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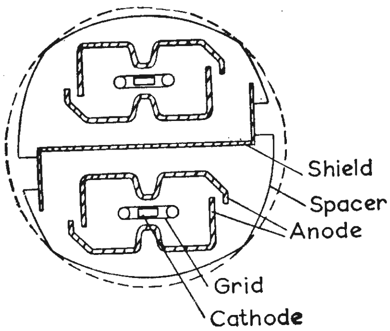
PCC189 - CASCODE R.F. AMPLIFIER VALVE

THE MULLARD PCC189 double-triode will be encountered in a number of present-day v.h.f. television tuner units. In these units, the triode sections of the valve are connected as a cascode r.f. amplifier; that is, a grounded-cathode section is coupled in series with a grounded-grid section.

The new valve is particularly suited for this application because it is characterised by low noise, good stability and efficient screening. These properties result from the special construction illustrated in the diagram. Frame grids are used in both sections. These enable small spacings to be used between the electrodes, so that the time taken for electrons to travel from the cathode to the anode is very short, and good high-frequency performance is consequently ensured. The anodes of both sections are fashioned so that high values of mutual conductance are obtained without incurring increased values of anode-to-grid capacitance. Each section of the valve possesses a variable-mu characteristic. Good cross-modulation performance can thus be achieved and interference between the sound and vision carriers eliminated.



PCF806-OSCILLATOR MIXER FOR V.H.F.

The new Mullard PCF806 is a triode-pentode intended for use in the mixer stage of v.h.f. tuners, and is to be found in many of the latest television receivers. The triode section is designed to operate as an oscillator and the pentode section as a mixer.

The design of the triode section is similar to that of earlier oscillator valves, but a slightly higher value (5·5mA/V) of mutual conductance has been achieved. This, together with a low amplification factor (17), results in a high effective slope. Consequently excellent oscillator operation—especially at the higher frequencies—can be achieved with the PCF806.

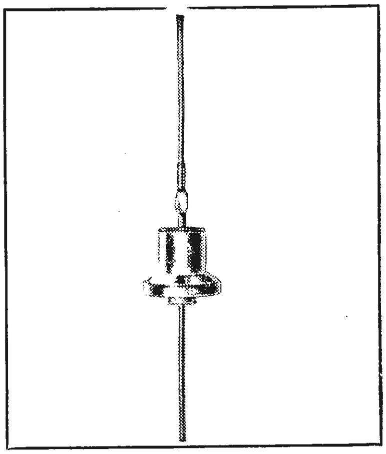
WHAT'S NEW IN THE NEW SETS

These articles describe the latest Mullard developments for entertainment equipment

The frame grid used in the pentode section results in a mutual conductance of 12mA/V at the working point, and a conversion conductance of 4·5mA/V. This value of conversion conductance is about double that obtainable with mixer valves not using frame grids, and the increase has been achieved without any reduction in input impedance. Hence the conversion gain of the PCF806 is also about double that of the earlier valves. Use of the PCF806 means that the design of the subsequent i.f. amplifier stages can be simplified and the standard of receiver performance considerably improved.

NEW ENVELOPE FOR BY100

In the past year, many television receivers have appeared which incorporate the Mullard



BY100 silicon mains rectifier. Mullard have now designed a new envelope for this device which is more convenient for wiring into receivers, and it is intended that this new version will eventually supersede the old version completely.

The difference in envelope is that the threaded stud surrounding the output connecting leads is now absent so that the new rectifier, although a straightforward replacement, is considerably shorter than the old. Electrically however, the studless version is identical with the earlier version. It has a maximum recurrent peak inverse voltage rating of 800V, at which the reverse current is 10μA. It will deliver an average forward current of 550mA at 50°C, while its maximum peak current rating is 5A. At this peak, the forward voltage drop is 1·5V.

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Batteries off the Ground

FIXED installations like domestic television receivers and digital computers are fortunate in being able to draw their power supplies from the electricity grid, but the increasing growth of electronic aids to air transport, guided missiles and space exploration has resulted in an intensification of research into compact, self-contained sources of power of high efficiency and reliability.

Progress from the damp paper of Volta's pile through the open glass jars and messy solutions of Daniell, Grove, Bunsen and Leclanché to the modern hermetically sealed nickel-cadmium cell, which consumes its own reaction by-products, has been continuous but intermittent. At the beginning of the 19th century Faraday laid the foundations of electro-chemistry, and before the end of this period advances in physical chemistry had firmly established the thermodynamic basis of battery operation. Simultaneously the *ad hoc* experimenters and neo-alchemists were trying every probable and even unlikely combination of electrodes and electrolytes in the hope of finding a battery bonanza. A search of the literature rarely fails to uncover a precedent for any "new" development. It is the application of new materials and, on the scientific side, the appreciation of the electronic basis of chemical reactions that have produced the recent advances of battery technology. Micro-porous plastic separators have taken the place of cedarwood in car batteries and sintered nickel electrodes are largely responsible for the success of sealed alkaline batteries. Some of the rare elements, e.g. indium, have fallen in price as the result of the demands of semiconductor manufacturers, and can now be considered for use in batteries—at least for those required in small quantities for special applications.

Further progress has been dictated by the change of environment in which batteries are called upon to operate. The neglected Leclanché jars on a shelf under the stairs were reliable enough and continued to ring the door bell for years until the electrolyte ran dry, but an aircraft battery must not froth over on charge while waiting to take off on a sub-tropical airfield, or freeze into hibernation half an hour later at 30,000ft. The solution to this problem does not come from one direction only. Apart from the quality of the battery itself there is the possibility of automatic control of charging rate, and also the conservation of heat in the thermal mass by lagging. This will extend by many hours the flight time before battery efficiency begins to deteriorate. The same problem, with opposite sign, is present in low-level attack aircraft in which battery compartment temperature may rise to as much as 55°C, but here the temperature difference is smaller and with normal batteries no trouble is likely to occur within the flight duration time.

The missile battery has a short life, if not a merry one: it must stand dry for months in the restricted space allotted to it, and after the electrolyte has been injected it must withstand a far more severe buffeting than any aircraft battery, and a change from ground level to space environment in a mattter of minutes. The satellite battery, on the other hand, must go on working unattended for years if communication satellite systems using present known techniques are to be regarded as an economic proposition. Float batteries are essential to a sustained service with relatively low orbits, when for part of the time the solar-cell primary sources of energy are in the earth's shadow. The sealed nickel-cadmium alkaline cell now provides a reliable answer, and it is the solar cell which is proving to be the weaker link (one up for the wet chemists, though the solid-state boys may yet have the last word).

It seems to us that the high-energy particles which cause the damage in solar cells might be put to work in *nuclear* batteries, supplemented if necessary by a gram or two of radioactive material. If the future of space propulsion lies in nuclear power, so may the sources of auxiliary electrical power.

Forty Years Ago

IT was on 14th November, 1922, that the Marconi Company formally handed over station 2LO to the newly formed British Broadcasting Company and so initiated a service of news, entertainment and instruction through the medium of sound which has continued, and we hope will always continue, to satisfy all those who have ears to hear. Although television makes the more popular appeal these days, let us never forget that there are many, apart from those who are unfortunately blind, for whom the things of good report in life—poetry, plays and music—enter the mind's eye through the ear.

TV TAPE TIME STABILITY

By AUBREY HARRIS,* A.M.I.E.E., A.M.Brit.I.R.E.

WITH an electronic signal recording system—for example, magnetic tape, magnetic disc, direct disc or thermoplastic film—an electrical waveform is transferred to a physical medium. The accuracy of the positioning of the recorded signal on the medium is in many cases of great importance and must be carefully controlled, both in the recording process and also during reproduction. Unless certain precautions are taken, there will be very slight discrepancies between the actual and correct recording-medium/head relationship.

If the amplitude of the error is large enough, the effect becomes objectionable and gives rise to the "wow" and "flutter" effects so well known on audio recordings.

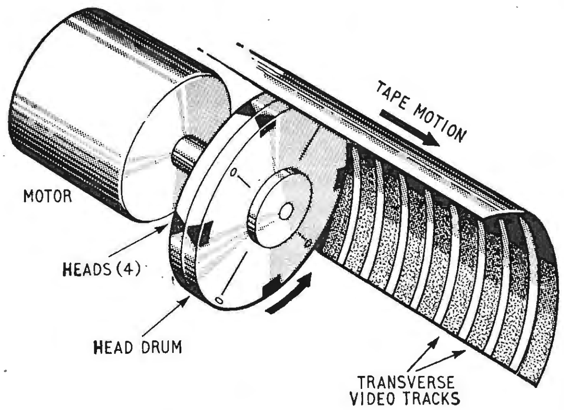


Fig. 1. Principle of rotary head recording. The upper and lower edges of the video tracks are later erased for recording of audio tracks.

In the past five or six years, much has been made of tape recording for broadcast television applications, using the rotary head technique. With this type of recorder the waveform is laid down *across* the tape by a set of four miniature heads mounted on the periphery of a two-inch diameter drum (Fig. 1). The drum is rotated steadily at 15,000 r.p.m. (250 r.p.s.) while the two-inch wide tape is advanced past the rotating heads at the relatively low speed of 15 in/sec. The recorded video pattern is thus in a series of bands, each band containing about 16 television lines.

The amount of time-modulation on the output signal of a television recorder must be reduced to the absolute minimum, in order to keep picture geometry as accurate as possible and also so that the output waveform is absolutely synchronous with, and is timed correctly to the other sources of programme material in the studio centre—for example, the cameras, film reproducers, caption scanners, etc.

It will be appreciated that the timing accuracy of the uncorrected waveform is dependent upon the rotational stability and angular positioning of the rotating head drum. An angular displacement of 5 minutes of arc represents a time period of approximately 1 microsecond. This calls for a high degree of precision in the mechanical components associated with the head drum, and considerable attention must be paid to the continuity and constancy of phase of the drum-motor drive-signal.

Types and Causes of Timing Errors

Many factors contribute to the output waveform timing errors in a television recorder; among them are:

(a) Alignment discrepancies of the head assembly.
(b) Variations of tape dimensions due to temperature or humidity changes.
(c) Variations in motor loading caused by bearings and slip rings.
(d) Non-symmetrical motor fields.
(e) Variations in head-to-tape friction.
(f) Modulation of the signal waveform by the recording carrier and/or by electrical noise.

The first two factors above occur due to mechanical alignment or dimensional differences between the recording and reproduction processes, and cause distortion in the regular geometry of the picture.

The other factors produce repetitive or random errors and cause displacement of one or more lines in the picture, or are seen as slight oscillatory or random horizontal motion of the reproduced image on the television screen. The terms "jitter" and "drift" are used to describe these errors. Jitter is applied to errors occurring at a frequency of greater than about 1 c/s; drift indicates errors at frequencies below this figure.

Subjectively, jitter is the more perceptible of the

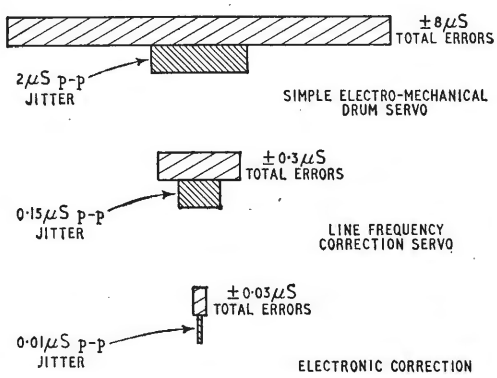


Fig. 2. Approximate limits of various methods of time error correction.

*Ampex (Great Britain) Ltd.

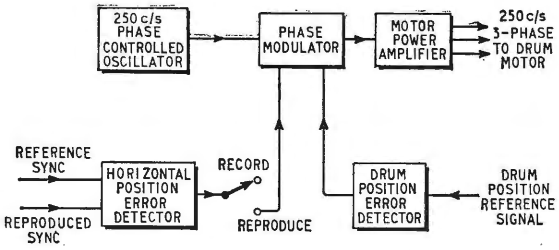


Fig. 3. Simplified block diagram of television signal synchronizer.

two types of error, for although the magnitude of the displacement is small, the frequency at which it often occurs (5 to 10 c/s) is such that the effect is particularly noticeable to the eye. The amplitude of the drift error (5 to 10 microseconds, uncorrected) is many times that of the jitter error, but as its period is very low, it is seldom discernible by observation of the picture; it is most troublesome when an attempt is made to cut or mix from a truly synchronous signal source to the picture from the recorder.

It is perhaps useful at this point to indicate the magnitude of timing errors with other forms of tape recording. Figures for typical audio recorders may be 0·1% to 0·2% r.m.s.; instrumentation tape recorders 0·2% to 0·3% peak-to-peak; digital tape handlers between 2% and 4% peak-to-peak. A typical television recorder with electromechanical servo control may have a jitter on the output signal, relative to a stable reference, of between 0·2 and 0·3 microseconds peak-to-peak; this corresponds approximately to $\pm$0·15% to $\pm$0·2% timing error in the waveform (on the 625-line system). The magnitude of the drift error (after correction) is similar.

Error Correction

There are two basic methods used for minimizing and eliminating timing errors, namely:

(a) Electromechanical correction and
(b) Electronic correction.

The first type of correction operates by the modification of the phase of the signal driving the head drum motor. This is a continuous process and correction down to the order of a few tenths of a microsecond is possible. However, by the addition of *electronic* correction, the time stability is increased to the region of between 10 and 20 nanoseconds with respect to a stable reference source. This finer correction is obtained purely electronically by passing the reproduced signal through a variable delay-line. A portion of a reproduced waveform which is detected to be in advance of its correct time position is made to suffer a delay. The amount of delay applied is exactly equal to the advance error detected. Conversely, the delay is reduced for a retarded portion of the waveform. The timing accuracy of an overall system with both electromechanical and electronic correction is some ten times better than with electromechanical correction only.

The degree of control of timing errors by electromechanical and electronic correction methods is shown in Fig. 2.

The error apparent in a reproduced, uncorrected television waveform contains timing discrepancies introduced both during recording and also those produced on reproduction. The method of correction is based on the determination of the timing error between the reproduced signal and a reference (usually the studio synchronizing generator). A correcting voltage is then produced which is a function of this timing error and is used after suitable processing to correct the error.

For practical reasons, it is not possible to sample errors at a greater rate than 250 c/s during the recording process.

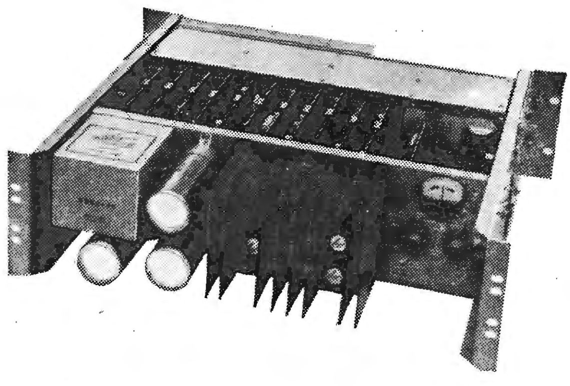
During reproduction, the sampling rate is at the line repetition frequency; it is possible also to reduce timing errors which were introduced in the recorded pattern. The higher rate of sampling is made by the comparison of the line synchronizing pulses, derived from the reproduced video signal, with similar pulses produced by a reference source.

The correction devices make use of both closed-loop and open-loop principles. The closed-loop equipment provides electromechanical correction by controlling the head drum velocity and its position. Open-loop correction is effected by electronic variable delay of the reproduced video signal.

Television Signal Synchronizer

Electromechanical correction of timing errors is made by the television signal synchronizer. This unit generates a 250 c/s signal which is used to feed the input of a 160-watt power amplifier. The 3-phase output of this amplifier drives a hysteresis synchronous motor to which the rotary head drum is directly coupled. Thus an alteration of the phase of the 250 c/s driving signal modifies the angular position of the head drum. A simplified block diagram of the unit is shown in Fig. 3.

Due to inherent mechanical characteristics, the motor and head drum have a natural resonance at a frequency in the region of 5 to 10 c/s. During recording, this resonance is damped by feeding back to a phase modulator, in the drum drive path, a signal derived from a photo-electric sensing device on the drum motor. This signal is 250 c/s and its phase is exactly related to the angular position of the drum motor. The gain in the feedback path is



The time element compensator. Solid state devices are used throughout, mounted on etched board plug-in cards. The delay lines are enlosed in the casting at the rear of the unit. The low-voltage power supplies are mounted on the front of the chassis.

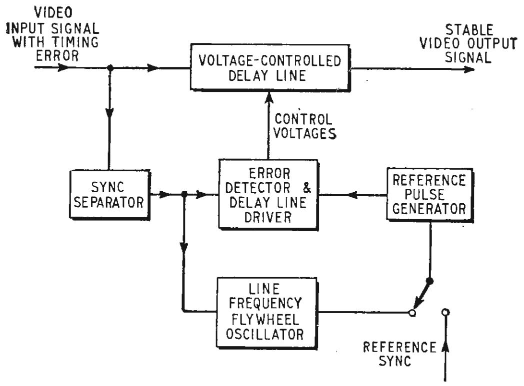


Fig. 4. Simplified block diagram of time element compensator.

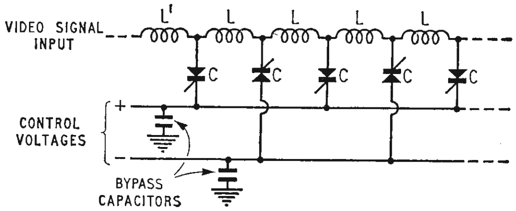


Fig. 5. Section of voltage-controlled delay line. L and L' are inductive circuit elements and C variable-capacitance semiconductor diodes.

adjusted to ensure critical damping of the motor.

The same damping arrangement is used on reproduction, but additionally the phase modulator is fed with a correction voltage which is a function of the horizontal jitter and drift error obtained from the output of a comparator which determines the remaining degree of error by sampling at the line repetition rate.

Time Element Compensator

Further, finer correction of timing errors in the reproduced output signal of the recorder is provided by this unit which is totally electronic in operation. The principle involved is as follows: the video signal at the output of the demodulator is passed through a lumped-constant delay line. The capacitive elements of the delay line are variable-capacitance semiconductor diodes; by varying the potentials applied to these diodes, their capacitance may be changed. This alters the constant of the line so that the total delay of the line is changed, with a change of the applied control voltage. A block diagram is shown in Fig. 4.

The video signal containing timing errors, enters the time element compensator and is fed through the delay line, and also to a sync separator. The output of the sync separator (containing sync signals having the same timing error as the video signal) is one input of a time comparator circuit, which is the error detector. The other input of this is from a reference pulse generator, which may be switched to produce pulses from either an external reference source or an internal line frequency flywheel oscillator. The output of the comparator is a varying voltage, the polarity and amplitude of which is a measure of the departure from consistency of timing of the reproduced waveform. After being passed through phase splitting driving stages, the correction voltages are applied to the voltage-controlled delay line.

When the internal reference is in use, individual reproduced synchronizing pulses are compared with the *average* rate of the reproduced synchronizing pulses, integrated over a period of 2 to 3 milliseconds. Under this condition of operation, the time element compensator will only correct for geometrical errors in the picture.

With the external reference source in use, all repetitive, random and geometric errors up to a maximum of 1 μsec are corrected to less than $\pm 0{\cdot}03$ μsec.

Voltage-controlled Delay Line

The complete voltage-controlled delay line assembly consists of four individual but similar lines, each made up of 11 sections. A diagram of a part of one line is shown in Fig. 5. Anti-phase correction voltages are applied as reverse bias to each set of diodes, thereby changing their capacitance and consequently the delay of the lines. The nominal impedance of the voltage-controlled delay line is 500 ohms, and the signal level at the input is approximately 0·25 V peak-to-peak.

The four delay lines are connected in series in two groups, known as the 'A' variable delay and the 'B' variable delay. (Fig. 6.) The total static delay through each group is 1·5 μsec. The output of the error detector is so phased that a greater delay is applied to "early" signals, and the delay line is made to have less delay for a "late" signal.

An interesting feature is that the delay control voltage is applied directly to the A lines and is subject to a 1·5 μsec delay before being applied to the B lines. The additional 1·5 μsec fixed delay at the output of the B driver and phase splitter exactly match the static or mean delay of the whole of the

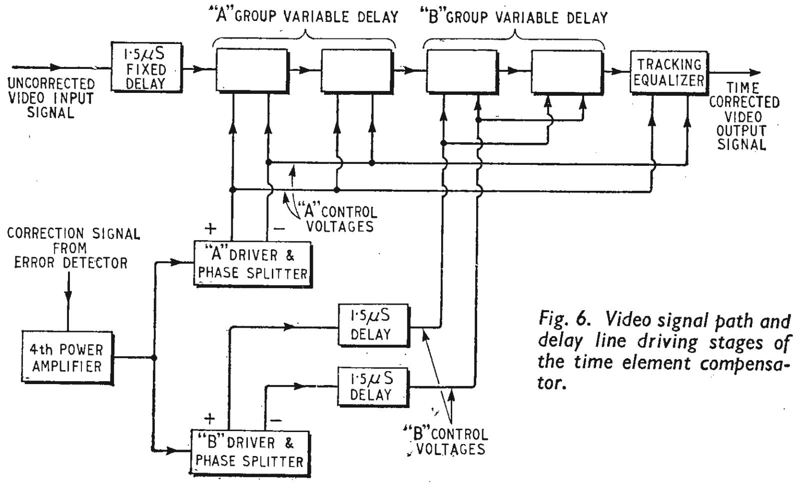


Fig. 6. Video signal path and delay line driving stages of the time element compensator.

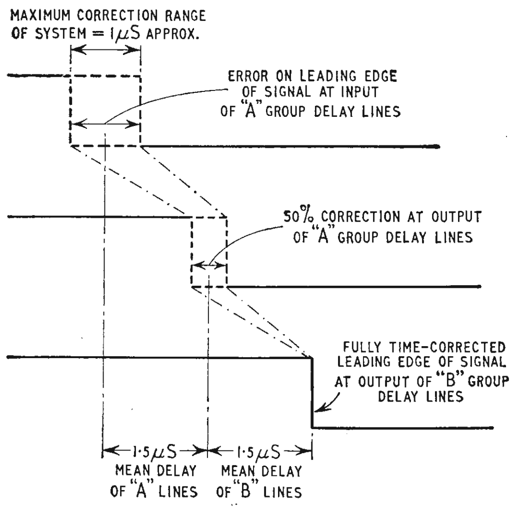


Fig. 7. Two-stage reduction of time error.

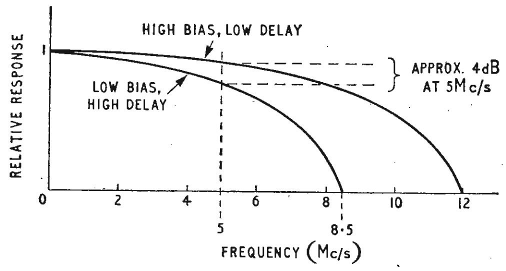


Fig. 8. Frequency characteristics of delay line (before correction by tracking equalizer) showing variation of response under high and low bias conditions. The response is made level between 0 and 5 Mc/s under all bias conditions by the tracking equalizer.

A group variable delay line. In other words, the correction voltage is delayed by a period equal to the transmission time of the video signal through the A delay lines. By this means, correction is made *twice* over the *same* portion of the video waveform, and thus the actual delay operation takes place during only a small time interval of the signal. Half of the maximum total correction range of the system (which is approximately 1 μsec) is carried out in the A lines and the remaining half in the B lines. (Fig. 7.)

A further fixed delay of 1·5 μsec is inserted in the video circuit. This is to compensate for the difference in timing due to the unequal path lengths traversed by the signals fed to the error detector and the delay lines respectively. Over the working range of the voltage-controlled delay line, the frequency response and the impedance of the line vary, due to the change in value of the capacitive elements.

The cut-off frequency under the non-correction state is about 10 Mc/s. During a low-delay/high-bias condition, $f_c = 12$ Mc/s; under high-delay/high-bias conditions, $f_c = 8{\cdot}5$ Mc/s (Fig. 8). At 5 Mc/s the response differs by about 4 dB between the high bias and low bias states. In order to counteract this change in video response, a variable equalizer circuit is incorporated at the output of the voltage-controlled delay line. This is known as a tracking equalizer and is controlled by the bias voltages which are fed to the A delay lines.

The change in impedance due to the variation of the capacitive elements is of the order of ±20%. This variation produces reflections in the delay line; however, these reflections are kept very low and in practice are not objectionable, since a reflection must traverse the line twice before it comes out; careful design of the line ensures the maximum attenuation of such reflections. The reflections over the operating range do not exceed 0·5% in one sense and 0·8% in the other.

The Error Detector and Delay Line Driver

The primary correction voltage for the delay line is generated in the error detector, where the pulses from either the external source (reference sync) or the internal line frequency oscillator are made to produce a sawtooth waveform with a 5 μsec negative slope (Fig. 9). The output of the sync separator (containing the timing error of the input video signal) is gated to allow only the leading edge of each line synchronizing pulse to pass. This is used to trigger a 2 μsec sampling pulse generator, thus the leading edge of the sampling pulse is modulated by the time error of the original video signal.

The time-modulated pulses derived from the video signal are made to sample the steady 5 μsec sawtooth

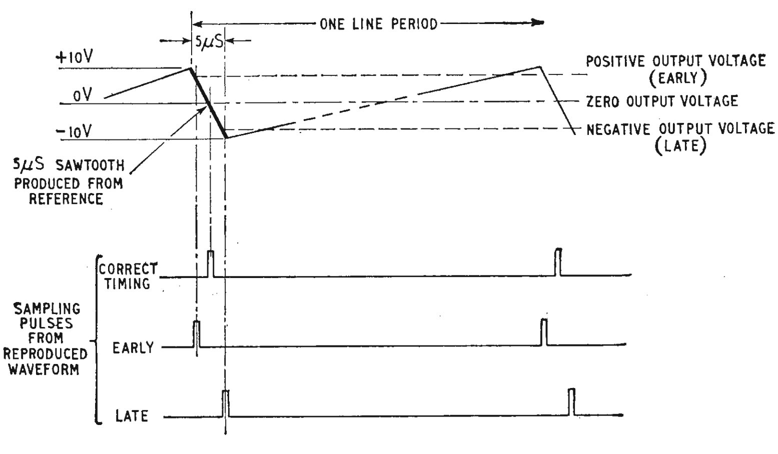


Fig. 9. Principle of error detector showing derivation of control voltages for delay line by timing variations of reproduced waveform. (Timebase not to scale.)

waveform and an output is obtained, the amplitude and phase of which is a function of the timing error. This correction voltage is fed to a driving amplifier before being applied to the delay lines.

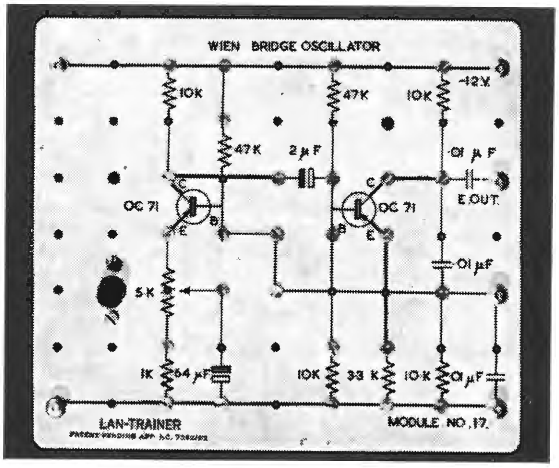
The driving amplifier which feeds the bias voltages to the voltage-controlled delay line has a non-linear characteristic, because the variation of delay does not change linearly with the applied bias voltages. This is due to two factors. First, the variation in delay time of a delay line is proportional to the square root of the capacitance per unit section. Second, the capacitance of the diodes, used in the delay lines, varies inversely (approximately) according to the square root of the applied voltage, over a small range. Thus, the delay time varies inversely as the fourth root (approximately) of the bias voltage on the delay line diodes. In order to obtain a linear time-delay/correction-voltage relationship, the characteristic of the driving amplifier is modified so that its output voltage varies as the fourth power of the input voltage. The circuit used is similar to that commonly employed to obtain non-linear characteristics in gamma control amplifiers. After being processed in this amplifier, the correction signal is fed to two phase-splitting stages. Each phase-splitter provides two correction bias voltages at 0° and 180° phase relative to its input, for driving the A and B group delay lines (see Fig. 6). The correction voltages for the B group delay lines are fed through 1·5 μsec delay lines as noted above. Additionally, the tracking equalizer is fed from the A group phase splitter.

Acknowledgment

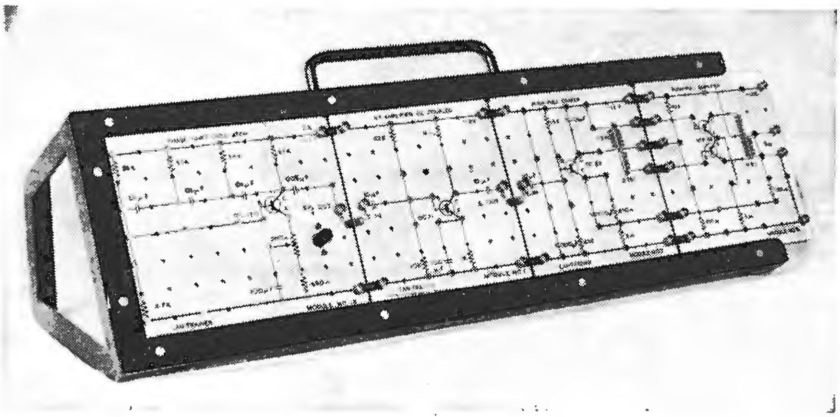
The equipment described in this article was developed in the television engineering laboratories of the Ampex Corporation, Redwood City, California. The author wishes to express his thanks to the designers of the equipment for permission to make use of the information contained herein.

TEACHING PRACTICAL ELECTRONICS

IT has long been the practice, when a "practical" lesson is to be taken, for the teacher and laboratory assistant to make up several sets of equipment, either on breadboards or chassis, and at the end of the period to dismantle the equipment and start again on something else. Even if the students themselves assemble the chassis as part of the lecture, it has been felt that both they and the lecturer could be more profitably occupied, and several alternative schemes have been proposed. The latest and certainly one of the neatest of these is called the Lan-Trainer, and is made by Lan-Elec Ltd., 97, Farnham Road, Slough, Bucks. A set of over 50 sub-circuits (amplifiers, multivibrators, frequency-dividers, etc.) are assembled on perforated panels or modules, the circuit diagram being printed on the front of the panel; all component connections are accessible from the front. The circuit components are standard commercial parts and are fitted with mounting lugs at centres corresponding to the hole grid. The individual modules, which may be modified very easily, can be mounted in racks and assembled to make complete receivers, instruments, etc., the only connections required being the necessary inter-module screw links. Manuals are provided with each modular card, giving a description of circuit operation and experimental information.



Module carrying oscillator unit. Screw heads are component terminations.



Complete oscillator and push-pull amplifier assembled on rack.

Reducing Interference in

Untuned-I.F. Receivers

By M. G. SCROGGIE, B.Sc., M.I.E.E.

IN the June 1956 issue under the title "Unconventional F.M. Receiver" was published a design based on the pulse-counter low-distortion discriminator described earlier (April 1956). A crystal-control modification of the same receiver was described in *Wireless World* in the April 1958 issue, and other versions have appeared elsewhere and apparently have been widely used.

An unusual feature of my design was the choice of an i.f. in the region of 0.15 Mc/s, so that an untuned RC amplifier could be used, dispensing with the need for alignment. The tuning of the single r.f. stage is fixed, and broad enough to cover all three B.B.C. channels in any one location It has been found by some that although there is no perceptible

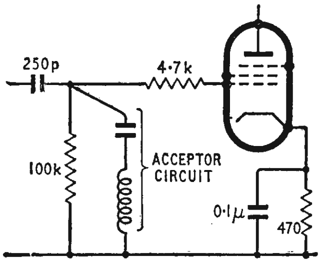


Fig. 1. Acceptor circuit centred on 2.2 Mc/s for reducing i.f. gain at interfering frequencies.

interference when only the two outer channels (Home and Light) are working, with all three there is a rustling or chattering background.

Having been otherwise occupied, I have not had time until recently to look into this complaint. The first thing was to ascertain the cause. The particular receiver available for test was crystal-controlled for the three Wrotham channels (89.1, 91.3 and 93.5 Mc/s), the third harmonic frequency of the oscillator being 0.12 Mc/s less than the wanted channel frequency in each case. The table shows in the first column these third-harmonic frequencies for the three crystals; in the second, the channel carrier frequencies; in the third the resulting difference frequencies supplied to the i.f. amplifier (the sum frequencies are ignored as being too high to be significant); and the last column the difference frequencies that would be created from these by the non-linearity of the limiter, supposing they were still present at appreciable strength there.

Oscillator third harmonic frequency		Carrier frequency	Difference frequency present in i.f.	Difference frequency created by limiter
Tuned to 3	91.18	L 89.1 3 91.3 H 93.5	2.08 0.12 2.32	2.20, 1.96 2.44, 2.20
Tuned to H	93.38	L 89.1 3 91.3 H 93.5	4.28 2.08 0.12	2.20 2.20, 1.96
Tuned to L	88.98	L 89.1 3 91.3 H 93.5	0.12 2.32 4.52	2.44, 2.20 2.20

It is seen that when switched to any of the three channels there are several frequencies at or near 2.2 Mc/s, and frequency modulation of them might well account for the type of interference experienced. While these signal components would be *relatively* weak, it should be noted that the gain of the i.f. amplifier described is of the order of 36 dB at 2 Mc/s. Furthermore, when switched to the 91.3 Mc/s channel there are difference products of 2·08 and 2.32 Mc/s, which, if there was non-linearity early in the i.f. amplifier, would create 0.24 Mc/s, at which the i.f. gain is only about 1 dB below maximum.

If this analysis is correct, an appropriate cure should be one or more acceptor circuits centred on 2.2 Mc/s, as in Fig. 1. A practical difficulty is that the capacitance must be small—less than 10pF—so as not to reduce the i.f. bandwidth appreciably, and then the self-capacitance of the inductor is of comparable magnitude, with undesirable results. In fact, the effect of the first acceptor tried was a substantial increase in interference! This was traced to the self-capacitance of the coil increasing its apparent inductance at the interference frequency, whereby it tuned the parallel stray capacitance of the amplifier

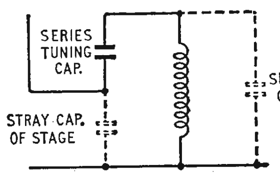


Fig. 2. How the implementing of Fig. 1 can give a contrary result if the inductance of the coil is calculated without allowing for its self-capacitance.

stage as a rejector circuit to that frequency, as shown in Fig. 2.

After compensating for this effect by reducing the inductance to 334μH, the interference was reduced to an acceptable level with 8-9pF series tuning capacitance. The coil was a small low-capacitance multi-layer type.

An alternative cure, theoretically less desirable but audibly acceptable and simpler to apply, was 10 pF across the 0·1 MΩ grid leak of each of the two i.f. stages.

Third International Symposium on Batteries

BOURNEMOUTH, 2nd-4th OCTOBER, 1962

OVER 300 specialists from all parts of the world attended the three-day conference on batteries last month. Held under the auspices of the Inter-departmental Committee on Batteries (a body charged with the duty of co-ordinating research in Government establishments and in industry) this Symposium is regarded as of sufficient importance to be opened by a Government scientist of the highest rank. This year it was Sir Solly Zuckerman, C.B., F.R.S. (Chief Scientific Adviser to the Government on Defence) who underlined the importance of batteries in modern weapon systems and praised the reliability now achieved in the special batteries used in missiles. This was in no small measure due to the international exchange of information, and our American friends with their vast resources had contributed much to battery reliability in space and to the development of solar cells, though their recent disturbance of the Van Allen belt had made the results from these devices somewhat less predictable! While noting that this year the majority of papers still dealt with batteries depending on electro-chemical action he pointed to the rising interest in fuel cells and other forms of energy exchange and hinted that in 1964 the title of the Symposium might be expanded to include these.

Lead-Acid Batteries. In spite of their long establishment as the principal type of storage battery and of the vast amount of research which has gone into their development, lead-acid cells are still capable of providing intriguing problems and fascinating answers. The whole of the first day was devoted to the lead-acid battery and one of the most interesting papers was that by Dr. M. Barak (Chloride Electrical Storage Company) who showed coloured cine-microphotographic films of the surface of a pure lead electrode when taken through a complete cycle from cathodic to anodic conditions. The formation of the dark brown film of lead dioxide when the plate was positive and of the formation of a network of white lead sulphate crystals over the surface on discharge were observed, at least by delegates in the front seats (the projector lamp power was below par) and the evolution of gases was seen to be confined to preferred sites. This not only explains some anomalies in the Tafel relationship (that the hydrogen overpotential varies as the logarithm of the current densitv), since the effective area of the electrode is obviously not its superficial area, but throws some light on the migration of atomic hydrogen over the surface to points of nucleation where combination to molecular hydrogen takes place.

The electron microscope has been called in to investigate electrode structure and Mrs. Jeanne Burbank (U.S. Naval Research Laboratory) gave as an example its successful use in finding the essential difference between two commercially produced lead-cadmium cells for "float" operation. The lead dioxide particles in the unsuccessful battery were found to be granular with smooth contours and rounded edges, resulting in poor cohesion, whereas in the successful battery they were prismatic crystals of the tetragonal βPbO_2 with twinning and many branch arms resulting in a stable interlocked mass. The texture and capacity of the inferior battery plates could be improved by prolonged shallow discharge cycles which produces prismatic $PbSO_4$ spines on the amorphous active material particles.

Normally, both α (orthorhombic) and β (tetragonal) crystal forms of PbO_2 are present together in lead accumulator plates and the proportions and distribution of the two forms are factors determining the performance of the battery. Dr. E. Voss and Dr. J. Freundlich (Accumulatoren Fabrik A.G.) reported detailed investigations of the discharge capacities of lead dioxide electrodes formed exclusively of one or other of these forms. They found that although the instrinsic maximum capacity output per unit areas (true surface) is higher for the α variety, the capacity per unit weight is higher for the β form. The stability and self-discharge properties of α and βPbO_2 in sulphuric acid have been investigated by P. Ruetschi, J. Sklarchuk and R. T. Angstadt (The Electric Storage Battery Co., U.S.A.) and they find that pure αPbO_2 was less stable, in spite of a smaller true surface area.

Alkaline Batteries. The kinetic basis for the successful operation of sealed nickel-cadmium cells was discussed by U. B. Thomas Jr. (Bell Telephone Laboratories) who pointed out that in theory no gas was involved and no net chemical change was involved in a complete charge/discharge cycle of any nickel-cadmium cell, but that on overcharge gas would be evolved continuously. Although theoretically the oxygen and hydrogen would eventually be reduced and oxidized at the negative and positive plates, the process is impracticably slow. In practical sealed cells the first step taken by the designer is to provide excess chargeable material in the negative plate to ensure that under any operating conditions the plate is never fully charged and no free hydrogen is evolved. Overcharging then results only in oxygen evolution at the anode. Rapid diffusion to the cathode is ensured by a short path (probably 10-30 microns) through a film of electrolyte on the porous surface of the negative plate. There is a rise of pressure due to the oxygen evolution, which as Dehmelt and von Dohren (Accumulatoren Fabrik A.G.) have pointed out increases the rate of oxygen reduction (to water). The steady state is dependent on complete sealing and any loss of oxygen or of water would result ultimately in the failure of the cell.

At the previous Symposium, R. T. Doran (Royal Aircraft Establishment) reported work on the addition of the hydroxides of compatible elements such as magnesium and cobalt to the nickel hydroxide electrode. This work has been extended to scandium and indicates that not only can the working potential be raised, but that the electrode capacity improves with age. The high cost of scandium limits practical applications but it is thought that the improved performance may be of value in satellites.

Fuel Cells. The difference between the open-circuit voltage of an oxygen fuel cell and the working voltage on load may be considerable and is due primarily to polarization phenomena at the electrode surfaces. The higher the chemical reactivity of the fuel the better will be the working voltage, and in this respect pure hydrogen gives the best results. However, the future of fuel cells undoubtedly depends on the use of more easily stored liquid fuels and Dr. M. I. Gillibrand and G. R. Lomax (Chloride Electrical Storage Co.) gave theoretical reasons for their choice of hydrazine (N_2H_4) and described a practical cell, comparable in size to a car battery giving a power output of about 60 watts (100 amperes at 0.6 volts). The open-circuit voltage of the cell was 1.0 volt. Fuel cost is, of course, high and in response to a questioner the authors estimated present running costs at £3 per kW-hr.

Solar Cells. Damage from high-energy particles in the Van Allen belt is a subject which is very much to the fore these days, and a review of the facts and the results of some experiments using synchro-cyclotrons and linear accelerators to simulate conditions in space were presented by F. C. Treble (Royal Aircraft Establishment). The principal source of damage is not high energy electrons, which can be effectively stopped by cover plates of glass, quartz or sapphire, but by protons which are capable of producing lattice displacements in silicon at a threshold energy of only 190 electron-volts. These defects act as traps for the recombination of minority current carriers and so reduce the output from the cell. Calculation shows that effective screening from proton bombardment would incur a prohibitive weight penalty. Tests have shown that there is no significant difference between p-on-n and n-on-p silicon junctions as far as damage is concerned, but gallium arsenide, which is less dependent on minority carrier lifetime, shows to advantage. Cells of this type have the additional advantages of higher output voltage and lower sensitivity to temperature rise.

The problems of heat balance, and the reduction of range of temperature variation between light and dark by the regulation of absorption on the front face and re-radiation of heat from the back present an interesting study, as was evident from a paper by H. J. H. Sketch (also of R.A.E.). Taking as examples the design of solar cell systems for a celestial-stabilized astronomical satellite and an earth-stabilized communications satellite, he outlined the factors deciding the geometry of the solar-cell paddles which often seem to assume unnecessarily weird shapes.

TELEVISION IN IRELAND

THE Eire television service (Telefis Eireann) is due to bring into service on November 1st its 625-line transmitter at Kippure, near Dublin. The station will then be operating on both 405 and 625 lines and therefore the first dual standards station in the British Isles. The 405-line transmitter operates in Band III with a 3-Mc/s vision bandwidth (in a 5-Mc/s channel) and a.m. sound and the 625-line station (also in Band III) with a 5.5-Mc/s vision bandwidth (in an 8-Mc/s channel) and f.m. sound.

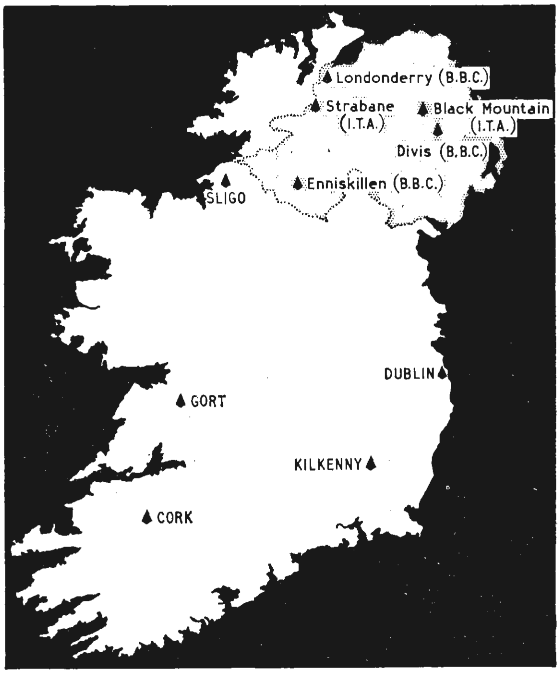
A dual standards station—at Truskmore, Sligo—is also to be introduced shortly. The 405-line transmitter is scheduled to be brought into service later this year and the 625-line transmitter early next year.

It is planned that the remaining three transmitters will operate on 625 lines only. They should be radiating the 625-line transmitter early next year.

In the table below we give operating characteristics of the transmitters at the first five stations. The two 405-line transmitters employ positive vision modulation and the 625-line transmitters negative modulation. The vision e.r.p. of all transmitters is 100 kW.

	Frequencies (Mc/s)		Polar'n
	Vision	Sound	
GORT (Maghera)	53.75	59.75	H
CORK (Mullaghanish)	175.25	181.25	V
DUBLIN (Kippure)*	184.75	181.25	H
KILKENNY (Mt. Leinster) ...	191.25	197.25	V
SLIGO (Truskmore)*	204.75	201.25	V
DUBLIN (Kippure)	207.25	213.25	H
SLIGO (Truskmore)	215.25	221.25	V

*405-line transmitters.



Locations of the Eire stations, together with those of the B.B.C. and I.T.A. (operational and planned) in Northern Ireland. Low-power transmitters at Larne and Newry are also planned by the B.B.C.

Microwave Components

I.E.E. CONFERENCE IN LONDON

ALTHOUGH complementing the recent microwave valve congress (see p. 462 of our last month's issue) the I.E.E. conference was rather smaller—thirty papers were given in three days. These will be printed in full in the conference report series of the Institution: only a selection (mainly illustrating general trends) are discussed in this report.

P-i-n Silicon Diodes for Power Modulation

One very noticeable new trend in this field is the increasing use of p-i-n silicon junction diodes for power modulation (variable attenuation or switching)—five of the papers dealt with various aspects of this topic. Essentially a p-i-n diode consists of a silicon wafer into which charge carriers can be injected. When a forward bias is applied these carriers drift through the crystal lattice colliding with imperfections. The time interval between collisions is much shorter than the period of a microwave r.f. cycle so that a microwave field acts like a d.c. field on the carriers, transferring energy to them. This energy is in turn transferred by collisions to the crystal lattice so that the effective resistance of the diode is then only a few ohms. This resistance can be increased to an effective open circuit (several thousand ohms) by sweeping the carriers out of the silicon by means of a reverse bias. It is this variable resistance which is used to provide power modulation: the p-i-n diode being coupled across the waveguide or coaxial microwave transmission line to give variable power absorption or reflection (as described in papers by Baker and Muskus). The diode also produces a small shunt capacitance across the line, but this can be tuned out by an added inductance. More sophisticated arrangements also described at the conference used two diodes in a 3dB coupler (see papers by Roberts and Robinson and by Anderson) or magic-T (see the paper by Anderson just mentioned and also the paper by Allen). A better overall match can be obtained by absorbing power in the diodes in the off or isolated condition rather than by reflecting it back to the source. As was pointed out by Anderson, wider bandwidths and better temperature compensation can be obtained by balanced arrangements.

The switching speed of such devices is determined by the time taken for the hole carriers to diffuse into or out of the intrinsic region. As was pointed out by Baker and also by Muskus, this time can be kept to a minimum by applying suitable excess bias voltages at the beginning and end of each switching pulse, switching times of the order of 0.5 μsec being readily achievable in this way.

The power-handling capacity of such devices was discussed in the paper by Baker already mentioned. The maximum continuous power which can be handled is limited to the order of 10W per diode by the necessity to keep the silicon temperature below 300°C. This is because above this temperature the intrinsic resistivity of silicon begins to increase and this increases the diode insertion loss (and thus the power absorbed). This leads to a further increase in resistivity, producing a runaway effect and eventual damage to the diode. Even so the power handling capacity is considerably greater than that of the point contact diodes which have also been similarly used for power modulation. The peak pulse power handling capacity is limited to about 500W by the diode reverse breakdown voltage. In addition the small amount of rectified current produced causes self biasing of the diode: this increases the insertion loss and this results in a limiting action at high peak powers.

Advantages of such modulators over other types (e.g., mechanical, ferrite or discharge-tube devices) mentioned in the paper by Muskus already referred to, are their light weight, small size, relative insensitivity to temperature changes, fast switching speed, low switching power requirements, and ruggedness.

Another important recent use of such diodes—to eliminate the need for a keep-alive discharge and to reduce spike leakage in T-R cells—has already been described in the Technical Notebook section of our August issue (p. 402).

A component which has a very wide variety of uses is the circulator. This is a multiport device in which power entering port 1 emerges only from port 2 (no power being transferred to any of the other ports) and so on in a cyclic manner. This enables incident and reflected power in a waveguide to be separated without loss, and is thus of use, for example, in permitting a common aerial for transmission and reception or for separating the output from the input in a single-port parametric amplifier. Circulators in which magnetized ferrites are used to provide phase shifts in rectangular guide or Faraday rotations in circular guide which are non-reciprocal (i.e., unequal in opposite directions) are well known. However, these devices require complicated transitions which tend to make them bulky and expensive. Recently much more compact forms of circulator have been developed using (magnetized) ferrite loaded waveguide junctions and these were discussed by Aitken and McLean and also by Penney. Apart from their compactness, two other advantages of this type of circulator are that they can be more easily made broadband and also that they are less sensitive to temperature changes.

Non-reciprocal phase-shifting devices using ferrites need, as is well known, a magnetic field for their operation. When a switchable device is required this field is usually provided by an electromagnet. It is, however, difficult to produce rapid switching with this arrangement owing both to the self-inductance of the electromagnet and to switching eddy currents set up in the waveguide wall. A different approach described by Nourse is to concentrate the field entirely in the ferrite, making use of the remanent magnetism in a square-loop material. This results in the additional advantage that no energy is required to maintain the phase shift at a given value, but only to change it. Computer ferrite rings were used, mounted on a dielectric rod along the centre of the waveguide: a central conductor was used to carry current pulses which switched the direction of remanent circumferential magnetization in the rings. A minimum switching time of 50 nanoseconds was achieved using this arrangement.

The recent advent of low-noise pre-amplifiers has focused attention on image channel rejection devices, since image noise can degrade the noise factor by 3dB. The obvious method of image suppression is, of course, to use a filter, but it is difficult to design and track such filters for a wide tuning range. Another approach is to use an image suppression mixer, and one was described at the conference by Endean and Robinson. The circuit uses a combination of ring mixers and 90° phase shifters which is actually the standard single-sideband modulator arrangement, though the design conditions are different.

BOOKS RECEIVED

Physical Principles and Applications of Junction Transistors, by J. H. Simpson and R. S. Richards. Written for the physicist and laboratory worker who is not an electronics engineer. A thorough treatment of the basic physics of semiconductors is given in the first part of the book, which then goes on to deal with circuit design, both generally and detail. Mathematics are not particularly obtrusive, and there is a voluminous list of classified references. Pp. 519. Clarendon Press: Oxford University Press, Amen House, Warwick Square, London, E.C.4. Price 63s.

Handbuch des Rundfunk-und Fernseh-Grosshandels 1962/3. Specifications and nominal prices of German radio and television receivers, car radios, tape recorders, record players, aerials, etc. and prices of some current semiconductor devices, batteries and valves. Published by the German Radio and Television Wholesalers Federation with the collaboration of the editorial department of *Funk Technik.* Pp. 460; Figs. 1,278. Verlag für Radio-Foto-Kinotechnik G.m.b.H., Eichborndamm 141-167, Berlin-Borgiswalde. Price DM7-50 (postage 88Pf).

Design of a Group of Plug-in Television Studio Amplifiers, by K. J. Austin. A B.B.C. Monograph, on the electrical and mechanical design of video and pulse amplifiers for distribution and loss-correction in coaxial lines. Pp. 18. B.B.C. Publications, 35, Marylebone High Street, London, W.1. Price 5s.

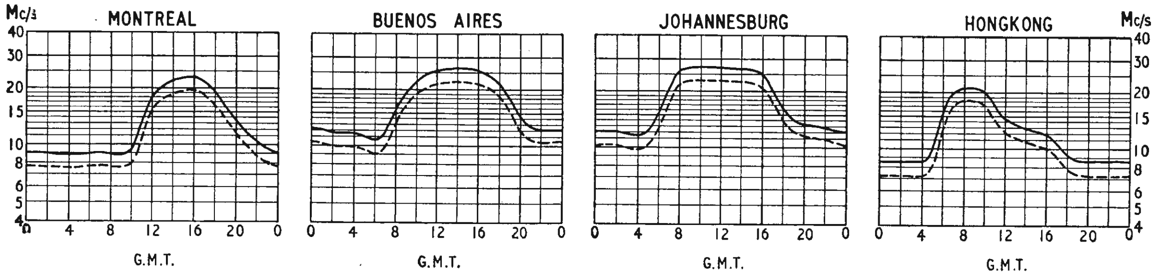
Introduction to Microwave Practice by P. F. Mariner. Adopts as far as possible a simple non-mathematical approach and makes use of a considerable number of explanatory diagrams. Topics discussed include transmission line and waveguide theory, measurement of reflection co-efficients, matching over broad and narrow bands, microwave instruments and components (with a special chapter on those employed in measurement) and the uses of ferrites. Three mathematical appendices prepare for more advanced work. Pp. 238. Heywood & Co., Ltd. (Book Sales Dept.), Tower House, Southampton Street, London, W.C.2. Price 50s.

British Instruments Directory and Buyer's Guide 1962. Information on firms, products, specifications and organizations. The glossary of terms has been extended to include a Russian section. Pp. 376. United Science Press Ltd., 9 Gough Square, Fleet Street, London, E.C.4. Price 63s.

Cabinet Handbook by G. A. Briggs. Deals with the more practical side of loudspeaker enclosure construction. The major part of the book is concerned with materials, processes and actual construction, while design in a general way is confined to two chapters. A whole chapter is devoted to the startling effects of an amplified electric guitar on loudspeakers, and cures for the ensuing troubles. Recommendations are given for room resonance damping. Pp. 112. Wharfedale Wireless Works Ltd., Idle, Bradford. Price 7s 6d.

SHORT-WAVE CONDITIONS

Prediction for November



——— PREDICTED MEDIAN STANDARD MUF
— — — PREDICTED OPTIMUM TRAFFIC FREQUENCY

The full-line curves indicate the predicted median standard maximum usable frequencies for the month of November.

The dashed-line curves show the optimum frequencies which are intended to allow for day-to-day variations from the monthly median.

These curves have been prepared by Cable & Wireless Ltd. from information supplied by the Radio Research Station, Slough.

WORLD OF WIRELESS

Forty Years of Broadcasting

TO mark the fortieth anniversary of the start of broadcasting in the U.K. by the British Broadcasting Company on November 14th, 1922, a book* has been published by the B.B.C. It opens with references to the early experiments of W. T. Ditcham from Chelmsword, and Capt. P. P. Eckersley from 2MT Writtle, for it was from the latter station that the first regular (though experimental) broadcast transmissions in the U.K. were made early in 1922. The original 2LO transmitter set up in Marconi House, London, in May that year for further experiments and demonstrations was taken over by the B.B.C. and became its first station. Older readers may recall that when 2LO (which had an output of 100W) was originally licensed it was for speech only and that at the end of every 7 minutes transmission, there had to be 3 minutes interval.

Lest we should be accused of parochialism we should mention that stations had been set up in other parts of the country. In Birmingham the Western Electric Company's station 5IT (originally operated in London) was taken into service by the B.B.C. on November 15th. In Manchester Metrovick had been operating 2ZY and this was taken into service within a week of the inauguration of the London transmitter.

To return to the B.B.C. book. Its three main sections cover broadcasting up to 1939, war-time developments and the post war period.

*"B.B.C. Sound Broadcasting: Its Engineering Development," B.B.C., 35, Marylebone High Street, London, W.1. Price 5s.

"Wireless World" Diary

A MINE of information, the radio engineer's *vade mecum*, and many other eulogistic clichés have been used by reviewers to describe the "*Wireless World* Diary" now in its 45th year of publication. We will confine ourselves to saying that we feel sure that those who have had a copy in past years will want to know that the 1963 Diary is now available. In its 80 reference pages are all the usual sections including addresses of organizations, formulæ, graphical and letter symbols, base connections for some 400 valves and a host of other useful general and technical information. The Diary costs 6s 2d (plus 1s 1d purchase tax) in leather, or 4s 6d (plus 9d) in rexine. Postage is 4d.

Dip. Tech. Awards

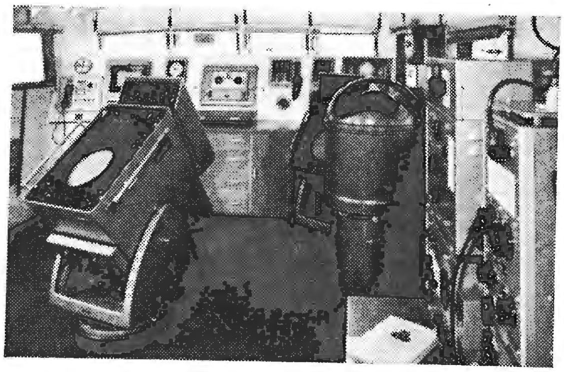
IN its latest Report the National Council for Technological Awards lists 18 courses in electrical engineering recognized as leading to the Diploma in Technology. In 17 of these courses students can, by a choice of optional subjects in the final part of the course, specialize in electronics and communications. Up to 31st March, 1962, the Council had conferred 409 "Dip. Techs." in electrical engineering—nearly 40% of the total of 1092 diplomas conferred—and of the 409 diplomates 155 specialized in electronics and communications or related subjects. Of the 6,200 students at present following Dip. Tech. courses 1,624 are studying electrical engineering.

The Report lists also the organizations which have provided industrial training for students upon whom the diploma was conferred during the year under review. Of the 62 "electrical" firms, 36 provided training for students studying electronics and communications.

The College of Technologists report five candidates for membership (M.C.T.) working in the field of electrical engineering; of these four are working on problems in electronics.

No Public U.K. Radio Show Next Year.—Radio Industry Exhibitions Ltd. has announced that it will not be organizing a National Radio Show at Earls Court in 1963 but intends to resume the series of public radio exhibitions in 1964. Next spring it is expected that manufacturers will stage a show for the trade only. Plans under consideration for the 1964 Radio Show include a suggestion that it should be international. B.R.E.M.A. have stated that no joint industry participation is contemplated in next year's international radio and television exhibitions on the continent.

The British Institute of Recorded Sound, founded in 1948 with the aim of forming a national archive of sound recordings of all kinds and to serve as a centre for their study, is to receive from the Government a grant in aid for an experimental period of three years at the rate of £10,000 in 1963-64, £12,500 in 1964-65 and £15,000 in 1965-66. The address of the Institute's Secretary is 38 Russell Square, London, W.C.1.



"Elettra III," Marconi's new yacht for the development and demonstration under seagoing conditions of marine communications equipment and navigational aids. She carries 12ft. and 6ft. slotted aerials for the Argus and Hermes demonstration radars as well as a small scanner for her own Consort set.

Les Platons, Jersey, v.h.f. sound station is changing the frequency of its Third Programme transmitter from 94.45 Mc/s to 94.75 Mc/s from November 1st. Under the Stockholm Plan of 1961 which came into effect on September 1st, this transmitter should have changed to 94.7 Mc/s. However, France in planning her f.m. sound broadcasting network required to alter the frequency of the Lille station from 92.2 Mc/s to 94.7 Mc/s. The U.K. therefore agreed at an international meeting held earlier this year at Kleinheubach, Germany, between west European signatories to the Stockholm Plan, to offset Les Platons by 50 kc/s to avoid co-channel interference.

At Brookmans Park, Herts., the B.B.C. have brought into operation a new 50-kW transmitter, which radiates the Light Programme on 1214 kc/s (247m) and serves the London area and parts of the Home Counties.

Portree, Isle of Skye, is now being served by a temporary low-power v.h.f. sound relay station transmitting the Scottish Home Service on 93.9 Mc/s, pending completion of a combined television and v.h.f. radio transmitting station at Penifiler early in 1964.

Hearing Aid Tests.—Reference is made in the annual report of the Royal National Institute for the Deaf to the work of its laboratory. In addition to its development work it provides a personal hearing aid test service whereby anyone can have any type of hearing aid tested if they suspect poor performance. The laboratory still provides facilities for testing and reporting on manufacturers' prototype hearing aids and, since the publication of the report, has announced the introduction of an audiometer calibration service.

"Network," a half-hour colour film made by the A.E.I. Film Unit, tells a true story of telecommunications. A ship stops at sea in the Gulf of Guinea; a casting has fractured. A replacement is flown from England and in less than 30 hours the ship is under way again. A vast network of telecommunications by radio, telephone and teleprinter carried the messages and makes such things possible. The film also shows the mass production of assorted telecommunications equipment in A.E.I. factories. It is available on loan from the A.E.I. Film Library, Crown House, Aldwych, London, W.C.2.

Components Used in Automatic Control Systems is the subject of an informal discussion which has been arranged by the Institution of Mechanical Engineers in conjunction with the Institution of Electrical Engineers, for November 20th next. Registration is required and applications should be addressed to the Secretary of the Institution of Mechanical Engineers, 1 Birdcage Walk, Westminster, London, S.W.1.

Electro-acoustics meeting of the Brit. I.R.E. at which Professor E. J. Richardson will speak on noise control, will be held on October 29th and not 31st as originally announced. The meeting begins at 6.0 at the London School of Hygiene & Tropical Medicine, Keppel Street, W.C.1.

Heathkit high-fidelity equipment is to be demonstrated by Daystrom Ltd. at the Russell Hotel, Russell Square, London, W.C.2, on November 3rd to 7th next, from 11 a.m. to 8 p.m.

"Stereo Broadcast Decoders."—In Fig. 6 of last month's article under this title (p. 489) the two 0.001μF capacitors near the extreme right hand side of the diagram should be 0.0001μF. In Fig. 8 the connections to some of the grids of the ECH84 heptode section are shown incorrectly: in each heptode grids 2 and 4 are internally connected and should go to the 22kΩ common h.t. dropping resistor (shown next to R_2) and grid 3 should be fed from the respective phase changing network terminated by the 470kΩ resistor.

"Sine v Square Waveform."—The output quoted at the end of the fourth paragraph of this article (October, p. 464) should be 650W.

B.A.T.C. Officers.—Grant Dixon, who has been chairman of the British Amateur Television Club since its very early days and has fostered its growth from some 20 members in 1947 to its present membership of about 500, has retired from the chairmanship but remains on the committee. He is succeeded by John Ware. Other committee members are J. Tanner (G3NDT/T), G. Sharpley (G3LEE), M. J. Sparrow (G3KQJ/T), I. Waters (G3KKD/T), L. A. F. Stockley, D. S. Reid, M. Lilley and J. Royle (G3NOX/T).

Micro-miniaturization.—With a view to securing some degree of co-ordination among manufacturers in the field of micro-miniaturization, an exploratory meeting to discuss future liaison and joint consultation between the members of the seven participating organizations was held on September 12th. The meeting was convened by the R.E.C.M.F. and attended by representatives of the following: B.R.E.M.A., E.E.A., S.B.A.C., S.I.M.A., T.E.M.A. and V.A.S.C.A.

Kenya's television station was formally opened in Nairobi on October 1st. Patrick Jubb, director-general of the new Kenya Broadcasting Corporation, stated that they are more concerned with television for educational purposes than for entertainment.

Amateur radio operators in New York State are to be allowed to have their station call letters and numbers registered as car numbers "in recognition of their service in rescue work and other emergencies."

"Toute la Radio", our well-known Paris contemporary of which E. Aisberg is director, is changing its title to *Toute l'électronique* with the November issue. The journal is in its 29th year of publication.

News From Industry

British Electronic Industries Ltd., the holding merger company for Pye Ltd. and E. K. Cole Ltd., report that trading profit for the year to March 31st last has dropped to £2.6M as compared with £3.2M for the previous year. After taxation, there is a loss of £155,112, as against a previous profit of £491,851. Individually Pye's pre-tax profit for 1961-62 is down from £1.39M to £0.89M, while E. K. Cole's loss has increased to £746,180 (£665,429).

Collaro Name Change.—Collaro (manufacturers of record changers, etc.) and Magnavox Electronics (marketing a range of radiograms, record players and tape recorders), both subsidiary companies of the American Magnavox Company, have been merged to form the Magnavox Electronics Company, with headquarters at Alfred's Way, By-Pass Road, Barking, Essex.

Lintas, the advertising agency which is part of the Unilever organization, has incorporated a comprehensive Marconi-RCA closed-circuit television system in its new premises in New Fetter Lane, London. TV auditioning, previewing "commercials", and general experimental work are facilitated by the equipment which will work on 405- or 625-line standards. A central control room houses two telecine channels and programme outputs are fed, after amplification and processing, to standard television receivers in offices throughout the building.

A.E.I. Automation Ltd. has been formed to develop and supply systems of automation for industrial applications in which computers will often be the main integrating apparatus. The company will undertake research, development and design, installation and commissioning, and sales and contract management. Headquarters of the new company will be in the grounds of Booths Hall, Knutsford. The chairman of A.E.I. Automation is Sir Eric Jones, managing director is J. L. Russell (see Personalities) and the chief engineer is J. L. W. Churchill.

British Relay Wireless & Television Ltd.—Trading profit for the year ended April 30th last, at £3,934,798, is some £417,000 up on the previous financial year. Chairman, Sir Robert Renwick, in his report, states that in the London area where the company serves 16 boroughs, they have been completing the line links whereby the 15 main stations and 42 sub-stations are integrated into what is believed to be the largest relay system in the world. He also reports the acquisition of a retail business trading in the London area as Gem Radio & Electrical Services and operating from 25 shops.

Electric & Musical Industries Ltd.—Group sales for the year to June 30th, 1962, were £82.58M, compared with £82.44M previously. Trading profits were £7.39M, against £7.84M for 1960-61. Profits and turnover of the Electronics Division reached a new peak, E.M.I. (Australia) profits increased substantially, and there was a further increase in the profits of Pathé Marconi in France. In the U.S.A. the profits of Capitol Records fell substantially, said to be partly due to the cost of the first year of its entry into the electronics field.

Bush and Rank Cintel.—Consolidated trading profit for the 53 weeks to June 30th was £370,266 as compared with £480,711 for the preceding year. After certain charges and taxation the net profit attributable to the parent company is £177,851 (£189,777 for 1960-61).

Radio Rentals' offer to acquire the ordinary share capital of the Dawes Radio Group Ltd. has received acceptances in respect of approximately 93% of such shares and the offer has been declared unconditional. The Manchester based Dawes Radio Group controls a chain of retail outlets for radio, television and electrical appliances in the North-west.

Marconi Marine have made arrangements with Kelvin Hughes (a division of S. Smith & Sons (England) Ltd.) to offer the latter's radar type 14/9 and 14/2 to U.K. shipowners on a rental basis only, so as to meet the special requirement of those needing a medium-priced radar. Marconi Marine service facilities will be available throughout the world for Kelvin Hughes radar supplied in this way.

Thorn Electronics Group.—A new group has been formed by Thorn Electrical Industries Ltd. to consolidate their various engineering and commercial activities in electronics, including TV studio equipment, closed circuit television, industrial instrumentation, process control equipment, radiation counters, and special c.r.t.s for instrument use.

Telcon Metals Ltd., a member of the B.I.C.C. Group, has established its own manufacturing plant in Italy for the production, from special alloys, of laminations and magnetic cores as used in the electrical and electronics industries. The new factory is at Agrate, 10 miles from Milan, and will operate under the name of Magnetic Cores (Italy) S.p.A.

Wideband limiter-discriminators developed by R.H.G. Electronics Laboratory Inc. are now available in the U.K. through Livingston Laboratories Ltd., 31 Camden Road, London, N.W.1. (Tel.: Gulliver 4191.)

Digital Measurements Ltd., 25 Salisbury Grove, Mytchett, Aldershot, Hants. (Tel.: Farnborough 2634) announce their appointment as sole European agents for the Military & Computor Electronics Corporation (MACE) of Florida, U.S.A.

Wire strain gauges manufactured by Tokyo Sokki Kenkyujo Co. Ltd., of Tokyo, Japan, are to be marketed in the U.K. by Electro Mechanisms Ltd., of 218-221 Bedford Avenue, Slough, Bucks. (Tel.: Slough 22228.)

The McMurdo Instrument Co. Ltd. have moved their precision wire-wound resistor production from Kingston to their Portsmouth factory.

Protona, the Hamburg, Germany, manufacturer of the Minifon magnetic-wire recorder, which is marketed in the U.K. by the Clarke & Smith division of Electric & Musical Industries Ltd., has been absorbed by Telefunken, the West German electronics concern. A spokesman for Clarke & Smith said that they had been advised by both Protona and Telefunken that the U.K. marketing arrangements for the Minifon product would continue as at present.

The Ministry of Aviation has placed orders with Marconi's W/T Company for the supply and installation of two airways radar links, one to serve Ringway (Manchester) and the other at Heath Row (London); also a dual channel surveillance radar for the U.K. Air Traffic Services Station, R.A.F. Hack Green; and three dual 500kW, 50cm radars for the U.K. Airways System installations at St. Annes, Lancs., Ash, Kent, and Ventnor, I.O.W.

Aveley Electric Ltd. have been appointed sole U.K. agents for the complete range of products manufactured by Nucletron & Electronic G.m.b.H., Munich, Germany, The Nucletron range comprises stabilized high voltage power supplies for all types of photomultiplier tubes, and nucleonic devices, mu-metal screens and holders to fit all well-known photomultipliers, c.r.t.s and other special purpose tubes.

The Potter Instrument Company, whose recent exhibition in London introduced this American company's high density magnetic tape system to British Government specialists and industrialists, are soon to establish a servicing and spare parts depot in the London area. For the present all enquiries should be addressed to the company at 7-9 St. James's Street, London, S.W.1.

Fine Tubes Ltd. new plant at Estover, near Plymouth, was inaugurated on October 5th. It is anticipated that the new 70,000 sq ft mill will increase production of the precision drawn small diameter thin-wall tubing by up to 50%. Electronic applications for the tubing include use as cathodes, coaxial tubes and sleeving for tungsten leads in all types of electron tubes.

Addison Electric Co. Ltd. announces that an American subsidiary company has been registered in the name of Addison Electric Co. Inc. The address is 1101 Bristol Road, Mountainside, New Jersey, U.S.A. Addison Electric is a member of the Muirhead Group.

W. G. Pye & Co. Ltd. is to manufacture, in the U.K., pH electrodes to the specification of Dr. Werner Ingold, of Zurich, Switzerland. These arrangements will not affect production of the established range of Pye electrodes but they will enable Pye to offer a widened range.

OVERSEAS TRADE

The New Zealand Broadcasting Service is already making active preparations to cover the Queen's visit next February. Another television outside broadcast vehicle—the third in eight months—has been ordered from Marconi's. The latest O.B. unit is to be based at the Wellington station which, like the other three New Zealand television stations at Auckland, Christchurch and Dunedin, is equipped with Marconi vision and sound transmitters.

A Scorpion radar set for use in testing the British Aircraft Corporation's Bloodhound II guided missile at the Woomera rocket range, was recently despatched to Australia by A.E.I.'s Electronic Apparatus Division. It is described by A.E.I. as their largest ever single consignment—over 80 tons—of electronic equipment for overseas.

Sweden's present air defence system is making extensive use of early warning and height finding radar equipment manufactured by Decca Radar Ltd. and installed by their Swedish subsidiary, Decca Navigator och Radar AB.

Personalities

Viscount Mills, K.B.E., has rejoined the board of directors of Electric & Musical Industries Ltd. He was previously a director of E.M.I. from July 1955 to January 1957, and rejoins the board after five years during which he has been successively Minister of Power, Paymaster-General and Minister without Portfolio in the Cabinet. The E.M.I. directors also announce the resignation of **Clifford Metcalfe,** C.B.E., from the board. Mr. Metcalfe was managing director of E.M.I. Electronics Ltd. from 1956 to 1961 and has been a member of the E.M.I. board since December, 1956.

The Plessey Company has announced that the **Earl of Kilmuir,** P.C., G.C.V.O., has accepted an invitation to join the board of the company with a view to succeeding to the chairmanship upon the retirement of the present chairman, Sir Harold Wernher, at the end of the year. A former Attorney General and Home Secretary, Lord Kilmuir was Lord High Chancellor from 1954 until June this year.

E. Eastwood, C.B.E., Ph.D., M.Sc., M.I.E.E., has been appointed group director of research for the English Electric group of companies, to be responsible for the co-ordination of all research work in the group and for the direction of the English Electric's Nelson Research Laboratory at Stafford, the Mechanical Engineering Laboratory at Whetstone, and the group research work carried on at the Marconi Laboratories at Great Baddow, Chelmsford. Following Dr. Eastwood's appointment **G. D. Speake,** M.A., has been appointed chief of research of the Marconi Company. Mr. Speake joined Marconi's W/T Company in 1950, when he was engaged in radar systems research. In 1956 he was appointed chief of the microwave physics group of the research division. He has been deputy chief of research since 1960.

C. E. Strong, O.B.E., B.A., M.I.E.E., F.I.R.E., head of the Radio Division of Standard Telephones & Cables Ltd. since 1938, has now been appointed to the executive staff at S.T.C.'s headquarters in London. Mr. Strong, in his new position, will act as the company's representative on Government and industrial bodies and also assist in the planning of development work on navigational and radio equipment in Standard Telecommunication Laboratories. He is succeeded by **K. P. Wood,** B.Sc.(Eng.), M.I.E.E., who was at one time senior lecturer in communication engineering and head of the Electrical Engineering Department at the Medway College of Technology. He later became an executive director of Cossor Communications Co. Ltd. He joins S.T.C. from British Communications Corporation Ltd., where he was director and general manager.

J. Langham Thompson, M.Brit.I.R.E., has resigned from the board of Ether Langham Thompson Ltd., of which company he was deputy chairman. Mr. Thompson has also resigned his directorship in J. Langham Thompson Ltd., the company he founded in 1946 and of which he was chairman and managing director. Aged 56, Mr. Thompson was with A. C. Cossor for seven years before joining McMichael Radio in 1934 as chief engineer. After war service as officer i/c wireless section, of the Inspectorate of Electrical and Mechanical Engineering until 1942, he practised as a consulting engineer until 1946. He is a vice-president of the Brit.I.R.E. and the Institute's representative on the Parliamentary and Scientific Committee.

D. H. C. Scholes

J. Wallace

D. H. C. Scholes has been appointed an executive director of the Plessey company and technical co-ordinator for the Plessey Group of Companies. Mr. Scholes, who began his career with Marconi's W/T Company, joined Plessey in 1946 as a senior engineer. In 1950 he became chief radio engineer and two years later chief engineer, Telecommunications Division. Since 1960 he has been chief engineer, Electronic and Equipment Group.

J. Wallace, D.L.C. (Eng.) Hons., has been appointed to the board of the M-O Valve Co. Ltd. as production director. He was previously with Standard Telephones & Cables Ltd. as manufacturing manager of the Microwave Systems Division.

The Electrical Research Association announces that it has appointed **G. Siddall** to be head of its newly formed Electronics Research Department. Mr. Siddall has been engaged on research and development in connection with the production of thin films for the past 12 years, at the A.E.I. Research Laboratories (Manchester) and afterwards with Edwards High Vacuum Ltd., Crawley. As deputy head of the department, the Electrical Research Association has appointed **R. M. Hill,** Ph.D., an Associate of the Royal College of Science and Technology, Glasgow. Dr. Hill has been an I.C.I. Research Fellow at the Clarendon Laboratory, Oxford, where he was investigating special applications of gaseous optical masers.

Antiference announce the appointment of **R. S. Roberts,** M.Brit.I.R.E., as technical consultant. Mr. Roberts is principal lecturer of the Department of Telecommunications, Northern Polytechnic. He has been continuously engaged in industry since 1923, becoming a consultant when he left Wright & Weaire (where he was chief engineer) in 1936 to join the newly formed Radio Department of the Northern Polytechnic. During the war he was seconded to the Ministry of Aircraft Production for the development of radar.

T. S. Littler, Ph.D., director of the Wernher Research Unit on Deafness, who was associated with the development of the Medresco hearing aid, has accepted the appointment of honorary scientific adviser to the Royal National Institute for the Deaf. He succeeds **P. Denes,** Ph.D., M.Sc., A.M.I.E.E., of London University, who is going to America.

C. Powell, originator of the two-range Decca technique and of many ideas for using hyperbolic systems as an aid to air and marine surveying, has been awarded the fellowship of the Institute of Navigation. After wartime service in the Army Operational Research Group he joined the Decca Navigator Company in 1946. He contributed an article on "Radio Aids to Hydrography" in our July 1960 issue.

Grp. Capt. A. D. Jackson has been appointed Controller of R.A.F. Telecommunications at H.Q., Signals Command, with the acting rank of Air Commodore. After studying at University College, London, he joined the R.A.F. in 1939 at the age of 24. He took a specialist signals course and served abroad before being posted to No. 90 (Signals) Group (the forerunner of Signals Command). He subsequently commanded R.A.F. Bawdsey, Suffolk. From 1949 to 1952 he was on the staff of the U.S.A.F. Air University, Alabama, on exchange posting, and returned to the R.A.F. Technical College, Henlow, Beds. Since December 1960, Grp. Capt. Jackson has been at the Air Ministry.

Sqn. Ldr. C. V. Colbon has joined Wayne Kerr Laboratories Ltd. in the position of Government liaison officer following his retirement from the Royal Air Force after 36 years' service. For the past three years he was at the Air Ministry where he was responsible for electronic test equipment policy.

Sqd. Ldr. C. V. Colbon

A. Cormack

A. Cormack, B.Sc., A.M.I.E.E., has been appointed chief engineer of G.E.C. (Electronics) Ltd., Coventry. Mr. Cormack was a member of the scientific staff at the Hirst Research Centre of the G.E.C. for 14 years prior to this appointment.

W. J. Bates, Ph.D., B.Sc., F.Inst.P., has been appointed to the board of R. & J. Beck Ltd. in the capacity of technical director. During the war Dr. Bates did research work in the Admiralty Signals Establishment and in a department of the Ministry of Aircraft Production. Later he carried out research work on interferometry, leading to a Ph.D. degree, and then joined the staff of Bristol University as lecturer in the Physics Department.

J. L. Russell, M.I.E.E., has been appointed managing director of A.E.I. Automation Ltd. Aged 42, he joined Metropolitan-Vickers Electrical Company as an apprentice in 1936, and attended Manchester College of Technology of which he is an associate. In 1947 he entered the newly formed Electronic Control Engineering Department and became section leader, Special Contracts, in 1949, and assistant chief engineer of the Special Applications Division of the Electronics Engineering Department in 1955. He is chairman of the Measurement and Control Group of the N.W. Centre of the I.E.E.

E. K. Cole, C.B.E., founder and former chairman and managing director of E. K. Cole Ltd. until its merger in 1960 with Pye to form British Electronic Industries (from which company he resigned last year), has joined Robinson Rentals as advisory chairman. Bedford based, Robinson Rentals operates a chain of nearly 100 shops engaged in the TV receiver hire trade.

F. C. Thompson, Ph.D., formerly manager of the radar division (which incorporated the magnetron and cathode-ray tube departments) of the English Electric Valve Co. Ltd., has now assumed the position of an assistant general manager. This appointment covers responsibility for all internal company maintenance and service departments, the piece part production plant and a newly formed display tube department which comprises cathode-ray tubes and memory tube production. The remainder of radar division—magnetron department—has been combined with the microwave research division to form a microwave tube division under the managership of **J. Dain,** M.A. (Cantab.), M.I.E.E. **E. Allard,** B.Sc., A.M.I.E.E., previously assistant to the managing director, has been appointed development manager covering all company and Government development projects, all company patent, research and development licensing and finished product inspection.

C. L. Wolsey has joined Morganite Resistors Ltd. as chief engineer, electronic components, to take over the management of the electrical test laboratory and the applications laboratory. Mr. Wolsey was formerly with Murphy Radio Ltd. **G. G. Davison** has been appointed chief design engineer, in charge of Morganite Resistors' design and development departments. **M. G. B. Mason,** former technical manager, has resigned.

R. B. D. Knight, M.A. (Oxon), D.Phil., A.Inst.P., has been appointed to the board of Digital Measurements Ltd., as director of research.

OUR AUTHORS

J. Somerset Murray, B.A., A.M.I.E.E., who writes in this issue on tape recorder circuits, first contributed to *Wireless World* in 1925 while he was at Emmanuel College, Cambridge, where he graduated in 1927. He subsequently did a year's research in acoustics. He was for a short time with P. K. Turner, of loudspeaker fame, who was at one time Editor of our sister journal *Wireless Engineer* (now *Industrial Electronics*). From 1939 to 1947 Mr. Murray was with S.R.D.E. as liaison officer on behalf of the Inspectorate of Electrical & Mechanical Engineering. Since 1948 he has been a consultant specializing in communications engineering and instrumentation.

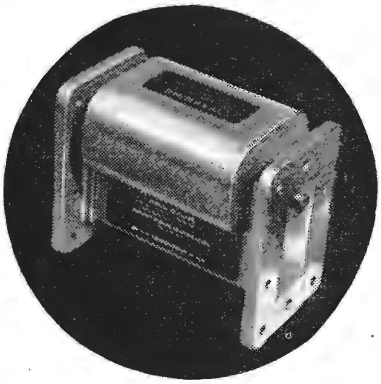
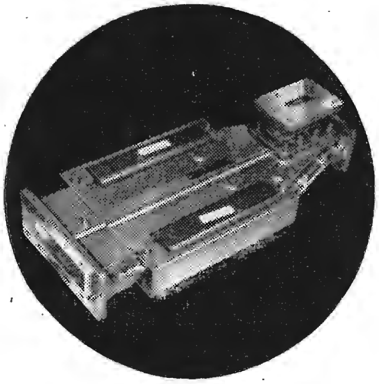
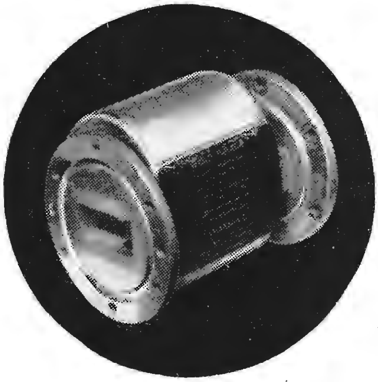
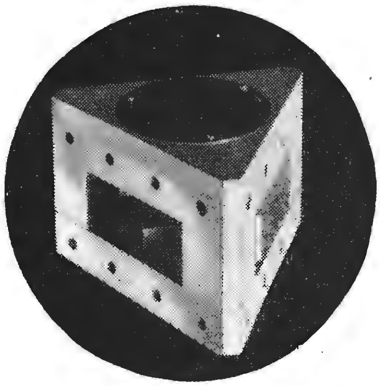
Aubrey Harris, A.M.I.E.E., A.M.Brit.I.R.E., contributor of the article on page 516, has been with Ampex since 1958, initially in the U.S.A. and for the past year with the U.K. subsidiary at Reading, Berks, where he is senior engineer. He was at the Post Office Research Station, Dollis Hill, for a time before joining the Marconi Company in 1952.

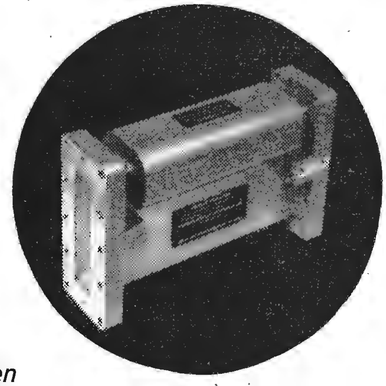
N. C. Baust, Grad.I.E.E., author of the article describing a current-controlled Schmitt trigger, has been with English Electric Aviation since 1957. He has been particularly concerned with the development of transistor digital computers. Prior to joining E.E.A. he was a student apprentice with the Electrical Research Association.

OBITUARY

E. W. Berth-Jones, B.Sc., whose death occurred recently, was a well-known figure in sound recording circles. Having obtained a science degree at Manchester University he joined E.M.I.'s Abbey Road studios in 1937, and was then occupied with design and general research work, eventually becoming chief engineer of the Recording and Relay Division of the old Gramophone Company. In 1961 he transferred to E.M.I. Electronics and became chief engineer of the Broadcasting and Recording Equipment Division.





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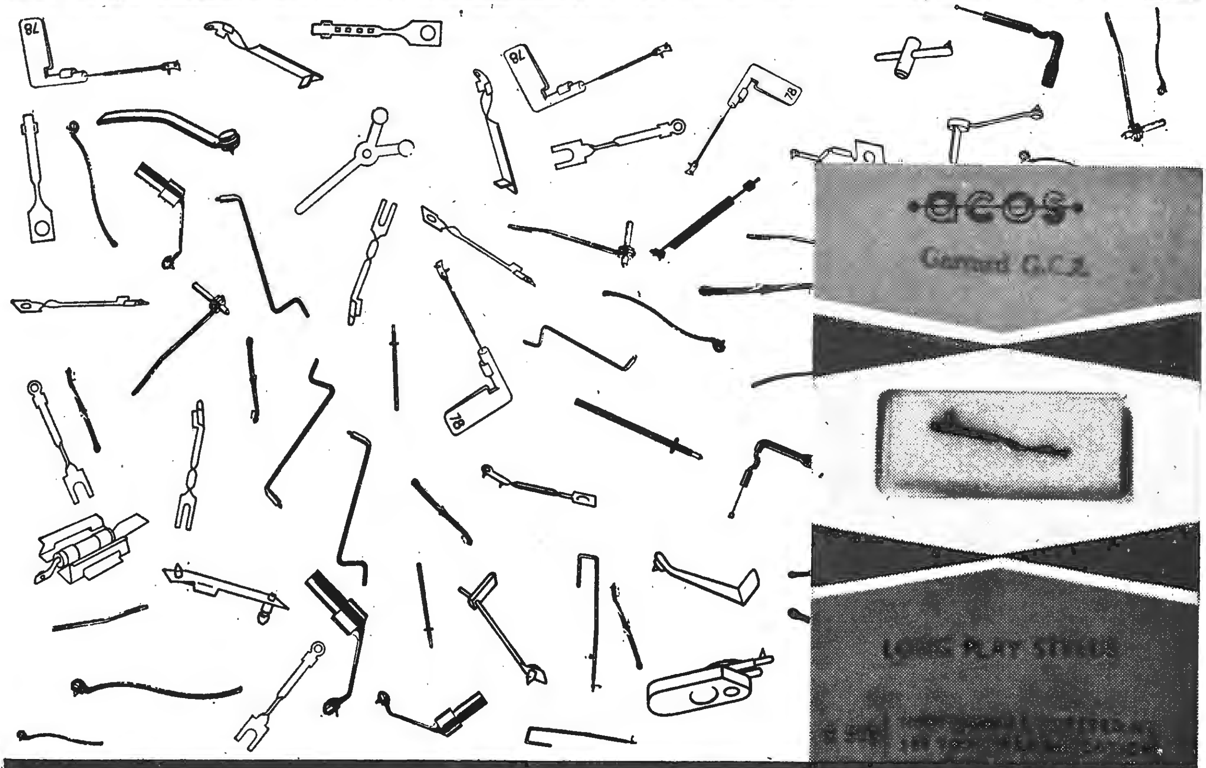

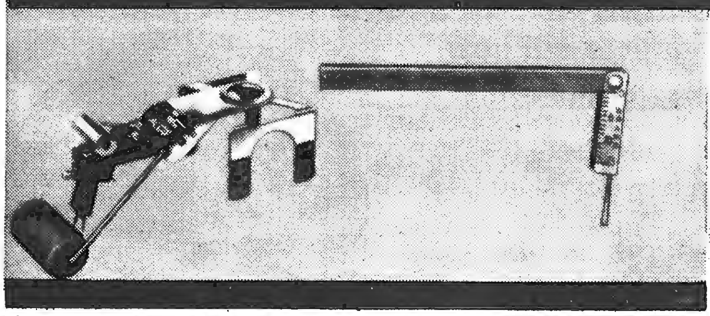

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TAPE RECORDER CIRCUITS

1.—USE OF HEAD TERMINATING RESISTOR

By J. SOMERSET MURRAY, B.A.(Cantab.), A.M.I.E.E

A TAPE head is an iron-cored inductor with a very small linear gap across which is impressed the magnetization of the element of tape in contact with it, or which itself applies a magnetic field to the tape in the recording process. Like all inductors it has both self-capacitance and resistance. The resistance is partly due to the copper loss, but the losses due to the magnetic circuit are more important at high frequencies, giving the inductance a $Q(=X/R)$, tending to a maximum as the frequency is increased.

The self-capacitance resonates with the inductance to provide a limiting frequency of operation. This article shows how these inevitable defects can be controlled to achieve a linear response in both recording and playback. It is shown that the recording and replay responses are identical and therefore the solutions are similar. The discussion is limited to the effects of the stray circuit elements and does not deal with tape and gap losses. These are more properly dealt with by pre-distortion in the recording chain and equalization in the replay. The problem discussed is simply that of obtaining a linear frequency response for the field applied to the tape and for the magnetization reaching the core.

Record head and bias supply circuit:—The bias supply to the head must not interfere with the signal supply or vice versa. Sometimes an additional winding on the head is used for the bias supply but this cannot be the right solution since, by transformer action, bias e.m.f.s. are generated in the signal coil and will cause current to flow at bias frequencies in the signal circuit. Again, there will be a signal current induced in the bias winding. This may modulate the oscillator, which must operate in class B or class C and therefore will be non-linear, and this may result in intermodulation. The bias feed problem is solved by using a single winding and feeding the signal to one end and the bias to the other, arranging a capacity to resonate with the head at some frequency between the wanted band and the bias frequency. This is easier when the bias frequency is high. The circuit is shown in Fig. 1.

The signal current traverses the head and returns to earth through the secondary winding on the bias oscillator transformer. At signal frequencies this winding adds only a small inductance to that of the head and can be neglected. Any signal voltage appearing across the oscillator primary will be very small compared with the anode-to-anode bias frequency voltage. Nevertheless rectification will take place in each anode and this should be examined. The instantaneously increased anode voltage on one anode increases the power output from that anode. At the same time the other anode gives less power. The result is an approximately constant output at the bias frequency. Only if this modulation is non-linear will third-order modulation terms be generated which will not cancel. However, if the coil is centre tapped, the second-order currents will not transfer themselves to the secondary. Note that the use of attenuators employing dissipative elements is inadmissible since these may cause the impedance of the bias oscillator secondary winding to be other than inductive at signal frequencies. The bias current may be varied coarsely by taps; with a fine control, covering the taps, by variation of the d.c. supply to the oscillator. The path of the bias current is then through the head inductance and the capacitor C back to earth.

The capacitor C_0 is the sum of the coil self-capacitance and that of the cable. It acts differently

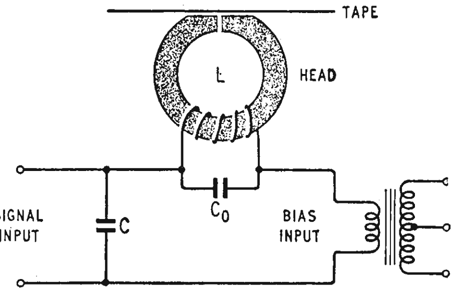


Fig. 1. Method of feeding signal and bias to a recording head.

in the two directions: since the impedance at signa frequencies of the bias winding will be many times less than that of the head, C_0 is effectively in parallel with C to signal frequencies; at bias frequencies the L C_0 C combination is a reactance transformer and all this does is to modify the voltage required from the bias winding without affecting the mode of operation.

The capacitor C allows the bias current to return to earth but develops a voltage at bias frequency across it. The bias voltage at C is many times less than that at the other end. The resulting dissipation of bias frequency power in a resistor R, destined to ensure a constant current, will be small and acceptable. This circuit requires no traps, the ladder L C network attenuating the bias current on the signal side.

The recording circuit is now seen to be, in its simplest elements, as in Fig. 2. The signal voltage V produces a current I in the head which we must make as nearly independent of frequency as possible.

Replay circuit:—If the generator and load change places, by the reciprocity theorem (see for example, W. L. Everitt, *Communication Engineering*, 2nd edition, McGraw-Hill, 1937, p. 52) the transfer

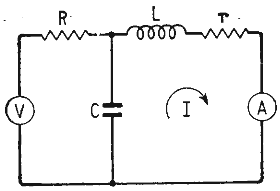


Fig. 2. Recording circuit corresponding to Fig. 1.

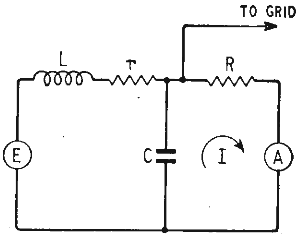


Fig. 3. Replay circuit obtained by interchanging generator and load in circuit of Fig. 2.

impedance is the same and the same output current will flow. Fig. 3 shows the replay circuit in which this inversion has taken place. L is now the replay head. If E = V the current I must be identical in the two circuits of Figs. 2 and 3. For replay purposes a voltage will appear across R (= IR), or the current in R may be used to drive a current-operated amplifier.

At this point we must examine the relationships between L, C and R to obtain the best frequency response. From the reciprocal symmetry it is obvious that the same solution applies to both the recording and replay functions of a single head. With three-headed deck assemblies, where the functions are separated, the performance of the complete system may be improved by use of a wide gap for recording and a narrow gap for replay. Also there will be a choice of more than one value of head inductance for the two functions. Nevertheless the L C R - relationship will be valid in all cases.

Maximally flat response:—This is a term introduced by V. D. Landon (*R.C.A. Review*, vol 5, p. 347, Jan. 1941). The general shape of the response of the circuit in Fig. 3 is well known to be of the type shown in Fig. 4 where the control of the peak depends on the relationship between L, C and R. For the purpose of tape recording we shall not be far wrong in taking the maximally-flat condition as our aim. This is produced when the slope of the response curve is zero for most of the band and is nowhere positive.

The transfer characteristic of a T network is

$$\frac{E}{I} = Z_T = \frac{Z_1Z_2 + Z_2Z_3 + Z_3Z_1}{Z_3} \quad (1)$$

Putting $Z_1 = r + j\omega L$, $Z_2 = R$ and $Z_3 = 1/j\omega C$, we have

$$Z_T = j\omega CR(j\omega L + r) + R + r + j\omega L \quad (2)$$

r is considered to be small and from now on will be neglected. We are concerned with the modulus of Z_T and its derivative.

$$|Z_T| = \sqrt{R^2(1 - \omega^2LC)^2 + \omega^2L^2} \quad (3)$$

Landon's definition of a maximally-flat transfer impedance is this expression in which the terms in ω^2 have been eliminated. Squaring and expanding we have

$$|Z_T|^2 = R^2 - 2R^2\omega^2LC + R^2\omega^4L^2C^2 + \omega^2L^2 \quad (4)$$

The ω^2 terms vanish when $\omega^2L^2 = 2R^2\omega^2LC$ or

$$\text{when } L = 2R^2C \quad (5)$$

It is interesting to note that equation (5) may be derived from equation (4) by equating the derivative of $|Z_T|^2$ with respect to ω^2 to zero at $\omega = 0$.

When R or C is adjusted to make equation (5) true the transfer impedance becomes

$$Z_T = R\sqrt{1 + \omega^4L^2C^2} \quad (6)$$

At the frequency of resonance, f_0, $\omega^2LC = 1$, i.e., $Z_T = R\sqrt{2}$ and the transfer loss is 3dB; above this the response falls at 12dB/octave. At any other frequency, f, the expression for Z_T becomes

$$Z_T = R\sqrt{1 + \left(\frac{f}{f_0}\right)^4} \quad (7)$$

This expression for Z_T may be used to determine the frequency at which the response has fallen by any given amount. For instance, it may be decided that the upper band edge (f_1) should be where the response has fallen 1dB. We then put $Z_T/R = 1.122 = \sqrt{1 + (f_1/f_0)^4}$ and obtain $f_1/f_0 = 0.71$, which puts the resonant frequency at 1.4 times the band edge. Thus a 20kc/s audio band requires a resonance at, at least, 28kc/s. It will be better to put this slightly higher at say 35kc/s since we have not yet examined the effect of the Q of the head.

Recording circuit values:—With this frequency settled—at 35 kc/s—we can derive the values for a 7mH record head as follows: $C = 1/4\pi^2f_0^2L = 2{,}950$pF, $R = \sqrt{L/2C} = 1{,}088\,\Omega$.

The signal voltage needed will be very moderate, certainly not more than 4 or 5V peak.

Playback values:—The replay head of, for instance, 120mH, will be given the same 35kc/s resonant frequency and have the same -1dB loss at 20kc/s. This will require C = 172pF and R = 18.7kΩ.

Since the gap and hysteresis losses will require slight additions to the "normal" equalization, and these can only be found by experiment, it is probable that the normal values as calculated above will be a good starting point.

Correction for finite Q of heads:—In the analysis shown above the small resistive component in the inductive branch has been neglected. We must now discover by how much the inclusion of this resistance will impair the validity of the results.

One way of accounting for the effect of r is to make use of the fact that the Q of a head is never very high and is nearly constant over the upper part of the wanted band.

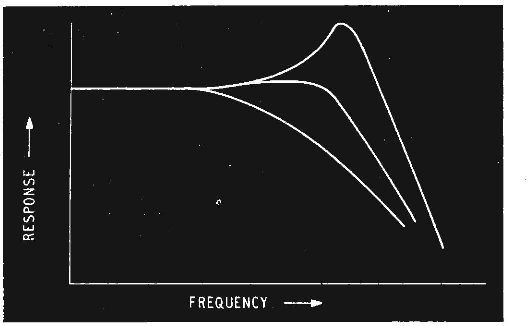


Fig. 4. General shape of response of replay circuit of Fig. 3.

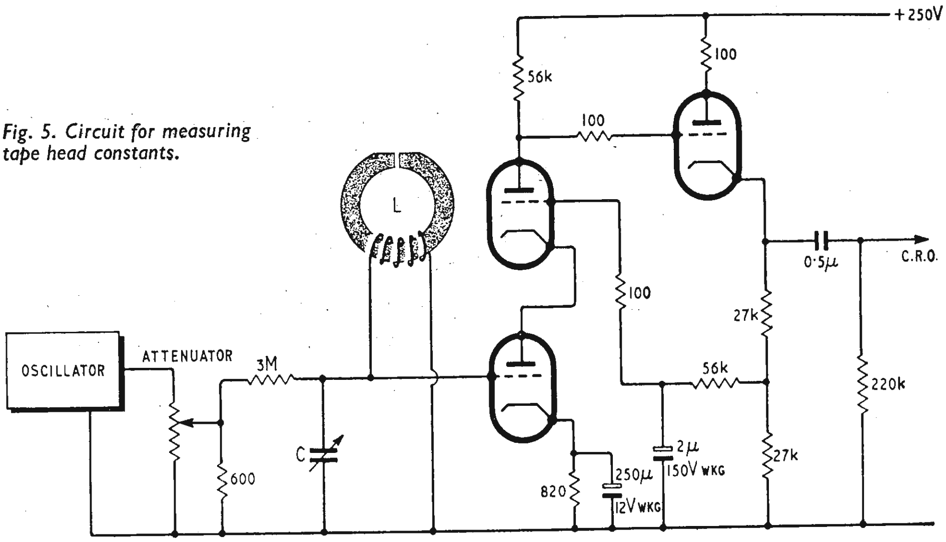


Fig. 5. Circuit for measuring tape head constants.

Bearing in mind that the value of Q which is significant will be that measured at the upper band edge we may write $\omega L/Q$ for r in equation (2). When this is done we obtain for $|Z_T|^2$ an expression which contains the first power of ω in addition to the second and fourth. Obviously we can only put the sum of the first and second order terms equal to zero at one frequency, so the flat part of the frequency response will no longer be perfectly flat. Nevertheless, we can still make the flat part approximate very closely to perfection if we choose R so that the sum of the first and second order terms is zero at the resonant frequency, f_0.

$$|Z_T|^2 = \omega^4 C^2 R^2 L^2\left(1 + \frac{1}{Q^2}\right) + \omega^2\left[\frac{L^2}{Q^2} + L^2 - 2CR^2L\right] + \frac{2\omega LR}{Q} + R^2 \qquad (8)$$

Equating the sum of the first and second order terms to zero we get

$$\omega^2\left[\frac{L^2}{Q^2} + L^2 - 2CR^2L\right] + \frac{2\omega LR}{Q} = 0 \qquad (9)$$

Substituting $\omega = 1/\sqrt{LC}$ gives

$$2\frac{C}{L}R^2 - 2\frac{R}{Q}\sqrt{\frac{C}{L}} - \left(1 + \frac{1}{Q^2}\right) = 0 \qquad (10)$$

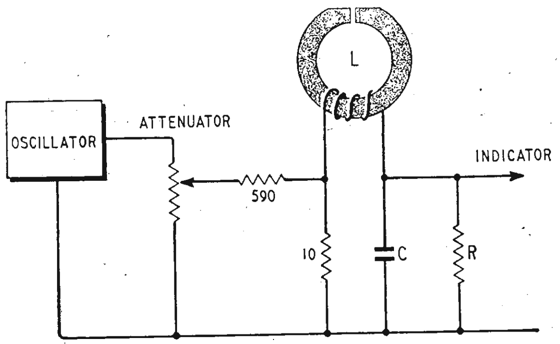


Fig. 6. Circuit for measuring replay response of tape head with terminating resistor.

From this

$$R = \sqrt{\frac{L}{C}}\left[\frac{1}{2Q} + \frac{1}{2}\sqrt{2 + \frac{3}{Q^2}}\right] \qquad (11)$$

and this is the practical expression for R. It reduces to equation (5) when Q is put equal to infinity.

Experimental confirmation:—Perhaps a few notes on the application of these principles to an actual head may not be out of place. The method is not difficult and uses commonly-available laboratory equipment. Anyone who has access to an audio oscillator and a cathode-ray-oscilloscope or a valve voltmeter can do it. I have recently pensioned off an old replay head and replaced it with a Bogen UK 104; the procedure was as follows:—

Adjusting a replay system:—This inexpensive head of nominal inductance 120mH was fitted to the deck. A pair of leads in an earthed shield were taken to a small laboratory flat amplifier. This amplifier besides providing a voltage gain of about 80 times has a low output impedance and can drive a cathode-ray-oscilloscope such as the Serviscope D 31 without loading the head. The circuit is shown in Fig. 5.

An oscillator was connected via a terminated attenuator to a 3MΩ resistor which was connected to the live side of the head and to the grid of the cascode. The first measurement is the resonant frequency. This proved to be 28kc/s with the four feet of lead used. By shortening the lead to the minimum practical length of just under two feet the resonance was lifted to 32kc/s. This meant that the target had to be slightly lowered to a value for the upper band-edge of 18kc/s.

The inductance and capacitance were now measured by the usual h.f. technique of adding capacitance until the resonant frequency had been exactly halved to 16kc/s. This gave L = 100mH and C = 245pF. To measure the Q, the capacity was adjusted to resonate the head at 18kc/s. The response was now found to be 3dB down at 16kc/s and at 21kc/s, giving a Q of 3.6.

This value of Q was a little disappointing, but is

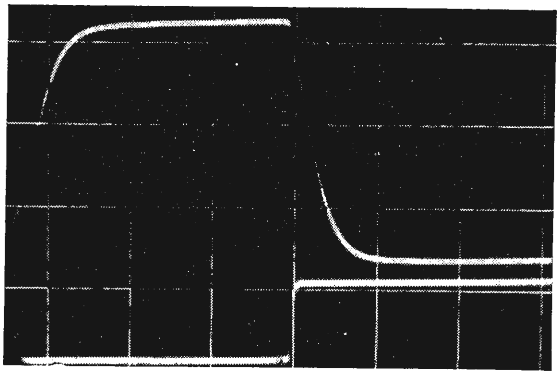


Fig. 7. Square-wave replay response of tape head with correct terminating resistor.

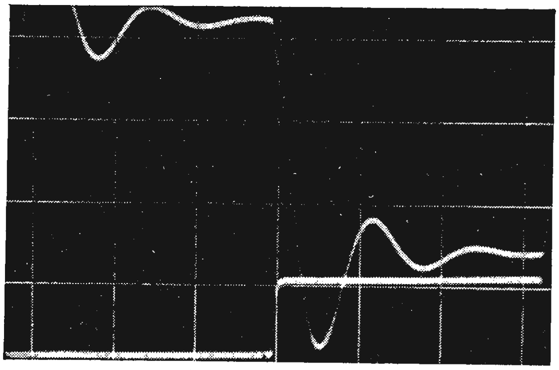


Fig. 8. Square-wave response of tape head without terminating resistor.

common in replay heads at present available. The recording head Q is often much better since it can afford to have a wider gap and sometimes even a rear gap (which is useful in maintaining the inductance at a constant value). It should be remembered that any change in the replay head inductance causes a real loss in sensitivity with frequency, since the inductance is a measure of the flux linkage with the pole pieces. For the present this will be regarded as just another loss to add to the gap loss for subsequent correction.

We can now check the response after adding the appropriate value of R. From the values of L, C and Q found above, we get a value for R from equation (11) of 17.9kΩ. The nearest value, 18kΩ, is then inserted between the hot side of the pair of leads and earth. The circuit is rearranged for the measurement of the response. Instead of the 3-MΩ resistor, the drive to the head is taken to the earthy side of the pair from a low resistance, about ten ohms, inserted at the earthy end of the 600-Ω resistor which terminates the attenuator, as in Fig.6. The frequency response was measured and gave the expected result as shown in the following table:—

Frequency (kc/s) ...	18	23	28	32	38	41	53	66	95
Response fall (dB) ...	1	2	3	4	5	6	9	12	18

The agreement with anticipation is highly satisfactory, the only significant divergence being the 4dB instead of 3 at 32kc/s!

For those with square-wave generators the value of R may be found empirically by adjusting it until the response shows the combination of minimum rise-time without any overshoot as in Fig. 7. This picture was taken by that method and gave the same value of R as calculated. The horizontal trace speed is 20μ sec/cm. The repetition frequency was about 8kc/s and the rise-time is seen to be approx 12μ sec corresponding to the calculated 3dB bandwidth of 32kc/s.

Conclusion:—From this set of arguments we can say that there are certain conditions which should be satisfied by any head assembly if a flat response is to be expected from this element in the system. First, the resonant frequency connected to the pre-amplifier must be at least 1.5 times the maximum wanted frequency. Next, it should have a reasonable Q at that frequency. Finally, the appropriate load resistor should be used. This can be found from equation (11) or by the square-wave approach. The second photograph (Fig. 8) shows what happens to the pulse response when the resistor is omitted altogether! It is otherwise identical with Fig. 7.

The defect indicated by the 36% overshoot has two actions on the noise characteristics of a tape system. In the first place, it makes an excessive demand on the linearity of the tape itself. Without it the recording level could be raised 2.7dB before the distortion limit were reached on programme type signals. On the replay side, the overshoot represents an unwanted peak in the response outside the band, but where there is likely to be a considerable lift in the replay amplifier in an attempt to overcome head losses, etc. This will pick up noise from the tape and add to the general level of power being transmitted through the amplifiers, which must increase the intermodulation distortion noise in the reproduction.

Radio Communications Exhibition

THIRTY exhibitors, some showing equipment from overseas, have taken space at the International Radio Communications Exhibition which opens at Seymour Hall, Seymour Place, London, W.1, on October 31st, for four days. Previously known as the Radio Hobbies Exhibition, this is the 15th in the series sponsored by the Radio Society of Great Britain.

A feature of this year's show is the series of technical lectures which will be given in the afternoons at 3.0. The subjects are v.h.f. aerials (31st), transistor power supplies (1st) and communication receivers (2nd).

Admission to the exhibition, open daily from 10 to 9, costs 3s. It will be officially opened by Capt. C. F. Booth, of the G.P.O., at 12.0 on the first day.

Amateur Radio Mobile Society
Avel Products
British Amateur Television Club
City and Guilds of London Institute
Daystrom
Electroniques (Felixstowe)
Enthoven Solders
Green & Davis
J. Beam Aerials
K.W. Electronics
Minimitter Company
M-O Valve Company
National Trade Press
Newnes
Philpott's Metalworks
Post Office Engineering Dept.
Radar & Electronics Association
Relda Radio
Royal Air Force
Royal Naval Reserve
R.S.G.B.
Salfords Electrical Instruments
Selray Book Company
Short Wave Magazine
65th Signal Regiment, T.A.
Sound Vision Service (Electrical)
Webbs Radio
Wireless World and *Industrial Electronics*
Withers (Electronics)

LC Oscillator Design

1—CONSTANT CURRENT SWITCHING FOR SINE-WAVE GENERATION

By R. C. FOSS, B.Sc., Grad. I.E.E., and M. F. SIZMUR, B.Sc.

FOR many years the thermionic valve was thought of almost exclusively as an amplifier. It may be because of this historical legacy that there is a tendency to regard an amplifier as straightforward and reasonably simple to design while waveform generating circuits are regarded as complex. Unfortunately, the treatment given in many textbooks tends to confirm this view. It is difficult to find any work dealing with the unifying principles of waveform circuit design.

One of the most important of such principles is that it is often more profitable to consider valves or transistors not as amplifiers but as switches. In a previous article published in *Wireless World*[1] the authors attempted to show that reliable design of such waveform circuits should be based on switches in which the states of the switch are determined mainly by the passive circuit components and do not depend critically on the precise valve or transistor parameters. In that article it was shown that the long-tailed-pair is a useful basic switching circuit satisfying this requirement. Whilst it may be obvious how such switches can be used in the design of multivibrators, monostable circuits[2] and similar devices it is probably less obvious how they may be applied to LC sine-wave oscillator design. It is the object of this article to show that such oscillators may be profitably based on these waveform and switching concepts.

The conventional approach to oscillators is based on linear amplifier concepts. For analysis it is assumed that just sufficient of the output of an amplifier must be fed back in the correct phase to the input. In practice, much more positive feedback than this is generally used and amplifier non-linearity effectively reduces the loop gain to unity at some, usually ill-defined, amplitude of oscillation. A form of non-linearity which is commonly used is automatic Class-C bias. As oscillation amplitude builds up, grid current biases back the amplifier until equilibrium conditions are reached. As poorly defined non-linearities are thus involved in the operation of conventional oscillators, any attempt at their detailed quantitative design is both difficult and of doubtful value. It should not be surprising therefore that oscillator design can be profitably approached from the point of view of switching circuits with defined non-linearities. The design procedure then amounts to considering the desired sine-wave output as a waveform to be generated by the proper application of waveform circuit principles.

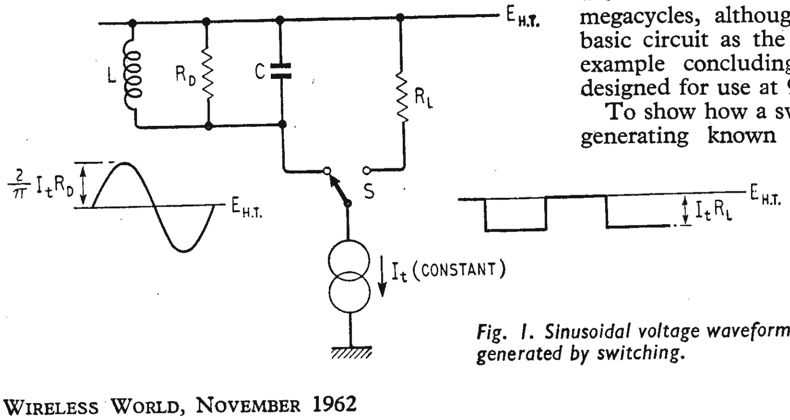


Fig. 1. Sinusoidal voltage waveform generated by switching.

Precision Design

The use of bottoming transistor switches makes possible LC sine-wave oscillators with exceptional power conversion efficiencies, such oscillators being first described by P. J. Baxandall.[3] Long-tailed-pair switches may also be used in oscillators which, whilst less efficient are simpler and so more suitable for general use.

It will be shown that these oscillators are in many ways superior to conventional oscillators. In particular they are *designable*. They may be designed to give a waveform whose frequency, amplitude, and harmonic content are determined mainly by the *passive* components of the circuit. This statement must be qualified by remarking that the shortest time in which it is possible to operate a valve or transistor switch eventually becomes an appreciable fraction of the cycle time, as the working frequency is raised. The operation then depends more on the characteristics of the active elements used and a simple theory gives less accurate results. It is not possible to state absolutely the maximum frequency at which the design methods for long-tailed-pair oscillators cease to be valid as this depends on the degree of accuracy required and the frequency limits of the active circuit elements. As a general guide the techniques to be described are most useful at frequencies up to a few megacycles, although an oscillator using the same basic circuit as the 100kc/s oscillator in the design example concluding this section was successfully designed for use at 90Mc/s using simple theory.

To show how a switch can be used in an oscillator generating known sinusoidal waveforms consider Fig. 1. A source of constant current, I_t is switched by the switch S into either of two paths. In one path there may be a load resistor, R_L and in the other is a parallel tuned circuit. Now suppose that the switch S be operated at a frequency corresponding to the resonant frequency of the tuned circuit given by:—

$$f = \frac{1}{2\pi\sqrt{LC}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

The current waveshape fed to the parallel tuned circuit will then be a square wave with a peak-to-peak value of I_t. By Fourier analysis this square wave may be represented as the sum of a sine wave at fundamental frequency whose peak amplitude is:—

$$\hat{i}_1 = \frac{2}{\pi} . I_t \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

and sine waves at odd harmonic frequencies:—

$$\hat{i}_n = \frac{2}{n\pi} I_t \text{ where } n = 3, 5, \text{ etc.} \quad \ldots \quad (3)$$

Given a tuned circuit with a reasonable value of Q (say 10 or greater) the harmonic components will be effectively by-passed, but a resonant dynamic resistance, R_D say, will be presented to the fundamental component, i_1. Thus a nearly sinusoidal voltage will be generated whose peak value is given by:—

$$\hat{v}_1 = \frac{2}{\pi} . I_t R_D \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

In Appendix 2 it is shown that the harmonic voltages are given by:—

$$\hat{v}_n = \hat{v}_1 . \frac{100\%}{(n^2 - 1) Q} \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

Where $\hat{v}_n$ is the nth harmonic voltage (peak)

$\hat{v}_1$ is the fundamental voltage (peak)

Q is the working Q of the tuned circuit.

Therefore, for Q = 10 say, the voltage across the tuned circuit is very nearly sinusoidal, with a third harmonic content of $\lessdot$1.3% and with higher order harmonics still more heavily attentuated.

It has been shown that the operation of the switch gives square waves of current which are converted to sinusoidal voltage in the tuned circuit. The system becomes a true oscillator when this sine-wave is used to operate the switch, giving self-maintained sine-wave oscillations. Provided the working Q of the tuned circuit is known, the amplitude and harmonic content of the sine-wave produced, as well as its frequency, will be known. An ordinary valve oscillator loads its tuned circuit with a variable R_a and grid current, making the working Q ill-defined. It is therefore worth noting that a generator of constant current, when connected to the tuned circuit, does NOT load it and it can also be arranged that the regenerative connection operates the switch with negligible loading on the tuned circuit.

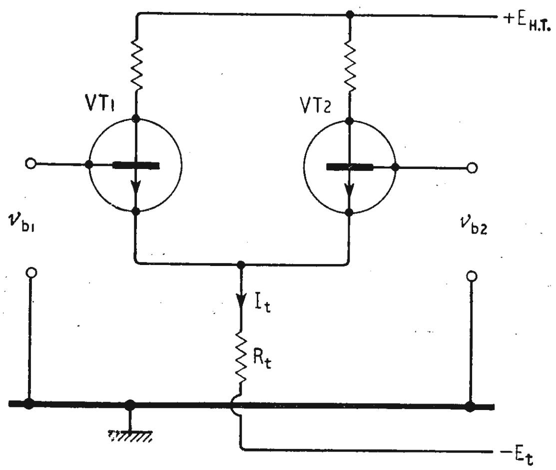


Fig. 2. Basic long-tailed pair.

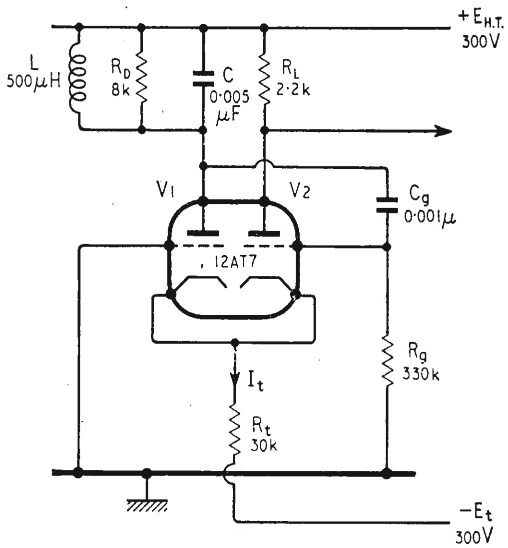


Fig. 3. 100kc/s sinewave oscillator.

Either a valve or transistor long-tailed-pair may be used in practice in place of the switch S and the constant current source shown in Fig. 1. As the authors in their previous article[1] described the operation of a basic circuit using valves, the brief description here is related to the transistors of npn polarity shown in Fig. 2.

In Fig. 2, suppose R_t is connected to a negative supply voltage large compared with the values of v_{b1} and v_{b2} when the circuit is operating. The current I_t will be given very nearly by $I_t \simeq E_t/R_t$ independent of the remainder of the circuit. This current is switched between VT1 and VT2 by varying the base-to-base voltage. Suppose the base of VT1 is earthed, that is, v_{b1} is zero. If VT1 and VT2 are assumed to be germanium transistors, the base-emitter voltage to just cut off either transistor may be taken as zero volts. The base-emitter voltage necessary if a transistor is to conduct current I_t may be found in characteristic curves and is typically 200mV. Therefore if v_{b2} is made +200mV say, the common emitter potential will be zero volts and VT1 will be just cut off. Any further increase in v_{b2} produces little change in I_t while v_{b2} is still small compared with $-E_t$. Conversely if v_{b2} is made negative by about 200mV, VT2 will be cut off and VT1 will conduct I_t which will not change while v_{b2} is made still more negative. It is convenient in long-tailed-pair circuits to define a parameter v_s, the voltage necessary to just hold the switch in one direction*.

With germanium transistors the value of v_s may be often taken as about 200mV. With silicon transistors the value of v_s is usually similar. While V_{BE} for a silicon transistor conducting common values of I_t is about 0.7V, V_{BE} to cut off the transistor is about 0.5V. The switch operates as before but with the common emitter voltage about 0.5V lower.

It should be noted that a transistor pair operates in

*The parameter v_s corresponds to the "effective grid base", $(e_c - e_b)$ of Ref. 1.

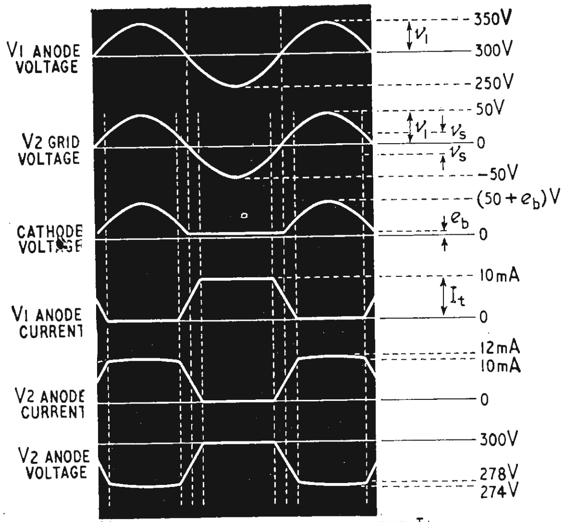


Fig. 4. Waveforms of circuit of Fig. 3.

a similar manner to a valve pair, in that it is the differential input voltage which operates the switch. In a properly designed valve pair, grid current is avoided by ensuring that the ON valve has sufficient anode-cathode volts to conduct I_t while still operating with negative grid bias.

A transistor pair cannot be operated without base current in the switch input circuit. This current is not, however, directly analogous to valve grid current because the base current in the ON transistor, given by $I_t/(\beta + 1)$, (where β is the common emitter current gain of the transistor) will be substantially constant for I_t constant. Also, for a transistor with known minimum β the maximum base current which the circuit must tolerate is known if care is taken to avoid bottoming the transistor. Just as a valve pair must not " run out of anode volts " so a transistor pair must avoid bottoming its transistors.

Practical Circuit

The operation of a practical oscillator circuit should now be clear. Fig. 3 shows an oscillator using a double triode valve. The circuit in this form which resembles a White multivibrator is due to P. B. Vanderlyn and E. L. C. White[4]. When the grid of V2 is more negative than the grid of V1 by an amount greater than v_s, V1 conducts the tail current. Conversely, with the grid of V2 more positive than the grid of V1 by the same amount, V2 conducts the tail current. A sinusoidal voltage is coupled to the grid of V2 by C_g and R_g with time constant $C_g \cdot R_g$ much greater than the period of oscillation. If the amplitude of this sine wave, $\hat{v}_1$, is much greater than v_s, anode current waveshapes in V1 and V2 will be approximately square. In V1, the amplitude of the current waveshape will be independent of $\hat{v}_1$ and for all practical purposes independent of anode voltage provided this is always sufficient to avoid grid current. With a tail resistor as shown in Fig. 3, the anode current in V2 rises to a maximum value of approximately $(\hat{v}_1 + E_t)/R_t$ as the common cathode point follows the positive swing at the grid of V2. Care needs to be taken to ensure that V2 has sufficient anode-cathode voltage to conduct this current without grid current flowing, operating with negative grid bias at all times.

Before considering how the values shown on Fig. 3 are determined in a design example it is worth comparing the operation of this circuit with the idealized switch configuration shown in Fig. 1. Firstly, no current flows in the grid circuit of V2 apart from the current in R_g. The assumption that a practical switch may be operated with negligible loading on the tuned circuit is therefore justified. Secondly, the closely defined anode current of V1 has a very high source impedance. (It may easily be shown that it is given by $R_a + (\mu + 1) R_t$ for V1 conducting I_t and naturally tends to infinity with V1 cut off.) As shown in Fig. 4 the waveshape of this current has an approximately trapezoidal form with rise and fall times depending on the value of $\hat{v}_1$ relative to v_s. By Fourier analysis[5] it may be shown that for $\hat{v}_1 = 10v_s$ the difference between the fundamental component of this trapezoid and the fundamental component in a square wave of the same amplitude amounts to about a half per cent. Even with $\hat{v}_1 = 3v_s$ the error is still less than six per cent. Harmonic components are affected to a greater extent giving a harmonic content less than the predicted value.

As a design example the component values shown on Fig. 3 may now be calculated. Suppose it is desired to generate a sine wave of 50V peak across a tuned circuit resonant at 100kc/s. At this frequency, the dynamic resistance including the applied load is to be 8kΩ.

With the values of inductance and capacitance shown this corresponds to a loaded Q of 25 as $\omega L = 320\,\Omega$ at 100kc/s. To obtain 50V peak it will therefore be necessary to drive the tuned circuit with a fundamental component of current of just over 6mA peak. From equation (2) it is seen that this corresponds to a tail current of 10mA. This in turn sets R_t at 30kΩ with a −300V supply. The coupling components C_g = 1nF and R_g = 330kΩ apply the 50V peak sine-wave to the grid of V2. For the 12AT7 v_s for I_t = 10mA is in the region of 5-8 volts making the trapezoid currents a sufficiently good approximation to the assumed square wave. Using equation (5) the third harmonic content in the output may be estimated. Putting $n = 3$ and $Q = 25$ shows that the third harmonic voltage will be about ½% or less. Up to this point the design has scarcely needed to consider the valve parameters at all. As a final check though, valve curves should be consulted to ensure that no grid current flows. V1 must conduct 10mA with a minimum anode-cathode voltage of 250V while V2 must conduct just less than 12mA with an anode-cathode voltage of about 220V. This performance is within the capabilities of a 12AT7 valve. Often practical circuits can be made to have large safety margins and work satisfactorily with valves whose performance is well below nominal limits.

When this circuit was built using the values shown in Fig. 3 the amplitude of the sine wave generated was measured using a suitable waveform monitor and found to be 50V peak. This was perhaps a little fortuitous; tolerances on R_t and R_D of about 5% might have given a result anywhere in the range of about 45 to 55 volts peak.

The similarity between this circuit and the White

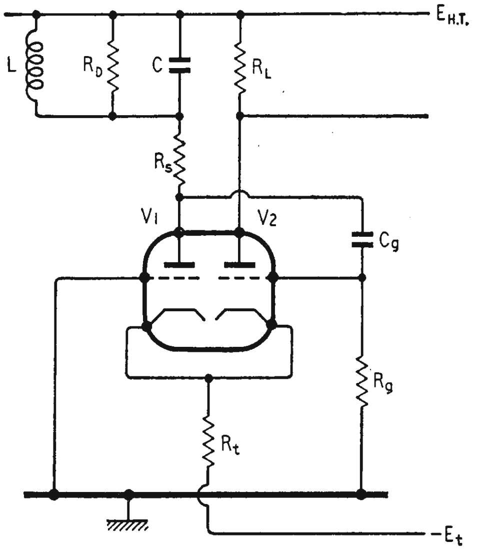


Fig. 5. R_s gives improved square wave at V2 anode.

multivibrator has already been noted. In the application for which this circuit was intended an approximately square wave locked to a stable frequency was required and good use was made of the waveshape obtainable at the anode of V2. In the language of long-tailed-pairs this is a "free" anode taking no part in the regenerative action of the oscillator so the output taken from this point is well isolated from the rest of the circuit. It is worth noting that with a second tuned circuit the fundamental or an odd harmonic at this point could be selected. With the resistor R_L 2.2kΩ as the anode load for V2 the output voltage I_t R_L, is about 22V for nominal I_t, and its rectangularity can be improved in several ways. To prevent the "bowing" of the base of the waveshape due to change in I_t the resistor R_t could be replaced by a valve giving a better approximation to a constant current although this refinement is not usually necessary. To obtain more rapid rise and fall times the circuit can be slightly modified in a way due to F. W. Cutts[6] as shown in Fig. 5. The additional resistor R_s placed in series with the tuned circuit is small compared with R_D but sufficient to give a step of voltage I_tR_s greater than $2v_g$. This step of voltage switches the circuit by a regenerative action similar to the White multivibrator. The rise and fall times at the anode of V2 are then limited only by stray capacitance.

Just as the conventional LC oscillator exists in many basically similar variations[3] so the long-tailed-pair may be made the basis of a whole family of oscillators. For example, the use of inductive coupling is an obvious step which makes the design procedures more flexible and is particularly useful in transistor variants of these circuits, as will be shown next month.

This section may be profitably concluded by listing the principal advantages of the long-tailed-pair family.

(1) The circuit can be easily designed to give a known waveform including a good estimate of the harmonic content. The performance is largely independent of the parameters of the active elements used.

(2) The fundamental/harmonic content ratio of the current waveform fed to the tuned circuit is higher than in many conventional oscillators where this waveform is a Class-C pip. The maximum odd harmonic content is independent of the loop gain and even harmonics are ideally entirely absent and in practice very small. As well as lower harmonic content this also means that LC ratios are often lower than in conventional circuits.

(3) The full Q of the tuned circuit may be realized because neither the circuit feeding current to it nor the regenerative path with its limiting action place any appreciable load upon it.

(4) Because valve or transistor parameters are less involved in the maintenance of oscillations and because LC ratios tend to be low, frequency stability can be exceptionally good.

(5) An auxiliary output largely isolated from the remainder of the circuit may be made available. A square wave can be conveniently obtained at this point.

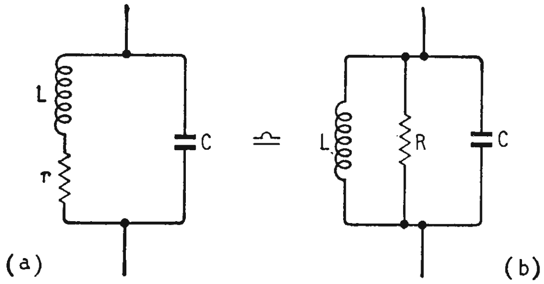
(6) In some forms of the circuit only a single untapped inductor with a single parallel tuning capacitor are required. If inductive coupling is used tight coupling rather than coupling through an arbitrary mutual inductance will usually be employed.

(7) The circuit cannot "squegg" as the limiting action has no associated time-constant.

(8) The amplitude of oscillation is directly proportional to tail current over a useful range. This property is used in a design example for an amplitude modulated transistor oscillator to be described next month.

APPENDIX I

An approximate relationship between series and shunt damping resistance in a parallel tuned circuit is useful for $Q \geqslant 5$ (a full discussion of this and similar relationships is given in *Electrical Engineering Circuits* by Skilling, and other textbooks).



The dynamic resistance at resonance of circuit (a), R, is given by:—

$$R = \frac{L}{Cr}$$

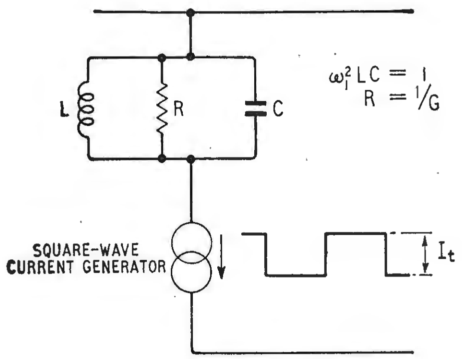
The magnification factor:—

$$Q = \frac{\omega L}{r} = \frac{1}{\omega Cr}$$
$$R = Q\omega L$$
$$R = Q \cdot Qr$$
$$R = Q^2 r$$

The series loss resistance r can therefore be represented by a shunt loss resistance R. With the oscillator circuits described any load coupled to the tuned circuit appears reflected in parallel with R and the total shunt resistance R_D can easily be calculated. In many practical oscillators the unloaded Q of the tuned circuit is made large and R may often be neglected in comparison with the reflected load or a deliberately applied parallel damping resistor.

APPENDIX II

The relative amplitudes of harmonic voltages across a parallel tuned circuit excited by a square wave of current may be derived as follows:—



Let n be the harmonic number.
i_n be the peak value of the n^{th} harmonic current.
Y_n be the admittance of the circuit at the n^{th} harmonic frequency.
B_n be the susceptance of the circuit at the n^{th} harmonic frequency.

$$B_n = B_C - B_L$$

$$B_n = n\omega_1 C - \frac{1}{n\omega_1 L}$$

$$B_n = \frac{1}{n\omega_1 L} \cdot (n^2\omega_1^2 LC - 1)$$

But $\omega^2 LC = 1 \therefore B_n = \dfrac{n^2 - 1}{n} \cdot \dfrac{1}{\omega_1 L}$

But $Q = \dfrac{R}{\omega_1 L} = \dfrac{1}{G\omega_1 L}$

$$\therefore B_n = \frac{n^2 - 1}{n} \cdot GQ$$

$$Y_n = \sqrt{B_n^2 + G^2}$$

$$Y_n = G\sqrt{\left(\frac{n^2 - 1}{n}\right)^2 \cdot Q^2 + 1}$$

At resonance $n = 1$

$$Y_1 = G$$

At harmonic frequencies

$$\frac{Y_n}{Y_1} = \sqrt{\left(\frac{n^2 - 1}{n}\right)^2 \cdot Q^2 + 1}$$

Q is typically 10 or greater and $n = 3, 5, \ldots$

$$\therefore \frac{Y_n}{Y_1} \simeq \left(\frac{n^2 - 1}{n}\right) \cdot Q$$

The harmonic voltage V_n is given by

$$V_n = i_n / Y_n$$

and $\dfrac{V_n}{V_1} = \dfrac{i_n}{i_1} \cdot \dfrac{Y_1}{Y_n}$

By Fourier analysis the peak value of i_n is given by

$$\hat{i}_n = \frac{2}{n\pi} \cdot I_t$$

$$\therefore \frac{\hat{i}_n}{\hat{i}_1} = \frac{1}{n}$$

$$\therefore \frac{V_n}{V_1} = \frac{1}{n}\left(\frac{n}{n^2 - 1}\right) \cdot \frac{1}{Q}$$

$$\frac{V_n}{V_1} = \frac{100}{(n^2 - 1)\,Q}\ \% \quad (n = 3, 5, \ldots)$$

For example:—
Taking $Q = 10$ the percentage third harmonic is:—

$$\frac{V_3}{V_1} = \frac{100}{8.10} = 1.25\%$$

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3. Baxandall, P. J., "Transistor Sine-Wave L/C Oscillators," *Proc. I.E.E.*, Vol. 106, Part B, Supp. No. 16 (Transistor convention).
4. Vanderlyn, P. D., White, E. L. C., British Patent Specification No. 540,941.
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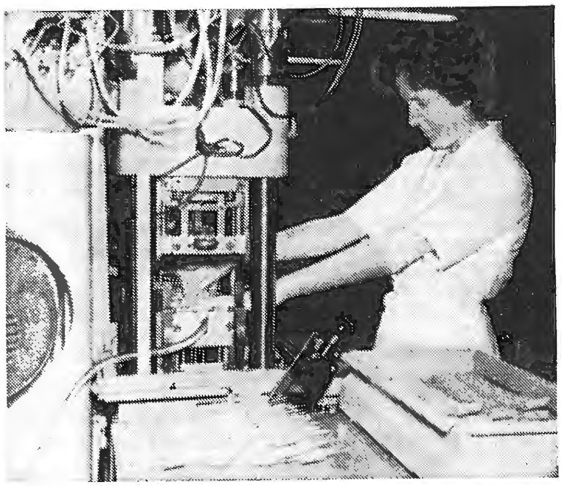
INJECTION MOULDED CAPACITORS

THE battle against ingress of moisture is continued by Dubilier with the introduction of their "Bluecon" and "Greycon" capacitors. Composed of paper and foil, and metallized paper respectively, the windings are impregnated with polystyrene and sealed, by injection-moulding, inside a pre-formed polypropylene shell. This process is considerably cheaper than the established cast-resin encapsulation, and has the added advantage over earlier types that the lead-out wires are supported over their internal length: high-frequency vibration failure is thereby avoided. Prior to encapsulation, the exposed metal at the ends of the windings is tinned and the lead-out wires butt-soldered in position. The inductance of the capacitor is greatly reduced and a very strong joint is obtained.



Capacitors being encapsulated at Dubiliers' Liverpool factory. "Bandoliers" of half-shells are loaded with windings: the space round the capacitor is filled and the other half-shell is formed by injection-moulded polypropylene.

LETTERS TO THE EDITOR

The Editor does not necessarily endorse opinions expressed by his correspondents

New Phase-Splitter

I READ with interest Mr. A. R. Bailey's article " New Phase-Splitter " in your September issue. I believe, however, that his circuit could be further improved by simple modifications.

The ratio of screen to anode current in a pentode varies somewhat from valve to valve, and also in the same valve in different parts of its characteristics. Negative feedback applied to the cathode of a pentode as in V1 of Mr. Bailey's Fig. 8 tends to ensure that the signal component of the cathode current is a faithful replica of the input signal. It would appear better to arrange the feedback so that the signal component of the anode current only is forced to follow the input signal. This can be done by removing the signal component of the screen current from the cathode resistor, i.e. by returning the screen decoupling condenser to the cathode instead of to earth. A similar argument could be applied to V2. This qualitative argument is admittedly inadequate to decide which arrangement would in fact give less distortion, and since it appears from inspection of the published characteristics of the EF86 that the screen current is a more linear function of the grid voltage than is the anode current, there is probably little to choose between the two arrangements on grounds of distortion alone.

If, however, the screen decoupling condensers of both V1 and V2 were returned to their respective cathodes as suggested, two other advantages would result:

(1) The input capacity of V1 would be very substantially reduced, since the signal potential of the screen as well as that of the cathode, would approximate to the grid signal potential. The input capacity of V2 would also be somewhat reduced.

(2) A balanced output would be obtained when the two anode loads of V2, were made equal instead of slightly unequal, moreover the balance would be less upset if valves of varying screen-to-anode current ratio were used for V2.

Bristol, 8. D. F. GIBBS
H. H. Wills Physics Laboratory,
University of Bristol.

I WAS much interested by Mr. Bailey's article on phase-splitters in your September issue. The long-tailed pair circuit described has many advantages but there is a rather odd snag that came to light in the course of developing a special amplifier using the more usual double-triode variation.

I found a very intractable hum in my amplifier. This was eventually traced back to the h.t. rectifier, which was choke-fed.

At the instant of conduction the transformer waveform is distorted and this results in a high-frequency pulse appearing in the heater circuit. Because of the high impedance in the cathode circuit of the phase-splitter this pulse can appear across the load via the heater-cathode capacity.

The effect is almost undetectable if the more usual capacitor-input rectifier circuit is used. A cure for it is the use of a separate heater transformer for this and other cathode-coupled stages.

With a choke-input filter the resultant hum (if this is the right term) is otherwise unacceptable but with a capacitor-input filter it may pass as background noise. However, an improvement in this can result from feeding any cathode-coupled stages from a separate heater transformer.

Swindon. T. S MARSHALL

I SHOULD like to discuss one or two points arising from Mr. A. R. Bailey's article on a " New Phase-Splitter " in the September issue.

(1) Can we safely establish the pentode and the triode at reasonable working points with the wide range of supply voltage quoted? I suspect that we may find the triode held very near cut-off. Certainly I am sure that there is no hope of getting the even-harmonic balance we get with a twin triode.

(2) The statement attributed to Crowhurst is a description of something which is obvious if stated more formally. Near instability implies that $(1-\mu\beta)$ in the gain expression, $\mu_f = \mu/(1-\mu\beta)$ is not the large number we usually have, but corresponds to the positive feedback in the region near the Nyquist point. With a 6dB gain margin, for example, $(1-\mu\beta) = \frac{1}{2}$ and the distortion of the basic amplifier at the 180° frequency is doubled. The feedback must be negative and maintained negative and high to the limit of audibility: the Duerdoth margin is a good guide.

(3) The long-tailed pair usually implies a high heater-cathode voltage. I prefer to raise the heated line to about +20 volts so that the first stage heater-cathode diode is biased into cut-off and heater emission cannot contribute to the current in the cathode resistor which is, so often, part of the feedback path. This reduces the stress on the l.—t. pair and also gives some improvement with some valves. There is a lot to be said for using an n-p-n transistor as the common cathode, resistor to give, since it can be in the common-base mode for a.c., a very high coupling impedance combined with low voltage drop.

I should be tempted to examine whether the pentode triode working points could be tied together by the sort of technique used in establishing the working conditions of starved amplifier, were it not for Mr. Bailey's Fig. 8. It is instructive to compare this with the circuit described by W. A. Ferguson in the May and June 1955 issues of *Wireless World*, in which an EF86 drives an ECC83 phase-splitter. (Fig. 1 p. 280 June 1955). As a point of detail, the EF86 screen is there decoupled to the cathode, but it is the anode circuit which is particularly revealing.

(4) Most circuits use a 100kΩ anode load for the EF86 and a 4.7kΩ resistor to give a 20 : 1 step. The difference between the two circuits is that Mr. Bailey uses a 500pF capacitor, whereas Mr. Ferguson used one of 47pF. Mr. Bailey has been worrying about the Miller effect at 25kc/s. Then he introduces a step circuit which is 3dB down at $\omega CR = 1$ with $C = 500.\ 10^{-12}$ and $R = 100{,}000$, so that $\omega = 2.10^4$ or around 3,000c/s. At 1200c/s he has about 75° contributed by this step circuit and has lost 12dB of his gain, and of his feedback. Moreover, at 25c/s his source impedance to the grid of the l.—t. pair is getting down towards the 4.7kΩ and if the overall response is sketched out it becomes apparent that the ECC83 Miller effect will begin to show some influence only in the region of 200-400kc/s. By this point, however, we are concerned with a whole lot of other factors like the CR response at the anodes of the l.—t. pair and the amplifier is, to use a metaphor painful to transistor circuit designers, back in the melting pot.

The loop characteristics of the 1955 circuit confirm this treatment and show that it is quite practicable to have

20dB of feedback left (from 30dB) at 30,000c/s. They also show, in textbook style, a positive feedback defect at the low frequency end, although it will be rare for us to be troubled by this 2c/s effect and easy to overcome it.

If it were necessary we could make use of a number of positive feedback tricks to reduce the effective input capacity of the l.—t. pair. The capacitance from A2 to G1 is obvious, but a capacitance from A1 to G2 with a 4.7 kΩ resistance in the G2 lead is an interesting variant which might repay study. However, if we use a step circuit, and since we shall need at least one we can do this, at the input to the phase-splitter the Miller effect is one of the least of our worries.

London, W.8. THOMAS RODDAM

IF Mr. Bailey cares to connect a 2pF capacitor from V2 anode to V1 grid in his circuit of Fig 5 he will find that the high-frequency disadvantages he mentions are eliminated.

Another place where neutralization is usually effective is in the output stage itself where, even when pentodes are used, a 1pF anode-grid capacitance is not uncommon. In this case a series 20-30kΩ resistor will prevent possible r.f. instability.

Woodford Green, Essex. P. A. DOUVALETELIS

IN his article in the September issue, Mr. A. R. Bailey states that the Concertina Phase Splitter suffers from unbalance at high frequencies due to the different output impedances at the anode and cathode.

It has been shown in this journal in a previous article (W. T. Cocking, Feb., 1948, p. 62) that provided the anode circuit and cathode circuit load impedances are equal in this circuit, the voltage outputs are equal. This is clear when it is realized that the anode current flows through both these circuits in series. Thus even at high frequencies where the circuit capacitances cannot be neglected, no voltage unbalance will occur with the usual circuit arrangements. The two outputs will fall off with " break point " frequencies

$$\frac{1}{2\pi R_L C_k} \quad \text{and} \quad \frac{1}{2\pi R_L C_a} \quad \ldots \quad \ldots (1)$$

where C_a is the anode-earth circuit capacitance and C_k is the cathode-earth circuit capacitance.

If $C_a = C_k$ (as it usually is since the major part of the capacitances consist of the grid-cathode capacitances of the following identical stages) the two outputs fall from the same frequency, remaining balanced. A calculation of the break point frequencies using the formula (1) shows that for typical values (100kΩ, 20pF) a frequency of order 100kc/s is obtained.

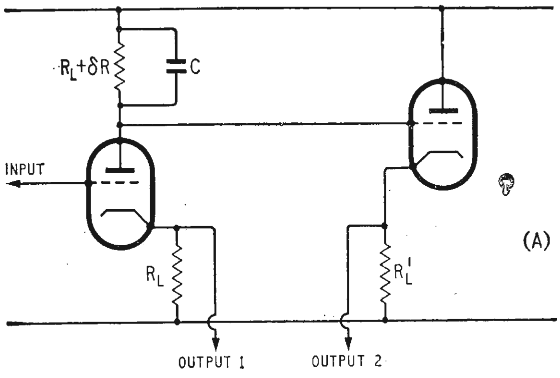
A further advantage of the concertina circuit is that it can be readily direct coupled to the previous stage (as shown in Mr. Bailey's article (Fig. 3)).

Analysis shows that Miller effect is negligible and there is a very high input impedance at the grid of the phase-splitting valve.

(Input impedance is approximately $2C_{ag} + C_{gk}/\mu$ which is approximately $2C_{ag}$ with the usual meaning for the symbols. As C_{ag} is of order 3pF for a triode, $2C_{ag}$ is usually negligible in comparison with the anode-earth capacitance of the previous stage.)

If it is desired to obtain equal output impedances then equal *low* output impedances can be obtained by using a cathode follower following the anode circuit, as shown in the accompanying diagram (A). This circuit will provide equal very low output impedances (of order 100Ω). The two outputs will remain balanced at high frequencies whilst the low output impedance is also of advantage in reducing the effects of " grid blocking."

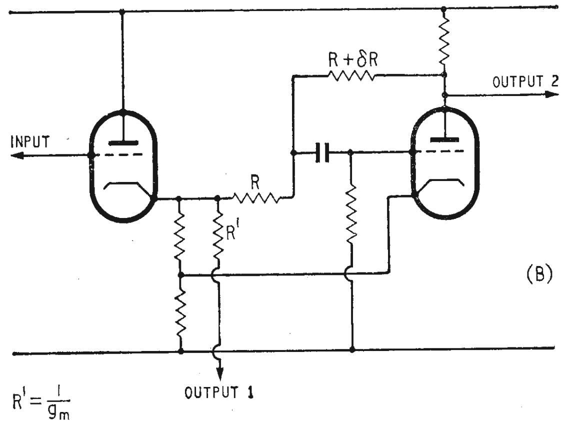
Mr. Bailey's article contains similar comments about the " floating " paraphrase phase splitter (Fig. 2 of the original article). Equal output impedances with very high impedance and no Miller effect can be obtained in this circuit with good balance at high frequencies by using a cathode follower as the first valve and (if it is required to have equal output impedances) a resistor in the output of the cathode follower to equalize the output impedances (B). Again very low output impedances are obtained. (Output impedance approx. $2/g_m$) of order 100Ω.



$$\frac{R_L+\delta R}{R_L} = \frac{\mu+1}{\mu}$$

C = CAPACITANCE IN PARALLEL WITH OUTPUT 1.

R_L' CHOSEN TO GIVE EQUAL CURRENTS IN BOTH TRIODES.



$$R' = \frac{1}{g_m}$$

These circuits are thus capable of providing most of the advantages claimed for Mr. Bailey's circuit whilst at the same time providing a lower output impedance. Mr. Bailey's circuit does however have the advantage of a higher gain.

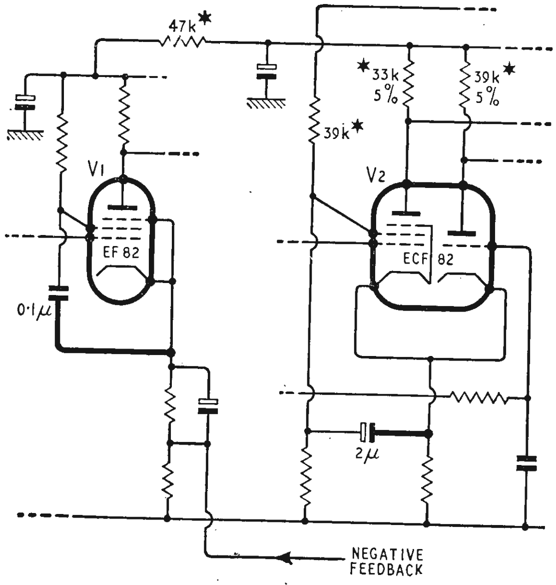
Cardiff. P. WILLIAMS
Department of Applied Physics,
Welsh College of Advanced Technology.

The author replies:—

IN replying to these letters I would first like to thank Mr. Gibbs for his helpful comments. I have tried returning the screen decoupling capacitors to their respective cathodes and there is an improvement as he predicts. In view of the advantages to be gained I would suggest making the modifications to the original circuit shown in the accompanying diagram, in which asterisks indicate altered values.

The bandwidth is slightly increased to 190 kc/s and the distortion is definitely reduced for large output swings. The effect of these two is barely noticeable on the overall amplifier performance but the balance obtained is far better for valves made by differing manufacturers. This enables the output valve distortion to be held to a minimum.

I was interested in Mr. Marshall's comments on choke-input filters. Luckily, however, the capacitor input filter is normally cheaper and therefore preferred.



In addition the input choke usually hums audibly due to magneto-strictive effects. Nevertheless it is a warning that is well worth bearing in mind.

Turning now to the letter from Mr. Thomas Roddam, I think that it will be preferable to deal with each of his points in turn.

(1) This may be due to a misprint (mine) that was corrected in the last edition of *Wireless World*. With the correct value of 220 kilohms for the anode feed resistor of V1 (or using the modified circuit just given) the two d.c. anode potentials of the phase-inverter balance to within 10 volts over the range of h.t. supply quoted. The distortion of the original circuit was far outweighed by that of the output valves and the distortion of the modified circuit is even lower.

(2) Unfortunately the matter is not quite as simple as Mr. Roddam would lead us to believe. The effect of near-instability that is well into the supersonic region can apparently be detected aurally. This subjective effect can be heard with amplifiers that have an impeccable performance within the normally accepted audio band of frequencies. I am not at all certain of the reason for this, but I would expect that the findings of D. L. Pimonow (reported in *Wireless World*, October 1962, p. 493) would have some bearing on the matter.

(3) I agree about the desirability of a positive bias on the heater-chain but feel that the addition of a n-p-n transistor and its associated components is unnecessary for an audio amplifier. There could easily be cases where such a circuit would have positive advantage but I do not think that this is one of them.

(4) In fact the circuit used commercially has a 40 to 1 step and uses a 2.7 kilohm series resistor. The 4.7 kilohm value was quoted in the article to allow for transformers of a lower resonant frequency. The step has been started fairly low down to obtain a high stability margin. If it is felt that a reduction in the h.f. distortion is preferable to the increased stability then the capacitor could be reduced to 250 pF and the series resistor increased to 10 kilohms.

Incidentally the loop gain of the Ferguson amplifier, referred to by Mr. Roddam, has fallen by 4 dB at 12 kc/s. This is more than would be expected from the values of the step network and the extra amount fits in with the predicted Miller capacitance.

I would definitely disagree with Mr. Roddam when he states that Miller effect is one of the least of our worries. "Know thy phase-shifts" has been my motto for feedback amplifiers and has always stood me in good stead. Once the transformer resonant frequency is reached there is then a severe overall phase retard. The step-network should have finished by this point or it will be adding further phase retard. An additional phase retard due to Miller effect in the phase-inverter can then well take the feedback loop close to instability even if not actually causing it. The effect of a capacitive load across the output further aggravates the matter, and it is well known that few amplifiers can stand appreciable capacitive load without instability.

If, however, the additional phase-shift due to Miller effect is absent, then the gain and phase margins can be made far better. One certainly cannot disregard phase-shifts about 400 kc/s; the gain margin of the amplifier in which this circuit is used is -26 dB but occurs at a frequency of 2.7 Mc/s. The phase margin is 90 degrees. Such performance could not be obtained with the normal long-tailed pair.

The last point is the use of positive feedback to remove the Miller effect. This is mentioned by both Mr. Roddam and Mr. Douvaletelis and was in fact the first thing that was tried in an attempt to improve the performance. (*Hi-Fi News*, July 1962, p. 90). Unfortunately it only made matters worse. This was traced to the unbalance in the input impedances of the output valves at high frequencies. This unbalanced loading of the phase-splitter causes unbalance in the drive to the output valves and consequently removes the neutralizing. The same is the case with neutralization of the output stage. So far I have found no advantage to be gained by neutralization and even if it could be made to work it would render the amplifier performance very susceptible to component changes.

So far as improvement in performance of amplifiers is concerned, the proof of the value of a modification is in the results obtained. By using this circuit it is possible to apply far more feedback without instability. With high resonant frequency transformers this improvement has been in the region of 15 to 17 dB. This, to me, seems to be a very worthwhile improvement.

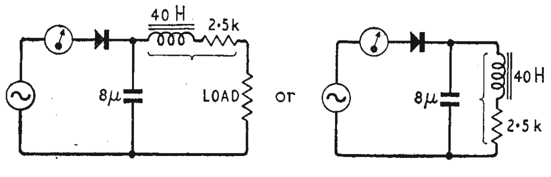
ARTHUR R. BAILEY

Rectifier Instruments

YOUR correspondent, Mr. J. Baggs (October 1962) speaks of the "serious errors" to which rectifier instruments are liable. This does not do justice to what are, in general, very useful instruments.

Rectifier instruments tell the "truth" (apart from frequency errors, errors due to rectifier non-linearity and those errors common to all moving-coil instruments) if they are scaled in mean (or average) quantities. The mean value of a sine wave is 0.9 times the r.m.s. value, so this would not suit most users, who would be disconcerted if their voltmeter indicated 216 volts when connected to the 240-volt mains. Consequently, as Mr. Baggs states, most rectifier instruments are scaled to indicate 1.11 times the mean value, i.e. the r.m.s. value on a sinusoid. For many purposes this is a reasonable and useful assumption, but "errors" are produced if the measured waveform is not the sinusoid assumed by the scaling.

Mr. Baggs' rectifier circuit would seem to have been:—



In either case the milliammeter is passing a considerable d.c. component. One suspects that the rectifier instrument was, in fact, a universal AVO meter. These instruments

(*continued on page 543*)

"Belling-Lee"

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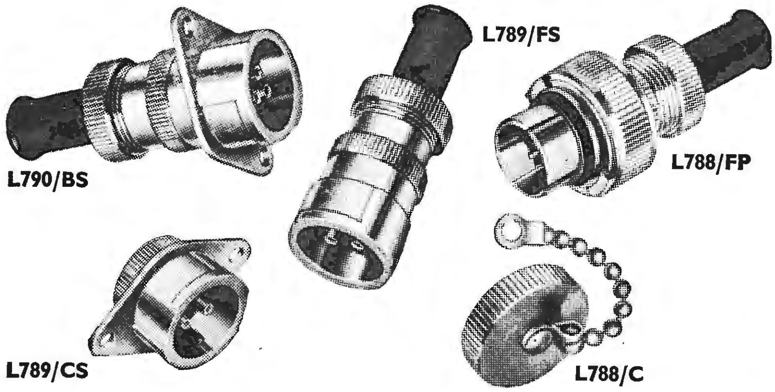
Lightweight Screened Plugs & Sockets



A range of lightweight screened connectors for low and medium frequency applications. The housings are polarised, and each accommodates either an internal plug or socket assembly, enhancing versatility. A spring-loaded locking ring is incorporated to prevent disengagement by accident, shock, or vibration, but screen continuity under static conditions is not impaired if the ring is left undone.

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Free Plug	L.788/FP	L.789/FP	L.790/FP
Joint Service No.	5935-99-943-8177	5935-99-972-8052	5935-99-911-1090
Free Socket	L.788/FS	L.798/FS	L.790/FS
Joint Service No.	5935-99-927-8082		
Bulkhead Socket	L.788/BS	L.789/BS	L.790/BS
Fixed Socket	L.788/CS	L.789/CS	L.790/CS
Joint Service No.	5952-99-943-8807	5935-99-972-8053	5935-99-911-1092

L.788/C
Protection Cover for range
(fixed members)

Current rating: 5A. **Breakdown voltage (D.C.) between poles** L.789 > 3kV. L.790 > 2kV., **poles/chassis** L788 5kV., L789 > 2kV., L790 > 2kV.
Insulation resistance: > 6×10^4 megohms. **Contact resistance:** < 2 milliohms.
Panel thickness: Mounting below panel 0.5-4mm. (0.020-0.156 in.) Can also be mounted above panel. **Cable size:** Up to 6.1 mm. (0.24 in.) overall. **Materials:** Aluminium housing; Synthetic rubber cable sleeve; Phenolic resin insulation; Contacts, silver-plated brass. **Average weight:** FP 13 gm. (0.46 oz.), FS 11.6 gm. (0.41 oz.), CS 6.2 gm. (0.22 oz.), BS 12.5 gm. (0.44 oz.).

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No. 44 of a series:

TERMINALS.

The basic requirements of a terminal are that it should make efficient connection, be robust, and easy to operate. It may or may not need to be insulated: sometimes a high order of insulation resistance may be required ($10^7 M\Omega$ or more) to avoid loading a high impedance circuit unduly, but a terminal, being a hook-up device, generally does not have to withstand a very high potential.

These requirements can be met by running a nut along a threaded stem onto a clamping platform. For ease of operation, it should be possible to tighten the nut securely with the fingers, without using tools. The clamping platform and nut can each be protected by insulation if need be, except in the clamping gap.

Such a terminal is simple and efficient, but can be improved by adding various refinements. For instance, when idle, the nut can become detached and lost; it is better to make it captive on the stem. Also, when attaching stranded conductors, difficulty is liable to be experienced by strands becoming trapped in the threads and jamming the nut, however tightly the strands are twisted initially. It is better to have the clamping gap free from threads, for there will be a tendency for strands to become separated as the nut is tightened, and even a solid conductor may become displaced. One way of preventing this is to fit a claw or cup washer; another is to provide a slot or cross-hole in the stem within the clamping gap—for a given width, a cross-hole weakens the stem less than a slot does.

A useful extra is the provision of a socket in the top of the terminal into which a wander plug can be inserted. However, perhaps the ultimate refinement is to provide the clamping zone with a means of piercing insulation, as has been done in the " Belling-Lee " Universal Terminal developed for the Services. This makes for maximum speed in connection by obviating the need for stripping the end of an insulated conductor, which can be a tricky operation under fire or in a tank on the move!

It may come as a surprise to some of our readers to realise that there are so many different facets to such a simple device, which most of us tend to take for granted. All these features are incorporated in " Belling-Lee " terminals, and are typical of the thought which goes into the design of all our components.

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use an internal current transformer on the a.c. ranges, the output from the secondary of which is rectified and displayed on the milliammeter.

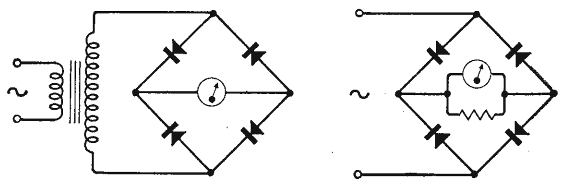
Hence only the a.c. component is indicated, whereas an r.m.s. sensing instrument (moving iron, electrodynamic, thermocouple, etc.) would indicate the r.m.s. value of the a.c. and d.c. components together. Further, the d.c. component in the primary of the c.t. might saturate the core, thus tending to produce an even lower reading of the instrument. A rectifier instrument not incorporating a current transformer *and calibrated on d.c.* would indicate correctly the mean value of the a.c. and d.c. components. Both r.m.s. and mean answers would be "correct": it would depend on the purpose of the measurement which was more useful.

It is not the nature of the circuit *per se* which causes rectifier instruments to read "wrongly" they may give an answer different from that of an r.m.s. sensing instrument merely because the r.m.s. and mean values *are* different, or for conventionally scaled instruments because the form factor (r.m.s. value/mean value) is not 1.11 as assumed in the scaling operation.

Hatfield, Herts. P. M. CLIFFORD
Standard Laboratory,
De Havilland Aircraft Co. Ltd.

THE letter from Mr. J. Baggs on rectifier meter errors is timely, but rather tends to overstate the case against this very useful instrument.

An error of 50%, while possible, is unusual, and must signify a pretty fantastic departure from a sine wave in the circuit quoted by him, since so extreme a case as 20% 2nd harmonic displaced by 180°, only results in an error of 9.2%, according to the published figures of a well-known rectifier manufacturer. It is true that a good moving-iron meter will read the r.m.s. of almost any low-frequency input, a.c. or d.c., or even a mixture of both,



(a) (b)

but it is surely asking a lot of a rectifier meter to interpret such a signal correctly in terms of mean value only.

The accurate measurement of current by means of rectifier instruments is, anyway, full of pitfalls for the unwary, not the least of which is the ratio of the impedance of the circuit to that of the instrument. A rectifier inserted into a low-impedance circuit will itself distort the waveform yet still indicate the mean of the rectified current on the moving-coil meter to which it is coupled, so that a calibration made on the assumption of a sinusoidal input no longer holds good. The lower the circuit impedance, the greater will be the error. This restriction does not, of course, apply to rectifier voltmeters, because the ohmic resistance usually incorporated "swamps out" the non-linear characteristic of the rectifier and hence its effect on the supply waveform.

Mr. Baggs does not specify exactly the type of rectifier meter he used, but for the measurement of currents above 10mA it is safer to use the better class instrument; that is, one which incorporates a step-down current transformer as in (a) rather than a directly coupled rectifier feeding a d.c. shunted moving coil as at (b).

The reason for this is that, as well as stepping up the current, the transformer steps down the impedance of the disturbing influence of the rectifier, and therefore gives a more accurate result, especially in low-impedance circuits. Where a step-down transformer cannot be used, i.e., in cases where the current to be measured is of the same order as that of the moving coil itself, a current reading by means of a rectifier meter can never be relied on if the total circuit voltage is less than 10 volts.

Highenden Valley, Bucks. T. H. FRANCIS

Sine Wave or Square

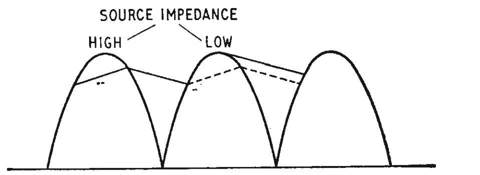
MR. VALCHERA has made, in the October issue, a plausible case for the use of square wave inverters for operating equipment normally driven by the sine wave of the a.c. mains. A misprint (65W for 650W) will probably not have deceived many readers. I am prepared to agree that there are few cases where equipment cannot be operated from square waves although I suggest that the small loss of power in a.c. motors should be read as an appreciable increase in running temperature.

Harmonics of the 50 c/s square wave can prove more troublesome than he suggests in some applications. Not only is there what many term the psophometric gain but in oscilloscope applications there is some danger of push-push high frequency energy appearing on the supply line and making its way into the supply lines inside the oscilloscope.

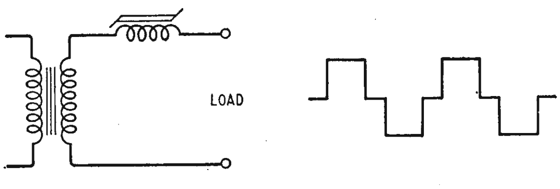
The most serious problem, however, is that although the equipment will accept a square wave the inverter will not produce one when the equipment is connected. I showed some little while ago how we could study the behaviour of an inverter by considering the interaction of a load line and a negative resistance "ratchet tooth," one of which is made to rotate by the introduction of a reactance in the circuit. If the equipment itself offers a reactive impedance to the inverter a self-oscillating inverter may operate in a false mode. This may be one in which the dissipation is high or the current demand excessive. The situation may be self-correcting or the load may be such that it cannot accept energy at the false mode frequency and the system will lock in this stalled condition. This effect is found with some motor drives and with fluorescent lamp and ballast circuits. A dangerous condition will be met, for example, if the filter shown in Mr. Valchera's Fig. 7 is left open-circuited by switching off the equipment without switching off the inverter, since this network presents a short-circuit at its upper band edge.

I cannot accept the significance of Fig. 2. The waveform which I find convenient for examination is that shown in the accompanying diagram (a) from which it will be apparent that even with a rather high impedance (by rectifier circuit standards) the reservoir voltage ripple is closely connected with the approximation to peak voltage.

An important solution which is not mentioned is the use of a saturable reactor to correct the waveform to that shown in (b). For a given load the r.m.s./peak ratio can



(a)



(b)

be chosen to be 1.4. Such a reactor can improve inverter performance, reduces the harmonic content of the output voltage by virtually removing all third harmonic and has other advantages outside the scope of a letter. The disadvantage is that it results in a pulsating current from the battery. The effect of tuning the battery line to the second harmonic merits examination.

Mr. Valchera will, I think, agree that square-wave inverters are by no means as simple as they look.

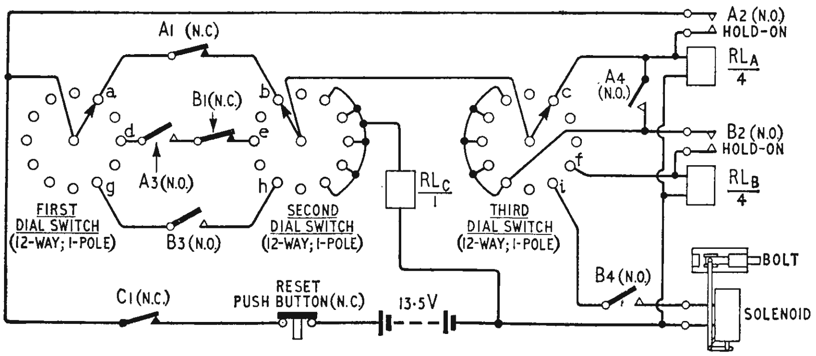
London, W.8. THOMAS RODDAM

Combination Lock

I READ with much interest the article by Dr. V. J. Phillips in the September issue, and have constructed the combination lock to the circuit given.

I find, however, that it is nothing like as difficult to "pick" as Dr. Phillips suggests. The fact that the relays are self locking and will not release until the reset button is pressed means that it is not necessary to pause at each combination. Thus if the intruder starts with the dials at 1.1.1. and continues dialing systematically (i.e. rotate dial 3 from 1-12, move dial 2 on one step, rotate dial 3 again and so on) he will in due course have set RL_A and RL_B and further systematic dialling will operate the solenoid.

The time this takes will depend on the actual digits used on dial number 1 (i.e. *a.d.g.*). If these happen to be in numerical sequence (say for instance 3.6.10.) the lock will be opened just before reaching the 1,444th combination. If *a.d.g.* are respectively 8.11.4 by the time the intruder has completed 1,728 combinations he will have set RL_A and RL_B and will only have to dial something less than 4 × 144 combinations to operate the solenoid.



Since it takes only 10 minutes to dial 1,728 combinations, the lock could be opened in about 13 minutes. At the very most it would take 30 minutes.

In actual practice with *a.d.g.* in sequence it took 7 minutes and with *a.d.g.* at 5.2.8. it took 15 minutes.

Fortunately, however, there is a simple modification which will render systematic dialling quite ineffective. This consists of adding a further relay which operates like the reset button if dials 2 and 3 are rotated sequentially. The circuit diagram (above) shows this modification; RL_C and contacts C.1 (N.C.). It will be seen that as soon as RL_A is locked on, continued rotation of dials 2 and 3 will soon operate RL_C and RL_A will be released.

Normal operation of the lock by the owner will not be affected since he would naturally move the dials straight from one combination to the next and would not be likely to have dials 2 and 3 set to any of the contacts which would operate RL_C.

It will be noted that the modification shown in Fig. 5 of Dr. Phillips' article has been incorporated. This is advisable as otherwise RL_A could be set prematurely.

There is one other point which should be noted. If the switches used are of the make-before-break type, contact *i* should not be adjacent to contact *f* as shown in Fig. 2, since dialling *a.b.f.* and turning switch 3 slightly would operate the solenoid.

Westcliff-on-Sea. K. G. HARLAND

The author replies:—

I WOULD like to thank Mr. Harland for pointing out something which I omitted to mention in my article, namely, the obvious undesirability of using make-before-break type contacts on the rotary switches.

His modification for foiling attempts at systematic dialling, which is similar in some respects to the burglar alarm system which I suggested, is an admirable one, and one well worth incorporating in a lock of this sort. With the connections as shown in his sketch, one must be careful to rotate the second dial in an anti-clockwise direction to avoid tripping off RL_C, but this is no disadvantage—indeed, it provides extra security and is to be welcomed on that account.

Another modification which would prevent sequential dialling (and which was in fact suggested to me by one of my students), consists of the connection of a thermal delay switch across RL_A. When RL_A has been set by dialling the first number, the thermal delay switch would begin to operate, and after a given time would open a set of "Reset" contacts, thereby de-energizing RL_A. A quick measurement suggests that a delay of about 10 seconds would be sufficient, since the second and third numbers can be dialled easily within this time. It is true that if the relay clicks could be heard, a systematic dialling procedure might still be possible by constant resetting of RL_A, but this would be a frustrating and time-consuming process. If all relays etc. were inaudibly mounted then operation by sequential dialling would be impossible.

Readers of my article may remember that I suggested various modifications in order to decrease the chances of the lock being opened, one method being the addition of another set of three dials. A thought which strikes me as I write is that it is not really necessary to have a complete extra set of dials to do this, because if a "holding-on" relay etc. is provided after the "*g.h.i.*" operation of the first three numbers, one can add an extra set of wafers to the three existing switches.

As I suggested in the original article, there must be very many modifications and improvements possible on the simple basic lock, and anyone building a lock of this kind will almost certainly wish to incorporate some ideas of his own.

V. J. PHILLIPS

"M. & B. Tablets"

"FREE GRID" has given me so much instruction and entertainment over the years that I feel diffident about taking issue with him now. But does he really consider Baird's transatlantic television accomplishment in 1928 a parallel to the spanning of the Atlantic by wireless telegraphy in 1901?

With respect, I submit that Baird's claim to fame rests (perhaps a little precariously) on his being the first to demonstrate that television was a practical proposition, and this feat is suitably commemorated on tablets.

The transatlantic television demonstration of 1928 was a horse of a different colour, for in this no new technical ground was broken. By this time Baird was working with thoroughly proven apparatus as far as the conversion of visual images into electrical impulses was concerned, and furthermore he had at his disposal a proven method of conveying them across the Atlantic for by 1928 radio-telephonic communication over such a distance was a commonplace. As his low-definition system demanded no more bandwidth than that for sound broadcast transmission, no problem existed in the matter of accommodating his signals on the carrier. In short, no element of technical uncertainty was involved.

As a parallel (dangerous word!)—if only red-painted aircraft had been flying the Atlantic for years, would one demand a memorial to the first blue-painted one to cross?

By contrast, Marconi, in 1901, was working in completely unknown territory. Although at the start of his career he, like Baird, had been using apparatus devised by others (with the notable exception of his application of the elevated aerial), the 1901 experiment marked a complete breakaway from the laboratory-type equipment previously used for transmission. The Poldhu transmitter, 100 times more powerful than any hitherto, was designed and built with no precedents to act as a guide; neither had anyone the slightest experience of transferring such a power into an aerial system—and beyond all else, none knew better than Marconi that he was attempting the theoretically impossible, in the light of then-existing knowledge. For sheer nerve and audacity it is doubtful whether the project has ever had an equal; as an experiment it must surely rank as one of the most significant ever made.

Nevertheless, so far as I am aware, no publicly subscribed or national memorial to Marconi exists in this country. The Poldhu obelisk was erected by the Marconi Company although it, and the Poldhu site, have subsequently been presented to the National Trust.

Chelmsford. W. J. BAKER

U.K. Citizens Radio Service ?

I HOPE that if Mr. Storey's suggestion (Oct. issue, p. 496) is followed up, any new service will be started on some band other than 26.96—27.28 Mc/s which has for many years been used for the control of models.

This hobby/sport is followed by many thousands of enthusiasts, in fact a recent *international* contest was held for radio control model aircraft at R.A.F. Kenley. This was supported by teams from no less than thirteen countries. This indicates not only an international following, but a world-wide use of this band.

This is not a novelty pastime; among these participants are many boffins who have, and are, contributing usefully to the full size research programmes by the use of their "model" control systems.

Watford. W. P. HOLLAND
Editor, *Radio Control Models & Electronics.*

THE remarks made by G. E. Storey on a U.K. Citizen's Radio Service in the October issue cause me to comment.

I should like to remind him that the band 70.2 to 70.4 Mc/s is at present available for use by amateurs and is becoming increasingly popular as a substitute for the old 5-metre amateur band now taken over for television. Any attempt to use this band for any other purpose than by qualified amateurs would meet with vigorous opposition.

Regarding his remark on Bands I, II and III interference surely, on a properly designed transmitter, the problem of second harmonic interference hardly applies.

The lowering of the technical standards of the transmitter as the writer suggests and the possible generation of spurious emissions should be avoided at all costs in these already crowded and noisy radio-frequency bands. Domestic television receivers are bad enough with their interference radiations.

Southampton. J. H. CROYSDALE (G3OZV)

AS a radio technician employed by a Government department I frequently operate v.h.f. transmitters in the course of my duties. Should I purchase one of these sets from one of your advertisers as Government surplus, I am not permitted to operate it for amateur transmission without passing the G.P.O. morse test.

A variety of unskilled business radio operators are free from this restriction. A correspondent in your columns recently even suggested inaugurating a "Citizen's Radio Band" to operate in the amateur band.

On the h.f. amateur bands morse is necessary to avoid interference with other services. But surely it is an anachronism on v.h.f.?

Could not the Post Office consider permitting "telephony only" amateur operation in a segment of the 70.2—70.4 Mc/s band for a trial period?

Shrewsbury. E. O. T. SABIN

Information Classification and Retrieval

I AGREE with your correspondent Mr. Tams, writing in the October issue; the problem of filing references into an information system is a most important one, and was insufficiently emphasized in my paper.

This matter did however receive careful consideration when we devised our system, and I do not agree with Mr. Tams that it need necessarily be unjustifiably expensive. The edge-notched card system has some advantages in this respect.

The following are the salient points of the classifying procedure—one very important objective being "spreading of the load."

1. Relevant papers, articles, etc., are brought to the attention of a number of senior engineers each of whom holds blank cards, and a copy of the index. This is done by circulating certain journals, and by the assistance of an American high-speed abstracting service; each engineer covers a range of subjects.
2. Abstracts are written or typed on to cards, subject code numbers are written in, and the appropriate "common variable" and "composition" holes are marked in pencil.
3. The cards are passed to the system operator who slots the code number and other holes, also the date, and the author's name.
4. The cards are scrutinized by the "editor," a senior engineer with an overall technical background, and are returned to the operator who places them in a drawer in any order.

In many organizations, and previously in this one, circularized journals tend to be read casually, if at all, and only carefully when the subject is of particular interest to the reader—any notes made become part of a "private" system. Now, the engineer reads selectively in the general interest, and his notes are made on a card in the form of an abstract. The system lends itself well to this, because classification, which needs no further processing (except slotting) can be carried out at a number of remote points—a considerable advantage for "load spreading" in the small organization. The costs of this operation are only marginally greater than the "private system" method.

Much work is going on towards the elimination of the index in information systems at N.P.L., in the United States and elsewhere. Mr. Tams has mentioned the wasteful operation of coding subjects and the possible use of syntax etc., arising from a study of semantics.

The writer has observed systems incorporating such ideas[1, 2], but has not so far observed nor can visualize a method which is readily self-evident to engineers, can cope with the synonym and order of language difficulty, and which does not involve expensive equipment or a department whose sole concern is classification and retrieval.

London, W.5. A. E. CAWKELL
Cawkell Research & Electronics Ltd.

[1]Cline: "An Edge-notched Index Card System for Mechanical Sorting", paper read at the American Chemical Society meeting, New York, 1954.

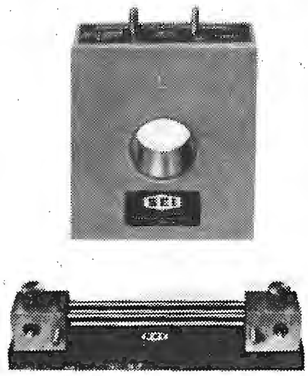
[2]Thoma, J.A.: "Simple and Rapid Method for Coding Punched Cards", *Science*, Vol. 137, No. 3526, page 278 (1962).

MANUFACTURERS' PRODUCTS

NEW ELECTRONIC EQUIPMENT AND ACCESSORIES

Selectest Range Extension

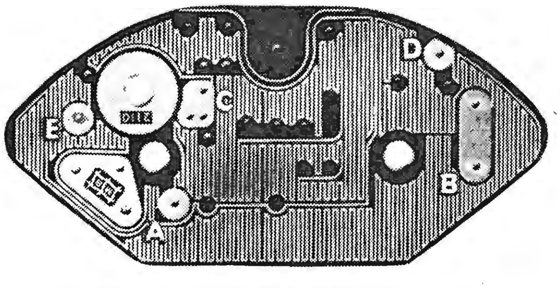
DESIGNED for use with the Salford Selectest test-meters, a number of accessories have been introduced to extend current measuring ranges. Encapsulated transformers extend the a.c. range of the Super K and Super 50 models to maxima of 300A and 250A respectively, while the new d.c. ranges are 750A and 250A, obtained by means of alloy shunts. Both shunts and transformers are designed for tropical conditions, and the accuracies are to B.S.89/1954. The meters and accessories are manufactured by Salford Electrical Instruments, Ltd., Peel Works, Silk Street, Salford 3, Lancs.



Current-range extension transformer for the Salford Selectest (top). Ratio is selected by the number of conductors threading the core. The shunt for use at d.c. is shown below.

Printed Components

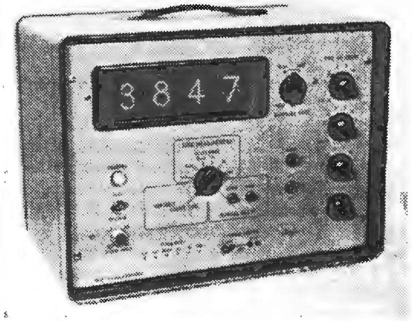
INDUCTORS, resistors and capacitors may be printed on a new type of board announced by Mills and Rockleys Electronics Ltd., Swan Lane, Coventry. The printed boards are double-sided, the foil used being either copper or resistive material such as cupro-nickel or nickel-chrome, and a further extension of this practice is the production of step-variable resistors, using graphite tracks. The example shown carries three resistors between points AB, AC and DE, the reverse side carrying wiring and a heat sink for a power transistor.



Printed resistor network by Mills and Rockleys.

Digital Tachometer

A LARGE proportion of digital counters and frequency-meters are usable as revolution-counting instruments if one is prepared to perform the arithmetic necessary to convert the reading into r.p.m., as the counting time is based on a decimal system. An instrument which fulfils both the conventional frequency and time measurement functions and which also has the advantage of a variable time-base for the measurement of other parameters has been introduced by Southern Instruments, Frimley Road, Camberley, Surrey. The time base is variable between 1 msec and 10 sec in 1 msec steps, and frequencies up to 120kc/s can be handled. "Start" and "Stop" signals can be applied from an external source and clock pulses are available at the front panel. A "delta" type of display is used, wherein the in-line display changes only when the count has reached a different total.

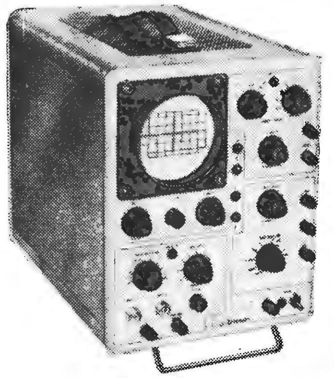


Southern Instruments MI 157 Counter.

General-Purpose Oscilloscope

CONTINUING their series of low-priced laboratory oscilloscopes, Dartronic have introduced a wide-band instrument (0-15 Mc/s, very conservatively rated) which offers all the flexibility and most of the trimmings of far more expensive equipment. Time-base speeds are from 0.2μsec to 1sec/cm, adjustable by a single control (the photograph shows a prototype on which two knobs were used), and all the usual triggering and synchronizing modes are available together with a manual single-shot switch. A 4kV p.d.a. flat-faced tube is employed, and power supplies are stabilized. If a reduction in bandwidth can be tolerated, a built-in pre-amplifier provides a sensitivity of 10mV/cm instead of 100mV/cm from 3c/s to 6Mc/s.

The standard Model 415 is accompanied by the 417

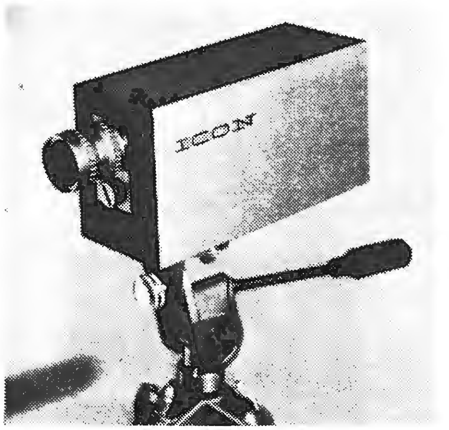


Dartronic 415/417 15Mc/s oscilloscope. Panel lamps to right of screen are beam locators.

which incorporates a 250nsec delay line. A leaflet can be obtained from Dartronic at 3-7, Windmill Lane, London, E.15.

Domestic Television Camera

CLOSED circuit television with "no knobs" is announced by Nottingham Electronic Valve Co., Ltd., East Bridgford, Notts. Known as the "Nev Icon," the new camera is either mains or battery powered, and works on 405-line or 625-line, 50c/s, or 525-line, 60c/s standards, the output being 5 mV on any of the Band I frequencies. Transistors are used exclusively and the camera is designed to take standard 16mm cine camera



NEV ICON low-cost c.c.t.v camera. Body size is $4\frac{1}{4} \times 3\frac{1}{4} \times 8$in.

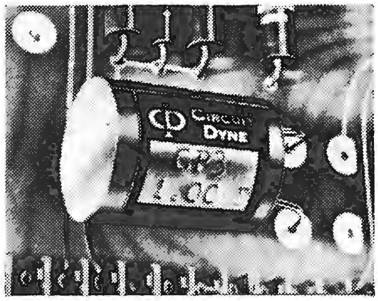
lenses. The price of the camera minus lens is £72, which brings it within range of domestic and educational use.

New Moving-Coil Microphone

THE familiar Acos Mic 39 is now available with an alternative to its normal crystal insert. This alternative is an (omni-directional) moving-coil system which includes a transformer to provide alternative high and low-impedance matching. The sensitivity is -80dB referred to 1V/dyne/cm^2 on open circuit for the low-impedance (200Ω nominal) connection and -54dB (also referred to 1V/dyne/cm^2) on open circuit for the high-impedance (50kΩ nominal) connection. The frequency response is within ± 3dB from 80 to 10,000 c/s and approximately 10dB down at 50c/s and 15kc/s. This microphone costs £7 10s and is manufactured by Cosmocord Ltd., of Eleanor Cross Road, Waltham Cross, Herts.

Current Regulator

A CURRENT dual of the established Zener diode is announced by the CircuitDyne Corporation, a subsidiary of Telonic. These devices exhibit a pentode-like current characteristic which is very similar to the Zener diode voltage characteristic, and are available in stabilized current ratings in the 5% range 1-20mA. Both symmetrical and polarized devices are obtainable, the CP2 and CN2 series operating from -55°C to 125°C

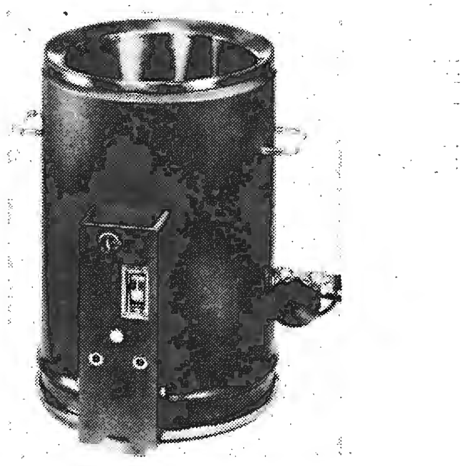


CircuitDyne Currector current stabilizers. Units are encapsulated and lead-out wires are either axial or single-ended.

at voltages up to 40V, and the CP3 and CN3 series in the range -55°C to 85°C at voltages from 2V to 25V. Current regulation of both series is ± 1% over the voltage range.

Fluidised Heating Bath

TRADITIONAL methods of heating electronic or electrical components for behaviour study at high temperature have relied to a large extent on immersion in liquids—oil or molten metal—or solid metal blocks. Ovens are also often used, but have the disadvantage that access to the component under test usually means that temperature control is lost. Liquids are not suitable for all purposes; for instance, oil tends to be dangerous at high temperatures, and molten metal can obviously not be used at temperatures below its melting point. Techne (Cambridge) Ltd. have developed, in

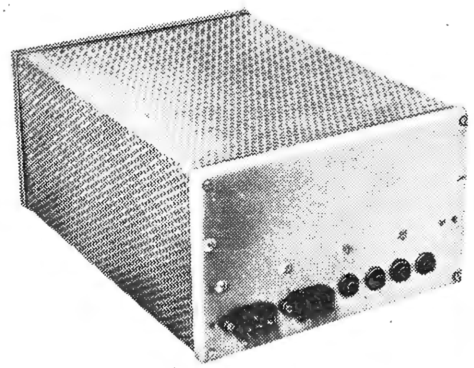


Tecam fluidized heating bath. Effective working space is about 20cm diameter and 15cm deep.

conjunction with H. Sutcliffe of Bristol University, a method of using sand, transformed by air pressure into a fluid. Air at 2 p.s.i. is forced through a porous plate upwards through the sand, which takes on the appearance of a boiling liquid. Heaters under the sand raise the temperature to a controllable maximum of 350°C, and this can be achieved in about 38 minutes. Temperature differential can be held to within ± 0.4°C. The heater, named the Tecam, is available from Techne (Cambridge) Ltd., at Duxford, Cambridge.

Frequency-stable Inverter

INTENDED to drive tape-recorder motors for computing and telemetry applications, a new inverter by M.L. Aviation Ltd. will provide sinusoidal outputs at either 400c/s or 60c/s within 0.01%. The master oscillator is crystal controlled, and feeds a chain of binary dividers, whose outputs are used in "AND" gates to fire silicon controlled rectifiers at the correct times to provide a three-phase output. Efficiency is about 70% and the unit will work from 28V d.c. to provide voltage-stabilized 100VA outputs at either 110V or 230V a.c., via a tuned transformer. The size of the equipment is

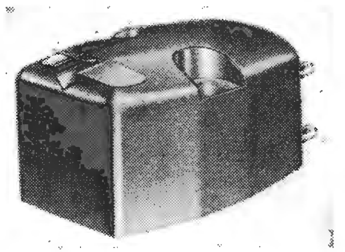


M.L. motor-drive Inverter.

$14 \times 10 \times 6$in and it costs £750 from M.L. Aviation Company Ltd., White Waltham Aerodrome, Maidenhead, Berks.

New Stereo Cartridge

AN effective stylus tip mass of only 0.5mgm and vertical and lateral compliances as high as 20×10^{-6} cm/dyne are claimed for the American Audio Dynamics ADC-1 moving-magnet stereo pickup cartridge (now being imported into this country by KEF Electronics). These parameters allow tracking weights as low as 0.75 gm to be employed with low-friction arms. The channel separation is better than 30dB up to 7kc/s and better

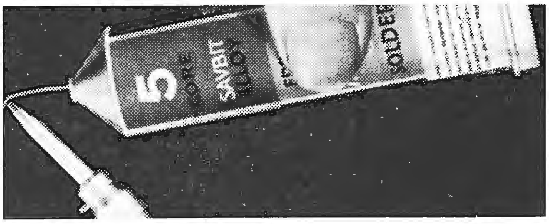


ADC-1 moving-magnet stereo pickup cartridge (imported by KEF Electronics).

than 18dB up to 20kc/s. The frequency response is within $\pm$ 2dB between 10 and 20,000c/s and the sensitivity within 2dB of $1\frac{1}{4}$mV cm/sec. A 0.6 mil radius diamond stylus is used. The price of this pickup cartridge is £25 10s from KEF Electronics Ltd., Tovil, Maidstone, Kent.

Solder Dispenser

SOLDER has normally been packed for the amateur in either rather unmanageable little coils or on reels. Many ingenious ways of using these packs have emerged, the problems being (a) holding the solder where it is required, and (b) the avoidance of tangling.

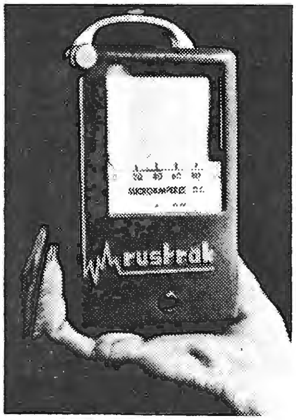


Multicore solder dispenser.

Multicore have avoided both these very neatly in a new dispenser which contains 16ft of 18 s.w.g. Savbit Alloy. The dispenser is an aluminium tube with a tapered nozzle, through which the solder can be drawn without tangling, and the base is sealed with a polythene cap. The dispenser is either held in the hand or stood on the bench. The new pack is made by Multicore Solders Ltd., Maylands Avenue, Hemel Hempstead, Herts.

Miniature Chart Recorder

ALTHOUGH rather smaller than the average multi-range test meter, the chart recorders introduced by Rustrak make no compromises and offer all the usual facilities provided by larger instruments. The writing process is completely dry on pressure-sensitive paper, the chart speed being from $\frac{1}{4}$ in to 60 in per hour, depending on the gear box used. Either a.c. or d.c. signals can be handled, maximum sensitivity being



Rustrak recorder. Accessory units use a similar, die-cast aluminium case.

50μA, and a range of amplifiers and range units are available to provide for measurement down to fractions of 1μA. Leaflets are obtainable from Rustrak Instruments, Lower Bevendean, Brighton 7, Sussex.

Current Controlled Schmitt Trigger

ASYMMETRICAL LOADING FOR HIGH OUTPUT CURRENTS

By N. C. BAUST*

A PROBLEM which sometimes arises in digital circuits is the designing of a threshold-sensitive switching unit. If the input signal is a voltage, one solution is to use a Schmitt trigger circuit. Generally this is a direct-coupled trigger circuit whose output potential takes up one or other of two levels, depending on whether the input level is above or below some chosen value. By choice the circuit may be designed so that the changes of state of the output occur for differing values of input. For instance the circuit may be designed so that to get the output to change to level A the input must pass through say +3V with a positive slope, whereas to change the output to the other level, B, the input must pass through say +1V with a negative slope. The difference between the two input threshold levels (2V in this case) is known as the " backlash ". The circuit would find use, for instance, in squaring a sine wave which has superimposed a comparatively low amplitude noise or ripple voltage.

Normally Schmitt trigger circuits are voltage controlled and are based on the long-tail pair configuration. Although the input backlash may be controlled by adjustment of the currents of the two active elements, generally the " tail " voltage is large compared with it, and consequently the active element currents are approximately equal. (The large tail voltage is desirable so that production tolerances on grid-base or base-base voltages have little effect on performance.) An occasion arose where a trigger circuit was necessary and the output anplitude from one side at least was to be well-defined and the load currents comparatively heavy. Unfortunately space restrictions were severe, so that component quantities had to be restricted, preferably not more than two transistors, and the signal source was a current rather than a voltage generator. Supplies available were ±27V and because of the temperature range to be covered silicon transistors were necessary. The signal source was a silicon solar cell, and under the restrictions the most suitable type was the Ferranti MS1AE. Since the cell was a novel element the initial design study was devoted to producing a conventional Schmitt trigger circuit controlled by it.

It was not clear from the data available at the time how much reverse voltage the cells would stand, nor what the response curve was for a reverse-biased unit. However, considering the simple equivalent circuit of Fig. 1(a) it was felt that it would not be unreasonable to extrapolate the characteristic curves in the negative voltage region as shown in Fig. 1(b). (A few precautionary investigations were made.) From Fig. 1(b) it can be seen that in the regions of reverse bias and low forward bias the units behave as constant-current sources for given values of illumination.

To keep the reverse bias low, the cell was placed across the base-emitter connections of the transistor. In this way each semiconductor would reverse-bias the other only by its own forward voltage drop: about 0.5V. A circuit was built with the form shown in Fig. 2. It was found that with certain intermediate light conditions, such as would occur when a shutter is gradually drawn between the cell and lamp, the circuit oscillated. It is possible to explain the behaviour in the following manner. Imagine

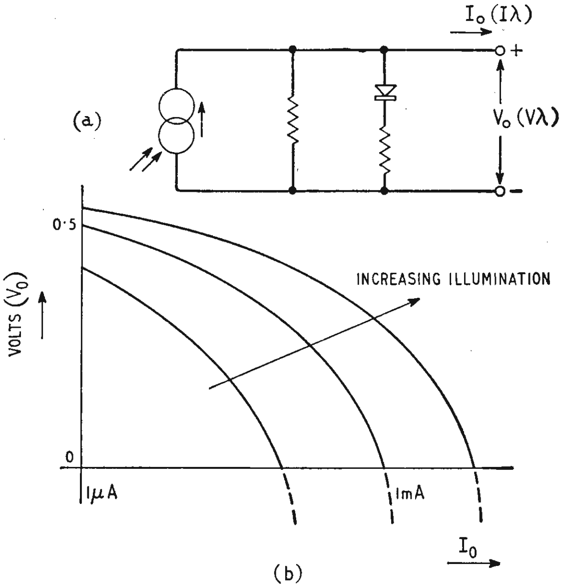


Fig. 1. (a) Equivalent circuit of solar cell. (b) Output characteristic of solar cell.

T1 conducting and zero illumination. There will be a voltage V_{e1} at the common-emitter connection. As the illumination is increased, base current for T1 reduces and the common-emitter potential rises. Eventually V_{ce} of T1 will equal the bias across R_3 and T2 will start to conduct. Current will be withdrawn from T1 and by regenerative action T2 will turn on. There will be a change in the common-emitter potential but, unless it is a drop in excess of the partial voltage across R_b due to the base current in T1 at the time of the change-over, the cell will be unable to supply enough current to maintain the new situation of T1 non-conducting, and the circuit will revert to its opposite state after a period determined by the cross-coupling time constants. It is

* English Electric Aviation, Ltd.

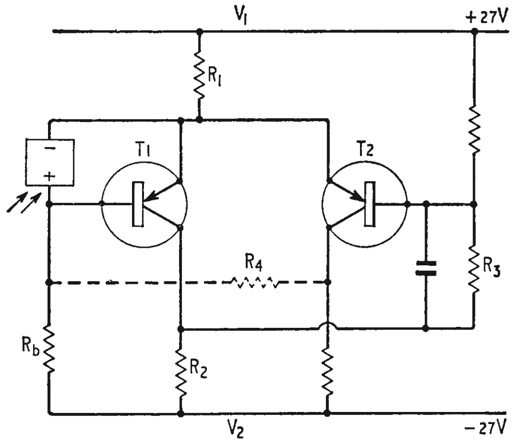


Fig. 2. Initial long-tailed pair circuit.

of course far more desirable that T1 be reverse-biased and not merely have its base current reduced to zero. This requirement is even more stringent. The situation can be explained a little more easily with the aid of Fig. 3.

Fig. 3 shows curves A and B for the cell, and C for the transistor $I_b - V_{be}$ characteristic. The current axis is positive to the right, with zero on the left for the cell, and zero on the right for the transistor. The same voltage scale is used for both. The current DE represents the current through R_b when the cell is not illuminated, and is therefore the maximum value of T1 base current. As the illumination increases, current in R_b is shared by the two units and the instantaneous values are obtainable from the position of a point moving along the line CGE. (Since the voltage across R_b increases for the initial rise in illumination the diagram is not strictly correct and the value DE should be stretched pro-rata as the cell output increases. However, the error is small if the initial voltage across R_b is large; it will be ignored here.)

Assume the illumination increases until the point reaches G, when T2 starts to conduct. T1 still has some forward bias and the illumination permits the cell output characteristic shown as AG. Consider now that at this level of illumination the circuit changes state. Unless the current through R_b is reduced to, say, DF, the cell cannot put a reverse bias across the transistor. For this reduction there must be a drop in the common-emitter potential, and obviously if the point is to move from G to H the maximum current through R_b must drop by 30 to 50%. Alternatively the load on the cell can be reduced by switching some of the current with resistor R_4. When T2 is bottomed little current flows through this resistor, yet a great deal flows through it when T1 is bottomed. R_b is still needed to provide an initial path for T1 base current as the illumination reduces. Hence with this arrangement the long-tailed pair circuit can be driven by the cell, but the collector currents of the two transistors are very similar, and are governed by the cell current. If the long tail is dispensed with and both emitters are returned to some voltage rail then a considerable difference between the transistor currents is possible, and an otherwise necessary extra amplifying stage may be dropped. As far as the job in mind was concerned a reduction of components, and therefore cost, followed. Note that the Zener diodes and biasing resistors of Fig. 4 are in lieu of supplies and the germanium diode in the figure is used to limit the output voltage to suit subsequent circuitry.

An analysis of the conditions for stable changeover is given in the appendix.

The circuit finally used is shown in Fig. 4 and the current locus diagram in Fig. 5.

For reasons which need not be discussed here it was convenient to use two 6.8V lines elsewhere (derived with Zener diodes) so the same two rails were used to define the output swing of T2. We have therefore the two emitters returned to the positive 6.8V line. Calculations showed that the maximum output current from the cell would be at least 120μA. As an initial assumption the levels at which the circuit was to trigger were chosen as 40 and 80μA, giving a cell current backlash of $80-40=40\mu$A.

Consider now the diagram of Fig. 5. DE represents the maximum value of base current for T1 (the maximum has to be restricted so that even with the maximum current gain for T1 the cell output will be sufficient to cause T1 to turn off). Starting at C with zero illumination, as the light intensity increases the cell's share of the current in R_b rises; the demarcation point travels along the transistor I_b—V_{be} curve CKGE until point G is reached. At this stage the remaining base current allows T2 to start to turn on. Assuming a high value of β for T1 at 35 and $R_2=68k\Omega$, the critical value of base current becomes approximately $\frac{1}{2}$ mA/35 $\simeq$ 15μA, therefore DE represents about 95μA as a first approximation. As the transistor turns off the locus moves rapidly along GH—the cell characteristic appropriate to the level of illumination. At H the circuit has changed state, resistor R_4 no longer passing any appreciable current, and the cell reverse biases T1. The cell current flows through R_b and any change in illumination changes the cell output voltage. Since the possible variation is so low and the voltage across R_b so large the locus is a straight line practically perpendicular to DE. Note that DF represents the current through R_b.

As the level of illumination falls the point moves down the " constant-current " line HF. At F base current starts to flow and the point moves along the T1 transistor characteristic once again, only at this time it is FJ. At J the base current reaches its critical value and T2 starts to turn off. The point moves rapidly along a line of constant illumination, JK, as the circuit changes state once more, and the base current increases rapidly by EF. Further

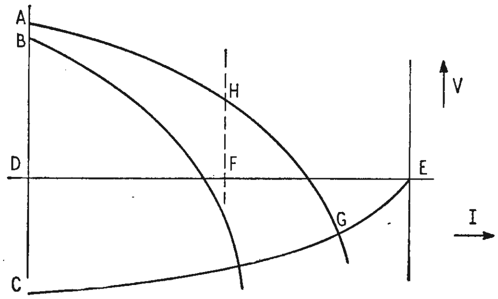


Fig. 3. Combined solar cell output and transistor input characteristic.

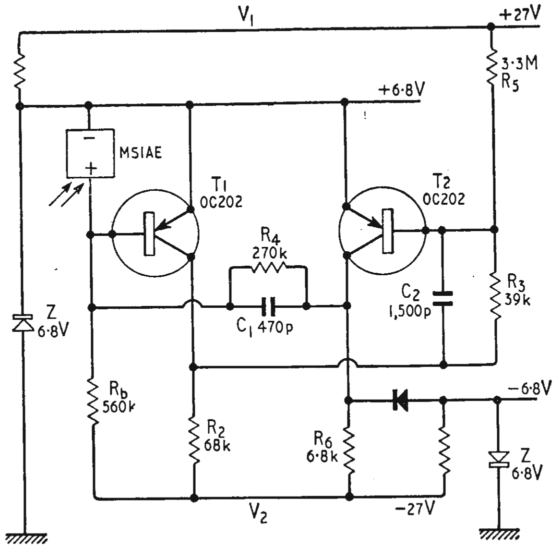


Fig. 4. Final trigger circuit.

movement of the point towards C or G will depend on the action of the illumination. We see then that the current CK should be about 40μA, and that DF should be greater than CK by the critical value of base current, i.e. a total of about 55μA. (Note for β as low as 20 the critical base current is 25μA and the cell illumination would therefore have to fall to 30μA to allow T1 to turn on.) Therefore $R_b \simeq 34V/55\mu A = 620k\Omega$. Allowing for tolerances R_b is therefore 560kΩ.

Once T1 turns on the maximum possible base current is represented by DE. Since this equals about 95μA, R_4 must pass approximately 35μA and $R_4 \simeq 14V/35\mu A \simeq 400k\Omega$. Allowing for tolerances 330kΩ would have been a suitable value. In practice, the value used was 270kΩ (which lets DE extend to 129μA under extremes of tolerance, but nevertheless permits operation). This low value of R_4 extends the backlash region which is important for the circuits where tolerance combinations would reduce the value permitted by the 330kΩ to a low level. (The backlash tends to counteract dither on the shutter mechanism.) To keep the backlash high with the particular value of R_b the collector current for T1 must therefore be in the $\frac{1}{2}$ mA region (see appendix), hence $R_2 = 34V/\frac{1}{2}\text{ mA} = 68k\Omega$.

For this particular circuit it was mandatory that the output impedance be as low as possible, consistent with a minimum of components. Base leakage current for T2 will not exceed 2μA at the highest ambient temperature at which the circuit has to operate, hence $R_5 < 20V/2\mu A = 10M\Omega$, but to produce a reasonable cut-off condition for T2, R_5 should be low. A compromise value of 3.3MΩ was chosen. Further, the maximum V_{ec} of T1, chosen as an OC202 for high gain, is 15V, thus the ratio $R_3 : R_3 + R_2 = 3 : 7$. Allowing for tolerances the nominal ratio must be less than this, and with R_3 as 39kΩ the ratio is 4 : 11. Having determined the values of base current source a quick calculation shows that for a β of 20 the maximum guaranteed collector current will be 5.3 mA. Therefore $R_6 > 32.5V/5.3$ mA. Allowing for tolerances $R_6 =$ 6.8kΩ.

The major criticisms of the circuit as shown are:

(a) T2 does not become truly reversed biased when T1 conducts. This could be corrected by altering values but only at the expense of reduced collector current from T2.

(b) The backlash is not particularly well defined; it is dependent on the current gain of T1. Further, the backlash which exists is based on the electrical output from the cell and not the input illumination; this is because of scanty knowledge of the lamp characteristics.

(c) The guaranteed steady-state current into an external load on this circuit is negligible. However, as used, it has no steady-state load. The transient load permitted is determined by the transient base current of T2. This is of adequate proportions, and is a function of the cross-coupling capacitor C_2. If a steady-state load has to be supplied either R_6 must be changed in value, or a buffer amplifier must be included.

The cross-coupling capacitors naturally do not affect the steady-state performance; their main function is to assist the regeneration action and incidentally produce fast rising edges. 470pF is quite sufficient and, being a value common to other associated circuitry, is the value fitted to the base of T1. For T2 the cross-coupling capacitor is not used only to "speed the rising edge" but also to ensure that T2 can handle its "external" transient load. This load has a time constant of approximately 150μS, a peak value of about $\frac{1}{2}$mA, and persists for less than 100μS. At the end of 100μS the

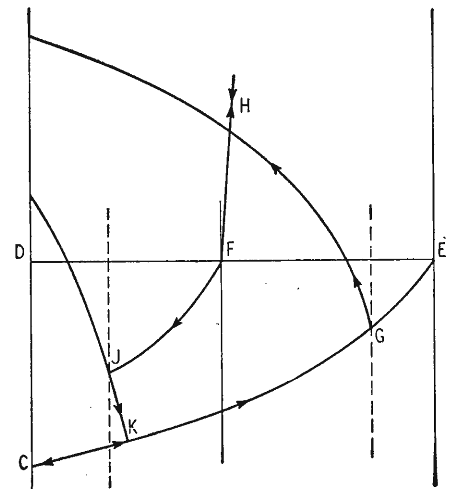


Fig. 5. Current locus diagram for circuit of Fig. 4.

external load is reduced to about 260μA, and with 1500pF at the end of the same period, the transient base current can still demand about 700μA collector current. On this analysis 470pF would just about suffice but the external load is in fact governed by two time constants, and a larger value capacitance is necessary. 1500pF is a suitable value which is also found in other associated circuitry and therefore is the value chosen. (Note that from the viewpoint of store simplification it would be better if both

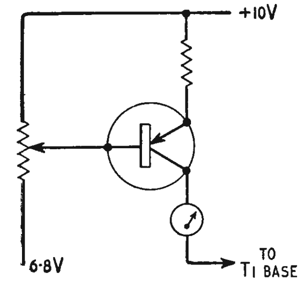


Fig. 6. Backlash measurement circuit.

capacitors were the same value, but the requirement of a fast negative-going edge on T2 output results in C_1 having to be not more than about 500pF nominal.)

The circuit has been tested for performance over the range −40°C to 100°C although, since the MSIAE cell is a "mounted" type, the upper level should have been restricted to 85°C. It has also been checked with full variation of the prime supplies and with the lamp voltage reduced by half. The lamp used is a 28V 40mA Atlas B.S.C. type run from a 20V line and is mounted 1cm away from the cell; light passes through a 0.05-in wide slit to reach the active surface. There is no mirror or lens along the light path, nor is there any provision for alignment of lamp and cell other than that due to the mechanical construction.

An attempt was made to check the degree of backlash by replacing the cell with a transistor as in Fig. 6 and removing the cross-coupling capacitors. Results for three completely different assemblies are given in the table below, the only common

Circuit	1	2	3	Mean	Calculated Value (27V H.T.)
I_{λ_1} μA	97	84	89	90	91
I_{λ_3} μA	57	48	52	52.3	54
T_2 output rise time (full amplitude) μS	12	6	12	10	—
T_2 output fall time (full amplitude) μS	1.3	1.3	<2	—	1.6

items being the ±27V supplies, and these were 26.3V as read on a Model 8 Avo, 100V range.

Conclusions

A trigger circuit which may be controlled by the output of a silicon solar cell has been described. Depending on the maximum light intensity available at the cell, a considerable degree of backlash may be built into the circuit, although the boundaries of the backlash range are dependent on the transistor current gain. This is principally because the circuit configuration permits the two transistors to have greatly different collector loads. The conventional long-tail pair circuit would permit better definition of backlash but also produce more nearly equal collector loads. Tests on three samples show reasonable agreement between theory and practice.

Acknowledgements

The author is grateful to English Electric Aviation Limited for permission to publish this material.

APPENDIX

(a) Change-over Stability

Consider Fig. 2 with T1 conducting, but that the circuit is just about to change state. The cell current, I_{λ_1} is given by

$$I_{\lambda1}=\frac{V_{e1}-V_{eb1}-V_2}{R_b}+\frac{V_{e1}-V_{eb1}-V_{c2}}{R_4}-\frac{I_{c1}}{\beta_1} \quad .. \quad (1)$$

where V_{e1} = emitter voltage, T1 ON
V_{c2} = T2 collector voltage
V_{eb1} = T1 emitter base voltage
I_{c1} = collector current } of T1
β_1 = current gain

Now consider that T2 has just turned on. The cell current becomes

$$I_{\lambda2} = \frac{V_{e2}+V_\lambda-V_2}{R_b} \quad .. \quad .. \quad .. \quad (2)$$

where V_{e2} = emitter voltage, T2 ON
V_λ = cell output voltage.

Obviously for the change (either direction) to be stable

$$I_{\lambda1} > I_{\lambda2} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3)$$

Now if the change of voltage across $R_b = \triangle V$ we have

$$\triangle V = (V_{e1}-V_{eb1})-(V_{e2}+V_\lambda) \quad .. \quad .. \quad (4)$$

Note $\triangle V \simeq V_{e1}-V_{e2}$ .. .. (5)

and is determined by the conduction states of T1 and T2. Combining equations 1, 2, 3 and 4 we get

$$\triangle V > R_b\left[\frac{I_{c1}}{\beta_1}-\frac{V_{e1}-V_{eb1}-V_{c2})}{R_4}\right] \quad .. \quad (6)$$

In the circuit used we find that nominally

$$V \simeq 6.2-7.2 = -1.0V$$

and R.H.S. of equation 6 $\simeq 560\left[\frac{\frac{1}{2}}{20}-\frac{6.2+7.5}{270}\right]$
$= -1.46V$

∴ equation 6 is satisfied.

(b) Change-over Threshold T1 → OFF, T2 → ON

The threshold level for T1 turn-off I_{λ_1} is

$$I_{\lambda1}=\frac{6.8-0.6+27}{560}+\frac{6.8-0.6-(-6.8-0.6)}{470} - \left[\frac{6.8-0.5+27}{20\times68}-\frac{27-(6.8-0.6)}{3339}\right]$$

$= 91\mu A$.

The term in square brackets allows $\frac{1}{2}$V for V_{ec} when T1 is just bottomed, and assumes $\beta = 20$.

(c) Change-over Threshold T2 → OFF, T1 → ON

With T2 conducting its base current will have some value and with T1 cut off the base current is I_{b2}. To just bottom T2 the necessary base current is

$$I_{b2}=I_{c2}/\beta_2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (7)$$

Hence as T1 varies from cut off to the threshold state the base current of T2 changes by the surplus

$$\delta I_{b2} = I_{b2}-I_{b2} \quad .. \quad .. \quad .. \quad .. \quad (8)$$

This change in T2 base current is due to T1 collector current; by current distribution in R_2 and R_3 the necessary T1 collector current is

$$\delta I_{c1} = \delta I_{b2} \cdot \frac{R_2+R_3}{R_2} \quad .. \quad .. \quad .. \quad .. \quad (9)$$

and consequently

$$\delta I_{b1} = \delta I_{c1}/\beta_1 \quad .. \quad .. \quad .. \quad .. \quad (10)$$

The cell current at this threshold is therefore

$$I_{\lambda3} \lessapprox \frac{V_{e1}-V_{eb1}-V_2}{R_b}-\frac{\delta I_{c1}}{\beta_1} \quad .. \quad .. \quad (11)$$

with terms as in (a) except for interchange of transistors. For the circuit considered we have

$$I_{\lambda3} = \frac{6.8-0.6+27}{560} - \frac{1}{20}\cdot\frac{68+39}{68}\left\{\left[\frac{6.8-0.6+27}{68+39}-\frac{27-(6.8-0.6)}{3300}\right] - \left[\frac{1}{20}\cdot\frac{6.8-.5+27}{6.8}\right]\right\}$$

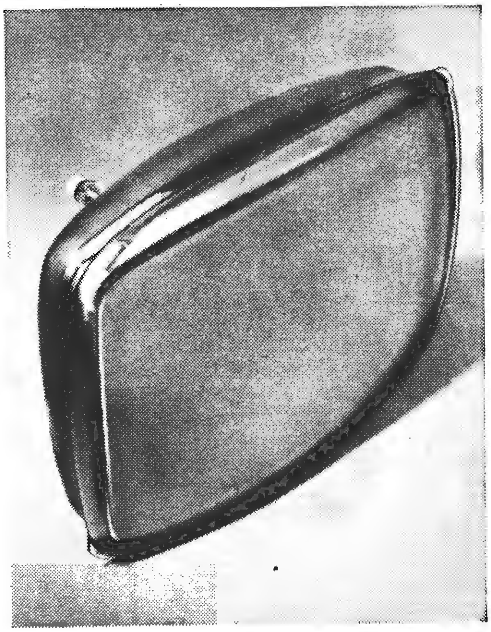
$$= 59.2 - \frac{107}{20.68}[304-245]$$

∴ $I_{\lambda3} = 54.5\mu A$.

As in (b) the square bracket term allows $\frac{1}{2}$V for V_e when T2 is just bottomed.

MAZDA

ALL GLASS twin panel tubes



110° 19" and 23" twin panel

SHORT TELEVISION TUBES

CME1906/A47-13W

CME2306/A59-13W

These new tubes differ from previous 19in. and 23in. tubes of the Mazda range in that each has a moulded safety panel of tinted glass bonded directly to the front face of the tube. In neck length and electrical characteristics these tubes are identical to CME1903 and CME2303 respectively, having 110° deflection angle and using magnetic deflection and electrostatic focus.

General details

Twin Panel	Tinted Grey Glass
Rectangular Face	110° Deflection Angle
Aluminised Screen	Silver Activated Phosphor
Electrostatic Focus	Magnetic Deflection
Short Neck	Straight Gun—non ion trap

External Conductive Coating
Heater for use in Series Chain
Heater Current I_h 0.3 A
Heater Voltage V_h 6.3 V

Design Centre Ratings

		CME1906	CME2306	
Maximum Second and Fourth Anode Voltage	$V_{a2,a4(max)}$	17	17	kV
Minimum Second and Fourth Anode Voltage	$V_{a2,a4(min)}$	13	13	kV
Maximum Third Anode Voltage	$V_{a3(max)}$	+1 to −0.5	+1 to −0.5	kV
Maximum First Anode Voltage	$V_{a1(max)}$	550	550	V
Maximum Heater to Cathode Voltage-Heater Negative (d.c.)	$V_{h-k(max)}$	200	200	V

Inter-Electrode Capacitances

Cathode to All*	C_{k-all}	3.5	3.5	pF
Grid to All*	C_{g-all}	8.5	8.5	pF
Final Anode to External Conductive Coating (approx.)	$C_{a2,\ a4-M}$	1250	2000	pF

*Including AEI B8H Holder VH68/81 (8 pin)

Typical Operation

Grid Modulation (Voltages referred to cathode)

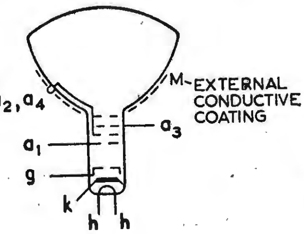
Second and Fourth Anode Voltage	$V_{a2,\ a4}$	16	16 to 17	kV
First Anode Voltage	V_{a1}	400	400	V
Beam Current		350	350	μA
Third Anode Voltage for Focus (Mean)	$V_{a3(av)}$	200	200	V
Average Peak to Peak Modulating Voltage		35.5	35.5	V
Grid Bias for Cut-off of Raster	V_g	−40 to −77	−40 to −77	V

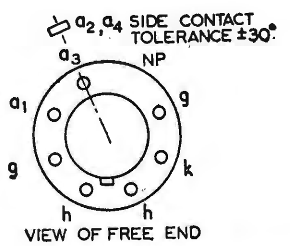
Maximum Dimensions

Overall Length	317	374	mm
Face Diagonal	491†	614‡	mm
Face Width	441	544	mm
Face Height	361	443	mm
Neck Diameter	29.4	29.4	mm

†The maximum dimension over the complete panel is 507 mm
‡The maximum dimension over the complete panel is 631 mm

Tube Weight
Nett (Approx.) 22.5 37.5 lbs.
Side Contact: CT8 (Cavity) **Base:** B8H





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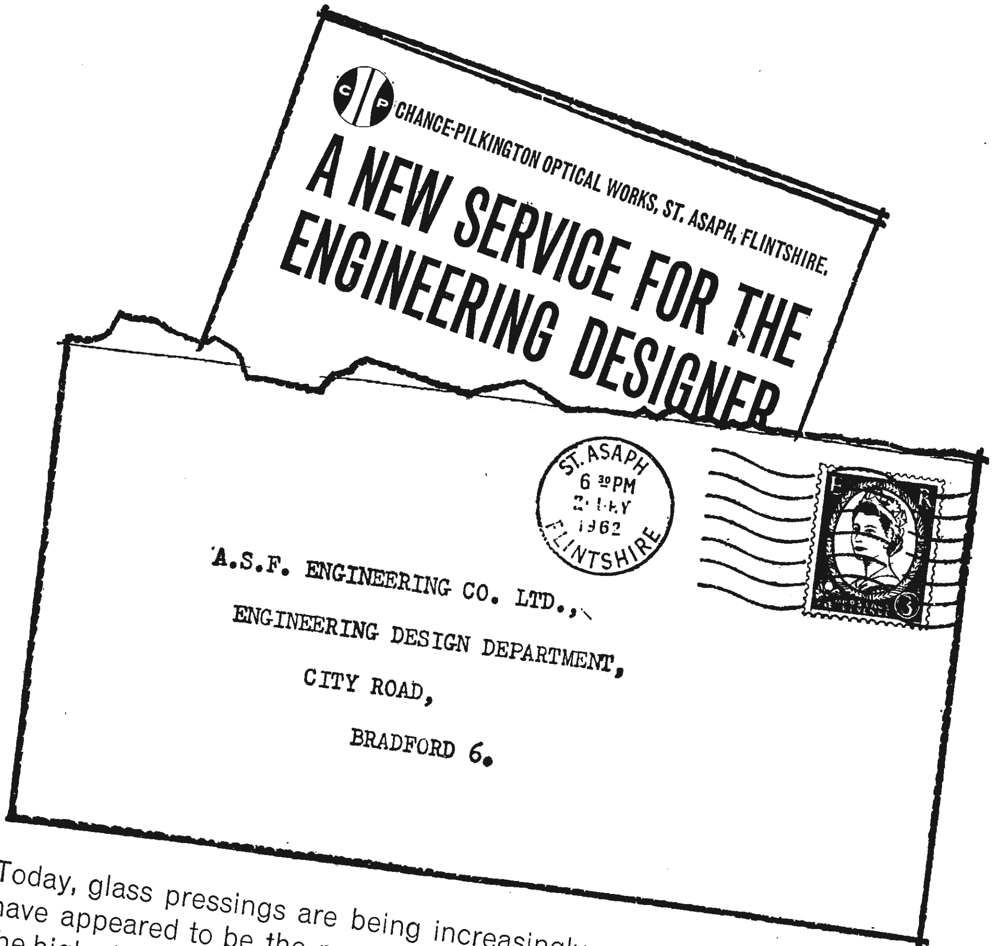
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Bootstrap D.C. Amplifier

PHASE-INVERSION IN TWO-STAGE AMPLIFIER

By J. F. YOUNG, C.G.I.A., A.M.I.E.E., A.M.Brit. I.R.E.

ARTICLES by A. R. Bailey[1] and by G. W. Short[2] have recently revived interest in bootstrap audio amplifiers. Ocasionally the use of such arrangements for d.c. amplification has been mentioned briefly[3], but this application of the bootstrap principle does not seem to have received as much attention as have audio applications. Consequently, it is of interest at the present time to examine the bootstrap amplifier in the d.c. form in which it has been found useful for industrial applications. In industrial electronics it is very important to keep reliability to a maximum and cost to a minimum, both of these objects best being achieved by simplicity.

The simple circuit which Short has christened the Bootstrap Follower appeared to have possibilities for d.c. amplification (i.e. for amplification in which coupling capacitors are avoided so that the gain does not fall off at low frequencies), but when we could obtain only single or double triodes and single pentodes there was little to be gained by using the circuit. However, when valves appeared on the market having a triode and a pentode with separate cathodes, both in the same miniature envelope, it was a different story.

As an example of such valves, the ECF80 triode-pentode has a high-gain pentode together with a medium-gain triode. It is very suitable for use as an amplifier plus cathode follower to give a low impedance output. Even used merely in this way such a valve made possible the simplification of industrial control as well as of audio and video equipment. However, in some cases an even higher gain than can be obtained reliably from the pentode alone is required. It is possible simply to use the two sections of a triode-pentode as a two-stage amplifier to obtain increased gain, the output then being in-phase with the input. In many applications this non-phase-inverting property is either advantageous or just unimportant. In other cases, such as negative feedback amplifiers or servo systems, it is sometimes necessary to add an extra amplifier stage merely to restore the phase inversion removed by the second stage. This can lead to additional problems, since every added stage introduces more high frequency phase shift and makes more difficult the prevention of undesired high frequency oscillation in the feedback system, in addition to increasing the size and complexity of the amplifier. If a normal grounded cathode amplifier stage as shown in Fig. 1 is considered, we really need to be able to add a non-phase-inverting stage in order to avoid the difficulty. It is, of course, possible to add the commonest form of non-phase-inverting stage, a cathode follower, but this would give no further increase of voltage gain.

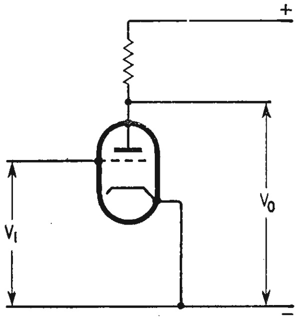


Fig. 1. Normal grounded-cathode amplifier.

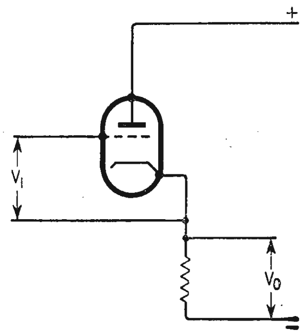
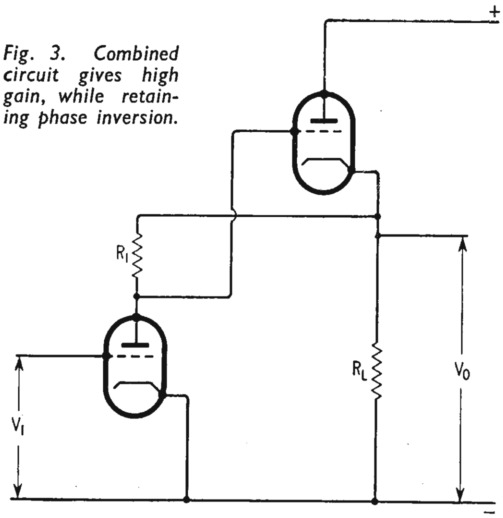


Fig. 2. Assuming zero-resistance power supply, this is identical to circuit of Fig. 1.

It is possible, however, to obtain a voltage gain without phase inversion from a stage which, like the cathode follower, has a grounded anode. In order to do so, the input must be arranged as shown in Fig. 2. Comparison of Fig. 2 with Fig. 1 shows that the only difference between the two is the order of connection of the power supply and the load resistor between the anode and cathode of the valve. This loaded-cathode amplifier circuit was often used in the early days of radio, when it was not unusual to connect earphones in the lead between the negative of the h.t. battery and the filament of the transformer-coupled a.f. amplifier. With the metal-cased earphones then in common use, no doubt the safety of the operator had something to do with the adoption of this arrangement! As time passed, it became customary always to put the load in the anode circuit, and loads in the cathode circuit disappeared until the bootstrap sweep generator came into wide use. However, the connection of Fig. 2 still has its uses in audio or d.c. amplifiers, whenever an extra stage of voltage amplification is required without phase inversion.

We are concerned here with d.c. amplification, and the basic circuit for direct coupling a grounded-anode voltage amplifier to a preceding stage is

Fig. 3. Combined circuit gives high gain, while retaining phase inversion.



shown in Fig. 3. Analysis of this circuit shows that, for a high value of voltage gain, it is necessary to have the largest possible values of R_1, of the amplification factor of the first stage and of the mutual conductance of the second stage. The first stage should therefore preferably be a pentode. In order to make it possible to have a large value of R_1 without running the first stage at too small an anode current, the positive end of R_1 can be directly coupled to the cathode of the second stage as shown in Fig. 4. This increases the voltage across R_1 and therefore increases the permissible value of R_1 for a given current I_1. The desirability of using a high value of R_1 can be seen, for example, from the equation (1) below. In an audio amplifier a capacitor can be used to provide the constant-voltage, while a gas-filled stabilizer tube or a Zener diode can be used in a d.c. amplifier. The similarity of this arrangement to a bootstrap sweep generator will be noted.

If a pentode is used in the second, grounded anode, stage, it is necessary to supply its screen grid at a constant voltage relative to its cathode. Either the constant voltage supplying R_1 in Fig. 4 or a separate floating supply can be used for this purpose. With the advent of the triode-pentode valve the use of a triode second stage became attractive since then the whole amplifier requires only one valve envelope.

Assuming that the second stage does not take grid current, an input current I_1 flows through resistor R_1 in Fig. 4. The output voltage is obtained across resistor R between the valve cathode and the negative supply line. The battery or some other device supplies a floating constant voltage. Analysis of this circuit shows that:—

$$V_o = \frac{I_1 R_a R}{R_a + R}\left(1 + \frac{\mu R_1}{R_a}\right) \quad \ldots \quad \ldots \quad (1)$$

where μ is the amplification factor and R_a is the anode resistance of the valve. If now R is very much greater than R_a, and if μ and R_1 are large, then the output voltage approaches $I_1 \mu R_1$. The current gain from the input current to the current in resistor R approaches $\mu R_1/R$, which can have a very high value. The input impedance of the circuit is:—

$$Z_1 = \frac{(\mu + 1)R_1 R + (R_1 + R)R_a}{R_a + R} \quad \ldots \quad (2)$$

while the output impedance with a constant current input is simply:—

$$Z_o = \frac{R_a R}{R_a + R} \quad \ldots \quad \ldots \quad (3)$$

If, however, the input is from a constant voltage source, the output impedance becomes:—

$$Z_o = \frac{R_a R}{(\mu + 1)R + R_a} \approx \frac{1}{g_{m2}} \quad \ldots \quad (4)$$

where g_{m2} is the valve mutual conductance. In fact, under the latter circumstances, the valve simply operates as a cathode follower. This fact can be of importance in some applications, since limitation of the input voltage, for example by a rectifier, will firmly restrain the output voltage. A typical application where this characteristic is useful is in the current limiting circuits of electronic motor controls. Here, if the current taken by the motor being controlled approaches an unsafe value, it is necessary to prevent further increase of the voltage applied to the motor by limiting the output of the control amplifier.

If the constant current input source to the grounded-anode triode amplifier is from a pentode, giving an anode current of $g_{m1}V_g$, then it can be seen from equation (1) that the overall voltage gain approaches $\mu g_{m1} R_1$. Because all three factors in this expression can have high values, a good overall gain with only a single phase inversion can be obtained. As soon as triode-pentode valves were available, it became possible to obtain this high gain with only a single miniature valve envelope.

Similar arrangements have been used by Jeffery[4] in a high gain a.f. phase splitter, and by Siskind[5] in an integrator. Both used a capacitor to replace the constant voltage coupling from the second cathode to the positive end of the first anode load. While both writers used two separate pentodes, Jeffery connected the second as a triode while Siskind had the screen of the second pentode capacitively coupled to its cathode. Kaufer[3] has mentioned the pentode-triode type of circuit and also the use of a neon tube to replace the capacitor in this arrangement. He considered the triode as providing a very high value of dynamic anode load for the pentode.

A practical d.c. amplifier circuit using an ECF80

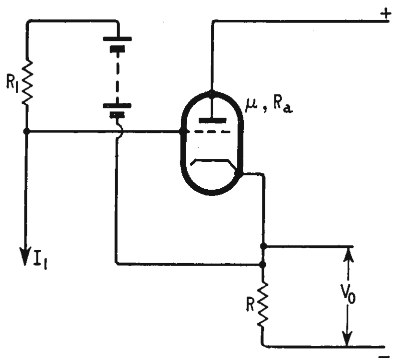


Fig. 4. I_1 is first-stage anode current. Battery blocks cathode from previous anode voltage.

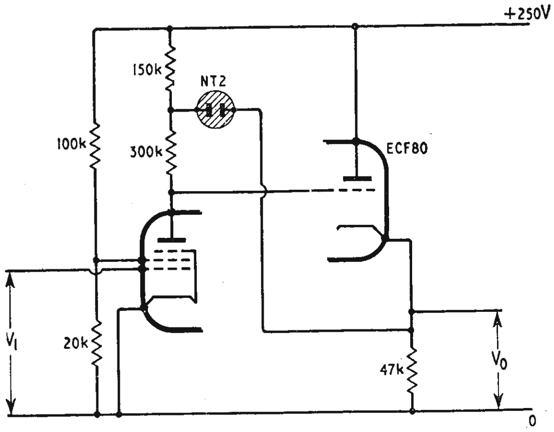


Fig. 5. Practical circuit. Constant voltage is offset by neon.

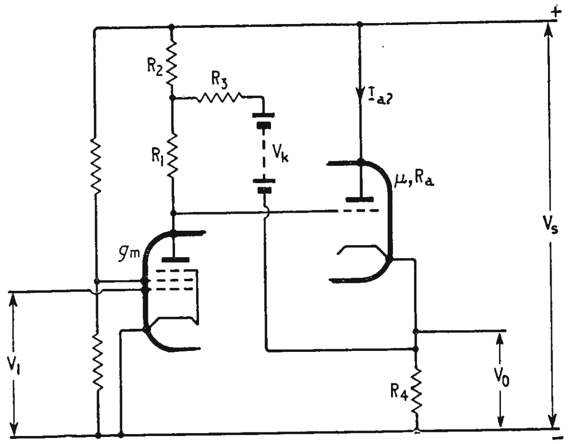


Fig. 6. Symbols used for gain calculation.

triode-pentode is shown in Fig. 5. Since the use of a high tension battery as shown in Fig. 4 is inconvenient, a cold cathode tube is used. The screen grid of the input pentode is operated from a potential divider in order to obtain a high gain, and the pentode anode current is low. Under these conditions, gains of about 400 are obtainable from the pentode alone. However, the total effective load in the cathode of the triode is formed from several resistors in parallel, and it is difficult to keep the value of the effective load high enough to obtain a good gain from the triode. The main reason for the difficulty is the operating current required by the cold cathode tube to prevent extinction at maximum positive output. Consequently, the overall gains obtained are much less than the maximum possible.

Analysis of this circuit, using the symbols shown in Fig. 6, gives for the gain:—

$$\frac{V_o}{V_1}=\frac{g_m R_4(\mu R_1 R_2+\mu R_2 R_3+\mu R_1 R_3+R_a R_2)}{R_a(R_2+R_3+R_4)+R_4[R_2+(1+\mu)R_3]} \quad (5)$$

If R_3 is small, the gain simplifies to:—

$$\frac{V_o}{V_1}=\frac{g_m R_2 R_4(R_a+\mu R_1)}{R_a(R_2+R_4)+R_2 R_4} \quad \ldots \quad \ldots \quad (6)$$

The anode current of the triode is given by:—

$$I_{a2}=\frac{g_m V_1[R_2 R_4-\mu(R_1 R_2+R_1 R_3+R_1 R_4+R_2 R_3)]}{R_a(R_2+R_3+R_4)+R_4[R_2+(1+\mu)R_3]} \quad (7)$$

The contribution of voltage V_k to the output voltage is:—

$$V_{ok}=\frac{V_k R_4(\mu R_2-R_a)}{R_a(R_2+R_3+R_4)+R_4[R_2+(1+\mu)R_3]} \quad (8)$$

It is interesting to note from the last equation that if only R_2 could be made equal to the inverse of the mutual conductance of the triode, variations of V_k would have no effect on the output voltage and the stability of the cold cathode coupling tube would be unimportant. Unfortunately in practice it is not possible to approximate to this result because R_2 must be large if a high gain is to be obtained.

The supply voltage term in the full expression for the output voltage is:—

$$V_{os}=\frac{V_s R_4[R_a+R_2+(\mu+1)R_3]}{R_a(R_2+R_3+R_4)+R_4[R_2+(\mu+1)R_3]} \quad (9)$$

From this it can be seen that R_a, R_2 and R_3 should be large compared with R_4 in order to minimize the effect of supply voltage variations on the output voltage. Thus the best results from all points of view will be obtained by making R_2 much greater than R_4; R_4 much greater than R_a and R_1 and μ must be as large as possible. Both the supply voltage and the cold cathode tube voltage should therefore be high.

Since the output voltage is in antiphase to the input voltage, the application of overall negative feedback to the circuit is quite easy. One way of doing this is shown in Fig. 7. Here, resistors R_x and R_y provide the negative feedback. The overall gain of such a circuit is:—

$$\frac{V_o}{V_1}=-\frac{AR_y}{(A+1)R_x+R_y} \quad \ldots \quad (10)$$

where A is the gain of the amplifier without feedback. The rate of change of overall gain with changes of A is:—

$$\frac{R_y(R_x+R_y)}{[R_x(A+1)+R_y]^2}$$

which is small if A is large. Thus, as is well known, the application of negative feedback in this way can give an amplifier with very stable gain characteristics. In addition, a good linearity is obtained because of the reduction of gain variations. The output impedance of such an arrangement is low while the

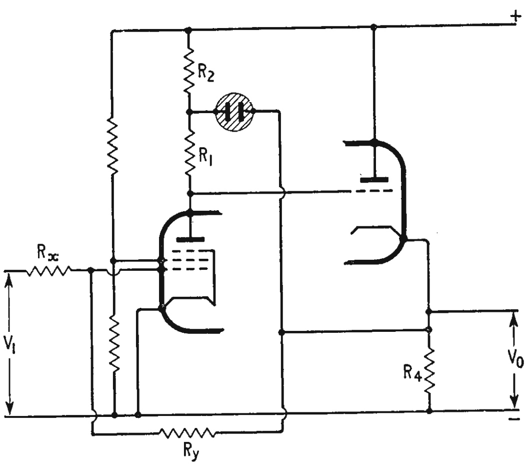


Fig. 7. Application of negative feedback to practical circuit of Fig. 5.

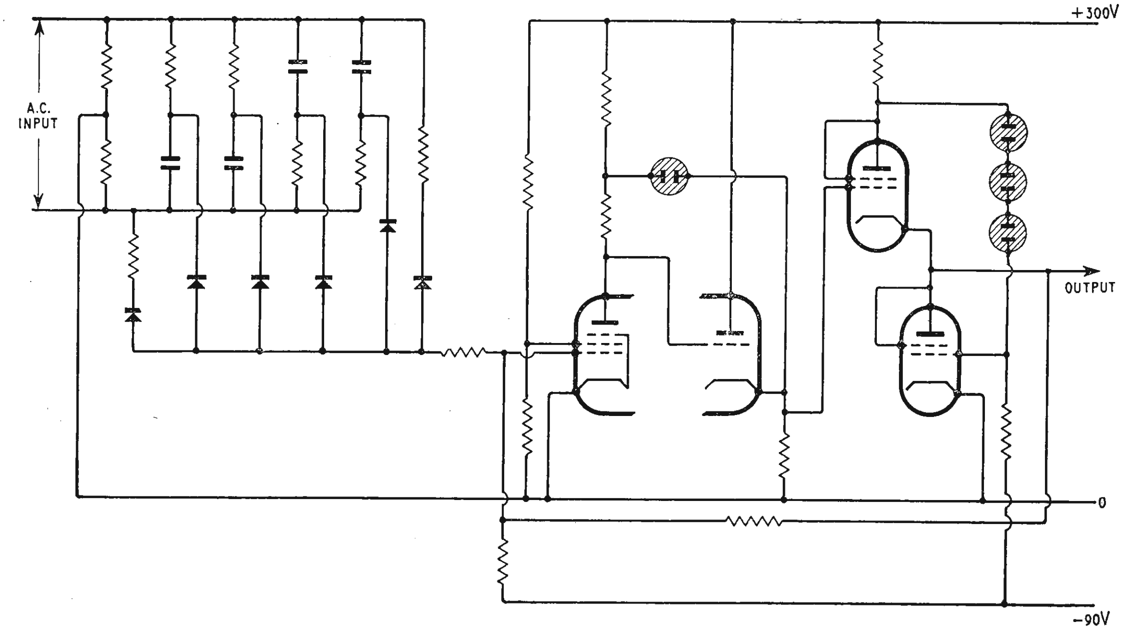


Fig. 8. Circuit to provide low impedance output for both increasing and decreasing input.

input impedance can be reasonably high. Resistor R_y should be large compared with R_4 in order to minimize the effect of the feedback resistors on the internal gain of the amplifier. It is usually desirable to keep the value of R_y high from another point of view, since if R_y is large so will be R_x and the input impedance will then be high.

The triode-pentode amplifier of Fig. 5 has given a gain of more than 1700 with a linear output voltage swing of over 100 volts. The mean output voltage level is about 120 volts, so one end of the heater is preferably connected to a point mid-way between the supply terminals in order not to exceed the heater to cathode voltage rating of the valve. Excellent linearity with a wider maximum output voltage swing can be obtained when negative feedback is added. For example, when the gain is reduced to about 100 by negative feedback, the linear swing is from over 200 volts to less than 50 volts of output.

The bootstrap d.c. amplifier is chiefly useful in that it gives a fairly high gain with one valve envelope and few components while retaining the phase inverting property of one stage and not requiring excessive supply voltages. The gain obtainable is limited mainly by the cold cathode tube coupling which is employed.

A typical industrial application of the bootstrap d.c. amplifier is shown in Fig. 8. Here, it was required to produce a d.c. output at very low impedance, the output being proportional to the amplitude of a 50 cycles per second alternating voltage. It was necessary to have a rapid response to variations of the alternating voltage, and therefore ripple due to rectification of the voltage had to be reduced as much as possible without using smoothing which would slow down the response. Consequently, the single phase alternating voltage is converted to a six phase supply by the use of phase shift circuits, and a six phase star connected rectifier (giving only a small 300 cycles per second ripple at its output) is used.

The feedback amplifier comprises a bootstrap d.c. amplifier to give a high internal gain, together with a double cathode follower to give a low output impedance regardless of whether the input voltage is increasing or decreasing. Heavy overall feedback is used. It will be seen that the bootstrap arrangement is ideally suited to this type of application.

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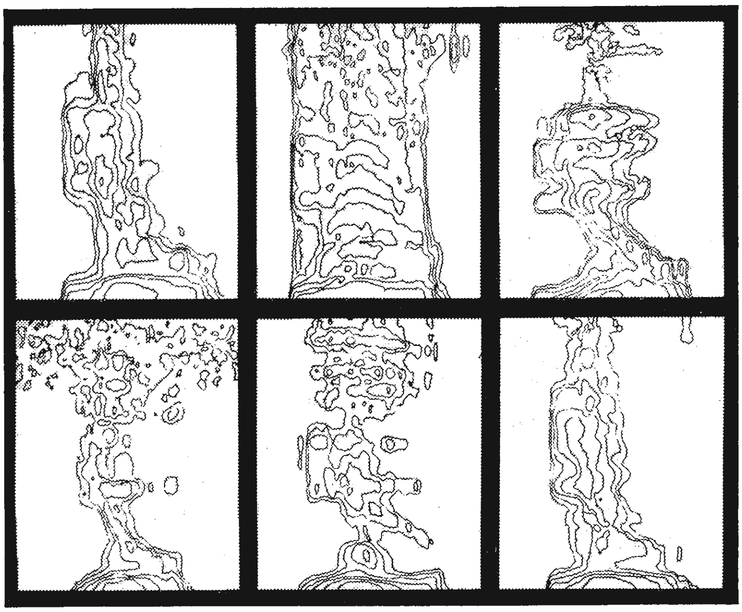
TECHNICAL NOTEBOOK

Radio-telescope image integration by analogue methods is suggested in a paper by D. J. McLean and J. P. Wild in the December 1961 issue of the *Australian Journal of Physics* (p. 489). To obtain sufficiently high resolutions in radio astronomy, interferometric methods involving large arrays of aerials have frequently to be adopted. Unfortunately, to observe an area of any size, such aerials then have to be electronically scanned over the area point by point. Such sequential scanning could be avoided by simultaneously forming all parts of the image by analogue means. One such method would be to set up a scaled-down optical array of pupils in an opaque screen geometrically similar to the original aerial array. The amplitude and phase of the radio signals from each aerial are used to modulate the transmission and aperture of the corresponding pupils. A parallel beam of light incident on the pupils can then be focussed by a lens placed behind them to form the required total image. Suitable "pupils" could possibly be made by splitting the beam into a number of parts which are reflected off vibrating mirrors before being recombined: the mirrors would be driven via piezo-electric crystals from the corresponding radio signals. The corrections necessary to avoid unwanted diffraction and side-lobe effects produced by the original aerial array can also be carried out optically. Unfortunately such a system would have two serious disadvantages. First, most of the light falls on the opaque screen between the pupils, while that which does pass through them becomes spread over a relatively wide area. The resulting image is thus very weak and "noise" in the light is likely to become serious. The second disadvantage is that the actual construction of the pupils is likely to involve considerable technical difficulties because of the short wavelength of light. The first disadvantage could be overcome if the light source and pupil array were replaced by an array of modulated sources. Nearly all the radiated energy would then be made use of in image formation so that the efficiency would be greatly improved and the noise correspondingly reduced. (Unfortunately, the necessary aerial corrections cannot then be carried out in the analogue system.) The second disadvantage could be avoided by using a suitably longer wavelength. Both these modifications could conceivably be incorporated in systems using either microwaves or ultrasonics, but unfortunately suitable image detectors are not as yet available in either case.

"Voiceprints" like fingerprints may be unique and so form a method of identification according to L. G. Kersta of the Bell Telephone Laboratories. Voiceprints are actually sound spectrograms (see figure) of a spoken word in which the time increases from left to right and the frequency from bottom to top. The sound intensity may be shown either as the density of the spectrogram trace or by plotting the spectrogram in contour form (see figure). Oddly enough, such "visible speech" seems to disclose fundamental patterns that cannot be distinguished by ear. In tests voiceprints were made of the same word spoken by several different persons, each person uttering the word several times. Trained subjects were then able correctly to group the "prints" corresponding to each voice in 97% of the cases. The illustration shows six contour voiceprints of the word "you" spoken by five different people, and it can readily be seen that the top left and bottom right prints correspond to the same person.



Neon Indicator for small-signal working has been developed by the Fuji Communication Apparatus Manufacturing Co. of Kawasaki. Normally, a neon tube requires several tens of volts to switch it from the "off" to the "on" states, because of the large series resistance required for current limiting. The TG121A, however, is a two-cathode tube, in which the anode current is split between the two, and any variation in the ratio of cathode currents has very little effect on the total. If a cathode bias resistor is inserted in one cathode (say K1), the current through this will be the smaller, and the associated cathode glow will be on K2. A negative going voltage applied to K1 (or a positive one to K2) will cause the glow to shift to K1, which is viewed through a window. The tube has a high impedance characteristic, and the signal applied between cathodes need only be about 5V for a full shift of the glow. The anode voltage is 110V.

Artificial hand which is automatically controlled without mechanical intervention was described by R. Tomović and G. Boni in the *I.R.E. Transactions on Automatic Control* for April, 1962. In prostheses previously developed, the only feedback loop has been the mainly visual one, and all control information has been supplied by the person himself, part of his brain being monopolized by the requirements of his prosthesis. In the hand described, sensory pressure pads are fixed to the finger tips, palms and wrists, and arranged to provide two feedback loops, one of which is adaptive. The pressure pads are resistive, and act as the "lower halves" of potential dividers, outputs being used to control the rotation of the driving servo motor. The fingers are connected in parallel and linked by springs, whereas the thumb is a separate mechanical element. Application of pressure to the hand causes the fingers and thumb to close, positive feedback *via* the pressure pads and a follow-up potentiometer ensuring a tight grip. Non-linearity of the system provides a limiting force. The cables driving fingers and thumb are attached to the motor pulley 90° apart, so that, depending on which direction the motor assumes, either the fingers or thumb will close first. In this way, a grasping or pinching action is obtained, depending on whether the palm or finger pressure pads are activated. Adaptive control of wrist lifting force is provided by pressure pads in the wrist, the outputs of which are arranged to control the reference phase of the servo motor.

Spark chamber nuclear particle track tracer (hodoscope) is being studied at the Clarendon Laboratory, Oxford. This device consists of a set of parallel metal plates spaced about a quarter of an inch apart with an inert gas between them. Any charged particle passing through the chamber ionizes the gas along its path. These ions are accelerated in a high electric field which is suddenly set up in the gas (by applying several kilovolts between alternate plates). This results in the generation of many more ions so that a spark appears between each pair of plates along the track of the original nuclear particle. Observation only of nuclear events of interest can be ensured by detecting the incident and product particles (in scintillation or Cherenkov counters) and only triggering on the high voltage supply when a suitable event occurs. A feature of this type of detector is that it is sensitive only for a very short time ($\approx 1\mu$sec).

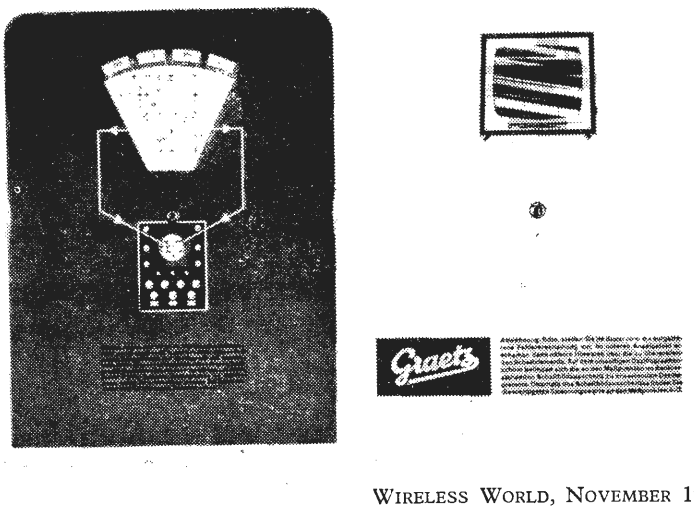
Gas-filled rectifiers which give a visual indication of approaching failure and which have a 50 to 100% longer than usual life are being manufactured in America by National Electronics and distributed in this country by Walmore Electronics. The longer life is obtained by using a cathode of the matrix nickelate type (produced by sintering a mixture of barium carbonate and powdered nickel on to a nickel base)—this is less easily destroyed by ion bombardment than a normal double or triple carbonate coated cathode. Visual indication of approaching failure is obtained by adding a small quantity of neon to the xenon gas normally used in such devices. Under normal tube operation a blue or light purple colour is produced in the tube by ionization of the xenon: the neon is not ionized owing to the higher potential required. However, near the end of the tube life the xenon gas becomes "cleaned-up" and the arc potential drop increases until it is sufficient to ionize the neon. This adds the characteristic bright red glow of ionized neon to the tube colour and so indicates approaching failure. When all the xenon is cleaned up the arc drop becomes sufficiently high to result in the destruction of the cathode emissive coating by (neon) ion bombardment.

Electrostatic recording is discussed in the April 13, 1962, issue of *Electronics* (p. 74). The recording signal (at a level $\approx$ 100V r.m.s.) is applied between a knife-edged electrode and flexible backing electrode, and an insulating tape drawn between these electrodes. This records a charge pattern on the tape which can be replaced by passing it between a similar pair of electrodes (giving $\approx$ 40mV into 10MΩ). Such recordings can be played back several hundred times with little loss of signal and have lives of several hundred years. To facilitate charge injection into the tape a d.c. bias ($\approx$1500V) is also applied to the signal electrode. The signal-to-noise ratio and distortion can be improved by the application of an additional reversed-polarity pre-bias (between similar electrodes) before the tape reaches the recording electrodes. The optimum pre-bias is found to be equal to the original bias so that an essentially neutral tape would be produced in the absence of a recording signal. Signal-to-noise and distortion can be still further improved by replacing the d.c. bias by a.c. (at a frequency $\approx$ 5Mc/s) applied, with the recording signal, between a pair of knife-edged electrodes on opposite sides of the tape. An atmospheric ion bath is used to neutralize net charge on the tape after recording and playback: this increases both the storage and replay life and also reduces print through. Signal resolution on the tape is influenced by the effective width of both the recording and playback heads. With simple knife-edged electrodes the minimum wavelength is about 0.7mil and at the tape speed used ($\approx 7\frac{1}{2}$in/sec) this gives an upper frequency response limit of about 10kc/s. Resolution can be improved by adding a close-fitting electrostatic shield on either side of the electrode: this reduces the resolution to the distance between the shields.

Laser peak power increase may be obtained by decreasing the total period in which the output power can be emitted. At the N.P.L. this is being done by using a rotating mirror for one of the reflectors (which pass light back and forth through the laser ruby rod). Laser action is then only possible in the short period in which the rotating mirror is in the correct position to produce such back and forth reflection.

TELEVISION FAULT TRACER

AN ingenious aid to the diagnosis and correction of television receiver faults, described as a Reparatur-Uhr (Repair Clock), has been devised by the Service Department of Graetz G.m.b.H., Altena/Westf., Germany. A circular card, printed on both sides, rotates inside a cover with windows back and front. On one side a series of typical picture faults appear in an aperture and below the most likely stage involved is indicated, together with the valve or diode type and the associated resistors or capacitors which should be checked. On the other side of the card the circuit diagram of the stage is shown with check points and voltages, and, where appropriate, the waveform which should be seen on an oscilloscope when connected to the points indicated.



FUNDAMENTALS OF FEEDBACK DESIGN

11.—DISTORTION

By G. EDWIN

THE question of the effect of feedback on the distortion of a system is one which offers the possibility of treatment at a number of different levels and is a topic which is probably more suitable for discussion than for analysis. The simplest treatment is the traditional textbook one which can be applied to both distortion and supply noise. Consider the amplifier to be split at the point at which the unwanted term appears: there will not, in fact, be such a single point, but the Superposition Theorem makes it quite permissible to consider each contribution independently. The circuit takes on the appearance shown in Fig. 62. The only signal to be considered is the noise or distortion term S which is introduced a "distance" μ_2 from the output of a system having a total forward gain of $\mu = \mu_1 \mu_2$ and feedback β. The resulting output is V and following this back we get an apparent input signal of βV and an output from the μ_1 section of $\mu_1\beta$V.

At the input of μ_2 we combine S and $\mu_1\beta$V to provide the input to the μ_2 section. Immediately, therefore, with the usual convention we must get for the output,

$$V = \mu_2 (S + \mu_1\beta V)$$

$$\text{or} \quad V = \mu_2 S/(1 - \mu_1\mu_2\beta)$$

When we put $\beta = 0$ we get $V_0 = \mu_2 S$, so that the effect of connecting the feedback path is to reduce the unwanted output in the ratio $V_0/V = (1 - \mu_1\mu_2\beta) = (1 - \mu\beta)$.

On the simple theory, therefore, an unwanted term which appears anywhere inside the amplifier will be reduced by the factor $(1 - \mu\beta)$. It is probably worth pointing out that for a given wanted signal output the application of feedback does not affect the levels inside the loop. All that is required is the provision of an appropriately changed input level. Thus the distortion generated will be the same. When there are internal feedback loops these, by altering the demands made on the other parts of the circuit, may alter the contributions made by these other parts and may affect the overall picture as a result of this interaction. Essentially, however, S remains unchanged, and the distortion fraction is improved by the factor $(1 - \mu\beta)$.

For conservatively designed systems this is a perfectly adequate result. When an amplifier producing 2% distortion has feedback applied to reduce the gain by 26dB the distortion will be reduced to 0.1%. So long as this simple statement is enough there is no need to consider the way in which the distortion is generated and reduced any further.

When an amplifier has been constructed on this simple basis it is found that at low signal levels it behaves according to the theory. However, as the input level is increased and the feedback without distortion rises, the effect of the negative feedback appears to be reduced, and above a fairly sharply defined knee the distortion with negative feedback increases very rapidly indeed. We shall discuss this effect shortly, but it is necessary to notice first that it is an effect which has some rather awkward implications. Most practical amplifiers incorporate internal feedback loops, which may be merely cathode resistors left undecoupled or may be more complex arrangements. These will have their own

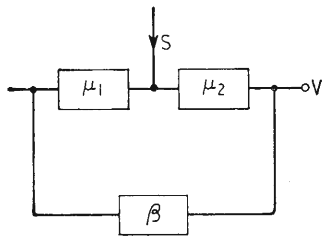


Fig. 62. Feedback amplifier with extraneous noise or distortion S introduced at a single point.

"knee" effects, so that some of the assumptions made in examining the simple theory more closely will not be true, even for the low levels of distortion we may be considering.

The first point which must be taken into account is the obvious fact that μ is not a constant: if it were the system would not produce any distortion and the whole discussion would be meaningless. Distortion is the result of the fact that the amplifier gain depends to some extent on the value of the instantaneous signal. Commonly the effect is that the value of μ is reduced for large signal excursions. It is this reduced value of μ which is appropriate for insertion in the expression $(1 - \mu\beta)$ which in turn determines the factor by which the distortion is to be reduced. Naturally this statement is only an approximation, but it is one which errs on the conservative side, as distinct from the usual approach which certainly leads to a result which is found by painful experience to be optimistic.

Three Types of Gain / Level Variation

We can distinguish three cases. A simple amplifier without internal feedback loops will have a smooth gain/level characteristic showing no sharp change of gain. Much of the effect will be due to the final stage which swings over the widest range, and the gain variation might be, perhaps, 2 to 1. But this is the sort of uncertainty we have in the forward gain μ anyway, and if the distortion is to be knocked down from 5% to 0.5% we must allow for low-gain stages and work with a design centre feedback of 26dB rather than 20dB. At high output levels the simple amplifier will overload, forcing μ right down to a very low value. We enter this region more sharply with the feedback applied, but although outside it the feedback is making the output waveform a true replica of the input, once overloading occurs and μ goes well down, the factor $(1 - \mu\beta)$ is small and we have no help from the feedback. The third case, the amplifier with internal feedback loops, shows this effect in an enhanced form. The overall forward gain will be pretty constant and the distortion low even without overall feedback up to the point where overloading begins. The internal feedback loops sharpen up this effect, of course, and the overall feedback can do nothing but make the clipped appearance of the waveform still more pronounced.

Gain / Frequency Variations

The second point which must be taken into account is the obvious fact that μ depends not only on amplitude but on frequency. When a signal of some frequency f_0 is applied to the input the output will contain components $2f_0$, $3f_0$, and so on. These are fed back to the input and are reduced by factors $(1 - \mu_{2f0}\beta)$, $(1 - \mu_{3f0}\beta)$ and so on. The very simple test of the feedback is to see how much the gain is reduced, but this is only the factor $(1 - \mu_{f0}\beta)$ and may be very much greater than the number which really should be used. Transistor power amplifiers in particular require careful study in this respect. The cheaper power transistors, used in common emitter configuration, have characteristic frequencies of about 4,000c/s so that at 8,000 c/s the stage gain is 6dB down and the stage phase-shift some 60°. Naturally this is in addition to the usual gain and phase terms associated with the rest of the amplifier and with the output transformer. The attempt to make a good amplifier using this sort of transistor leads to the introduction of step circuits which cut the gain even more at frequencies which may be as low as, in one design study, 2,000 c/s. Overall feedback can be used to produce an apparently good frequency response: the feedback can be described as being, say, 26dB at 1,000c/s, and yet at 5,000c/s it may be only perhaps 12dB while at 10,000c/s the feedback may be positive. The low harmonics of the lower audio frequencies will get the full benefit of the 26dB of negative feedback, but the higher harmonics of the higher audio frequencies will actually be increased.

Certainly this is an exceptional case, for slightly more expensive transistors of slightly lower power-handling capacity have very much higher cut-off frequencies. The principle, however, remains, and must be borne in mind whenever an attempt is being made to reconcile the theory with the experimental results in an amplifier in which feedback has been relied on to ameliorate an otherwise indifferent performance.

It is questionable whether the production of harmonics is, in itself, a serious matter in the reproduction of speech or music. The addition of harmonics to the note of the ocarina may produce a flute-like effect, but it might be considered that changes in the tone of the ocarina can be only for the better. Most musical instruments produce waveforms which are already rich in harmonics and the small changes in harmonic content with which we are dealing normally could be expected to be insignificant. Much more important is the effect of intermodulation. Both the calculation and the measurement of intermodulation are much more complex than the calculation and measurement of distortion. Thus discussions of intermodulation tend to be restricted to purely professional applications, although the search for high-fidelity reproduction is quite largely outside the normal run of professional amplifier design.

Intermodulation

Intermodulation is, in very simple terms, the distortion of one signal by another which is passing through the amplifier at the same time. Let us consider that our amplifier is set up to produce from a loudspeaker a level which is about average American conversation level. To sound equally loud a 30c/s tone will need to be some 15-20dB higher in level. We can normally expect, therefore, that the lowest frequency signals will have relatively very large amplitudes if we are to hear them. The peaks of these signals will drive the amplifier right up into the region where the gain is falling and the system will dwell there for several milliseconds every, for a 30c/s signal, 167 milliseconds. The much more audible signals between 500c/s and 1,500c/s will be presented, as it were, with this variable gain situation and will have their amplitudes modulated. If the low and almost inaudible signal is large enough, the middle frequencies may be chopped out completely, especially if the amplifier has a good deal of negative feedback.

In a sense, the distortion is transferred from the large low-frequency signal (of which it forms a small part in an insensitive region of the spectrum) to the smaller middle-frequency signal (where it appears as a large proportion of the whole in a particularly sensitive region). This mechanism, with its many factors, suggests that it will be difficult to find a single number which can be used as a measure of quality, and it is clearly this difficulty which hampers the development of cheap and simple test equipment.

Negative Feedback and Intermodulation

Negative feedback may, in marginal cases, result in a serious increase in intermodulation effects. For a fixed amount of distortion there may be a smooth and slightly curved input/output characteristic or a very linear characteristic terminated by a sharp clipping region. The effect on the small middle-frequency signal will obviously be much more serious when sections are chopped out, the condition we meet with the linear to clipping type of characteristic, the type we get when negative feedback is large.

A second form of intermodulation which does not raise quite the same problems is produced when two high audio-frequency signals produce a difference tone at a more easily heard frequency, especially if the result is discordant. Here negative feedback will usually be able to play its full part and it is thought that this is not a serious cause of poor subjective quality.

One cause of poor performance with negative feedback which does not lend itself to analysis and which is commonly neglected arises from the effect of feedback on the transient response. Addition of feedback usually produces an amplitude response which is flat over the working band and then cuts off rather sharply: the phase characteristic shows a relatively small phase shift over much of the band with a steep rise near the edge. The net effect is that the transient response often has a substantial initial overshoot. A physical interpretation is that when the signal is applied it passes straight through the amplifier to come out, rounded by the amplifier behaviour, after a short delay. But at this stage it is at the full gain μ since until the signal reaches the output it cannot join the β-path to return to the input. Then this high level goes back through the feedback to make an initial over correction, producing a new output level, and so on. We can see this analytically by writing the simple expansion:

$$e_{out} = e_{in}\,\mu/(1-\mu\beta)$$
$$= e_{in}\,\mu[1 + \mu\beta + (\mu\beta)^2 + (\mu\beta)^3 \ . \ . \ .]$$

Under suitable conditions the beginning of a transient can produce enough grid current to charge one of the capacitors and shift the working point of the amplifier. This is not a trouble which is met in sound design, for a transient ring is a sign of a rather uncontrolled response. Here, however, is another reason why care must be taken.

MARCONI MARINE "ARGUS" RADAR

INCORPORATING THE "BONUS" (BOW OR NORTH-UP STABILIZED) SYSTEM

IT has long been appreciated by navigators that if precious minutes, even seconds, are to be saved in evaluating a situation involving an apparently unchanging compass bearing, then compass stabilization must be applied to the radar display. Also, after an avoiding alteration of course, accurate observation of the further development of the situation, either on the p.p.i. or on a reflection plotter, again demanded a fixed north-up stabilization. North-up stabilization is, of course, an integral function of a true-motion radar display.

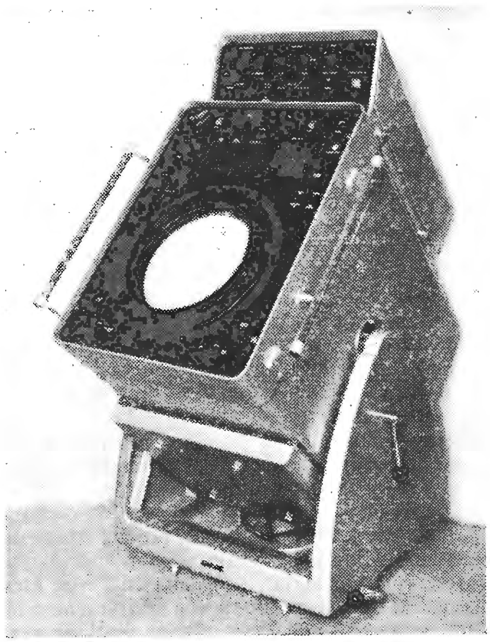
In direct conflict with this desideratum is the very understandable and human desire of the navigator to see, on the radar screen, a "bird's-eye view" of the situation as he sees it from the bridge of his ship. He does not want—and should not have—to turn himself round, physically or mentally, through a certain number of degrees in order to equate in his mind what he sees on radar with what he might see with his own eyes from the bridge of his ship.

This is liable to be at its most aggravating and confusing in the not uncommon situation where the navigator finds himself deriving information on the movements of other ships from his own eyes and ears as well as from the radar screen. This is the time when most of all he needs a radar picture in precisely the same orientation, i.e., ship's-head-up, as the real-life observations he makes through the wheelhouse window or from the wing of the bridge. At such a time seconds may be vital, and the difficulty of correlating, accurately and instantaneously, the radar and real-life views, when one may be upside-down in relation to the other, and with port and starboard reversed, cannot be over-emphasized.

True Motion With Ship's Head Up

To meet his conflicting requirements the navigator has until now been able to have either the benefits of compass stabilization and true motion, or the display with the natural ship's-head-up alignment; but not both together. True, some ships have gone to the length of fitting two independent radar installations to provide both forms of display simultaneously, but this is a costly and cumbersome expedient and one practicable only on ships with room to spare on the bridge. Even then the bow-up display, being unstabilized, suffers from the effect of yaw; and the possibility of a dangerously wrong conclusion being drawn from confusion of target identification on two separate screens must be considered.

In the new "Argus" the navigator's two opposed requirements are successfully reconciled, thanks to the employment of the BONUS technique. This is achieved by stabilization, not only of the bearing ring, but also of the tube itself which in the BONUS system rotates within the display unit in correspondence with every change of course. The navigator will now have the benefits of compass stabilization continuously, plus true motion if he wishes it, and will at the same time be able to switch to ship's-head-up or north-up as he wishes to provide him



In the Marconi Marine "Argus" radar, not only the rotation of the bearing ring but the tube itself can be compass-stabilized, giving rapid choice of either north-up or ship's-head-up, without smearing after-glow traces.

with the type of presentation he needs for swift and accurate appreciation of any situation.

Selection of natural ship's-head-up or north-up orientation does not affect any other function of the display; neither is any other display function affected by selection of either relative or true motion presentation. Compass coupling is engaged at all times to eliminate "picture smear" and to make possible the direct reading of true bearings and the accurate monitoring of situations showing unchanging compass bearings, even at the moment when a substantial course alteration is being made.

The great feature of the new "Argus" and its stabilized screen is its ability to give the navigator a full true motion picture but with own ship always moving up the display, i.e., always ship's-head-up. With a sharp alteration of course, the p.p.i. itself "alters course" so that the previous line of own ship is now at the appropriate angle to the fore and aft diameter of the display just as the ship's wake is to the line of its new course. All other targets, whether moving or stationary, also swing round on the screen as they do on the sea. For example, another ship may be observed fine on the port bow and on an approaching course. Own ship alters course 90° by going hard-a-starboard, but after this alteration is still, on the p.p.i., proceeding up the screen, while the other ship is now seen to the left-hand or port side of the screen and, provided her distance and speed are appropriate, will pass safely clear under own ship's stern. A later reversion by own ship to original course will again produce a correspondingly proportionate rotation of the p.p.i. and show the other ship clear away on the port quarter of the display—exactly as she would be seen by the eye from own ship's bridge. If the navigator then wishes to change to north-up presentation, perhaps to check on the chart his new position against an adjacent landmark, operation of a single switch is all that is necessary. He need not hesitate to switch from the ship's-head-up to the north-up mode of presentation (or vice-versa) since the change of presentation is completed swiftly and without any smearing of the picture and with no interruption of established after-glow tracks or plots.

The "Argus" will be available with or without true motion, although it is anticipated that the majority of orders will stipulate true motion facilities in order to take full advantage of the new BONUS technique. The production set has a 12-inch p.p.i. and uses a slotted waveguide aerial of the latest type with narrow beam width and a rotation speed of 25 r.p.m. The aerial is similar to that employed in the recently announced "Hermes" radar installation and is available in two types, one 6 ft long, the other 12 ft.

The transmitter has a power of 70 kW nominal, and will operate on dual pulse lengths to eight closely spaced display ranges of $\frac{3}{4}$, $1\frac{1}{2}$, 3, 6, 12, 18, 24 and 48 miles.

The "Argus" has been given, among many other refinements, a high definition screen, the improved slotted waveguide aerial, transistorized true motion circuitry and automatic re-setting of the true motion tracking system. The true motion tracking system can, of course, also be reset by a touch on a push-button at any time and in any direction to give an extended picture, but if the navigating officer is engaged upon other duties and own ship reaches a predetermined distance from the edge of the screen it will automatically return to the pre-selected point of origin and start tracking again. When not using true motion this push-button reset facility may be employed to off-centre own ship and hold an extended view in any direction for as long as may be required.

The eight ranges of the "Argus" are spaced to cover any situation which may occur and a reflection plotter can be fitted over the screen for detailed plotting of the courses and speeds of other ships within range.

Designed for installation on passenger and cargo liners, the aerial/transmitter system is capable of operating two display units each with independent selection of mode of presentation, sensitivity, anti-clutter, range, and ranging controls. Each display unit may be installed up to 1,000 ft in cable length from the transmitter/receiver unit which, for optimum performance, can be sited close to the aerial unit.

The "Argus" and the "Hermes" were developed as "side-by-side" projects, the aim being to produce a good conventional radar, the "Hermes," and a completely new stabilized-screen radar, the "Argus," which would employ the same aerial, transmitter/receiver, and power units, differing only in the choice of display unit and the facilities it offers. These displays are, in effect, fully compatible and may, if required, be installed in conjunction with each other, using one common aerial and transmitter/receiver.

Commercial Literature

Teleng have designed a new range of wide band amplifiers and channel converters for communal aerial systems, which is fully described in the latest and enlarged edition of their "U" series catalogue. Teleng Ltd., Church Road, Harold Wood, Romford, Essex.

Copper and resistance wires: an extensive range of copper instrument wires, Litz wires, Eureka and Constantan resistance wires, nickel-chrome wires, and copper wires with a variety of synthetic enamel coverings, are listed in a pamphlet available from Post Radio Supplies, 33 Bourne Gardens, London, E.4.

Principles of operation of the image orthicon television camera tube together with general data on the three $4\frac{1}{2}$-in tubes manufactured by English Electric Valve Company are given in an illustrated brochure available from the company at Chelmsford, Essex.

Moving coil panel meters, of Taylor manufacture, are described and illustrated actual size in a brochure available from Taylor Electrical Instruments Ltd., Montrose Avenue, Slough, Bucks.

Ultra-pure elements, single crystals and inorganic compounds are catalogued in a new publication by the Pure Elements Division of L. Light & Co. Ltd., Poyle Colnbrook, Bucks.

Lasky Hi-Fi Catalogue (second edition) contains 122 pages of specifications and illustrations of a wide range of equipment for the audio enthusiast. Copies (5s 6d post free) from Lasky's (Harrow Road) Ltd., 207 Edgware Road, London, W.2.

Civil communications equipment is the subject of a new 32-page illustrated brochure by Plessey. The range of units and terminal equipments described are suitable for use in C.W., M.C.W., R/T, facsimile transmission and F.S.K. civil applications. Copies of this publication, No. 248, from the Plessey Co. (U.K.) Ltd., Ilford, Essex.

Wire-wound Resistor Catalogue listing vitreous enamelled types with resistance values from 0.1 to 120,000 Ω and in wattage ratings from 4 to 14, is now available from the C.G.S. Resistance Co. Ltd., at Marsh Lane, Gosport Street, Lymington, Hants.

NOVEMBER MEETINGS

Tickets are required for some meetings; readers are advised, therefore, to communicate with the secretary of the society concerned.

LONDON

1st. Television Society.—"Characteristics of special vidicon camera tubes and their application" by A. C. Dawe at 7.0 at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, W.C.2.

6th. I.E.E.—"Electrical and electronic problems in Service aircraft" at 6.0 at Savoy Place, W.C.2. (Joint meeting with the Royal Aeronautical Society.)

7th. British Interplanetary Soc.—One-day symposium on "Ground support equipment" at 9.30 at Royal Aeronautical Soc., 4 Hamilton Place, W.1. (Fee 2gn.)

7th. I.E.E.—"V.H.F. diffraction problems introduced by the rough earth" by G. Millington at 5.30 at Savoy Place, W.C.2.

7th. Brit.I.R.E.—Discussion on "The teaching of communications in a modern context" at 6.0 at London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.

8th. I.E.E.—"Optical masers" by I. L. Davies at 5.30 at Savoy Place, W.C.2.

8th. Radar & Electronics Association.—"Satellite communication systems" by W. J. Bray at 7.0 at the Royal Society of Arts, John Adam Street, W.C.2.

13th. I.E.E.—"Some aspects of the use of computers in process control application" by J. F. Roth at 5.30 at Savoy Place, W.C.2.

14th. I.E.E.—"Stereophonic Broadcasting Systems" by Dr. G. J. Phillips at 5.30 at Savoy Place, W.C.2.

14th. Brit.I.R.E.—Papers on "Radio transmission of digital data" at 6.0 at London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.

14th. Society of Environmental Engineers.—"Analysing the results of vibration tests with f.m. tape recordings" by S. D. Hall at 6.0 in Mechanical Engineering Department, Imperial College, Exhibition Road, S.W.7.

15th. Electronic Organ Constructors' Soc.—Demonstration of transistor organs at 7.0 at the Arsenal Tavern, Blackstock Road, N.4.

16th. Institute of Navigation.—"Automation and marine navigation" by C. T. Clayton at 5.30 at Royal Institution of Naval Architects, 10 Upper Belgrave Street, S.W.1.

16th. Television Society.—"An 819-line transistor television receiver" by a member of La Compagnie Générale des Semi-conducteurs at 7.0 at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, W.C.2.

19th. I.E.E.—Discussion on "The future of global communications" opened by R. J. Halsey at 5.30 at Savoy Place, W.C.2.

20th. I.E.E. and I.Mech.E.—Discussion on "Components used in automatic control systems" at 5.30 at 1 Birdcage Walk, S.W.1.

21st. I.E.E.—Discussion on "Microminiaturization" at 5.30 at Savoy Place, W.C.2.

21st. Brit.I.R.E.—"Automatic biochemical analysis and particle counting" by Dr. J. F. Marten and M. I. Henderson at 6.0 at London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1. (Joint meeting with I.E.E. Medical Electronics Discussion Group.)

21st. British Computer Society.—"An introduction to P.E.R.T. [Programme Evaluation and Review Technique]" at 6.30 at Northampton College of Advance Technology, St. John Street, E.C.1.

21st. British Kinematograph Society.—"Technical requirements for a television news service" by W. H. O. Sweeny at 7.0 at Central Office of Information, Hercules Road, Westminster Bridge Road, S.E.1.

22nd. Institute of Phys. and Phys. Soc.—"The sun, the earth and radio" by J. A. Ratcliffe at 5.30 at 47 Belgrave Square, S.W.1.

26th. Brit.I.R.E. — "Transistorized television camera chain" by J. Wilson and A. W. Comer at London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.

28th. Society of Environmental Engineers.—"Sweep random vibration" by G. B. Booth at 6.0 in Mechanical Engineering Department, Imperial College, Exhibition Road, S.W.7.

28th. British Kinematograph Society.—"'Candid camera': technical and production problems" by R. Taylor at 7.0 at Mezzanine Cinema, Shell-Mex House, Strand, W.C.2.

29th. I.E.E.—Discussion on "Training for the profession of electrical engineering" opened by S. E. Goodall at 5.30 at Savoy Place, W.C.2.

BASINGSTOKE

8th. Brit.I.R.E.—"Applications of lasers to communications" by Dr. R. C. Smith at 7.0 at Basingstoke Technical College.

BIRMINGHAM

9th. Society of Instrument Technology.—"Quality measurement and control" by W. H. Topham at 7.0 at the Lecture Theatre, Bing Kendrick Suite, College of Advanced Technology, Gosta Green.

15th. Brit.I.R.E.—"Recent developments in computer storage" by Dr. D. B. G. Edwards at 6.15 at Electrical Engineering Department, The University.

BRISTOL

7th. Brit.I.R.E.—"Recent progress in research into the ionosphere" by G. M. Brown at 7.0 at the University Engineering Lecture Rooms, Queens Building.

13th. Television Society.—"SECAM colour television system" by G. B. Townsend at 7.30 at Royal Hotel, College Green.

CARDIFF

7th. Brit.I.R.E.—"Modern methods of investigating the brain and behaviour" by Dr. R. Cooper at 6.30 at the Welsh College of Advanced Technology.

CHELTENHAM

2nd. Brit.I.R.E.—"The application of automation in the pulping industry" by D. A. Mackintosh at 7.0 at North Gloucestershire Technical College.

EDINBURGH

7th. Brit.I.R.E.—"Line fault location" by P. A. James at 7.0 at the Department of Natural Philosophy, The University, Drummond Street.

GLASGOW

8th. Brit.I.R.E.—"Line fault location" by P. A. James at 7.0 at the Institution of Engineers and Shipbuilders, 39 Elmbank Crescent.

30th. Society of Instrument Technology.—"Navigational instruments" by Captain Watkinson at 7.15 at the Scottish Building Centre, 425 Sauchiehall Street, C.2.

LIVERPOOL

21st. Brit.I.R.E.—"The problems of blind landing of aircraft" by F. R. Gill at 7.30 at the Walker Art Gallery.

MANCHESTER

1st. Brit.I.R.E.—"Automatic marshalling yards" by R. M. Foulkes at 7.0 at Reynolds Hall, College of Science and Technology.

NEWCASTLE-UPON-TYNE

14th. Brit.I.R.E.—"Modern transmitter techniques" by W. J. Morcom at 6.0 at the Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road.

NOTTINGHAM

12th. Television Society.—"TV under the Camera" by H. J. Barton-Chapple at 7.30 at the Co-op Educational Centre, Broad Street.

14th. Brit.I.R.E.—"Communications satellites" by L. F. Mathews at 6.45 at the University of Nottingham.

PORTSMOUTH

14th. Brit.I.R.E.—"Applications of lasers to communications" by Dr. R. C. Smith at 7.0 at Portsmouth College of Technology.

21st. I.E.E.—"Review of syncro and servo components" by C. G. A. Woodford at 6.30 at College of Technology, Anglesea Road.

SCUNTHORPE

12th. Society of Instrument Technology.—"Transistorised continuous belt weighing and controlling" by W. E. Watts at 7.30 at Leggott Grammar School.

SOUTHAMPTON

13th. I.E.E.—"Charge definition of transistor parameters" by Dr. A. R. Boothroyd at 6.30 at The University.

27th. I.E.E. and Brit.I.R.E.—"Fields of application of absolute and incremental methods of numerical control" by K. J. Coppin at 7.0 at the Lanchester Building, The University.

UNBIASED

By "FREE GRID"

Breaching the Peace

IT is strange how the transistor set craze on beaches, buses and almost everywhere else has grown, and it is not surprising that letters have appeared in the national press about it, suggesting that it causes a breach of the peace in the ordinary sense of the phrase, and it is liable to do so in its legal sense. A correspondent in the *Daily Telegraph* who said he was able to jam transistor receivers by means of a simply constructed device was besieged by requests for details of it.

When broadcasting first began, most listeners possessed a device of this nature, namely, a simple regenerative receiver which, when improperly adjusted, used to cause enough whistles to mess up reception for other listeners. This effect used to be so serious that it often brought an anguished cry of "Please, don't do it," from P. P. Eckersley, who was in charge of the Writtle experimental station and later became the B.B.C.'s chief engineer.

Such jamming was not, of course, deliberate but it was very effective. To my warped mind it seems a pity that it is "agin the law" to indulge in it, more especially as in these days of Lilliputian apparatus it would be easy to construct a very compact pocket oscillator. This would effectively purge the surrounding area of the sound of the pestiferous outpourings produced by these portable purveyors of parasitic polyphony.

However, in my opinion, a far greater menace to the Queen's Peace is the ubiquitous Tom-Thumb tape recorder. Nobody knows when his off-the-record remarks are becoming on-the-record ones since these devices are so easily concealed in a pocket with the midget mike hidden in a gay buttonhole. The so-called candid camera is a big enough menace, but these tiny tape recorders are far worse.

I have been racking my brain to find a way of combating them. The ideal solution would, of course, be for us all to carry on us magnetic wipe-out apparatus which we could set into operation whenever we were holding a conversation. However, the Editor, who is one of those irritating practical down-to-earth sort of fellows, reminds me that for it to be effective, the magnetic field and the apparatus generating it would have to be of a similar order of magnitude to those associated with a cyclotron.

There must, however, be some solution to the problem, and I hope you can offer a few suggestions. Meanwhile I carry a large buzzer* in my bowler to cause acoustic interference and I always switch this on when I suspect my remarks are being taped; but it makes normal conversation a strain.

* Not to be confused with the bees that occasionally buzz in his bowler!—ED.

Defining Valves

NOWADAYS we talk about transistors and valves just as though a transistor were not a valve. We did the same sort of thing 40 years ago at the beginning of broadcasting when we spoke of crystals and valves even though a crystal was just as much a valve as the thermionic type.

No radio receiver has ever been produced—or, indeed, could be produced—which did not embody a valve, using the word in its literal sense as an open-and-shut device or, in other words, a door. Usually the incoming radio signals open the "door" to release a local source of energy. They did that in the coherer which was the very first practical radio receiver; I don't count the Hertz resonator which was only of academic interest.

In the coherer receiver, the feeble incoming signals closed a local circuit and so released a relatively large amount of power which was supplied by the usual electric cell. It was thus an amplifier as well as a receiver. In fact, the coherer set-up consisted of a couple of relays in cascade and so was entitled to be called a multi-valve receiver.

The second practical receiver, namely the magnetic detector, which made its commercial debut in 1902, was also very definitely an amplifier-receiver which triggered off a local supply of energy. In this case, however, the local energy was not supplied by an electric cell but by two permanent magnets. The field of these magnets passed through the slowly moving soft-iron band and owing to hysteresis, the lines of force were dragged after it for a short distance as it travelled along.

The weak incoming signals passed through the few turns of a coil in series with aerial and earth. This coil was wound on a small glass tube through the bore of which the moving soft-iron band passed, and therefore the incoming radio-frequency signals cancelled out the drawn-out magnetic field by the same principle as is used in the wipe-out coil of our tape recorders. The collapsing magnetic field cut through the secondary coil, which was wound over the primary, and so caused a hefty click in the phones which were connected to its ends.

The next detector was the Fleming thermionic diode. This was simply a non-return valve which rectified the incoming signals and did not control any local source of energy. A dry cell used to be employed to adjust the detector to its most resistive spot, namely its bottom bend, and exactly the same thing was done in the carborundum crystal.

I dare say there are some who would argue that because of this adjustment we were really deriving some of the energy in our phones from the dry cell. Certain schools of philosophy would point out that they were correct in the sense that if the bias were not applied no signals at all would be heard unless the transmitter were very close or very powerful. But that argument could not be used in the case of the ordinary galena crystal which we all used in our sets when broadcasting began, and which, of course, had no biasing cell.

It will be seen, therefore, that although all wireless sets are valve sets, the crystal and simple diode types are receivers pure and simple and not amplifiers as well. Now, I suppose, somebody will write to the Editor describing a wireless receiver which employs no valves of any kind. Needless to say, I shall be most interested.

Electronics and Electrotonics

MANY people have tried to define the word "electronics" but as far as I am aware no satisfactory definition has been found. The Editor dealt with the matter in the issue of May 1958, and I am not going to discuss it further now.

What I do want to know, however, is when the term "electronics" first appeared. I do not mean the adjective "electronic" so much as the noun "electronics," but I should like to know the debut dates of both. The dictionaries don't help me much although they all have a lot to say about the word "electron." Can anybody help me?

I greatly doubt that the word "electronics" was in use during Edwardian days, but I may be quite wrong. During my dippings into the dictionaries I was reminded of many things I had forgotten. I must confess that until the "O.E.D." reminded me of it I had forgotten that Faraday had coined the word "electrotonicity" to describe the current in the secondary of a transformer; presumably one with a lot more turns in the secondary than the primary winding, as the central "ton" is obviously an abbreviation of the Greek word "tonos," which means tension.

I wonder at what voltage "electricity" ends and "electrotonicity" begins. It is the sort of thing lawyers could argue about for hours in court as I don't suppose it has ever been authoritatively decided. I think Faraday missed a great opportunity in not coining also the word "electrotonics" to describe the phenomena associated with high-tension current.

For Men Only?

IT is interesting to note that the B.B.C. is now giving us one-channel stereo transmissions instead of the two-channel ones which use up such a lot of elbow room. I see that in some quarters the dreadful expression "multiplex stereo" is being used to describe the system. To my mind the word "multiplex" is all wrong when used in this connection as it is far too vague and utterly non-descriptive. Long ago I suggested the name homodic stereo for this type of transmission, as its literal meaning is "same road" and so, by inference, "same carrier" used for both components of the stereo programme.

Naturally nobody took any notice of me any more than they did when I suggested monodic as the correct word for that which some people call, with ludicrous inaccuracy, monaural, although most scribes nowadays use the far less offensive but not strictly accurate word monophonic which has apparently come to stay.

I won't say any more about that lest I raise another hellenic hullabaloo as I did a few years ago. I do think, however, that we might talk of homostereo (short for homodic stereo) to distinguish the system from two-channel stereo broadcasting. If the latter wants a distinguishing name it could be called diodic stereo. Heterodic stereo would clearly convey an entirely wrong meaning; and certainly not "biodic," please.

However, even with my suggested word "homostereo" to describe one-channel stereo there is obviously a risk that some ardent feminist might think that the B.B.C. was transmitting special stereo programmes "for men only."

1963 CONFERENCES AND EXHIBITIONS

Latest information on next year's events both in the U.K. and abroad is given below. Further details are obtainable from the addresses in parentheses.

LONDON

Jan. 14-17 — R.H.S. Halls
Physical Society Exhibition
(Inst of Physics & Phys. Soc., 47 Belgrave Square, S.W.1)

Mar. 25-27 — Savoy Place
H.F. Communication Convention
(I.E.E., Savoy Place, W.C.2)

April 18-21 — Hotel Russell
Audio Festival & Fair
(C. Rex-Hassan, 42 Manchester Street, W.1)

April 23-May 2 — Earls Court & Olympia
International Engineering Exhibition
(F. W. Bridges, Grand Buildings, Trafalgar Square, W.C.2)

May 6-9 — 1 Gt. George St., S.W.1.
Productivity and the Engineer Conference
(Institution of Production Engs., 10 Chesterfield St., W.1)

May 21-24 — Olympia
Components Exhibition
(Radio & Electronic Component Manufacturers' Federation, 21 Tothill Street, S.W.1)

May 27-31 — Olympia
Hospital Equipment International Exhibition
(Contemporary Exhibitions, 288 Regent Street, W.1)

June 12-22 — Olympia
International Plastics Exhibition and Convention
(*British Plastics,* Dorset House, Stamford Street, S.E.1)

Sept. 9-13 — Church House
Non-Destructive Testing Conference
(Institution of Mechanical Engs., 1 Birdcage Walk, S.W.1)

September — Savoy Place
Design and Use of Microwave Valves
(I.E.E., Savoy Place, W.C.2)

October — Savoy Place
Automatic Production in Electrical and Electronic Enginering Symposium
(I.E.E., Savoy Place, W.C.2)

Nov. 11-16 — Earls Court
Industrial Photographic & Television Exhibition
(Industrial & Trade Fairs, Commonwealth House, New Oxford Street, W.C.1)

DURHAM

April 23-25 — The University
Electronic Processes in Dielectric Liquids Conference
(Inst. of Physics & Phys. Soc., 47 Belgrave Square, London, S.W.1)

SCARBOROUGH

May 19-22 — Royal Hotel
R.T.R.A. Conference
(Radio & Television Retailers' Assoc., 19 Conway Street, London, W.1)

SOUTHAMPTON

April 16-20 — The University
Electronics & Industrial Productivity
(Brit.I.R.E., 9 Bedford Square, London, W.C.1)

OVERSEAS

Jan. 8-10 — Orlando
Millimeter and Submillimeter Conference
(I.R.E., 1 East 79 Street, New York 21)

Jan. 22-24 — San Francisco
Reliability and Quality Control
(R. Brewer, Hirst Research Centre, Wembley, Middlesex)

Jan. 30-Feb. 1 — Los Angeles
Military Electronics Convention
(I.R.E., 1 East 79 Street, New York 21)

Feb. 8-12 — Paris
International Electronic Components Exhibition
(F.N.I.E., 23 rue de Lübeck, Paris 16e)

Feb. 11-15 — Paris
Quantum Electronics Congress
(Société Francaise des Electroniciens et des Radio-électriciens, 10 avenue Pierre-Larousse, Malakoff)

Feb. 20-22 — Philadelphia
Solid-State Circuits Conference
(F. J. Witt, Bell Telephone Labs., Murray Hill, N.J.)

Mar. 20-23 — Los Angeles
Audio Convention
(Audio Eng'g. Society, Old Chelsea Station, New York 11)

Mar. 25-28 — New York
I.R.E. International Convention
(I.R.E., 1 East 79 Street, New York 21)

April 16-18 — New York
Optical Masers Symposium
(Polytechnic Institute, 55 Johnson Street, Brooklyn 1)

April 17-19 — Washington
Nonlinear Magnetics Conference
(I.R.E., 1 East 79 Street, New York 21)

May 7-9 — Washington
Electronic Components Conference
(J. E. Hickey, Chilton Co., 56 Street, Philadelphia 39.)

May 13-15 — Dayton
Aerospace Electronics Conference
(I.R.E., 1414 East 3 Street, Dayton, Ohio)

May 20-22 — Santa Monica
Microwave Theory and Techniques Symposium
(Dr. I. Kaufman, Space Technology Laboratories, 1 Space Park, Redondo Beach, Cal.)

May 20-24 — Melbourne
Radio and Electronic Engineering Convention
(I.R.E. Aust., 157 Gloucester Street, Sydney)

May 20-25 — Montreux
International TV Symposium & Exhibition
(J. H. Gayer, I.T.U., Place des Nations, Geneva)

June 4-5 — Philadelphia
Radio Frequency Interference Symposium
(I.R.E., 1 East 79 Street, New York 21)

June 11-13 — Los Angeles
Space Electronics & Telemetry
(J. R. Kauke, 1632 Euclid Street, Santa Monica)

June 19-21 — Minneapolis
Automatic Control Conference
(Prof. O. L. Updike, Department of Chemical Engineering, University of Virginia, Charlottesville, Virginia)

July 22-25 — Liège
Medical Electronics Conference
(Dr. A. Nightingale, St. Thomas's Hospital, London, S.E.1)

July 26-Aug. 10 — Sydney
Sydney Trade Fair
(Industrial & Trade Fairs, Commonwealth House, New Oxford Street, London, W.C.1)

Aug. 4-9 — Washington
Aerospace Support Systems Conference
(I.R.E., 1 East 79 Street, New York 21)

Aug. 20-23 — San Francisco
Western Electronics Show & Conference
(WESCON, 1435 La Cienega Blvd., Los Angeles)

Aug. 27-Sept. 4 — Basle
Automatic Control International Congress
(I.F.A.C. Congress, Kinkelstrasse 10, Postfach 289, Zurich)

Aug. 30-Sept. 8 — Berlin
German Radio Exhibition
(Berliner Ausstellungen, Charlottenburg 9, Berlin)

Sept. 2-7 — Basle
Industrial Electronics Exhibition
(Swiss Industries Fair, Postfach, Basle 21)

Sept. 15-19 — Tokyo
International Scientific Radio Union General Assembly
(U.R.S.I., 7 place Emile Danco, Uccle, Brussels)

Sept. 18-19 — East Lansing
Industrial Electronics Symposium
(L. J. Giacaletto, Michigan State Univ., E. Lansing, Mich.)

Sept. 30-Oct. 2 — Toronto
Canadian Electronics Conference
(I.R.E., 1819 Yonge Street, Toronto 7)

September — Paris
International Radio and TV Exhibition
(F.N.I.E., 23 rue de Lübeck, Paris 16e)

Oct. 7-9 — Utica
Communications Symposium
(I.R.E., 1 East 79 Street, New York 21)

Oct. 28-30 — Chicago
National Electronics Conference
(N.E.C., 228 N. La Salle Street, Chicago)

Nov. 10-14 — Atlantic City
Magnetism & Magnetic Materials Conference
(I.R.E., 1 East 79 Street, New York 21)

Nov. 12-14 — Los Angeles
Computer Conference
(I.R.E., 1 East 79 Street, New York 21)