kann in einem kleinen Speicher eine Zeit von 0,5 μs für den Lese/Einschreibevorgang erreicht werden.

Ausbackbare Ganzmetallhähne für Höchstvakuum von N. W. Robinson

Zusammenfassung des Beitrages auf Seite 759-760

Für eine wachsende Anzahl von Mikrowellen- und Bildverstärkerröhren wird es notwendig, ein Vakuum in der Grössenordnung von 10^{-10} Torr zu schaffen. Das kann durch eine von Alpert beschriebene Methode erreicht werden, für die ein fettloser Isolationshahn erforderlich ist. In diesem Beitrag wird ein ausbackbarer Ganzmetallhahn beschrieben und seine Leistung untersucht.