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Commentary

IT is only ten years since the first transatlantic telephone cable was laid, but already a strong body of opinion asserts that cable systems are obsolescent and will shortly be replaced by communication satellites. Conversely the protagonists of cable systems, although probably in a minority, hold that, for many years to come cables will be a more viable proposition for long distance communications. In fact neither of these extreme views is likely to be completely true; with the large growth in international trade and the rapid development of hitherto backward countries, the demand for high grade voice and data transmission channels appears to be insatiable and for many years to come it is likely that cables and satellites will live happily together and be complementary to one another. It is also likely that it will be many years before, either jointly or separately, they oust h.f. radio from the very prominent position which it holds; it is still the cheapest form of long distance communication and carries a considerable proportion of world-wide traffic.

The launching of the first earth satellite coincided, more or less, with the laying of the first transatlantic telephone cables and during the intervening ten years the growth of satellite technology has been tremendous. Even apart from the more glamorous aspects of manned-space flights, it has undoubtedly provided the most dramatic scientific advances ever known to man. Reliable long distance communication via satellites, of both voice and television is now an accepted part of daily life. So far, of course, satellites have provided only point-to-point communication and to realize their full potential it is necessary that they should become multiple access devices, that is they must be able to work between several ground stations simultaneously. When this is accomplished satellites will provide an easy and economic means of long distance communication to remote and comparatively sparsely populated areas where cables would not be economic. Plans for multiple access satellites, using frequency division multiplex, are already well under way and flight trials will soon be carried out in preparation for the Intelsat-3 multiple access global system due to be put into operation by Comsat early in 1968.

During this same period of time the technical progress in long distance underwater cables and repeaters has also been great. Only eight years ago 80 voice-grade channels was considered a large capacity whereas today 360 channels is accepted as standard while an American prototype cable provides 720 channels and one British development aims at 1280 channels. Interest in cables has recently been stimulated by the South African plans to provide a missing link in a scheme to connect cables round the world. Fulfilment of these plans depends partly on the success of negotiations for a cable between South Africa and India which could then be joined to the 6000 mile cable which is shortly to be laid between South Africa and Portugal. If South Africa's plans are successful the only section missing in the world system would be the comparatively short distance from Penang to Madras.

From a purely technical viewpoint, satellites and cables have their own peculiar advantages and disadvantages. One of the outstanding advantages of satellites is the very wide bandwidth and the consequent high number of voice-grade channels, which they can provide. In terms of television or high speed data transmission it is an advantage which is unlikely to be even remotely approached by cables. Unfortunately, this advantage is, in a way the satellites own worst enemy for the spectrum crowding which is now so severe in the h.f. bands may well, in time, have its counterpart, in satellite communications. In fact, commercial communications satellites are at present operating in the same frequency bands as line-of-sight relay systems on the ground and this, apart from other considerations, limits the amount of power that may be allowed to reach the ground from satellites in these shared frequency bands and prevents the use of higher power satellites which are already technically feasible. Since cables do not depend on radiated signals, this problem does not arise and thus the number of cables which can be employed is unlimited.

In those cases where either satellites or cables can be employed with equal competence, as will probably be the case in the majority of communication applications, the choice will be made almost entirely on economic grounds. The costs for satellite schemes are, as yet, rather uncertain. It has been said that the cost for 'Early Bird' is about £35 a circuit mile compared with £128 for an older submarine cable and £30 for the newer 360 channel circuits. Larger capacity cables are expected to bring prices down to £5 a mile which is similar to estimates for the Intelsat-3 global satellite scheme. In other words the choice of medium may well be governed by the type of traffic to be carried and the density of traffic to any given area.

Fortunately, Britain is in a reasonably favourable position whichever medium is adopted. British designed and manufactured underwater cables, repeaters and terminal equipment are in use throughout the world and with reasonable effort and forethought the U.K. can certainly maintain its pre-eminent position in the world markets for this class of equipment. So far as satellites systems are concerned Britain is not, and nor is she likely to be for some time, in a position to supply the actual satellite part of the system but the really big business will be in ground stations and here the experience gained from the satellite terminal station at Goonhilly Downs and from radio telescopes places this country in a very favourable position to emerge as one of the world's leading suppliers.

Electronically Controlled Starting of Aircraft Gas Turbines

By K. F. Bacon*, A.M.I.E.E.

In modern aircraft, it is necessary to relieve the crew of routine tasks. This article describes how one of these tasks, engine starting, can be performed automatically using an electronic control unit. The problems encountered in the design of one of these units are described in some detail and the ways in which they were overcome explained. The circuits of the final unit and its operation in the aircraft system are described and the possible applications to turbine starting outlined.

(Voir page 552 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 559)

With the increased complexity of modern aircraft there is an increasing tendency to relieve the pilot of the more routine tasks involved in flying the aircraft and to rely on automatic means of achieving the same purpose. Generally, it is found that by doing so, the particular tasks can be performed more efficiently and rapidly. Most of the attention has been directed to the flight phase including take-off and landing, but the pilot has many jobs to do before the aircraft is airborne. One of these jobs is the starting of the engines.

In military aircraft, it is important that the engines be started as rapidly as possible to enable take-off to be achieved within seconds of the pilot entering the cockpit. The pilot must, therefore, be relieved of as many routine tasks as possible to enable him to carry out a rapid pre-flight check.

In civil aviation, the large capital investment involved in the purchase of modern aircraft is increasing the pressures on airlines to turn them round at airports as rapidly as possible. This also leads to consideration of ways in which the time for pre-flight checks can be minimized.

This article describes ways in which automatic engine starting can be achieved and outlines the methods used in one piece of equipment designed for a specific aircraft.

Information Required

To enable the engine starting sequence to be controlled, certain information from the engines, starting means, aircraft and cockpit is required. The information from the engines is usually rotational speed and signals from warning devices such as fire detection, oil pressure switches, etc. Signals from the starting device can include centrifugal switches and limit switches, while the aircraft provides, for example, signals derived from the oleo indicating that the aircraft is airborne. The actual control of the starting sequence is initiated and stopped from cockpit switches operated by the pilot.

Speed Signals

Measurement of the speed of the engine is necessary so that the sequence of events involved in its starting can be closely controlled. For example, the ignition circuits and starting device may have to be turned off at preselected speeds. The source of this speed signal will vary from engine to engine depending on the engine accessories arrangement. On some engines there is a generator for the provision of electrical supplies and on others there may be a tachogenerator. To avoid problems of different tolerances etc., it is best that these devices generate a.c. so that their frequency can be used to determine engine speed. On some engines, a small signal is derived from a magnetic pick-up which effectively counts the teeth on one of the engine gear wheels. This pick-up was originally designed for use with engine control units which control the engine when it is running.

The main disadvantage of these devices is that their output is small during engine run up (particularly with the last device, where it is in the millivolt region) and this involves problems in the design of the measuring circuit which will be dealt with later. In general, it is true to say that to avoid troubles with noise and pick-up, the signal should be as large as possible. Troubles have been experienced in the past due to the use of speed sensing devices designed for use when the engine is running for engine starting, when the signal levels are very low.

Warning Circuits

The main warning signals required are fire and oil pressure. The fire warning may be derived from a number of devices such as fire-wires or thermocouples. Arrangements sometimes have to be made to change the datum of the warning during the starting sequence since the jet pipe temperature allowed may be higher during starting than it is during normal running. Warning of low oil pressure is usually obtained from conventional oil pressure switches.

Other Signals

Scattered about the aircraft is a number of actuators, limit switches, relays, push-buttons, toggle switches and other mechanical contacts which provide information necessary for the control of engine starting. Most of these are electromechanical and provide signals derived from the opening or closing of contacts. If these are fed into electronic circuits, then care must be taken to avoid problems due to noise and contact bounce.

Outputs to be Controlled

Any control unit for engine starting will have to feed signals to a large variety of control devices. These will include hydraulic, electrical, pneumatic and fuel valves and solenoids. Most of them involve the control of energy flow by means of an electrical solenoid but others include ignitors, starting cartridges, lamps and motors. Very few of these loads are purely resistive and most of them are regenerative. All loads possessing inductance generate large reverse voltages across their switching device when switched off. The lamps suffer from another disadvantage in that the inrush current is many times greater than the normal load current. The ignitor load, as well as being inductive, can also be oscillatory and may require current flow in two directions. All these factors must be seriously considered when the choice of switching device is made.

Practical Realization

This section will deal generally with the possible ways

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of achieving the results required from the information available to the control unit and more specifically with examples taken from such a unit designed for a two-engined military aircraft.

SPEED SIGNALS

As described previously, the speed of the engine is usually determined from the frequency of an electrical generator of some form which rotates at a speed proportional to engine speed. This generator should preferably provide a substantial voltage to feed to the control unit. The reason for this is obvious from the following considerations. The voltage of the generator will rise with speed and the speed measuring circuit may be required to accept a large voltage range if speeds at widely different levels are to be sensed. For example, the signal may be taken from a 200V, 400c/s alternator which has an automatic voltage regulator. At low speed the output from this alternator will be only the residual voltage of the machine (which may be less than 10V at half full speed). At a certain speed the regulator will take over and the voltage will rise instantly to 200V. Thus, if it is required to sense 10 per cent and 90 per cent full speed, the measuring circuit may have to accept voltage input ranges of more than 100 : 1. Because of this, and the necessity to ignore harmonics, some form of squarer or limiter is required on the input. Some help may be obtained from a low-pass filter since, at first, voltage is proportional to frequency. When the regulator takes over, however, this is no longer true.

On some engines, it may not be possible to use a 200V generator and one may have to use the output from a gear-tooth type counter. This produces an output which increases with engine speed to a level of about 1V at 2 to 3kc/s. Because of saturation the output flattens above a certain speed and may even fall slightly as the speed increases further. In the example quoted, this device consisted of a magnetic probe so installed that the magnetic circuit was influenced by the teeth of an engine gear wheel. It was originally designed for an engine control unit so that only the output at high engine speeds was considered. As the aircraft and engine developed, other coils were added to the probe to provide speed signals for such instruments as a crash recorder, thrust meter and starting unit. It rapidly became evident that such a device has serious short-comings when used as the input to the starting unit. The other devices connected to the probe are essentially analogue devices either for speed indication or servo control. The engine starting unit, however, is 'digital' in that when a certain speed is reached, the starting cycle is stopped. The electrical environment is very noisy since the control unit may be some distance from the engines, and interference can occur from many sources such as the 400c/s electrical system, ignitor noise etc. In one particular instance, interference of 800c/s was traced to a poorly smoothed d.c. signal fed into one of the coils of the sensing probes to check continuity. The affect of this noise is to cancel the starting cycle as soon as it begins. The low signal level of the probe simply aggravates the situation. Eventually, these problems were cured by great attention to details such as the starting unit input circuit, screening, earthing, sharing of the probe coils with other users and boosting its output to a more reasonable level.

This experience leads to the conclusion that a speed sensing device giving a reasonably large output is required for future systems.

In order to use the speed signal it must be converted to a d.c. level proportional to speed. It is necessary to square the input to the starting unit to produce a waveform independent of input waveform distortion. This waveform is then amplified before being applied to the frequency-to-d.c. convertor. The circuit for the input amplifier to the main engine speed sensing unit is shown in Fig. 1. The input transformer is fed from two identical separate coils on the one speed sensing probe. These are connected together at the amplifier to make a centre-tapped pair with the centre-tap earthed. This ensures that the failure of one coil or its leads only produces a fall in input to the amplifier and not a fatal loss (the system is designed to accept this without loss of performance). The probe unit is designed so that if one coil becomes short-circuited or has shorted turns, the output from the other coils on the probe does not fall sufficiently to cause failure.

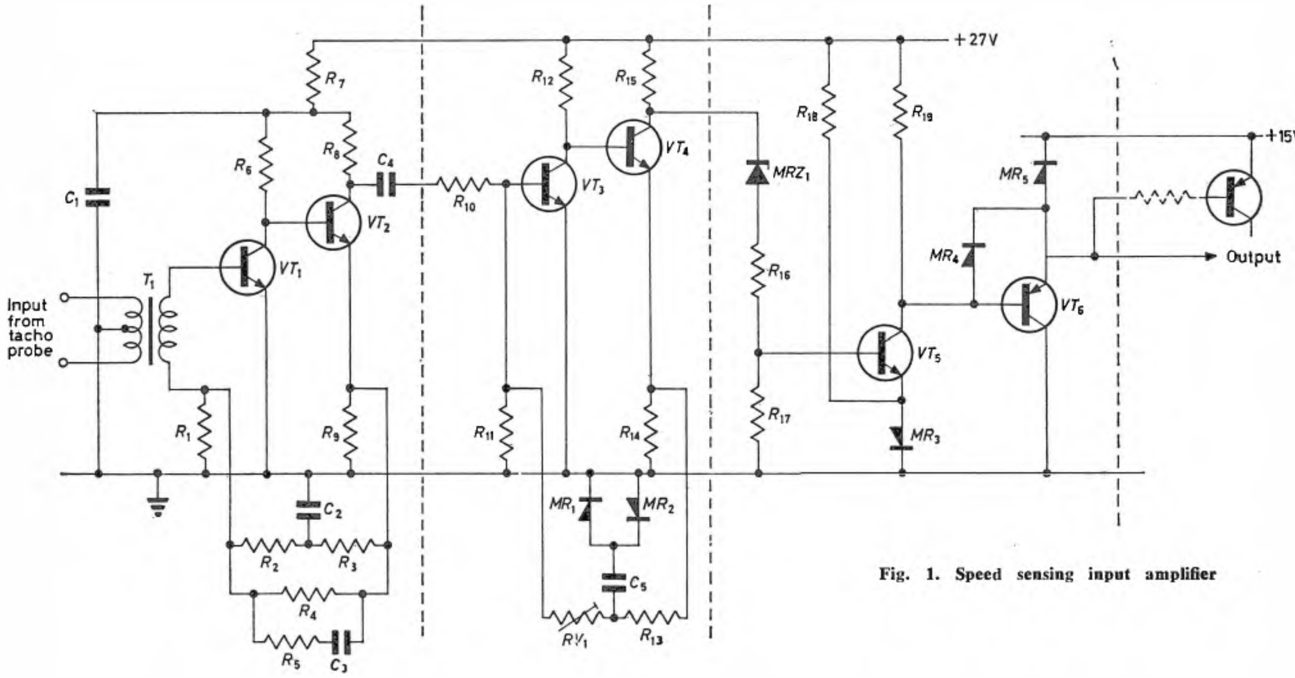


Fig. 1. Speed sensing input amplifier

The amplifier circuit consists of three sections:

(1) Frequency correcting amplifier.

(1) Squarer.

(3) Driver.

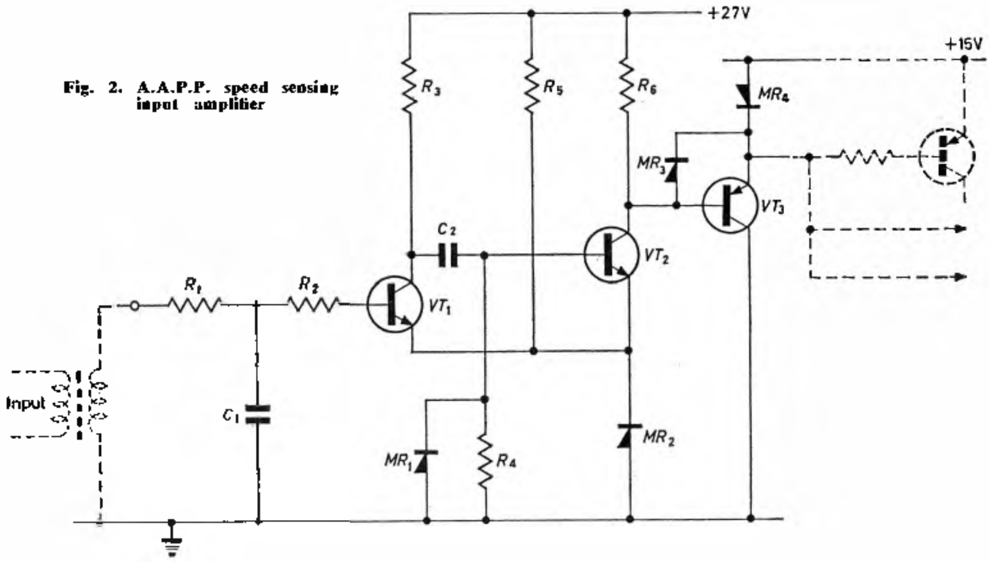


Fig. 2. A.A.P.P. speed sensing input amplifier

The frequency correcting amplifier changes the rising probe output into a nearly constant amplifier output. This ensures that high frequency noise is sharply attenuated and helps in the noise elimination problems. The circuit is a two-stage directly coupled amplifier with negative feedback. R_3 and R_3 provide feedback which is effective only at d.c. because of the decoupling effect of C_2. This ensures that the amplifier's operating point is stabilized and is independent of temperature. A.C. feedback is applied via R_4, R_5 and C_3 which ensure a gain fall-off of 6dB/octave. The output from this circuit is fed into the squaring circuit, which is again a two-stage directly coupled amplifier similar to the input stage. A large amount of feedback is applied via R_{13} and RV_1 so that the gain is very low. C_5 is ineffective at low signal levels because diodes. MR_1 and MR_2 are non-conducting. Thus, for small input signals very little output appears across R_{15}. When the signal increases to a level sufficient for MR_1 and MR_2 to conduct, the feedback circuit is decoupled and becomes ineffective. The gain of the amplifier then increases rapidly and produces a square-wave output. This threshold limiting device again helps to combat the noise problem. By the correct choice of fall-off characteristics and threshold levels, the output of the squarer can be muted until the speed of the engine, and thus the signal level, reaches a reasonable level. The speed sensing circuit is, therefore, effectively muted during the period when noise is most troublesome, that is with the engine ignitors operating and no input to the speed sensing circuits.

The d.c. operating point of the squarer is set up by RV_1 so that the output has a mark-to-space ratio of 2:1. This is because the frequency to d.c. convertor circuit which follows the amplifier operates most effectively with an unbalanced input. This will be described later. The output stage of the amplifier consists of an npn-pnp pair VT_5 and VT_6 which drive the frequency to d.c. convertor. Diodes MR_3, MR_4 and MR_5 provide cut-off bias to VT_5 and VT_6.

For sensing the speed of engines which possess a higher voltage generator, the simpler circuit of Fig. 2 is used. This provides a constant width pulse via the driver transistor VT_3. The pulse length is determined by the differentiating circuit C_2, R_4.

The output signals from these amplifiers are fed to diode pump frequency-to-d.c. convertors. In most cases a 5 per cent accuracy (over a large temperature range) is sufficient for main engine starting circuits. The complete circuit

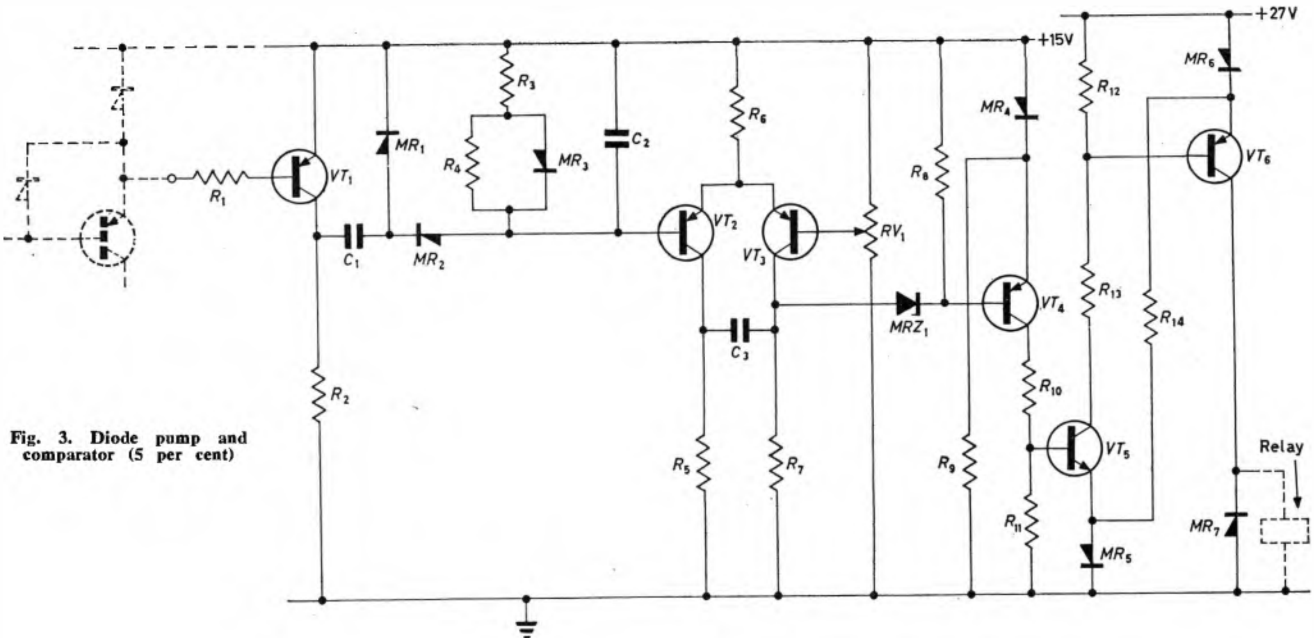


Fig. 3. Diode pump and comparator (5 per cent)

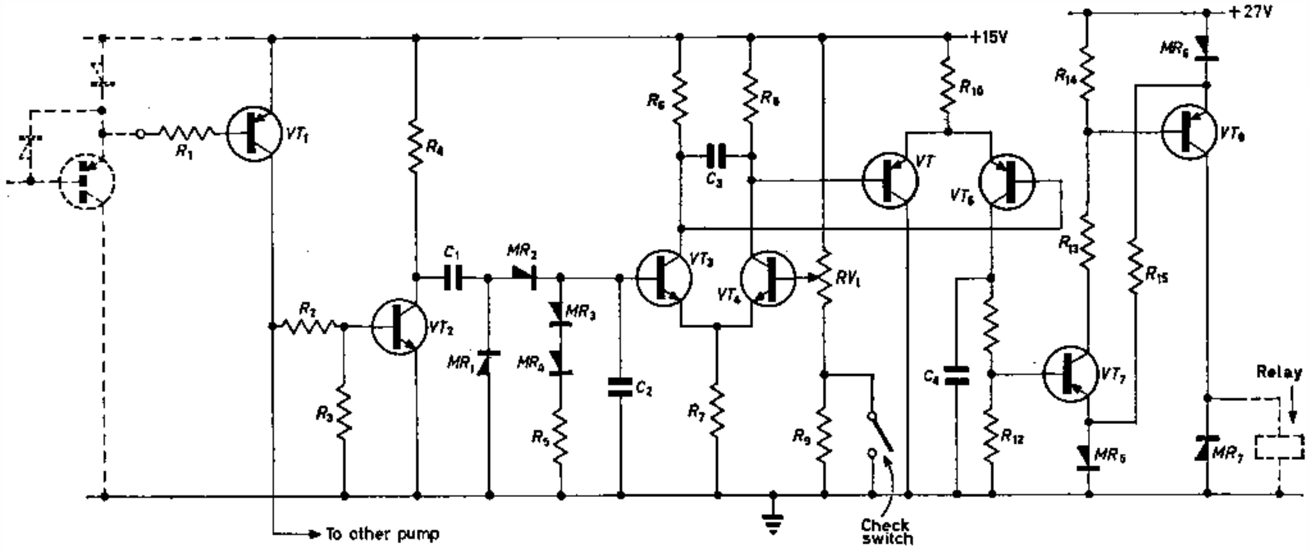


Fig. 4. Diode pump and comparator (1 per cent)

for a diode pump followed by comparator and output stage is shown in Fig. 3 The operation of the circuit is as follows: when VT_1 is cut-off C_2 charges via R_2, C_1 and MR_2. C_2 is very much larger than C_1 so that most of the supply voltage appears across C_1. When VT_1 conducts C_1 discharges via MR_1 since MR_2 is reverse biased by the charge on C_2. Therefore, when the next pulse arrives C_1 is discharged and ready for use again, and a charge builds up on C_2 due to the small increments of charge transferred from C_1. This continues until the leakage of charge via R_3 and R_4 equals the incoming average charge and the voltage across C_2 is thus dependant on the frequency of the incoming pulses. Diode MR_3 provides some compensation for the temperature drift of MR_1, MR_2 and the V_{ce} of VT_1. The optimum length of input pulse is determined mainly by the charging time-constant C_1R_2 and must be above a definite minimum to ensure correct operation.

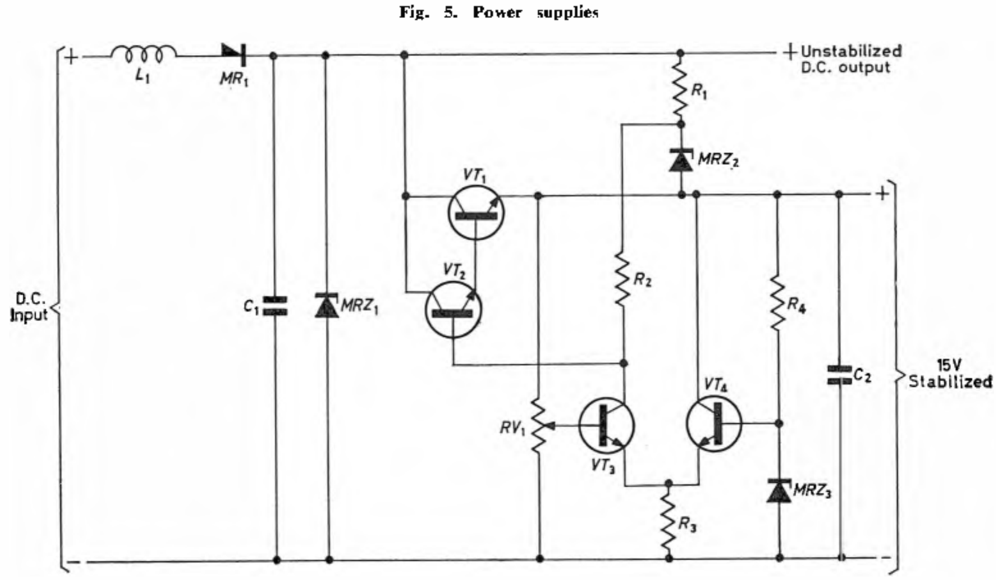
The output voltage is sensed by the long-tailed-pair VT_2 and VT_3. When the voltage reaches the same value as that on the base of VT_3, VT_2 conducts and VT_3 switches off. When this occurs VT_4, VT_5 and VT_6 conduct providing sufficient current to operate the output relay. C_3 removes the ripple which is obviously present due to the operation of the diode pump.

The voltage supply rail (+15V) to the diode pump and comparator is stabilized to give the required accuracy. By supplying these two circuits from the same supply some measure of compensation is inherent. The stabilizer is described later. For some systems a much more accurate speed sensing circuit is required. For example, standby power is provided in some aircraft by an auxiliary airborne power plant (a.a.p.p.) which serves to provide electrical, air and sometimes hydraulic power for ground use. This is usually a small high-speed turbine which may need special protective measures. Because of their high operating speed, a sensitive and accurate overspeed device is used to switch off the engine in case of governor failure. The accuracy of this circuit is often required to be better than 1 per cent. The circuit shown in Fig. 4 is an example.

The operation of this circuit is similar in principle to the one shown in Fig. 4 but with several important differences of detail. To obtain a switching transistor with a low and stable V_{co} it is necessary to use an npn device. It is, therefore, necessary to rearrange the circuit and drive the switch from an invertor VT_1. To improve the temperature stability, two diodes MR_3 and MR_4 are used, and the comparator consists of two long-tailed-pair stages. Since greater sensitivity is required of this comparator, it is necessary to filter the ripple more effectively, and C_4 is added.

So that the accuracy and operation of the circuit can be checked without overspeeding the engine, a check switch

Fig. 5. Power supplies



is provided across R_9. This changes the datum when closed so that the circuit just operates at normal engine speed. The same drive circuit can be used to drive other diode pumps and comparators for different speeds. Attempts to operate several comparators from a common diode pump have not proved successful because of the loading effect of long-tailed pair amplifiers when switched.

The power supply for this and the other circuits of a typical unit is shown in Fig. 5. This is designed to accept the normal 28V d.c. aircraft supply and provides an unstabilized output which varies with this supply (and may be quite low when starting) and a +15V stabilized output via a conventional series stabilizer. L_1, C_1 and MRZ_1 are for positive transient and spike suppression while MR_1 removes negative excursions.

Time Delays

The main engines are started by the ignition of the fuel/air mixture by electrical ignitors while the engines are rotated by a starter motor of some kind. The ignitors are, therefore, operated for a time period slightly greater than that for a normal start, the timer being cancelled as soon as the engine reaches a speed higher than the light-up speed. As the engine takes longer to start at low temperatures it is advantageous if the time period is made inversely proportional to temperature. The circuit for such a time delay is shown in Fig. 6. Operation of the time delay is initiated by the start push-button. Under quiescent conditions, the relay is de-energized and supplies base current to VT_1 via R_4. This makes VT_1 conduct and ensures that C_1 is charged to +15V. Current via R_5 and MR_1 switches on VT_3, rendering VT_4, VT_5 and VT_6 non-conductive. This is the stable 'off' condition. Closure of the push-button causes the relay to conduct and VT_1 and VT_3 to switch off because of their positive base voltages. Since VT_3 is off, VT_4, VT_5 and VT_6 conduct and hold the relay closed when the push-button is released. C_1 discharges via R_2 and the potential at the emitter of VT_2 follows the capacitor potential. VT_3 and VT_4 comprise a Schmitt trigger circuit which fires when the base potential of VT_3 reaches the critical voltage (set by RV_1). As soon as this happens, VT_4, VT_5 and VT_6 switch off and the circuit reverts to its stable 'off' state.

A wet tantalum capacitor is used for C_1 and as its leakage current increases at a greater rate than its capacitance with increasing temperature, the timer period falls at high temperatures. The timer period can be cancelled by a contact from a speed sensing circuit across R_{12}. This switches off VT_6 to de-energize the relay. Thus, this circuit is effectively a monostable circuit with a very large time-constant. Delays of up to 80sec were achieved in the unit used as an example. By the choice of the charging components, circuits for different delays can be achieved.

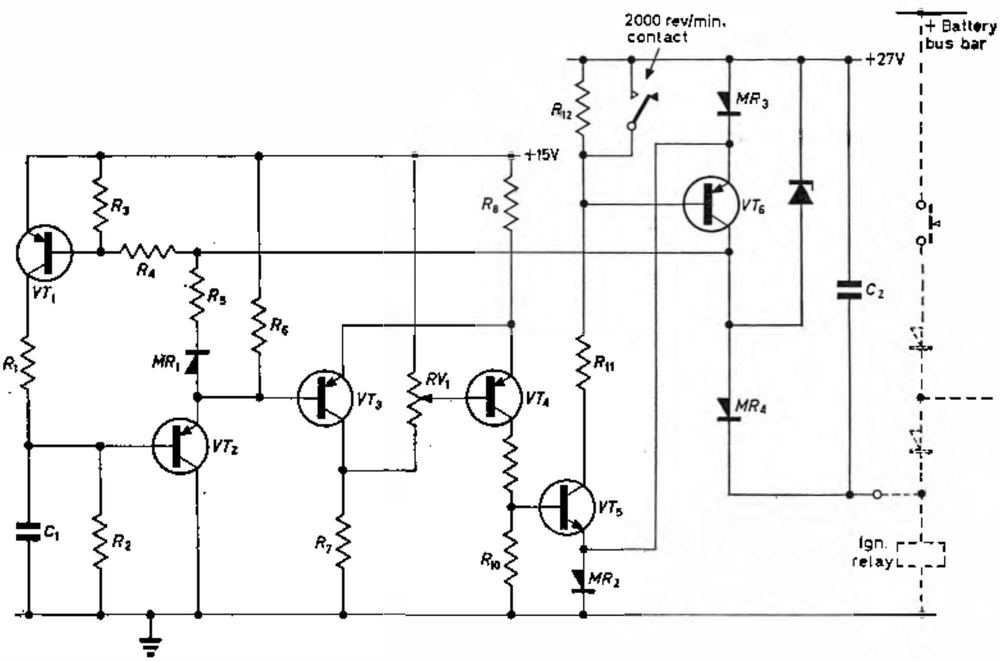


Fig. 6. Ignition time delay circuit

Output Circuits

In earlier examples of engine starting units, semiconductor circuits were used throughout for logic and output devices. With the state of semiconductors at the time, especially with regard to thyristors and power transistors, this meant that the overload capacity of the output devices was low. When such a unit was installed in engine test rigs this lack of capacity became a distinct disadvantage. Engine test rigs are not normally fully equipped with the aircraft standard solenoids, valves, actuators, etc. and are serviced by normal works electricians. The cost of engine down-time is very high on these rigs and it is quite possible for a 28V coil having the rated nominal current of 0·5A to be replaced with one of a previous type having a nominal current of 3A in the urgency of the moment. This could be disastrous to an engine starting unit with an output device designed for 0·5A. This damage occurred a number of times in the early days of the design and it was realized that this represented a fundamental weakness in the design although it was caused by improper use. The output devices must have sufficient overload capacity to enable them to be protected by normal fuses. It is quite common in aircraft practice for a minimum size of fuse to be specified to avoid mechanical failure due to vibration. This value may be high enough to make the use of semiconductor output devices unacceptable. At the present state of the art, this decision could be reconsidered due to the availability of larger transistors with higher overload capacities. The use of thyristors is also possible although the problems of commutation on the d.c. power supply and false triggering due to the noisy environment must be faced.

Because of all these factors, it was decided, for the example described, to change the design. Fortunately, at about this time, the design of aircraft relays had changed with several radical improvements. Among these was the miniature relay with separate sealed compartments for coil and contacts. (The cost of these devices, incidentally, quite often exceeds that of a transistor circuit). It was decided, therefore, that the new engine starting unit would be based upon these relays. Because of the multiplicity of output contacts, it was also decided that any logical operations would also be performed with the relay contacts. Speed sensing, power supplies and time delays, etc., would still be performed by the semiconductor circuits already described. This results in a hybrid circuit which may offend the purist but which provides a much more versatile unit, since it was designed to have redundant relay contacts to allow for easy modification which is usually necessary for such a unit designed in parallel with engines and aircraft. In any case, the input functions are mainly mechanical switches which would form the weakest link in the system.

System Design

Before describing the design of the engine starting system, it is necessary to describe briefly the various parts of the particular aircraft associated with the starting of the engines.

The main engines are fitted with a gearbox to drive the auxiliaries such as the hydraulics. This gearbox drives a constant speed unit for rotating the aircraft alternator to provide a constant 400c/s output. The constant speed unit can simply be described as a differential gearbox with the main shaft driven by the engine and with an alternator and an air motor on each side of the differential. When the engine speed is low, the air motor is driven by engine air to increase the alternator speed and when the engine speed is high the air motor is run as a compressor to decrease the alternator speed. By locking the alternator shaft and driving the air motor from an external air source, the engine shaft may be turned for starting. By locking the engine shaft, the alternator can be driven on the ground from the same source to test the electrical system. Thus, the control of this constant speed drive and gearbox is closely associated with the starting cycle.

The fuel in the engines is ignited by standard ignitor units which provide a high voltage discharge across points in the engine. These provide a major source of electrical noise and their connexions and cables must be carefully sited in the aircraft to reduce the attendant problems.

Also on this particular aircraft was an airborne auxiliary power plant consisting of a small jet engine lowered from the fuselage and operated on the ground to provide a source of compressed air for engine starting at airfields where an external source of air is unavailable. Connected to this engine is a small alternator for providing a small amount of electrical power for essential purposes. The starting of this engine is initiated by cartridges.

Protection of this engine against overspeed is vital since it is surrounded by fuel tanks. Because it normally runs at a high speed which is quite close to the bursting speed, a high accuracy speed sensing circuit is used which takes its input from the alternator.

Single Engine Start (Main Engines)

The starting of the main engines is based on two relays, the c.s.d. (constant speed drive) and ignition relays. The circuit associated with these relays is shown in Fig. 7. For a normal start the e.m.s. (Engine Motor Switch) is closed, the flt/m.o. (Flight/Motor Over) switch is in the flt position and the start/relight button pressed. The c.s.d. and ignition relays operate, the latter being cancelled after approximately 65sec or when the engine reaches 3 100 rev/min, and their contacts operate the starting circuits. The ignition circuits can also be operated from the double engine start circuits which are described later. For removing fuel from the engine after a 'wet' start, the flt/m.o. switch is placed in the m.o. position. Thus, only the c.s.d. relay is operated and the engine is turned without the fuel being ignited.

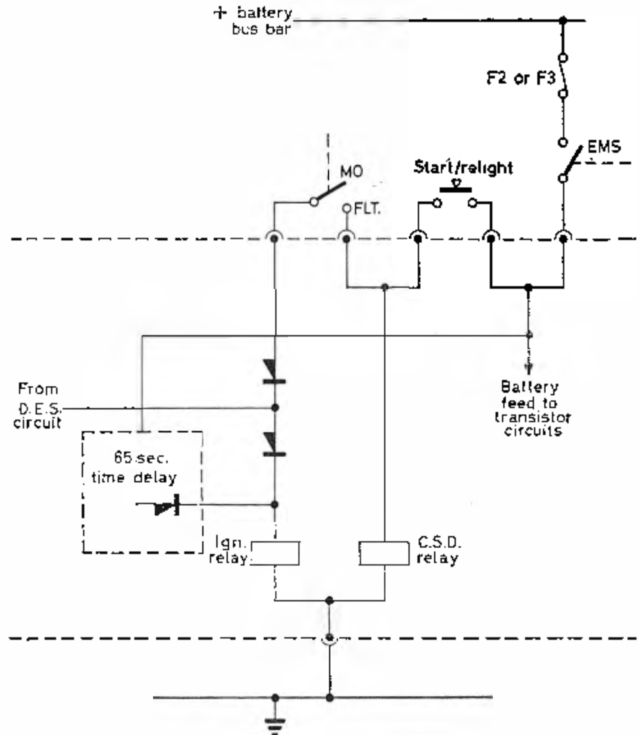


Fig. 7. Start relay circuits

The operation of the main starting sequence is shown in Fig. 8, which shows one half of the unit. The other half for starting the other main engine is identical. When the ignition relay operates power is fed via contacts *1C* and *1D* to the double ignitor unit. These circuits are duplicated for safety reasons.

An output is supplied to the air starter trolley (an external source of compressed air) via the ignitor relay contacts *1B* (from either engine). In addition, the air starter trolley is operated from a feed in parallel with each c.s.d. 'open' solenoid, so that air is provided whenever the c.s.d. motor is called upon to operate.

It is essential for rapid and easy starting that unnecessary loads are removed from the engine. For this reason, the hydraulic pumps are disconnected by the operation of the hydraulic pump off-loading solenoids (h.p.o.l.). These solenoids are operated via contact *1A* when the ignition relay is operated and de-energized when the engine speed reaches 3 100 rev/min. An alternative feed is provided via contact *2A* of the c.s.d. relay so that the h.p.o.l. solenoids are operated during a motor over cycle.

External to the unit, the feed to the h.p.o.l. solenoids is taken via an oleo contact. This contact is closed only when the aircraft is on the ground and prevents loss of hydraulic power in flight. Another oleo contact operates an oleo relay to enable this facility to be duplicated in the unit.

The aircraft possesses the facility of mechanically disconnecting each gearbox from its engine. This enables various auxiliaries to be driven with power from the other engine in the event of engine failure. This disconnexion is achieved by the disconnect actuator *M*. The actuator is operated by the gearbox disconnect switch in the cockpit and the h.p. cock relay. This relay is energized when the engine throttle lever is operated and is used to open the high pressure fuel cock. Thus, it is impossible to disconnect a gearbox if the engine is under power. Under normal conditions with the gearbox connected a feed is provided via the h.p. cock relay, c.s.d. relay and oleo relay to the c.s.d. close solenoid. (The c.s.d. is operated by two solenoids and has a mechanical latch. Thus, the c.s.d. remains closed when the feed to the close coil is removed and only operates when a supply is connected to the open solenoid). When the h.p. cock is energized, this feed is removed and closure of the c.s.d. relay closes the c.s.d. open solenoid. Thus the c.s.d. valve opens, air enters the air motor and the engine turns. When the engine speed reaches 5 300 rev/min, a permanent feed is applied to the c.s.d. latching mechanism.

Should an engine fail in flight, the throttle lever is closed, and if the engine is below 5 300 rev/min the c.s.d. open solenoid is de-energized. When the g.b. disconnect switch is operated, the c.s.d. open and close solenoids are thus de-energized and the actuator operates. When it reaches the end of its travel, the limit switches change over and power is fed again to the c.s.d. open coil. This facility also enables the c.s.d. to be run on the ground. In reverse, the disconnect switch is opened and the c.s.d. close coil is energized. The air motor stops and the actuator connects the gearbox again. The c.s.d. open coil is then energized and the system reverts to normal.

The injector relay is used for internal aircraft control purposes and is energized when either h.p.o.l. circuit is energized.

Auxiliary Airborne Power Plant (A.A.P.P.) Start Circuits

The a.a.p.p. provides compressed air and some electrical power and is automatically lowered from the aircraft fuselage when required. To prevent accidents, the engine can only be operated when a manually operated run/stop switch which is situated near the engine is closed. The circuits of the start system are shown in Fig. 9. The system is designed so that if any faults occur such as overspeed or overtemperature, the engine is shut down and remains inoperative until the circuits are reset.

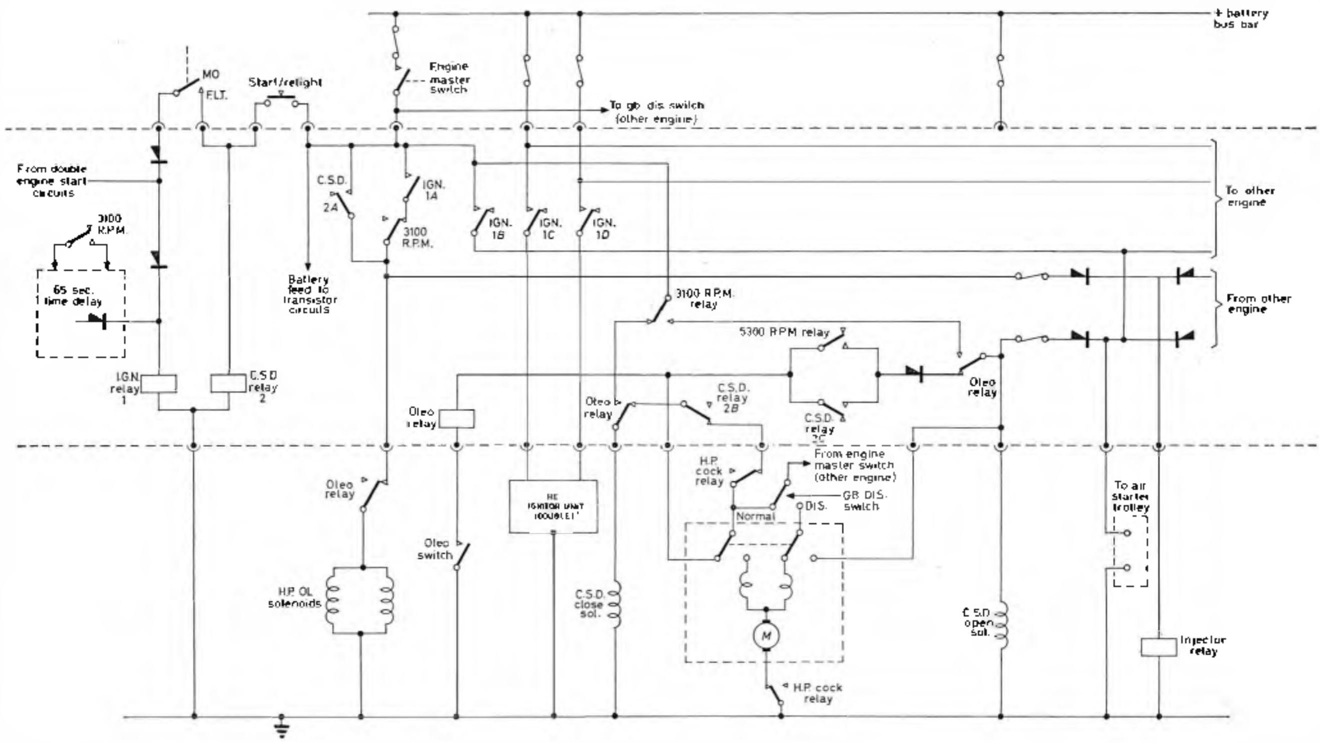


Fig. 8. Single main engine circuits

Selection of S_1 (a cockpit switch) to DOWN operates the down relay and low pressure fuel cock. The a.a.p.p. is lowered and the down limit switch closes. Moving S_1 to RUN energizes the ignition circuits via K_1, H_1 and F_2. When the engine reaches 8 000 rev/min, F_2 opens and the ignition circuits are de-energized. If the engine does not reach 8 000 rev/min in 6·5sec, relay L_E operates and latches via K_1 (J_1 is normally in the closed position and does not influence the circuit in this mode).

The high pressure fuel cock is energized with the ignition provided that no faults are present (i.e. K_2 and H_2 are as shown). A feed is also taken to an alternator bias circuit during the start sequence. This provides a small direct current to the alternator field to ensure an output from the alternator before its regulator is operative. This is necessary because the alternator is used to provide the speed sensing signals. When the engine is running down, an alternative feed is taken via F_3. Diode MR_{142} prevents a backfeed to the h.p. cock.

When the engine speed reaches 34 000 rev/min (the

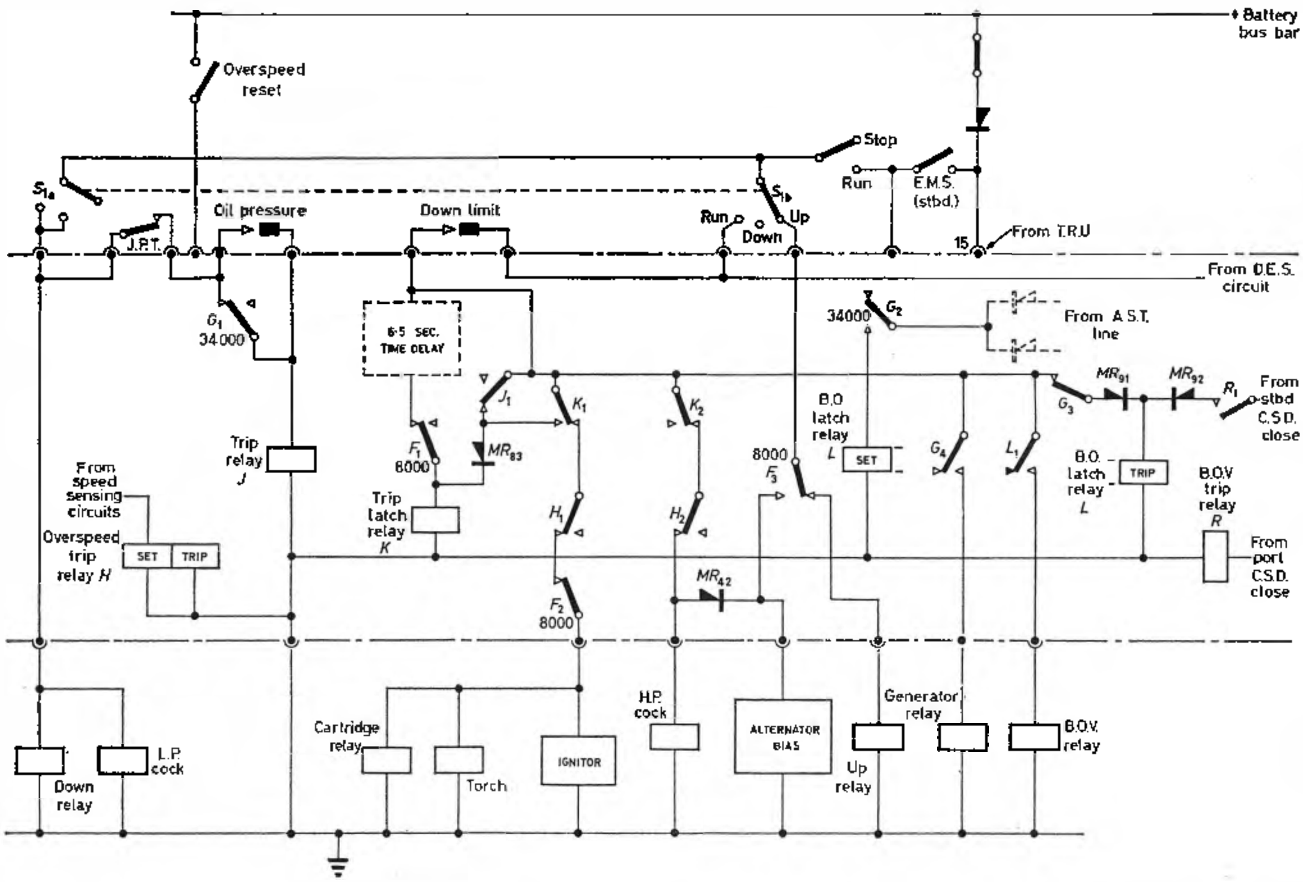
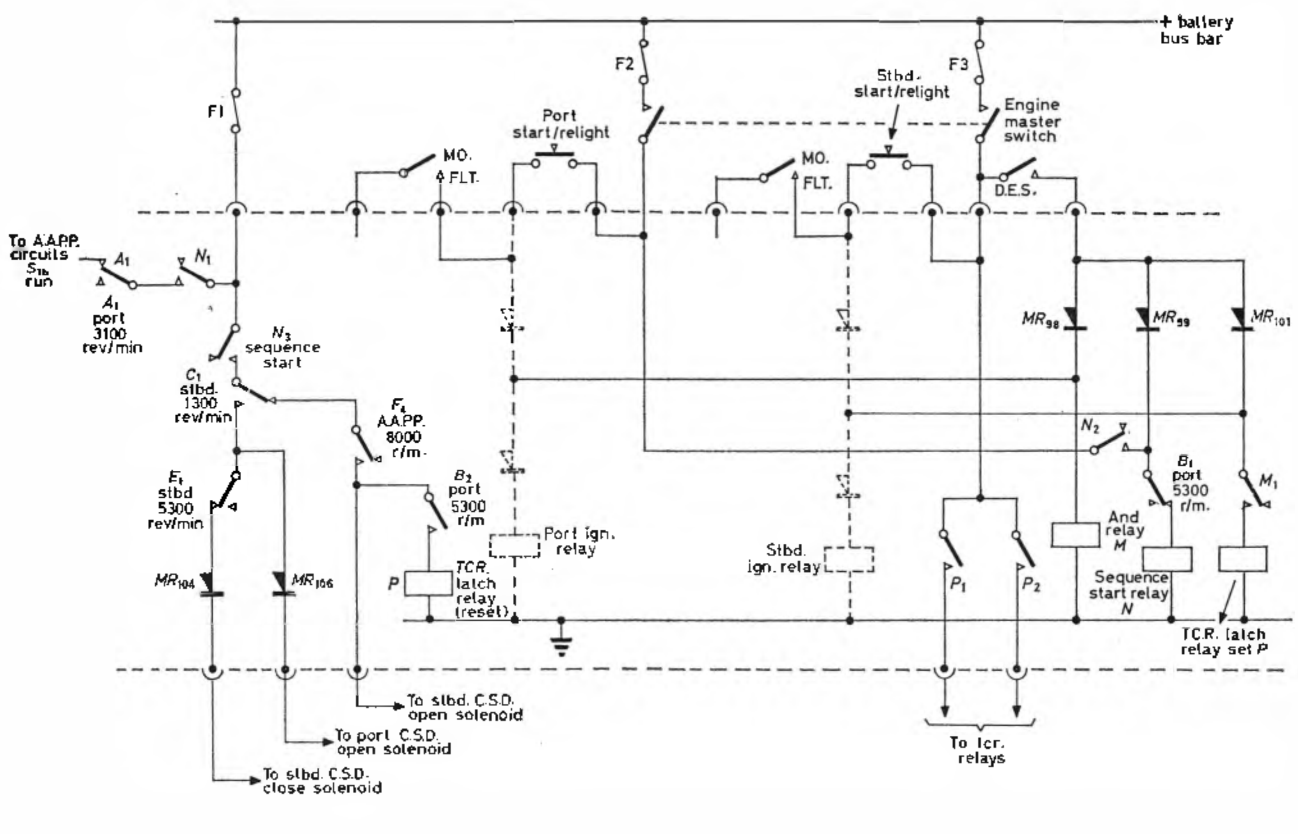


Fig. 9. A.A.P.P. circuits

Fig. 10. Double engine start circuits



normal running speed is 35 000 rev/min) relay LG closes and energizes the generator relay. This operates and connects the alternator to the electrical circuits in the aircraft which are to be operated from this source.

Relay G also operates the blow-off valve latch relay, L, provided that either main engine c.s.d. open coil is operated. This last condition is satisfied by feeds from the air starter trolley circuits. The blow-off valve is necessary to discharge the air from the a.a.p.p. when it is not required for engine starting. The b.o. latch relay is reset automatically by G_3 when the a.a.p.p. speed falls below 34 000 rev/min. Above 34 000 rev/min the relay is reset via relay R when both c.s.d.'s close coils are operated.

So that the a.a.p.p. can be operated continuously without draining the aircraft batteries, a transformer rectifier unit is included which operates from the alternator. A feed from this unit is fed via pin 15 to the a.a.p.p. control circuits.

To switch the engine off S_1 is moved to the UP position. Providing that the engine speed is less than 8 000 rev/min the UP relay is energized and the a.a.p.p. automatically retracts into the fuselage.

Protection is provided against high jet pipe temperature, low oil pressure and overspeed. The first two signals are provided by units outside the starting unit and appear simply as normally closed contacts. The trip relay, J, is normally energized and is de-energized if either the j.p.t. or oil pressure contacts open. To prevent operation during starts when the oil pressure is naturally low, the oil pressure switch is bypassed by G_1 below 34 000 rev/min. When relay J opens, the trip relay K is energized by J_1 and latches via K_1. This breaks the supplies to the ignition and fuel cock circuits and the engine runs down. The fault circuits can be reset by moving S_1 to the DOWN or UP positions. When the engine overspeeds, the overspeed trip relay H is operated and latched. This again switches off the ignition and fuel circuits. Reset is provided by a separate coil reset switch.

Double Engine or Sequence Start System

To enable the main engines to be started rapidly, provision is made for them to be started either simultaneously or sequentially. The former is used when an air starter trolley is available and is initiated by pressing both start/relight buttons simultaneously. The air supply from the a.a.p.p. is only sufficient to start one engine at a time and hence the sequential start is used when the a.a.p.p. is in operation. The circuit for these modes is shown in Fig. 10.

For rapid starting, the engine starting unit is linked to the main throttle control units by the throttle control latch relay (P). This provides a supply to the throttle control units which artificially simulates full throttle. The set coil of relay P is energized when either the double engine start (d.e.s.) switch is closed (via M_1) or both start/relight buttons are pressed simultaneously. The port button energizes relay M and the starboard button feeds M_1 to relay P. When both buttons are pressed simultaneously then both engines start using the sequence described for Fig. 9.

For a sequence start, however, assuming the a.a.p.p. is down, the d.e.s. switch initiates a sequence which starts the a.a.p.p., then the starboard and the port main engines and finally shuts down the a.a.p.p. This is designed to be as rapid as possible and leaves the pilot free to perform other cockpit checks. When the d.e.s. is closed, relay N operates and self-holds via N_2. Power is then applied via N_1 and A_1 to the a.a.p.p. start circuits (Fig. 8) and the a.a.p.p. starts as described earlier. When the a.a.p.p. feed reaches 8 000 rev/min sufficient air is available to start one engine and F_4 closes providing a feed to the starboard c.s.d. open solenoid. The starboard engine starts to turn and when it reaches 1 350 rev/min relay C is energized and the feed is removed from the c.s.d. open solenoid and taken to the close solenoid. The starboard engine is now self-sustaining and begins to gather speed. Meanwhile the port c.s.d. open solenoid is energized via C_1 and this engine starts to turn. When the starboard engine reaches 5 300 rev/min the feed to its c.s.d. close solenoid is removed and its open solenoid is energized in the normal way (Fig. 8).

When the port engine reaches 3 100 rev/min, A_1 operates and shuts down the a.a.p.p., since the port engine is now self-sustaining. When this engine reaches 5 300 rev/min, relay B operates and opens relay N via B_1. At the same time the t.c.r. latch relay is reset by the closure of B_2. At the end of this sequence, both main engines are running and the a.a.p.p. is losing speed. The pilot can then select a.a.p.p. UP and prepare for take-off. When the a.a.p.p. reaches 8 000 rev/min it will automatically retract and the aircraft can take-off. Thus, it is possible to imagine that with experience, the pilot could actually commence his take-off run as the a.a.p.p. runs down. For military applications this gives an extremely rapid and unhindered aircraft take-off sequence.

Construction

The unit is designed for mounting on standard aircraft equipment bay racking and consists of two compartments: the relay compartment and the electronic compartment. The relays are of the miniature sealed type and are soldered in. The electronic circuits are mounted on plug in printed wiring cards and are coated with a protective resin. The speed sensing circuits and time delay circuits were designed to allow for wide variations in setting because the exact times and speeds were not known at the time of the original design.

Conclusions

A brief description of the background and possibilities of automatic engine starting has been given and one particular example has been described in detail. This example represents a design which is roughly four years old and future designs may differ in many respects. For multi-engine aircraft the sequence may be complicated enough to warrant the use of microminiature techniques and consideration may be given once again to the use of semiconductor output circuits with large overload ratings. Such a system would be lighter, more reliable and smaller than a relay version and could even be cheaper.

Acknowledgments

Many people have been associated with this project and it is impossible to acknowledge them all. However, the following personnel and organizations must be recognized for the major part which they played.

Mr. J. P. Ratcliffe, his colleagues and staff of the British Aircraft Corporation both at Weybridge and Warton for the original system concept.

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Finally, acknowledgments are due to The M.E.L. Equipment Co. Ltd for permission to publish this article.

An Economical Ring Commutator for Oscilloscope Beam Switching

By A. F. Lewis*, B.Sc., Ph.D., M.B., B.Ch.

A simple four-way ring multivibrator is used to control eight push-pull transistor switches which provide simultaneous display of eight signals on a double beam oscilloscope.

(Voir page 552 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 559)

THERE is no greater help in the understanding of basic circuits or the mastery of complex circuits than the ability to visualize the component waveforms simultaneously. This prompted the design of an economical switching unit for undergraduate use to provide consecutive sampling of four input signals for display by a single beam oscilloscope. The addition of four more switches to this unit allows eight inputs to be displayed on a double beam oscilloscope.

Adequate switching is provided for signals from 100mV to 20V in amplitude, or if channel separation is required, to 4V in amplitude. For single shot viewing the frequency range is from 0 to 100c/s, for repetitive viewing the response is from 0 to 25kc/s (−3dB point).

Ring Commutator

A conventional four-way ring commutator consists of four binary pairs and an astable driver, together with a suitable commutating device. A considerably simpler circuit is shown in Fig. 1. This unit is a saturated ring monostable with diodes to isolate the collectors from the loading effect of the coupling capacitors. When switched on the circuit enters one of three possible states:

(a) All transistors saturated.

(b) Double pulse cycling with alternate transistors saturated.

(c) Single pulse cycling with all transistors except one saturated.

Condition (c) is initiated by simultaneous grounding of the bases of two adjacent transistors, the most reliable triggering points being found by selection.

In this state the collectors provide a sequential negative rectilinear pulse of 100μsec duration (Fig. 2(a)). Using medium speed switching transistors at a collector current of 10mA the rise time is 1μsec and the fall time 2μsec (Fig. 2(b)). This collector current gives the optimum switching time and ensures relatively light loading of the collectors by the beam switches.

When this commutator is used to drive a simple beam switch without means of beam suppression during switching, the amount of flare between channels depends upon the relation of switching time to sampling time. This figure is 3 per cent in this circuit and flare is at an acceptably low level.

The power requirement is 30mA at 18V centre-tapped which is conveniently provided by two 9V batteries.

No provision has been made to vary the sampling frequency in this unit although the coupling capacitors

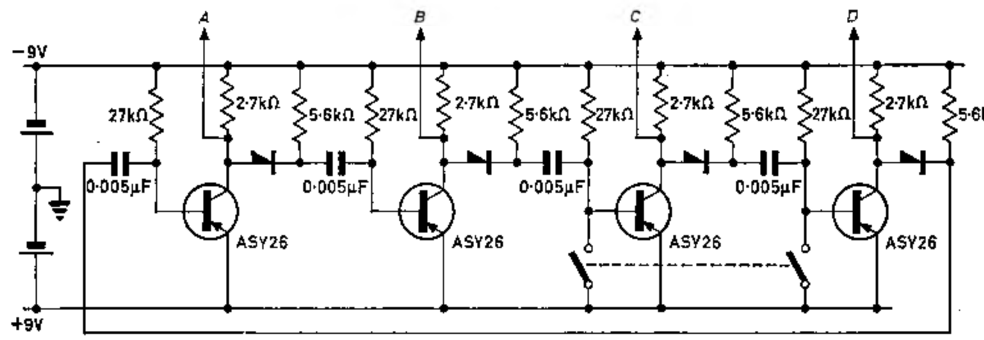

Fig. 1. The ring commutator

A pulse of 100μsec duration and 18V amplitude is available sequentially at the outputs A, B, C and D. Small signal germanium diodes are used to isolate the collector load resistors from the coupling capacitors

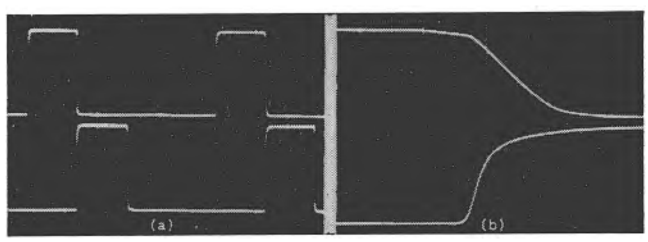

Fig. 2 (a). The output pulse from two consecutive collectors of the ring commutator

The duration is 100μsec and amplitude −18V

Fig. 2 (b). The leading (lower trace) and trailing (upper trace) edges of the ring commutator pulse

The time scale is 1μsec/cm grid marking

could be switched to deal with signals that prove to be harmonics of the switching frequency.

Beam Switch

The mechanical equivalent of the transistor beam switch is shown in Fig. 3. The switches S_1 are normally open and present infinite impedance to the signal source. The switches S_2 are normally closed and hold the junction point X at ground potential. During sampling the switches for that channel (channel A in Fig. 3) are reversed. The point X_a now assumes signal potential and the signal source is loaded by a resistance equal to $4R/3$, i.e. R_a in series with the parallel combination of R_b, R_c and R_d. This figure neglects the small shunting effect of the oscilloscope input impedance.

Because of the potential divider action of the resistances

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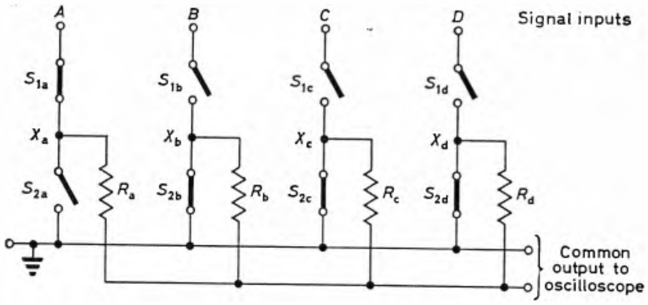


Fig. 3. The mechanical equivalent of the beam switching units
The resistors R_a, R_b, R_c and R_d are of equal value

R_a, R_b, R_c and R_d the oscilloscope sees the signal attenuated to 1/4 across a resistance of $R/3$, i.e. the parallel combination of R_b, R_c and R_d.

The transistor beam switch in Fig. 4 uses a complementary pair to replace switches S_1 and S_2. In the non-sampling state the +9V base input from the ring commutator holds the pnp transistor VT_1 (replacing S_1) in the cut-off state. The same input to the base of the npn transistor VT_2 (replacing S_2) holds it in the saturated state at a base current of 90μA. Because these transistors have at all times a finite resistance they act as a potentiometer across which the signal is applied. Thus in the non-sampling state the point X is very nearly at ground potential and the signal source is loaded by the resistance of the cut-off transistor VT_1. This value exceeds 1MΩ at 20°C. The ratio of the resistance of these two transistors in the non-sampling state determines the crosstalk of the switch. This value is less than 1 in 10^5.

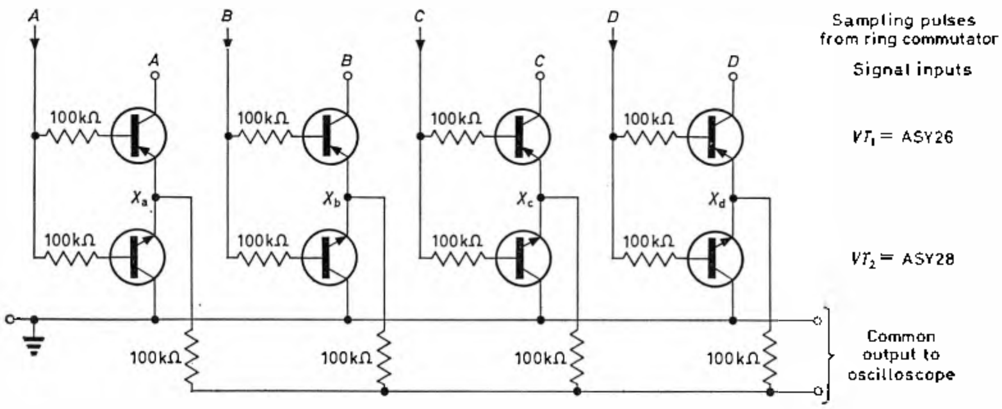


Fig. 4. The beam switching units using complementary transistors

During sampling the −9V pulse from the ring commutator reverses the state of the transistors VT_1 and VT_2 and the potential at X approaches signal potential. Again the ratio of the resistance of the two transistors determines the precise potential at X and signal attenuation at this point is less than 1 in 10^5. The input resistance is now $4R/3$ (equal to 133kΩ) shunted by the 100kΩ biasing resistor to the base of VT_1 in series with the collector load of the ring commutator. This gives a theoretical value of 58kΩ for input resistance during sampling which agrees closely with the experimental value.

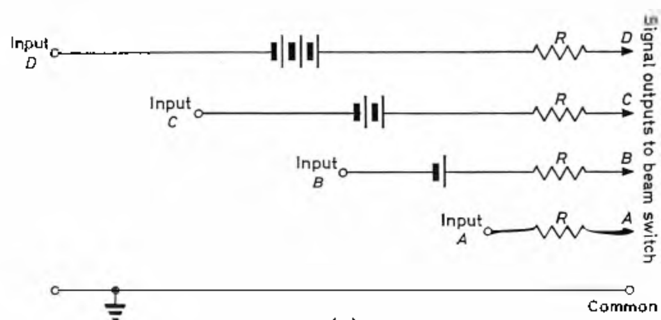
The output resistance is still $R/3$ which is 33kΩ and overall signal attenuation is to 25 per cent. This is very satisfactory for feeding into an oscilloscope but the input impedance of 60kΩ is too low for many applications unless a pre-amplifier is used. As further signal attenuation is easily compensated in the oscilloscope amplifier a series resistor may be used to achieve a suitable input resistance. A resistor of 90kΩ will raise the input resistance to 150kΩ and give overall signal attenuation to 10 per cent.

Channel Separation

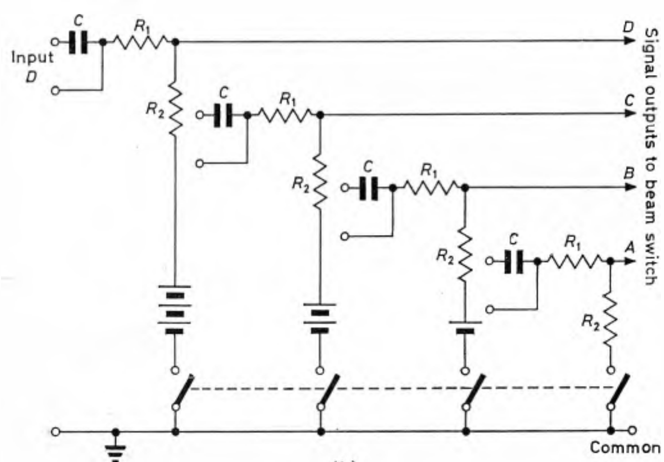
Shift controls for channel separation must be current operated and such a biasing network from a common power source produces unequal attenuation of the signals. As current requirements are very low, floating batteries in series or parallel with the signal inputs offer a simple and expedient solution.

Because of the leakage current in the transistor switch the junction point X (Fig. 4) assumes a potential of about −4V at zero input. The negative limit for the potential at X is −9V as a greater potential would eliminate the reverse bias on the base of VT_2 during sampling. Therefore the maximum negative signal that can be accepted is −5V before the operation of the switch is impaired. One channel is used without further shift voltage being applied and the other three channels have a positive shift voltage applied. The upper limit of positive applied voltage is the breakdown characteristic of the switching transistors. In this circuit the critical value is the reverse base-emitter

Fig. 5 (a). Simple d.c. channel separation circuit
Fig. 5 (b). A more versatile channel separation circuit for a.c. or d.c. signal input



(a)



(b)

voltage of VT_2, which limits the applied positive signal to $+15$V. This creates a potential at X of $+11$V and since the base bias is -9V the reverse potential across the base-emitter junction is 20V. These devices are undoubtedly conservatively rated and to date no device has failed despite 100 per cent overload in this respect.

The simplest channel separation technique is shown in Fig. 5(a). A battery of greater e.m.f. than the peak signal is connected in series with the input via a suitable resistor to attenuate the voltage across the input to within 4V, the channel width. Thus for the display of 9V signals the separation e.m.f. is obtained from batteries that are multiples of 15V and a series resistor of 180kΩ (equal to three times the input resistance) attenuates the 15V increment to just under the channel width of 4V.

In this example the largest battery of 45V would pass a maximum current of 200μA during sampling if the source resistance is zero. Note that this current passes through the source and a.c. coupling is not possible by this technique. However this method is simple and the average current drain from the three batteries of 45V, 30V and 15V is only 50μA, 25μA or 12μA respectively.

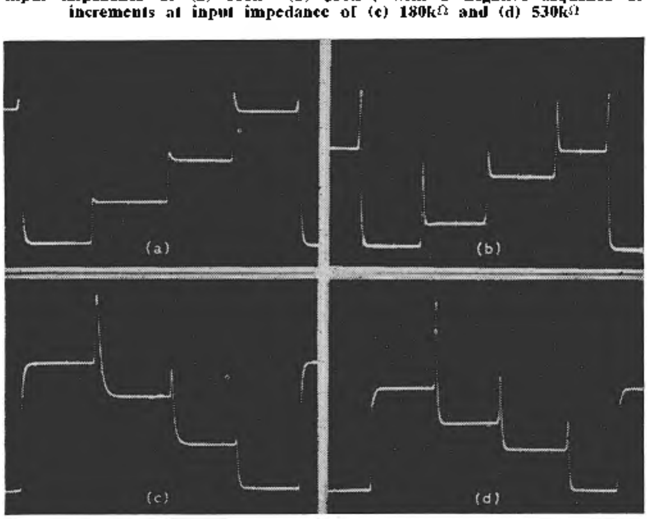
In Fig. 5(b) is shown a more versatile system which permits a.c. or d.c. coupling of the signals. The separation voltage and the signal are in parallel across the input. The signal is attenuated to a value within the channel width by the resistor R_1 and the battery e.m.f. is reduced to a multiple of 4V by the resistor R_2. As R_2 affords an additional attenuating path for the signal input the lowest practical value is 120kΩ and this means that the battery e.m.f. must be three times the channel separation voltage required, i.e. 12V to produce 4V separation. The average current drain from a 45V battery in this circuit is not more than 100μA.

Flare Between Channels

The effect of the ratio of switching to sampling time in the ring commutator has already been described. In addition flare is caused by a switching transient within the switch itself. As this transient is positive going it is possible to minimize its effect by careful choice of the timing sequence. When the sampling sequence is composed of positive separation voltage increments (Fig. 6(a)) the visible transient is much less than in the reverse order (Fig. 6(c)).

Raising the input resistance increases the amplitude of this transient (Fig. 6(b) and (d)). However it is always possible to produce a well contrasted trace for viewing or photography and variations in the transient are of little consequence.

Fig. 6. The sequential sampling of four separated channels at zero signal input, with a positive sequence of separation voltage increments at an input impedance of (a) 180kΩ (b) 530kΩ, with a negative sequence of increments at input impedance of (c) 180kΩ and (d) 530kΩ



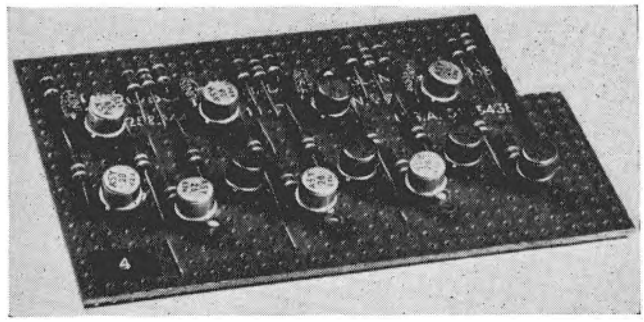


Fig. 7. The assembled ring commutator and switches for four channels
The Veroboard dimensions are 13 × 8cm

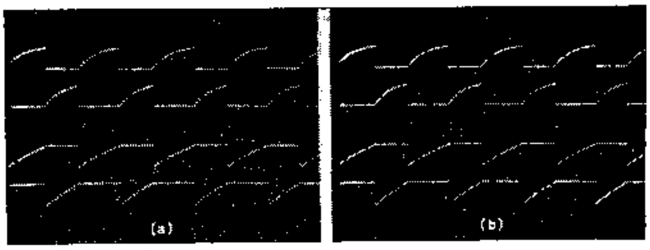


Fig. 8. The collector and base waveforms of an astable multivibrator operating at 100c/s displayed on a single beam oscilloscope
(a) With single shot viewing and (b) with repetitive viewing

Construction And Use

The whole unit is compactly and easily constructed upon Veroboard. Fig. 7 shows a four-channel unit constructed upon a piece of 0·15in pitch material measuring 13 × 8cm. No selection of components is necessary and 10 per cent tolerance resistors are adequate. The unit plugs into the appropriate power supply and channel separation network and is easily replaced should a fault arise. No overload protection is really necessary in handling signals from circuits operating at less than 25V. If necessary the input voltages can be limited to the range -5V to $+15$V by Zener diodes across the switch inputs.

It is necessary to synchronize the oscilloscope externally direct from one of the signal inputs otherwise triggering occurs on the switching pulse.

Both the switching transient and the channel separation voltage may have an adverse effect upon the functioning of the signal source. Discretion should be exercised in selecting the circuit points that will receive the highest channel separation currents and it is not advisable to attempt to record small base-emitter potentials without pre-amplification.

Biological signals with a low signal/interference ratio should be passed through a differential pre-amplifier before switching.

It is worth emphasizing that these results are obtained with low price medium speed germanium switching transistors so that the complete unit is produced for less than £5.

An additional use for the unit is as a chopper to allow d.c. measurements to be made on an a.c. oscilloscope.

Typical results of single shot and repetitive viewing are shown in Fig. 8 where the limits of definition are visible.

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Bistable Elements for Sequential Circuits

By J. J. Sparkes*

The problem of how to avoid the ambiguous drive conditions of the R-S flip-flop is discussed and all the possible and reasonable solutions are briefly compared as to their simplicity of construction and as to their versatility in sequential circuits. It is shown that even for quite simple sequential circuits the exclusive OR *flip-flop (or the E flip-flop) may be an improvement on the J-K flip-flop in that it not only leads to a comparable degree of simplification of the combinational circuits but also is simpler to construct, can be realized in both triggered and d.c. form and is identical for both positive and negative logic.*

(Voir page 553 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 560)

NOWADAYS most sequential switching circuits are centred round transistor switches or transistorized bistable elements. The simplest of the memory elements is the Eccles-Jordan bistable circuit, with or without pulse steering, usually called the *R-S* flip-flop. In this article bistable circuits in which logic inputs are used to steer trigger pulses will be called triggered flip-flops. An example of one is shown in Fig. 1(b). The circuits which are driven directly by their logic inputs, as in the example of Fig. 1(a), will be called d.c. flip-flops. In either case the logic inputs will be labelled *S* and *R*.

The operation of these circuits is briefly as follows. Each circuit has two levels of output, 6V and zero in this case, and all inputs are driven by circuits which switch between these two levels. Using positive logic, so that 6V is the significant voltage level, it is evident that when the set input *S* of the *R-S* d.c. flip-flop is excited (i.e. driven positive) output *A* goes positive and the flip-flop is in the '1' state. Similarly when the reset input *R* goes positive, the flip-flop is put in the '0' state and $\bar{A}$ goes positive. Evidently when *S* and *R* are both kept at zero volts, the flip-flop stays in its last state, but if both bias inputs *S* and *R* are driven positive together, the state of the flip-flop when *S*, *R* return again to zero will not in general be predictable and the memory function of the circuit will be lost. Thus if an *R-S* d.c. flip-flop is to be used as a storage or memory element both bias inputs should not be excited together.

The *R-S* triggered flip-flop differs from the d.c. flip-flop in that the state of the circuit does not respond immediately to the *S* and *R* inputs. They determine the next state of the flip-flop. The state of the circuit can only change when a negative-going pulse or voltage step is applied to the pulse input. With *S* excited (i.e. positive) the next state is '1'. With *R* excited the next state is '0' ($\bar{A} = 6$V), but again when both bias inputs are excited together the outcome is indeterminate and in sequential circuits this condition must be avoided.

The above limitation on the range of acceptable inputs can be removed by some of the alternative forms of flip-flop proposed by Phister[1] in 1957. Since then others[2,3] have proposed various other flip-flops in some of which the ambiguous state does not occur. In particular Wrzesinski[4] has shown that there are 108 basically different arrangements of two-input flip-flops, and has made some progress towards choosing the most useful one. The purpose of the present article is to consider a restricted number of configurations in which only the ambiguous state of the *R-S* elements is modified, and to compare them approximately both as to their complexity of construction and their versatility in sequential circuits.

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Selected Logical Functions of Bistable Elements

It is sensible when considering the possible next states arising from the four combinations of two binary inputs to arrange that the input $S\bar{R}$ always leads to a '1' state and that the input $R\bar{S}$ always leads to an '0' state. It is

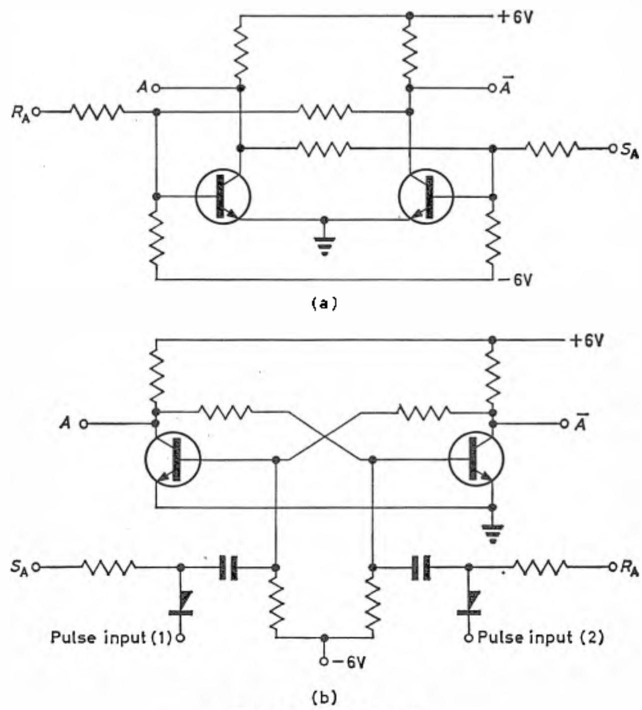


Fig. 1. Bistable Circuits
(a) An R-S toggle (or d.c. flip-flop)
(b) An R-S triggered flip-flop

also usually desirable to have one input combination which leaves the bistable element in its previous state. These arbitrary but reasonable restrictions lead to the following configurations.

TRIGGERED FLIP-FLOPS

Using Phister's notation, that Q^n is the state of the flip-flop after the n^{th} clock pulse and Q^{n+1} its state after the $(n+1)^{th}$ pulse, a table can be compiled which expresses all the possible logical functions a two input flip-flop might perform, as shown in Table 1.

The table also shows suggested code names for each type of flip-flop.

The *R-S* flip-flop and its inverse, the *R-S-I* flip-flop both have indeterminate states indicated by the question mark in the rows corresponding to the *SR* and $\bar{S}\bar{R}$ input condition respectively.

The *J-K* flip-flop was proposed by Phister and corresponds to Marcus's *S-R-SR* flip-flop. In this case the flip-flop changes when both inputs are excited together.

The Set Biased and Reset Biased flip-flops (S, R) were discussed by Maley and Earle[3]. Their function is self evident.

The exclusive flip-flop (E) appears not to have been discussed before. It remains unchanged for both SR and $\bar{S}\bar{R}$ input conditions.

All the Inverse flip-flops are logical inversions of the ones referred to above, (except the E flip-flop which is its own inverse).

The logical statements given in Table 1 assume one polarity of logic, either positive or negative. In Fig. 1 positive logic is being used, but the same circuits with pnp transistors and inverted diodes and supply voltages would give equivalent negative logic circuits. However, negative logic circuits used in a positive logic system exhibit the same logical functions as the Inverse circuits shown in the table, so that this table includes the function of both negative and positive logic circuits. In Table 1 negative logic circuits are analysed as if they were inverse positive logic ones since if circuits for all types of flip-flop are devised, using only positive logic, then another set also exists using negative logic in which all supply polarities are reversed, all npn and pnp transistors are replaced by their complements and all diodes are turned round.

Both Marcus and Phister also envisaged two flip-flops with a single bias input X (called T and D flip-flops). They also discussed *J-K* or *R-S* flip-flops with the X input added. These are called *R-S-T* and *J-K-T* flip-flops.

TABLE 1

INPUTS		Q^{n+1}								
R	S	*R-S*	*J-K*	S	R	E	*R-S-I*	*J-K-I*	*S-I*	*R-I*
0	0	Q^n	Q^n	Q^n	Q^n	Q^n	?	$\overline{Q^n}$	1	0
0	1	1	1	1	1	1	1	1	1	1
1	0	0	0	0	0	0	0	0	0	0
1	1	?	$\overline{Q^n}$	1	0	Q^n	Q^n	Q^n	Q^n	Q^n

TABLE 2

R	S	X	Q^{n+1}			
			T	D	*R-S-T*	*J-K-T*
–	–	0	Q^n	0		
–	–	1	$\overline{Q^n}$	1		
0	0	0			Q^n	Q^n
0	1	0			1	1
1	0	0			0	0
1	1	0			?	$\overline{Q^n}$
0	0	1			$\overline{Q^n}$	$\overline{Q^n}$
0	1	1			?	?
1	0	1			?	?
1	1	1			?	Q^n

The properties of these four flip-flops are indicated in Table 2.

The Trigger flip-flop, T, changes state when the input X is excited, while the Delay flip-flop D simply passes the input state into the flip-flop, and 'remembers' it.

The *R-S-T* and *J-K-T* flip-flops have extra indeterminate states because in some circumstances the inputs are in conflict. For example if the circuit is in state '1' an S input confirms it, but an X input contradicts it. There are, of course, many other arrangements of multiple input flip-flops but further consideration of them is beyond the scope of this article.

D.C. FLIP-FLOPS

Bistable elements with d.c. inputs cannot use their own outputs to control their inputs, since the 'next state' does not wait for a clock pulse; the state of the flip-flop changes immediately the appropriate input voltage level is applied so that use of a flip-flop's own output as an input is likely to set up an oscillatory condition.

This means that no logical rule of operation which specifies a change of state in *either* direction can apply to a single flip-flop since this involves feedback from output to input. In practice to achieve this kind of feedback with flip-flops, two flip-flops in cascade are used. In addition the d.c. D flip-flop only delays the input by the propagation delay of the circuit.

Thus the logical functions specified for *J-K* or T flip-flops must employ master-slave techniques when only d.c. flip-flops are used.

The Construction of Bistable Elements

Today there are many different ways being used to construct bistable elements using transistors and it is, of course, not possible to consider and compare them all. Indeed since this article is concerned only with circuit complexity resulting from modifying the logical function of the circuit it would be inappropriate to consider all the factors which lead to the modern high speed, noise resistant designs such as those involving high level transistor transistor logic. Consequently bistable circuits will be discussed in their simplest form and briefly investigated as to how much extra circuit complexity is needed to construct each modified flip-flop, as compared with an *R-S* flip-flop, using the same method of construction for each modification.

The method of construction used for this comparison is indicated in Fig. 2. Starting with a central *R-S* or *R-S-I* bistable element, whose inputs will now be referred to as S', R', combinational circuits connect S', R' to the two new circuit inputs S, R, so that the overall response of the network to the S, R inputs is of the required modified form (*J-K*, E, S, T etc.). These combinational circuits can be constructed in many ways but here NOR logic will mainly be considered.

The simple *R-S* flip-flops were illustrated in Fig. 1. The corresponding *R-S-I* flip-flop, using positive logic and npn transistors, is shown in Fig. 3. Here a negative going pulse or voltage step applied to the pulse input only reaches the base of a transistor if the corresponding bias input (R', or S') is at about zero volts (i.e. the more negative of the two available levels). This is the opposite of the *R-S* flip-flop and leads to indeterminacy if both R' and S' are at zero together. Thus the R-S-I function as indicated in Table 1 is obtained.

It can easily be shown that, for the E flip-flop using an R-S central flip-flop, the Boolean expression relating inputs S, R to inputs $S' R'$ are:

$$\left.\begin{aligned} S' &= S\overline{R} \\ R' &= R\overline{S} \end{aligned}\right\} \quad \dots\dots\dots\dots \quad (1)$$

Each input S', R' requires half an exclusive OR gate to drive it.

If NOR gates are used in the combinational circuits, their inputs are given by the Boolean expressions for $\overline{S'}, \overline{R'}$ namely:

$$\left.\begin{aligned} \overline{S'} &= \overline{S} + R \\ \overline{R'} &= S + \overline{R} \end{aligned}\right\} \quad \dots\dots\dots\dots \quad (2)$$

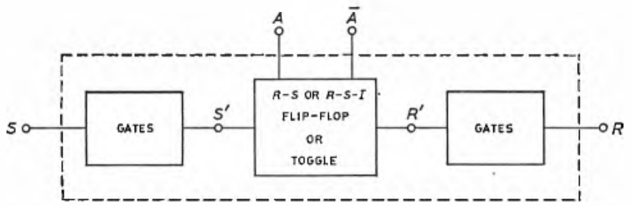


Fig. 2. A method for modifying the logical behaviour of bistable elements

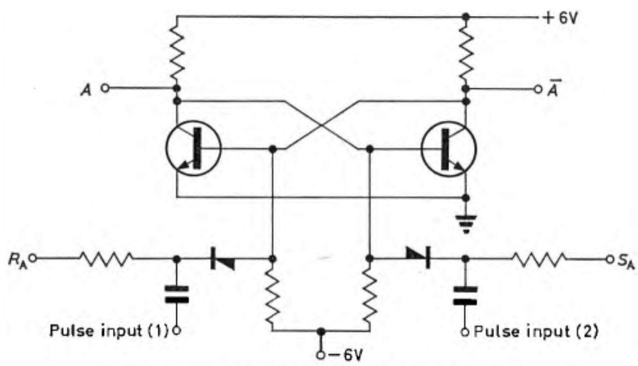


Fig. 3. An R-S-I triggered flip-flop

Using NOR gates to construct a J-K flip-flop the corresponding equations, using an R-S central flip-flop are:

$$\left.\begin{aligned} \overline{S'} &= \overline{S} + R\,A \\ \overline{R'} &= \overline{R} + S\,\overline{A} \end{aligned}\right\} \quad \dots\dots\dots\dots \quad (3)$$

While using an R-S-I central flip-flop they are:

$$\left.\begin{aligned} \overline{S'} &= R\,\overline{(\overline{S} + A)} \\ \overline{R'} &= S\,\overline{(\overline{R} + \overline{A})} \end{aligned}\right\} \quad \dots\dots\dots\dots \quad (4)$$

Similar expressions can be obtained for all the types of bistable elements listed in Tables 1 and 2.

Equations (3) are a little easier to realize in NOR logic than are equations (4). Thus a J-K flip-flop built round an R-S flip-flop requires fewer NOR gates than one using an R-S-I. If, similarly, the simplest arrangement for each type of flip-flop is taken and the NOR gates involved are counted the results of Table 3 are obtained.

The use of NOR gates is often expensive in terms of the number of components, so one would expect simplification to be possible, as illustrated in Fig. 4 for an E flip-flop. (In order to achieve usable fan-in, the gating transistors in Fig. 4(b) and 4(c), which are forms of inhibit gates, must be operated at a current level of about an order of magnitude less than that used in the R-S or R-S-I flip-flop.) In general, however, it may be assumed that the NOR gate component count gives some indication of the complexity involved in obtaining modified bistable elements.

TABLE 3

Type of Bistable ..	*JK*	*E*	*S*	*T*	*D*	*JKI*	*SI*
Number of *NOR* gates	6	4	2	3	1	6	4

Economy of Combinational Circuits

The differences between the various types of flip-flop are reflected in the complexities of the combinational circuits needed to drive them. In general the more complicated the flip-flop construction the simpler the input combinational circuits. But this greatly oversimplifies the effect upon circuit design of the use of different types of flip-flop.

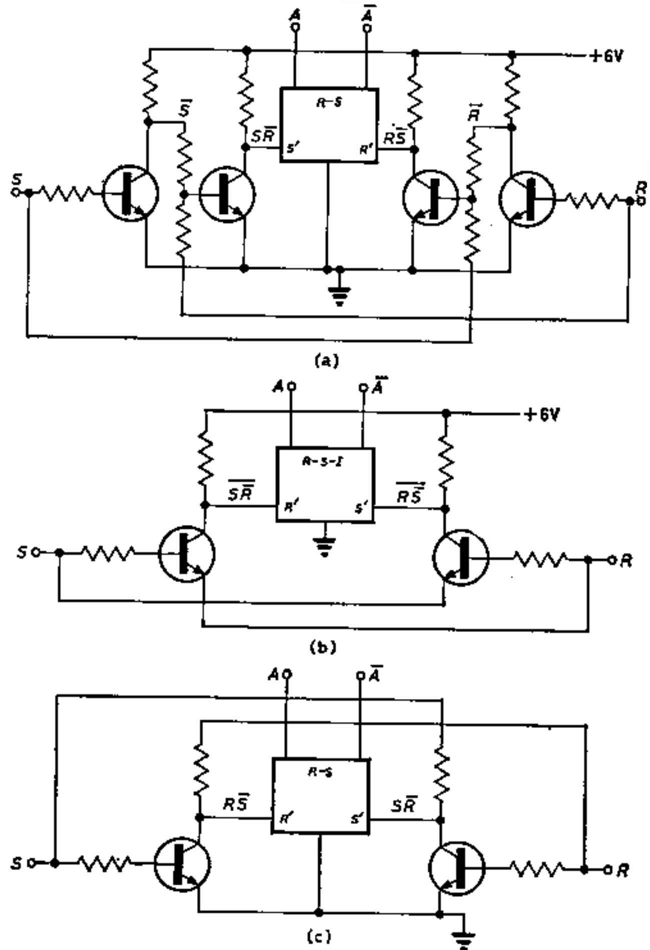


Fig. 4. Three realizations of the E flip-flop
(a) Using NOR gates
(b) and (c) Using two types of inhibit gates

The systematic design of sequential circuits usually culminates in a Karnaugh map[5], or its equivalent, two simple examples of which are shown in Fig. 5. (For completeness the derivation of these particular maps is outlined in the appendix.) The significance of different types of flip-flops can be illustrated with reference to these maps.

In Fig. 5, the rows and columns labelled A, B, P, Q are the outputs of four flip-flops, which can either be part of the sequential circuit, or can be flip-flops in quite

AB \ PQ	00	01	11	10
00	r	r	r	r
01	r	S	r	r
11	R	R	R	R
10	–	–	–	–

(a)

AB \ PQ	01	01	11	10
00	S	r	r	S
01	s	s	R	s
11	s	R	R	s
10	–	–	–	–

(b)

Fig. 5. The Karnaugh maps for the two flip-flops comprising a simple sequential circuit

(The design is outlined in the Appendix)

(a) flip flop A *(b) flip-flop B*

separate circuits. In this case, A and B form part of the sequential circuit, while P and Q provide inputs to it.

These particular maps are the input excitation maps of flip-flops A and B in the sequential circuit and each entry in the maps indicates what voltage (positive or zero), using positive logic, should be applied to the Set and Reset inputs of an R-S flip-flop, for all 16 combinations of the states of A, B, P, Q. (The notation used follows closely that of Zissos[6] which is somewhat simpler than that of Marcus and others.)

The entries have the following significance:

(1) A capital S means that the Set input of the flip-flop *must* be excited, because the state of the flip-flop has to be changed from '0' to '1' when the clock pulse arrives.

(2) A capital R means that the Reset input *must* be excited, so that the state of the flip-flop changes from '1' to '0'.

(3) A lower case s means that the flip-flop is already in state 1 and has to remain there. Thus the Set input *may* be excited because it will cause no change. Alternatively it can be left alone since this way, too, no change will occur.

(4) A lower case r similarly means that the Reset input *may* be excited.

(5) Blank entries appear where some combinations cannot occur. In Fig. 5 the state $A\bar{B}$ never occurs.

For example, when the circuit is in the state $A = B =$ '0' an input of $P =$ '1' and $Q =$ '0' must cause B to change to the '1' state but must leave A in the reset or '0' state.

The maps of Fig. 5 apply to an R-S flip-flop because there is only one entry in each element of the map. That is only one input is excited at any one time, and the indeterminate state of an R-S flip-flop is avoided. Any other two-input flip-flop can be driven in the same way, but, in addition, for these other flip-flops the possibility exists of economizing on combinational circuits by allowing double capital letter entries (i.e. both inputs excited) in the appropriate places. Fig. 6 shows the map of Fig. 5(a) redrawn for each type of flip-flop, including the single input ones, together with the Boolean equations describing the input combinational circuits.

The J-K flip-flop.

A change of state can be achieved by exciting both inputs. Thus any capital letter can be replaced by a double capital letter if this simplifies the combinational circuits. For this example, as shown in Fig. 6(b), the four-input AND gate for the Set input is reduced to a three-input one.

The E flip-flop

No change of state occurs if both inputs are excited. Thus any lower case entry can be replaced by a double capital letter entry. This offers a range of alternative combinational circuit arrangements as illustrated in Figs. 6(c), (d), (e). These different arrangements show a progression from a four-input AND gate for the Set input together with a single input for the Reset, to the other extreme of a single Set input and a four-input OR gate for Reset, as shown in Fig. 6(e). In this way it is possible to adjust the fan-in of each combinational circuit towards a uniform minimal value. (When fan-in and speed have to be traded this can be important.)

In addition to allowing this interchange of AND complexity for OR simplicity, the E flip-flop also makes overall economies in some combinational circuits possible. The kind of map for which the E flip-flop is of most value, and for which the J-K flip-flop is no help, is shown in Fig. 7. The E equations shown are obtained by changing all the r's to RS's.

It is, of course, always possible, using an E flip-flop, to arrange for one of the input circuit equations to contain only one letter on the right-hand side, (as shown in the E flip-flop equation for S_A in Fig. 7) by changing either all the r's to RS's or all s's to RS's. Unfortunately

Fig. 6. The Karnaugh maps of flip-flop A revised for use with the various types of modified flip-flops

(a) 'R-S' $(S=\bar{A}B\bar{P}Q;\ R=A)$ *(b)* 'J-K' $(S=B\bar{P}Q;\ R=A)$
(c) 'E' $(S=\bar{A}\bar{P}Q;\ R=A+\bar{B})$ *(d)* 'E' $(S=\bar{A}\bar{P};\ R=A+\bar{B}+Q)$
(e) 'E' $(S=\bar{A};\ R=A+\bar{B}+P+\bar{Q})$ **(f)** 'J-K-I' $(S=\bar{A};\ R=\bar{B}+P+\bar{Q})$
(g) 'T' $(X=A+B\bar{P}Q)$ **(h)** 'D' $(X=\bar{A}B\bar{P}Q)$

(a)

	$\bar{P}\bar{Q}$	$\bar{P}Q$	PQ	$P\bar{Q}$
$\bar{A}\bar{B}$	r	r	r	r
$\bar{A}B$	r	S	r	r
AB	R	R	R	R
$A\bar{B}$				

(b)

r	r	r	r
r	S	r	r
R	SR	R	R

(c)

r	RS	r	r
r	S	r	r
R	R	R	R

(d)

RS	RS	r	r
RS	S	r	r
R	R	R	R

(e)

RS	RS	RS	RS
RS	S	RS	RS
R	R	R	R

(f)

RS	RS	RS	RS
RS	S	RS	RS
R	~~R~~	R	R

(g)

	X		
X	X	X	X

(h)

	X		

	$\bar{P}\bar{Q}$	$\bar{P}Q$	PQ	$P\bar{Q}$
$\bar{A}\bar{B}$	S	r	S	r
$\bar{A}B$	r	S	r	S
AB	R	s	R	s
$A\bar{B}$	s	R	s	R

Fig. 7. The type of map for which the E flip-flop offers the greatest economy in combinational circuits

$$R-S\begin{cases} S=\bar{B}\bar{P}\bar{Q}+\bar{B}PQ+B\bar{P}Q+BP\bar{Q} \\ R=B\bar{P}\bar{Q}+BPQ+\bar{B}\bar{P}Q+\bar{B}P\bar{Q} \end{cases}$$

$$E\begin{cases} S=\bar{A} \\ R=B\bar{P}Q+BPQ+B\bar{P}Q+\bar{B}P\bar{Q} \end{cases}$$

the other equation is usually made much more complicated as a result.

The S flip-flop

Any S entry can be changed to an RS entry since if both inputs are excited together the flip-flop goes to the '1' state. This offers no advantage in the example under consideration.

The R flip-flop

Any R entry can be changed to RS. Fig. 6(b) for the J-K flip-flop is also the modified map for the R flip-flop. In general, of course, the J-K flip-flop provides the economies of both the S flip-flop and the R flip-flop.

Inverse flip-flops

An inverse flip-flop is only prevented from changing state if both inputs are excited together, or if the next clock pulse drives it into the state it already holds. Thus r's must now be interpreted as R's or changed to RS's,, and s's must either be interpreted as S's or changed to RS's, whichever is the more convenient. It is advisable not to rewrite the r's as R's or s's as S's since these rewritten capitals which involve no change of state can be confused with the capitals already in the map which prescribe a change of state. With the R-I and J-K-I flip-flops the possibility exists of exciting neither input in order to achieve a change of state.

For an S-I flip-flop any S entry can be crossed out. For an R-I flip-flop any R entry can be crossed out and for a J-K-I flip-flop any single capital letter entry (R or S, but not the 'no change' RS entries) can be crossed out. This is illustrated in Fig. 6(f) for a J-K-I flip-flop with one R entry erased.

(Crossed out entries must not be confused with blank entries on the map. A blank entry can be included in any excitation grouping. A crossed out entry must be included in none.)

All these inverse procedures are, of course, an alternative way of expressing a change from positive to negative logic (or vice versa).

The T flip-flop

The single bias input, X, of the T flip-flop must be excited only when a change of state is required. Thus the map and equations of Fig. 6(g) apply.

The D flip-flop

This flip-flop must always be driven into its next state, there is no excitation which means 'no change' whatever the previous state. Thus any set entry, whether lower case or capital, must involve excitation of the single bias input, X. Fig. 6(h) shows the particularly simple solution in this case, although a four-input AND gate is needed.

It is similarly possible to read the Karnaugh maps in order to write down the Boolean expression for the multiple input flip-flops[2], but as yet there is no clear cut way of optimizing the procedure.

Similar maps can be compiled to express the logical operations to be performed by d.c. flip-flops. The procedure for reading the maps is the same, but since there are fewer possible types of d.c. flip-flops there are not so many variations.

Conclusion

It is clear that there are a variety of modified forms of flip-flop which are capable of simplifying the combinational parts of sequential circuits, but that each type of flip-flop is suited to one class of circuit rather than to all classes. It is not possible as yet to single out one type of flip-flop as being the most versatile, although it is evident that only the E flip-flop (i.e. the Exclusive OR flip-flop) possesses a range of certain desirable properties in addition to the fact that it can offer economies in the number of components involved in the combinational circuits. These properties are:

(1) It makes possible the overall minimization of the fan-in of combinational circuits. That is, a large fan-in circuit can be reduced at the cost of increasing the inputs of a low fan-in circuit.

(2) It is a circuit which is identical in both negative logic and positive logic.

(3) It is relatively simple to construct—much simpler, for example, than the J-K flip-flop.

(4) It can be constructed both in triggered and d.c. flip-flop form (e.g. for either synchronous or asynchronous operation).

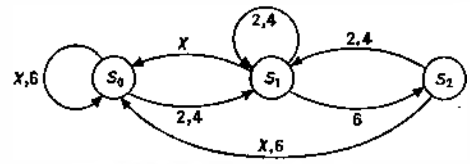
Fig. 8. The design of the sequential circuit

(a) The input code
(b) The sequence diagram
(c) The state excitation map and state assignment table
(d) The input excitation maps for flip-flop A and flip-flop B

(a)

P \ Q	0	1
0	2	6
1	4	X

(b)



(c)

PQ	00	01	11	10
S_0	S_1	S_0	S_0	S_1
S_1	S_1	S_2	S_0	S_1
S_2	S_1	S_0	S_0	S_1
—	—	—	—	—
	(2)	(6)	(X)	(4)

A	B
0	0
0	1
1	1
1	0

(d)

	$\bar{P}\bar{Q}$	$\bar{P}Q$	PQ	$P\bar{Q}$
$\bar{A}\bar{B}$	r	r	r	r
$\bar{A}B$	r	S	r	r
AB	R	R	R	R
$A\bar{B}$	—	—	—	—

$\bar{P}\bar{Q}$	$\bar{P}Q$	PQ	$P\bar{Q}$
S	r	r	S
s	s	R	s
s	R	R	s
—	—	—	—

For all these reasons the E flip-flop seems to be a network whose potentialities are worth exploring further, and which might well be worth development in integrated form.

Appendix

This appendix indicates how the Karnaugh maps in Fig. 5 were derived. A sequential circuit is required which gives an output whenever the sequences of numbers 2, 6 or 4, 6 occur on a punched tape carrying numbers 0 to 9.

There are binary signals emerging from the tape reader, indicating the presence of numbers 2, 4, 6 or X, where X is any other number. The numbers are recognized on the punched tape by simple combinational circuits. The binary signals from the tape reader can form the inputs to two flip-flops P, Q according to the map in Fig. 8(a). The outputs of P, Q form the inputs to the sequential circuit.

The sequence of operations is shown in the sequence diagram of Fig. 8(b). There are three necessary states of the circuit, S_0, S_1, S_2 corresponding to the initial state, the state when a '2' or '4' has been received, and the final state when a '6' has followed the '2' or '4'. The arrows indicate the transitions necessary for each input at each state.

The information of the sequence diagram is repeated in more convenient form in the state excitation table of Fig. 8(c).

The three states S_0, S_1 S_2 can be represented by three of the four possible combinations of the outputs of two flip-flops A, B. Which outputs are assigned to which states is arbitrary; those chosen here are also shown in Fig. 8(c).

The input excitation maps can now be prepared. In the state S_0, namely $\bar{A}\bar{B}$, the circuit changes to S_1 ($= \bar{A}B$) when the input flip-flops are in state $\bar{P}\bar{Q}$. Thus flip-flop B is changed to '1', while A is left at '0'. Therefore an r is placed in the top left-hand corner of the map for flip-flop A and an S is placed in the top left-hand corner of the map for flip-flop B. Similarly all the other elements of the map are entered as in Figs. 8(d). There is no state $A\bar{B}$, so the bottom row of the maps remains blank. These are the maps of Fig. 5.

Of course different state assignments and a different assignment of numbers to flip-flops P, Q would have given rise to different maps. No attempt has been made to minimize combinational circuits at this stage although improvements are easy to find.

The Boolean equations for the combinational circuits of flip-flop B are shown below.

$$\left.\begin{array}{l} S = \bar{Q} \\ R = PQ + AQ \end{array}\right\} R\text{-}S \text{ or } J\text{-}K,\ S, R \text{ flip-flops}$$

$$\left.\begin{array}{l} S = \bar{Q} + \bar{A}\bar{P} \\ R = Q \end{array}\right\} E \text{ flip-flop}$$

$$X = \bar{A}\bar{B}\bar{Q} + BPQ + AQ \quad \} \ T \text{ flip-flop}$$

$$X = \bar{Q} + \bar{A}B\bar{P} \quad \} \ D \text{ flip-flop}$$

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A New Method of Radio Receiver Construction

An advanced version of a module system, originally designed for the construction of television receivers, has now been applied to radio receivers by Hede Nielsens Fabriker of Horsens, Denmark.

With this method of construction a conventional a.m.-f.m. radio receiver consists of no more than five separate sub-units or modules which together comprise about 95 per cent of the total number of components.

These modules, all of standard size (approx. 2in$\times$1¼in $\times$¾in high), are fully screened and fitted with pin connexions. They are made on a mass production basis and are pre-aligned before plugging in to the receiver chassis.

The five standard modules consist of the following types:

Module No. 1, an f.m. tuner with three tuned circuits giving an overall gain of 33dB. The tuning capacitor is replaced by three silicon diodes, voltage controlled by a potentiometer. A band-pass filter is installed between the r.f. stage and the mixer.

When used with four i.f. modules, a sensitivity better than 1μV for an output of 5mV from the f.m. detector is provided.

Module No. 2 contains the a.m. mixer and an f.m.-i.f. amplifier stage providing an f.m. gain of 23dB and an a.m. gain of 32dB.

Module No. 3 is an i.f. amplifier stage for f.m. and a.m. together with an automatic gain control circuit which is applied to the a.m. mixer and the first a.m.-i.f. amplifier.

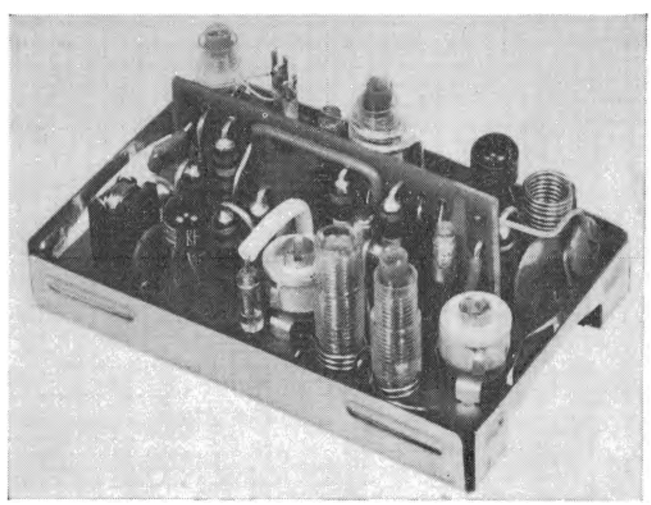
Module No. 4 is the f.m.-i.f. amplifier and a.m. detector stage.

Module No. 5 consists of an f.m. driver stage, a diode limiter and a ratio detector.

Modules with an output up to 5W are designed for the audio amplifiers required for the various types of receivers. For stereo broadcast reception a stereo decoder is housed in two standard modules.

Silicon transistors are used throughout.

One of the modules with the cover removed



Transfer Function Measurement Using Fast Pulses

(Part 1)

By R. C. French*, A.M.I.E.R.E.

The time waveform at the output of an electric circuit is uniquely determined by the time waveform at the circuit input and the circuit gain and phase against frequency response, i.e. its transfer function. Conversely, if the input and output waveforms are known then the transfer function can be calculated from them. A system for measuring circuit transfer functions has been developed in which the input and output waveforms are sampled in time and automatically recorded on punched paper tapes. A computer is used to calculate the circuit transfer function from the paper tapes and the results are printed on either a typewriter or high speed graph plotter in the form of gain, phase and group delay against frequency. The accuracy of the measurement is a function of the type of transfer function being measured; in the laboratory experiment described the highest accuracy obtained was ±0·3dB gain and ±2° phase with a dynamic range of 10dB. The system operates over the band up to 100Mc/s but much higher bandwidths are possible.

(Voir page 553 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 560)

THE method of measuring a circuit's transfer function described here stems from the idea of delta function testing. A delta function, which is a zero width spike of unit energy content, has frequency components of constant amplitude at all frequencies. Since the delta function illuminates the entire band uniformly, the circuit output waveform is uniquely determined by the circuit transfer function. A time delta function cannot be generated practically but a short pulse, which has spectral components with an approximately $(\sin x)/x$ envelope, can be used to approximate it. However, only a small part of the pulse spectrum at the low frequency end of the band contains components of sufficiently uniform amplitude. To cover the band up to 100Mc/s pulses only 0·1nsec wide would be required, and to obtain an adequate signal-to-noise ratio a large amplitude would be necessary. Many circuits would be non-linear with large input pulses and frequency components in the output due to input signals at the same frequency would be confused with signals generated by non-linearity from inputs at subharmonic frequencies.

The method described uses the alternative approach of a lower amplitude pulse with a greater width and recording the input waveform as well as the output waveform. The same equipment can be used to measure first the output waveform and then the input. The actual spectrum of the input is then taken into account in the calculation of the transfer function.

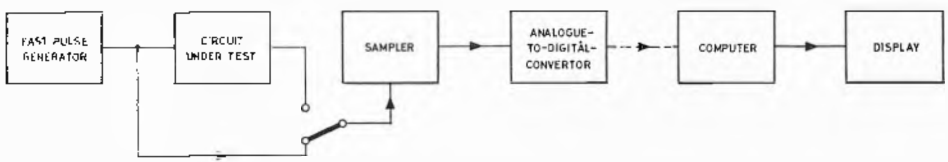


Fig. 1. The measurement system

The Measurement System

The measurement system is shown in Fig. 1. The measurement begins by sampling the time waveform of the fast pulse supplied by the fast pulse generator to the circuit input. The set of samples obtained is converted to digital form in the analogue to digital convertor (a.d.c.), and fed to the computer store. The sampler is then connected to the circuit output and the output waveform similarly sampled and fed via the a.d.c. to the computer. The parameters of the measurement are also given to the computer, i.e. the number of samples taken, the apparent time between samples, and the frequencies at which the transfer function is to be computed.

Both waveforms are Fourier analysed by the computer to yield two sets of spectral lines of defined amplitude and phase over the band of frequencies specified. The computer then takes the lowest spectral frequency and compares the amplitude and phase of the input waveform spectral line with the amplitude and phase of the output waveform spectral line. The gain and relative phase are stored and the calculation repeated for the remainder of the spectral frequencies in turn. The transfer function is then displayed in the form of amplitude and phase against frequency. The phase data is used to calculate phase delay or group delay.

Theoretical Considerations

The method of measuring transfer functions described involves time sampling of a waveform and subsequent Fourier analysis of the waveform using the sampled data. For these two operations to be valid a number of conditions must be fulfilled.

The Sampling Procedure

A time waveform is sampled by recording its instantaneous amplitude at equally spaced points in time. The sampling theorem[2] gives a maximum value to the time between samples t if the set of samples is to correctly describe the waveform. The sampling frequency F_S is defined as $1/t$ and the highest frequency component present in the waveform being sampled is called F_H. The sampling theorem states that:

$$F_S \geqslant 2\,F_H$$

In practice this gives a time between samples which is far too small for present samplers to resolve and much too high a data rate for available a.d.c. circuits. This difficulty is overcome by using a repetitive input pulse

* *Mullard Research Laboratories.*

of p.r.f. F_p, and only taking one sample from each pulse, as shown in Fig. 2. The first sample is made at time t after the start of the first pulse, and the second at time $2t$ after the start of the second pulse. The k^{th} sample is taken kt after the start of the k^{th} pulse. To construct a record consisting of N samples, N identical pulses have to be produced. In order to correctly position the k^{th} sample in the k^{th} pulse the beginning of the pulse must be accurately found, and the quantity $(k-1)t$ stored at the time the $(k-1)^{th}$ sample was made.

If the time between pulses is T_p the time between samples is $T_p + t$. However, assuming N identical input pulses, the set of samples is identical with a set taken from

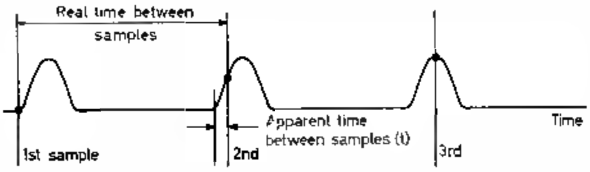


Fig. 2. Sampling procedure

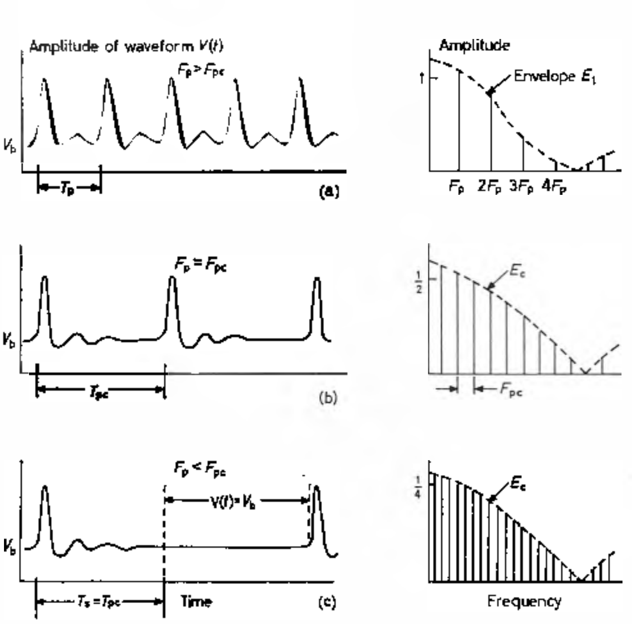


Fig. 3. Dependence of pulse spectrum on p.r.f.

only one pulse with a time between samples of t. The apparent time between samples is therefore t. The total time period over which samples are taken is $NT_p + Nt$ and the apparent sample period T_S is equal to Nt.

The sampling process reduces the volume of data recorded to the minimum necessary to compute the transfer function to a given accuracy. The principle of taking only one sample per pulse to reduce the data rate from F_S to approximately F_p can be extended by taking a sample say once every 10^{th} pulse to give a data rate of $F_p/10$.

Fourier Analysis

Fourier analysis is used to find the amplitude and relative phase of all the frequency components present in a time waveform. This can only be done if the shape of the time waveform is known, or can be obtained from a set of samples describing it. For a repetitive pulse at a p.r.f. F_p the first frequency component, neglecting the d.c. component, is at the frequency F_p. All remaining frequency components, or spectral lines, are at harmonics of F_p.

The amplitude of the spectral lines and the shape of their envelope E are shown in Fig. 3(a) for a typical repetitive pulse. The p.r.f. is high enough for the pulses to overlap; the disturbance due to a pulse lasting long enough to be observable when the next pulse is generated. The spectral line envelope E is given by:

$$E = kF(f)$$

where $F(f)$ is a function of frequency determined by the shape of the time waveform and k is a constant determined by the pulse duty ratio. The value of k determines the absolute magnitude of the spectral lines but not their relative size. If the p.r.f. of the pulses is reduced the spectral lines get closer together and the value of k decreases. The time waveform will change due to the decrease in the pulse 'overlap' and so the function $F(f)$ will change. As the p.r.f. is reduced further a p.r.f. is reached, called the critical p.r.f. F_{pc}, below which the overlap of the pulses is insignificant and can be ignored. Below F_{pc} the function $F(f)$ is independent of the p.r.f. and therefore the envelope E is also independent, neglecting the scaling factor k.

The waveform is shown in Fig. 3(b) at the critical p.r.f. together with the corresponding set of spectral lines and envelope E_c. Fig. 3(c) shows the pulses with a p.r.f. less than F_{pc}. The spectral lines are closer together than in Fig. 3(b) but the envelope is still E_c. Suppose that the waveform of Fig. 3(c) is sampled over the time interval $T_S = T_{pc}$ as shown. The waveform from the end of the period T_S to the beginning of the next pulse is everywhere at a constant voltage V_b. Since the waveform is known everywhere outside the period $T_S = T_{pc}$, there is no need to take actual samples outside this period. The set of samples taken during the period T_S can be supplemented with dummy samples, all having the value V_b, to produce a valid waveform having any p.r.f. from F_{pc} to zero, so that the Fourier analysis can be done with any spectral line spacing required. In this way the measurement resolution can be made as large as required without having to take more samples than are necessary to cover the period T_{pc} in which the pulse occurs.

Choice of System Mode

The system can be used either in the fast mode, with a p.r.f. greater than the critical value F_{pc}, or in the slow mode, $F_p < F_{pc}$. In the fast mode the apparent sample period must be exactly one or more p.r.f. periods T_p. The apparent time between samples must therefore be an integer sub-multiple of T_p.

$$t = T_p/N$$

In the slow mode, samples are only taken over the period $T_S = T_{pc}$ so that in general there is no need for an exact relationship between t and T_p which makes the system easier to build.

In order to illustrate the difference between the two modes an example will be taken of a measurement over the band 0 to 100Mc/s using a sampler capable of taking 100 samples. Assuming the fast pulse is filtered and that all components are negligible above 200Mc/s the parameters are:

$$F_H = 200\text{Mc/s} \qquad F_S = 400\text{Mc/s}$$

Then:

$$t = 2{\cdot}5\text{nsec} \qquad N = 100$$

so that:

$$T_S = Nt \qquad = 250\text{nsec}$$

The sampling theorem prohibits any increase in t so that T_S, the sampling period, is fixed by the value of N. In the slow mode the fast pulse and any ringing after it must be shorter than 250nsec. At the circuit input this condition is easily met, but the circuit output is largely determined by the transfer function and any transmission poles near the origin will produce slowly decaying oscillations which may easily last longer than 250nsec. The value of t can only be increased when measuring the output waveform if the circuit has a low-pass characteristic which adequately attenuates the high frequency components. However, there are many circuits which do not have a low-pass characteristic and it would be valuable to be able to measure broadband circuits with any type of transfer function. If the waveform lasts longer than T_S the measurement is invalid when using the slow mode.

Under the same conditions the fast mode can be used. The p.r.f. of the input pulses is adjusted so that:

$$F_p = 1/T_S = 1/Nt > F_{pc} \qquad = 4\text{Mc/s}$$

The output pulses, and possibly the input pulses, now overlap but any period of 250nsec from the waveform will give valid results in the Fourier analysis, so long as the waveforms have reached their steady state. The values used in the example gave an $F_p = 4$Mc/s and this is the spacing of the measurement frequencies. Such a resolution may not be adequate and it can only be improved by increasing the number of samples N. The rather small value of N used in the example could certainly be exceeded in practice so that a higher resolution could be achieved, together with a more convenient data rate than 4Mc/s for the a.d.c. circuits. Circuits having a difficult transfer function will generally need more samples if the system is used in the slow mode, than are needed to obtain sufficient resolution in the fast mode. Should it not be possible to obtain either adequate resolution in the fast mode or a long enough sampling period in the slow mode, then there are two methods of avoiding the difficulty. In the fast mode the p.r.f. can be set to a number of values in the region of the minimum allowed by the sampling theorem which will give some intermediate frequencies between the closest spacing otherwise achieved in a single experiment. A better approach is to filter the input pulse to exclude those components which produce long rings. The high frequency components can then be properly sampled by using a small t without the need for a large sampling period. The measurement can then be made at the frequencies at which ringing occurs but at a larger value of t permitted by low-pass filtering of the input. Such piecemeal approaches are, however, undesirable and a large enough value of N should be provided so that they are only used in extreme cases.

The experimental model described below was used in the slow mode because the equipment used for the a.d.c. and the temporary storage of the waveform samples were far too slow to permit realistic p.r.f.'s in the fast mode. Also the coherence between the F_S and F_p essential in the fast mode would have been very difficult with the sampler that was used. Both these problems could be overcome in modern digital custom built equipment.

Organization of Complete System

The organization of the complete system depends on whether the computer is conventional or on line time shared. With an on line computer data can be accepted direct from the measurement as it is performed.

Use of Conventional Computers

Conventional computers normally cannot be used on line for reasons of economy, so that their input data has to be stored temporarily until the computer is available for use. The most common medium for this temporary storage is punched tape or cards. Due to the low data rate of tape or card punches the system can only be run in the slow mode, unless the sampling process is extended to give very low data rates. However, as was shown above, the slow mode can have disadvantages. The use of direct electrical input through an interface matching unit (i.m.u.) would allow the use of a core store which, with its high writing speed, would make the fast mode possible with a conventional computer. Such a core store would typically need a capacity of only a few thousand bits, and could be quite economic as a temporary store. It would be possible to use the system in the fast mode, store the information at the necessarily high data rate in a core store, and then transfer the data to punched tape at low speed. It would then be possible to use the fast mode with a computer which did not have an i.m.u. The use of two stores in cascade is, however, rather expensive and cumbersome.

The use of a graph plotter output, which is possible with most computers, is advantageous as most gain or phase responses are required in graphical form eventually, so that a lot of manual plotting is avoided.

Use of On Line Computer

With an on line connexion the data can be written into the computer core store at high speed and the output displayed very rapidly on a c.r.t. display sited at the experiment. It is then possible to make circuit changes or adjustments and have an almost immediate display of the effect on the circuit gain and phase response. The parameters of the measurement are simultaneously read by the computer before each calculation from a number of variable controls mounted on the output display. The setting for example of the bandwidth required and its centre frequency could be done by potentiometers as in a conventional wobbulator; the voltage across the potentiometers being converted to digital form for feeding to the computer.

It is considered that a system in which the effect of changes can be seen straight away would be very valuable in developing all circuits in which the complete transfer function has to be controlled, for example video circuits, filters and data or television links.

The Computer Programme

The computer programme is written in Elliot Algol 1 and consists of four basic parts. These are read procedures to read in the data tapes and parameters. Numerical Fourier analysis of the input and output data, comparison of input and output spectra, and the output instructions.

The procedure which performs a numerical Fourier analysis[3] is given below:

```
procedure Fourier Trans (Flow, Fint, t, M, N, C, S, P);
  real Flow, Fint, t; integer M, N; array C, S, P;
  begin real V, V1, V2, F, CS; integer I, J, K;
    K: = 1;
    F: = Flow;
    for I: = 0 step 1 until M do
    begin
      V1: = 0;
      V2: = 0; CS: = cos (6·283*F*t);
      for J: = K step −1 until 1 do
      begin
        V: = 2*V1*CS + P [J] −V2;
        V2: = V1;
        V1: = V;
      end;
        C[I]: = checkr (2/N* (P[O]−V2+V1*CS));
```

$S[I] := \text{checkr}\,(2/N^*V_1^*\sin(6{\cdot}283^*F^*t));$
$F := F + F_{\text{int}};$
end;
end;

where F_{low} = Frequency of lowest spectral line
F_{int} = Frequency interval between lines
M = Number of lines
t = apparent time between samples
N = Number of samples of input and output waveform
P = $P[R]$ = array of samples of time waveform $k = 0 \to M$
C = $C[k]$ = array of cosine coefficients
S = $S[k]$ = array of sine coefficients

Each spectral line is calculated as a sine and cosine component which are later used to calculate the line amplitude and phase. After the analysis of the input waveform the sine and cosine components are stored in the arrays $S_1[k]$ and $C_1(k)$. The components of the output waveform Fourier spectrum are stored in the arrays $S_2[k]$ and $C_2[k]$. The parameter k denotes a spectral line number in the range $0 \to M$. The procedure contains two *for* loops, the first of which steps over the range of spectral lines. The second, which is nested in the first, steps over the waveform samples starting with the last sample and working back to the first. The value of the sine and cosine components of any particular spectral line are independent of the number of consecutive zero samples occurring at the end of the set of samples. This can be seen by examining the instructions in the inner *for* loop. It is this independence which allows the choice of any spectral line frequencies when the system is used in the slow mode.

The program first reads the parameters of the measurement and then the input waveform data. The Fourier analysis procedure is used to calculate $S_1[k]$ and $C_1[k]$. The output waveform data is then read and $S_2[k]$ and $C_2[k]$ calculated. The two sets of coefficients are then used to find the transfer function:

$$\text{Gain} = \underset{k=0\to M}{G[k]} = 8{\cdot}686 \ln\left[\frac{\sqrt{(S_1[k]^2 + C_1[k]^2)}}{\sqrt{(S_2[k]^2 + C_2[k]^2)}}\right]$$

$$\text{Phase} = \underset{k=0\to M}{\beta[k]} = \arctan\frac{C_2[k]}{S_2[k]} - \arctan\frac{C_1[k]}{S_1[k]}$$

$$\text{Group Delay} = \underset{k=1\to M}{GD[k]} = \frac{\beta[k] - \beta[k-1]}{2\pi F_{\text{int}}}$$

The calculation of group delay is an approximation which improves as F_{int} is decreased:

$$\frac{\beta[k] - \beta[k-1]}{2\pi F_{\text{int}}} = \Delta\beta/\Delta\omega \to d\beta/d\omega$$
$$\text{as } \Delta\omega \to 0$$

The value of F_{int} must be carefully chosen. Too large a value gives a poor approximation, and too small a value accentuates the error due to noise on the phase characteristic. Let the phase error at one frequency relative to its adjacent frequency be β_e. This error is assumed independent of line spacing, which in practice is approximately true. The error in the group delay GD_e is then obtained from:

$$GD + GD_e = \frac{\beta[k] - \beta[k-1] + \beta_e}{2\pi F_{\text{int}}}$$

when $GD_e = \dfrac{\beta_e}{2\pi F_{\text{int}}}$

so that GD_e increase as F_{int} decreases. The value of F_{int} should be no smaller than necessary to follow the fine structure of the phase characteristic.

The gain characteristic is set relative to the maximum gain in the band, which is set equal to 0dB. Any region where the gain is less than −30dB is discarded as the signal-to-noise ratio is too small for acceptable results. The phase characteristic is examined for phase jumps of 2π and these are removed to yield a phase characteristic free of discontinuities. The group delay characteristic is calculated from the corrected phase data and the minimum delay set to zero, all other delays are then relative to the minimum delay.

Finally the data is plotted on a graph plotter with axes scaled by the computer to ensure that the graphs fill the paper, consistent with sensible intervals on the axis.

The program was written for use with a system working in the slow mode with a conventional computer. Only those parts of the program controlling peripheral devices would require serious change if the fast mode were required. The use of an on line computer would require more extensive programming but the essential part of the program, the Fourier analysis and spectral comparison, would be valid.

Specification of Fast Pulse

The fast pulse used in the measurement must have a reasonably constant spectral line amplitude over the band, so that the best signal-to-noise ratio can be maintained. The envelope of the spectral lines of a typical pulse is shown in Fig. 3. The finite rise and fall times produce a small spectral line amplitude at the higher frequencies and there is a null in the spectrum at the frequency $1/T$ where T is the pulse width. The width must therefore be made sufficiently small so that:

$$1/T > F_H$$

where F_H = Highest measurement frequency.

Clearly no measurement can be made at the frequency $1/T$ as no energy is supplied to the circuit. The null at $1/T$ is repeated at integral multiples of $1/T$. If the system was used to measure only a narrow band of high frequencies a wide pulse could be used if its output was band-pass filtered with the passband lying between, say, $1/T$ and $2/T$. This would only be desirable if sufficiently short pulses could not be generated.

The discussion on 'choice of system mode' showed the need to use as low a sampling frequency F_S as possible for a given upper working frequency. The sampling theorem states that for a set of samples to describe correctly a time waveform the sampling frequency must be at least twice the maximum frequency present in the waveform. The frequency components of the fast pulse decrease only slowly with frequency, whereas the system requires the most rapid decrease possible. The fast pulse is therefore low-pass filtered with a steep cut-off filter, consistent with not producing excessively long rings. A practical system would require a range of filters so that the lowest possible sampling frequency could be used for any particular circuit.

The amplitude of the fast pulse should be large in order to achieve a good signal-to-noise-ratio in the sampler. The amplitude must not however be so large that the nonlinearity of either the circuit or the sampler materially alters the shape of the pulse. The analysis and computation are only valid for linear circuits.

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(To be continued)

Design and Group-delay Correction of a Low-pass Filter for use in Video Circuits

By J. W. Head*, M.A., F.Inst.P., A.M.I.E.E., F.I.M.A., and D. Howorth*, B.Sc.Tech., A.M.I.E.E.

This note discusses the detailed application of a previously published design procedure (including group-delay correction) to obtain a low-pass filter having a uniform response up to 5·5Mc/s, theoretically infinite attenuation at 8·25Mc/s and attenuation not less than 46dB above 8·25Mc/s. Practical results are given which were obtained when this procedure was used to design a filter that was significantly smaller than those of similar characteristics which were previously available.

(Voir page 561 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 568)

THE theoretical requirement, ignoring dissipation, was that the filter should be low-pass, with constant response (within 0·5 per cent) at frequencies up to 5·5Mc/s, give theoretically infinite attenuation at 8·25Mc/s and at least 46dB of attenuation at frequencies above 8·25Mc/s and be substantially corrected for group-delay distortion. The filter was also required to be of small physical size so that it might be included in standard video units: this would enable it to be used, for example, in the output of a derivative equalizer to remove any high frequency noise components outside the video band which might be increased in amplitude by the derivative generator or in the output of a video detector unit to remove unwanted carrier components. It was found that the response/frequency requirements could be met with the symmetrical attenuating network of Fig. 1. The determination of the inductance and capacitance values for the attenuating network follows closely the procedure of Davies and Head[1].

For design calculations, it is preferable to work with normalized units, so that the terminating resistances are taken as unit resistances and the upper limit of the pass-band is taken as $1/2\pi$. Further, it is considered justifiable to assume that all filter elements are purely reactive except the terminating resistances. The actual values of capacitances and inductances required when the terminating resistances are 75Ω and the upper limit of the pass-band is 5·5Mc/s are easily deduced.

Determination (in Normalized Units) of the Element Values of Fig. 1

It is necessary first to determine a normalized response function at frequency $\omega/2\pi$ of the form:

$$T(\omega^2) = \frac{\mu^2\{U(\omega)\}^2 + \{V(\omega)\}^2}{\{V(\omega)\}^2} \quad \ldots\ldots\ldots\ldots \quad (1)$$

which will meet the requirements, where μ is a constant. It will be assumed initially that $U(\omega)$ is of degree 7 and odd in ω, and:

$$V(\omega) = 1 - (4/9)\,\omega^2 \quad \ldots\ldots\ldots\ldots \quad (2)$$

In fact it is found that the requirements can then be met: had this not been so it would have been necessary to consider increasing the degree of $U(\omega)$ or $V(\omega)$ or both, or altering the ratio of the (theoretical) infinite-attenuation frequency to the nominal cut-off frequency. Equation (2) gives the simplest possible form of $V(\omega)$ which is even in ω and zero when $\omega = 1·5$. It is not obligatory to have $U(\omega)$ odd and $V(\omega)$ even, but this makes the network of Fig. 1 electrically symmetrical when the input and output terminating resistances are equal, and in this case the network turns out to be physically symmetrical as well, as is indicated by the notation of Fig. 1.

If the variations of V in the pass-band are ignored, the form of U which would give equal maxima in the pass-band and minima which are zero is:

$$U = \omega(\omega^2 - \sin^2 \pi/7)(\omega^2 - \sin^2 3\pi/7)(\omega^2 - \sin^2 5\pi/7) \quad \ldots\ldots \quad (3)$$

$$= \omega(\omega^2 - 0·18825)(\omega^2 - 0·61126)(\omega^2 - 0·95048) \quad \ldots\ldots \quad (4)$$

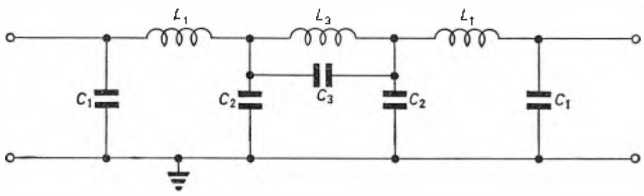

Fig. 1. General arrangement of the filter (without group delay correction)

since U is then $0·015625 \sin(7 \sin^{-1} \omega)$. By adjusting the zeros of U to take account of the variations of V in the pass-band so that the maxima of $|U/V|$ in the pass-band are approximately equal (as explained[1]) a more suitable expression for U is found to be:

$$U = \omega(\omega^2 - 0·2219)(\omega^2 - 0·658)(\omega^2 - 0·964) \quad \ldots\ldots \quad (5)$$

which gives the following key values of $|U/V|$ in the pass-band and of $|V/U|$ in the stop band:

TABLE 1

Key Values of $|U/V|$ in Pass Band and of $|V/U|$ in Stop Band

PASS BAND			STOP BAND		
ω	ω^2	$\lvert U/V \rvert$	ω	ω^2	$\lvert V/U \rvert$
0·272	0·074	0·0216	1·8708	3·5	0·00755
0·663	0·440	0·0205	2	4	0·01014
0·901	0·811	0·0194	2·121	4·5	0·00811
1	1	0·0172	2·236	5	0·00653

This is regarded as completely satisfactory, so that the maximum $|U/V|$ in the pass-band can be taken as 0·022 and the maximum $|V/U|$ in the stop band as 0·01.

Since the maximum permitted pass-band ripple is 0·5 per cent, it is necessary that:

$$\{1 + \mu^2 \times (0·022)^2\}^{\frac{1}{2}} < 1·005; \quad \mu < 4·5 \quad \ldots\ldots \quad (6)$$

and since the minimum stop band loss is 46dB:

$$\frac{0·01}{\mu} < 0·005 \qquad \mu > 2 \quad \ldots\ldots\ldots \quad (7)$$

so the obvious value for μ is $(2 \times 4·5)^{\frac{1}{2}} = 3$, and the resulting amplitude function is:

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$$T(\omega^2) = 1 + \frac{9\omega^2(\omega^2-0{\cdot}2219)^2(\omega^2-0{\cdot}658)^2(\omega^2-0{\cdot}964)^2}{(1-(4/9)\,\omega^2)^2} \quad \ldots (8)$$

Enough has been said previously[1] on the procedure for factorizing the polynomial in ω^2 obtained by clearing equation (8) of fractions; the result alone is relevant here, namely:

$$(1/9)\,(1-(4/9)\,\omega^2)^2\,T(\omega)^2 =$$
$$(\omega^2+0{\cdot}317726)\,(\omega^4-2{\cdot}339914\omega^2+1{\cdot}405925)\times$$
$$(\omega^4-0{\cdot}126132\omega^2+0{\cdot}282214)\,(\omega^4-1{\cdot}539480\omega^2+0{\cdot}881391) \quad \ldots\ldots\ldots\ (9)$$

Next, it is necessary to determine the A-matrix of the filter, and factorize this A-matrix to obtain the element values for Fig. 1. For this, one must deduce from equation (9) an expression $f(p)$ (where p is strictly d/dt but for the purposes of this note can be taken as $j\omega$) free from zeros in p with positive real parts. Corresponding to the factor $\omega^2 + 0{\cdot}317726$ of $T(\omega^2)$ in equation (9), $f(p)$ has a factor:

$$1 + p/(0{\cdot}317726)^{\frac{1}{2}} = 1 + 1{\cdot}774081p \quad \ldots\ldots\ (10)$$

and if $T(\omega^2)$ has a factor proportional to:

$$1 - 2r^2\omega^2\cos 2\theta + r^4\omega^4 \quad \ldots\ldots\ldots\ldots\ (11)$$

then $f(p)$ has a corresponding factor:

$$1 + 2rp\sin\theta + r^2p^2 \quad (0 < \theta < \pi/2) \quad \ldots\ldots\ (12)$$

Thus:

$$(1 + (4/9)\,p^2)\,f(p) =$$
$$(1 + 1{\cdot}774081p)\,(1 + 0{\cdot}149967p + 0{\cdot}843371p^2)\times$$
$$(1+0{\cdot}619550p+1{\cdot}065162p^2)\,(1+1{\cdot}821527p+1{\cdot}882395p^2) \quad \ldots\ldots\ldots\ldots\ (13)$$

$$= \{1 + 9{\cdot}882258p^2 + 16{\cdot}156501p^4+6{\cdot}872368p^6\} +$$
$$\{4{\cdot}365125p + 15{\cdot}153440p^3 + 13{\cdot}402877p^5 + 3p^7\} \quad \ldots\ldots\ldots\ldots\ (14)$$

$$= E_1(p) + O_1(p), \text{ say}$$

where $E_1(p)$ denotes the terms of even degree in p and $O_1(p)$ denotes the terms of odd degree. If further a corresponding expression* is derived from equation (5):

$$g(p) = 3p(p^2 + 0{\cdot}2219)\,(p^2 + 0{\cdot}658)\,(p^2 + 0{\cdot}964) \quad \ldots\ (15)$$

then the required A-matrix can be taken as:

$$A = \frac{1}{1 + (4/9)\,p^2}\begin{bmatrix} E_1(p) & O_1(p) - g(p) \\ O_1(p) + g(p) & E_1(p) \end{bmatrix} \quad \ldots (16)$$

$$= \frac{1}{1 + (4/9)\,p^2}\times$$
$$\begin{bmatrix} 1+9{\cdot}882258p^2+16{\cdot}156501p^4 + 6{\cdot}872368p^6 & 3{\cdot}942864p+12{\cdot}170739p^3 + 7{\cdot}871177p^5 \\ 4{\cdot}787386p+18{\cdot}136141p^3 + 18{\cdot}934577p^5+6p^7 & 1+9{\cdot}882258p^2+16{\cdot}156501p^4 + 6{\cdot}872368p^6 \end{bmatrix} \quad \ldots\ldots\ (17)$$

The routine for factorizing the A-matrix (17) has been fully described[2]; the factors are:

$$A = \begin{bmatrix} 1 & 0 \\ 0{\cdot}873084p & 1 \end{bmatrix}\begin{bmatrix} 1 & 1{\cdot}423257p \\ 0 & 1 \end{bmatrix}\begin{bmatrix} 1 & 0 \\ 1{\cdot}519803p & 1 \end{bmatrix}\begin{bmatrix} 1 & \dfrac{1{\cdot}094698p}{1+4(p^2/9)} \\ 0 & 1 \end{bmatrix}$$
$$\times\begin{bmatrix} 1 & 0 \\ 1{\cdot}519803p & 1 \end{bmatrix}\begin{bmatrix} 1 & 1{\cdot}423257p \\ 0 & 1 \end{bmatrix}\begin{bmatrix} 1 & 0 \\ 0{\cdot}873084p & 1 \end{bmatrix} \quad \ldots\ldots\ldots\ldots\ (18)$$

* *The coefficient of p^7 in equation (15) is made 3 so that $O_1(p) - g(p)$ shall be of lower degree than $O_1(p) + g(p)$. This permits realization with as few inductances as possible.*

TABLE 2

Phase Calculations Associated with the Circuit of Fig. 1 and Correcting All-Pass Networks

ω	θ_1	θ_2	θ_3	θ_4	$\theta(\omega)$	$K\omega$	$\phi(\omega)$
0	0	0	0	0	0	0	0
0·1	10°3½′	0°52′	3°35′	10°26′	24°56′	121°37·2′	48°20′
0·1865	—	—	—	—	—	—	90°
0·2	19°32′	1°47′	7°22½′	21°30′	50°11½′	243°14·4′	96°31′
0·37	33°17′	3°35′	15°1′	42°14′	94°7′	449°59·6′	177°56′
0·3744	—	—	—	—	—	—	180°
0·38	33°59′	3°43′	15°33′	43°39′	96°54′	462°9·4′	182°38′
0·56	44°51′	6°31′	27°31′	68°7′	147°0′	681°4·3′	267°2′
0·5664	—	—	—	—	—	—	270°
0·57	45°21′	6°43′	28°22′	69°29′	149°55′	693°14·0′	271°39½′
0·76	53°26′	12°32′	50°45′	93°7′	209°50′	924°18·7′	357°14′
0·7667	—	—	—	—	—	—	360°
0·77	53°48′	13°0′	52°19′	94°38′	213°45′	936°28·4′	361°22′
1	60°35½′	43°45′	96°0′	115°51′	316°12′	1216°12′	450°0′
1·05	61°46′	65°46′	105°0′	119°21′	351°53′	1277°0·6′	462°34′
1·1	62°52′	97°5′	112°58′	122°31½′	395°26½′	1337°49·2′	471°11′

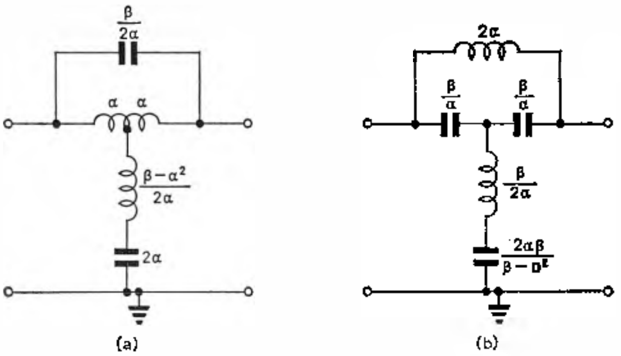


Fig. 2. Standard elements for realizing the delay-correcting network associated with the factors in equation (24)

(a) for $\alpha^2 > \beta$ *(b) for* $\alpha^2 < \beta$

From equation (18), the element values in Fig. 1 can be deduced in normalized units; they are:

$$C_1 = 0{\cdot}873,\ C_2 = 1{\cdot}520,\ C_3 = 0{\cdot}406$$
$$L_1 = 1{\cdot}423,\ L_3 = 1{\cdot}0947 \qquad \ldots\ldots (19)$$

Group Delay Correction (in Normalized Units)

From the transfer function $f(p)$ given by equation (13), it can be deduced that the phase change $\theta(\omega)$ resulting from passage through the network is given by:

$$\theta(\omega) = \tan^{-1}(1{\cdot}774081\omega) + \tan^{-1}\frac{0{\cdot}149967\omega}{1-0{\cdot}843371\omega^2} + \tan^{-1}\frac{0{\cdot}619550\omega}{1-1{\cdot}065162\omega^2} + \tan^{-1}\frac{1{\cdot}821527\omega}{1-1{\cdot}882395\omega^2} \qquad \ldots (20)$$

at (normalized) frequency $\omega/2\pi$. Three all-pass sections are now added to provide a total phase change $2\phi(\omega)$ in such a way that:

$$\theta(\omega) + 2\phi(\omega) \simeq K\omega \quad \ldots\ldots\ldots\ldots\ldots (21)$$

where K is a constant. K has been chosen so that $\theta(\omega)$ shall be 450° when $\omega = 1$; this gives $K = 1216{\cdot}2°$ (per radian of ω).

If the respective terms of the right-hand side of equation (20) are called θ_1, θ_2 and θ_3 and θ_4, Table 2, with a little judicious interpolation, gives the values of $\theta(\omega)$ and $\phi(\omega)$ for various values of ω, and in particular the values of ω for which $\phi(\omega)$ is a multiple of 90°:

The calculations for $\omega = 1{\cdot}05$ and $1{\cdot}1$ check that $\phi(\omega)$ continues to increase at a reasonable rate when $\omega \simeq 1$ and the uncorrected group delay is greatest. Hence the transfer function $W(p)$ for the correcting network can be taken as:

$$W(p) = \left(1 + \frac{p^2}{0{\cdot}1865^2}\right)\left(1 + \frac{p^2}{0{\cdot}5664^2}\right)\left(1 + p^2\right) + \lambda p\left(1 + \frac{p^2}{0{\cdot}3744^2}\right)\left(1 + \frac{p^2}{0{\cdot}7667^2}\right) \qquad \ldots (22)$$

and λ can be chosen so that the phase is correct when $\omega = 0{\cdot}1$. This gives:

$$\lambda = \tan 48°\,20' \times \left[\frac{\text{real part of } W(p)}{\text{imaginary part of } W(p)}\right]_{p=j0{\cdot}1} = 8{\cdot}411 \quad \ldots (23)$$

To determine the correcting networks, it is then necessary to factorize:

$$W(p)\times 0{\cdot}1865^2\times 0{\cdot}5664^2 = p^6 + 1{\cdot}139026p^5 + 1{\cdot}355593p^4 + 0{\cdot}829212p^3 + 0{\cdot}366752p^2 + 0{\cdot}093854p + 0{\cdot}0111585$$

$$\therefore W(p) = (1 + 6{\cdot}2589p + 12{\cdot}5802p^2)(1 + 1{\cdot}7501p + 4{\cdot}5467p^2) \times (1 + 0{\cdot}4020p + 1{\cdot}5669p^2) \quad \ldots\ldots\ldots\ldots (24)$$

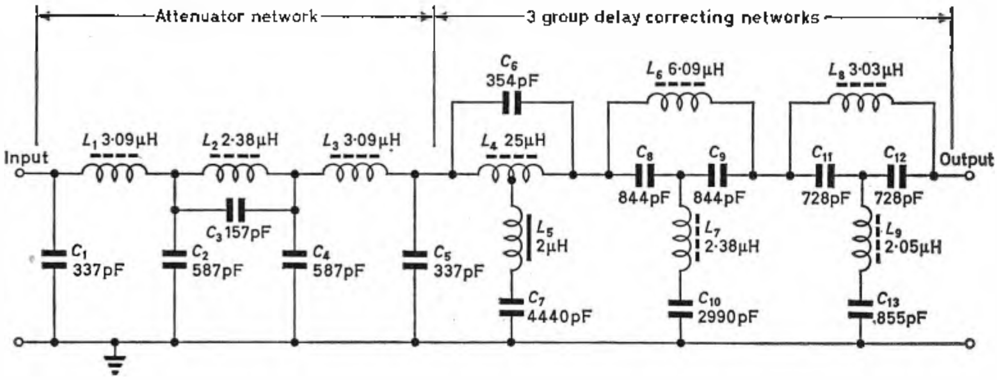


Fig. 3. Practical circuit diagram of a low-pass filter with a characteristic impedance of 75Ω and a cut-off frequency of 5·5Mc/s

A possible realization for group delay correction is then three bridged-T networks in cascade, each having the form of Fig. 2(a), (in normalized units). For practical reasons, however, it may be preferable to use the equivalent form of Fig. 2(b) when $\alpha^2 < \beta$. The values of α, β associated with the factors of $W(p)$ in equation (24), the primary inductance $(\beta + \alpha^2)/2\alpha$, the mutual inductance $(\beta - \alpha^2)/2\alpha$, the coupling coefficient $(\beta - \alpha^2)/(\beta + \alpha^2)$ and the bridging capacitance $\beta/2\alpha$ are given in Table 3:

Practical Construction

Fig. 3 shows the practical circuit diagram of a low-pass filter which has a cut-off frequency of 5·5Mc/s and a characteristic impedance of 75Ω and which has been

TABLE 3
Elements for Phase-Correcting Networks

Values from factors in equation (24) α	β	Primary inductance $(\beta+\alpha^2)/2\alpha$	Mutual inductance $(\beta-\alpha^2)/2\alpha$	Coupling $\frac{\beta-\alpha^2}{\beta+\alpha^2}$	Bridging capacitance $\beta/2\alpha$
6·2589	12·5802	4·134	−2·124	−0·514	1·005
1·7501	4·5467	2·174	0·424	0·195	1·299
0·4020	1·5669	2·150	1·748	0·813	1·949

derived from the theoretical circuits shown in Figs. 1 and 2 and the data given in equations (19) and Table 3.

An experimental attenuating network was constructed and tested to ensure that its attenuation/frequency characteristic was within specification. The specified and measured characteristics of the attenuating sections are shown in Figs. 4(a) and 4(b) respectively. In order to check the impulse response of the complete filter, the experimental attenuating network was cascaded with three 'Cintel Adjustable Phase Correctors' which had been set to the calculated values. It was found empirically that the impulse response of the filter could be improved by a slight adjustment of the settings of the phase correctors and the networks shown in Fig. 3 have in fact the corrected values.

The completed filter was required to occupy a space no greater than 4in × 2½in × 1in and this was achieved by the use of miniature components mounted on a small printed circuit board as shown in Fig. 5. The individual sections of the filter are connected together by means of soldered links which can easily be removed to facilitate the setting-up procedure. Prior to assembly each inductor was adjusted to be within ±1 per cent of its calculated value by means of an inductance bridge and each capacitor was selected to be within ±1 per cent of its specified value by means of a capacitance bridge. In order to increase the probability of being able to select the required value of capacitance from a given batch, provision was made on the printed circuit board for two capacitors to be used in parallel. With the type of miniature core assembly used, all the inductors had a magnification factor (Q) in excess of 100 over the whole of the pass-band except that of L_4 which had a magnification factor of 70 at low frequencies and 40 at the cut-off frequency.

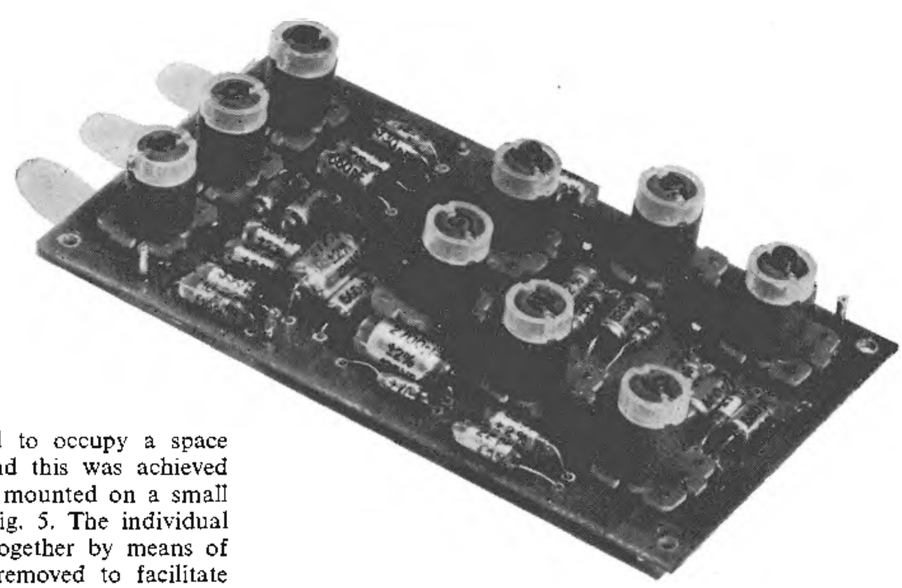


Fig. 5. The complete 5·5Mc/s low-pass filter

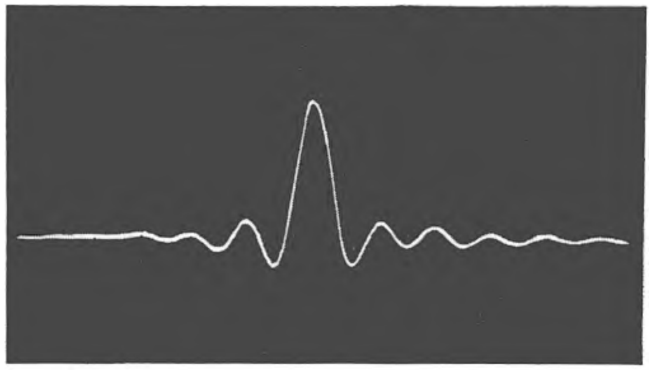
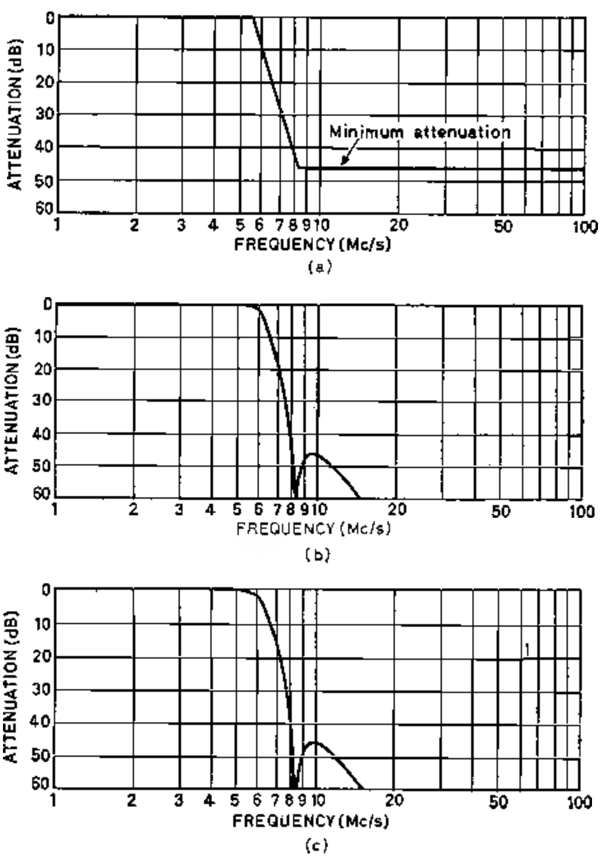


Fig. 6. The impulse response of the 5·5Mc/s low-pass filter shown in Fig. 5.

Fig. 4. Required and realized attenuation/frequency characteristics
(a) Specified attenuation/frequency characteristic
(b) Attenuation/frequency characteristic of attenuation network
(c) Attenuation/frequency characteristic of complete filter



After assembly the individual networks of the filter were adjusted as follows. First L_2 was adjusted to give maximum attenuation at 8·25Mc/s. The inductors L_1 and L_3 were then adjusted to give the terminated attenuating network an impedance as close as possible to 75Ω in the pass-band, consistent with the impedance being symmetrical. This was carried out by driving the terminated attenuating network from a 75Ω video sweep-frequency generator and observing the voltage developed across the input terminals. Symmetry was checked by reversing the connexions. Subsequently, each of the three group-delay correctors were adjusted in turn to have an impedance of 75Ω in a similar manner. The soldered links were then replaced and the overall performance of the filter measured.

Performance

The measured attenuation/frequency characteristic of the complete filter is also shown in Fig. 4(c). The attenuation of the filter at the high-frequency end of the pass-band is principally caused by the relatively low magnification-factor of the inductor L_4. It is difficult to see how this inductor could be made to have a sufficiently high magnification-factor while maintaining its small physical size. However, the attenuation could be corrected by the addition of a suitable resistor across C_7 or, alternatively, by a constant resistance equalizing section in cascade; this, of course, would introduce a significant insertion loss. The impulse response of the filter was tested by driving it with a rectangular 5nsec pulse. Fig. 6 shows the resulting $\sin x/x$ output pulse and it is estimated that the departure from constant group-delay in the pass-band is no greater than ±25nsec. The delay of the filter is approximately 0·55μsec.

Eight filters of the type described in this article were constructed (with parameters scaled to suit the American 525/60 television standard) with a laboratory effort of 12 man-days, for use in the BBC colour transmission of the 1966 General Election.

Conclusions

The filter described in this note, which was the result of a successful marriage of theoretical calculation and empirical adjustment, has a performance more than adequate for the video applications for which it was intended. There is no reason why the method should not be used to produce filters with a higher rate of cut-off at the expense of more filter elements (and a greater physical size) or to produce filters with a lower rate of cut-off and a lower number of filter elements (and smaller physical size). In either case the minimum attenuation in the stop band may be specified by the designer.

Acknowledgment

The authors wish to express their thanks to the Director of Engineering of the British Broadcasting Corporation for permission to publish this article.

REFERENCE

1. DAVIES, R. E., HEAD, J. W. The Design and Realisation of L.C. Filters of Specified Performance. *Instrum. Rev. 12*, 322, 371 and 424 (1965).

Ground Satellite Station at Ascension Island

A satellite communications ground station has been designed, built and brought into full operation by The Marconi Co. Ltd. at its experimental station at Rivenhall, Essex, in a record time.

Work was started in October last year and after a full range of acceptance tests, the station is now dismantled and being shipped to the Ascension Island in the South Atlantic midway between Africa and South America where it is to be re-erected and in full operation by the end of September for hand-over to Cable & Wireless Ltd.

This satellite ground station is to be linked with Intelsat II, a satellite which is expected to be launched in October at an altitude of 22 300 miles over the Atlantic and so provide communications with the Goddard Space Flight Centre in Maryland via the Andover, Maine, earth station.

It is the first stage of the entry of Cable and Wireless Ltd into space communications, and is part of the APOLLO project for the landing of a man on the moon in 1970. Two more ground satellite stations in connexion with the APOLLO project are being constructed for installation at Grand Canary and Carnarvon, Australia.

There is already an American satellite tracking station on Ascension Island which will be used for the APOLLO project and which will be connected to the Cable and Wireless ground station, thus providing a communication link between the APOLLO space craft and the Goddard Centre.

The station uses a 42ft diameter, steerable, parabolic dish aerial with a focal length of 12ft and with a centre area accuracy to within 0·030in. Even at the edge, the aerial is within 0·075in of the ideal paraboloid.

The dish surface is made up of 2in thick aluminium honeycomb sandwich panels, mounted on an aluminium backing framework. This very rigid structure is attached to an elevation mounting, which is in turn mounted on a patform which can turn through 370° in azimuth. The majority of the transmitter and receiver equipment is mounted on this patform, and both the received and the transmitted signals pass to the ground installation through coaxial cables at the intermediate frequency of 70Mc/s.

The transmitter operates in the 6 000Mc/s frequency band, with a final output power in the region of 15kW. The output klystron, mounted in the transmitter cabin on the back of the dish structure, feeds power through a waveguide system to a horn mounted in the surface of the dish. A Cassegrain system is used, with a small hyperbolic sub-reflector at the focus of the main dish.

The receiver, operating in the 4 000Mc/s frequency band, uses the same Cassegrain feed system and circular waveguide horn. It is also mounted directly on the back of the dish aerial. A cryogenic parametric amplifier, operating at −253°C in a helium system, is used to reduce the noise temperature of the receiving system.

Power supplies and all major elements of both the transmitting and receiving systems are fully duplicated, to ensure maximum reliability. The standby equipment is maintained in a fully prepared condition, and switching to standby for any unit can be effected within a fifth of a second.

The aerial attitude control system is of the auto-flow type. The Cassegrain sub-reflector is mounted slightly off centre and is rotated to produce a conical scanning motion of the aerial beam. This produces variations in the received signal when the axis of the aerial is not exactly pointing at the satellite. These variations are used as an 'error signal' to drive the aerial into the correct position.

The station provides seven 4kc/s channels and Cable and Wireless Ltd has now received enquiries from America for a further 12 voice channels between Ascension Island and North America using the same 42ft dish aerial. This will require the installation of a further £100 000-worth of equipment which will not be used for the APOLLO project.

Cable and Wireless Ltd is extending its satellite communication systems by inviting tenders for three more satellite ground stations.

The first of these is to be installed at Hong Kong and to operate via the Intelset II Pacific satellite with other ground stations in the Pacific area covering Australia, South East Asia, the Far East and North America; and its main purpose is to meet the very heavy demand for telephone channels in the Pacific area which cannot now be supplied by the SECOM and other cable networks in that area.

Cable and Wireless Ltd propose to install its second ground station at Bahrain in the Persian Gulf, but have no firm plans at the moment for the site of the third station.

The 42ft parabolic dish aerial and its associated transmitting and receiving equipment mounted on a temporary tripod during the construction and testing at Rivenhall.

Frequency Transformations for A.C. Networks

By J. G. Advani* and M. N. Neelkantan*

The transformation $p = (2/3)\dfrac{s(s^2+1)}{(s^2-0{\cdot}25)}$ *for an a.c. integrator has been suggested by Prabhakar and is claimed to have ½ per cent error in linearity and symmetry within the frequency range 50 per cent up and down the carrier frequency. It is shown here that this error is about 8 per cent in this frequency range. A new transformation (for a.c. integrator) that gives about 3 per cent error in this frequency range is also presented.*

Another transformation for obtaining transfer functions of a.c. compensating networks from the transfer functions of d.c. compensating networks is also suggested. This transformation gives ½ per cent error in linearity and symmetry within the frequency range 10 per cent up and down the carrier frequency.

The networks for an a.c. integrator and an a.c. lag-lead network using RC and active elements are synthesized.

(Voir page 553 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 560)

THE envelope transfer function for an a.c. system can be obtained from the corresponding d.c. transfer function using the frequency transformation. The ideal transformation is:

$$p = s - j\omega_0 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where ω_0 is the carrier frequency of the a.c. system. This transformation evidently yields an unrealizable function due to the presence of the imaginary coefficient. A commonly used transformation is:

$$p = \frac{s^2+\omega_0^2}{2s} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

Generally most of the a.c. systems operate with a data frequency of about 10 per cent of the carrier frequency, and in this range the error in linearity due to the transformation of equation (2) is found to be as much as 5·5 per cent.

For an a.c. integrator to perform satisfactorily the frequency transformation should be linear within the range $0{\cdot}5\,\omega_0$ to $1{\cdot}5\,\omega_0$. In this range the error in linearity due to the transformation of equation (2) is about 50 per cent.

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To reduce the error in linearity Prabhakar[1] has suggested the transformation:

$$p = (2/3)\frac{s(s^2+1)}{s^2-0{\cdot}25} \quad \ldots\ldots\ldots\ldots \quad (3)$$

p and s in equation (3) are normalized frequency variables.

It is claimed in that article that the transformation of equation (3) has only ½ per cent error in linearity within the desired range of 50 per cent of the carrier frequency on either side of ω_0. It is the contention of the present authors that the error in linearity is 8 per cent and this occurs at $\omega/\omega_0 = 0{\cdot}8$ as shown in Table 1.

A.C. Integrator

Transformation

The frequency transformation (p and s are normalized frequency variables)

$$p = \frac{0{\cdot}5162\, s(s^2+1)(s^2+4{\cdot}45)}{(s^2+4)(s^2-0{\cdot}1836)} \quad \ldots\ldots\ldots \quad (4)$$

is found to be linear within 3 per cent in the frequency range from $0{\cdot}5\,\omega_0$ to $1{\cdot}5\,\omega_0$. $p(j(\omega/\omega_0))/j$ is tabulated for different values of ω in Table 2. The percentage error

TABLE 1

$\dfrac{\omega}{\omega_0}$	$\dfrac{p(j\omega/\omega_0)}{j}$	ERROR (per cent)
0·5	−0·5	0
0·6	−0·42	5·00
0·7	−0·322	7·33
0·8	−0·216	8·00
0·9	−0·1076	7·6
1·0	0	0
1·1	0·1055	5·5
1·2	0·208	4·00
1·3	0·308	2·67
1·4	0·405	1·25
1·5	0·5	0

TABLE 2

$\dfrac{\omega}{\omega_0}$	$\dfrac{p(j\omega/\omega_0)}{j}$	ERROR (per cent)
0·5	−0·5	0
0·6	−0·4098	2·45
0·7	−0·3087	2·9
0.8	−0·2046	2·3
0·9	−0·1014	1·4
1·0	0	0
1·1	0·09931	0·69
1·2	0·1973	1·35
1·3	0·2952	1·6
1·4	0·395	1·25
1·5	0·5	0

from linearity for transformations of equations (3) and (4) is plotted in Fig. 1.

SYNTHESIS OF A.C. INTEGRATOR

The transfer function for an a.c. integrator using the transformation of equation (4) is:

$$T = \frac{1}{0{\cdot}5162}\,\frac{(s^2+4)(s^2-0{\cdot}1836)}{s(s^2+1)(s^2+4{\cdot}45)} \quad \ldots\ldots (5)$$

This transfer function is synthesized by the following two procedures:

(1) The network configurataion using a current negative impedance convertor as suggested by Yaganisawa[2] is shown in Fig. 2.
The networks for the four admittances for the realization of the transfer function of equation (5) are shown in Fig. 3.

(2) The network configuration using three thermionic valves as suggested by Bobrow and Hakmi[3] is shown in Fig. 4.

The networks for the four impedances for the realization of the transfer function of equation (5) are shown in Fig. 5.

A.C. Compensating Networks

TRANSFORMATION

The transformation suggested for a.c. compensating networks is:

$$p = \frac{0{\cdot}113\,(s^2+1)(s^2+14{\cdot}21)}{s(s^2+4)} \quad \ldots\ldots\ldots (6)$$

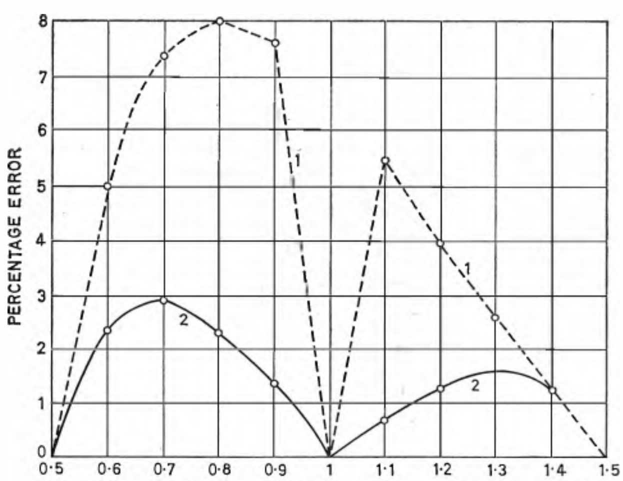
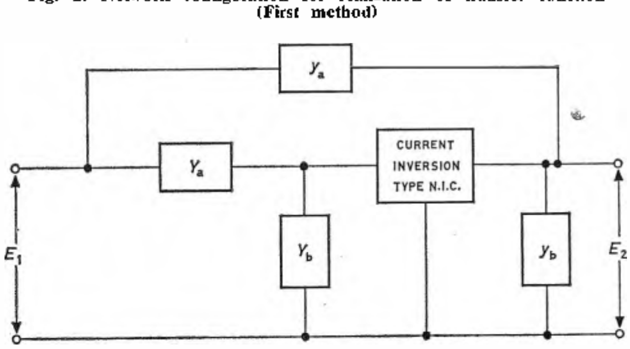


Fig. 1. Percentage error from linearity of curve (1) equation (3); curve (2) equation (4)

Fig. 2. Network configuration for realization of transfer function (First method)



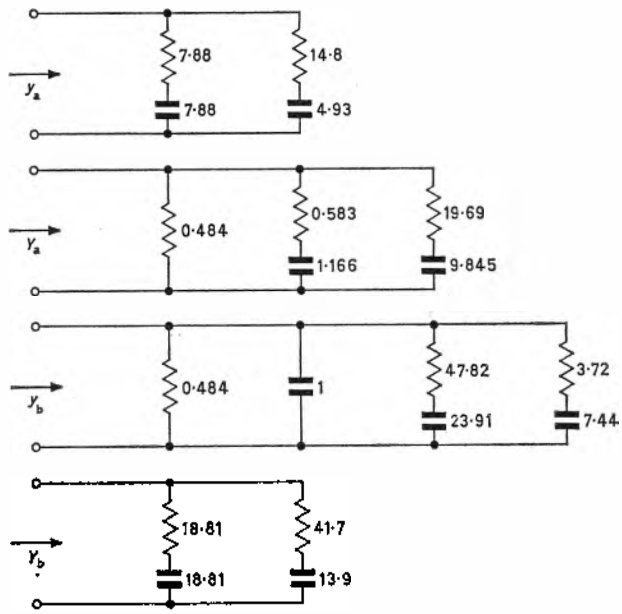


Fig. 3. Admittance networks for a.c. integrator
(Values in mhos and Farads)

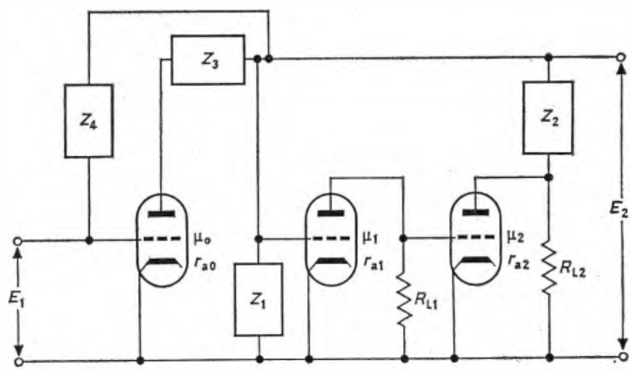


Fig. 4. Network configuration for realization of transfer function (Second method)

Fig. 5. Impedance networks for a.c. integrator

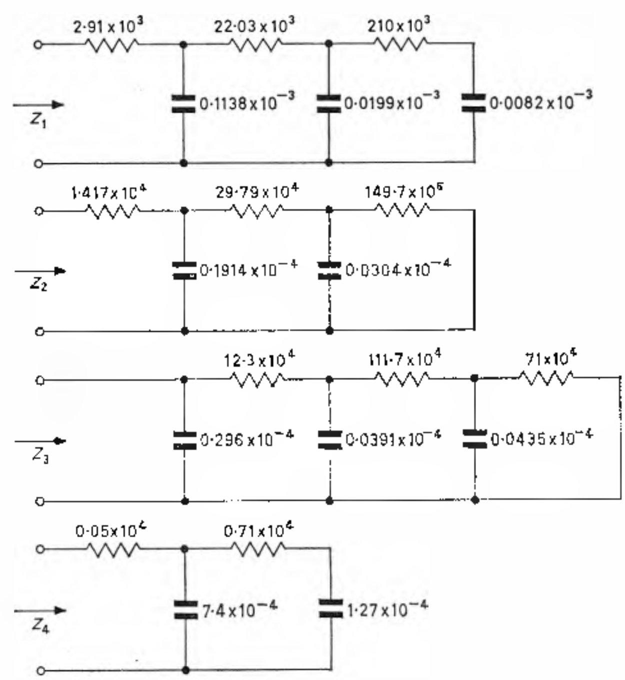


TABLE 3

$\frac{\omega}{\omega_0}$	$\frac{p(j\omega/\omega_0)}{j}$	ERROR (per cent)
0·9	−0·1002	0·2
0·93	−0·0699	0·14
0·95	−0·04983	0·34
0·97	−0·02986	0·47
1·00	0	0
1·03	0·02989	0·37
1·05	0·0499	0·2
1·07	0·07004	0·06
1·1	0·1005	0·5

p and s are normalized frequency variables.

The maximum error in linearity for this transformation is found to be $\frac{1}{2}$ per cent in the frequency range from $0{\cdot}9\,\omega_0$ to $1{\cdot}1\,\omega_0$. The values of $p\,(j(\omega/\omega_0))/j$ at different points in this frequency range are given in Table 3.

Synthesis of A.C. Lag-Lead Networks

The transfer function for a typical lag-lead is given in equation (7).

$$T_{dc} = \frac{1+4s}{1+40s} \times \frac{1+30s}{1+3s} \quad \ldots\ldots\ldots (7)$$

Using the transformation of equation (6) the transfer function of an a.c. lag-lead network would be:

$$T_{ac} = \frac{[s(s^2+4)+0{\cdot}452\,(s^2+1)\,(s^2+14{\cdot}21)] \times [s(s^2+4)+3{\cdot}39\,(s^2+1)\,(s^2+14{\cdot}21)]}{[s(s^2+4)+4{\cdot}52\,(s^2+1)\,(s^2+14{\cdot}21)] \times [s(s^2+4)+0{\cdot}339\,(s^2+1)\,(s^2+14{\cdot}21)]} \quad \ldots\ldots (8)$$

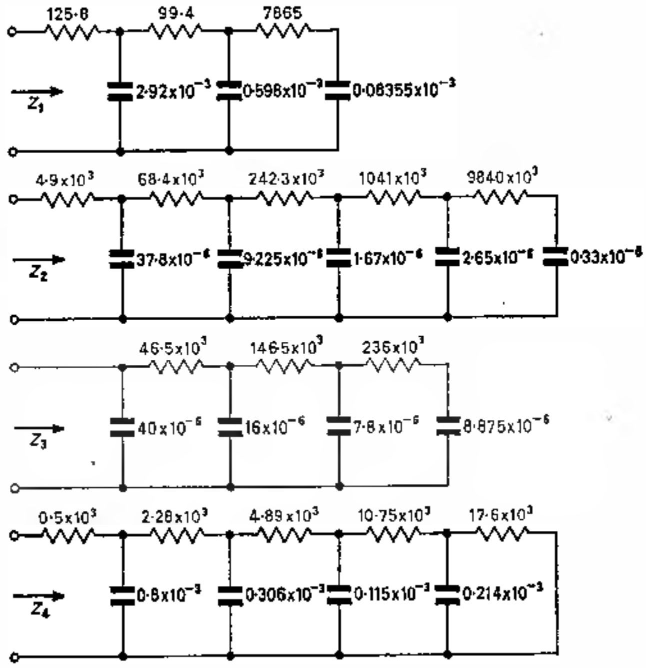


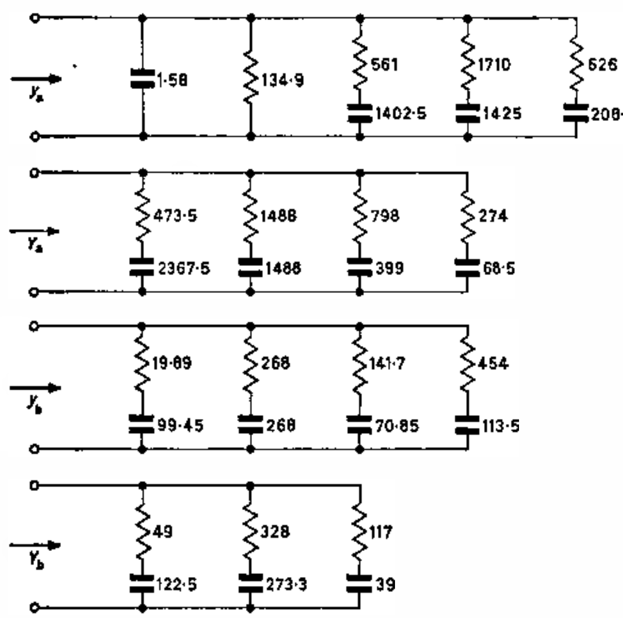
Fig. 7. Admittance networks for a.c. lag-lead network

The transfer function of equation (8) is synthesized by the two methods given previously. The *RC* networks for the four admittances of Fig. 2 are shown in Fig. 6 and the *RC* networks for the four impedances of Fig. 4 in Fig. 7.

Conclusion

The improvement in linearity achieved from the modified transformation can be noted from Fig. 1. The maximum error within the desired frequency range for an a.c. integrator has been reduced from 8 per cent to about 3 per cent. This is achieved at the cost of extra passive elements required for the a.c. integrator circuits. It is still possible to find transformations giving even lower errors in linearity. But it is expected that such transformations would be more complex and would require more passive elements in the integrating circuits.

Fig. 6. Admittance networks for a.c. lag-lead network



The error due to the transformation of equation (6) is only $\frac{1}{2}$ per cent, while that of the original transformation of equation (2) is as much as 5·5 per cent within the frequency range from $0{\cdot}9\,\omega_0$ to $1{\cdot}1\,\omega_0$.

Acknowledgment

The authors acknowledge the facilities offered by Professor L. B. Shah, Dean and Dr. S. M. Sen, Vice-Dean and Head of the Electrical Engineering Department, Faculty of Technology and Engineering, M. S. University of Baroda for carrying out the present work.

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A Graphical Method for Designing Wide Range Resistance Transducer Bridges

By Hubert S. Scar* M.D.

A graphical method is given for designing Wheatstone bridges that correct most of the non-linearity of resistance transducers like thermistors when used over wide ranges. The method indicates the amount of residual error, its sign, and where in the range of the transducer the error will occur. It consists of matching the transducer characteristic curve with one of a family of bridge characteristic curves. The application to the design of wide range thermistor thermometers with linear output is given.

(Voir page 553 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 560)

A REVIEW of analytical methods for linearizing the output of non-linear resistance transducers with Wheatstone bridges is given by Scott[1]. A graphical method is presented here for use over a wider range. This is possible by working with the actual transducer response rather than working with an expression approximating it. The curve representing this response is superimposed on a family of curves representing the response required for linear output for various bridge configurations. The curve of this family most closely approximating the response obtainable from the transducer represents the desired bridge configuration. Deviations from exact match not only show where, in the range, errors of indication will occur, but also will show the sign and amount of error.

The family of bridge curves is derived as follows:

The half symmetrical bridge labelled as shown in Fig. 1 is used. The simultaneous mesh equations are solved for the transducer resistance R rather than for I_3 as is usually done. A bridge excitation source is assumed whose internal resistance can be neglected.

$$R_{th} = C\,[E - I_3\,(A+2M)]/E + I_3\,(A+2M+2C) \quad \ldots\ldots \quad (1)$$

Equation (1) is plotted with I_3 as the independent variable since each value of current may be assigned to a specific value of the physical variable being measured by the transducer. For instance, if a linear output is desired in a thermistor thermometer, the temperature scale may be evenly spread along a linear I_3 axis. In order to superimpose curves, all quantities in the bridge equation and all transducer resistances must be normalized by dividing by the transducer resistance at bridge balance (or the value of C in Fig. 1). Normalized values are indicated in the equations below by lower case letters:

$$r = \frac{e - I_3\,(a + 2m)}{e + I_3\,(a + 2m + 2)} \quad \ldots\ldots\ldots\ldots \quad (2)$$

If $(a + 2m) \equiv D$, then:

$$r = \frac{1 - (I_3/e)\,D}{1 + (I_3/e)\,(D + 2)} \quad \ldots\ldots\ldots\ldots \quad (3)$$

Fig. 2 shows a family of curves of equation (3). Thermistors will be used as an example of a non-linear transducer whose output it is desired to linearize with a Wheatstone bridge. Fig. 3 shows typical curves relating the temperature and normalized resistance changes obtainable, from three types of thermistor material (obtainable from Fenwal Electronics Co.). The curvature even though plotted on semi-logarithmic scales, indicates the well known deviation of thermistor resistance from a true exponential over a wide temperature range. The curve for type B material approximated by 3 straight lines is shown superimposed on a particular bridge curve in Fig. 4. The maximum separation is seen to occur at a temperature of 20°F where the bridge output will indicate a temperature about 2° too low and at 70°F where the bridge will indicate a temperature about 2° too high.

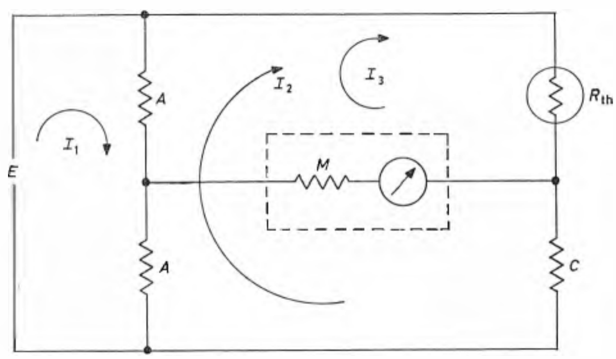

Fig. 1. Half-symmetrical bridge described in equation (1)

Details of this method will be illustrated by the following example:

A Fenwal Electronics Co. GB32J1 bead thermistor made of type B material (Fig. 3) was chosen for the sensor of a linear thermometer for the range 0 to 100°F.

A family of bridge curves was plotted on semi-logarithmic paper according to equation (3) resulting in Fig. 2. The thermistor curve was then plotted on a transparent overlay. The resistance ratios available from the manufacturer's literature were normalized to 0°F as is equation (3), and were laid out on the ordinate to the same scale as was used for the bridge curves. The abcissa represents temperature and there is a free choice of its scale to the I_3/e scale of the bridge curves. Two rules that are useful in choosing this relation are first, the thermistor curve should lie near some bridge curve in the middle of the group. Secondly, the curves should intersect (i.e. the bridge indication should be correct) at a point about 20 per cent below full scale. Using the first of these rules, the curve of the bridge of $D = 0{\cdot}7$ was chosen for the first trial. According to the second rule, this bridge should read correctly at 80°. The chosen thermistor provides a normalized ratio of 0·135 at this temperature. Referring to the curve for $D = 0{\cdot}7$ at this ratio, $I_3/e = 0{\cdot}8$. Therefore for the first approximation the relation to be

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used between I_3/e and temperature is that 0·8 units on the I_3/e scale represents 80°; each 0·1 unit on the I_3/e scale represents 10°. When the thermistor curve resulting was plotted and the transparent overlay was placed over the bridge curve, the maximum errors were read from the separation of the curves as follows:

Actual temperature	*Error* (°F)
20	−1
60	+2
100	−4

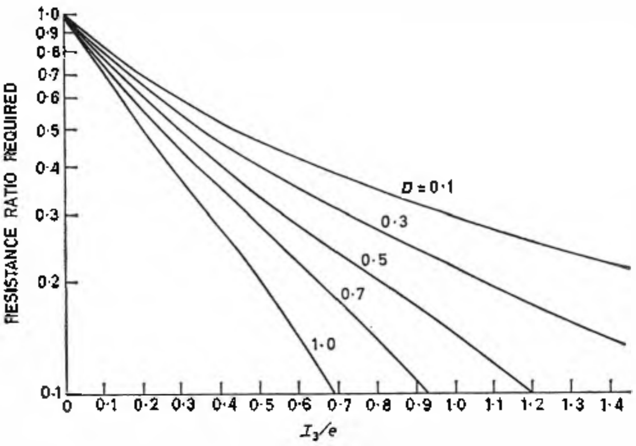

Fig. 2. Family of curves obtained from equation (3) showing the normalized resistance characteristic required by the bridge for linear output current

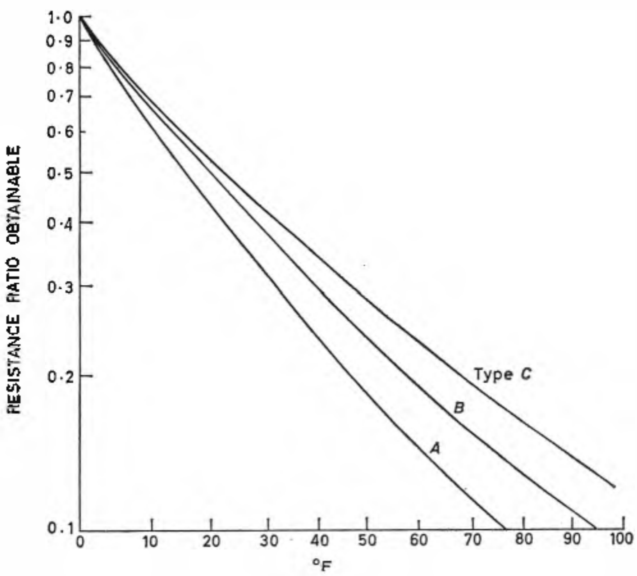

Fig. 3. Family of curves for three types of thermistor material showing the normalized resistance characteristic obtainable

In an effort to reduce these errors, comparison with neighbouring bridge curves can be made by replotting the thermistor curve with the abcissa stretched or compressed. Tedious replotting can be avoided and the same effect approximated by rotating the overlay about the point it has in common with the bridge curves, that is $X = 0$, $Y = 1$.

In this manner comparison with the curve for $D = 1$ was made, revealing that, although the errors were reduced at the low temperature end, they were increased at the high temperature end and were unacceptable. Rotating the overlay in the opposite direction revealed a promising fit with the bridge for $D = 0{\cdot}5$, and the thermistor curve was replotted. Now $I_3/e = 1{\cdot}03$ at 80°; 0·13 unit represents 10°.

Replotting was simplified in this second trial, by approximating the thermistor curve by three straight lines. The following maximum errors were indicated.

Actual Temperature	*Error* (°F)
20	−2
100	$+3\frac{1}{2}$

A slight further reduction was obtained by equalizing the positive and negative errors. Inspection reveals that this can be done by a slight compression of the temperature scale and the result is shown in Fig. 4. Maximum errors of $2\frac{1}{2}$° occurred at 20° and 80°.

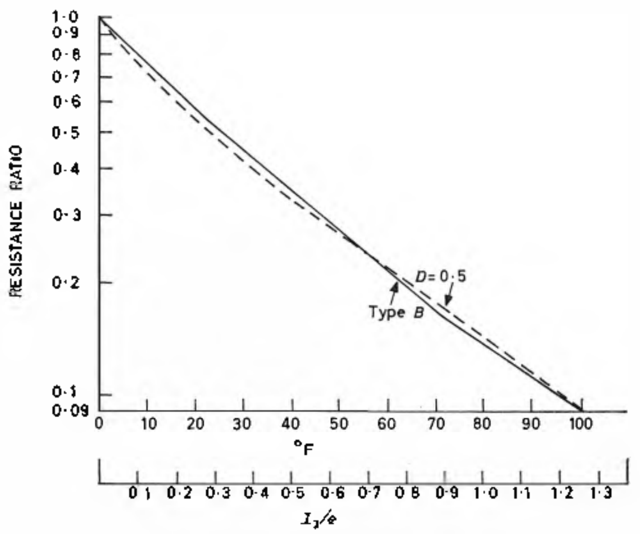

Fig. 4. Comparison of thermistor characteristic (approximated by three straight lines) and the bridge characteristic for D-0·5

The actual values of the bridge arms, meter sensitivity and resistance, and battery voltage were then calculated by substitutions in equations (2) and (3). When the bridge was constructed, maximum errors were obtained at 85°F and 23°F and were one degree above those predicted.

It should be noted that a trade-off is possible between meter sensitivity (that is I_3 at full scale) and the supply voltage. Reduction of the required sensitivity, requires increasing the supply voltage e, which in turn is limited to a value which will not cause an unacceptable error due to self-heating of the thermistor.

This maximum allowable voltage can be derived from the expression relating applied voltage and resultant self-heating[2].

$$E = (DTR_{th})^{1/2}$$

D = dissipation constant of thermistor (W/°C)

T = temperature rise allowable due to self-heating

R_{th} = thermistor resistance at the highest temperature in the range

$R_o = C$ = thermistor resistance at bridge balance.

Dividing both sides by R_o normalizes resistance and voltage. After assuming an equal voltage drop across the adjacent bridge arm, the maximum allowable normalized bridge voltage is given by:

$$e = 2\,(DTR_{th})^{1/2}/R_o$$

There are two methods of improving the closeness of fit of the thermistor and bridge curves that were not applied in the example in the interests of clarity. First the bridge characteristic curves might be changed either by use of non-linear resistors in the bridge arms or by use of a linear resistor in series with the supply voltage causing a gradual decrease in e as the excursion from balance increases. The ends of the bridge curves then become more horizontal. The second and more feasible approach, is use of resistors in series or parallel with the thermistor. The resistance of these networks at different temperatures may then be normalized as was done for the thermistor alone. Adding a resistance in series with the bead thermistor of the example above, with a normalized value of 0·1 (actual value 1 400Ω) reduced errors to less than $1\frac{1}{2}°$.

Acknowledgment

This work was supported in part by Grant Number 679 of the Massachusetts Heart Association.

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Regulation Factor for Temperature Compensated Zener Diode References

By H. Banasiewicz*

Temperature compensated Zener references of the same temperature coefficient often differ in one or more of the three basic parameters—Zener voltage, bias current and slope resistance. It is shown how these figures may be combined to give direct indication of the regulation capability of a device against bias current changes.

(Voir page 553 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 560)

IN designing stable voltage reference sources using temperature compensated Zener units[1], the engineer is often faced with a multitude of devices of different temperature coefficient, power rating, package, long term stability, price and of course operating voltage, bias current and slope resistance. The first five parameters are relatively easy to sort out, but having done that one is then usually trying to pick a unit giving maximum regulation against change of bias current. Low slope resistance is the natural characteristic to look for, but soon it becomes obvious that operating voltage and bias current[2] affect the issue.

Consider the simple circuit of Fig. 1

where V is the Zener operating voltage

v is the Zener breakdown voltage

I is the Zener bias current

r is the Zener slope resistance (at I)

$dV/dI = r$

$\therefore \ \Delta V = r\Delta I$ for small increments .... (1)

The Regulation Factor (*R.F.*) of the circuit (taken as the ratio of fractional change in I to fractional change in V) is:

$\dfrac{\Delta I/I}{\Delta V/V}$ and substituting from equation (1)

$$R.F. = \frac{\Delta I/I}{r\Delta I/V} = V/rI \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

Equation (2) shows that for a given slope resistance r, *R.F.* is directly proportional to operating voltage V and inversely to bias current I.

For most temperature compensated Zener references, bias current cannot be altered without upsetting temperature coefficient specification.

All three parameters of equation (2) are associated with the device itself and form a sort of 'Figure of Merit' as regards regulation.

Table 1 shows some better known references with *R.F.* values added.

In the case of current drive, *R.F.* value is a direct measure of 'dilution' of fractional input changes.

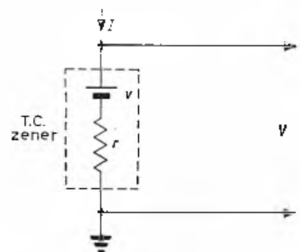

Fig. 1. Approximate equivalent circuit of a temperature compensated Zener reference

For example, using IN935, *R.F.*(min) = 60; a 6 per cent bias current change results in $\dfrac{6 \text{ per cent}}{60} = 0{\cdot}1$ per cent or better, change in operating voltage V.

In practice it is very common to drive a temperature compensated reference using a resistor and a Zener diode, sometimes also temperature compensated as in Fig. 2. The *R.F.* value is still very useful here and may be applied as before, provided changes in Zener voltage E are translated into current changes through R.

This may be done as follows:

In Fig. 2

$$I = \frac{E - V}{R} \quad \therefore dI/dE = 1/R$$

* *G. & E. Bradley Ltd.*

or $\Delta E = R\Delta I$

$$\therefore \frac{\Delta I/I}{\Delta E/E} = \frac{R\Delta I/(E-V)}{R\Delta I/E} = \frac{E}{E-V} \quad \ldots\ldots\ldots \quad (3)$$

Thus percentage ΔI = percentage $\Delta E \; \dfrac{E}{E-V}$

Suppose $E = 2\text{V}$ as is common practice, and percentage $\Delta E = 6$ per cent, then percentage $\Delta I = 6\,(2V)/(2V-V) = 12$ per cent. Using IN935, change in output voltage is $\dfrac{\text{percentage } \Delta I}{R.F.(\text{min})} \leqslant 12/60 \leqslant 0{\cdot}2$ per cent.

Effect of Load

If a load resistor R_L is added in Fig. 2 (current I being increased to maintain correct bias of temperature compensated Zener) it can be taken into account by using the effective (Thevenin) value E', instead of E, where:

$$E' = E\,\frac{R_L}{R + R_L}$$

since percentage $\Delta V = \dfrac{\text{percentage } \Delta I}{R.F.}$ by definition,

$$= \frac{\text{percentage } \Delta E}{R.F.} \cdot \frac{E'}{E' - V}$$

$$= \frac{\text{percentage } \Delta E}{R.F.} \cdot \frac{E}{E - V\,((R/R_L)+1)} \quad \ldots\ldots\ldots \quad (4)$$

Regulation Factor at Different Bias Currents

Occasionally it may be preferable to operate a temperature compensated Zener reference using bias current different from the value specified for the particular type. For example, when further improvement in temperature coefficient[3] is required.

Obviously, V will change only by a small amount and the effect may be ignored. The slope resistance, however, is heavily affected by the bias current I. Examination of the published data[4] shows that for IN821 series, the slope resistance at 3mA and 20mA is 20Ω and 3·2Ω respectively. At 7·5mA it is 7·5Ω. However, the product $r \times I$ is 0·060, 0·064 and 0·056 respectively. For most purposes it is a constant. Similar results are obtained for other types and therefore in practice it is reasonable to regard Regulation Factor as a constant, independent of bias current in a particular device.

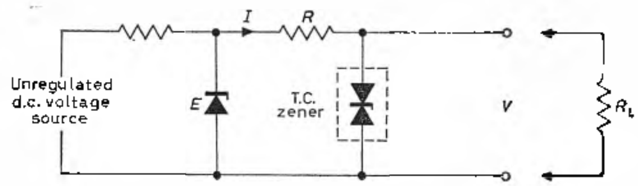

Fig. 2. Temperature compensated Zener reference driven via resistor and Zener diode

TABLE 2

	r (Ω)	r d.c. (Ω)
KS67	27	34
ZS11	60	83
IN430	10·8	11·5
IN3498	13·1	13·4

Thermally Induced Resistance

The foregoing comments were limited to temperature compensated Zener references. These devices, almost by definition could be expected to be just as good against self heating as against ambient temperature.

To test the validity of this assumption a few types were subjected to a test of slope resistance, first at 50c/s and then at d.c. Table 2 shows the results. Clearly the thermally induced[5] component is present to a varying degree. Apparently, not all of the Zener junction temperature change is passed to the forward junctions and vice versa. As might be expected, an integrated unit such as a Transitron IN3498 is virtually free from this effect.

Conclusions

The Regulation Factor for temperature compensated Zener references is shown to be a useful tool for selection purposes as well as circuit performance. Although there is little to choose between many of the devices, the very knowledge of this is helpful to the circuit designer.

R.F. figures in Table 1 are derived from the manufacturers data where the slope resistance is the a.c. value. In the absence of information as regards thermally induced resistance, it may be wise to check for this effect experimentally before final choice is made.

TABLE 1

TYPE		V (nom.) (VOLTS)	I (mA)	r (max.) (Ω)	$R.F.(\text{min.}) = \dfrac{V(\text{nom.})}{I \times r(\text{max.})}$
IN	429	6·2	7.5	20	41
	821	6·2	7·5	15	55
	821A	6·2	7·5	10	83
	3498	6·2	7·5	15	55
	3501	6·35	7.5	12	71
	430	8·4	10·0	15	56
	935	9·0	7·5	20	60
	941	11·7	7·5	30	52
	2620	9·3	10·0	15	62
	4565	6·4	0·5	200	64
	4570	6·4	1·0	100	64
	4775	8·5	0·5	200	85
	4780	8·5	1·0	100	85
	4765	9·1	0·5	350	52
	4770	9·1	1·0	200	45
SV	8227	12·6	7·5	30	56
SV	3171	7·0	10·0	10	70
KS	67	9·0	5·0	35	52
ZS	11	11·0	5·0	120	18
BZY	78	5·35	11·5	20	23
FC	107	6·65	0·1	750	89
GREA	7	9·0	4·0	40	56

Acknowledgment

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The Future of Graphic Communications

By T. M. C. Lance

In this article the methods of communicating graphic information are reviewed and some of the future uses of these techniques are discussed.

(Voir page 553 pour le résumé en français; Zusammenfassung in deutscher Sprache auf Seite 560)

THE initial object of this article was to collect data on the techniques and devices used for the electronic transmission of graphic information over various communication circuits.

It quickly became evident when studying the possible extension of these processes in the course of time that enormous possibilities are now within our grasp. The author was intrigued by studying the sudden upsurge in the scientific press of such forward-looking papers as "Telecommunications in 1984," by Dr. Pierce of the Bell Telephone Laboratories.

1984 was the year chosen by a playwright to over-dramatize the anxiety of those who fear the future. However far the business of telecommunications may have progressed at that time, one need have no fears that it will not be primarily based on present day thinking. After all 1984 is less than twenty years ahead, and putting aside the political or social attributes of 'big brother,' it is quite a thought that the electronics prophesied are now passing out of the laboratory stages.

Another intriguing thought brought out in these studies is that there must be an underlying law of civilization, which makes the standard of living for a community proportional to a power law of the communication facilities. This advantage is available to any foresighted nation with the courage and technical skill to grasp the opportunities —the basic ingredient is "know-how."

Firstly, the world is hungry for education at all levels, from the desire for literacy in the emerging countries to the emphasis on higher education in the more developed countries. Telecommunications, and in particular television will play a vital role in filling these needs, and new techniques in teaching will have to be studied hand in hand with the development of suitable equipment.

In this country, the present day provision of programmes to schools by both the Broadcasting Authorities is a pointer to the future and the 'University of the Air' is not so much a pipe-dream as critical politicians are apt to assume.

The major technical problem is that with the four entertainment programmes anticipated by the public, at least two of which will carry colour, there is no availability of channels for broadcasting television for other purposes. A temporary solution and one suited to countries less densely populated than the United Kingdom, would be to transmit educational programmes outside normal hours and anticipate the extensive use of video recorders at the receiving terminals. Such recorders are now appearing on the domestic market and are available at approximately one twentieth the cost of the professional magnetic machines and are sufficiently simplified to be operated in conjunction with normal domestic television receivers to record 'off-air' programmes.

More realistically for this country, however, the multi-channel cable networks now being operated by the relay companies will have to be expanded to cover more completely the areas of concentration of population, with microwave spurs to the lesser populated areas. Small pilot schemes for cable distribution of video programmes exclusively to schools are already being set up in the Home counties by educational authorities, who are even at this early stage finding that the time allocated by the Broadcasting Authorities is insufficient to meet their requirements.

This country leads in the design of cable equipment and has accrued considerable experience of distributing television over large networks. Granted that these are mostly located in places where direct reception has proved difficult for their subscribers, but the satisfactory performance encourages belief that we have already the beginning of a National Electronic Grid. With the existing multi-channel wired television relay systems, up to seven broadcast television programmes, pay television, and three sound programmes can be distributed simultaneously to large numbers of domestic subscribers. It is not too fantastic to look ahead to the time when such a nation wide grid is a *fait accompli* and will provide not only the services mentioned above but a daily news-sheet and video-telephone.

Eventually, with the national grid in operation it will be possible to reallocate the radio frequency spectrum which is now occupied to a great extent by entertainment and make it available for essential services, the need for which will surely expand at the same time.

After basic education for schools and technical teaching, the medical field is the most obvious application of closed-circuit television, because visual aids have such fundamental impact in the instruction of surgery and diagnosis. Students can now derive the benefit of watching surgery performed by specialists in their own or distant hospitals, and although the same instruction can be given by films prepared for this purpose, experience has shown that the actuality of the instantaneous demonstration is initially important. Television channels are already being used during off-hour periods for the transmission of refresher courses for medical practitioners, but this service should for ethical reasons be transmitted on closed-circuit. It has been proposed to use the security of Pay-Television circuits to overcome this difficulty.

It must also be remembered that the cinema industry has always been keenly observant of the developments in television as they have been interested in the possibility of distributing their programmes by electronic means to groups of cinemas. It would be advantageous to show a film programme simultaneously to all the houses in one of four groups of cinemas, but so far this application of closed-circuit television has been retarded by the inadequacy of the projection equipment to produce large screen pictures in colour with sufficient brightness and definition. Surveys have been made periodically by the industry of the problems of distribution, and it is quite clear that cable circuits will have to be used in the major areas of population. This is because the erection of tall buildings would prohibit the employment of microwaves as originally planned, and the requirement of absolute security from any form of interference. The problem of the projection within the cinemas is gradually being solved by the developments associated with the Eidophor television projector. At the present time this produces a picture which is adequate for sports events, but with the eventual solution of colour and distribution a complete recasting of the cinema industry can be visualized. Intimate cinemas in community centres will be built where the attraction of group viewing of programmes and particularly of sporting events and adult education will add a new form of electronic entertainment.

All the applications of television referred to so far are examples where bandwidth compression cannot be employed, since the displays must contain full information, as regards definition, contrast graduation, and smooth movement, and in most cases colour will be found essential.

The introduction of long distance telephone cables has demonstrated that when high quality circuits of reliable performance are established between countries and continents, then international traffic expands rapidly, and with the traffic, trade expands. Submarine telephone cables and experimental satellite communication now provide a partial global speech and Telex service, as well as television exchanges of programmes between continents. Expansion of these circuits will lead to a demand for many additional forms of information transfers such as digital data between business houses and groups, and the facsimile reproduction of personal letters, photographs, and drawings.

The methods of picture telegraphy as employed by the newspapers for inter-office use within that industry are well known, and are completely satisfactory where the time of transmission can be measured in minutes, a restriction imposed by the bandwidth of the circuits available. In all these methods the image to be transmitted is scanned and synthesized by mechanically operated equipment.

In general, this process calls for the preparation of a photographic print of the matter to be transmitted, this print being adjusted to the dimensions of the transmitting scanner. While this is quite convenient in a newspaper office the facility is not always available to others.

Electronic scanning with either 'flying spot' cathode-ray tubes or with Videcon type camera tubes has, therefore, certain advantages for facsimile transmission in applications where the initial photographic step or the rigid dimensions of the scanned area are not convenient. Even in these cases, however, the limitation of telephone circuits will still render the transmission time to be very slow and music or carrier lines should be used, since for a given information content, time and bandwidth are interchangeable.

There is also the great advantage that extracts from books, record cards, and drawings can be scanned directly. This is in effect, a 'one frame' television system, but the same equipment can be employed to give continuous scanning with very low repetition rate for applications where a reduced definition image is acceptable. Such an application would be the transmission of electrical meter readings to a central station, and quite intelligent television pictures of boiler gauges and of flame conditions in furnaces can be transmitted.

The advantage of a television type display for these purposes as against digital analysis is that any fault or random interference during transmission only gives a distorted picture rather than erroneous information which might lead to a dangerous situation if misinterpreted.

In a similar manner the relaying of radar displays such as plan position indications, where the movements are relatively slow, can be effected by optically viewing the image on the radar cathode-ray tube by a storage Videcon camera tube with a long persistence.

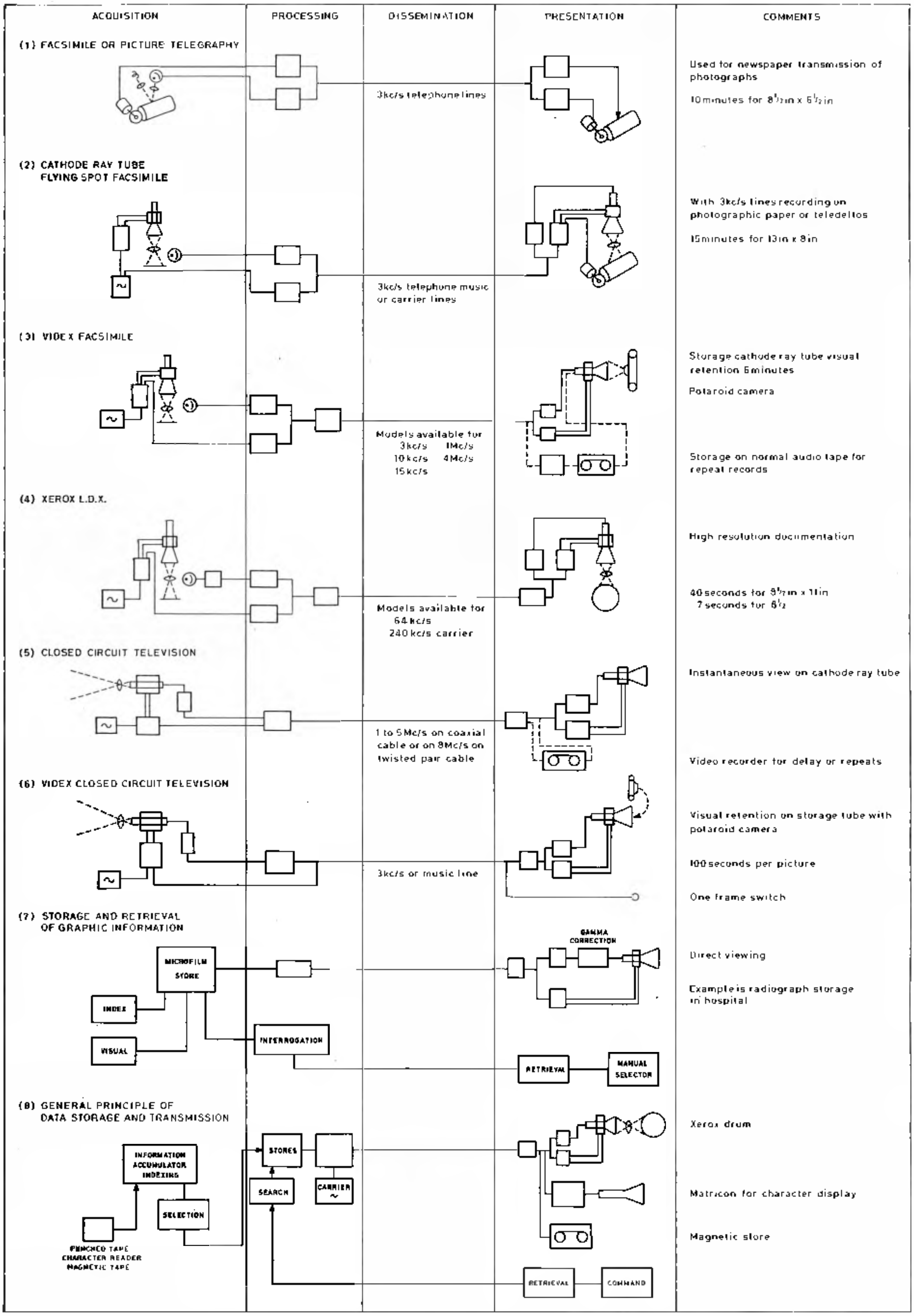
In whatever way the images are analysed the major problem of these systems lies in the receiving terminal. The incoming slow scan signals can be reconstituted and displayed on the screen of a cathode-ray tube with a persistent phosphor but the image will fade within a few minutes and be lost. There are, however, now available storage tubes where the image remains visible for a considerable time, or scan convertor tubes where the image is stored as an electron charge pattern on a screen within the tube and is read off by scanning means at a different scan repetition rate so that it can be viewed as a normal television image. Both storage tubes and scan convertors are extremely expensive with performances severely limited in definition and graduation of the final image. This has led to the development of the only practical solution, which is to form the incoming information on a more orthodox cathode-ray tube with a phosphor suitable for recording and thus render the images permanent either by rapid process photography or by the xerographic process. With either of these recording methods the image on the cathode-ray tube can be optically enlarged to produce a full size replica of the original. Since the desirable feature of such a facsimile system should be to produce a replica of the original information as near as possible to the original and as quickly as possible, these latter two processes are the most interesting until a process is invented by which the electrical signals are directly recorded on paper. Using a rapid photographic process equipment is now available which will produce black and white prints in six seconds on wide paper and if required on a continuous paper feed.

Here again there is the opportunity to utilize this process for medical purposes. In the same way that it is now possible to transmit electrocardiographs over television lines for consultation purposes, one can imagine the advantages of being able to confer with a distant specialist by displaying before him a televised radiograph. The problem to be solved here is not so much one of definition but of maintaining the low degree of contrast in fluoroscope images throughout the transmission, which would demand a signal-to-noise ratio considerably better than is achieved in present day television.

Photographic recording equipment is now manufactured which will produce positive monochrome or two colour images on 35 or 16mm film again with automatic processing and with built-in optical projection on to large screens. This is one method of presenting radar information, for example, to a large number of observers, while a smaller version of the equipment with medium size but

TABLE 1

Graphic Communication Systems



ACQUISITION	PROCESSING	DISSEMINATION	PRESENTATION	COMMENTS
(1) FACSIMILE OR PICTURE TELEGRAPHY		3kc/s telephone lines		Used for newspaper transmission of photographs 10 minutes for 8½ in x 6½ in
(2) CATHODE RAY TUBE FLYING SPOT FACSIMILE		3kc/s telephone music or carrier lines		With 3kc/s lines recording on photographic paper or teledeltos 15 minutes for 13 in x 8 in
(3) VIDEX FACSIMILE		Models available for 3kc/s 1Mc/s 10kc/s 4Mc/s 15kc/s		Storage cathode ray tube visual retention 6 minutes Polaroid camera Storage on normal audio tape for repeat records
(4) XEROX L.D.X.		Models available for 64kc/s 240kc/s carrier		High resolution documentation 40 seconds for 8½ in x 11 in 7 seconds for 8½
(5) CLOSED CIRCUIT TELEVISION		1 to 5Mc/s on coaxial cable or on 8Mc/s on twisted pair cable		Instantaneous view on cathode ray tube Video recorder for delay or repeats
(6) VIDEX CLOSED CIRCUIT TELEVISION		3kc/s or music line		Visual retention on storage tube with polaroid camera 100 seconds per picture One frame switch
(7) STORAGE AND RETRIEVAL OF GRAPHIC INFORMATION MICROFILM STORE INDEX VISUAL	INTERROGATION		GAMMA CORRECTION RETRIEVAL MANUAL SELECTOR	Direct viewing Example is radiograph storage in hospital
(8) GENERAL PRINCIPLE OF DATA STORAGE AND TRANSMISSION INFORMATION ACCUMULATOR INDEXING SELECTION PUNCHED TAPE CHARACTER READER MAGNETIC TAPE	STORES SEARCH CARRIER ~		RETRIEVAL COMMAND	Xerox drum Matricon for character display Magnetic store

very bright displays suitable for daylight viewing, have been installed on the navigational bridges of a number of ships. These displays are also used in harbour shipping control stations where the actual radar scanners may be located several miles distant and where the signals are transmitted over carrier cables. In addition to the facility of observing movements of shipping with only a few seconds delay, there is the further advantage that the films can be preserved for subsequent examination.

Relatively few of the graphic presentations for transmission require a grey scale, being mostly from line drawings, typescript, tabulations, or long hand writing such as signatures, all of which can be considered as black information on a white field. These forms of images lend themselves very conveniently to pulse modulation which enhances the signal-to-noise content of the transmission and leads to clean images on reception over long distance circuits.

A further interesting application of low speed television is in the security protection of buildings. Here storage television cameras are installed and maintained in continuous operation trained on check positions which are illuminated with either infra-red or normal lighting. Although the cameras are arranged to receive continuous optical images, the scanning is inhibited until a command signal is sent from the receiving terminal, whereupon one frame scan of the stored image is transmitted. This image is either viewed on a storage tube, but more conveniently it is recorded by photography with a Polaroid camera from a normal cathode-ray tube. The operator at the receiving position is then able to remove the photograph in a few seconds and compare it with an album of authorized persons before granting admission to a caller. This principle of 'freezing' one frame on a television signal, or of storing one frame at the transmitter until commanded by a discrete signal to radiate the information to the receiving station, is of course well established in the operation of weather, lunar and planetary satellites. This application of television for security purposes within a building or enclosure such as an aircraft hanger was referred to here because it is one application where optical communication links could be used provided that unobstructed line-of-sight paths can be established from the cameras to the control position. At present, using a newly developed semiconductor device as the light source and a similar device for the light detector, ranges up to 200ft are achieved with only a fraction of a watt drive power. The possibility of modulating laser light sources and coupling these with optical waveguides are at the present time being studied. Who can predict the future extensions and derivations of this combination?

Single optical circuits with bandwidths of 100Mc/s are now within the realms of probability for the main trunk routes for communication within each country, with the advantage that none of the radio-frequency spectrum will be occupied and that the transmissions will be wholly contained within the waveguide with complete security, and freedom from interference.

In addition to the graphic forms of information which have been reviewed above, the outputs from Teleprinters, Telex services, or computer type devices can, of course be transmitted over similar circuits. In these cases the information is handled as groups of pulses, each group being identified with a character In general the groups are derived from punched tapes which are passed through the transmitting equipment at a higher speed than that at which they are originally prepared. At the end of the circuit the signals are either decoded into an electrical typewriter or tape punch if the rate of handling is suited to mechanical devices, or is recorded and then decoded at a slower speed.

For direct recording at very high speed the characters electronically recognized by the groups of pulses received, can be formed on a high brightness, high resolution cathode-ray tube and recorded optically line by line on paper using the Xeronic process. Alternatively, a tube called the Matricon specially developed for this purpose can be used and this considerably reduces the amount of electronic equipment required for the decoding

Most of this is common knowledge to the communication engineers, but the important point to realize is the great need for a considerable expansion of wideband circuits to carry the aggregate of all this information globally.

The provision of inter-continental cables obviously is capable of handling a vast amount of traffic but it is patently clear that satellite relaying is the most feasible means available and which is capable of expansion for the future.

The first television images from America to be transmitted by satellite were received at Goonhilly Down in 1962 via Telstar I, with a power of 225 watts. The received signal was so feeble that large and expensive aerial assemblies were necessary, with extremely sensitive receivers. The availability of reception was about twenty minutes a day with the particular attitude of the satellite. This factor has since been improved by satellites placed in different orbits. But if the satellites are 'synchronized' so as to appear stationary to the observers on the earth, then a twenty four hour service can be expected. It has been predicted that three such satellites spaced equidistant on an equatorial circle would be sufficient to cover the globe with the exception of the latitudes more than 75° which includes the polar regions, in which areas the traffic would be almost nil.

Early Bird is the first example of this project and has been in commercial operation for communications between Europe and America since 28 June 1965. It is interesting to note that during the first six months the transmissions have earned £720 000 for the operators.

It has also been shown that a Syncom type satellite on a stationary orbit at an altitude of 500km with a re-radiating power of 3 to 10kW dependent on the system of modulation, would be sufficient to be received on domestic aerials and would cover a surface of the earth equivalent to Brazil or Indonesia It is quite possible that such a satellite might be the most economic means of furnishing a television coverage, and an educational service together with certain commercial communications for countries with difficult terrain and scattered populations.

Having established such a chain of satellites, then complete coverage of the globe would have been achieved. So far only the Americans and Russians are capable of placing such satellites in position because of their experience in rocketry, but the French are also planning towards this target. It is to be hoped that the British Government realizes the possibilities and the implications of this great expansion of telecommunications over the whole world, and will try to find some means of establishing effective participation in space communications

It is quite clear that the World Power which dominates global communications will enjoy privileges in the establishment and expansion of all the national systems of internal communications which together with the tolls and marketing of associated apparatus will be worth thousands of millions of pounds up to 1984.

LETTERS TO THE EDITOR

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

An R.F. Gate with no Pedestal Voltage

DEAR SIR,—In r.f. gates, coupling of the gating signal to the output results in the output signal riding on a pedestal voltage, as shown in Fig. 1. Normally this pedestal is unwanted, and an economic shunt gate in which it is variable through zero has been evolved, and is shown in Fig. 2.

Consider the gate in the 'on' condition, when both transistors are cut off

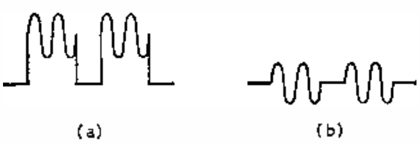


Fig. 1. (a) Output with pedestal (b) Output with no pedestal

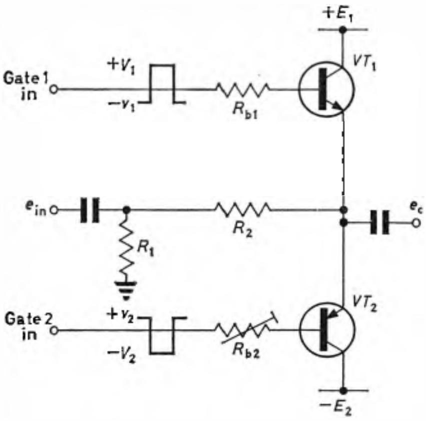


Fig. 2. Shunt gate

due to their bases residing at the potentials $-v_1$ and $+v_2$. The emitter potential is effectively at zero due to the resistive return to zero through R_1 and R_2. The impedance seen from R_2 looking into the two emitters is that of two back-biased diodes in parallel (R_b). If R_2 is chosen to be much less than this impedance, little attenuation of e_{in} results as it goes through the gate and $e_o \simeq e_{in}/[1 + R_2/R_b]$. The difference in emitter leakage currents flows through R_1 and R_2 and develops a small potential v_o at the junction of the emitters.

Application of the gating voltages $-V_1$ and $+V_2$ to the bases of the transistors turns the transistors on, in which state the gate is said to be 'off.' The impedance looking from R_2 into the emitters is now very low and large attenuation of the input signal now occurs on its passage through the gate and $e_o \simeq e_{in}/[1 + R_2/R_f]$ where R_f is the parallel impedance of the two forward-biased emitter junctions. The 'on' to 'off' attenuation ratio of the gate is $\dfrac{1 + R_2/R_f}{1 + R_2/R_b}$ and is typically 10^3 to 10^4.

The actual values of $-V_1$ and $+V_1$, and the current gains of the transistors, are taken into account when choosing the values of R_{b_1} and R_{b_2}. One of these resistors, for example R_{b_2}, is made variable in order that the emitter current of the pnp transistor, VT_2, may be made less than, equal to, or greater than that of the npn transistor, VT_1. Thus a standing voltage due to the difference in emitter currents flowing through $R_1 + R_2$, and so variable through zero, is developed at the junction of the emitters. In practice, a residual voltage v_o is established during the 'off' period of the gate to centralize the base-line with respect to the gated input signal. Capacitive coupling from the output removes the common standing potential.

A practical circuit is in use in the laboratory in which $R_1 = R_2 = 15\text{k}\Omega$, $R_{b_2} = 50\text{k}\Omega$ variable in series with $10\text{k}\Omega$ and $R_{b_1} = 22\text{k}\Omega$. The transistor VT_1 is a P346A and VT_2 is a P346A. $+E_1$ and $-E_2$ are 9V and the gating changes from $V_1 \simeq V_2 \simeq 2\text{V}$ to $v_1 \simeq v_2 \simeq 1\text{V}$. The actual values of the gating voltages V_1, V_2, v_1 and v_2 are not critical, but v_1 and v_2 must be sufficiently large to avoid clipping on the peaks of the input signal. At the same time the voltage difference between v_1 and v_2 must be less than the V_{EBO} rating of either transistor.

Yours faithfully,

D. ANDERSON,

Solid State Physics Laboratory,
Edinburgh University.

A New Type of Analogue Divider

DEAR SIR,—The arrangement of a divider using two integrators and two switching devices is shown in Fig. 1.

X and Y are two variable d.c. voltages whose quotient is to be displayed. The output of the first integrator which is fed by X linearly increases with time, its slope being directly proportional to X. If the integrator is short-circuited and reset when the output voltage reaches a value V_o say, then the output waveform is a sawtooth voltage, with a repetitive sweep time $= K_1V_o/X$ where K_1 is the integrating time-constant.

The input Y is fed to the second integrator and the output voltage of this integrator rises at a rate proportional to Y. If the second integrator, which has a time-constant K_2, is short-circuited and reset at the same time as the first, by a synchronizing switch, the output of the second integrator will have the form

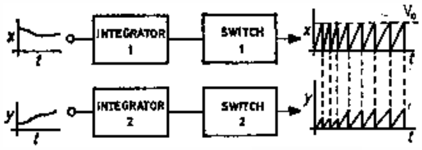


Fig. 1. A new type of analogue divider

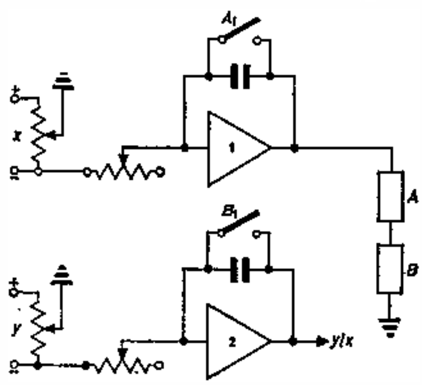


Fig. 2. Connexion diagram of the divider

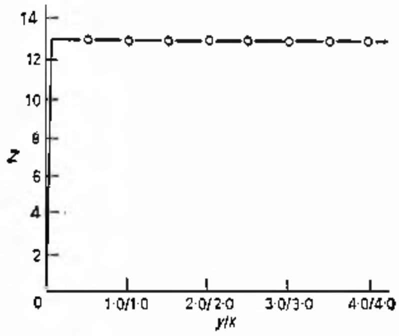


Fig. 3. Output voltage against input amplitude ratio

shown in Fig. 1 reaching a peak value $V_2 = (K_1/K_2).(Y/X).V_o$. The value of V_2 is therefore the required quotient.

A divider on the above principle has been made, and the circuit diagram is shown in Fig. 2.

X is the divisor and Y is the dividend and the quotient Z is displayed on a moving-coil voltmeter. The two integrators are of the conventional type using high gain operational amplifiers with series resistances and capacitive feedbacks. The switches used for resetting the integrators are high speed electromechanical relays and they are shown in the figure.

The characteristics of the divider are expressed in Figs. 3 and 4. The accuracy is better than 2 per cent for ranges of inputs from 0·25 to about 4V and for ratios less than 4.

The present model is a slow speed divider since it uses electromechanical switching elements, but its speed of

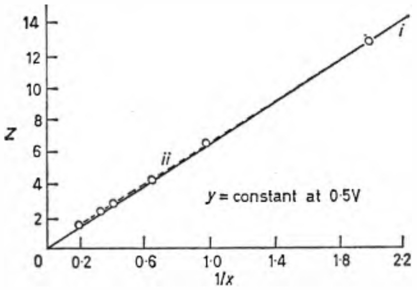


Fig. 4. 1/x against output voltage

i—Theoretical; ii—experimental

operation can be increased about a hundred times by employing transistor or valve switching circuits.

Yours faithfully,

A. Prabhakar
M. K. Garimella,
University College of Engineering,
Osmania University,
Hyderabad-7,
India.

Simple Temperature-Compensated D.C. Amplifier

Dear Sir,—Transistor pairs for d.c. amplifiers may be matched for equal β-values and base-emitter voltages, but there usually remains some difference in temperature coefficient.

A simple temperature-compensated direct coupled d.c. amplifier is shown in

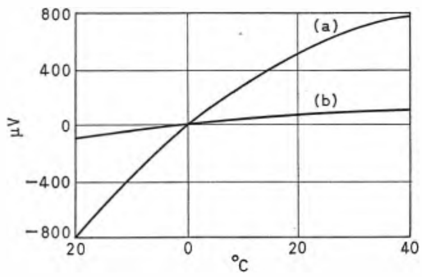


Fig. 2. Typical temperature drift (Equivalent input voltage against temperature)

(a)—without and (b)—with temperature compensation

Fig. 1. The transistors for the long-tailed pairs are selected for approximately equal β, and the temperature compensation is accomplished by a silicon diode. The correct amount of compensation is provided by adjusting a voltage divider.

A compensation circuit for the input offset voltage (due to base current) is also provided. This may be omitted if an offset voltage of 2mV can be tolerated.

With the component values shown, the voltage gain is 500. Temperature drift is easily reduced to less than 5μV/°C referred to the input (see Fig. 2).

Yours faithfully,

J. Sandstad,
University of Oslo,
Norway.

Point on a Cycle Switch

Dear Sir,—I was interested in R. A. Stevens' article on a 'Point on a Cycle Switch' in the May 1966 issue. We have developed a similar unit at Derby in which the switching and timing are achieved in much the same way as in Mr. Stevens' circuit. The unit is to be used to investigate the current transient which occurs when an alternator is short-circuited. In this case a three-phase switch is required and this raises problems in isolation between the three switches. These problems have been overcome by firing the thyristors via three mercury wetted reed switches, all being operated from the same coil. Three separate batteries are required to provide the gate currents.

The reeds were matched for pull-in ampere-turns to within ±6 per cent thus ensuring almost simultaneous closing even when operated by a coil in which the rate of rise of current is relatively slow due to the inductance. When this precaution is taken, the reeds all operate within 0·5msec which is adequate for the purpose for which the unit was designed.

Yours faithfully,

H. R. Beastall,
Derby & District College of Technology.

The author replies:

Dear Sir,—Mr. Beastall's comments are very interesting.

The only comment I have to make is that we had thought along the same lines for the three-phase switch, and had come to the conclusion that some form of electromagnetic switch was inevitable. I wonder if Mr. Beastall could provide more details about the switch type and the operating coil.

The only other possibility would seem to be to convert the output of the switching unit to a train of pulses and feed each thyristor via a transformer. This has the disadvantage pointed out in the article that one of the thyristors may not fire on the first pulse.

Yours faithfully,

R. A. Stevens,
Welsh College of Advanced Technology.

PUBLICATIONS RECEIVED

DESIGNERS' GUIDE TO ICI SILICONE RUBBERS is a new booklet produced by the Nobel Division of Imperial Chemical Industries Ltd. It describes the wide range of applications for ICI silicone rubbers throughout industry; from a designer's point of view, the booklet also gives a comprehensive analysis of the technical properties of these rubbers.

Copies of this booklet are available from Imperial Chemical Industries Ltd, Imperial Chemical House, Millbank, London, S.W.1.

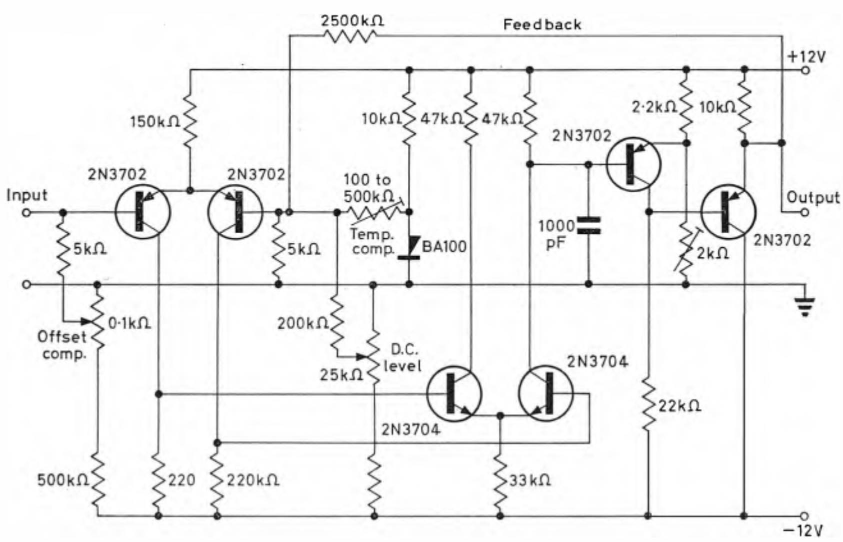
APPLICATION REPORTS 1, 2, 3 and 4 deal respectively with 'Optical Spectrometer Measurements with the Phase Sensitive Detector', 'Measurement of Hall Effect Using a Phase Sensitive Detector', 'Measurement of the Edge Shift in Gallium Arsenide Using a Phase Sensitive Detector' and 'The A.C. Operation of Thermistor Bridges'.

Copies of these reports are available, free of charge, to scientists, engineers and teachers interested in this type of measurement. They are obtainable from Brookdeal Electronics Ltd. Myron Place, Lewisham, London, S.E.13.

GLASS-TO-METAL SEALS MANUAL gives details of a range of compression seals, diode housings, transistor headers, B7 valve headers, miniature relay bases, etc.

Copies of the manual are available from Wesley Coe (Cambridge) Ltd, Scotland Road, Cambridge.

Fig. 1. Temperature compensated amplifier



NEW EQUIPMENT

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

(Voir page 547 pour la traduction en français; Deutsche Übersetzung Seite 554)

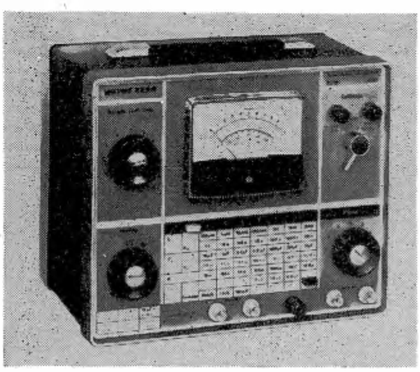
AUTOBALANCE COMPONENT BRIDGE

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Illustrated below)

A new component bridge by Wayne Kerr, the B421 'Compact,' gives direct meter readings of *L, C, R*, series or shunt losses, leakage current of electrolytics, and tolerance as the percentage difference between any two components. The Autobalance circuit gives extreme simplicity of operation: no manual balancing is required and even changing values can be observed since the bridge is continuously balanced by an electronic nulling process.

Source and detector (1000c/s) are built-in and the solid-state circuits can be powered from normal a.c. supplies or batteries. On resistance and capacitance measurements a switched backing-off control permits a measurement accuracy of 0·25 per cent to be achieved on all seven ranges. Coverage is 1μH to 100H, 0·01pF to 10μF and 10mΩ to 100MΩ.



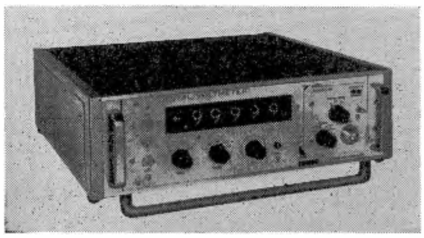
EE 96 751 *for further details*

DIGITAL VOLTMETER

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustrated above right)

With a scale of 99999 and an accuracy under optimum conditions of 0·001 per cent f.s.d. ±0·0025 per cent of reading, the DM2023 potentiometric digital voltmeter has been designed to meet those more exacting applications where accuracy, stability and a high degree of input/output isolation may be required. In common with all other Dynamco instruments, the DM2023 has a guaranteed performance over the full range of



10° to 40°C for a period of 1 year without resort to any external calibration equipment.

By carefully shielding and isolating the circuits from power earth it has been possible to achieve common mode rejection figures of 180dB at d.c. and 110dB at 50c/s. These figures apply without the 50dB filter in circuit, i.e. at 50 readings per second, and with any print-out equipment referred to ground.

The DM2023 has a wide range of operating modes, including full accuracy maximum and minimum operation, and its versatility is further extended by the use of plug-in input modules. The two modules now available are a 10μV to 1kV manual ranging unit and a 10μV to 1kV automatic ranging unit. Additional units to give higher sensitivity, a.c. measurement and current measurement will be announced shortly.

EE 96 752 *for further details*

TAPE CONVERTOR

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Illustrated right)

A.P.T. Electronic Industries Ltd has now added a tape convertor to its range of 'APTEC' data processing equipment.

The tape code conversion set, designated AT.3101C has been designed to accept paper tape in any specified code regime and produce a corresponding output tape in any other specified code regime.

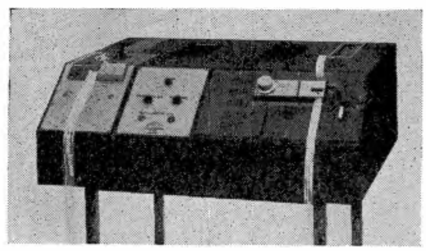
The tape reader will accept 5, 6, 7, or 8 level tapes, or edge punched cards and the punch will produce output tape or edge punched cards at a maximum conversion rate of 20 codes per second.

The machine is suitable for use where data in one code is to be converted into a different code structure. For example data received by the Telex network (in C.C.I.T.T. No. 2 code) may readily be converted into a computer input code, thus enabling Telex lines to be used for low speed data transfer.

Other uses include the conversion of tapes produced by tape controlled typewriters or business machines into tapes suitable for direct computer input. It will also be useful where computer programmes in one code structure are required to be used on computers which employ different input codes.

The operation of the tape convertor is a simple matter of placing the input tape into the tape reader and depressing the start control. This will cause conversion to proceed, the punch producing the appropriately coded output tape. A validity check system ensures that only those code combinations which are recognized by the convertor input can be translated in the punch. Any non-valid code combination causes the system to stop with the error lamp illuminated. It is therefore impossible for error codes to produce output punching of any description.

Additional controls are provided in order to enable the punching of any code combination, and some frequently used combinations, e.g. stop, or delete, may be punched by means of single switches.



EE 96 753 *for further details*

THIN FILM DIGITAL CIRCUITS

Ferranti Ltd, Ferry Road, Edinburgh, Scotland

Ferranti Ltd is now marketing a new range of low cost compatible digital logic and accessory circuits constructed from passive thin film and discrete semiconductor devices. Passive components are deposited on the surface of a glass substrate, hermetically sealed with one or more fused glass superstrates and interconnected with superimposed photoetched silver and gold plated copper patterns.

Two ranges are available, both operating at 1Mc/s and 5Mc/s. The 1200 series is the general purpose range operating from −30°C to +100°C, while the 1300 series is aimed at military applications primarily as the temperature range is extended to −55°C to +125°C.

Principal features of the circuits are the high noise immunity level of ±3·5V, the 6 to 12V operating voltage, and the small size of ½in × ⅜in × ⅜in high. Two rows of leads on 0·075 or 0·100in centres (both versions are available) with unambiguous location facilitate insertion and design of the conventional printed circuit card. The open lead configuration provides high packing density as modules can be stacked side by side and can be interconnected easily.

The following types of thin film digital circuits are immediately available:

(1) J-K and R-S Flip-Flops.

(2) Schmitt Triggers.

(3) Delay Multivibrators.

(4) Multi Input and Double Gates.

(5) Buffers.

EE 96 754 *for further details*

ANALOGUE-TO-DIGITAL CONVERTOR

The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex

(Illustrated below)

A new analogue-to-digital convertor, believed to be one of the fastest being manufactured in Europe, is now in production at the Crawley plant of The M.E.L. Equipment Co Ltd. This a.d.c., Philips type PR7830, has a maximum sampling speed of 20kc/s, and its price compares very favourably with that of American high speed instruments.

The PR7830, is of the successive approximation type and its main application is in fast data acquisition systems for industrial and research establishments. It accepts an input of plus or minus 10V and converts this to a 12-bit pure binary coded output, with an additional bit to identify polarity. It has a 400kc/s internal clock, but can also be used with an external clock of 1 to 400kc/s.

An important feature of the instrument is that it is very simple to use. It will accept any waveform for the digitize-command input and the external clock input if connected. It will also accept a phase shift between digitize-command and external clock pulses, and ignores further digitize-command pulses during digitization.

The resolution is 1 in 4096 and accuracy remains within 0·1 per cent over a simultaneous variation of temperature from 0 to 40°C and mains

supply between plus or minus 10 per cent of nominal.

There is a neon tube display of parallel output, and circuits are completely solid state. Overall dimensions are 19in wide × 5¼in high × 18in deep and the construction allows the internal circuit boards to be laid out to allow easy access to all components.

EE 96 755 *for further details*

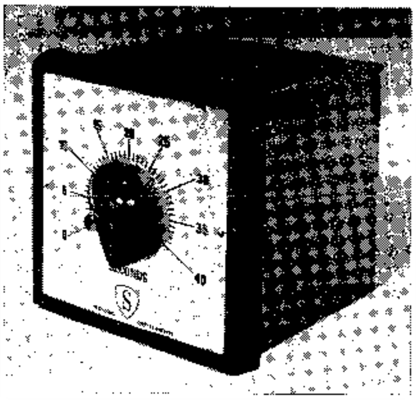
DELAY TIMER

Sterling Instruments Ltd, Sterling Works, Crewkerne, Somerset

(Illustrated below)

This model will control any timed operation and is especially applicable in the photographic, chemical and process industries.

The pointer is moved to whatever time is required for the machine to operate. The red button is then pressed and the machine will turn itself on automatically. After this time it will turn



itself off, and the red button must be depressed to restart the cycle.

Standard models are available where the delay can be specified from 0 to 10sec up to 0 to 12h. Other models up to 0 to 24h delay can be supplied to special order at extra cost.

The timer operates on a.c. mains and is run by the well known Sterling synchronous meter. Timers can be provided suitable for use over a wide range of voltages up to 440V.

EE 96 756 *for further details*

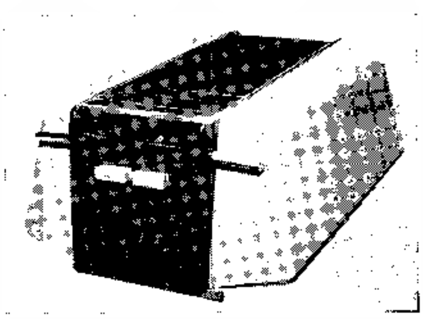
OSCILLOGRAPH RECORDER

Graphic Instruments (Research) Ltd, Trout Road, West Drayton, Middlesex

(Illustrated above right)

Graphic Instruments has added a new galvanometer oscillograph recorder to its range of recording instruments. Designated 'Graphicorder 9-90', the new instrument is a small high resolution, high precision, nine channel direct writing galvanometer oscillograph trace recorder, designed for use as a low cost high quality laboratory and field instrument. For the first time, as a result of a new approach to oscillograph optics, the benefits of the tungsten iodine lamp over the traditional high pressure mercury vapour lamp are available to users.

This recorder features dual power



supplies, either 200 to 250V a.c. or 24V d.c., a 9cm × 100ft direct print out chart, 9 channels of interchangeable galvanometer elements with frequency responses up to 2500c/s flat to ±5 per cent or sensitivities from 5µA/cm. Control of the instrument is by two levers which operate fan, lamp, chart, on/off switch, and instant 10:1 speed changes. The lamp may be switched on and off at will as the tungsten iodine lamp does not require stabilized power supplies, which latter cause delay in use and re-use after switching off. Basic chart speeds are readily set between the range 2mm/sec to 1280mm/sec. There is provision for plug in printed circuit boards for galvanometer circuit matching series and parallel resistors, multi-scale timing units and electronic chart speed control.

Graphic offers a range of galvanometers in four classes for use in the 'Graphicorder 9-90' with specifications suitable for such widely diversified applications as student instruction and supersonic flight recording.

EE 96 757 *for further details*

SELECTIVE NULL DETECTOR

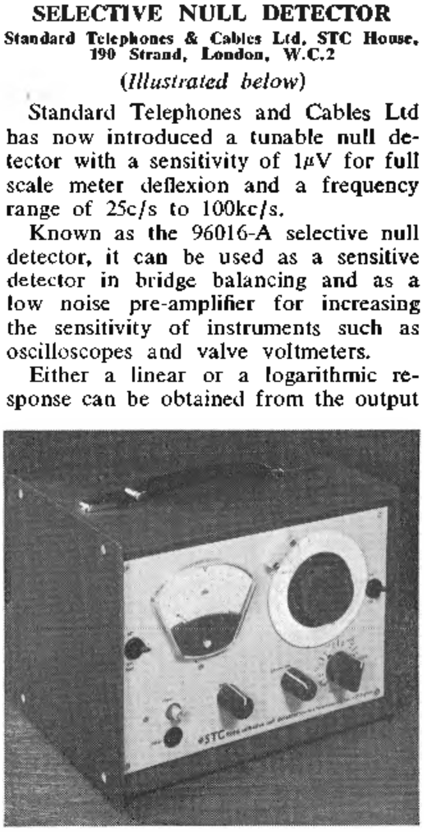
Standard Telephones & Cables Ltd, STC House, 190 Strand, London, W.C.2

(Illustrated below)

Standard Telephones and Cables Ltd has now introduced a tunable null detector with a sensitivity of 1µV for full scale meter deflexion and a frequency range of 25c/s to 100kc/s.

Known as the 96016-A selective null detector, it can be used as a sensitive detector in bridge balancing and as a low noise pre-amplifier for increasing the sensitivity of instruments such as oscilloscopes and valve voltmeters.

Either a linear or a logarithmic response can be obtained from the output

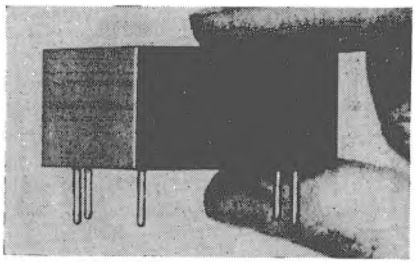


amplifier by operating a switch, thus simplifying the search for a null point by widening the scale in the logarithmic condition. Provision is made for using external filters to extend the scope of the instrument, which is particularly suitable for field use because it operates from two 6V internal dry batteries.

The instrument has an input impedance of 25kΩ to 1MΩ depending on gain control setting and on output impedance of about 300Ω in series with 5pF. Maximum output is about 1V r.m.s.

Overall dimensions are 10¾in wide × 9in high × 9¼in deep and the weight is 15lb.

EE 96 758 *for further details*



ENCAPSULATED POWER SUPPLIES

Roband Electronics Ltd, Charlwood Works, Lowfield Heath Road, Charlwood, Horley, Surrey

(Illustrated above)

These encapsulated power supplies provide a 2A (maximum) stabilized output at voltages between 6V and 24V with a maximum dissipation of 10W.

Any voltages in the range can be selected at will by means of a single external resistor and the re-entrant current overload can be similarly selected.

The small size of the unit, only 2¼in^3, means that it can be wired direct to printed circuit boards or mounted in any other way.

Apart from its obvious rugged advantages, electronically the performance is as good as conventional stabilized supplies with the added advantage that any number can be paralleled across a single unstabilized rail giving any number of discreet stabilized output voltages and currents required. This facility provides complete isolation against cross-coupling of electronic circuits.

The stabilization ratio is 1 000 : 1, the d.c. output resistance 15mΩ and the output impedance up to 100kc/s less than 0·5Ω. The maximum input voltage is 40V.

EE 96 759 *for further details*

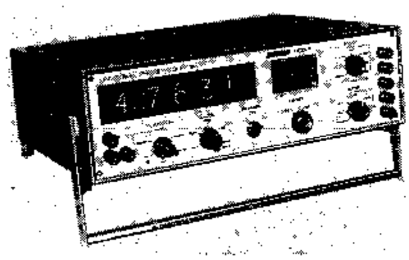
PRE-SET DIGITAL COUNTER/TIMER

Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(Illustrated above right)

This versatile pre-set counter/timer, model 805R, is intended primarily for use in industrial environments.

Digital speed and speed ratio measurements, time, period and frequency measurements to at least 1Mc/s are possible. The digital time-base is fully



variable in 1msec steps from 1msec to 100sec enabling the five-digit in-line 'storage' read-out to display measurements in any desired engineering units such as metres/minute, gallons/hour or revolutions/second.

All-solid state silicon circuits are used for reliability, with plug-in printed-circuit cards to ease maintenance. Operating ambient temperature is from −10°C to +50°C, and the counter is robustly constructed, being available for either bench or rack mounting.

The inbuilt crystal has a stability of ±1 part in 10^6 per month ±1 part in 10^6 short term. Sensitivity is 100mV r.m.s., and input impedance 10 or 100kΩ according to function. The input signal for chronometer operation may be positive or negative voltage transition, contact opening or closure (direct or via paralysis circuit) or manual push-button.

The model 805R may be used as a frequency meter, ratio meter, chronometer for time interval measurement, or delay generator, with delay variable between 1μsec and 100sec. Optional features available include 1 248 b.c.d. output, serial output for digital printer, remote display, and supplementary switching modules which can be mounted externally and operated individually by push-button.

EE 96 760 *for further details*

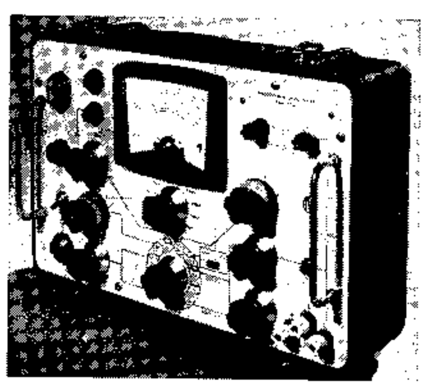
TRANSISTOR TESTER

Avo, Avocet House, Archcliffe Road, Dover, Kent

(Illustrated below)

The Avo transistor and diode tester type TT 537 provides, in one instrument, facilities for both transistor and diode testing. The tester is a compact, simple to operate, direct reading instrument providing an accurate and convenient method for the measurement of transistor and diode characteristics.

Provision is made for the rapid and accurate measurement of transistor h_{fe}



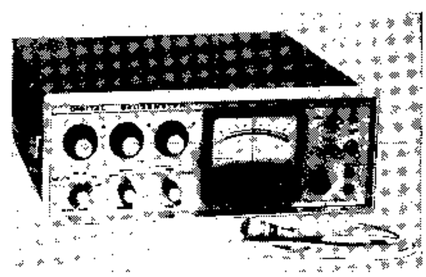
up to 1 500 at a frequency of approximately 1kc/s and the measurement of leakage current with a first indication of 1μA for both pnp and npn low medium power germanium or silicon transistors.

Both the forward and reverse characteristics of diodes can be measured, the reverse characteristics at voltages up to 1kV under current limiting conditions.

The design of the instrument enables accurate measurements to be made with the minimum of adjustments and setting-up. The layout of the panel controls is such that operation of the instrument, which centres around a function switch and a transistor diode selector switch, is largely self-explanatory. The panel markings are colour-coded and these features together with the protective devices incorporated in the instrument provide a suitable tester for use not only by engineers and technicians but also by unskilled personnel.

Collector voltage is stabilized and is continuously variable from 0 to 12V and is monitored on a panel meter, as is the collector current.

EE 96 761 *for further details*



DIGITAL GAUSSMETER

Distributed by: Livingston Laboratories Ltd, Greycaines Estate, Bushey Mill Lane, North Watford, Hertfordshire

(Illustrated above)

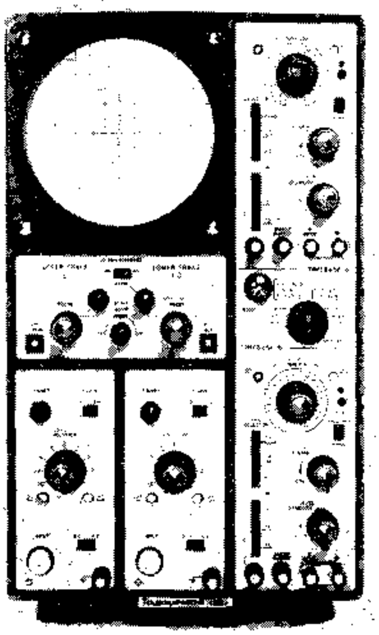
The Bell model 660 is believed to be the first commercially available solid state digital gaussmeter. Utilizing the Hall effect it features a three digit in-line read-out, providing a resolution of better than 1 part in 1 000. With full scale ranges from 100 gauss to 100 kilogauss and a basic accuracy of ±0·3 per cent ±0·05 gauss the instrument is suitable for most research, development and production applications, including the determination of magnetic fields in magnetrons, travelling wave tubes, n.m.r. systems and many other uses.

In addition to the digital model, the Bell 660 can be used as a direct reading gaussmeter over the wider range—1 gauss to 100 kilogauss at an accuracy of 1·5 per cent.

The Hall probes employ temperature compensation networks to achieve a linearity of −0·002 per cent/°C. Axial, transverse, differential and X-Y-Z probes are available for use with the gaussmeter.

A factory calibration is assigned to each probe to facilitate quick interchangeability while maintaining instrument accuracy.

EE 96 762 *for further details*



DOUBLE BEAM OSCILLOSCOPE

Telequipment Ltd. 313 Chase Road, Southgate, London, N.14

(Illustrated above)

In the D56 double beam oscilloscope two independent time-bases are provided, together with a versatile switching arrangement. This allows one time-base to be coupled to the upper beam with the other simultaneously coupled to the lower; alternatively either may be coupled to both beams. With a two-gun 5in spiral p.d.a. tube and separate brightness/focus controls for each beam, this facility enables two unrelated waveforms to be viewed at the same time, and for adjustment to either display to be made without affecting the other.

Each time-base has 22 calibrated positions plus variable control from 0·5μsec/cm to 5sec/cm. Single shot with lock-out is also provided.

An additional feature of the D56 is that time-base 'B' can be switched to act as a delay to time-base 'A'. This allows a selected portion of the signal displayed on the lower trace to be expanded and shown simultaneously on the upper trace. Superimposed on the lower beam is a 'bright-up' marker which can be moved over the entire length of the trace by the delaying sweep control. The portion of the signal covered by the marker is determined by the setting of the main sweep and is displayed across the entire length of the upper trace.

Two marker facilities are provided. One, a television line marker, uses a 0·5V pulse output from time-base 'A' which may be fed to a video monitor for line identification during video waveform investigations. Pulse width is equal to sweep length. In addition, internally generated 10Mc/s marker pips are available on the upper beam for rise time measurement.

Two identical Y amplifiers have bandwidths from d.c. to 15Mc/s (−3dB) and maximum sensitivities of 10mV/cm. They are compensated for optimum pulse response.

Accurately calibrated attenuators provide direct readings of input voltages from 100mV/cm to 50V/cm (d.c. to 15Mc/s), and 10mV/cm to 5V/cm (d.c. to 500kc/s). Variable gain control covers the range between attenuator steps.

Two modes provide steady synchronization for all waveforms. The time-base is locked automatically to any frequency from a few cycles per second to 1Mc/s at any input level. H.F. synchronization covers the range from 1Mc/s to above 15Mc/s, and a considerable reduction in jitter content has been achieved.

The X amplifier has a bandwidth of d.c. to 500kc/s (−3dB) with maximum sensitivity 100mV/cm. X expansion is continuously variable up to 10 screen diameters approximately, and the trace expands symmetrically about the screen centre. X-shift control positions any portion of the expanded trace; there is direct access to 'X-amp', and a sweep output facility.

EE 96 763 *for further details*

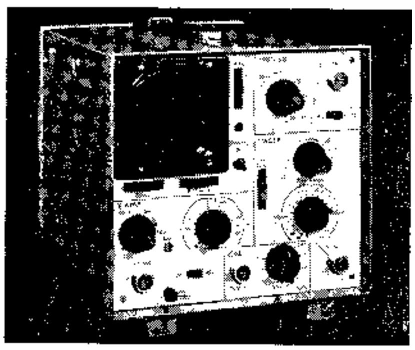
PORTABLE OSCILLOSCOPE

Marconi Instruments Ltd. St. Albans, Hertfordshire

(Illustrated below)

This new transistorized oscilloscope (type TF2203) is specially aimed at those applications demanding a portable, compact and low-cost instrument. It has a Y amplifier bandwidth of 15Mc/s with a sensitivity of 50mV/cm. A calibrator is provided, enabling each attenuator setting to be checked.

Major features of the specification include: tunnel-diode triggering giving



auto and triggered operation well beyond the Y amplifier bandwidth; optional mains or battery operation, employing a power supply system conferring complete independence of mains voltage variation over a very wide range.

Small size, low weight (15 lb) and rugged construction make this instrument suitable for production area use, 'on site' installation and maintenance applications, and for teaching, demonstrations and monitoring.

EE 96 764 *for further details*

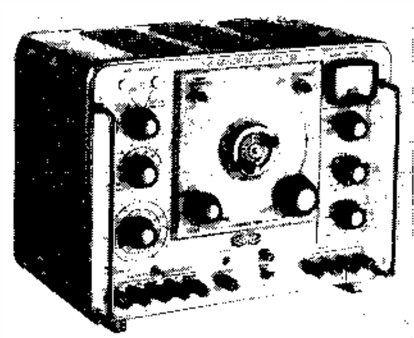
SWEEP SIGNAL GENERATOR

Airmec Ltd. High Wycombe, Buckinghamshire

(Illustrated below)

The sweep signal generator type 352 covers the frequency band 20c/s to 200kc/s in two ranges, 20c/s to 20kc/s and 200kc/s. Each of these ranges may be swept by either manual tuning or motor drive; the degree of sweep is infinitely variable. from 270° (corresponding to the whole of the range in use) down to 5°, by means of limit stops, and the sweep speed under motor drive is controllable within wide limits.

Output level, which is stabilized to within ±0·5dB over the whole frequency range, is variable from 10mV to 10V in 1dB steps by using the coarse and fine attenuators. The output may be selected as balanced or unbalanced, floating or referred to earth, from a 600Ω source impedance.



An important feature is the built-in logarithmic detector, which provides a d.c. output and a meter indication both of which are proportional to the logarithm of the amplitude of the applied frequency.

The X scan output provides a direct voltage proportional to the logarithm of the instantaneous frequency, suitable for driving the X amplifier of an oscilloscope such as the Airmec type 279; adjustable frequency marker pulses are also available at sockets on the rear panel.

EE 96 765 *for further details*

A.C. AMPLIFIERS

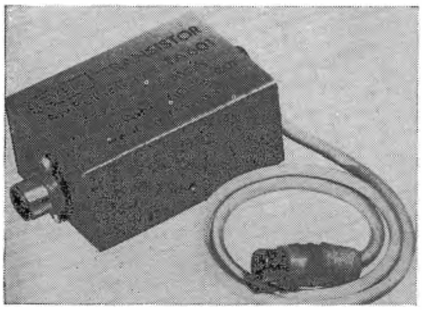
Levell Electronics Ltd. Park Road, High Barnet, Hertfordshire

(Illustrated on page 542)

These amplifiers are designed to increase the sensitivities of oscilloscopes and electronic voltmeters over a wide band of frequencies. Types TA401 and TA601 have preset gains of 40dB and 60dB whereas type TA605 has the gain switched in 10dB steps from 20dB to 60dB.

Silicon planar transistors are used with considerable negative feedback to stabilize against variations in temperature and supply voltage.

The gain of type TA605 amplifier is controlled by a switch with six positions marked 'OFF and 0dB', '20dB', '30dB', '40dB', '50dB' and '60dB'. In the first position, the input and output connexions are joined together with



the amplifier isolated and switched off. The gain in the other positions is controlled by an internal preset potentiometer that may be set for correct gain at any level.

The power supply is a self-contained long life battery, but an a.c. mains power unit is available as an alternative to fit in type TA605 amplifier.

EE 96 766 *for further details*

CABLE CONTINUITY TESTER

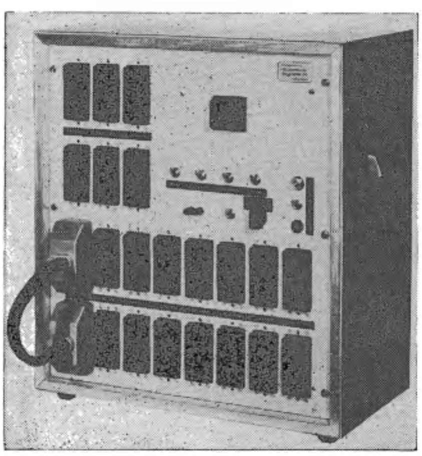
Technivision Engineers Ltd, Braywick House, Maidenhead, Berkshire

(Illustrated below)

This continuity tester is capable of checking up to 944 wires in a cableform or electrical circuit. Its primary function is to indicate any open-circuit in a cableform or similar wiring assembly immediately and automatically. With the addition of a small discriminator circuit it will detect a high resistance point, such as a dry soldered joint.

Faults are indicated on a panel-mounted electromechanical counter which will 'stop' immediately a fault is detected. The tester can also be used to scan automatically 944 circuits for checking values of inductance, capacitance and resistance in conjunction with a suitable instrument, such as a bridge or a universal type measuring instrument. This it will do at a speed of two circuits per second. Alternatively, and in order to allow an operator to record measurements, this function can be slowed to a rate of one circuit per 8 seconds, before automatically passing to the next circuit.

An interesting feature of this piece of equipment is that it can also be used



as a wiring aid. In this role, it is connected via a suitable cable to the piece of equipment being wired. The wiring sequence of the equipment is suitably programmed and arranged so that the tester does not progress to the next circuit until the previous wire is correctly connected between its own predesigned terminations. A changeover switch is provided to allow an inspector to operate and check the assembly at any time 'in situ'.

EE 96 767 *for further details*

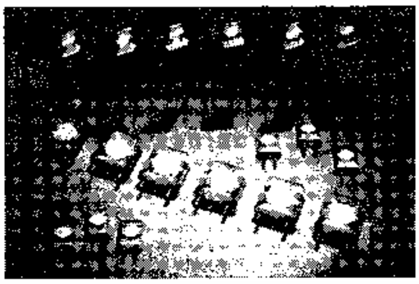
QUARTZ CRYSTALS

Standard Telephones & Cables Ltd, Components Group, Footscray, Sidcup, Kent

(Illustrated below)

In addition to its extensive range of two-pin quartz crystal units encased in metal envelopes, STC Quartz Crystal Division has developed and started quantity production of units encased in evacuated envelopes of glass.

Three styles are produced, designated types 4444, 4445 and 4446 respectively.



They meet DEF specification styles L, M and N, are are equivalent to MIL-C-3098 styles HC-27/U, 26/U and 29/U.

The evacuated glass construction offers an order of improvement in ageing characteristics plus long term reliability, and enables units to operate in ambient temperatures up to 150°C.

Closer frequency tolerance is also possible, and this feature, coupled with the low ageing characteristic, makes these units suitable for 12½kc/s channel-spacing mobile radio.

Type 4444 is available in frequencies from 1·6Mc/s to 100Mc/s; types 4445 and 4446 cover the band 5Mc/s to 100Mc/s.

EE 96 768 *for further details*

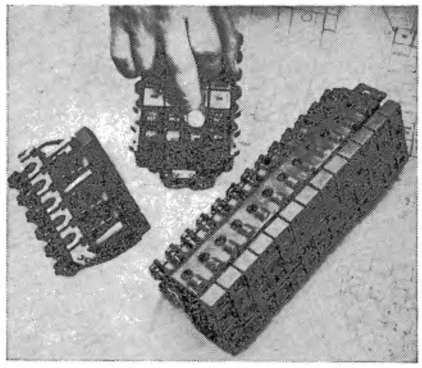
PUSH-BUTTON KEYBOARDS

Elliott-Automation Ltd, 167 Great Portland Street, London, W.1

(Illustrated above right)

A new system for producing push-button keyboards of virtually any layout and extent for switching electrical and electronic systems has been developed by Elliott-Automation Ltd.

The new keyboard is based on a simple, light alloy investment casting forming a mounting for four rows of 13 keys and microswitches. The casting can readily be cut either across or along the four rows to allow for any pattern and number of keys.



Each key assembly is simply threaded and held by pegs on to a small post glued into its socket in the main casting. A distinct 'click' feel is given to key operation by a small permanent magnet which momentarily holds the key against finger pressure. The mechanism has been successfully tested over a wide range of temperature and vibration conditions.

Casting and keys can be quickly and cheaply made up to customers' requirements. Colour, labelling and size of the plastic key tops are optional, as is the type and quality of miniature microswitches and a number of other features.

The groups of keys are designed for conventional mounting in a front panel of a console or case.

Additional features include a simple mechanical interlock to prevent more than one key in a group on one casting being pressed, and a limited amount of interface logic for which the circuit boards can be accommodated in the switch mountings.

EE 96 769 *for further details*

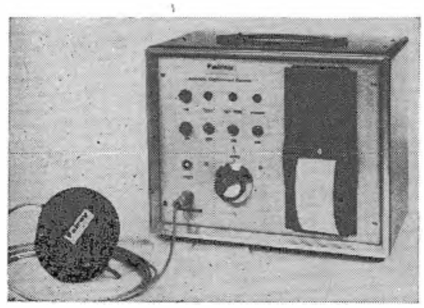
AUTOMATIC DISPLACEMENT RECORDER

Fairey Engineering Ltd, Stockport, Cheshire

(Illustrated below)

Known at D.A.T.A.R. (Digital Auto Transducer and Recorder), this equipment measures linear displacement to a resolution of 0·0001in and prints-out the results automatically against a timed programme or by remote control.

Compact and easily set up, D.A.T.A.R. employs a phase-sensitive pulse counting system to measure movement both positive and negative. The maximum print-out rate is 25 units of resolution per second, and the maximum print-out rate is four readings per second. Automatic



print-out may be supplemented at any time during a test run by manual operation.

The two basic units of the D.A.T.A.R. system are the transducer (available with resolutions of 0·0001in, 0·001in and 0·01mm and maximum displacement ranges of 0·5in, 2·0in and 12·0mm respectively) and the recorder unit which carries out the function of pulse processing, counting, timing and print-out.

Overall accuracy is guaranteed at ±2 units of resolution or ±1 per cent, whichever is the greater. Four controls are used for setting up the equipment and initiating the automatic recording of test results, which are presented digitally on a tear-off paper strip, to provide a permanent record.

The transducer is slightly larger, but similar in configuration to a dial gauge, weighs 1½ lb and operates satisfactorily over a temperature range from 0°C to 40°C. The recorder unit measures 13in by 10½in by 9in and weighs 29 lb. The whole system is transistorized for reliability and low power consumption (normal a.c. mains supply).

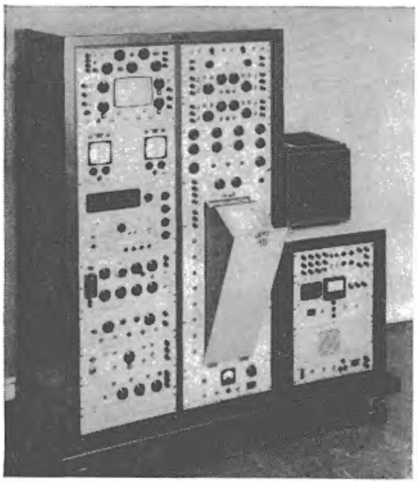
EE 96 770 *for further details*

OSCILLOGRAPH RECORDER

Southern Instruments Ltd. Camberley. Surrey

(Illustrated below)

This new low-cost oscillograph recorder for the mechanical and electrical engineering industries, the MRE.141, can be particularly useful for manufacturers of turbines, rotary and reciprocating engines of all types. It records simultaneously four channels of information, together with time, event and duration of event on 70mm or 120mm film or paper. It incorporates logic in all modes of operation, preventing initiation of the event or accidental exposure. Single shot or repetitive records can be taken and calibration signals may be recorded before, during or after the event. Triggering is performed by input voltage level, pre-set time delay, or counting cycles of an input signal. Programmes can be arranged for monitoring at discreet intervals in times ranging from one millisecond to two days. Eight functions can be recorded simultaneously in *X-Y* presentation, with additional time and event markers, and recording cameras are included for single shot, continuous feed and drum operation. An accuracy of 1 per cent can be obtained and it is possible to record from d.c. up to a frequency of 250kc/s. There is also room to include combinations of 2Mc/s f.m. and d.c. pre-amplifiers and other ancillary panels.



EE 96 771 *for further details*

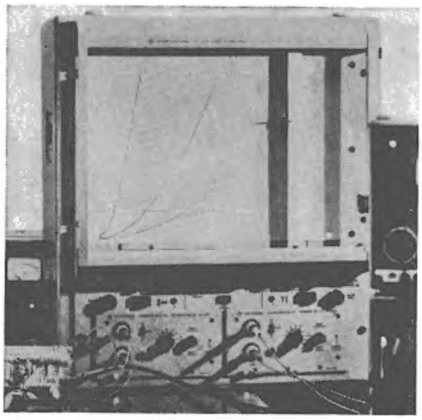
XY PLOTTER

Distributed by: Aveley Electric Ltd, South Ockendon, Essex

(Illustrated below)

A high recording speed of 1·25m/sec and a sensitivity of 0·1mV/cm to 10V/cm are characteristics of the new Rohde and Schwarz *XY* plotter type ZSK. The instrument works on the principle of the automatically balancing compensator and records on a DIN A3 are linear in both directions.

The measured values may be on different potential levels since the inputs



for *X* and *Y* voltages are ungrounded and galvanically separated from each other. Without any further additions, mathematical operations such as multiplying and squaring are possible. Rack mounting units for linear conversion of a.c. voltage or frequency into d.c. voltage increase the versatility of the plotter.

EE 96 772 *for further details*

CURRENT TRANSFORMERS

Atkins, Robertson & Whiteford Ltd, Industrial Estate, Thornliebank, Glasgow

(Illustrated above right)

Atkins, Robertson & Whiteford Ltd is producing a range of ring current transformers designed for economical metering with industrial grade ammeters to BSS 89(1954). These transformers offer a price advantage over class "C" BSS 3938(1965), the type frequently used for ammeter measurements.

Economy in design has been achieved by using standard diameters and by

reducing the rated burden to a level suitable for modern panel mounting ammeters.

The maximum ratio error throughout the operating range is less than 1 per cent of full scale and although the transformers are not intended for wattmeter measurements the phase difference does not exceed 100 minutes.

Secondary terminals are fitted as standard and a mounting bracket can be incorporated at a small extra charge.

These transformers are rated between 1VA and 15VA according to ratio.

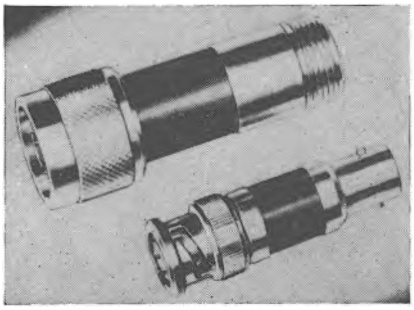
EE 96 773 *for further details*

COAXIAL ATTENUATORS

Greenpar Engineering Ltd, Station Works, Harlow, Essex

(Illustrated below)

The Electronics Division of Greenpar Engineering Ltd has introduced a new range of precision coaxial attenuators at competitive prices. They are specially designed for accuracy and stability over a wide temperature range. The standard attenuation range is 1, 2, 3, 6, 10, 12 14 and 20dB, all of which are available in a choice of three connector interfaces, series 'N', 'BNC' and 'TNC'. The bodies and the impedance matching sleeves are silver-plated brass, enclosing 'T' networks of thin film close tolerance, disks and rods, set in p.t.f.e. insulators, with gold-plated contacts: climatic seals are made of silicon rubber.



EE 96 774 *for further details*

SHORT NEWS ITEMS

A Conference on Air Traffic Control Systems Engineering and Design is to be held at the Institution of Electrical Engineers on 13 to 17 March 1967.

Sponsored by the professional group on radio navigation and radio location of the IEE Electronics Division and the radar and navigational aids group of the Institution of Electronic and Radio Engineers, the conference will cover:

Information Services

- Communications for air traffic control
- Navigational aids for air traffic control
- Primary radar systems and equipment
- Secondary radar systems and equipment

Data Handling and Displays

- Display for air traffic control
- Methods of character production and presentation
- Tracking systems and automatic tracking
- Transmission and storage of radar information
- The role of the computer in air traffic control

The Organizing Committee invites the submission of contributions for consideration for inclusion in the conference programme. Authors should, in the first place, submit synopses of approximately 250 words to the Joint Conference Secretariat by 20 August 1966.

Registration forms and other details will be available from the Joint Conference Secretariat, IEE, Savoy Place, London W.C.2,

The Electron Microscopy and Analysis Group of The Institute of Physics and The Physical Society with the Electron Microscopy Section of The Royal Microscopical Society is arranging a conference on 'Advances in Electron Microscopy' to be held at the Institution of Electrical Engineers, London, from 26 to 28 September 1966. The object of the conference is to report and discuss the latest advances in electron microscopy. The first two days will be devoted to reports and discussions of the most significant advances and the most interesting new material presented at the Sixth International Congress for Electron Microscopy to be held in Kyoto, Japan in August 1966.

Further details and application forms are now available from the Meetings Officer, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

The British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semi-Conductor Manufacturers' Association (VASCA) announce the following sterling value of their members' sales during the three months ended 31 March, 1966:

Valves and tubes	£11·7M
Semiconductor devices	8·2M
Total	£19·9M

The rapid developments in computer use at the Imperial College have led to the decision to set up a new type of organization, in the form of a Centre for Computing and Automation, which will be outside the departmental structure and will act as a focus for co-operation between departments while developing the interdisciplinary nature of the subject. The Centre, comes into being at the beginning of next session and replaces the College Computer Unit which was established in 1964 in the Electrical Engineering Department.

A permanent staff will be attached to the Centre, which already includes the IBM 7090 installation together with the 1401 and other ancillary equipment; there is also the link to the University of London Atlas machine. The main IBM 7090 installation is now working at capacity. It is staffed to operate three full shifts five days a week, and is available at weekends for those users who supply their own operating staff.

In addition to the services available to departments, the Centre will be concerned with post graduate studies. Facilities for research will be provided and a post graduate course in Computing Science will be introduced in October this year. Students attached to the Centre will all be post graduates, but the staff will provide lectures for undergraduates on computing methods and other allied subjects.

An International Congress and Exhibition on Instrumentation and Automation in the paper, rubber and plastics industries is to be held at Antwerp, Belgium on 24 to 28 October 1966.

Registration forms and other details are obtainable from the Secretariat of the P.R.P. Automation, Ingenieurshuis, Jan van Rijswijcklaan 58, Antwerp 1, Belgium.

The Electronic Engineering Association, the Scientific Instrument Manufacturers' Association and the British Electrical and Allied Manufacturers' Association have now formed a joint EEA/SIMA/BEAMA Traffic Engineering and Instrumentation Group.

The aim of this co-ordination is to eliminate unnecessary duplication of activities and to ensure unity of purpose in negotiations with Government departments and other branches of industry. SIMA will provide secretarial services for the new group, which already has a membership of 35.

Discussions about the possibilities of closer collaboration in other spheres of common interest are in progress between the three organizations.

The International Electronic Components Exhibition at Paris (Salon International des Composants Electroniques will be held on 5 to 10 April 1967 at the Parc des Expositions, Porte de Versailles.

There will be no measuring instruments section at this Exhibition in view of the International Exhibition of Measuring, Testing, Control and Automation (MESUCORA) which is to be held on 14 to 21 April 1967 at the Palais de la Dafense, Paris.

An International Conference on Electronics and Space is to be held on 10 to 15 April 1967 at the Unesco Centre Paris.

A revision of B.S.2133B, 'Ceramic dielectric capacitors Type 2 for use in telecommunication and allied electronic equipment' has now been published by the British Standards Institution. It covers fixed capacitors for radio-frequency currents not exceeding 1A or for a loading not exceeding 10VA (reactive). The standard is divided into three sections.

Section 1 defines terms, gives a classification into categories according to B.S. 2011 and gives standard values and tolerances for rated capacitance and voltage.

Section 2 establishes uniform requirements for judging the capacitors' suitability for use over stated ranges of temperature and humidity; their ability to withstand specified dynamic conditions, shock or vibration for instance, likely to be met in transit or in service; and their ability to withstand normal assembly processes. The necessary test methods are specified. Reference is required to B.S.2011.

Section 3 which is still under consideration will state standard dimensions for the capacitors.

An appendix gives an example of a suitable circuit for the voltage proof test.

The standard is identical in technical content and layout with IEC Publication 187, except where indicated to the contrary.

Copies of B.S.2133 may be obtained from the B.S.I. Sales Branch, 2 Park Street, London, W.1. Price 15s each (postage 1s. extra will be charged to non-subscribers).

Pye Telecommunications Ltd has supplied and installed telementry equipment for the Grafham Water Scheme of the Great Ouse Water Authority, Huntingdonshire.

Three Telescan equipments, situated at the Intake Pumping Station, the Dam Pumping Station and at Flitton Booster Station, send measurement information continuously in digital form on water pressures, rates of flow and levels to the Control Room over cables buried alongside the pipelines.

A monitoring system is incorporated which transmits continuously the state of all plant and provides warning if any faults occur.

The main telemetry equipment in the Control Room at the Treatment Works routes and processes all incoming information for display on the control panel and for recording. Signals can be transmitted from the Control Room to operate remote pumping plant as required.

These signals are coded and transmitted automatically when the appropriate switches on the control panel are operated. The controls include the starting and stopping of pumps, control of pump speeds and the operation of valves on the pipelines.

The M-O Valve Co. Ltd is to take over the marketing as well as the manufacture of certain professional cathode-ray tubes previously handled by Mullard Ltd. This arrangement, involving some 50 tube types, will become effective on 1 August, 1966, and M-O V. will take over the supply of all requirements at that date.

Under this arrangement, the M-O. V. Co. Ltd. will be able to offer a comprehensive range of modern instrument, radar and monitor cathode-ray tubes for professional electronic equipment.

Standard Telephones and Cables Ltd, London, is to supply a 760-mile submarine cable system linking the United States Air Force Eastern Test Range base at Patrick, Cape Kennedy, with Grand Turk Island in the Bahamas. The main contractor to the USAF for the project is Federal Electric Corporation of New Jersey; the total value of the project is £2·15M and the value to STC —for cable, repeaters, equalizers and terminal equipment—is over £1·5M.

The cable will be used to support the USAF and NASA test programmes on the Eastern Test Range and will be capable of handling administrative, record, and high speed data communications. It will form part of the global network required for manned space exploration. Intermediate shore stations will be located at Grand Bahama Island and San Salvador.

The cable system will be capable of carrying 270 simultaneous telephone conversations, or the high speed data equivalent, from Cape Kennedy to Grand Bahama Island, and 60 channels from Grand Bahama to San Salvador and Grand Turk.

Multiplexed signals travelling along the coaxial cable are amplified at intervals of 30 nautical miles between Cape Kennedy and Grand Bahama Island to provide the 270 channel system, and at 9·7 nautical mile intervals between Grand Bahama Island, San Salvador and Grand Turk to give the 60 channel system.

HUCO, an international consortium led by Hughes Aircraft Company of the U.S.A., has been selected as the lowest bidder to construct the £100M NATO Air Defence Ground Environment (NADGE) project. In addition to Hughes, the consortium consists of Compagnie Française Thomson-Houston of Paris, France; The Marconi Co. Ltd, Chelmsford, England; Selenia S.P.A., Rome, Italy; Hollandse Signaal Apparaten, Hengelo, Netherlands, and Telefunken AG, Ulm, Federal Republic of Germany.

The NADGE system, estimated to take four to five years to complete, will provide NATO with a completely integrated early warning and weapon control system, extending in depth from Norway to Turkey and consisting of a network of long-range radars to provide the continuous surveillance needed to detect the first hint of airspace penetration.

The nuclei of this electronic system are real-time general-purpose computers. The system takes advantage of high-speed computers to provide display for command and control of air defence weapons.

Advanced data-display equipment serves the functions of data gathering (detecting, tracking, height-finding, target identification and target-size-analysing); and data utilization (threat analysing, weapons assigning, and weapons controlling).

Once a target is acquired by radar, the information is electronically transmitted by data link to the Command Centre where it first appears on a display console in the form of a target 'blip'.

At the same time, the information is transmitted to a video processor—special electronic equipment which determines whether the 'blip' is an actual target, enemy jamming efforts, or simply video clutter.

The information is next transmitted to a correlator or a computer memory unit, whose job is to record and remember a particular 'blip' among other targets, remember whether it is a real target or clutter, and remember, as the 'blip' moves—whether it is the same target or a new one.

From the correlator, the information is sent back to the original console in the form of digitized 'track symbology'. This symbology is superimposed over the raw video input, providing the operator with two means of tracking the target—the raw incoming video data and the track symbology from the computer.

The entire sequence of events must transpire within thousandths of a second.

Actual identification of the target may be accomplished in several different ways, including voice identification, identification through comparison of coded electronics or by computer-compared information about the target; or the air defence Commander may call upon interceptor aircraft for on-the-spot visual identification.

The Council of the Society of Instrument Technology (S.I.T.) has set up a National Technical Committee to replace the previous sub-divisions under four main specialized committees namely Automation, Measurement Technology, Control and Systems Engineering, which in the past, have been the main focus of the formal papers of the Society and to be the single organization on behalf of the Society for all matters of a technical nature.

Already this new Committee has been considering a Symposium on 'Numerical Control of Manufacturing Processes' in conjunction with interested member bodies of U.K.A.C. They are also considering the Aerospace field.

The Guided Weapons Division of British Aircraft Corporation and Hughes Aircraft Company of the United States announce that they have signed a long-term licence and technical assistance agreement on space vehicle design, development and manufacture. The agreement covers navigation, meteorological, experimental and scientific satellites.

This new agreement is of great assistance to BAC as prime contractor, in collaboration with leading European companies, in tendering for the Thor Delta series of scientific satellites for ESRO. The experiments in this series of satellites are primarily to study the relationship between the earth and the sun during periods of maximum solar activity and to study stellar astronomy and the measurement of cosmic rays.

The BBC has extended its stereophonic sound transmission by the daily broadcasting of two or three programmes on the u.h.f. band in the Third Network.

The transmissions will be on the pilot-tone system which is already established for stereophonic services in the United States, Germany, France, Italy and Holland.

The stereo transmissions will be available at first only to listeners within the service area of the Third Network v.h.f. transmitters at Wrotham and Dover (South East England), but it is planned to extend them to the Sutton Coldfield v.h.f. transmitter (Midlands) in approximately twelve months and to the Holme Moss v.h.f. transmitter (North of England) in approximately fifteen months.

A five-nation European group, to be called the European Space Team (EST), has been formed to work on space developments under the leadership of Elliott-Automation with Compagnie Française Thomson Houston (CFTH) in France, Fokker in the Netherlands, Allmanna Svenska Elektriska AB (ASEA) in Sweden and Fabbrica Italina Apparecchi Radio (FIAR) in Italy. The new association combines the resources and capabilities of some of the most advanced and experienced space research companies within the member countries of the European Space Research Organization, producing a coherent international team capable of undertaking the design and production of complete satellite systems.

The General Electric Company of the U.S.A. has been retained as consultant to the whole team.

The first objective of EST is to bid for the design and construction of the TD-1 and TD-2 scientific satellites to be launched for ESRO by the U.S. National Aeronautics and Space Administration from California in 1969 and 1970. The launch vehicles will be Thor Deltas—hence the initials TD. The EST offers an established foundation of industrial strength and achievement in space technology in support of Europe's forthcoming space projects.

The TD-2 satellite, which is to be completed before TD-1, is to be launched during a period of high solar activity expected in October 1969. TD-1 will follow approximately six months later. Both will carry a considerable number of experiments, provided to ESRO by European scientists and concerned variously with the study of solar-terrestrial relationships, auroral phenomena, stellar astronomy and cosmic rays.

The Data Handling Division of the Plessey Automation Group has produced a remote control and telemetry system known as Dictum II. This system is intended for use by public utility authorities, pipeline installations and other industrial complexes, or in any situation where a number of outstations must be controlled and/or monitored at some distance from central office for applications which would otherwise involve the high capital cost of direct wire.

Dictum II operates by providing an encoding and decoding facility for transmitting information on a number of parameters to be measured, together with the appropriate control instructions. Transmission may be over physical circuits or any standard transmission channel such as radio or carrier.

The system will continuously monitor the dynamic operation conditions and the state of plant at a number of outstations associated with the system. The information is presented at the central control in a form suitable for visual display, print-out and information logging, or computer control.

Optimal signalling speeds, in the range 50 to 1200 bits per second, permit flexibility of application to match closely the functional operating time needs of the user's system, with maximum economy in the use of bearer channels. The use of the lower signalling speeds will, in general, permit voice communications over the same bearer channel.

The system is fail-safe to loss of power supplies and transmission bearer channels.

Cossor Electronics Ltd and Elliott-Automation Ltd have combined to produce a new range of radar equipment to improve flight safety on the world's airways.

Under the designation CE 70 (Cossor-Elliott 70), the two companies are now to offer secondary surveillance radar equipment to meet the widest range of requirements, from simple manual installations to complete air traffic systems. Even the simplest of these installations is designed so that it can subsequently be integrated with any data processing system.

Eurocontrol has placed an order with a group of companies consisting of the Marconi Co. Ltd, Compagnie Française Thomson-Houston (France) and Siemens (W. Germany) for the manufacture and installation of the secondary radar systems at Shannon, Brussels and Berkenfeld (W. Germany).

This secondary radar system is the SECAR type jointly developed by Marconi and Compagnie Française Thomson-Houston.

The installation at Brussels will cover a major part of one of the most complex and crowded airways systems in the world. Display facilities will be provided at the Brussels Area Control Centre, for the time being, but will ultimately be transferred to a new Control Centre which is planned for Maastricht in Holland.

The installation at Shannon, on the west coast of Eire, will cover one of the principal entry and exit points for North Atlantic air traffic. The displays will be located initially in the existing Oceanic Control Centre, under the control of Eurocontrol operators.

It will be recalled that Eurocontrol was set up in 1963 to co-ordinate the control of air traffic in the very densely populated European airlines. Seven countries are at present members of the organization: Belgium, Eire, France, West Germany, Holland, Luxembourg and the United Kingdom.

Some six months ago, Eurocontrol took over full responsibility for the control of all civil aircraft flying in the upper airspace, from 20 000ft, or in some areas 25 000ft, up to 40 000ft.

The majority of international flights occupy this part of the airspace except for the take-off and landing phases. Control of the upper airspace by a single authority has made it possible to devise a control structure based on the requirements of air traffic, and independent of the additional complication of national boundaries.

The Industrial Control and Electronics Division of BEAMA is studying the possibility of producing recommendations for standard interfaces to be used in coupling peripheral equipment to the central computer in industrial control processes. This work is being carried out in association with BSI's activities on data-processing interface and data-acquisition systems. The study will initially be confined to interfaces and peripheral equipment.

The need for standardization has increased rapidly during recent years due to the number of companies now marketing industrial control systems. Standardization will mean ultimately compatibility of systems and system components.

The work will be carried out by the Process Control Computers Committee of the BEAMA Industrial Electronic Equipment Section. The Committee was set up one year ago, to investigate the problems involved in applying computer centred automation systems and where possible take collective action in their solution, and secondly to condition the user environment to the acceptance of these techniques, which will make a significant contribution to the national productivity.

Already a considerable amount has been accomplished by the Committee on the contractural side and warranty and guarantees recommendations for both hardware and software have been published. In addition the Committee is constantly monitoring work on software and will ultimately publish a report on computer languages. Other aspects of the Committee's work include liaising with other trade Associations, the British Productivity Council, British Computer Society and arranging for industry representatives to attend international exhibitions to promote the British Electronics Industry.

Arrangements are now being made to hold a Symposium in the early summer of 1967. The Symposium will be aimed at presenting to potential users i.e. senior management responsible for formulating company policy—the economic benefits to be derived from the efficient use of computer systems and the organizational and management problems involved. Examples will be given of successful applications of computers to industrial process control.

NOUVELLES Réalisations

Traduction des pages 538 à 543

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

PONT DE COMPOSANTS À ÉQUILIBRAGE AUTOMATIQUE

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Illustration à la page 538)

Le nouveau pont de composants que vient de mettre au point la société Wayne Kerr Co. Ltd, le B421 "Compact," donne des lectures directes de L, C, R, de pertes en série ou de dérivation, de courant de fuite d'électrolytiques et la tolérance en tant que différence de pourcentage entre n'importe quels deux composants. Le circuit Autobalance est d'une simplicité de fonctionnement extrême, nul équilibrage manuel n'étant nécessaire, et même les changements de valeurs peuvent être observés car le point est à équilibrage continu au moyen d'un processus d'annulation électronique.

La source et le détecteur (1000 Hz) sont incorporés et les circuits constitués de corps solides peuvent être entrainés à partir de batteries ou d'alimentations normales en alternatif. Pour la mesure de capacité ou de résistance, une commande à commutation assure une précision de mesure de 0,25 % sur toutes les sept gammes. Le pont est prévu pour des valeurs de 1 μH à 100 H, 0,01 pF à 10 μF et 10 mΩ à 100 MΩ.

EE 96 751 *pour plus amples renseignements*

VOLTMÈTRE NUMÉRIQUE

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustration à la page 538)

Le voltmètre numérique potentiométrique DM2023 dont l'échelle est de 99999 et la précision dans des conditions optima de 0,001 % de déviation sur la totalité de l'échelle $\pm$0,0025 % de la lecture, a été conçu pour répondre aux utilisations exigeant un degré élevé de précision, de stabilité et d'isolement d'entrée/sortie. De même que tous les autres instruments Dynamco, le DM2023 a une performance garantie dans la gamme complète de 10° à 40° C pendant une période d'un an, sans qu'il soit nécessaire de recourir à n'importe quel matériel d'étalonnage extérieur.

En protégeant et en isolant les circuits par rapport à la masse de puissance, on a pu réaliser des chiffres de rejet de mode commun de 180 dB au courant continue et de 110 dB à 50 Hz. Ces chiffres s'appliquent sans le filtre de 50 dB en circuit c'est à dire pour 50 lectures par seconde et avec n'importe quel matériel d'impression par rapport à la masse.

Le DM2023 a une gamme étendue de modes de fonctionnement, comprenant le fonctionnement minimum et maximum de précision totale, et sa souplesse d'emploi est étendue par l'utilisation de modules d'entrée à fiches. Les deux modules livrables actuellement sont un élément manuel de 10 μV à 1 kV et un élément automatique de 10 μV à 1 kV. Des éléments supplémentaires d'une sensibilité plus élevée pour la mesure du courant alternatif seront offerts à la vente prochainement.

EE 96 752 *pour plus amples renseignements*

CONVERTISSEUR DE BANDES

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Illustration à la page 538)

La société A.P.T. Electronic Industries Ltd vient d'ajouter un convertisseur de bandes à sa gamme de matériels de traitement de données APTEC.

Ce convertisseur de code de bandes, portant la référence AT.3101C, a été conçu pour recevoir des bandes de papier dans n'importe quel régime de code spécifié et produire des bandes à sortie correspondante dans n'importe quel autre régime de code spécifié.

Le lecteur de bandes peut recevoir des bandes à 5, 6, 7 ou 8 niveaux, ou des cartes perforées sur les bords, et le perforateur produira une bande de sortie ou des cartes perforées sur les bords à une vitesse de conversion maximum de 20 codes par seconde.

La machine peut être utilisée pour la conversion de données en un code déterminé en une structure de code différente. Par exemple, les données reçues par le réseau Telex (dans le code No. 2 C.C.I.T.T.) peuvent être aisément converties en un code d'entrée de calculatrice, permettant ainsi aux lignes Telex d'être utilisées pour le transfert de données à vitesse réduite.

D'autres utilisations comprennent la conversion de bandes produites par téléscripteur contrôlés par bande ou machine de bureau en bandes pour entrée directe de calculatrices. Le convertisseur sera également d'utilité lorsque les programmes de calcul en une certaine structure de code doivent être utilisés sur des calculatrices employant des codes d'entrée différents.

La mise en oeuvre du convertisseur de bandes se réduit à l'introduction de la bande d'entrée dans le lecteur de bande et à la pression de la commande de mise en marche. La conversion s'effectuera alors, la perforatrice produisant la bande de sortie de code approprié. Un système de contrôle assure que seules les combinaisons de code reconnues par l'entrée du convertisseur soient traduites dans la perforatrice. Toute combinaison de code qui n'est pas valable provoque l'arrêt du système et éclaire la lampe d'erreurs. Il est donc impossible à des codes erronés de produire une quelconque perforation de sortie.

Des commandes supplémentaires permettent la perforation de n'importe quelle combinaison de code et certaines des combinaisons d'usage fréquent, telles que l'arrêt ou l'effacement, peuvent être perforés au moyen de commutateurs simples.

EE 96 753 *pour plus amples renseignements*

CIRCUITS NUMÉRIQUES À PELLICULE MINCE

Ferranti Ltd, Ferry Road, Edinburgh, Scotland

La société Ferranti Ltd vend maintenant une nouvelle gamme de circuits à logique numérique compatible et accessoires à bas prix construits avec des pellicules minces passives et des dispositifs semi-conducteurs. Les composants passifs sont déposés sur la surface d'une couche sous-jacente en verre, hermétiquement scellée à l'aide d'une ou de plusieurs couches de verre et reliés entre eux par des réseaux en cuivre doré et argenté superposés à gravure photoélectrique.

Deux gammes sont prévues, fonctionnant toutes deux à 1 MHz et 5 MHz.

Le série 1200 est constituée par la gamme universelle fonctionnant entre −30° C et +100° C, tandis que la série 1300 répond principalement aux applications militaires, sa gamme de température s'étendant de −55° C à +125° C.

Les principales caractéristiques de ces circuits sont leur niveau d'immunité à bruit élevé de ±3,5 V, la tension de régime de 6 à 12 V et leur format réduit: 1,27 cm × 0,95 cm × 0,85 cm de hauteur. Deux rangées de câbles à entr'axes de 0,075 pouce ou 0,100 pouce (les deux versions existent), avec localisation sans ambiguité, facilitent l'insertion et l'étude de la carte à circuit imprimé classique.

Le montage à câble ouvert assure une grande densité d'entassement car les modules peuvent en effet être entassés côte à côte et reliés entre eux aisément.

Les types suivants de circuits numériques à pellicule mince peuvent être obtenus immédiatement de la société Ferranti Ltd:

(1) J-K et R-S Flip-Flops.
(2) Déclencheurs Schmitt.
(3) Multivibrateurs à retard.
(4) Portes doubles et à plusieurs entrées.
(5) Tampons.

EE 96 754 *pour plus amples renseignements*

CONVERTISSEUR ANALOGIQUE/NUMÉRIQUE

The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex

(Illustration à la page 539)

Un nouveau convertisseur analogique numérique, considéré comme l'un des plus rapides qui soient construits en Europe est actuellement en cours de production à l'usine de Crawley de la société M.E.L. Equipment Co. Ltd. Ce convertisseur Philips, type PR7830, a une vitesse d'échantillonnage maxima de 20 kHz et son prix se compare très favorablement à celui des instruments américains à action rapide.

Le PR7830 est du type à approximation successive et sa principale application est dans les systèmes rapides d'obtention de données pour les établissements industriels et de recherche. Il peut recevoir une entrée de ±10 V qu'il convertit en une sortie codée binaire pure de 12 chiffres, avec un chiffre binaire additionnel pour identifier la polarité. Il comporte une montre intérieure de 400 kHz mais il peut également être utilisé avec une montre extérieure de 1 à 400 kHz.

Une des caractéristiques les plus importantes de cet instrument est sa très grande simplicité d'emploi. Il peut recevoir n'importe quelle forme d'onde pour l'entrée de commande des chiffres et l'entrée de la montre extérieure. Il peut également recevoir un ordre de déphasage entre les chiffres et des impulsions de montre extérieures. Durant la digitisation, il ne tient pas compte d'impulsions de commande supplémentaires.

La résolution est de 1 dans 4096 et la précision reste à 0,1 % dans une variation de température simultanée de 0 à 40° C et d'alimentation secteur de ±10 % de la valeur nominale.

Il comporte un affichage à tube de néon à sortie parallèle et les circuits sont entièrement constitués de corps solides. Les dimensions hors-tout sont de 48,2 cm de large × 13,33 cm de hauteur × 45,72 cm de profondeur. La construction permet de retirer les plaquettes de circuit imprimé pour avoir accès à tous les composants.

EE 96 755 *pour plus amples renseignements*

MINUTERIE À RETARD

Sterling Instruments Ltd, Sterling Works, Crewkerne, Somerset

(Illustration à la page 539)

Cet instrument peut contrôler n'importe quelle opération minutée et il est particulièrement applicable aux industries photographiques, chimiques et de procédés.

L'aiguille est fixée à n'importe quel temps voulu pour que la machine puisse entrer en action. Le bouton rouge est alors pressé et la machine se mettra en marche automatiquement. Lorsque ce temps a été dépassé, elle s'arrêtera et le bouton rouge devra être pressé à nouveau pour recommencer le cycle.

Les modèles standard prévoient des délais de 0 à 10 sec allant jusqu'à 0 à 12 h. D'autres modèles de délais de 0 à 24 h peuvent être obtenus sur commande spéciale moyennant un supplément de prix. La minuterie fonctionne sur courant secteur alternatif et elle est entraînée par moteur synchrone Sterling. Des minuteries peuvent être fournies pour l'utilisation dans une gamme étendue de tensions allant jusqu'à 440 V.

EE 96 756 *pour plus amples renseignements*

ENREGISTREUR OSCILLOGRAPHIQUE

Graphic Instruments (Research) Ltd, Trout Road, West Drayton, Middlesex

(Illustration à la page 539)

La société Graphic Instruments (Research) Ltd vient d'ajouter un nouvel enregistreur oscillographique galvanométrique à sa gamme d'instruments d'enregistrement. Le "Graphicorder 9-90" est un nouvel instrument à haute résolution de grande précision et à neuf canaux à lecture directe, conçu pour servir d'instrument de haute qualité et à bas prix pour les laboratoires et sur le terrain. Pour la première fois, par suite d'une nouvelle méthode d'optique oscillographique, les avantages de la lampe à iodine au tungstène par rapport à la lampe classique à vapeur de mercure de haute pression sont mis à la disposition des utilisateurs.

Cet enregistreur comprend des alimentations doubles, soit de 200 à 250 V c.a. soit de 24 V c.c., un diagramme d'impression directe de 9 cm × 30,48 mètres, neuf canaux d'éléments galvanométriques interchangeables à réponse de fréquence allant jusqu'à 2500 Hz linéaires à ±5 % ou des sensibilités à partir de 5 μA/cm. La commande de l'instrument s'effectue par deux leviers qui actionnent le ventilateur, la lampe, le diagramme, le commutateur d'arrêt/marche et les changements de vitesse instantanée de 10:1. La lampe peut être allumée ou éteinte à volonté car la lampe à iodine de tungstène ne nécessite pas une alimentation stabilisée, cette dernière causant un retard dans l'emploi et dans la réutilisation après avoir été éteinte. Les vitesses de base du diagramme sont facilement réglées entre la gamme de 2 mm/sec et 1280 mm/sec. On peut également insérer des plaquettes de circuit imprimé pour circuit galvanométrique et résistance parallèle, minuterie à plusieurs échelles et commande électronique de vitesse du papier.

La société Graphic offre une gamme de galvanomètres de quatre catégories pouvant être utilisés dans le Graphicorder 9-90 avec des spécifications appropriées à des applications aussi diversifiées que l'instruction des étudiants et l'enregistrement de vols supersoniques.

EE 96 757 *pour plus amples renseignements*

INDICATEUR DE ZÉRO SÉLECTIF

Standard Telephones & Cables Ltd, STC House, 190 Strand, London, W.C.2

(Illustration à la page 539)

La société Standard Telephones and Cables Ltd vient de créer un indicateur de zéro accordable à sensibilité de 1 μV pour une déviation sur la totalité de l'échelle et à gamme de fréquence de 25 Hz à 100 kHz.

L'indicateur de zéro sélectif 96016-A peut être utilisé comme détecteur sensible dans un équilibrage de pont et comme préamplificateur à faible bruit pour augmenter la sensibilité d'instruments tels que les oscilloscopes et les voltmètres électroniques.

On peut obtenir soit une réponse linéaire soit une réponse logarithmique de l'amplificateur de sortie en actionnant un commutateur et en simplifiant ainsi la recherche d'un point zéro en élargissant l'échelle dans l'emploi logarithmique. On peut, en outre, utiliser des filtres extérieurs pour étendre l'utilité de l'instrument qui est particulièrement indiqué pour l'emploi sur le terrain en raison du fait qu'il fonctionne sur deux batteries sèches internes de 6 V.

L'indicateur a une impédance d'entrée de 25 kΩ à 1 MΩ, suivant le réglage du contrôle de gain, et une impédance de sortie d'environ 300 Ω en série avec 5 pF. La sortie maxima est d'environ 1 V efficace.

Le détecteur mesure 276 mm de large × 229 mm de haut × 248 mm de profondeur et son poids est de 6,8 kg.

EE 96 758 *pour plus amples renseignements*

ALIMENTATION ENCAPSULÉE

Roband Electronics Ltd, Charlwood Works, Lowfield Heath Road, Charlwood, Horley, Surrey

(Illustration à la page 540)

Ces blocs d'alimentation encapsulés fournissent une sortie stabilisée de 2 A (au maximum) à des tensions allant de 6 V à 24 V avec une dissipation maxima de 10 W.

Toutes les tensions dans la gamme peuvent être obtenues à volonté au moyen d'une seule résistance extérieure et la surcharge de courant en retour peut être obtenue de la même manière.

L'encombrement réduit du bloc, qui n'est que de 37,5 cm³, lui permet d'être branché directement à des plaquettes de circuit imprimé ou monté de n'importe quelle autre manière.

En dehors de sa grande robustesse, sa performance du point de vue électronique est aussi bonne que celle des blocs d'alimentation classiques. En outre, n'importe quel nombre de ces blocs peut être fixé en parallèle sur un seul rail non-stabilisé assurant ainsi n'importe quelle quantité voulue de tension et de courant de sortie stabilisée. Cet avantage garantit l'isolement complet contre le couplage croisé de circuits électroniques.

Le rapport de stabilisation est de 1 000 : 1, la résistance de sortie de tension continue est 15 mΩ et l'impédance de sortie jusqu'à 100 kHz est inférieure à 0,5Ω. La tension d'entrée maxima est de 40 V.

EE 96 759 *pour plus amples renseignements*

COMPTEUR-MINUTERIE NUMÉRIQUE PRÉRÉGLÉ

Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(Illustration à la page 540)

Ce compteur-minuterie préréglé, modèle 805R, est un appareil d'une grande souplesse d'emploi, principalement conçu pour l'utilisation à des fins industrielles.

Il peut effectuer des mesures de vitesse numérique et de vitesse de rapport, ainsi que des mesures de temps, de périodes et de fréquence jusqu'à 1 MHz au moins. La base de temps numérique est entièrement variable en plots de 1 msec, de 1 msec à 100 sec, permettant à l'indicateur à cinq chiffres en ligne d'afficher des mesures en n'importe quelle unité voulue telle que mètres/minute, gallons/heure ou tours/seconde.

Il est entièrement composé de circuits au silicium constitués de corps solides d'une parfaite fiabilité et comporte des cartes de circuit imprimé interchangeable afin d'en faciliter l'entretien. La température ambiante de fonctionnement s'étend de $-10°C$ à $+50°C$. Le compteur est de construction robuste et peut être fourni pour montage sur banc d'essai ou sur bâti.

Le cristal incorporé a une stabilité de $\pm$ une partie dans 10^6 par mois ± 1 partie dans 10^6 à court terme. La sensibilité est de 100 mV efficaces et l'impédance d'entrée varie de 10 à 100 kΩ suivant la fonction. Le signal d'entrée pour le fonctionnement chronométrique peut être une transition de tension positive ou négative, une ouverture ou une fermeture de contact (directe ou au moyen d'un circuit de paralysie) ou par bouton-poussoir manuel.

Le modèle 805R peut être utilisé comme fréquencemètre, compteur, chronomètre pour la mesure des intervalles de temps ou comme générateur de retard, avec des retards variant de 1 μsec à 100 sec. Il comporte des accessoires facultatifs comprenant la sortie limite de 1248, la sortie en série pour imprimeur numérique, l'affichage à distance et des modules de commutation supplémentaires pouvant être montés à l'extérieur et utilisés individuellement par bouton-poussoir.

EE 96 760 *pour plus amples renseignements*

CONTRÔLEUR DE TRANSISTORS

Avo, Avocet House, Archcliffe Road, Dover, Kent

(Illustration à la page 540)

Le contrôleur de transistors et de diodes Avo, type **TT 537**, permet d'assurer, à l'aide d'un seul instrument le contrôle de transistors et de diodes. C'est un instrument compact et d'un fonctionnement simple dont les indications à lecture directe constituent une méthode précise et pratique pour la mesure des caractéristiques de transistors et de diodes.

On peut ainsi mesurer rapidement et avec précision l'indice de haute fréquence de transistors jusqu'à 1500 à une fréquence d'environ 1 kHz ainsi que le courant de fuite avec une première indication de 1 μA de transistors au silicium ou au germanium de faible puissance moyenne, tant pnp que npn.

On peut mesurer les caractéristiques d'avances et d'inversion de diodes ainsi que les caractéristiques d'inversion à des tensions allant jusqu'à 1 kV dans des conditions de limite de courant.

La conception de l'instrument permet d'effectuer des mesures précises avec un minimum de réglage. La disposition des commandes du panneau est telle que l'utilisation de l'instrument, dont le fonctionnement est centré sur un commutateur de fonctions et un sélecteur de diodes-transistors, s'explique de soi dans une très grande mesure. Les marques du panneau sont à code de couleurs et ces caractéristiques ainsi que les dispositifs de protection incorporés à l'instrument, en font un contrôleur d'une grande utilité, non seulement pour les ingénieurs et les techniciens mais également pour le personnel non-entraîné.

La tension au collecteur est stabilisée et elle est à variation continue de 0 à 12 V; cette tension est contrôlée à partir d'un instrument de mesure fixé sur le panneau qui contrôle également le courant au collecteur.

EE 96 761 *pour plus amples renseignements*

GAUSSMÈTRE NUMÉRIQUE

Distributeurs: Livingston Laboratories Ltd, Greycaines Estate, Bushey Mill Lane, North Watford, Hertfordshire

(Illustration à la page 540)

Le gaussmètre numérique constitué de corps solides, modèle Bell 660, est considéré comme le premier instrument du genre pouvant être obtenu dans le commerce. Utilisant l'effet Hall, il comporte une indication en ligne à trois chiffres, fournissant une résolution supérieure à une partie dans 1000. Grâce à ses gammes complètes de 100 gauss à 100 kilogauss et une précision de base de $\pm 0{,}3\%$ $\pm 0{,}05$ gauss l'instrument convient pour la plupart des applications de recherche, de mise au point et de développement, y compris la détermination des champs magnétiques dans les magnétrons, les tubes à ondes progressives, les systèmes à indication numérique et de nombreux autres usages.

En plus du modèle numérique, le gaussmètre Bell 660 peut être utilisé comme instrument à lecture directe dans la gamme de 1 gauss à 100 kilogauss avec une précision de 1,5%.

Les sondes Hall utilisent des réseaux à compensation de température pour réaliser une linéarité de $-0{,}002$ pour cent/° C. Des sondes axiales, transversales, différentielles et de X-Y-Z jusqu'à 30 kg peuvent être fournies pour l'emploi avec le gaussmètre.

Chacune des sondes est soumise à un étalonnage à la fabrique pour faciliter l'interchangeabilité rapide tout en maintenant la précision de l'instrument.

EE 96 762 *pour plus amples renseignements*

OSCILLOSCOPE À DOUBLE FAISCEAU

Telequipment Ltd, 313 Chase Road, Southgate, London, N.14

(Illustration à la page 541)

L'oscilloscope à double faisceau D56 comporte deux bases de temps indépendantes ainsi qu'un dispositif de commutation à usages multiples. On peut ainsi coupler une des bases de temps au faisceau supérieur tandis que l'autre base est accouplée simultanément au faisceau inférieur; en variante, l'une ou l'autre peuvent être accouplées aux deux faisceaux. Grâce au tube d'accélération de poste déviation à spirale de 12,5 cm et deux canons et aux commandes séparées de brillance/concentration pour chaque faisceau, on peut observer en même temps deux formes d'ondes non-reliées et effectuer l'ajustement à l'un ou l'autre des affichages.

Chacunes des bases de temps comprend 22 positions étalonnées plus une commande variable de 0,5 μsec/cm à 5 sec/cm.

Une caractéristique supplémentaire du D56 est que la base de temps 'B' peut être branchée de manière à faire fonction de dispositif à retard de la base de temps 'A'. On peut ainsi étaler et montrer simultanément sur la trace supérieure une partie choisie du signal affiché sur la trace inférieure. Un marqueur "d'éclairage" superposé au faisceau inférieur peut être déplacé sur toute la longueur de la trace par la commande de balayage de retard. La partie du signal couverte par le marqueur est déterminée par le réglage du balayage principal et elle est affichée sur toute la longueur de la trace supérieure.

Deux marqueurs sont prévus, à savoir: un marqueur de lignes de télévision utilisant une sortie d'impulsions de 0,5 V à partir de la base de temps 'A' qui peut être injectée à un contrôleur de vidéo pour l'identification de ligne pendant les recherches de forme d'onde vidéo. La largeur d'impulsion est égale à la longueur de balayage, de plus, des tops marqueurs de 10 MHz produits intérieurement peuvent être déclenchés sur le faisceau supérieur pour la mesure du temps de montée.

Les deux amplificateurs Y identiques ont des largeurs de bande allant de la tension continue à 15 MHz (−3 dB) et des sensibilités maxima de 10 mV/cm. Ils sont compensés pour assurer une réponse d'impulsion optima.

Les atténuateurs étalonnés avec précision fournissent des lectures directes de tension d'entrée de 100 mV/cm à 50 V/cm (c.c. à 15 MHz), et de 10 mV/cm à 5 V/cm (c.c. à 500 kHz). La commande de gain variable couvre la gamme entre plots atténuateurs.

Deux modes assurent une synchronisation stable pour toutes les formes d'ondes. La base de temps est automatiquement fixée à n'importe quelle fréquence à partir de quelques cycles par seconde jusqu'à 1 MHz, à n'importe quel niveau d'entrée. La synchronisation HF couvre la gamme de 1 MHz à plus de 15 MHz et une réduction considérable du vacillement de la base de temps a été réalisée.

L'amplificateur X a une largeur de bande du courant continu à 500 kHz (−3 dB) avec une sensibilité maxima de 100 mV/cm. L'étalement X est à variation continue jusqu'à environ 10 diamètres d'écran et la trace s'étale symétriquement autour du centre de l'écran. La commande de déplacement de X fixe la position de n'importe quelle partie de la trace étalée; on peut avoir accès direct à une commande de courant et de balayage.

EE 96 763 *pour plus amples renseignements*

OSCILLOSCOPE PORTATIF

Marconi Instruments Ltd, St. Albans, Hertfordshire

(*Illustration à la page 541*)

Ce nouvel oscilloscope transistorisé (type TF2203) a été spécialement conçu en vue d'applications exigeant un instrument portatif, compact et à bas prix. Il a une largeur de bande d'amplificateur Y de 15 MHz d'une sensibilité de 50 mV/cm. Le calibreur dont il est muni permet la vérification de chaque réglage d'atténuateur.

La spécification comprend en particulier le déclenchement par diode-tunnel, permettant le fonctionnement automatique, et par déclenchement bien au-delà de la largeur de bande d'amplificateur Y; le fonctionnement sur batterie ou sur courant secteur, à l'aide d'un système d'alimentation donnant une indépendance complète des variations de tension secteur sur une gamme très étendue.

L'encombrement réduit, le faible poids (8 kg) et la construction robuste de cet instrument en font un appareil particulièrement indiqué pour l'emploi dans la zone de production, pour l'installation sur place et pour les applications d'entretien, ainsi que pour l'enseignement, les démonstrations et le contrôle.

EE 96 764 *pour plus amples renseignements*

GÉNÉRATEUR À SIGNAL DE BALAYAGE

Airmec Ltd, High Wycombe, Buckinghamshire

(*Illustration à la page 541*)

Le générateur à signaux de balayage, type 352, couvre la gamme de fréquence de 20 Hz à 200 kHz en deux gammes: 20 Hz à 20 kHz et 200 kHz. Chacune de ces gammes peut être balayée soit par accord manuel soit par entraînement de moteur; le degré de balayage est à variations infinies, de 270° (correspondant à l'ensemble de la gamme utilisée) jusqu'à 5°, au moyen d'arrêts de limite, et la vitesse de balayage par entraînement de moteur peut être contrôlée dans des limites étendues.

Le niveau de sortie, stabilisé à ±0,5 dB près sur l'ensemble de la gamme de fréquence, est variable de 10 mV à 10 V en plots de 1 dB par l'emploi des atténuateurs de réglage précis et approximatif. On peut obtenir, à partir d'une impédance de source de 600 Ω, une sortie équilibrée ou déséquilibrée, flottante ou rapportée à la masse.

Le générateur comprend en particulier un détecteur logarithmique incorporé qui fournit une sortie de courant continu et une indication d'instrument de mesure, qui sont toutes deux proportionnelles au logarithme de l'amplitude de la fréquence appliquée.

La sortie de balayage X fournit une tension directe proportionnelle au logarithme de la fréquence instantanée, convenant pour entraîner l'amplificateur X d'un oscilloscope tel que l'oscilloscope Airmec type 279; des impulsions réglables de marqueur de fréquence peuvent également être obtenues à l'aide des prises du panneau arrière.

EE 96 765 *pour plus amples renseignements*

AMPLIFICATEURS DE COURANT ALTERNATIF

Levell Electronics Ltd, Park Road, High Barnet, Hertfordshire

(*Illustration à la page 542*)

Ces amplificateurs ont été conçus pour augmenter les sensibilités d'oscilloscopes et de voltmètres électroniques sur une gamme étendue de fréquences. Les types TA401 et TA601 ont des gains préréglés de 40 dB et de 60 dB tandis que le type TA605 a un gain commuté en plots de 10 dB de 20 dB à 60 dB.

Les transistors planaires au silicium utilisés fournissent une contre-réaction considérable de stabilisation du courant contre les variations de température et de tension d'alimentation.

Le gain de l'amplificateur type TA605 est contrôlé par commutateur à six directions marquées "arrêt," "0 db," "20 dB," "30 db," "40 dB," "50 dB," et "60 dB." Dans la première de ces directions, les connexions d'entrée et de sortie sont reliées lorsque l'amplificateur est isolé et éteint. Le gain dans les autres positions est contrôlé par un potentiomètre préréglé interne pouvant être réglé pour assurer un gain correct à n'importe quel niveau.

L'alimentation est fournie par une batterie autonome à longue durée de vie mais un bloc d'alimentation secteur en courant alternatif peut être fourni en variante pour être incorporé à l'amplificateur type TA 605.

EE 96 766 *pour plus amples renseignements*

CONTRÔLEUR DE CONTINUITÉ DE CÂBLE

Technivision Engineers Ltd, Braywick House, Maidenhead, Berkshire

(*Illustration à la page 542*)

Ce contrôleur de continuité peut vérifier jusqu'à 944 fils métalliques dans un câble ou un circuit électrique. Sa fonction principale est d'indiquer tout circuit ouvert dans un câble ou dans un assemblage de fils similaires, de manière automatique et immédiate. En lui ajoutant un petit circuit discriminateur, il peut détecter un point de haute résistance telle qu'une jointure à soudage sec.

Les défauts sont indiqués sur un compteur électro-mécanique monté sur panneau qui s'arrête immédiatement lorsqu'un défaut est repéré. Le contrôleur peut être utilisé également pour balayer automatiquement 944 circuits afin de vérifier les valeurs d'inductance, de capacitance et de résistance, en liaison avec un instrument approprié tel qu'un pont ou un instrument de mesure universel. Ce contrôle s'effectue à la vitesse de deux circuits par seconde. En variante, et afin de permettre à l'opérateur d'enregistrer les mesures, cette fonction peut être réduite à la vitesse d'un circuit par huit secondes avant de passer automatiquement au circuit suivant.

Une caractéristique intéressante de cet appareil est le fait qu'il peut être utilisé également comme accessoire de bobinage. Dans ce rôle, il est relié à travers un câble approprié au matériel bobiné. La séquence de bobinage de l'appareil est programmée et disposée de façon à ce que le contrôleur ne passe pas au circuit suivant avant que le fil métallique précédent ne soit correctement relié entre ses propres terminaisons pré-conçues. Un commutateur à permutation est prévu pour permettre à un inspecteur de contrôler l'assemblage à n'importe quel moment sur place.

EE 96 767 *pour plus amples renseignements*

CRISTAUX DE QUARTZ

Standard Telephones & Cables Ltd, Components Group, Footscray, Sidcup, Kent

(*Illustration à la page 542*)

En plus de sa gamme étendue de cristaux de quartz à deux broches, enfermés dans des enveloppes métalliques, la division des cristaux de quartz de la société Standard Telephones & Cables Ltd a mis au point et commencé la production en série d'éléments enfermés dans des enveloppes de verre dont on a fait le vide.

Trois modèles sont fournis, à savoir les types 4444, 4445 et 4446 respectivement. Ils répondent aux spécifications DEF L, M et N et sont équivalents au type militaire MIL-C-3098, styles HC-27/U, 26/U et 29/U.

La construction en verre dont on a fait le vide constitue une amélioration des caractéristiques de vieillissement et assure une fiabilité à long terme qui permet d'utiliser ces éléments dans des températures ambiantes allant jusqu'à 150° C.

Une tolérance de fréquence plus rigoureuse est également possible et cet avantage, allié aux caractéristiques de vieillissement amélioré, rendent ces éléments indiqués pour le matériel radioélectrique mobile à espacement entre voies de 12,5 kHz.

Le type 4444 est prévu pour des fréquences de 1,6 MHz à 100 MHz; les types 4445 et 4446 couvrent la bande de 5 MHz à 100 MHz.

EE 96 768 *pour plus amples renseignements*

TABLEAUX DE COMMANDE À BOUTONS-POUSSOIRS

Elliott-Automation Ltd, 167 Great Portland Street, London, W.1

(*Illustration à la page 542*)

Un nouveau système pour produire des tableaux de commande à bouton-poussoir de pratiquement n'importe quelle disposition et grandeur pour la commutation de systèmes électriques et électroniques a été réalisé par la société Elliott-Automation Ltd.

Ce nouveau tableau est basé sur un moulage en alliage léger de construction simple, formant un support pour quatre rangées de 13 commandes et microrupteurs. Le moulage peut être facilement coupé transversalement ou le long des quatre rangées afin de pouvoir lui donner n'importe quelle forme pour n'importe quelle nombre de clés.

Chacun des assemblages à clés est fileté simplement et retenu par des goupilles sur une petite broche fixée dans sa douille sur le moulage principal. Un sensation distincte de déclic est donnée à chaque opération de clé par un petit aimant permanent qui protège momentanément la clé contre la pression des doigts. Ce mécanisme a été mis à l'épreuve avec succès dans une gamme étendue de températures et de conditions de vibration.

Le moulage et les clés peuvent être réalisés rapidement et économiquement suivant la demande du client. La couleur, l'étiquetage et les dimensions des clés en matière plastique sont facultatifs de même que le type et la qualité des microrupteurs miniature et un certain nombre d'autres caractéristiques. Les groupes de clés sont conçus pour le montage conventionnel sur le panneau frontal d'une console ou d'un coffret.

Le tableau comprend en outre un dispositif d'inter-verrouillage mécanique simple qui empêche la pression de plus d'une clé sur un groupe d'un seul moulage, ainsi qu'un degré limité de logique entre les faces pour lesquelles les plaquettes de circuit peuvent être inserées dans les montages de commutation.

EE 96 769 *pour plus amples renseignements*

ENREGISTREUR AUTOMATIQUE DE DÉPLACEMENT

Fairey Engineering Ltd, Stockport, Cheshire

(*Illustration à la page 542*)

Le D.A.T.A.R. (transducteur et enregistreur automatique numérique) mesure le déplacement linéaire jusqu'à une résolution de 0,0001 pouce et imprime automatiquement le résultat suivant un programme minuté ou par commande à distance. Compact et facilement réglé, le D.A.T.A.R. emploie un système de comptage par impulsions sensibles aux phases pour mesurer les mouvements positifs et négatifs. La vitesse d'impression maxima est de 25 unités de résolution par seconde, et la vitesse d'impression maxima est de 4 lectures par seconde. L'impression automatique peut être complétée à n'importe quel moment durant le contrôle par le fonctionnement manuel.

Les deux éléments de base du système D.A.T.A.R. sont le transducteur (fourni avec des résolutions de 0,0001 pouce, 0,001 pouce et 0,01 mm et des gammes de déplacement maxima de 0,5 pouce, 2 pouces et 12 mm respectivement) et l'enregistreur qui exécute le traitement des impulsions, le comptage, le minutage et l'impression.

La précision hors-tout est garantie à ±2 unités de résolution ou ±1 %, suivant celle de ces deux valeurs qui est la plus importante. Quatre commandes sont utilisées pour régler l'appareil et déclencher l'enregistrement automatique des résultats de contrôle qui sont présentés numériquement sur une bande de papier pouvant constituer un enregistrement permanent.

Le transducteur est légèrement plus grand mais de configuration analogue à une jauge à cadran; son poids est de 0,48 kg et il fonctionne de manière parfaitement satisfaisante dans une gamme de température de 0° C à 40° C. L'enregistreur mesure 33 cm × 26,67 cm × 22,86 cm et son poids est de 13 kg. L'ensemble du système est transistorisé pour lui assurer une plus grande fiabilité et sa consommation électrique est réduite (alimentation secteur normale en alternatif).

EE 96 770 *pour plus amples renseignements*

ENREGISTREUR OSCILLOGRAPHIQUE

Southern Instruments Ltd, Camberley, Surrey

(*Illustration à la page 543*)

Ce nouvel enregistreur oscillographique à bas prix, portant la référence MRE.141, a été conçu pour l'industrie mécanique et électrique et il peut être particulièrement utile pour les constructeurs de turbines et de machines rotatives de tout type. Il enregistre simultanément quatre canaux d'information, ainsi que le temps, l'événement et la durée de l'événement sur papier ou film de 70 mm ou 12 mm. Il comprend des dispositifs logiques dans tous les modes de fonctionnement, empêchant le déclenchement de l'événement ou l'exposition accidentelle. Des enregistrements à répétition ou à une seule épreuve peuvent être effectués et des signaux d'étalonnage peuvent être enregistrés avant, pendant ou après l'événement. Le déclenchement s'effectue par niveau de tension d'entrée, par retard de temps préréglé ou par cycle de comptage d'un signal d'entrée. Les programmes peuvent être prévus pour le contrôle à intervalles distincts à des durées allant de 1 milliseconde à deux jours. Huit fonctions peuvent être enregistrées simultanément dans la présentation X-Y, avec des marqueurs supplémentaires de temps et d'événement, et des caméras d'enregistrement sont prévues pour le fonctionnement à un seul coup, à alimentation continue ou par tambour. Une précision de 1 % peut être obtenue et on peut enregistrer à partir du courant continu jusqu'à une fréquence de 250 kHz. On peut également inclure des combinaisons de pré-amplificateurs à modulation de fréquence et à courant continu de 2 MHz et d'autres panneaux auxiliaires.

EE 96 771 *pour plus amples renseignements*

APPAREIL DE REPORT X Y

Distributeurs: Aveley Electric Ltd, South Ockendon, Essex

(*Illustration à la page 543*)

Une vitesse d'enregistrement élevée, soit 1,25 m/sec, et une sensibilité de 0,1 mV/cm à 10 V/cm sont les caractéristiques du nouvel appareil de report Rohde et Schwarz X Y, type ZSK. L'appareil fonctionne suivant le principe du compensateur à équilibrage automatique et effectue des enregistrements linéaires dans les deux sens suivant la norme DIN A3.

Les valeurs mesurées peuvent être à des niveaux de potentiel différents puisque les entrées de tension X et Y ne sont pas mises à la masse et sont galvaniquement séparées l'une de l'autre. Des opérations mathématiques telles que la multiplication et le calcul des racines carrées sont possibles sans ajouter un accessoire quelconque. La souplesse d'emploi de l'appareil de report est encore accrue par les éléments pour montage sur bâti pour la conversion linéaire de tension alternative ou de fréquence en tension continue.

EE 96 772 *pour plus amples renseignements*

TRANSFORMATEURS DE COURANT

Atkins, Robertson & Whiteford Ltd, Industrial Estate, Thornliebank, Glasgow

(Illustration à la page 543)

La société Atkins, Robertson & Whiteford Ltd produit une série de transformateurs de courant circulaires, conçus pour la mesure économique à l'aide d'ampèremètres industriels suivant la norme britannique BSS 89 (1954). Ces transformateurs présentent des avantages de prix par rapport aux modèles de la classe "C," BSS 3938 (1965), c'est à dire la fréquence du type utilisé pour la mesure par ampèremètre.

L'économie de construction a été réalisée par l'emploi de diamètres normaux et par la réduction de la charge à un niveau convenant pour les ampèremètres de montage sur panneau moderne.

L'erreur de rapport maximum sur toute la gamme de fonctionnement est inférieure à 1 % sur la totalité de l'échelle et bien que les transformateurs ne soient pas prévus pour les mesures de Watts, la différence de phase ne dépasse pas 100 minutes.

Le transformateur est muni de bornes secondaires et un support de montage peut être fixé moyennant un petit supplément de prix.

La puissance nominale de ces transformateurs se situe entre 1 VA et 15 VA, suivant le rapport.

EE 96 773 *pour plus amples renseignements*

ATTÉNUATEURS COAXIAUX

Greenpar Engineering Ltd, Station Works, Harlow, Essex

(Illustration à la page 543)

La division d'électronique de la société Greenpar Engineering Ltd a mis au point une nouvelle gamme d'atténuateurs coaxiaux de précision offerts à des prix compétitifs. Ces atténuateurs ont été spécialement conçus afin de présenter une grande précision et une haute stabilité dans une gamme étendue de températures. La gamme d'attenuation normale est de 1, 2, 3, 6, 10, 12, 14 et 20 dB, pouvant tous être prévus dans un choix de trois inter-faces de connecteurs séries "N," "BNC" et "TNC." Le corps de l'atténuateur ainsi que le manchon d'équilibrage d'impédance est en laiton argenté et comprend des réseaux "TEE" à tolérance de film mince très rigoureuse, ainsi que des disques et des tiges, fixées dans des isolateurs en matière plastique isolante thermo-résistante avec contacts dorés; les scellements climatiques sont en caoutchouc au silicium.

L'impédance de ces atténuateurs est de 50 Ω ou 75 Ω et la gamme de fréquence s'étend du courant continu à 4 GHz. Le taux d'ondes stationnaires est inférieur à 1,05 à 1 GHz et inférieur à 1,20 à 4 GHz. La dissipation de puissance continue maxima est de 1 W.

EE 96 774 *pour plus amples renseignements*

Résumés des Principaux Articles

Démarrage de turbines à gaz d'avions par commande électronique par K. F. Bacon

Résumé de l'article aux pages 498 à 506

Dans les avions modernes, il est nécessaire de dispenser l'équipage de tâches routinières. Cet article décrit la manière dont l'une de ces tâches, la mise en marche des moteurs, peut être effectuée automatiquement par commande électronique. Les problèmes que posent l'étude de ces commandes sont décrits en détail ainsi que la manière dont ils ont été résolus. Les circuits des commandes finalement mises au point et leur fonctionnement dans le système de l'avion sont examinés et leur application éventuelle au démarrage des turbines est indiquée.

Un commutateur annulaire économique pour la commutation de faisceaux oscilloscopiques par A. F. Lewis

Résumé de l'article aux pages 507 à 509

Un multivibrateur annulaire simple à quatre directions est utilisé pour la commande de huit commutateurs à transistors push-pull qui fournissent un affichage simultané de huit signaux sur un oscilloscope à double faisceau.

Eléments bistables pour circuits séquentiels par J. J. Sparkes

Résumé de l'article aux pages 510 à 515

L'auteur étudie le problème qui consiste à éviter les conditions d'excitation ambigues de l'univibrateur ou flip-flop R–S; toutes les solutions possibles et raisonnables sont brièvement comparées en fonction de leur simplicité de construction et de leur souplesse d'emploi dans les circuits séquentiels. Il démontre que, même pour des circuits séquentiels fort simples, le flip-flop exclusif OR (ou le flip-flop E) constituerait un perfectionnement par rapport au flip-flop J.K. car il donne lieu non seulement à un degré comparable de simplification des circuits de combinaison mais il est également plus simple à construire puisque, en effet, il peut être construit aussi bien pour le régime synchrone que pour le régime asynchrone et il est identique tant pour la logique positive que pour la logique négative.

Mesure de la fonction de transfert à l'aide d'impulsions rapides par R. C. French

Résume de l'article aux pages 516 à 519

La forme d'onde de temps à la sortie d'un circuit électrique est déterminée uniquement par la forme d'onde à l'entrée du circuit et par le gain et la phase par rapport à la réponse de fréquence, c'est à dire sa fonction de transfert. Inversement si les formes d'onde d'entree et de sortie sont connues, la fonction de transfert peut alors être calculée d'après elles. On a réalisé un système pour mesurer les fonctions de transfert de circuit, suivant lequel les formes d'onde d'entrée et de sortie sont échantillonnées dans le temps et automatiquement enregistrées sur bandes de papier perforées. Une calculatrice est utilisée pour calculer la fonction de transfert de circuit d'après les bandes de papier et les résultats sont imprimés soit sur une machine à écrire soit sur un appareil de report graphique à action rapide sous la forme de gain, phase et retard de groupe par rapport à la fréquence. La précision de la mesure est fonction du type de la fonction de transfert mesurée; au cours de l'expérience en laboratoire dont il est question dans l'article le degré de prècision le plus élevé obtenu fut de ±0,3dB de gain et ±2° de phase avec une gamme dynamique de 10dB.

Le système fonctionne dans la bande jusqu'a 100MHz mais des largeurs de bande beaucoup plus étendues sont possibles.

Réalisation et rectification par temps de propagation de groupe d'un filtre passe-bas pour circuits vidéo par J. W. Head

Résumé de l'article aux pages 520 à 524

Cet article traite de l'application détaillée d'un processus de réalisation publié précédemment (comprenant la rectification par temps de propagation de groupe) et permettant d'obtenir un filtre passe-bas ayant une constante de réponse de 0,5% près dans la bande passante (jusqu'à 5,5MHz), une atténuation théoriquement infinie à 8,25MHz et une atténuation non inférieure à 46dB au-dessus de 8,25MHz. Les résultats pratiques indiqués furent obtenus lorsque ce processus fut utilisé pour étudier un filtre sensiblement plus petit que ceux de caractéristiques analogues que l'on pouvait obtenir précédemment.

Transformation de fréquence pour réseaux de courant alternatif par J. G. Advani et M. N. Neelkantan

Résumé de l'article aux pages 525 à 527

La transformation $p=\frac{2}{3}\,\frac{s(s^2+1)}{(s^2-0,25)}$ *pour un intégrateur de courant alternatif a été proposée par Prabbakar et elle aurait 1/2 pour cent d'erreur de linéarité et de symétrie dans la gamme de fréquence 50% d'une extrémité à l'autre de la fréquence porteuse. Cet article montre que cette erreur est d'environ 8% dans cette gamme de fréquence. Une nouvelle transformation (pour intégrateur de courant alternatif) donnant une erreur d'environ 3% dans cette gamme de fréquence est également présentée.*

Une autre transformation pour obtenir des fonctions de transfert de réseaux de compensation en alternatif à partir des fonctions de transfert de réseaux de compensation en continu est également suggérée. Cette transformation donne une erreur de linéarité et de symétrie de ½% dans la gamme de fréquence de 10% d'une extrémité à l'autre de la fréquence porteuse.

Les réseaux pour un intégrateur de courant alternatif et un réseau de courant alternatif à conducteur de retard utilisant des éléments R C et actifs sont synthétisés.

Une méthode graphique pour réaliser des ponts transducteurs à résistance de gamme étendue par H. S. Sear

Résumé de l'article aux pages 528 à 530

Il s'agit dans cet article d'une méthode graphique pour l'étude de ponts de Wheatstone. Cette méthode obvie en grande partie à la non-linéarité de transducteurs de résistance tels que les thermistors lorsqu'ils sont utilisés dans des gammes étendues. Cette méthode indique l'importance de l'erreur résiduelle et son indice, ainsi que le point précis dans la gamme du transducteur où elle se produira. La méthode consiste à adapter la courbe de caractéristiques du transducteur à l'une des courbes de caractéristiques de pont. L'application de cette méthode à l'étude des thermomètres à thermistors de gamme étendue avec sortie linéaire est indiquée.

Le facteur de régulation pour les références à diodes zéner à compensation de température par H. Banasiewicz

Résumé de l'article aux pages 530 à 531

Les références zéner à compensation de température du même coefficient de température diffèrent souvent en ce qui concerne un ou plusieurs des trois paramètres de base: la tension zéner, le courant et la résistance de pente. L'auteur montre comment on peut combiner ces chiffres afin qu'ils puissent donner une indication directe de la capacité de régulation d'un dispositif contre les changements de courant de polarisation.

L'avenir des communications graphiques par T. M. C. Lance

Résumé de l'article aux pages 532 à 535

Cet article étudie les méthodes de communication d'informations graphiques et examine certains des usages futurs de ces méthodes.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 538 bis 543

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Automatisch abgleichende Messbrücke

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Abbildung Seite 538)

Die neue Bauelemente-Messbrücke B421 "Compact" von Wayne Kerr gibt direkte Messgeräteanzeigen für L, C, R, Serien- und Parallelverluste, den Reststrom von Elkos und Toleranz als Prozentdifferenz zwischen jeweils zwei Bauelementen. Die selbstabgleichende Schaltung macht die Bedienung äusserst einfach: man braucht nicht von Hand zu nullen, und da die Brücke laufend durch elektronisches Nullen abgeglichen wird, kann man sogar Wertänderungen beobachten.

Quelle und Anzeige (1000 Hz) sind eingebaut, und die Festkörperschaltungen werden wie üblich durch Wechselstromanschluss oder Batterien gespeist. Bei Widerstands- und Kapazitätsmessungen gestattet ein schaltbarer Regler Messungen mit 0,25 Prozent Genauigkeit auf allen sieben Bereichen. Der Messumfang ist 1 μH...100 H, 0,01 pF...10 μF und 10 mΩ...100 MΩ.

EE 96 751 *für weitere Einzelheiten*

Digitalvoltmeter

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Abbildung Seite 538)

Das Kompensations-Digitalvoltmeter DM 2023 hat bei einer Skala von 99999 unter optimalen Bedingungen eine Messunsicherheit von 0,001 Prozent des Skalenendwertes ±0,0025 Prozent der Anzeige und wurde für anspruchsvollere Anwendungen, bei denen es auf Genauigkeit, Konstanz und hochwertige Eingang-Ausgangtrennung ankommt, entwickelt. Wie bei allen anderen Dynamco-Instrumenten sind auch für das DM 2023 alle Kenndaten für ein Jahr über den vollen Bereich von 10...40° C ohne Rückfallen auf externe Eichvorrichtungen garantiert.

Durch sorgfältiges Abschirmen und Trennen der Schaltungen von der Netzerde wurde Gleichtaktunterdrükkung von 180 dB bei Gleichstrom und 110 dB bei 50 Hz erreicht. Diese Werte gelten ohne Einschaltung des 50-dB-Filters, d.h. bei 50 Anzeigen je Sekunde, und Beziehen irgendwelcher Druckwerke auf Masse.

Das DM 2023 hat viele Betriebsarten, einschliesslich Maximum- und Minimumbetrieb bei voller Messgenauigkeit; seine Vielseitigkeit lässt sich durch Eingangseinschübe noch erweitern. Die beiden bereits lieferbaren Einschübe sind ein Modul für automatische Bereichschaltung von 10 μV...1 kV und ein Modul für manuelle Bereichschaltung von 10 μV...1 kV. Weitere Einschubmoduln für höhere Empfindlichkeit, Wechselstrommessungen und Strommessungen werden in Kürze angekündigt werden.

EE 96 752 *für weitere Einzelheiten*

Lochstreifen-Umsetzer

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Abbildung Seite 538)

A.P.T. Electronic Industries Ltd hat ihr APTEC-Programm für Datenverarbeitungsgeräte durch einen Lochstreifen-Umsetzer erweitert.

Das Lochstreifen-Umsetzgerät AT 3101C ist für Eingabe von Lochstreifen in einem beliebigen spezifizierten Code ausgelegt und erzeugt einen entsprechenden Ausgangslochstreifen in einem anderen spezifizierten Code.

Der Streifenleser kann Streifen im 5-, 6-, 7- oder 8-Kanalcode oder Randlochkarten lesen und der Ausgangs-Locher Streifen oder Randlochkarten mit einer Höchstumsetzgeschwindigkeit von 20 Zeichen/Sekunde lochen.

Die Einrichtung ist überall dort für Einsatz geeignet, wo ein Code in eine andere Code-Struktur umgewandelt werden muss. So können z.B. über das Telex-Netzwerk in CCITT-Code Nr. 2 empfangene Daten ohne weiteres in einen für Eingabe in eine EDV-Anlage geeigneten umgesetzt werden, wodurch die Benutzung von Telex-Leitungen für Datenübertragung mit niedriger Geschwindigkeit möglich wird.

Ausserdem können u.a. von lochstreifengesteuerten Schreibmaschinen und Büromaschinen erstellte Lochstreifen in Streifen für direkte EDV-Eingabe umgesetzt werden. Das Gerät wird auch dort nützlich sein, wo ein Computerprogramm in einer Codestruktur in einem Computer mit anderer Eingabecodestruktur Verwendung finden soll.

Die Bedienung des Lochstreifenumsetzers ist einfach; der Eingangslochstreifen wird in den Leser eingelegt und die Starttaste gedrückt. Danach erfolgt die Umsetzung, und der Locher erzeugt den entsprechenden Ausgangslochstreifen. Eine Gültigkeitskontrolle sorgt dafür, dass nur vom Untersetzereingang erkannte Codekombinationen umgewandelt werden. Jede ungültige Codekombination stoppt die Anlage durch Aufleuchten von Fehlerlampen, was es unmöglich macht, fehlerhafte Codes irgendwelcher Ausgangslochung hervorzurufen.

Zusätzliche Organe ermöglichen das Lochen jeder Codekombination; einige oft angewendete Kombinationen wie z.B. Stopp oder Streichen können mittels Einzelschalter gelocht werden.

EE 96 753 *für weitere Einzelheiten*

Dünnfilm-Digitalschaltungen

Ferranti Ltd, Ferry Road, Edinburgh, Scotland

Ein neues Angebot der Ferranti Ltd umfasst preisgünstige kompatible Digitallogik- und Zusatzschaltungen, die aus passiven Dünnfilmelementen und diskreten Halbleitern aufgebaut sind. Passive Bauelemente werden auf die Oberfläche eines Glassubstrats aufgedampft, mittels eines oder mehrerer Superstrats aus Sinterglas hermetisch abgeschlossen und durch überlagerte, photogeätzte, versilberte oder vergoldete Kupferleitungszüge verbunden.

Es gibt zwei Baureihen, die beide bei 1 MHz und 5 MHz arbeiten. Baureihe 1300 ist für militärische Anwendung, hauptsächlich im Bereich $-55°$ C... $+125°$ C gedacht.

Hauptmerkmale dieser Schaltungen sind der hohe Störfestigkeitspegel von $\pm 3{,}5$ V, die Arbeitsspannung von 6...12 V und die kleinen Abmessungen von 12,7 $\times$ 9,5 $\times$ 9,5 mm hoch. Anschlussdrähte werden in zwei Reihen und wahlweise mit Rastermass 1,9 oder 2,54 mm geliefert; eindeutige Führung erleichtert Einsetzen und Aufbau der herkömmlichen Druckplatten. Die Anordnung der Anschlüsse gestattet hohe Packdichten, da Moduln nebeneinander gestapelt und ohne Schwierigkeiten verdrahtet werden können.

Dünnfilm-Digitalschaltungen folgender Typen sind sofort von Ferranti Ltd erhältlich:

(1) K-K- und R-S-Flip-Flops
(2) Schmitt-Triggerschaltung
(3 Verzögerungsmultivibrator
(4) Mehrfacheingang und Doppeltore
(5) Puffer.

EE 96 754 *für weitere Einzelheiten*

Analog-Digitalwandler

The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex

(*Abbildung Seite 539*)

Bei M.E.L. Equipment Co Ltd in Crawley ist die Fertigung eines neuen Analog-Digitalwandlers angelaufen, der einer der schnellsten in Europa hergestellten sein soll. Dieser Philips-Wandler Typ PR 7830 hat eine maximale Abtastfrequenz von 20 kHz; im Vergleich mit schnellen amerikanischen Wandlern liegt der Preis sehr günstig.

Der PR 7830 arbeitet nach dem Stufenkompensationsprinzip und findet hauptsächlich in Datenerfassungssystemen für Industrie und Forschung Anwendung. Eingaben von ± 10 V werden in einen rein binär codierten 12-Bit-Ausgang umgewandelt; ein zusätzliches Bit gibt die Polarität. Da der interne Taktgeber mit 400 kHz arbeitet, kann man das Gerät mit externen Taktgebern von 1...400 kHz einsetzen.

Wesentlich ist, dass das Gerät so einfach in der Anwendung ist. Es akzeptiert jede beliebige Wellenform als Eingabebefehl für die Digitalumsetzung und den externen Taktgeber, falls ein solcher angeschlossen ist. Es kann auch eine Phasenverschiebung zwischen dem Digitalumsetzungsbefehl und externen Taktimpulsen tolerieren; weitere Befehle zur Digitalumsetzung bleiben während der Wandlung unbeachtet.

Die Auflösung ist 1:4096, und die Genauigkeit bleibt während gleichzeitiger Temperaturänderung von 0° auf +40°C und Netzschwankungen von ± 10 Prozent innerhalb 0,1 Prozent.

Der Parallelausgang wird auf Neonröhren dargestellt; alle Schaltungen sind in Festkörpertechnik ausgeführt. Die Abmessungen sind 483 $\times$ 134 $\times$ 457 mm tief; die Konstruktion gestattet Auslegung auf internen Schaltungsplatten, um den Zugang zu allen Bauelementen zu erleichtern.

EE 96 755 *für weitere Einzelheiten*

Verzögerungsschaltuhr

Sterling Instruments Ltd, Sterling Works, Crewkerne, Somerset

(*Abbildung Seite 539*)

Dieses Modell ist für die Steuerung jedes zeitgebundenen Vorgangs, vor allem in den Sektoren Fotografie, Chemie und Verfahrenstechnik geeignet.

Der Zeiger wird auf die Zeit eingestellt, für die das Gerät laufen soll. Nach Drücken der roten Taste startet es dann automatisch und stoppt nach Ablauf der Zeit. Für den Start eines neuen Zyklus muss die rote Taste wieder gedrückt werden.

An lieferbaren Standardmodellen lässt die Verzögerung sich von 0...10 s bis zu 0...12 h spezifizieren. Andere Modelle bis zu 0...24 h können gegen Sonderauftrag und Zuschlag hergestellt werden.

Die Schaltuhr arbeitet mit Wechselstromnetzanschluss und wird durch den bekannten Sterling-Synchronmotor getrieben. Schaltuhren sind für Spannungen bis zu 440 V lieferbar.

EE 96 756 *für weitere Einzelheiten*

Direktschreiber

Graphic Instruments (Research) Ltd, Trout Road, West Drayton, Middlesex

(*Abbildung Seite 539*)

Graphic Instruments hat ihr Schreiberprogramm durch einen neuen Galvanometer-Direktschreiber erweitert. Das neue Instrument—der Graphicorder 9-90—ist ein direktschreibender Galvanometerkurvenschreiber kleiner Abmessungen, mit hoher Auflösung und Genauigkeit mit neun Kanälen, der besonders für hohe Qualität in Laborarbeiten und Aussendienst bei niedrigem Aufwand konstruiert wurde. Als Ergebnis einer neuen Einstellung zu den Problemen der Schreiberoptik stehen Anwendern erstmalig die Vorteile der Wolfram-Jodlampe über die herkömmliche Hochdruck-Quecksilberlampe zur Verfügung.

Der Schreiber hat eine Doppelstromversorgung entweder für 200...250 V$\sim$ oder 24 V—, einen 9 cm breiten und 30,5 m langen Direktschreib-Registrierstreifen, auswechselbare Galvanometerwerke mit Frequenzgang von ± 5 Prozent bis zu 2500 Hz oder Ablenkfaktoren von 5 μA/cm in 9 Kanälen. Die Bedienung des Gerätes erfolgt über zwei Hebel, die Ventilation, Lampe, Registrierstreifen, Ein-Ausschalten und eine sofortige Vorschubänderung von 10:1 steuern. Die Lampe kann jederzeit ein- und ausgeschaltet werden; die Wolfram-Jodlampe benötigt keine Konstantstromversorgung, die üblicherweise nach Ausschalten und vor Wiederbenutzung eine Verzögerung verursacht. Der Grundstreifenvorschub ist ohne Schwierigkeiten zwischen 2 mm/s und 1280 mm/s einstellbar. Druckschaltungsplatten mit Serien- und Parallelwiderständen zur Anpassung von Galvanometerschaltungen, Mehrfach-Zeitmassstäben und elektronischer Vorschubregelung können eingesteckt werden.

Für den Graphicorder 9-90 bietet Graphic eine Auswahl von Galvanometern in vier Klassen an, deren technische Daten den Anforderungen der unterschiedlichsten Anwendungen vom Unterricht bis zum Überschallflug genügen.

EE 96 757 *für weitere Einzelheiten*

Selektiver Abgleichindikator

Standard Telephones & Cables Ltd, STC House, 190 Strand, London, W.C.2

(*Abbildung Seite 539*)

Standard Telephones & Cables Ltd hat einen neuen abstimmbaren Abgleichindikator mit 1 μV Vollausschlagempfindlichkeit und einem Frequenzbereich von 25 Hz...100 kHz eingeführt.

Der selektive Abgleichindikator 96016-A lässt sich als empfindlicher Anzeiger für den Brückenabgleich und als rauscharmer Vorverstärker zur Erhöhung der Empfindlichkeit von Instrumenten, z.B. Oszillografen und Röhrenvoltmetern, einsetzen.

Durch Bedienung eines Schalters kann der Ausgang von linearer auf logarithmische Anzeige umgeschaltet und durch Dehnung der Skala im logarithmischen Zustand die Suche nach dem Nullpunkt vereinfacht werden. Zur Erweiterung des Anwendungsbereiches des Instruments, das durch Speisung aus zwei internen 6-V-Trockenbatterien besonders für den Aussendienst geeignet ist, kann man externe Filter anschliessen.

Die Eingangsimpedanz des Gerätes liegt zwischen 25 kΩ und 1 MΩ und ist von der Verstärkungseinstellung abhängig; die Ausgangsimpedanz ist etwa 300 Ω in Serie mit 5 pF. Die Höchstausgangsspannung ist 1 V_{eff}.

Die Aussenabmessungen sind 276 × 229 × 248 mm, das Gewicht 6,8 kg.

EE 96 758 *für weitere Einzelheiten*

Eingekapselte Stromversorgungen

Roband Electronics Ltd, Charlwood Works, Lowfield Heath Road, Charlwood, Horley, Surrey

(*Abbildung Seite 540*)

Diese eingekapselten Stromversorgungen geben bei Spannungen zwischen 6 V und 24 V und einer Höchstverlustleistung von 10 W einen konstanten Ausgangsstrom von maximal 2 A ab.

Jede beliebige Spannung innerhalb des Bereiches kann mittels eines externen Widerstandes und der Stromüberlastungsschutz auf ähnliche Weise eingestellt werden.

Da der Baustein nur 41 cm^3 gross ist, kann er direkt in Druckschaltungsplatten eingelötet oder auf jede andere Art montiert werden.

Abgesehen von seiner robusten Konstruktion ist der Baustein elektronisch so leistungsfähig wie herkömmliche Konstantstromquellen, mit dem zusätzlichen Vorteil, dass er sich in beliebiger Anzahl parallel an eine nicht stabilisierte Einzelspeiseleitung legen lässt, wenn eine Anzahl diskreter Konstantspannungen und -ströme benötigt werden. Dadurch werden elektronische Schaltungen zur Verhinderung von Verkopplung vollisoliert.

Die Regelung ist 1000:1, der ohmsche Ausgangswiderstand 15 mΩ und die Ausgangsimpedanz bis zu 100 kHz niedriger als 0,5 Ω. Die höchste Eingangsspannung ist 40 V.

EE 96 759 *für weitere Einzelheiten*

Vorwahl-Digitalzähler-Zeitmesser

Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(*Abbildung Seite 540*)

Dieser vielseitige Vorwahlzähler-Zeitmesser 805R ist in erster Linie für industriellen Einsatz gedacht.

Digitalmessungen von Geschwindigkeit und Übersetzungsverhältnis, Zeit, Intervall und Frequenzen bis zu 1 MHz sind möglich. Die digitale Zeitbasis ist in 1-ms-Stufen von 1 ms bis zu 100 s einstellbar, was fünfstellige, einzeilige "Speicher"-Anzeige in beliebigen technischen Einheiten wie z.B. Meter/Minute, Gallonen/Stunde oder Umdrehungen/Sekunde gestattet.

Für grösste Zuverlässigkeit und einfachen Service werden durchweg Silizium-Festkörperschaltungen mit Einsteck-Druckschaltungsplatten verwendet. Der Bereich der Betriebsumgebungstemperatur ist $-10°...+50°$ C, die Konstruktion robust. Der Zähler ist für Labortisch oder Gestelleinbau lieferbar.

Der eingebaute Quarz hat eine Konstanz von $\pm 1 \times 10^{-6}$ je Monat, $\pm 1 \times 10^{-6}$ kurzfristig. Die Empfindlichkeit ist 100 mV_{eff}, die Eingangsimpedanz je nach Funktion 10 oder 100 kΩ. Das Eingangssignal für Chronometerbetrieb kann aus einem positiven oder negativen Spannungsübergang, Kontaktöffnen oder -schliessen (direkt oder über Verriegelungsschaltung) bestehen oder manuell gestastet werden.

Das Modell 805R kann man als Frequenzmesser, Quotientenmesser, Chronometer für Zeitintervallmessungen oder zwischen 1 µs und 100 s einstellbaren Verzögerungsgenerator einsetzen. Wahlweise gibt es u.a. 1-2-4-8 binärkodierte Dezimalausgänge, Serienausgabe für Digitaldrucker, Fernanzeige und zusätzliche Schaltmoduln, die extern montiert und einzeln durch Drucktaste betätigt werden können.

EE 96 760 *für weitere Einzelheiten*

Transistor-Tester

Avo, Avocet House, Archcliffe Road, Dover, Kent

(*Abbildung Seite 540*)

Der Transistor- und Diodentester Avo TT537 gestattet Prüfen von Transistoren und Dioden mit einem Gerät. Der Tester ist kompakt, einfach zu bedienen, und ein direktanzeigendes Messgerät bietet eine genaue und bequeme Methode zum Messen von Transistor- und Diodenkenndaten.

Transistor-h_{FE}-Werte bis zu 1500 können bei etwa 1 kHz Frequenz schnell und genau gemessen werden; Reststrommessungen von Germanium- und Siliziumtransistoren mittlerer Leistung erfolgen sowohl für pnp- wie npn-Typen mit einer ersten Anzeige von 1 µA.

Kenndaten von Dioden sind sowohl in Durchlass- wie Sperrichtung bei Spannungen bis zu 1 kV und unter Strombegrenzungsbedingungen messbar.

Die Konstruktion des Gerätes gestattet genaue Messungen mit Minimumnachregelung und -einstellung durchzuführen. Die um den Funktionsschalter und einen Transistor-Diodenwähler zentrierte Arbeitsweise des Gerätes bedarf wegen der Anordnung der Bedienelemente auf der Frontplatte kaum einer Erklärung. Die Markierungen auf der Frontplatte sind farbkodiert, wodurch das Gerät zusammen mit den eingebauten Schutzvorrichtungen nicht nur für Anwendungen von Ingenieuren und Technikern, sondern auch für ungelerntes Personal geeignet ist.

Die Kollektorspannung ist stabilisiert und kontinuierlich zwischen 0...12 V regelbar; Kollektorspannung und -strom werden mittels eines eingebauten Messgerätes abgelesen.

EE 96 761 *für weitere Einzelheiten*

Digitalgaussmesser

Vertrieb: Livingston Laboratories Ltd, Greycaines Estate, Bushey Mill Lane, North Watford, Hertfordshire

(*Abbildung Seite 540*)

Es wird angenommen, dass das Bell-Modell 660 der erste kommerziell greifbare Festkörper-Digitalgaussmesser ist. Das den Hall-Effekt ausnutzende Gerät hat eine einzeilige, dreistellige Anzeige, deren Auflösungsvermögen besser als 1×10^{-3} ist. Mit Bereichendwerten von 100 Gauss bis zu 100 Kilogauss und einer Grundmessunsicherheit von ±0,3 Prozent ±0,05 Gauss ist dieses Instrument für die meisten Anwendungszwecke in Forschung, Entwicklung und Fertigung, einschliesslich der Messung magnetischer Felder in Magnetrons, Wanderfeldröhren, magnetischen Kernresonanzsystemen usw., geeignet.

Ausser der Digitalausführung gibt es den Bell 660 auch als direktanzeigenden Gaussmesser mit grösserem Messumfang von 1 Gauss bis zu 100 Kilogauss mit 1,5 Prozent Messunsicherheit.

Die Hall-Tastköpfe sind mit temperaturkompensierten Netzwerken ausgerüstet, die eine Linearität von $-0{,}002$ %/°C gewährleisten. Für Einsatz mit diesem Gaussmesser sind Axial-, Transversal-, Differential- und X-Y-Z-Tastköpfe lieferbar.

Jeder Tastkopf wird im Werk geeicht, um die Messgenauigkeit des Gerätes auch beim schnellen Auswechseln der Köpfe aufrechtzuhalten.

EE 96 762 *für weitere Einzelheiten*

Zweistrahl-Oszillograf

Telequipment Ltd, 313 Chase Road, Southgate, London, N.14

(*Abbildung Seite 541*)

Im Zweistrahl-Oszillograf D56 gibt es zwei unabhängige Zeitablenkungen mit vielseitigen Schaltanordnungen. Das ermöglicht, eine Zeitbasis mit dem oberen Strahl, die andere gleichzeitig mit dem unteren zu koppeln; andererseits kann die eine oder andere mit beiden Strahlen gekoppelt werden. Die 5-Zoll-(127 mm)-Röhre mit wendelförmiger Nachbeschleunigung und zwei Elektronenstrahlsystemen hat für jeden Strahl getrennte Helligkeitsregelung und Fokussierung, was gestattet, gleichzeitig zwei in keiner Beziehung stehende Wellenformen zu beobachten und jede Darstellung ohne Beeinflussung der anderen nachzustellen.

Jede Zeitablenkung hat 22 geeichte Positionen plus Feinregelung von 0,5 µs/cm...5 s/cm. Einmalablenkung mit Einschaltsperre gibt es auch.

Eine weitere Eigenschaft des D56 ist, dass Zeitablenkung "B" so umgeschaltet werden kann, dass sie als Verzögerung der Zeitablenkung "A" wirkt. Dadurch wird es möglich, ausgewählte Abschnitte der durch die untere Spur dargestellten

Signale gleichzeitig gedehnt auf der oberen Spur darzustellen. Eine der unteren Spur überlagerte Aufhellmarke lässt sich mit Hilfe des Verzögerungsreglers der Zeitablenkung über die ganze Spurenlänge verschieben. Der durch die Marke überdeckte Signalabschnitt wird durch Einstellung der Hauptzeitablenkung eingestellt und über die gesamte Länge der oberen Spur dargestellt.

Zwei Markierungseinrichtungen sind vorgesehen; eine entnimmt der Zeitablenkung A einen 0,5-V-Impuls, der während der Untersuchung der Videowellenform für Zeilenkennzeichnung in einen Videomonitor gespeist werden kann. Die Impulslänge entspricht der Länge der Zeitablenkung. Ausserdem gibt es intern erzeugte 10-MHz-Marken für Anstiegszeitmessungen auf der oberen Spur.

Zwei identische Y-Verstärker haben Bandbreiten von 0...15 MHz (−3 dB) und Ablenkkoeffizienten von max. 10 mV/cm. Sie sind für optimale Impulswiedergabe kompensiert.

Genau geeichte Abschwächer gestatten Direktablesen von Eingangsspannungen zwischen 100 mV/cm und 50 V/cm (0...15 MHz) und zwischen 10 mV/cm und 5 V/cm (0...500 kHz). Zwischen den Abschwächerstufen gibt es eine Verstärkungsregelung.

Gleichbleibende Synchronisierung aller Wellenformen kann in zwei Betriebsarten erreicht werden. Die Zeitablenkung synchronisiert automatisch mit jeder Frequenz von einigen Hertz bis zu 1 MHz und bei beliebigem Eingangspegel; HF-Synchronisierung überstreicht den Bereich von 1 MHz bis zu über 15 MHz mit wesentlich reduziertem Zittern.

Der X-Verstärker hat eine Bandbreite von 0...500 kHz (−3 dB) mit maximalem Ablenkkoeffizienten von 100 mV/cm. Die X-Dehnung ist kontinuierlich bis zu etwa 10x Schirmdurchmesser regelbar, und die Spur wird symmetrisch um die Bildmitte gedehnt. Durch Verschiebung in X-Richtung lässt sich jeder Teil der gedehnten Spur einstellen. Der X-Verstärker ist direkt zugänglich, und die Zeitablenkung kann entnommen werden.

EE 96 763 *für weitere Einzelheiten*

Tragbarer Oszillograf

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Abbildung Seite 541)

Dieser neue transistorierte Oszillograf TF2203 ist vor allem für Anwendungszwecke bestimmt, die ein tragbares, kompaktes und preiswertes Gerät verlangen. Er hat eine Y-Verstärkerbandbreite von 15 MHz mit einem Ablenkkoeffizienten von 50 mV/cm. Jede Abschwächereinstellung lässt sich mit einem vorgesehenen Eicher prüfen.

Zu den Haupteigenschaften zählen u.a. Tunneldiodentriggerung für automatischen und getriggerten Betrieb weit über die Y-Verstärkerbandbreite hinaus, sowie wahlweiser Batterie- oder Netzbetrieb mit einer Stromversorgung, die selbst von grossen Netzschwankungen unabhängig macht.

Wegen seiner kleinen Abmessungen, niedrigem Gewicht (6,75 kg) und robuster Bauart eignet sich dieses Gerät für Einsatz im Werk, Installation und Wartung an Ort und Stelle, Unterricht, Demonstration und als Monitor.

EE 96 764 *für weitere Einzelheiten*

Wobbler

Airmec Ltd, High Wycombe, Buckinghamshire

(Abbildung Seite 541)

Der Wobbler 352 überstreicht 20 Hz... 200 kHz in zwei Bereichen: 20 Hz...20 kHz und 200 kHz. Beide Bereiche können entweder manuell durchgestimmt oder mit Motorantrieb durchlaufen werden; der Wobbelgrad ist zwischen 270° (dem gesamten benutzten Bereich entsprechend) und 5° mittels kontinuierlich verschiebbarer Anschläge veränderlich, und die Wobbelgeschwindigkeit ist bei Motorantrieb auch zwischen breiten Grenzen regelbar.

Der über den Gesamtfrequenzbereich innerhalb ±0,5 dB konstantgehaltene Ausgangspegel ist mittels Grob- und Feineinstellung in 1-dB-Stufen zwischen 10 mV und 10 V regelbar. Der Ausgang von einer 600-Ω-Quellenimpedanz ist entweder symmetrisch oder unsymmetrisch, erdfrei oder auf Erde bezogen wählbar.

Ein wesentliches Merkmal ist der eingebaute logarithmische Demodulator, dem ein Gleichstromausgang und eine Messgerätanzeige entnommen werden, die bei dem Logarithmus der Amplitude der angelegten Frequenz proportional sind.

Der X-Abtastungsausgang gibt eine Gleichspannung ab, die dem Logarithmus der Momentanfrequenz proportional ist und den X-Verstärker eines Oszillografen wie z.B. des Airmec 279 treiben kann; einstellbare Frequenzmarkenimpulse können Buchsen auf der Rückseite entnommen werden.

EE 96 765 *für weitere Einzelheiten*

Vorverstärker

Levell Electronics Ltd, Park Road, High Barnet, Hertfordshire

(Abbildung Seite 542)

Diese Verstärker wurden für Erhöhung der Empfindlichkeit von Oszillografen und elektronischen Voltmetern über einen grossen Frequenzbereich entwickelt. Die Typen TA401 und TA601 haben eine voreingestellte Verstärkung von 40 dB bzw. 60 dB, während Typ 605 eine in 10-dB-Stufen zwischen 20 dB und 60 dB wählbare Verstärkung hat.

Silizium-Planartransistoren werden verwendet, und zur Stabilisierung gegen Temperatur- und Speisespannungsschwankungen wird stark gegengekoppelt.

Die Verstärkung von Typ TA605 wird mittels eines sechsstufigen Schalters gewählt, dessen Positionen mit “Aus und 0 dB”, “20 dB”, “30 dB”, “40 dB”, “50 dB” und “60 dB” markiert sind. In der ersten Position wird der Eingang mit dem Ausgang verbunden, der Verstärker abgeschaltet und isoliert. Die Verstärkung in den anderen Positionen wird durch einen voreingestellten Spannungsteiler gesteuert, der für jeden Verstärkungspegel einregelbar ist.

Speisung erfolgt aus einer Langlebensdauerbatterie, lässt sich aber für den Verstärker TA605 durch einen Wechselstromnetzanschluss ersetzen.

EE 96 766 *für weitere Einzelheiten*

Leitungsprüfer

Technivision Engineers Ltd, Braywick House, Maidenhead, Berkshire

(Abbildung Seite 542)

Dieser Leitungsprüfer kann bis zu 944 Leiter in Kabelbäumen oder elektrischen Schaltungen prüfen. Seine Hauptaufgabe ist, sofort und automatisch Leiterunterbrechungen in Kabelbäumen und ähnlichen Leitungssätzen anzuzeigen. Mit einer kleinen Diskriminatorschaltung als Zusatz können Punkte hohen Widerstandes wie z.B. kalte Lötstellen entdeckt werden.

Fehler werden auf einem elektromechanischen Zähler auf der Frontplatte angezeigt, der stoppt, sowie ein Fehler entdeckt wird. Man kann den Prüfer auch zusammen mit einem geeigneten Messgerät, z.B. einer Brücke oder einem Universalinstrument, für das automatische Abtasten von 944 Schaltungen und Prüfen der jeweiligen Selbstinduktions-, Kapazitäts- und Widerstandswerte einsetzen. Das geschieht mit einer Geschwindigkeit von zwei Schaltungen pro Sekunde, kann jedoch auf eine Schaltung jede acht Stunden herabgesetzt werden, um der Bedienung Registrieren von Messergebnissen zu ermöglichen.

Diese Ausrüstung ist auch als Hilfsmittel für die Verdrahtung zu verwenden und wird in dieser Rolle mittels eines geeigneten Kabels mit dem zu verdrahtenden Instrument verbunden. Die Verdrahtungsfolge wird in geeigneter Weise programmiert und so angeordnet, dass der Prüfer nicht weiterschaltet, bis die Anschlüsse der letzten Schaltung korrekt angeschlossen sind. Ein Umschalter gestattet dem Inspektor, den Zusammenbau jederzeit ohne Ausbau zu betreiben und zu prüfen.

EE 96 767 *für weitere Einzelheiten*

Quarzkristall

Standard Telephones & Cables Ltd, Components Group, Footscray, Sidcup, Kent

(Abbildung Seite 542)

Ausser dem breiten Fertigungsprogramm für zweipolige Quarzkristalle in Metallgehäuse hat der STC-Bereich Quarzkristalle Elemente in evakuierten Glaskolben entwickelt, deren Grosserienfertigung nunmehr angelaufen ist.

Es werden drei Ausführungen unter den Typenbezeichnungen 4444, 4445 und 4446 hergestellt, die den Anforderungen der DEF-Pflichtenblätter Ausführungen L, M und N sowie den MIL-C-3098-Ausführungen HC-27/U, 26/U und 29/U entsprechen.

Die Konstruktion im evakuierten Glaskolben hat bessere Alterungseigenschaften sowie höhere Langzeitzuverlässigkeit und gestattet Betrieb in Umgebungstemperaturen bis zu 150° C.

Ausserdem werden engere Frequenztoleranzen möglich, wodurch sich diese Elemente zusammen mit den niedrigen Alterungseigenschaften für den mobilen Sprechfunk mit 12,5 kHz Kanalabstand eignen.

Typ 4444 ist in Frequenzen von 1,6...100 MHz lieferbar, Typen 4445 und 4446 überstreichen das Band 5...100 MHz.

EE 96 768 *für weitere Einzelheiten*

Drucktastengesteuerte Tastatur

Elliott-Automation Ltd, 167 Great Portland Street, London, W.1

(Abbildung Seite 542)

Ein neues System für die Erzeugung drucktastengesteuerter Tastaturen fast jeder Anordnung und Ausdehnung für das Schalten elektrischer und elektronischer Systeme wurde von Elliott-Automation Ltd entwickelt.

In der neuen Tastatur bildet ein Präzisionsformgussstück den Träger für vier Reihen von je 13 Tasten und Mikroschaltern. Das Gussstück kann ohne Schwierigkeiten entweder entlang den vier Reihen oder quer zu ihnen so zugeschnitten werden, dass sich Anordnungen in allen Mustern und Zahlen unterbringen lassen.

Jeder Tastenzusammenbau wird einfach auf einen in das Gussstück eingeklebten Stiel aufgeschoben und mit Stiften befestigt. Bei Betätigung der Taste fühlt man einen deutlichen "Klick," der durch einen kleinen Dauermagneten hervorgerufen wird, der erst dem Fingerdruck auf die Taste entgegenwirkt. Der Mechanismus wurde erfolgreich über einen grossen Temperaturbereich und unter schweren Vibrationen geprüft.

Sowohl das Gussstück wie die Tasten lassen sich schnell und billig entsprechend Kundenanforderungen herstellen. Farbe, Beschriftung und Grösse der Plastik-Tastenkappen sind genauso wie Typ und Qualität der Miniaturmikroschalter und eine Anzahl anderer Konstruktionsmerkmale Kundenwünschen vorbehalten.

Die Tastengruppen sind für Einbau in die Bedienfelder von Konsolen oder Gehäusen ausgelegt.

Weitere Konstruktionsmerkmale sind u.a. eine einfache mechanische Verriegelung, um zu verhindern, dass mehr als eine Taste in einer Gruppe auf einem Gussstück gedrückt wird, und begrenzte Schnittstellenlogik, die auf Schaltungsplatten in den Schaltergussstücken untergebracht werden kann.

EE 96 769 *für weitere Einzelheiten*

Automatisches Wegregistriergerät

Fairey Engineering Ltd, Stockport, Cheshire

(Abbildung Seite 542)

Diese als D.A.T.A.R. (Digital Auto Transducer and Recorder) bezeichnete Ausrüstung misst lineare Verschiebungen mit 0,0025 mm Auflösung und druckt die Ergebnisse automatisch entsprechend einem Zeitprogramm oder ferngesteuert.

Das kompakte und leicht einstellbare D.A.T.A.R.-System misst positive und negative Bewegung mit Hilfe eines phasenempfindlichen Impulszählsystems. Maximal können 25 Auflösungseinheiten je Sekunde und 4 Messungen je Sekunde ausgedruckt werden. Zusätzlich zum automatischen Ausdrucken kann man jederzeit einen manuell gesteuerten Test durchführen.

Das D.A.T.A.R.-System besteht aus zwei Grundgeräten, und zwar dem Messwertwandler (mit Auflösungsvermögen von 0,0025 mm, 0,025 mm und 0,25 mm und Wegbereichen bis zu 12,7 mm, 50,8 mm bzw. 12 mm lieferbar) und dem Registriergerät, das Impulsverarbeitung, Zählen, Zeitmessen und Drucken ausführt.

Eine Messunsicherheit von ±2 Auflösungseinheiten oder ±1 Prozent mit Gültigkeit des grösseren Wertes ist garantiert. Für die Einstellung der Ausrüstung, Einleitung der automatischen Registrierung der Testresultate in Digitalform auf Abreissstreifen als Dauerprotokoll gibt es vier Bedienelemente.

Der Messwertwandler ist etwas grösser als eine Messuhr, in der Anordnung der letzteren jedoch ähnlich, wiegt 675 g und hat einen Betriebstemperaturbereich von 0...40° C. Die Abmessungen des Registriergerätes sind 330 × 267 × 229 mm, das Gewicht 13 kg. Das ganze System ist für Zuverlässigkeit und niedrigen Stromverbrauch (Wechselstromnetzanschluss) konstruiert.

EE 96 770 *für weitere Einzelheiten*

Direktschreiber

Southern Instruments Ltd, Camberley, Surrey

(Abbildung Seite 543)

Der neue preisgünstige Direktschreiber MRE.141 für Maschinenbau und Elektrotechnik kann besonders für Hersteller von Turbinen, Umlauf- und Kolbenmaschinen aller Art sehr nützlich sein. Er registriert gleichzeitig mit vier Informationskanälen und zusätzlich Zeit, Vorgang und Vorgangsdauer auf 70 mm oder 120 mm breitem Film oder Papier. Für alle Betriebsarten eingebaute Logik verhindert Einleitung des Vorgangs oder zufällige Belichtung. Es können einmalige oder wiederholte Registrierungen vorgenommen werden; Eichsignale kann man vor, während oder nach dem Vorgang registrieren. Triggerung erfolgt durch Eingangsspannungspegel, Vorwahlzeitverzögerung oder Zählen von Perioden des Eingangssignals. Das Gerät lässt sich für Überwachung in diskreten Zeitintervallen von einer Millisekunde bis zu zwei Tagen programmieren. Acht Funktionen können gleichzeitig in X-Y-Darstellung zusammen mit Zeit- und Vorgangmarken registriert werden; für Einmal-Registrierung, kontinuierlichen Vorschub und Trommelbetrieb lassen sich Registrierkameras vorsehen. Die erreichbare Genauigkeit ist 1 Prozent, und man kann von 0...250 kHz registrieren. Platz für Nachrüsten mit einer Kombination von 2-MHz-FM- und Gleichstromvorverstärkern sowie weiteren Hilfsfeldern ist vorhanden.

EE 96 771 *für weitere Einzelheiten*

X-Y-Schreiber

Vertrieb: Aveley Electric Ltd, South Ockenden, Essex

(Abbildung Seite 543)

Hohe Registriergeschwindigkeit von 1,25 m/s und Empfindlichkeit von 0,1 mV/cm bis zu 10 V/cm sind Eigenschaften des X-Y-Schreibers ZSK von Rohde & Schwarz. Das Gerät arbeitet nach dem Prinzip des automatischen Abgleichkompensators und schreibt linear in beiden Richtungen auf DIN A3.

Die gemessenen Werte können unterschiedliche Potentialpegel haben, da die Eingänge für X- und Y-Spannungen erdfrei und galvanisch voneinander getrennt sind. Ohne weitere Zusätze sind mathematische Operationen wie Multiplizieren und Quadrieren möglich. Gestelleinbaueinheiten für die lineare Umsetzung von Wechselspannungen oder Frequenzen in Gleichspannungen erhöhen die Vielseitigkeit des Schreibers.

EE 96 772 *für weitere Einzelheiten*

Stromwandler

Atkins, Robertson & Whiteford Ltd, Industrial Estate, Thornliebank, Glasgow

(Abbildung Seite 543)

Atkins, Robertson & Whiteford Ltd stellt eine Reihe von Ringstromwandlern her, die für wirtschaftliches Messen mit industriellen Strommessern (Norm BSS 89 (1954)) konstruiert sind. Diese Wandler haben einen Preisvorteil gegenüber denen der Klasse C nach BSS 3938 (1965), der oft für Strommessungen benutzte Typ.

Verwendung genormter Durchmesser und Herabsetzung der Belastbarkeit auf einen für moderne Schalttafelinstrumente geeigneten Pegel führten zu Einsparungen.

Über den ganzen Betriebsbereich ist der grösste Übersetzungsfehler 1 Prozent des Skalenendwertes, und obwohl die Wandler nicht für Wattmessungen gedacht sind, überschreitet die Phasendifferenz 100 Minuten nicht.

Sekundäre Anschlussklemmen sind vorhanden, und ein Montagewinkel kann gegen Aufschlag vorgesehen werden.

Die Wandler haben je nach Übersetzungsverhältnis Nennbelastbarkeiten zwischen 1 VA und 15 VA.

EE 96 773 *für weitere Einzelheiten*

Koaxialabschwächer

Greenpar Engineering Ltd, Station Works, Harlow, Essex

(Abbildung Seite 543)

Der Bereich Electronics der Greenpar Engineering Ltd hat eine neue Serie von konkurrenzfähigen Koaxial-Präzisionsabschwächern herausgebracht. Sie wurden besonders für Genaugkeit und Konstanz über einen breiten Temperaturbereich entwickelt und kommen in den Standardwerten 1, 2, 3, 6, 10, 12, 14 und 20 dB. Alle sind für die drei Steckverbindungen der Typen N, BNC und TNC lieferbar. Die Körper und Impedanzanpassungshülsen bestehen aus versilbertem Messing und umhüllen TEE-Dünnfilmnetzwerke, in PTFE-Isolation eingebettete Stifte und Scheiben mit vergoldeten Kontakten, sowie klimatische Dichtungen aus Silikonkautschuk.

Die Impedanz dieser Abschwächer ist 50 Ω oder 75 Ω, der Frequenzbereich 0...4 GHz. Das Stehwellenverhältnis liegt unter 1,05 bei 1 GHz und unter 1,20 bei 4 GHz. Die maximale Dauerverlustleistung ist 1 W.

EE 96 774 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Elektronisch gesteuertes Anlassen von Flugzeug-Gasturbinen von K. F. Bacon

Zusammenfassung des Beitrages auf Seite 498-506

In modernen Flugzeugen ist es unerlässlich, die Besatzung von Routineaufgaben zu entlasten. Der Beitrag beschreibt, wie eine dieser Aufgaben—das Anlassen von Triebwerken—mittels eines elektronischen Steuergerätes automatisch durchgeführt werden kann. Die beim Entwurf eines solchen Gerätes auftretenden Probleme werden eingehend beschrieben und der Weg zu ihrer Lösung erläutert. Die Schaltungen des endgültigen Gerätes und seine Arbeitsweise im Flugzeugsystem werden beschrieben und die Anwendungsmöglichkeiten beim Anlassen von Turbinen umrissen.

Ein wirtschaftlicher Ringkommutator für das Strahlumschalten in Oszillografen von A. F. Lewis

Zusammenfassung des Beitrages auf Seite 507-509

Acht, mittels eines einfachen Vierweg-Ringmultivibrators gesteuerte Gegentakt-Transistorschalter ermöglichen gleichzeitige Darstellung von acht Signalen auf dem Schirm eines Zweistrahloszillografen.

Bistabile Elemente für Folgeschaltungen von J. J. Sparkes

Zusammenfassung des Beitrages auf Seite 510-515

In einer Besprechung des Problems der Vermeidung zweideutiger Steuerzustände für R-S-Flip-flops werden alle möglichen und sinnvollen Lösungen kurz in bezug auf ihren unkomplizierten Aufbau und ihre Vielseitigkeit in Folgeschaltungen verglichen. Es wird gezeigt, dass selbst für ganz einfache Folgeschaltungen der ausschliessliche ODER-Flip-flop (oder der E-Flip-flop) besser sein kann als der J-K-Flip-flop, da er nicht nur zu einer vergleichbaren Vereinfachung von Kombinationsschaltungen führt, sondern auch einfacher zu konstruieren ist, sowohl für synchrones wie asynchrones Arbeiten gebaut werden kann und für positive wie auch negative Logik identisch ist.

Messen der Übertragungsfunktion mit schnellen Impulsen von R. C. French

Zusammenfassung des Beitrages auf Seite 516-519

Die Zeitwellenform am Ausgang einer elektrischen Schaltung ist eindeutig durch die Zeitwellenform am Schaltungseingang und den Frequenzgang der Verstärkung und Phase, d.h. ihre Übertragungsfunktion, bestimmt. Umgekehrt kann die Übertragungsfunktion mittels bekannter Ein- und Ausgangswellenformen berechnet werden. In einem entwickelten System zum Messen der Schaltungsübertragungsfunktionen werden die Ein- und Ausgangswellenformen zeitmässig abgetastet und automatisch auf Lochstreifen registriert. Die Übertragungsfunktion wird mit Hilfe des Lochstreifens in einer Datenverarbeitungsanlage berechnet und die Ergebnisse entweder auf einer Schreibmaschine ausgeschrieben oder mittels eines schnellen Kurvenschreibers als Verstärkung, Phase und Gruppenlaufzeit über der Frequenz aufgetragen. Die Messunsicherheit ist eine Funktion des gemessenen Übertragungsfunktionstypes; in beschriebenen Laborversuchen war die erzielte höchste Genauigkeit ±0,3 dB für Verstärkung und ±2° für Phase mit einem dynamischen Bereich von 10 dB. Das System ist für das Band bis zu 100 MHz geeignet, es sind aber viel grössere Bandbreiten möglich.

Entwurf und Gruppenlaufzeitkorrektur eines Tiefpasses für Videoschaltungen von J. W. Head

Zusammenfassung des Beitrages auf Seite 520-524

Der Beitrag beschreibt eingehend eine bereits früher veröffentlichte Entwurfsmethode (einschliesslich Gruppenlaufzeitkorrektur) für einen Tiefpass, dessen Frequenzgang im Durchlassbereich (bis zu 5, 5 MHz) innerhalb 0, 5 Prozent konstant ist und der bei 8,25 MHz eine theoretisch unendliche, über 8,25 MHz nicht weniger als 46 dB Dämpfung hat. Angegebene praktische Ergebnisse beim Entwurf eines Filters, das wesentlich kleiner war als soweit mit gleichen Eigenschaften zur Verfügung stehende, wurden durch Anwendung dieser Methode erreicht.

Frequenztransformation für Wechselstromnetzwerke von J. G. Advani und M. N. Neelkantan

Zusammenfassung des Beitrages auf Seite 525-527

Für die von Prabhakar vorgeschlagene Transformation $p = \frac{2}{3} \cdot \frac{s(s^2+1)}{(s^2-0{,}25)}$ *für einen Wechselstromintegrator wird innerhalb plus/minus 50 Prozent der Trägerfrequenz ein maximaler Linearitäts- und Symmetriefehler von 1/2 Prozent beansprucht. Es wird gezeigt, dass dieser Fehler in diesem Frequenzbereich etwa 8 Prozent beträgt. In diesem Beitrag wird eine neue Transformation für Wechselstromintegratoren vorgeführt, die in diesem Frequenzbereich einen Fehler von etwa 3 Prozent gibt.*

Eine andere Transformation zum Erlangen von Übertragungsfunktionen kompensierender Wechselstromnetzwerke von Übertragungsfunktionen kompensierender Gleichstromnetzwerke wird auch vorgeschlagen. Diese Transformation gibt innerhalb des Frequenzbereiches von ±10 Prozent der Trägerfrequenz 1/2 Prozent Fehler in Linearität und Symmetrie.

Netzwerke eines Wechselstromintegrators und eines Wechselstromphasenschiebers mit R, C und aktiven Elementen werden abgeleitet.

Eine grafische Entwurfsmethode für Widerstandswandler-Grossbereichbrücken von H. S. Sear

Zusammenfassung des Beitrages auf Seite 528-530

Eine für den Entwurf von Wheatstone-Brücken geeignete grafische Methode, die durch Widerstandsmesswandler wie z.B. Thermistoren bei Einsatz über einen breiten Bereich hervorgerufene Nichtlinearität weitgehend korrigiert, wird beschrieben. Das Verfahren bestimmt den Restfehlerwert, sein Vorzeichen und an welcher Stelle im Messwandlerbereich der Fehler auftreten wird. Nach diesem Verfahren wird die Kennlinie des Messwandlers mit einer Kurve des Kennlinienfeldes der Brücke zusammengepasst. Die Anwendung dieser Entwurfsmethode auf Thermistor-Thermometer grossen Messumfangs mit linearem Ausgang wird gegeben.

Regelfaktor für Zenerdioden-Bezugsquellen mit Temperaturausgleich von H. Banasiewicz

Zusammenfassung des Beitrages auf Seite 530-531

Temperaturkompensierte Zenerdioden-Bezugsquellen mit gleichem Temperaturkoeffizienten unterscheiden sich oft hinsichtlich eines oder mehrerer der drei Grundparameter Zenerspannung, Strom und Widerstandskomponente der Elektrodenimpedanz. Es wird gezeigt, wie diese Werte in Kombination einen direkten Hinweis auf die Regelfähigkeit einer Einrichtung gegen Änderung des Vorspannungsstroms geben können.

Die Zukunft des grafischen Fernmeldewesens von T. M. C. Lance

Zusammenfassung des Beitrages auf Seite 532-535

Die Übertragungsmethoden für grafische Information werden in diesem Beitrag zusammengefasst und einige der zukünftigen Anwendungsmöglichkeiten dieser Technik besprochen.