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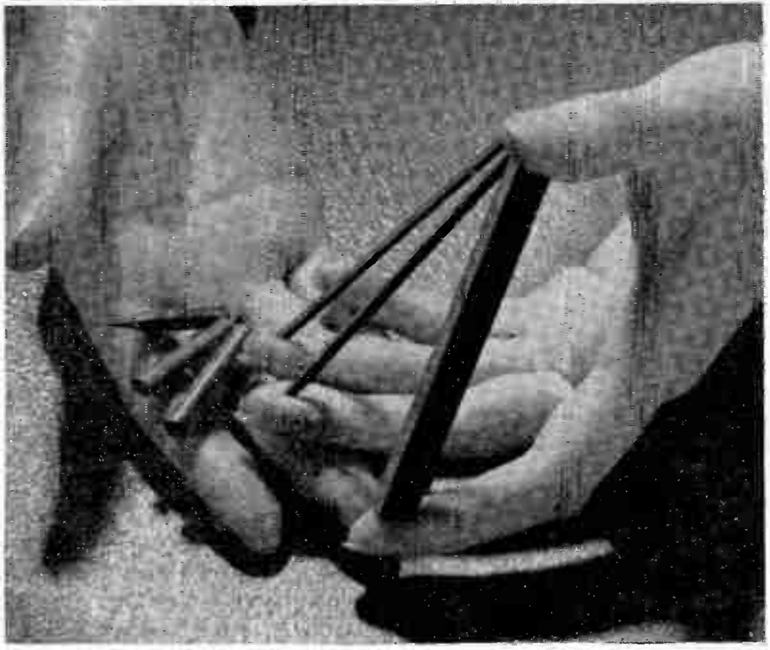

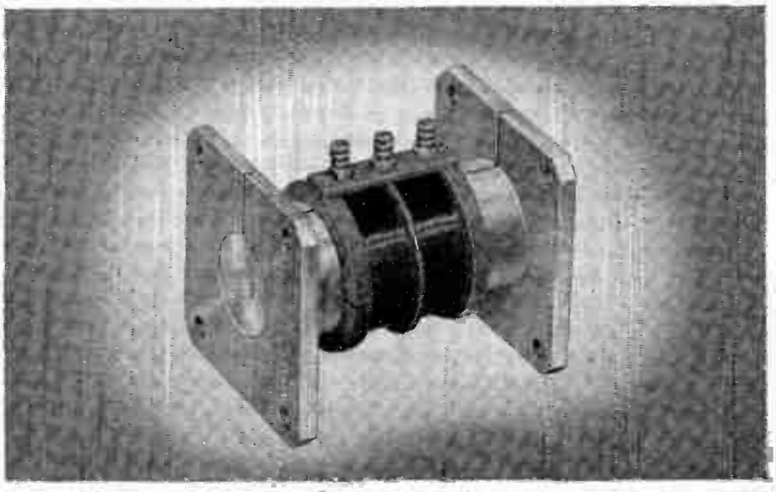

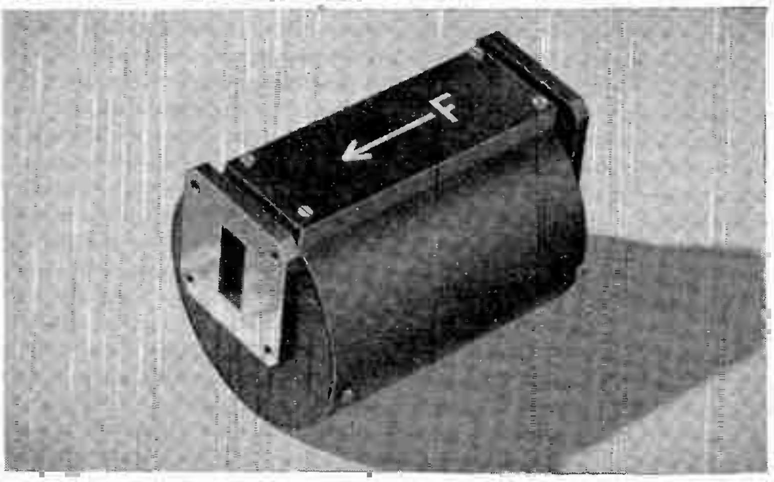

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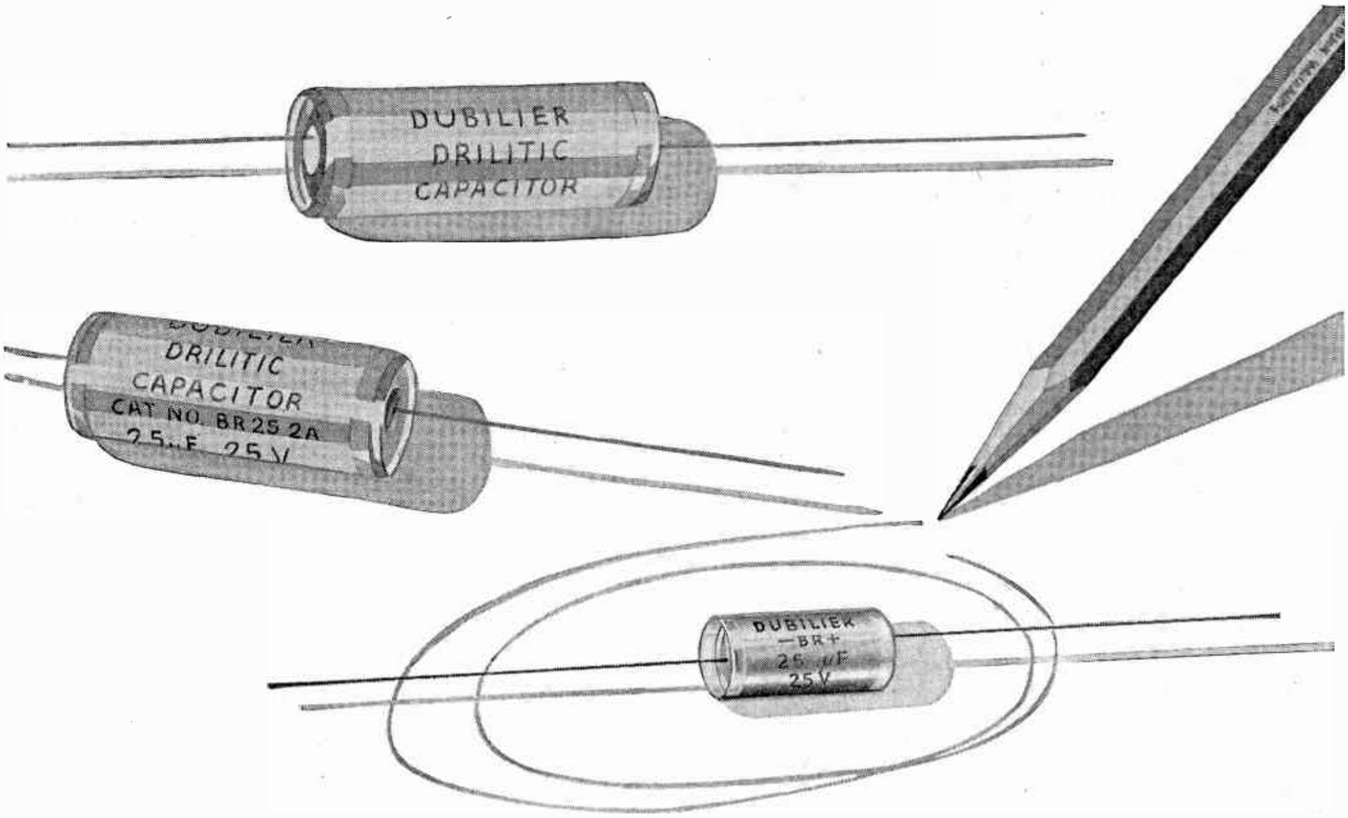

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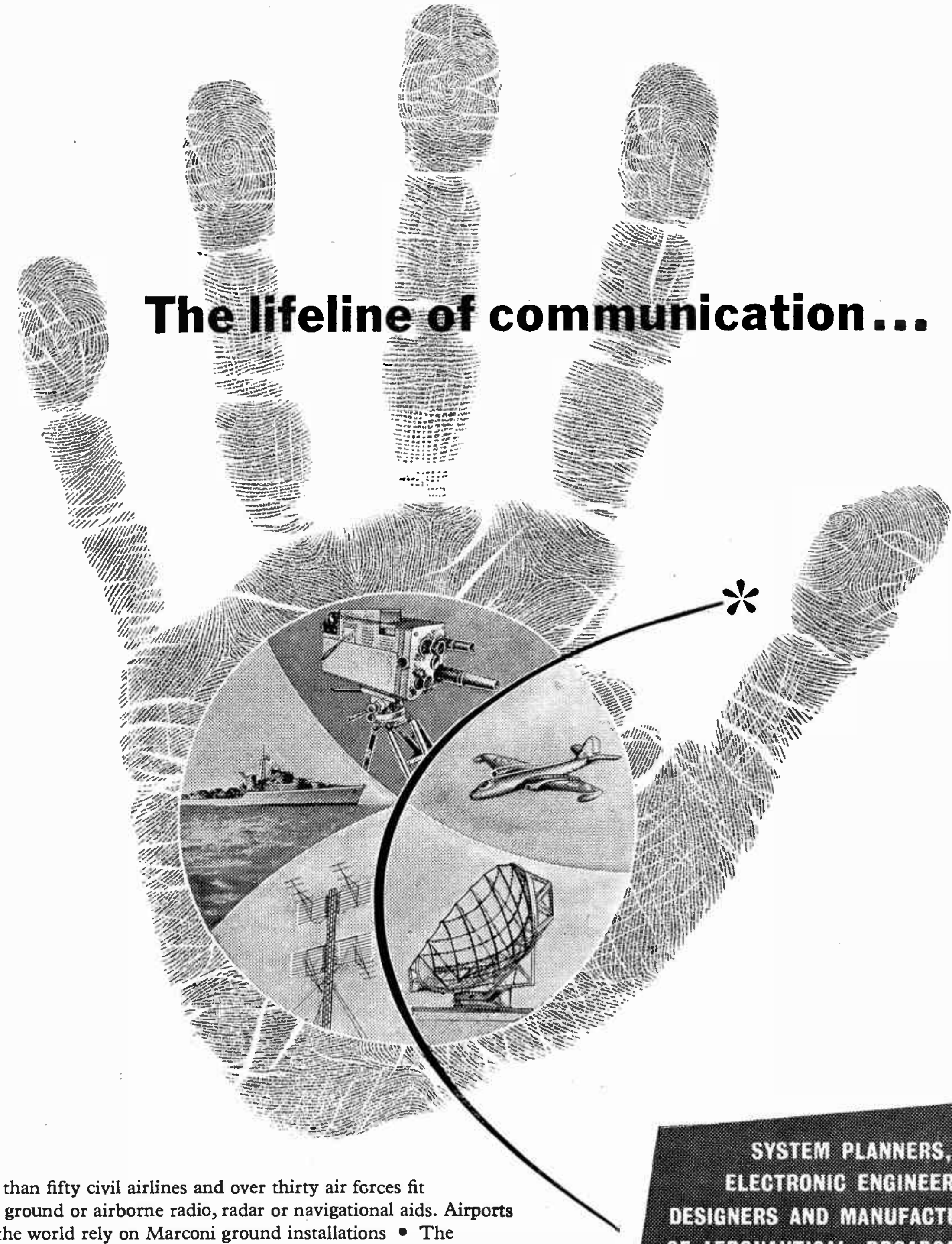

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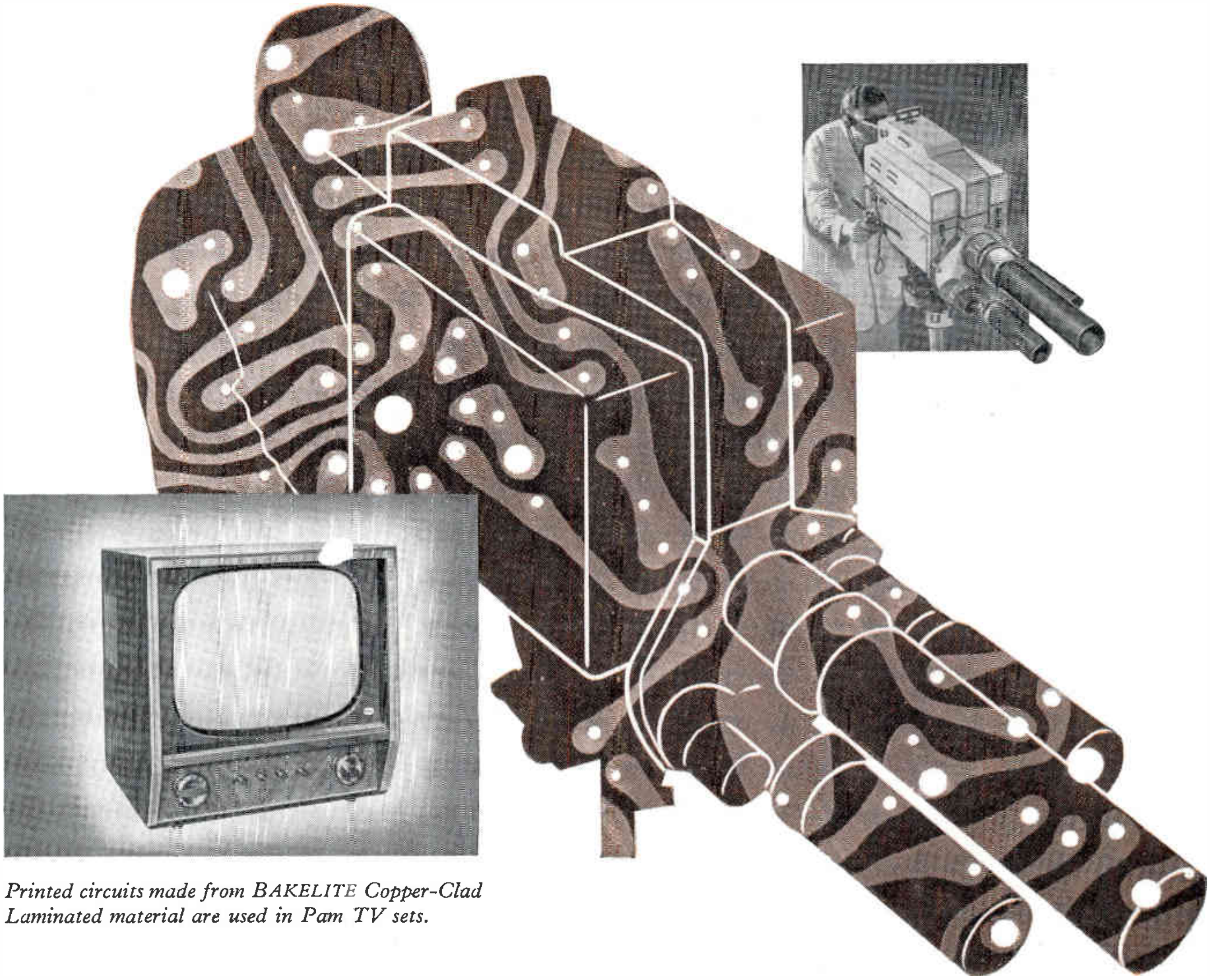

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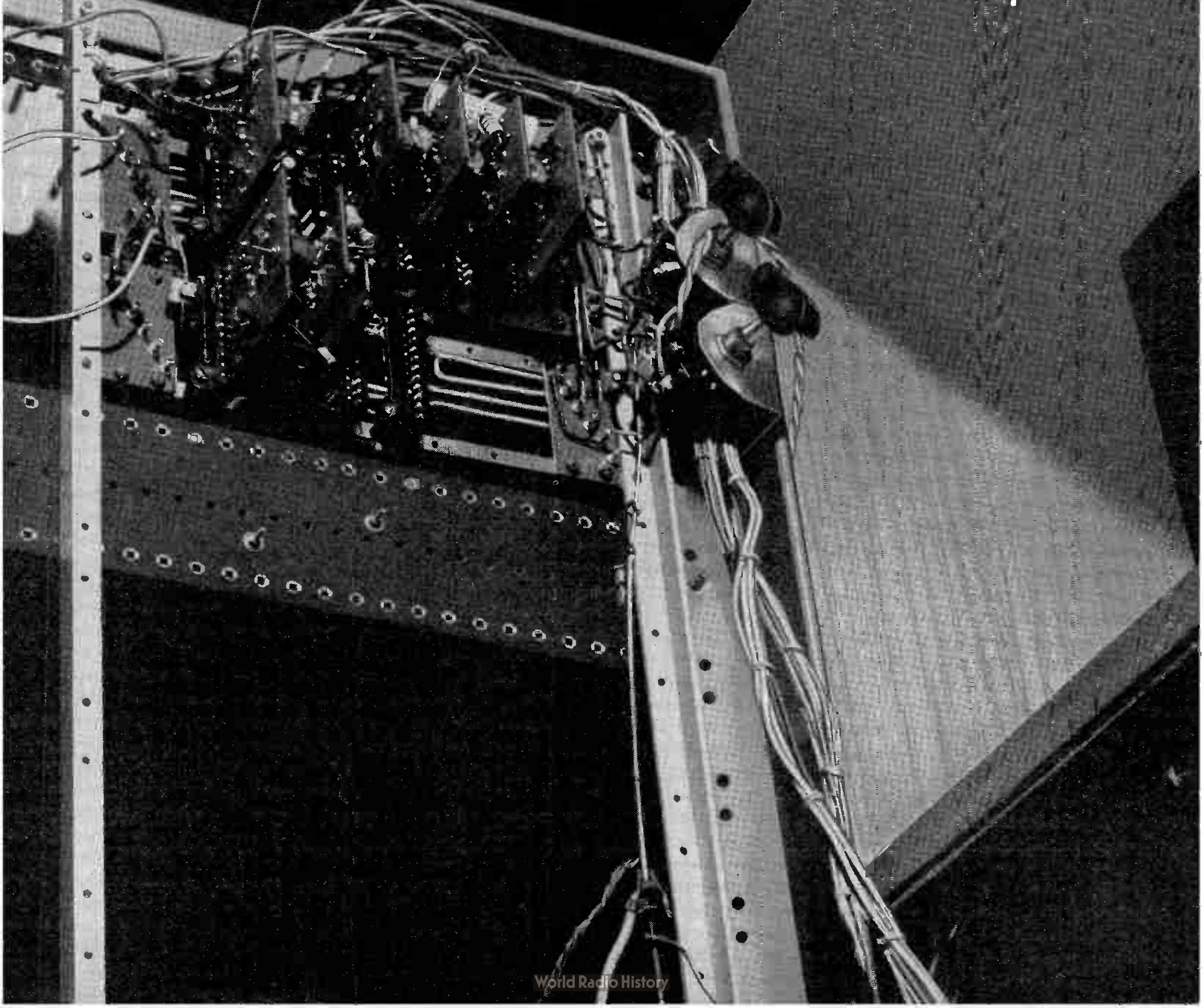

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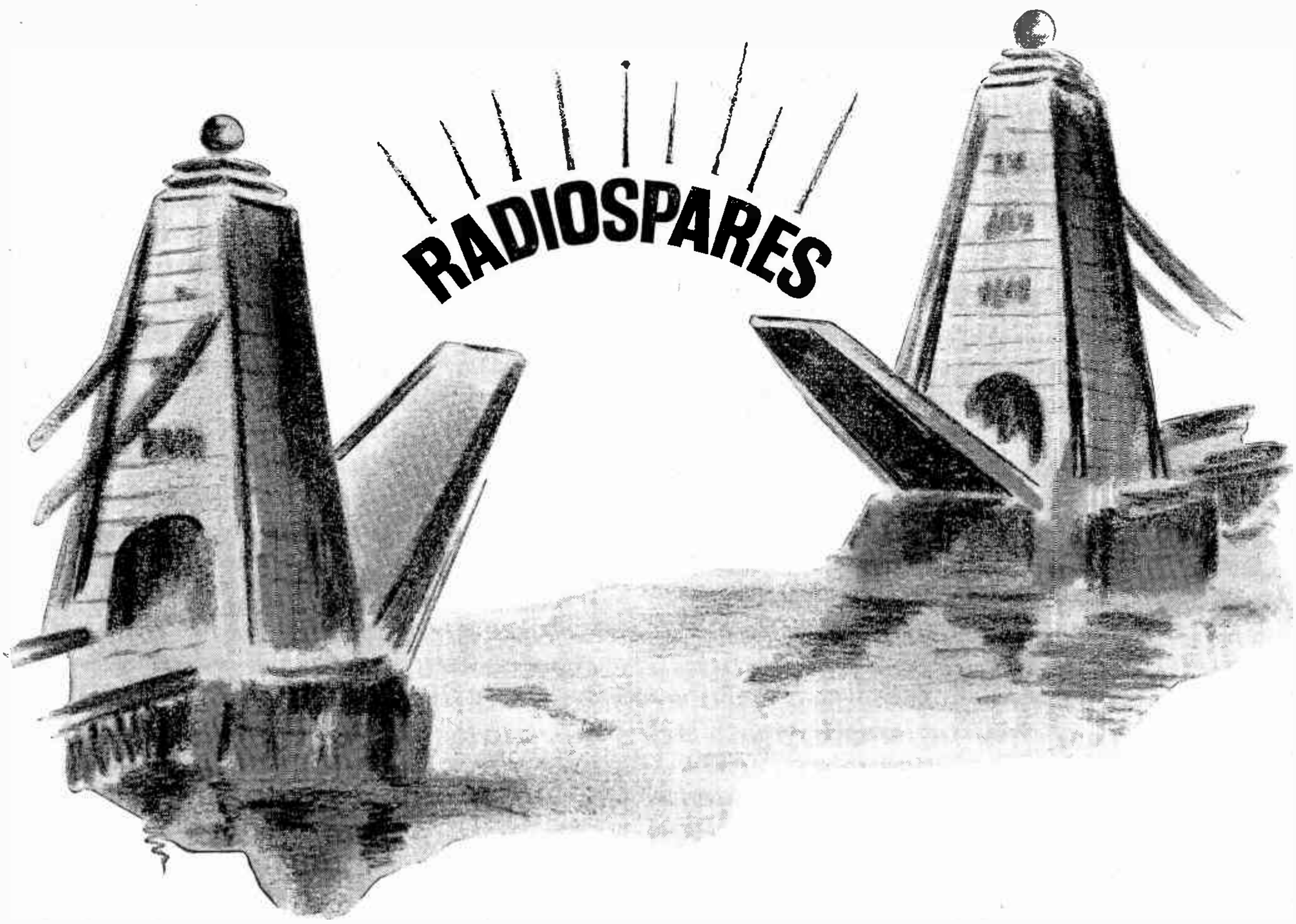

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VOLUME 35 NUMBER 1

JANUARY 1958

ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

Pulse Testing of A.F. Amplifiers

THE idea of using pulse waveforms for the testing of a.f. amplifiers is advocated from time to time, presumably to save the labour of extensive steady-state measurements. It is not always realized, however, that if such a test is to be of any value the shape of the pulse employed must be chosen carefully.

An a.f. amplifier is intended to handle only frequencies within the audible range and, in normal use, its input signal will not contain components of frequency outside that range. If the pulse used for testing has a spectrum extending beyond the audio range, an amplifier which passes only frequencies in that range will necessarily distort it. The distortion of the output pulse is then no indication of any fault in the amplifier; it may still be able to deal perfectly with any audio signal.

This may appear obvious, but we have come across cases where a pulse of rise-time 0·05 μsec has been used for testing an amplifier. Because the output waveform exhibited marked overshoot, it was said that the overshoot would tend to overload the output valves and so limit the maximum undistorted output.

We regard this criticism as quite invalid. A rise-time of 0·05 μsec demands frequencies up to at least 20 Mc/s for proper reproduction and it is unreasonable to expect any a.f. amplifier to deal faithfully with such a pulse. The criticism would be a valid one only if the overshoot occurred with a test pulse which had a spectrum confined to the audio range.

We referred last month to the use of pulse tests in television and an article in that issue discussed the kind of pulse needed. It was shown to be a sine-squared pulse with a duration at half amplitude equal to the reciprocal of the cut-off frequency of the communication channel.

Applied for a.f.-amplifier testing, this would call for a sine-squared pulse of half-amplitude duration 50 μsec, for this would have a spectrum up to 20 kc/s only. A first-class a.f. amplifier should pass such a pulse without distortion, and any distortion at all would have some significance. It is, however, difficult to correlate pulse distortion with the audible effects of distortion and we are somewhat doubtful about the usefulness of such tests on a.f. amplifiers. The television case is different, for the test waveform is of the same nature as a television signal and it is easy to relate pulse distortion to picture distortion.

New Types of D.C. Amplifier—Part 1

THE CASCADE-BALANCE SYSTEM

By D. J. R. Martin, B.Sc., A.Inst.P.*

SUMMARY. *In the amplifier described, the zero variations in the first stage due to supply-voltage and temperature fluctuations are balanced against those in the second stage, instead of the variations being balanced between valves in a stage as is usual. An essential feature of the system is the application of the Owen-Prinz method of drift correction to the first stage, which is duplicated to preserve bandwidth.*

The outstanding advantage of the system over the parallel-balance type of amplifier is the relative reduction in zero variations (including those due to valves ageing) by a factor of well over 100, *rendering stabilized heater-supplies redundant. Incidental merits of the amplifier are a very high input resistance and a frequency response level to* 25 *kc/s.*†

The principal limitation on the performance of conventional direct-coupled amplifiers is in the matter of zero variations or 'drift'. This usually arises from several causes, the chief among these being supply-voltage variations and changes in amplifier components—especially valves—due to temperature fluctuations and ageing. The effects of supply-voltage variations may be reduced by the use of stabilized supplies, while the effects of temperature changes may be minimized by employing certain compensation techniques within the amplifier. Further improvement in both respects is obtained by using pairs of valves in balanced circuits.

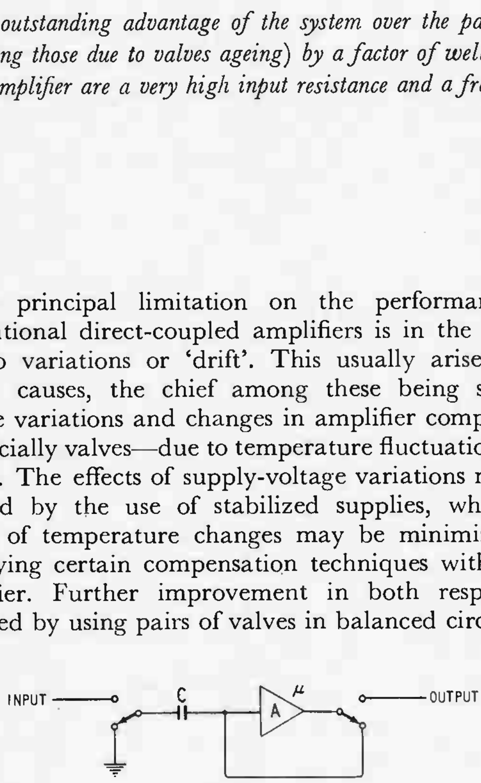


Fig. 1. *The Owen-Prinz method of drift correction*

An alternative approach to the problem of drift lies in the use of automatic drift-correction methods. These involve circuits which sense any zero variations in the amplifier and correct them. In the past, two such methods have been described. The first of these employs a comparatively simple principle and needs few additional components. It suffers from the disadvantage that the amplifier must be periodically removed from service while the drift is checked and corrected; the bandwidth of the amplifier is therefore reduced and depends on the frequency with which the correcting operation can be carried out. The method seems to have been independently devised by C. E. Owen[1] and D. G. Prinz[2]; it is accordingly termed the Owen-Prinz method in these articles.

In the second, and more common, type of drift-corrected amplifier the signal channel is not interrupted at any time, and the bandwidth of the amplifier is not affected. To achieve this, however, a subsidiary amplifier has to be employed in the correcting operation. This additional amplifier must itself be drift-free, though it need not respond to frequencies above a few cycles per second; in practice, a contact-converter type of amplifier has hitherto been the inevitable choice in this position. This method of drift correction was also invented by Owen[1] and has been described by many later workers (see, for example, references 3–6).

The two amplifiers described in these articles make use of the simpler Owen-Prinz method in ways which avoid its bandwidth limitation and lead to novel balancing features. The first is intended for single-sided input signals; the second has a differential input circuit with a high rejection factor against in-phase components. Both amplifiers are generally superior in performance to a conventional parallel-balance amplifier, and their long-term stability, in particular, is over a hundred times better. They do not require stabilized heater supplies.

The Owen-Prinz Method of Drift Correction

For a full analysis of this method, reference may be made to the original publications[1, 2]; but for present purposes the following brief description will be found adequate.

A block diagram of the basic circuit is shown in Fig. 1, in which A represents a direct-coupled amplifier of gain μ. It is essential to the operation of the principle that the amplifier to be drift-corrected causes a net phase-reversal of the signal; in other words, if normal methods of coupling are used, the amplifier must contain an odd number of stages. Another requirement of the amplifier is that the available output-voltage swing shall embrace zero; this will usually require potentiometer coupling to the output circuit, using a negative d.c. supply.

* *Formerly British Scientific Instrument Research Association, Chislehurst; now Mining Research Establishment.*
† *Patent Application No. 28573/56 covers the system described.*

With the switches in the position shown, the input and output terminals of the amplifier itself are directly connected, and the input and output potentials are therefore the same. Assuming a linear amplifier, there is only one potential that satisfies this condition and, since the negative feedback connection is inherently stable, the amplifier settles at it. The capacitor *C* charges to this value, which may be termed the error potential since it approximates very closely to the potential required to give zero output. If the amplifier should suffer any disturbance tending to cause a change in output indication, the 100% negative feedback connection reduces the actual effect by a factor $(\mu + 1)$; this constitutes the drift-correcting property of the circuit.

Operating the switches puts the amplifier into service. The capacitor *C*, charged to the error potential, is interposed in series with the signal to provide zero correction. Should any disturbance occur in the amplifier, causing a false change in output indication, correction will automatically be applied when the switches are next moved to the 'correcting' position, resulting in a modified charge on the capacitor. To maintain the correct charge against leakage and changing requirements frequent operation of the switches is necessary. In fact, the process is best made automatic, by some such means as a change-over relay operated by a multivibrator.

The Owen-Prinz system as it stands seems to have found little use in practice. The main objection to it is, of course, the fact that the signal channel is interrupted while the correction is taking place.

The Cascade-Balance Principle

The new systems utilize the Owen-Prinz method of drift correction, making use of a fundamental property of it which has not previously had a precise significance; this is the fact, already mentioned, that the drift is reduced by a factor almost exactly equal to the gain of the amplifier to which the method is applied.

In its basic form the cascade-balance amplifier comprises two identical single-sided stages connected in cascade in the normal fashion. In the absence of correction, the effects of supply-voltage variations and temperature changes would be approximately the same in each stage; but referred to a fixed point in the amplifier, such as the input terminal, variations in the first stage would predominate by a factor equal to the stage gain, since they suffer a further stage of amplification. However, the Owen-Prinz method is applied to the first stage only. The zero variations in that stage are thus reduced by a factor equal to the stage gain; in their ultimate effect, therefore, they are now approximately equal to the variations in the second stage. Because of the phase reversal in each stage, they are also opposite in sign; they therefore tend to cancel out and, in practice, may be made to do so almost exactly by the use of balancing techniques such as are normally applied to parallel-balance amplifiers.

Since the drift reduction is confined to one stage, the disadvantage of the periodic interruption of the signal channel may be overcome simply by duplicating the first stage, it being arranged that the two counterparts are taken out of service for correction alternately.

A block diagram of the system is shown in Fig. 2, where A, B and C are the identical stages. The automatic switching must be so arranged that neither of stages A and B is taken out of service for correction until the other is restored to service; in other words, there must be a short 'overlap' period in which both stages are operating in parallel.

An amplifier designed on this principle may easily have an inherent zero-stability over a hundred times better than that of a conventional parallel-balance amplifier, and the single-sided nature of the circuit enables negative feedback to be applied. The practical design of such an amplifier is dealt with in the succeeding discussion.

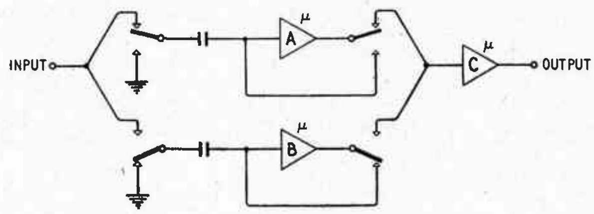


Fig. 2. *Block diagram of the cascade-balance system*

The Basic Stage

Apart from incidental differences in detail, the output stage and the twin input stages must be identical if the full advantages of the cascade-balance system are to be realized. The design of the basic stage is therefore governed by a combination of requirements, which may be listed as follows:

(1) the stage gain must be high—preferably over 100—as this is a direct measure of the drift reduction;

(2) a low value of grid current is necessary in the first stage to avoid significant changes in capacitor potential between correcting operations;

(3) potentiometer coupling and a negative supply are necessary to obtain an output zero-level near earth potential;

(4) an adequate output-voltage swing must be available from the second stage without overloading.

The circuit adopted is shown in Fig. 3. The first two requirements are met by operating a pentode under low-current conditions, and grid current is further reduced by under-running the heater of this valve. For the second-stage position some of the component values are modified slightly, as shown by the references in brackets; this increases the available output-voltage swing without spoiling the inter-stage balancing properties.

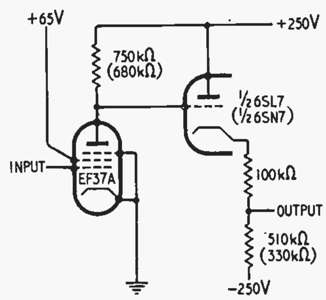


Fig. 3. *Circuit of basic stage*

Negative Feedback

In such an amplifier as this, negative feedback is a practical necessity. Apart from its usual benefits, it enables the gain of the amplifier to be controlled without disturbing the cascade-balance conditions.

Overall feedback offers the greatest advantage, and may be applied either direct to the input terminal (parallel feedback) or to the cathode circuit of the first stage (series feedback). The former method is usually adopted in some special cases such as analogue-computer applications where a current-summation circuit is required; it may be used here without complication except that a phase-reversing stage will need to be introduced and precautions taken to preserve feedback stability.

The preferred method for most applications is to introduce the feedback voltage across a small resistor inserted in the cathode return circuit of the twin input stages. The position is somewhat complicated by the fact that while each stage is in the 'correcting' phase the feedback to it must be removed or nullified or it will interfere with the correcting operation. An exact analysis shows that the effect of the interference is to superimpose on the correction potential a voltage proportional to the feedback and slightly larger in magnitude. It may be counteracted by returning the correcting capacitors to a suitable compensating potential during the correcting phase instead of to earth. The arrangement is illustrated in Fig. 4.

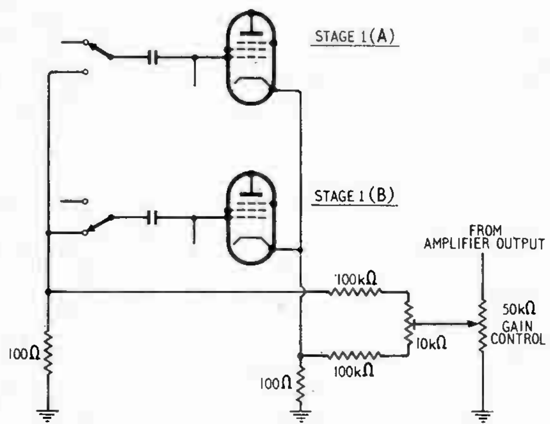


Fig. 4. *Method of applying negative feedback without disturbing drift-correction process*

Adjustment of Gain and Zero Level

The overall gain of the amplifier must be controlled by negative feedback, as already stated. In addition, some means of matching the individual gains of the first stages is required, preferably without disturbing the zero setting of the amplifier.

Over the small range which is adequate such a control may be effected in the screen circuit of a stage, by variation of the resistance presented to the screen. The method is illustrated in Fig. 5. R_z passes the screen current and is arranged to drop 20 volts or so. R_x and R_y are relatively low in value and their ratio chosen so that the potential drops across R_x and R_z are approximately equal. The gain may then be controlled by R_v without seriously affecting the standing potential on the screen.

Control of zero level also may be conveniently effected in the screen circuit. As well as the main zero control, a differential adjustment is necessary to ensure that the same zero level is maintained while each of the input stages is operating.

The Automatic Switching

Switching Frequency

The choice of switching frequency is governed by entirely different considerations from those prevailing with contact-converter types of amplifier, where frequencies below 50 c/s are rarely used. The minimum allowable switching frequency in the present case is decided by the stability requirements together with two other factors: these are the maximum rate of drift likely to be encountered and the rate of leakage of the charge on the capacitors providing the correcting potential.

Assuming that the h.t. supplies can if necessary be stabilized, or at least provided with extremely large smoothing constants, the quickest fluctuations likely to need correction are those arising from heater-voltage changes; the ultimate lower limit to the switching frequency is set by the maximum step-change in mains voltage to be allowed for and the thermal lags in the valves. In practice, about five complete switching cycles per second is sufficient to maintain a zero stability of 100 μV against instantaneous heater-voltage changes of 0·5%.

The error introduced by the leakage of charge from the correcting capacitors is a function of the capacitance employed and the input current of the valves as well as the switching frequency. At five switching cycles per second, the effect may be kept well within 100 μV by using capacitors of 2–8 μF without selecting valves for low grid current.

Contact Sequences

The switching operations fall naturally into two groups, associated respectively with the twin input stages. A prior requirement of the system is that at least one of the input stages shall be in operation at any given time; the switching must therefore be so arranged that none of the contact actions in the one group changes to the 'correcting' condition until all the contact actions of the other group have adopted the 'operating' condition.

The sequence of contact actions within each group itself is also important if errors and switching transients are to be avoided; the requirements will be apparent from a consideration of Fig. 2. The change-over contacts must be of the normal 'break-before-make' type, as indicated; it is obviously undesirable for the input signal to be momentarily earthed or the input connections of successive stages to be bridged. The connection of the output of the first stage to the input of the second stage must be the last to be made and the first to be broken in the change-over operation; in other words, the output of the first stage must not be passed on unless the signal is applied to the stage. The remaining requirement is that the signal connection shall not be made until the grid-to-output connection has been broken; this is to prevent the charge on the capacitors from being

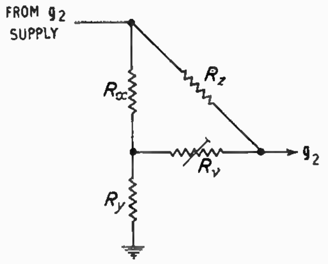


Fig. 5. *Screen network for adjustment of stage gain*

disturbed during the change-over process, and also to avoid momentary shunting of the signal by the output impedance of the stage. It will be noted that the complete switching cycle is reversible.

Switching Methods

At the relatively low switching frequencies which are adequate, the use of standard Post Office relays is possible and is probably the most convenient means to adopt. The overlap requirement of the two contact groups is then simply met by using two relays; one is allocated to each of the input stages and the overlap obtained by suitable adjustment of the release lags. The requirements of contact sequence within each group may be met by suitably setting the fixed contact springs. To obtain the necessary alternation of energization, additional contacts and electrical interlocking of the relays may be used, in a self-running mode of operation. Otherwise the relays may be driven from a multivibrator.

As an alternative to relays, cam-operated contacts may be used; their main advantage would be ease of sequence adjustment.

The Complete Amplifier

Fig. 6 shows the circuit of a typical cascade-balance amplifier, excluding details of h.t. feed arrangements. The dotted boundary defines the duplicated portion; stage A only is shown and the connections to stage B indicated, references in brackets applying to stage B.

Capacitors C_3 and C_4 are for high-frequency compensation; if maximum bandwidth is not required they may be omitted but, in that case, it may be necessary to shunt R_{12} by a 50-pF capacitor to ensure feedback stability at low gain settings.

A connection has been brought out from the feedback chain and marked 'guard potential'. This potential follows the signal voltage and is available for the connection of screens and metal objects when input capacitance is to be minimized.

Resistors R_3 and R_4 have been included to prevent the grids from floating momentarily during the change-over operation; in their absence, a resulting large switching transient in the stage out of service might cause interference in the signal channel through stray couplings. Their loading on the input circuit is negligible since they are returned to a potential which closely follows the signal; the actual effect is in fact negative and if there is any risk of the input being open-circuited the connection of these resistors should be tapped a little way (about 2%) down R_9 to reduce the loop gain below unity and so safeguard stability. The externally-available guard potential cannot be used for this connection since it includes the standing voltage drop across R_{10}.

It should be understood that although the grid of the first stage has no leak resistor while in service it must not be regarded as a floating electrode; for the size of the capacitor ensures that the grid potential is effectively tied during the short time between correcting operations, and the capacitor may be considered as a short-circuit to anything but the correcting potential.

With the specified value of 8 μF for capacitors C_1 and C_2, leakage effects during the operating phase are not discernible. A considerable reduction in capacitance is possible before the resulting sawtooth reaches the quoted 100-μV noise maximum, but the valves then need a longer warming-up period before grid current falls low enough. The capacitors themselves should be metal-cased and the casings connected to the guard potential; reliable paper-dielectric smoothing capacitors have suitably low leakage tendencies.

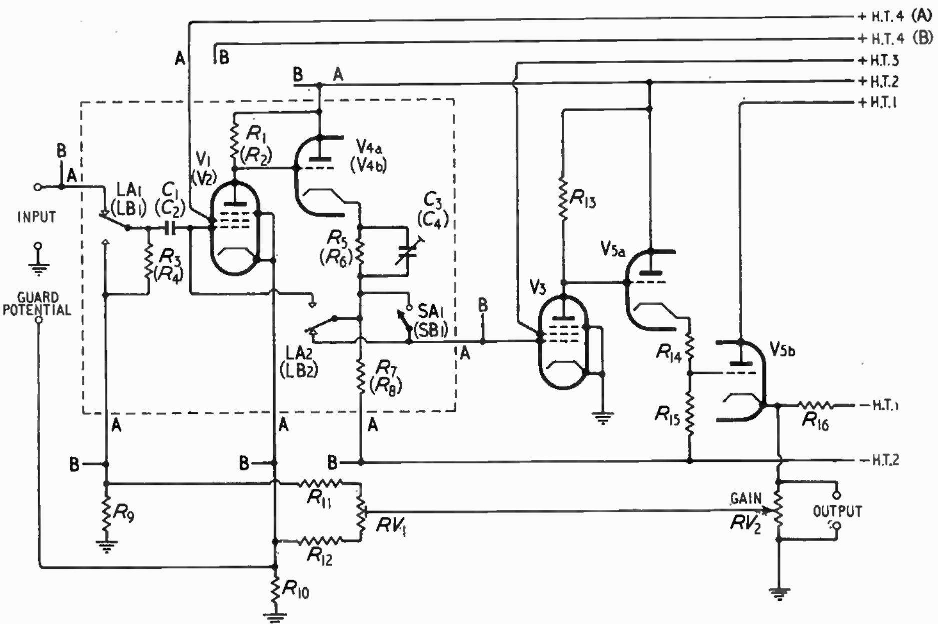


Fig. 6. *Circuit of cascade-balance amplifier (Note: relay contacts are shown in operated condition).*
R_1, R_2 = 750 $k\Omega$ (2%) *high-stability carbon;* R_3, R_4 = 2·2 $M\Omega$; R_5, R_6 = 100 $k\Omega$ (5%); R_7, R_8 = 510 $k\Omega$ (5%); R_9, R_{10} = 100 Ω (2%) *wirewound;* R_{11}, R_{12} = 91 $k\Omega$ (2%) *wirewound;* R_{13} = 680 $k\Omega$ (2%) *high-stability carbon;* R_{14} = 100 $k\Omega$ (5%); R_{15} = 330 $k\Omega$ (5%); R_{16} = 50 $k\Omega$ (2 *W*); RV_1 = 10 $k\Omega$ *wirewound;* RV_2 = 50 $k\Omega$ *wirewound;* C_1, C_2 = 8 μF *paper;* C_3, C_4 = 75–125 *pF; SA, SB = 2-circuit change-over toggle switch (see also Fig. 7);* V_1, V_2, V_3 = *EF37A;* V_4 = *6SL7;* V_5 = *6SN7*

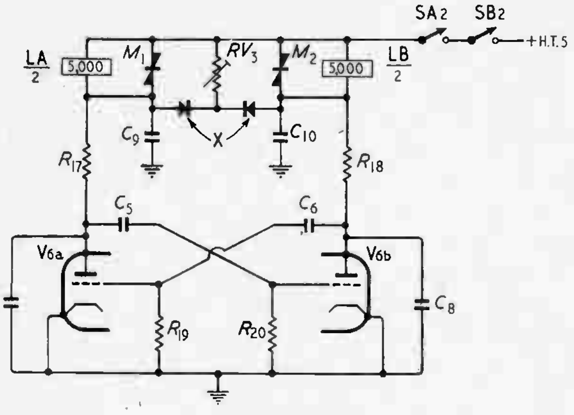


Fig. 7. *Circuit of multivibrator for driving relays;* R_{17}, R_{18} = 33 $k\Omega$ (1 *W*)*;* R_{19}, R_{20} = 3·3 $M\Omega$*;* RV_3 = 50 $k\Omega$*;* C_5, C_6 = 0·02 μF*;* C_7, C_8 = 0·02 μF*;* C_9, C_{10} = 0·25 μF*;* M_1, M_2 = *Metrosil disc, 1-in diam.* (*annular*), 25 *mA at* 50 *V* (*Metropolitan-Vickers*)*; X = full-wave rectifier,* 50 *V*, 10 *mA; SA, SB* (*see Fig. 6*)*; LA, LB = relay, type* 3,000*;* 5,000-Ω *coil, platinum contacts, 2-pole change-over;* V_6 = *6SN7*

In Fig. 7 is shown the circuit of the multivibrator which is used to drive the relays. The variable resistor RV_3 provides a control of the release lag of the relays and thus enables the overlap in the operation of the input stages to be adjusted. The Metrosil discs M_1 and M_2 reduce the effects of h.t. changes on this overlap time, and would be omitted if the multivibrator h.t. line were stabilized. The oscillation frequency is about 5 c/s.

The multivibrator may be switched off by either of the switches SA and SB. Both input stages then rest in the correcting condition, and the grid of the second stage is bypassed to one or the other of these according to which switch is operated. In this state both signal and feedback are ineffective, though the cascade-balance principle operates; preliminary checks on zero-level and stability, and some of the final adjustments, may be simply made in this condition of the amplifier.

The relays used are of the standard Post Office type. As already explained, the contact springs may need some adjustment to ensure the correct switching sequence. These checks are best carried out before the relays are installed, and it is also advisable at this stage to see that the relays are as closely matched as possible in spring tensions.

In the layout of the amplifier it is advisable to keep the input circuits away from the multivibrator; otherwise, interference may arise from the steep-fronted waveform. The field of the relay coils is comparatively harmless, as the high-frequency currents are bypassed through capacitors C_7, C_8, C_9 and C_{10}, which should be grouped with the other multivibrator components. It may be found necessary to screen the multivibrator valve and possibly its associated circuits as well. With certain layouts it may also be necessary to screen the output valve (V_5) to preserve stability at high gain settings.

Power Supplies

The effects of all normal changes in heater supplies and of slow changes in h.t. supplies are subject to a twofold reduction: the Owen-Prinz method of drift correction divides them by a factor of 100 or so; and the cascade-balance feature reduces them by a further similar factor. A sudden change in h.t. voltage, however, could momentarily escape correction, resulting in an unacceptable transient in the amplifer output. It is quite practicable to prevent such occurrences by using extremely large time constants (of the order of 10 seconds) in the h.t. smoothing circuits, and electronic 'slugging' circuits have been devised for the purpose; but it is probably more straightforward to use conventional stabilized supplies. In any case, it is recommended that the screen supplies be stabilized, this function being served very simply by a single neon tube.

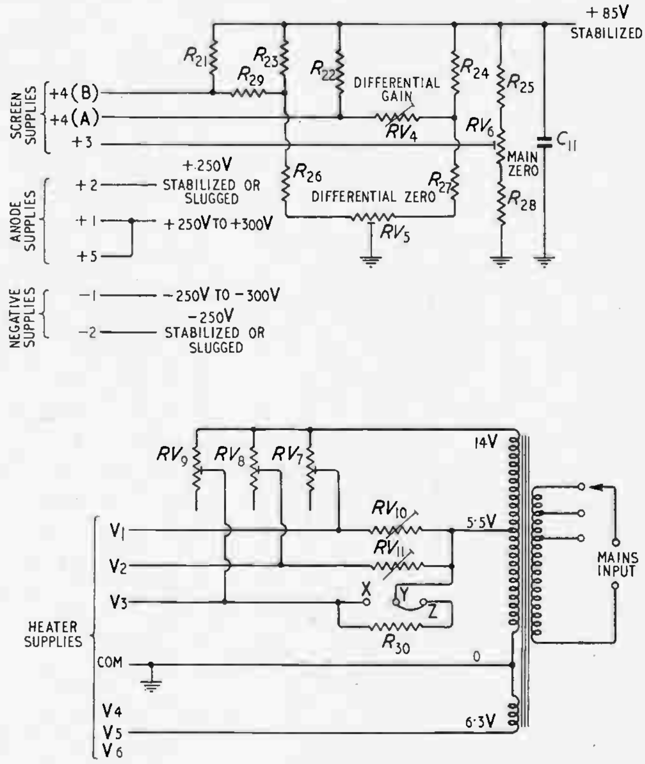
Stabilization of the heater supply will not normally be necessary, though steps may need to be taken to obtain an improved match between stages for the effects of supply-voltage changes; the method has been described in a separate article[7].

The recommended supply arrangements are shown in Fig. 8, which also includes the gain and zero adjustments already mentioned.

Tests and Adjustments

Before the amplifier is switched on for the first time, variable resistors RV_3, RV_7, RV_8, RV_9, RV_{10} and RV_{11} should be set to their maximum values; the other controls should be set half-way. Then, with the heaters only switched on, resistors RV_7, RV_8 and RV_9 should be adjusted for nulls in the voltages across RV_{10}, RV_{11} and R_{30} respectively.

Fig. 8. *H.t. and heater supply arrangements;* R_{21}, R_{22} = 220 $k\Omega$ (5%) *high-stability carbon;* R_{23}, R_{24}, R_{25} = 10 $k\Omega$ (5%) *wirewound;* R_{26}, R_{27}, R_{28} = 33 $k\Omega$ (5%) *wirewound;* R_{29} = 68 $k\Omega$*;* R_{30} = 27 Ω *wirewound;* RV_4 = 250 $k\Omega$*;* RV_5 = 10 $k\Omega$ *wirewound;* RV_6 = 5 $k\Omega$ *wirewound;* RV_7, RV_8, RV_9 = 50 Ω (*or suitable fixed value*) *wirewound;* RV_{10}, RV_{11} = 100 Ω *wirewound;* C_{11} = 0·5 μF



With the multivibrator switched off by means of SA or SB, the h.t. supplies may then be applied and a 0–50-V moving-coil d.c. meter connected between the output terminals. Provided that RV_5 is centrally set, it should be possible to adjust the output voltage to zero by RV_6. The second multivibrator switch should now also be operated. If there is a change in output reading, RV_5 should be adjusted in conjunction with RV_6 until the zero adjustment holds regardless of whether SA or SB (or both) are operated.

At this stage the response to a.c. signals may be checked; for this purpose temporary grid leaks of about 2·2 MΩ are connected from the grids of stage 1 to the grid of stage 2. A low-frequency signal applied to the input terminals is then routed through each of the input stages in turn by operating the relevant multivibrator switch and manually closing the corresponding relay. The output is observed on an oscilloscope and the gains of the two channels are equalized by adjustment of RV_4. The gains may then be matched at a much higher frequency (say 20 kc/s) by means of C_3 and C_4. In this condition of the amplifier, RV_1 may be set if a pair of headphones is available; these are interposed in series with one of the correcting capacitors while a 1,000-c/s signal is passed through on the other channel with the gain control turned to minimum (maximum feedback), and RV_1 adjusted for minimum sound.

The adjustments for stability against heater-supply variations are fairly straightforward and not at all critical. The leaks used in the preceding adjustments are no longer required; with either SA or SB operated, small changes in heater-supply voltage are made by means of the primary taps of the mains transformer and the effects on the output indication noted. If a rise in supply voltage is followed by a rise in (positive) output voltage, then the portion of RV_{10} or RV_{11} (as the case may be) in circuit should be reduced a little and the test repeated; vice versa for a contrary effect. When the optimum adjustment has been established the other input stage may be compensated in the same way. If it is found that RV_{10} or RV_{11} is apparently too small in maximum value, then the link Y–Z should be transferred to X–Y and the balancing operations begun again. On the other hand, if a negative value of resistance seems to be required in RV_{10} or RV_{11} then the link should be removed altogether. It is possible by adjustment of RV_7, RV_8 and RV_9 to equalize the thermal lags of the valves[7] but this will not normally be necessary.

With the input terminals of the amplifier shorted, the multivibrator should now be switched on by restoring both SA and SB to normal, and the output of the amplifier observed on an oscilloscope with a slow time base. If violent transients at twice the switching frequency are seen to occur, their probable cause is a momentary interruption of the signal channel as the input stages interchange. To remedy this, RV_3 should be reduced until a definite overlap is obtained; this adjustment should be made with the h.t. supply to the multivibrator at its lower limit. Transients may also result from incorrect contact sequences in the relays.

The amplifier is now in a condition to pass a.c. or d.c. signals applied to the input. If the gain control is set to minimum a gain of just under 1,000 will be obtained; the maximum is about 40,000 and will not normally find practical use. With a 0–10-V moving-coil d.c. meter connected to the output, the zero (RV_6) and differential zero (RV_5) adjustments may be checked again and also the differential gain adjustment (RV_4).

A final check should be made on the adjustment of RV_1. To do this, a fairly large 50-c/s signal is applied to the amplifier and the output watched closely on a d.c. meter. If RV_1 is incorrectly set, small random fluctuations of the output indication will occur, coinciding with the switching operations. Another method involves using RV_1 to match accurately the respective gains measured on d.c. and on a.c. at about 1,000 c/s.

Whenever the amplifier is switched on from cold it should be allowed to idle with the multivibrator inoperative for about ten minutes while the input valves settle down to a condition of low grid current.

Performance

The characteristics and performance of the typical cascade-balance amplifier described may be summarized as follows:

Gain: approximately 900 to 40,000 (continuously variable).

Stability of zero-level: better than 100 μV referred to the input for slow mains-voltage variations of $\pm$ 10% and step changes of 0·5%, and all normal temperature and long-term effects.

Input grid current: less than 10^{-9} A.

Required warming-up time: 10 minutes.

**Stability of gain:* 1%.

**Frequency response:* level within $\pm$ 1% over the range 0 c/s to 25 kc/s.

**Output impedance:* about 20 Ω.

**Maximum undistorted output-voltage swing:* — 20 to + 40 volts.

**at a gain setting of* 1,000.

Further Development

As it stands, the cascade-balance system provides only for a single-sided input signal, whereas for many applications a differential-input circuit is desirable. This deficiency could be remedied by suitably redesigning the basic stage circuit to accept a differential signal. However, the incorporation of differential-input facilities opens up entirely fresh possibilities in the basic nature of the system and enables some minor disadvantages of the present amplifier to be eliminated. A completely revised system is therefore preferable and this will be described in the concluding part of this article.

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(*To be continued*)

A Feedback Circuit Equivalence

RECONCILING DIFFERENT POINTS OF VIEW

By A. W. Keen, M.I.R.E., A.M.I.E.E.*

SUMMARY. *Application of shunt feedback over a bootstrap-type amplifier yields a circuit configuration which is essentially identical with that of the conventional (i.e., grid input-anode output) shunt feedback amplifier. Despite this similarity, the former circuit and its variants are often regarded as cases of positive feedback whereas, in the latter the feedback is invariably considered to be negative. This paper seeks to resolve this apparent inconsistency and show that both points of view may be justified. A notable feature is the use of transfer network representations and their transformations to illustrate and confirm the equivalence.*

There is repeated evidence in the literature of apparent inconsistency in the description of the kind of feedback (i.e., whether negative, positive, or both) occurring in the class of circuits typified by the bootstrap integrator. Thus, O. S. Puckle[1] divides circuits developed for the 'linearization of electrostatic deflection potentials by means of feedback' into two groups according as the feedback is positive or negative, and includes the bootstrap integrator among the latter. F. C. Williams, on the other hand, in a well-known paper,[2] places this circuit in a section entitled 'Amplifiers With Positive Feedback', stating that the circuit 'is sometimes regarded as a case of positive feedback'. In the same paper, and in a more recent book,[3] he shows that the basic circuit of the bootstrap is, except for a trivial shift of the earth connection from the cathode to the anode side of the valve, identical with that of A. D. Blumlein's Miller-type integrator, which is invariably considered to be an application of shunt-type negative feedback. Despite this known relationship, V. W. Hughes and R. M. Walker, in the same book,[4] classify the bootstrap integrator as a 'circuit involving both positive and negative feedbacks', but include the Blumlein-Miller circuit in a section devoted to purely negative feedback arrangements. These apparent inconsistencies recur in the more recent literature.

It is clear, therefore, that if these various treatments are all factually correct a paradox exists; namely, that one and the same circuit may equally well be regarded as a case of positive or of negative feedback, or both. In this paper it is proposed to clarify this situation, by confirming the equivalence of these different points of view, both algebraically and geometrically, using the transfer network type of representation used in the servo-mechanism literature. In order to avoid the phase shift which occurs in the integrator circuits, due to the presence of a capacitor, a circuit having aperiodic feedback through an entirely resistive network will be used as the typical example for analysis. Before considering this circuit it may prove helpful to restate the feedback concept, and to introduce the transfer network method of representation.

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The Feedback Concept

The feedback concept may, in a broad sense, be applied to any network but is most useful in those con-

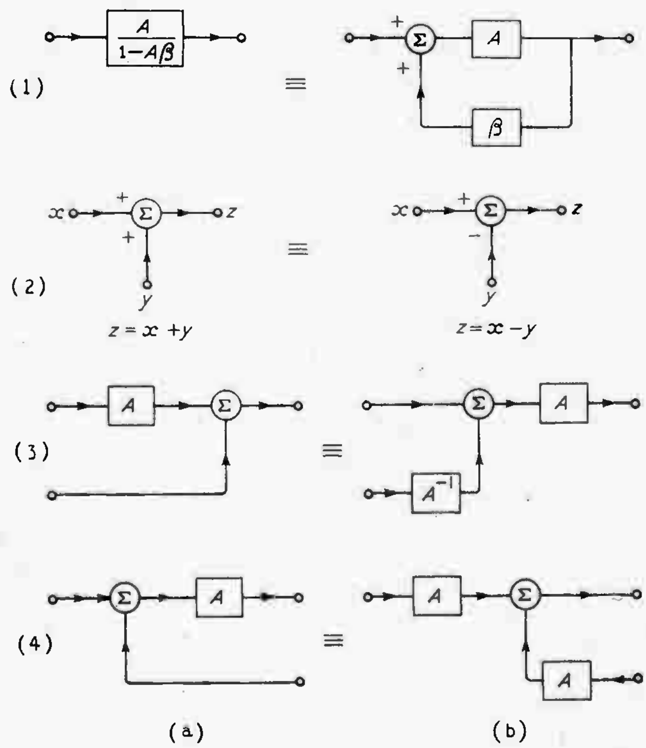
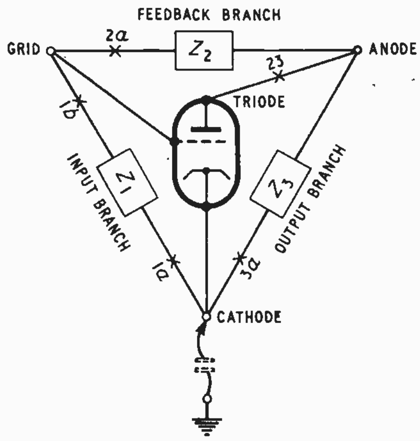


Fig. 1. *Simple transfer networks and their transformations*



	VARIANT	SIGNAL INPUT	H.T. INPUT	EARTH TO	
COMMON-CATHODE OPERATION	NORMAL	1a	3a	CATHODE	(a)
	BOOTSTRAP	1a	2b	ANODE	(b)
	EARTHED GRID	1b	2a	GRID	(c)

Fig. 2. *A regular form of the triode shunt-shunt feedback amplifier, showing its three practical variants*

taining a substantially unilateral transmission element, such as an amplifier valve, bridged by a bilateral network or by a second unilateral device acting in the opposite direction. One can then distinguish a unidirectional closed signal transmission path, through the unilateral element and back round the external circuit. The open-circuit value of the voltage gain round this loop (called the 'internal gain') is of great practical importance since, when its magnitude is sufficiently great, the gain from input to output (called the 'external gain') closely approximates the reciprocal of the return part of the loop, which is a relatively stable quantity. For this effect to be useful, however, the network must remain stable in the sense that an adequate margin must be secured against possible oscillation of the circuit. According to the Nyquist criterion, avoidance of sustained oscillation requires that the polar plot of the internal gain must not encircle the critical point (1, 0).

In general, the feedback may be said to be negative or positive according as the real part of the loop gain is negative or positive. In the following sections the argument will be conducted in terms of purely resistive networks, so that there will be no quadrature component of loop gain, and the latter will be purely positive or negative, thereby making the feedback purely positive or negative also. Any signal voltage injected into the feedback loop will be transmitted round it either with or without exact inversion (i.e., negative or positive feedback, respectively), depending on the way in which the connections are made.

Transfer Networks

In dealing with feedback systems containing several feedback loops (multiple feedback), as frequently happens in analogue computers and servomechanisms, a method of representation known as the transfer or operator network has been found useful[5]. A transfer network is a block diagram whose building blocks are mathematical operators which account for the operations performed on the signal by the corresponding parts of the actual system. It is a graphical illustration of the algebraic form of the transfer function of the system. All the elements are assumed to act unilaterally in the direction of signal flow—hence the alternative name of 'signal flow graph', which has occasionally been employed. An important feature is the use of summing point symbols to account for the combination of signals, such as input and feedback, in preference to multiple inlets to blocks. A simple example is provided by a single feedback loop; its transfer diagram is shown in Fig. 1 (1), where the summing point takes the input signal v_i and the fraction β of the output voltage v_o, and supplies the forward-acting section (A) of the loop with a total input $v_i + \beta v_o$. The total output is, therefore,

$$v_o = A\,(v_i + \beta . v_o)$$

from which may be obtained the familiar external gain formula:

$$v_o/v_i = A/(1 - A\,\beta)$$

In a feedback loop the summing point must provide either the sum or the difference of the two signals depending on the required sign of the open-circuit loop gain (generally negative), as indicated in Fig. 1 (2).

In the course of analysing these networks a number of useful transformations have become apparent, two of which are shown in Fig. 1 (3), (4). In an application of (4) to be used later (Fig. 8) an outer feedback loop is transformed into an inner one by combining the two summing points; this procedure involves modifying the β factor of the outer loop by a suitable factor in order to preserve its feedback value unchanged.

Typical Class of Circuits

A set of circuits typical of those with which this paper is concerned is shown in Fig. 2. The bootstrap feedback amplifier, which will be used in subsequent sections as an example, may be derived from the more conventional form of shunt-feedback amplifier specified at (a) simply by earthing the anode end, rather than the cathode end,

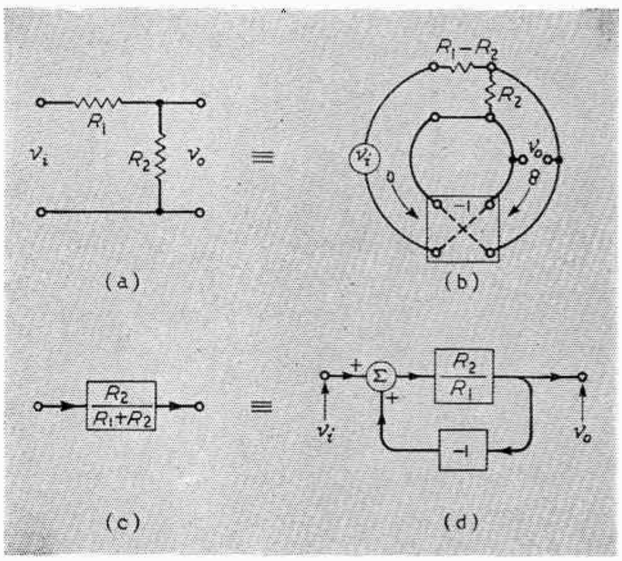


Fig. 3. *Feedback interpretation of a simple passive network*

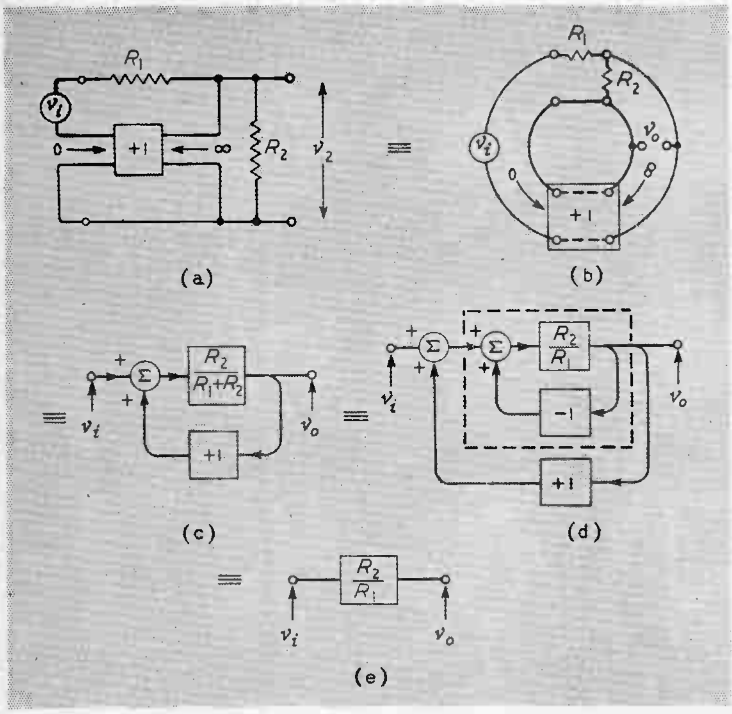


Fig. 4. *Addition of unit–β positive feedback to the network of Fig.* 3

of the load impedance. This change has no effect on the magnitude of the gain, but reverses its sign, and leaves the source of input entirely floating. It is equally possible to shift the earth lead to the grid terminal, leaving the load floating, but this circuit is rarely used. Both variants (b) and (c) call for the interposition of a two-winding transformer to allow the use of an unbalanced source or load. All three variants have the same gain and impedance properties, since the location of the earth point does not affect the basic circuit action. For aperiodic feedback, as for amplification, both impedances would be resistances; for integration Z_1 would be a resistor and Z_2 a capacitor, in which case (a) and (b) are the Blumlein-Miller and bootstrap forms, respectively. From this point onward only form (b) will be considered, with resistive impedances, and will be derived first in positive feedback form and then in negative feedback form, the two forms being found to be identical in configuration.

Positive Feedback Approach

As a preliminary step the simple transmission network shown in Fig. 3 (a) will be considered. In so far as the back-e.m.f. developed across R_2, i.e., the output voltage v_o, opposes the input voltage v_i, this network may, in a broader sense than is generally employed, be regarded as a negative-feedback system (b). Thus, taking the general formula for the external gain of a single-loop feedback system, viz.,

$$v_o/v_i = A/(1 - A\beta)$$

with $\beta = -1$ (because the entire output is, in effect, fed back) one obtains:

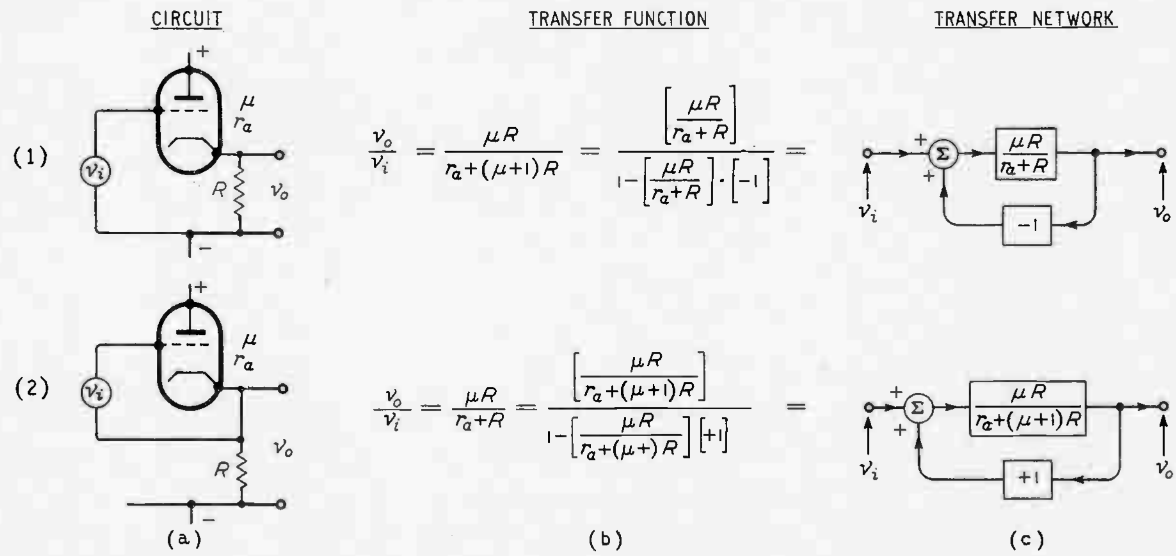
$$A/(1 + A) = R_2/(R_1 + R_2)$$

whence $$A = R_2/R_1.$$

One may, therefore, replace the direct form of the transfer network shown at (c) by the equivalent feedback form (d). If, by appropriate circuit modification, the transmission network (a) could be arranged to provide a transfer ratio of R_2/R_1 the change would be tantamount to the introduction of just sufficient positive feedback to cancel out exactly the 'negative feedback' due to the load back e.m.f. This result may be achieved by feedback through a unilateral non-inverting network having unity voltage gain, with infinite input impedance and zero output impedance,[6] as shown in circuit form in Fig. 4 (a) and in feedback loop form at (b). If the forward-acting part of the loop is substituted from Fig. 3 by its equivalent negative feedback loop the entire network will reduce to a single unilateral transfer stage of voltage gain R_2/R_1, as illustrated by the sequence of transfer network diagrams (c), (d) and (e) in Fig. 4.

Having established the scheme of Fig. 4 (a) as a

Fig. 5. *The cathode follower* 1 (*a*)*; with its transfer function* 1 (*b*)*; and transfer network representation* 2 (*c*). *The latter requires a bootstrap amplifier* 2 (*d*) *in the forward acting part of the loop whose transfer function and network are shown at* 2 (*b*) *and* (*c*)



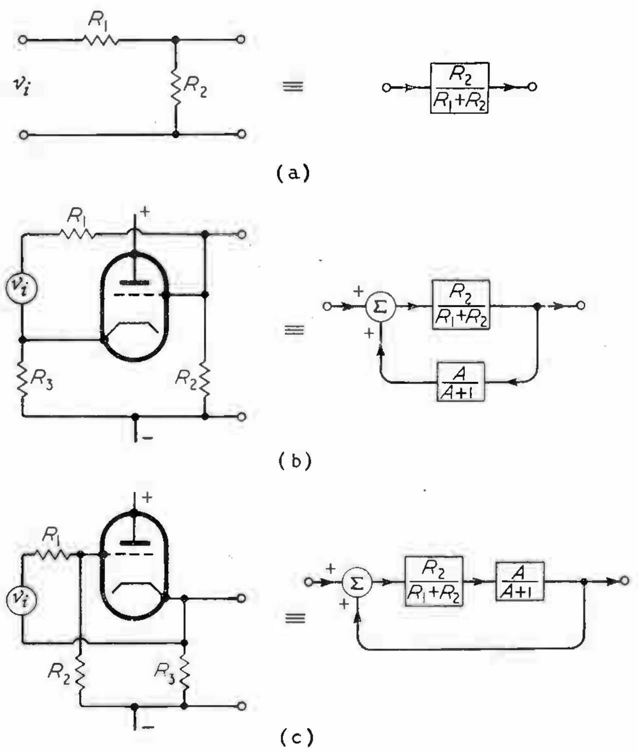


Fig. 6. *Positive-feedback derivation of the bootstrap variant of Fig. 1*

positive-feedback system, practical circuits may be obtained by replacing the hypothetical feedback stage by actual circuits having similar properties. A close approximation is provided by the cathode-follower stage shown in Fig. 5 (1) (a), where rearrangement of its transfer function into feedback form at (1) (b) shows that this circuit may itself be treated as a localized feedback loop, as shown schematically by the transfer network at (1) (c). This will be used later, in the course of transforming the positive feedback circuit being derived, into the negative feedback form obtained in the next section. Substitution of Fig. 5 (1) (a) into Fig. 4 (a), which has been repeated in Fig. 6 (a), gives the circuit of Fig. 6 (b). Output may equally well be taken from the output side of the cathode-follower without modifying the nature of the feedback, as shown at (c), resulting in the bootstrap form of Fig. 2. Thus not only the bootstrap but the earthed-cathode (Miller feedback), and the earthed-grid variants of the common-cathode shunt-feedback amplifier are interpretable as positive-feedback loops.

Negative Feedback Approach

Instead of applying positive feedback over a passive network, thereby increasing its gain, one may equally well obtain the same resultant voltage transfer by applying the required amount of negative feedback to an active network which has more than the required amount of gain. In the present case, this needs to be done by rearrangement of the circuit without changing its actual connections. Beginning with a valve connected as a bootstrap amplifier [see Fig. 5 (2)], as in Fig. 7 (a), negative feedback may be applied through a two-element resistive attenuator (R_1, R_2), as shown at (b). This circuit has a gain which, with increase of μ, approaches the reciprocal of the feedback ratio; i.e.,

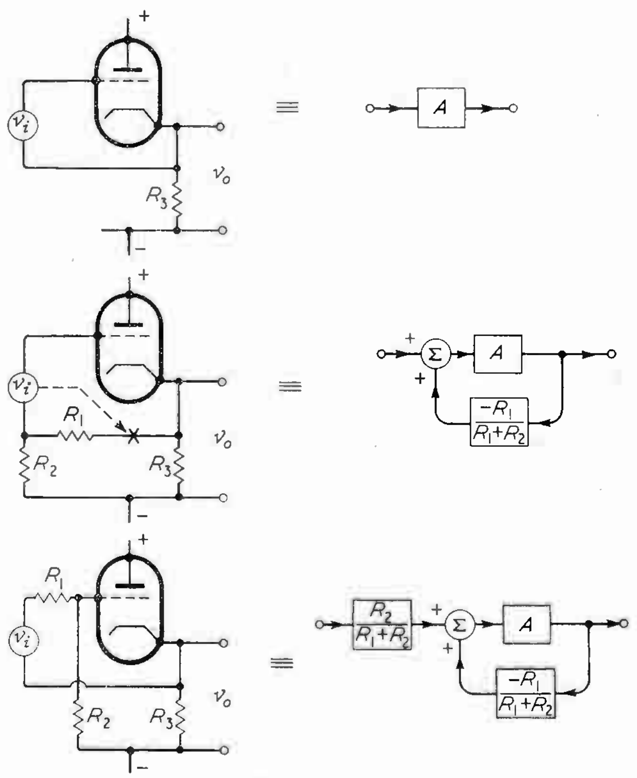
$$\frac{v_o}{v_i} \rightarrow + \frac{(R_1 + R_2)}{R_1} \text{ as } \mu \rightarrow \infty.$$

The nature and magnitude of the feedback will be unchanged if the source of input is shifted in the manner indicated by the arrow in (b), with the result shown at (c). In its new position the input signal does not appear in full between grid and cathode, as before, but is reduced by R_1, R_2 in the ratio $R_2/(R_1 + R_2)$. Since this reduction takes place outside the feedback loop the corresponding transfer network must be modified in the manner shown. With increase of μ the limiting value of gain is now the product of the signal reduction factor into the reciprocal of the feedback ratio; i.e., the simple ratio R_2/R_1. The circuit obtained at (c) is, of course, identical with that of Fig. 6 (c), as required, despite the different derivation. The circuit under consideration has, therefore, been found to be, with equal validity, an application of positive or of negative feedback.

Correlation of Derivations

One may confirm the equivalence of the two methods of approach by showing that the transfer network representation of the positive-feedback form given in Fig. 6 (c) transforms into that of the negative-feedback form shown at Fig. 7 (c)—or vice versa, using the loop form of the bootstrap (Fig. 5). Starting with Fig. 6 (c), which has been reproduced at (a) in Fig. 8, the unilateral voltage transfer network $A/(A + 1)$ is replaced by its negative-feedback equivalent Fig. 8 (b). In the circuit

Fig. 7. *Negative-feedback derivation of the same circuit as in Fig. 6*



$$A = \frac{\mu R_3}{r_a + R_3}, \ (R_1 + R_2) \gg R_3$$

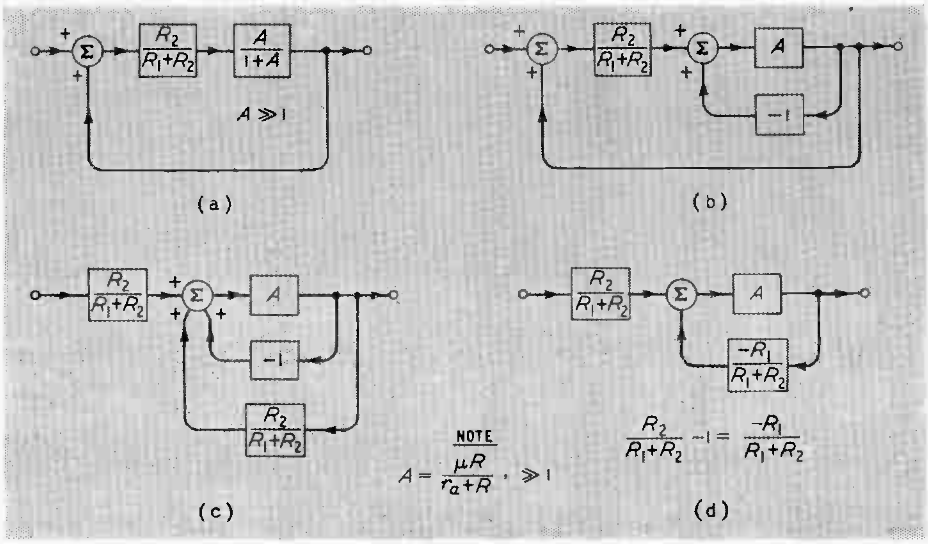


Fig. 8. *Transformation of the transfer network of Fig.* 6 (*c*) *into that of Fig.* 7 (*c*)

under consideration the unity-gain network is a cathode-follower, which is a bootstrap amplifier with 100 per cent negative feedback, as shown in Fig. 5 (c). The transfer network representation now has a localized negative feedback loop within a more extensive positive feedback loop; this is the composite feedback interpretation given by Hughes and Walker.[4] Next, the positive feedback is shifted over the attenuator R_1, R_2 to the internal summing point, and its value reduced by the factor $R_2/(R_1 + R_2)$ in the feedback path, in order to preserve the value of the positive-feedback loop gain Fig. 8 (c). The two feedbacks are then combined prior to the summing point by adding the feedback factors -1 and $R_2/(R_1 + R_2)$ [see Fig. 8 (d)]. The resultant ratio being negative; viz., $-R_1/(R_1 + R_2)$, the network reduces to pure negative form, with a signal reduction attenuator outside the feedback loop, as in Fig. 6 (c). Finally, one may check these network transformations algebraically, as follows:

$$\text{(a)}\quad \frac{\left(\dfrac{R_2}{R_1 + R_2}\right)\left(A'\right)}{1 - \left(\dfrac{R_2}{R_1 + R_2}\right)\left(A'\right)}$$

where $A' = A/(1 + A) \approx 1$ for $A \gg 1$

$$\rightarrow\text{(b)}\quad \frac{\left(\dfrac{R_2}{R_1 + R_2}\right)\left(\dfrac{A}{1 + A}\right)}{1 - \left(\dfrac{R_2}{R_1 + R_2}\right)\left(\dfrac{A}{1 + A}\right)}$$

$$\rightarrow\text{(c)}\quad \left(\frac{R_2}{R_1 + R_2}\right)\frac{A}{1 - \left\{(-1) + \left(\dfrac{R_2}{R_1 + R_2}\right)\right\}A}$$

$$\rightarrow\text{(d)}\quad \left(\frac{R_2}{R_1 + R_2}\right)\frac{A}{1 + \left(\dfrac{R_1}{R_1 + R_2}\right)A}$$

In conclusion, it may be pointed out that the treatment given applies equally to the other two practical variants of the circuit considered (Fig. 1).

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EXPERIMENTAL TELEVISION TRANSMISSIONS

A Band V television transmitter has been installed by the B.B.C. at Crystal Palace for experimental purposes. Operating at 654.25 Mc/s, the vision transmitter has a peak-white output of 10 kW and the sound transmitter is 2.5 kW. Manufactured by E.M.I. Electronics Ltd., both transmitters have low-level modulation with high-power klystron amplifiers for the final stages which give a power gain of about 100 and are driven by cathode-modulated amplifiers.

The outputs of the two transmitters are combined in a filter bridge and the power is conveyed up the aerial tower by an elliptical waveguide of some 12 in. by 6 in., the power loss in this being no more than 1.5 dB. The aerial system is mounted on top of the Crystal Palace tower with its centre line at 691 ft above ground level. There are four helical aerials mounted vertically above one another and each is centre-fed by a 5-in. coaxial feeder, the feeders being junctioned and fed by the waveguide near the base of the aerial system. The helical aerials are wound in opposite directions from the centre so that the vertical components of radiation cancel, leaving a horizontally-polarized field. The effective radiated power is 125 kW.

The transmitter is being used at present mainly for propagation tests and is radiating the normal programme of the television service. These are, of course, 405-line transmissions. About March 1958 it is intended to transmit 625-line pictures, derived from a flying-spot film scanner supplied by Cinema-Television Ltd., and conforming to C.C.I.R. standards. These standards, of course, imply the use of negative modulation and f.m. sound.

The purpose of the experiments is not merely to find out whether the propagation characteristics of Band V are suitable for television, it is also to determine whether the 625-line system has any advantages over the 405-line under normal reception conditions as they exist for the normal viewer. By using duplicate copies of films, it will at times be possible to radiate simultaneously the same programme with 405 lines on Band I and 625 lines on Band V and to obtain a direct comparison of the performance on two receivers side by side.

RADAR IN THE RAIN

When I consulted B. J. Mason's recent book on *The Physics of Clouds* last month, it was only to check dutifully on some points in the theory of the Wilson cloud chamber. But this is not the sort of subject one can dip into without becoming interested, and I found in the latter part of the book so good an example of the way in which radio engineering permeates the instrumentation of pure science that it seemed very suitable for a 'Fringe' discussion. The application is as old as radar itself—indeed, writing ten years ago in the Pilot Press book "Electronics", R. A. Smith mentioned its use for the analysis of cloud structures, for the study of the mechanism of rain formation, and for a detailed examination of frontal conditions.

Scattered Radiation and its Reception

The usual calculation for the power P_r at the receiver (as given, for example, by Smith in the book referred to above) introduces the ideas of the 'scattering area' of the target and the effective aperture of the receiving aerial. The only difference between a water-drop target and an aircraft from this point of view is that one is treated as an assemblage of dielectric spheres, and the other as a continuous conducting sheet. In an electric field, each water molecule develops a dipole moment proportional to the field-strength. The Clausius-Mossotti (or Lorentz-Lorenz) equation gives its value, in unit field-strength, as proportional to $(\epsilon - 1)/(\epsilon + 2)$, where ϵ is the permittivity of the bulk material which, in a radiation field, depends on the wavelength. This calculation is given on p. 98 of von Hippel's "Dielectrics and Waves", and in Chapter 18 of Bleaney and Bleaney's "Electricity and Magnetism". Proceeding from molecules to drops, the unit-field dipole moment of a drop of diameter D is proportional to its volume, and is

$$\left(\frac{\epsilon - 1}{\epsilon + 2}\right)\left(\frac{D}{2}\right)^3.$$

For the scattering, provided that D is very much smaller than the wavelength λ of the radiation, the classical Rayleigh '$1/\lambda^4$' formula, as derived in the standard works on light, applies. The whole purpose of the operation in the end is to find out something about the structure of the target by a method which usually determines only the effective target area directly. So the difficult part of the theory is completed when the target area has been calculated in terms of the dipole moment of a drop and the Rayleigh scattering. Taking the simplest possible case, the radiation is supposed to fall on an assembly of small spheres, all of the same D; in fact, since D is at most a few millimetres and λ is between 3 cm and 10 cm, the necessary condition $D \ll \lambda$ is satisfied.

Suppose the transmitter, emitting pulses of centimetre radiation with power P_0 (of the order of 100 kW) concentrates this in a narrow beam, which gives an aerial gain G as compared with an isotropic radiator. Then the power received per unit area of target at distance R is $P_0G/4\pi R^2$. If this beam falls on a target of *effective area* T and is scattered back to an aerial of effective area A at the same place as the transmitter, the power received there is $P_r = ATP_0G/(4\pi R^2)^2$, the extra $4\pi R^2$ factor arising as the inverse square law operates on the way back without benefit of gain.

Next, to calculate T in terms of the properties of the particles and their population-density. Let V be the volume traversed by the beam, and n the number of particles per unit volume. The dipole moment of a sphere in unit field is given by

$$\left(\frac{\epsilon - 1}{\epsilon + 2}\right)\left(\frac{D}{2}\right)^3$$

and the radiation scattered per unit solid angle is proportional to the square of this quantity. It is also inversely proportional to the fourth power of the wavelength λ. The quantity T is not an actual area, but is more like what the highbrows call a 'scattering cross-section'; for

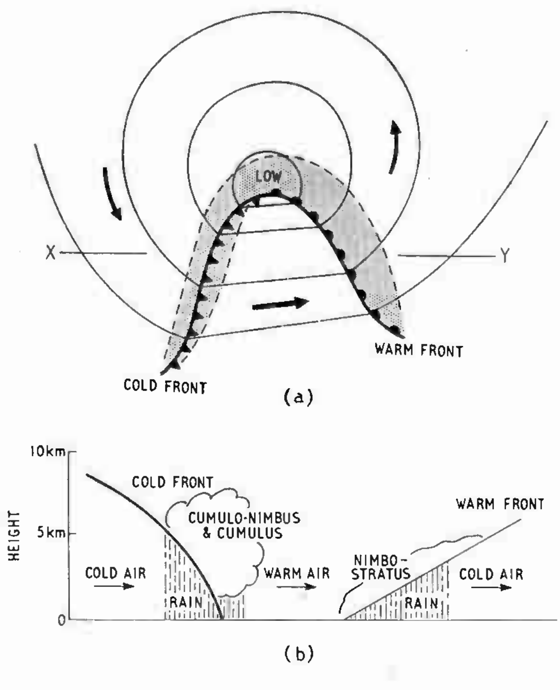


Fig. 1. *(a) The usual isobar pattern of a frontal depression, showing cold and warm fronts, the general direction of air circulation, and (shaded lightly) the regions where rain is falling; (b) A vertical section through (a), along the line XY. Only the chief radar-target clouds and precipitation are shown. At the cold front, these are massive cumulo-nimbus or cumulus clouds, and heavy rain-showers; at the warm front, a more extensive area of layer rain-clouds (nimbo-stratus) and steady rain*

as the formula for P_r shows, $T/4\pi R^2$ represents the fraction of the incident power which gets back to the receiver from the target. Anyhow, calculations similar to those given in the reference books quoted above lead to the value

$$T = \frac{nV\pi^5}{\lambda^4} \frac{(\epsilon - 1)^2}{(\epsilon + 2)^2} . D^6,$$

and thus give

$$P_r = \pi^3 \frac{P_0 AG}{16R^4\lambda^4} \frac{(\epsilon - 1)^2}{(\epsilon + 2)^2} nVD^6.$$

The assumption that D is the same for all the particles in an actual cloud or rain-shower is a gross over-simplification and, if the information is available, then the factor nVD^6 should be replaced by a summation involving the 'drop-spectrum' of the n and D for each size of particle. Many ways of determining the size-distribution have been tried, from simply collecting the drops on dyed filter-paper, vaselined slides, or flour-covered trays, to a mass-spectrometer analogue in which they are dispersed by a horizontal air-current as they fall. Perhaps the most elegant so far is that of Mason and Ramanadham, in which the drops fall through a zone strongly illuminated by visible or ultraviolet light; the energy reflected by each drop depends on its area, and after focusing on a photocell causes a corresponding voltage pulse; an electronic pulse-sorting circuit does the rest. The trouble is, of course, that the precipitation received on the ground has altered its character on the way down, and the only way of being sure about the distribution at a given level seems to be to send an aircraft with some straightforward sampler to find out.

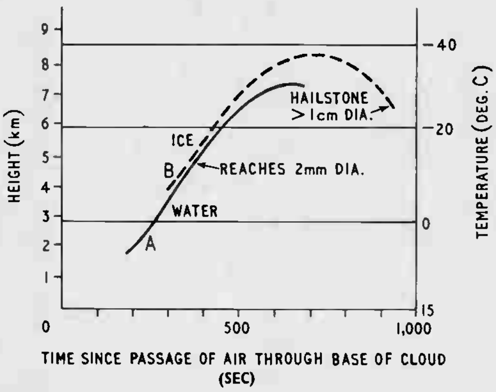
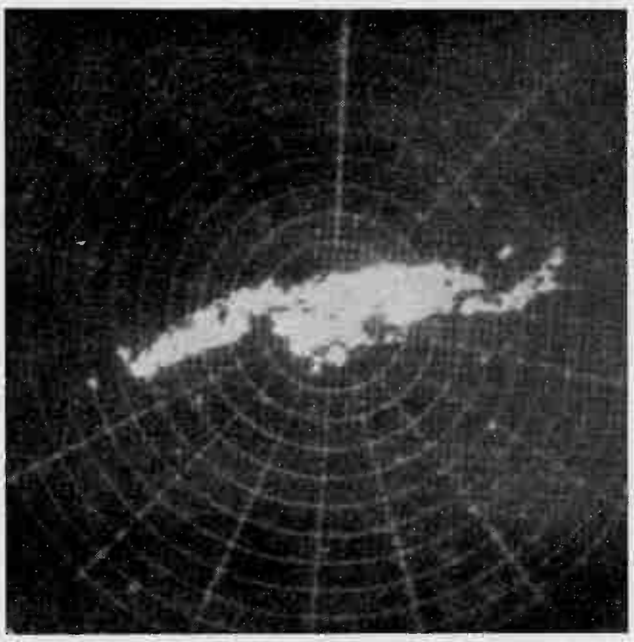


Fig. 2. *Representing the rate of ascent of water-drops and hailstones, which grow all the time as they rise in the ascending air current within a cumulus cloud. The radar echo can distinguish between them, and also by the enhanced intensity can show the level at which the falling hailstone has melted appreciably*



(*Courtesy of Dr. B. J. Mason and Meteorological Office*)

Fig. 3. *Radar echo from a cold front, displayed on p.p.i. The structure of Fig.* 1 (*b*) *should give such relatively narrow but intense echo-regions*

Finally, the gain G, assuming that the whole of V is covered by the beam, can be written as $2\pi^2\tau c/V$, where τ is the pulse duration and c the wave-velocity (the velocity of light). The minimum detectable power is of the order of 2×10^{-13} watt; in order to get as high a P_r as possible, λ should be short and τ should be long. On the other hand, short waves suffer greater attenuation, and pulses of long duration locate the target less precisely and so give less resolution of detail. Most of the figures given by Mason are for $\lambda = 3{\cdot}2$ cm and $\lambda = 10$ cm, apparatus employing the former being the more sensitive by a factor of about a hundred, as the $1/\lambda^4$ term suggests. He also states that the optimum λ for centimetre-wave work would lie between these two figures, at 5·6 cm. Attenuation by air, water-vapour, and light clouds formed of droplets less than 100μ in diameter are between them relatively trivial even at 3·2 cm; but the percentage attenuation in heavy rain is about thirty times as great at 3·2 cm as at 10 cm. In some circumstances, then, the 10-cm radar may prove the more sensitive.

Information from the Echo

The Clausius-Mossotti term $(\epsilon - 1)^2/(\epsilon + 2)^2$ in the expression for P_r works out, for the centimetre range, to a little more than 0·9 for water, and to 0·18 for ice—which is some five times less. The intensity of the echo thus tells (if we are sure enough about n and D) whether the reflecting particles are liquid or frozen. An ice sphere coated to a depth of one-fifth of its radius with water reflects as strongly as water.

Spherical particles inflict no change on the plane of polarization of the radiation; the reflection from non-spherical particles—prolate or oblate spheroids—contains a perpendicular component, or is partially depolarized. The polarization at the receiver thus gives a clue as to the shape of the droplets, reinforced by the fact that attenuation is greater for spheroids than for spheres of the same volume. Ice-particles show effects of the same kind, but much less in magnitude.

Finally, there is a relationship between the rate of precipitation, as it would be measured by a rain-gauge, and the intensity of the echo from falling rain; but this is rather a poor relation.

Fig. 1 shows the familiar frontal depression isobar pattern, with the air circulation anti-clockwise, cold air undercutting the warmer air along the spiked 'cold front', and warm air rising more gradually over cooler air at the spotted 'warm front'. Below is a simplified weather picture along a vertical section through the warm sector from X to Y. Along the cold front, the main clouds are cumulus or cumulo-nimbus, tall heavy rain-clouds which are precipitating strongly. The warm sector itself may be covered with low stratus cloud, with or without rain. At the warm front, more rain-clouds, usually nimbo-stratus which spread horizontally rather than vertically—and below them of course more rain. Because of the clouds and the rain-curtains, the fronts themselves are detectable at distances of the order of 100 km. It might be added that the *rain* droplets within the cloud, and falling from it, will have diameters between about 1 mm and 5 mm. Outside these limits, smaller ones fall too slowly to be classed as rain, while larger ones break up in flight; much larger particles of ice or hail may be present.

Fig. 2, based on a diagram in the article by B. J. Mason and F. H. Ludlam in Vol. XIV of the Physical Society's 'Reports on Progress in Physics' (1951) shows the growth-lifetime and ascent of a typical small water-droplet (A) and ice-particle (B) in the ascending current of air within a large cumulus cloud. At the end of the full line, A has become a supercooled water droplet of radius 1 mm and begins to fall. At the end of the dotted line, B is a large hailstone of several millimetres diameter, and is well on the downward path; B will start to melt a few hundred metres below the 0°C level, and will thereafter shed large water drops in its wake. Echoes from a vertically-directed radar beam, with the information they can give about the nature of the reflecting particles, detect this kind of thing as it is happening.

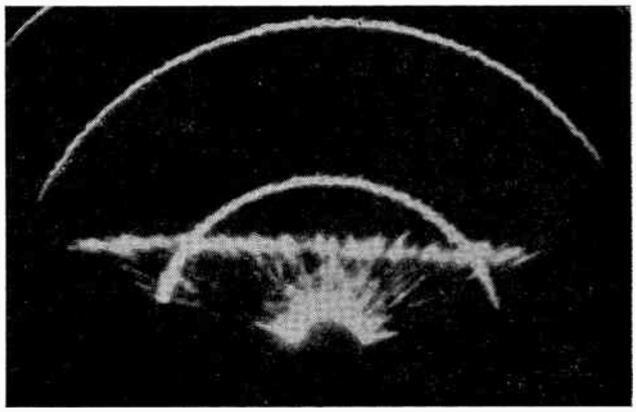
Methods of Observation

Three distinct methods are used. First, the straightforward p.p.i. display (Fig. 3) which simply detects and locates the cloud, front, or rainfall. Secondly, the range-height indicator (r.h.i.), of which Fig. 4 shows a typical record; the extensive horizontal trace is that of the enhanced reflection at the 'melting-zone' level, where hailstones or snowflakes are well on with their melting on the downward journey. Thirdly, the amplitude recorder (the 'A-scope') which displays the amplitude of the reflected signal against the range, or against the altitude if the beam is vertical (Fig. 5). In the picture, altitude is read downwards; at the 'top' is the mark of a calibrating signal, injected to correct for non-linearity. Quantitative information more or less equivalent to that of Fig. 5 can also be obtained by a pulse-integration method, dealing with the echo from one level at a time; but conditions are changing so rapidly throughout a cumulus that only a simultaneous picture of the whole is really valuable.

Further Outlook

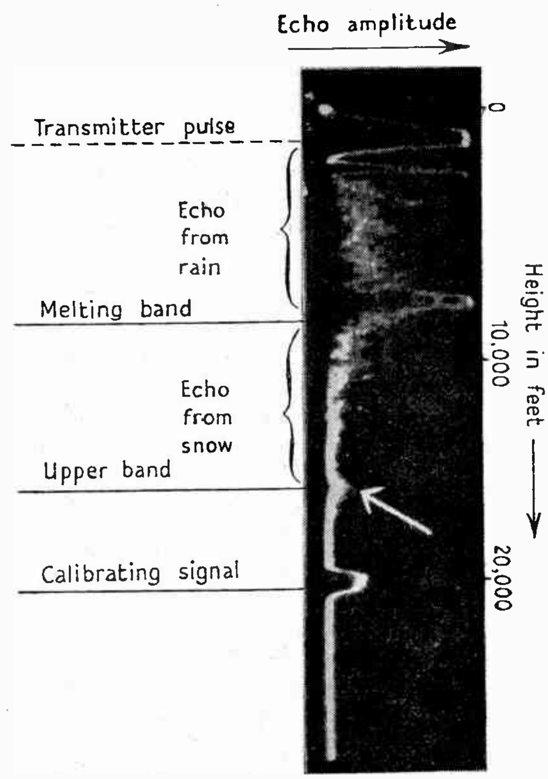
The primary aim of the work which has been so briefly outlined here is to find out more about the physical processes that lie behind the weather. The chief value to the meteorologist in his day-to-day work would seem to be in the continuous minute-by-minute reports it can give on rapidly-changing situations, and the steady tracking of fronts and storms and hurricanes. It may be that 10-cm radar can give as good a general picture of the rainfall over a large area as a thinly distributed chain of rain-gauges would, but it seems doubtful in view of the uncertainty about the drop-spectrum whether an individual estimate of rate of rainfall can be reliable.

For the purposes described here, the 3·2 and 10-cm wavelengths appear to be satisfactory; this is fortunate, since a good deal of existing equipment of this kind was available to be adapted to the work. But these wavelengths are useless for particles below the usual raindrop size. With millimetre equipment, however, the investigations could be scaled down to fine clouds, mists, and smaller particles; and future progress will probably exploit the millimetre-wavelength range for these purposes.



(*Courtesy of Dr. B. J. Mason*)

Fig. 4. *Precipitation from a warm front, shown on the r.h.i. display. The range circles are at two-mile intervals. The upper horizontal echo is that from the melting band*



(*Courtesy of Dr. I. C. Browne and Dr. B. J. Mason*)

Fig. 5. *Conditions within a warm front, A-scope display. The beam was directed vertically. Note the difference in intensity between the echoes from rain and snow, and the very strong echo from the melting band*

Polyphase Oscillators

By A. S. Gladwin, D.Sc., Ph.D., F.Inst.P.*

SUMMARY. *A set of polyphase voltages can be produced either by a symmetrical polyphase oscillator or by tapping off at suitable points on the feedback network of a single-phase resistance-capacitance oscillator. Circuit arrangements suitable for both types of generator and for an odd and even number of phases are described. It is shown how spurious oscillation in the symmetrical type of oscillator may be avoided.*

A formula connecting the oscillation frequency with the magnitude of the harmonic voltages is derived and is applied to estimate the relative performance of the two types of oscillator with various circuit arrangements. The single-phase type of oscillator is found to be greatly superior to the symmetrical polyphase form in respect of frequency stability and freedom from harmonics.

The process by which the oscillation amplitude builds up is examined and it is shown that for the symmetrical type of oscillator the build-up time is much greater than was formerly supposed. The discrepancy is related to the frequency change which occurs during the build-up period.

A polyphase oscillator forms the basis of an electronic polyphase supply system and is particularly convenient for variable-frequency and for very-high or very-low frequency operation. 3-phase systems are the most common but 2-phase supplies are often required in connection with bridges and other measuring instruments.

A set of polyphase voltages can be produced in two main ways: by a symmetrical polyphase oscillator, or from a single-phase oscillator by the use of phase-shifting networks. A symmetrical n-phase oscillator is made up of n identical sections, each consisting of an amplifier and a coupling network, connected in tandem to form a closed loop. Fig. 1 shows, for example, a 3-phase oscillator of the type described by van der Mark and van der Pol[1]. The phase shift between the voltages at successive anodes is $2\pi/3$ and the frequency can be varied by adjusting the values of R or C. This particular circuit, however, is prone to spurious oscillation at high frequency and it cannot be made to operate with an even number of phases.

If a phase-shifting network is used in conjunction with a single-phase oscillator the network requires readjustment when the frequency is changed. A logical step is to incorporate the phase-shifting network into the oscillator as part or whole of the frequency-determining system. Fig. 2 shows the a.c. equivalent circuit of a well-known form of RC oscillator. A phase shift of nearly π exists between the voltages at points 0 and 3. Hence, by tapping at suitable points on the resistors R_1, R_2 or R_3, voltages having any phase between 0 and π with respect to the voltage at point 3 can be obtained. Thus, to produce a 3-phase system, voltages with relative phase shifts of 0, $\pi/3$ and $2\pi/3$ would be tapped off. The phase of the second voltage would be changed to $-2\pi/3$ by means of a phase-reversing amplifier, and the magnitudes of the three voltages adjusted to give a balanced 3-phase system. A number of other arrangements are also possible.

In the symmetrical type of generator the anode currents also form a balanced polyphase system and the current of oscillation frequency in the h.t. supply line is zero. This simplifies the problem of decoupling, especially at very low frequencies. However, with neither form of generator would the load be connected directly to the oscillator because of the adverse effect on the frequency and amplitude; an intermediate amplifier would always be used. Since the anode currents in the amplifiers also form a balanced system and are generally much larger than the currents in the oscillation generator, any residual alternating current in the supply line due to the use of a single-phase oscillator would be a small fraction of the total direct current. In practice, the symmetrical type of oscillator has little advantage in this respect.

The object of the paper is to compare the various types of symmetrical polyphase and single-phase oscillators with respect to such factors as frequency stability and harmonic generation. These questions are particularly important for resistance-capacitance oscillators, since, because of the poor frequency-discriminating properties of the network, the oscillator may generate large amounts of harmonics unless the operating conditions are carefully adjusted.

1. Conditions for Oscillation

Oscillators are essentially nonlinear devices but it is convenient to consider first the behaviour of the circuit when the alternating voltages and currents are so small that the amplifiers can be regarded as linear. The circuit equations are then linear differential equations

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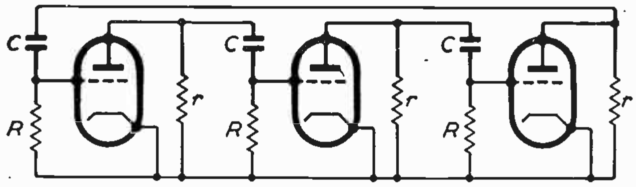

Fig. 1. *A.C. equivalent circuit of symmetrical 3-phase oscillator*

with constant coefficients the solutions of which are of the form $\Sigma\, V_n \exp(a_n t) \cos(\omega_n t + \theta_n)$. If all the values of a_n are negative, oscillations cannot build up from a small value. It does not follow that oscillation is impossible, for the circuit may be able to oscillate with a large amplitude though not with a small one. Again, if two or more of the values of a_n are positive there is the possibility, depending on the initial conditions, of oscillation at one or other of the corresponding frequencies, or of oscillation at two or more frequencies simultaneously. In what follows the oscillator is assumed to be 'well-behaved'; i.e., whatever the initial conditions the oscillation eventually reaches the same final values of amplitude and frequency. The problem is then to determine the conditions under which one, and only one, of the values of a_n is positive. Any small initial disturbance then initiates an oscillation which builds up smoothly to the steady-state regime.

Fig. 3 shows the a.c. equivalent circuit for one section of an n-phase oscillator made up of n such identical sections. In practice, no grid current can be permitted because of its serious effect on the oscillation frequency. This requirement arises from the fact that in polyphase oscillators the frequency is determined as much by the resistive as by the reactive elements in the coupling network. Hence the output impedance Z_0 has no effect; only the open-circuit input and transfer impedances Z_i and Z_t need be considered.

In complex notation the relation between the two grid voltages in Fig. 3 is

$$v_{2g} = Av_{1g}$$

$$\text{where } A = -\mu Z_t/(r_a + Z_i) \quad \ldots \quad \ldots \quad \ldots \quad (1.1)$$

and μ and r_a are the amplification factor and anode (slope) resistance of the valve. For voltages of the form $V \exp(at) \cos(\omega t + \theta)$ the values of Z_t and Z_i are to be taken at the complex frequency $p = a + j\omega$.

Proceeding to the following stages, $v_{3g} = Av_{2g} = A^2 v_{1g}$ etc., and since the n^{th} stage is followed by the first stage, $v_{1g} = Av_{ng} = A^n v_{1g}$. Hence

$$A^n = 1 \quad \ldots \quad \ldots \quad \ldots \quad (1.2)$$

The possible values of p are the roots of this equation and the criterion for an oscillation to build up is that two, and only two, of the roots should be complex conjugates with positive real parts. All other roots must be negative or have negative real parts. This requirement can be expressed in terms of the behaviour of A^n at real frequencies ($p = j\omega$); for the number of roots of the equation with positive real parts is equal to the number of times which the locus of A^n in the complex plane encircles the point (1,0) as ω varies from $-\infty$ to ∞. The criterion for oscillation to build up is therefore that the locus should encircle the point twice. This is Nyquist's criterion and the conditions under which it is valid are always met in circuits of the type under consideration. Since the value of A for a negative value of ω is the complex conjugate of its value for ω positive, only that part of the locus corresponding to $0 < \omega < \infty$ need be drawn.

It is simpler, however, to express the criterion for oscillation in terms of the locus of A rather than that of A^n. Equation (1.2) can be written as

$$A = 1^{1/n} = \exp(-j2\pi k/n) \quad k = 1, 2, \ldots n.$$

This set of equations will have two roots with positive real parts if the locus of A encircles once two of the points $\exp(-j2\pi k/n)$ having complex conjugate values as ω varies from $-\infty$ to ∞. Again, because of the symmetry of the locus, the range of ω can be restricted to 0 to ∞.

The relation between consecutive grid voltages is

$$v_{2g} = Av_{1g} = v_{1g} \exp(-j\phi)$$

$$\text{where} \quad \phi = 2\pi h/n \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1.3)$$

and h is the value of k corresponding to the point enclosed by the locus of A with $0 < \omega < \infty$. Thus the grid voltages are all of equal amplitude and differ only by a constant phase shift.

As the oscillation builds up the nonlinear properties of the amplifier are evoked; the effective value of r_a

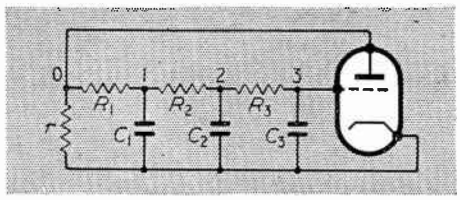

Fig. 2. *A.C. equivalent circuit of single-phase oscillator*

changes and the locus shrinks until it touches the point $\exp(-j\phi)$. If the steady-state harmonic voltages are negligibly small the steady-state frequency ω_1 can be calculated from the equation

$$A = -\mu Z_t/(r_a + Z_i) = \exp(-j\phi)$$

with $p = j\omega_1$. Let $Z_t = R_t + jX_t$ and $Z_i = R_i + jX_i$. By equating the real and imaginary parts separately two equations are obtained, and, by eliminating r_a between these, the frequency-determining equation is found to

$$R_t \sin\phi + X_t \cos\phi + X_i/\mu = 0 \quad \ldots \quad \ldots \quad (1.4$$

2. Circuit Arrangements

The criterion for oscillation build-up developed in the previous Section is now applied to examine the performance of oscillators having various forms of coupling network. Fig. 4 shows the coupling network of the van der Mark–van der Pol type of oscillator. C_a and C_g represent the anode-earth and grid-earth capacitances; the effect of the anode-grid capacitance is discussed later.

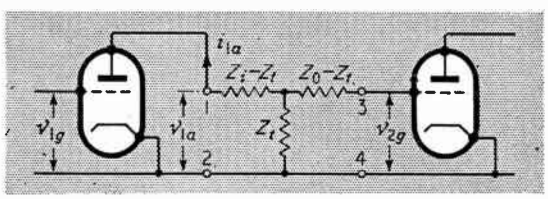

Fig. 3. *One section of polyphase oscillator*

The impedances Z_t and Z_i are easily evaluated and it is a straightforward matter to show that the locus of A as ω goes from 0 to ∞ is a circle passing through the origin with its centre on the real axis. Fig. 5 shows such a locus together with the points 'a', 'b' and 'c' given by $\exp(-j2\pi k/3)$ with $k = 1, 2, 3$, corresponding to a 3-phase oscillator.

There are two frequencies at which the phase of A is $\pm 2\pi/3$; a low frequency ω_1 determined mainly by the product RC, and a high frequency ω_2 determined mainly by the product $r(C_a + C_g)$. It is immediately obvious that the locus must enclose either no point, in which case there is no oscillation, or both points 'a' and 'b', in which case two oscillations of frequencies near to ω_1 and ω_2 build up simultaneously. Capacitance between grid and anode has the effect of modifying slightly the shape of the locus at high frequencies by decreasing the magnitude of A. Thus, by careful adjustment of the gain, point 'b' could be made to lie outside the locus while point 'a' remained inside. Single-frequency operation would then be secured, but the difference in gain required for this is so small that the discrimination cannot be achieved in practice.

Unwanted oscillation at high frequency is usually suppressed by increasing artificially the value of C_a or C_g in one of the coupling networks. At low frequencies this has little effect and the oscillator continues to behave as a symmetrical system, but at high frequencies the symmetry is destroyed and the loop gain round the circuit is so much reduced that the high-frequency oscillation is unable to start. A more fundamental approach to the problem is to design the coupling network so that only one frequency of oscillation is possible. Fig. 6 shows an a.c. equivalent circuit, and Fig. 7 a practical embodiment, of a coupling network which ensures single-frequency operation. The locus of the stage gain A and the three critical points corresponding to a 3-phase oscillator are shown in Fig. 8.

Inspection of Fig. 5 shows that the network of Fig. 4 is unsuitable when the number of phases, n, is even. For, as the diameter of the locus is increased by increasing the amplifier gain from zero, the point 'd' $(-1,0)$ corresponding to $k = \frac{1}{2}n$ is the first to be enclosed. An oscillation then builds up at a frequency near to ω_3 and with a phase shift of π between successive grid voltages. A set of polyphase voltages cannot be obtained. Coupling networks suitable for an even number of phases are therefore more complicated. A network suitable for 4-phase operation is shown in Fig. 9, and Fig. 10 shows the corresponding locus of A with the four critical points $\exp(-j2\pi k/4)$ $k = 1, 2, 3, 4$. When the circuit constants are chosen to make $LC_1 = C^2R^2$, the frequency of oscillation is $\omega_1 = 1/CR$ (with $\mu = \infty$ and $r_a = \infty$).

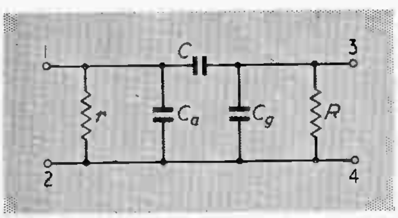

Fig. 4. *Coupling network for oscillator of Fig. 1*

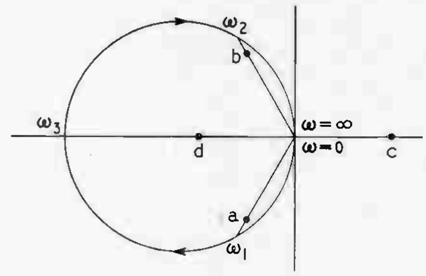

Fig. 5. *Locus of stage gain A for circuit of Fig. 4*

Single-phase oscillators have already been mentioned. Although a voltage of any specified phase between 0 and π can be obtained by tapping on the resistors R_1, R_2 and R_3 of Fig. 2 it is usually more convenient to choose the element values so that the voltages at the junction points 1, 2, 3 have the required phase differences. In Fig. 2, for example, let

$$C_1 = C_2 = C_3 = C,\ R_3 = R,\ R_2 = 2R,\ R_1 = 5R/3 - r(1 - 35/\mu) \quad \ldots \quad (2.1)$$

Taking the grid voltage as the reference, the voltages at points 3, 2, 1 are then

$$v_3 = V\cos(\omega_1 t),\ v_2 = 2V\cos(\omega_1 t + \pi/3)$$
$$v_1 = 10V\cos(\omega_1 t + 2\pi/3)$$

where $\omega_1 CR = 3^{1/2}$. By reversing the phase of v_2 and adjusting the magnitudes of v_2 and v_3 a set of 3-phase voltages is obtained. The three capacitances need not be equal although equality would usually be desirable in a variable-frequency oscillator. For example, let $C_3 = C$, $C_2 = aC$, $C_1 = bC$, $R_3 = R$, $R_2 = R/d$, $R_1 = R/f - r$, and $\mu = \infty$. Then if $d = a - \frac{1}{2}$ and $f = b - (2a - 1)(a + 1)/(4a + 1)$, the voltages at points 3, 2, and 1 are $v_3 = V\cos(\omega_1 t)$, $v_2 = V_2\cos(\omega_1 t + \pi/3)$, $v_1 = V_1\cos(\omega_1 t + 2\pi/3)$ where $V_2 = 2V$ and $V_1 = 2V(4a + 1)/(2a - 1)$. By making $a \gg 1$ and $b \gg a$, the voltage gain required of the amplifier can be reduced from the value of about 35 to the minimum value of about 8. In a similar way the element values can be chosen to give phase shifts of $\pi/3$ between the voltages at points 0, 1, and 2. This arrangement gives a somewhat greater voltage output.

Similar results can be derived for the shunt-resistance type of oscillator obtained by interchanging the positions of the resistors and capacitors in Fig. 2. In this network the phase shift increases (instead of decreasing) between the input and output terminals. Such a network can be combined with that of Fig. 2 as shown in Fig. 11 to enable a set of 3- or 6-phase voltages to be obtained without the use of phase-reversing amplifiers. If $r \ll R$ the phase shift increases by $\pi/3$ between consecutively numbered points and the voltage magnitudes are in the ratios 35 : 10 : 2 : 1 : 2 : 10 Another way of obtaining a set of 3-phase voltages is to add a fourth section to the right-hand side of the coupling network of Fig. 2 to give an additional phase shift of $\pi/3$.

A 2-phase system can similarly be obtained by tapping on R_2 in Fig. 2 and taking the quadrature voltage from point 0 or 3, or alternatively by tapping on R_3, and taking the quadrature voltage from point 1. When the element values given by (2.1) are used and R_2 is tapped at 5/6 of the way between points 1 and 2, the voltage at the tapping point is $v_q = -(5/3^{1/2})V\sin(\omega_1 t)$ when $v_3 = V\cos(\omega_1 t)$.

Reich[2] and Fleming[3] have proposed an arrangement in which cathode-follower buffer stages are interposed between the three sections of the feedback network of a single-phase oscillator as shown in Fig. 12. The three resistances can be made equal and likewise the three capacitances. The overall gain required is slightly more than 8 and this is provided by the final amplifier stage. By suitable choice of resistance values a set of balanced 3-phase voltages can be taken from the points 1, 2 and 3, the resistors between 4 and 5 and 3 and 5 having been removed. When these resistors are retained a 2-phase output can be obtained at points 1 and 5.

In order to obtain very low frequency oscillations without using excessively large values of resistance and capacitance Smiley[4] has proposed a symmetrical form of 3-phase oscillator one section of which is shown in basic form in Fig. 13. This uses feedback from the anode circuit (Miller effect) to increase the effective value of C. The idea may also be applied to the cathode-follower-coupled single-phase oscillator as shown in Fig. 14. Here the feedback is in the cathode circuit and its effect is to increase the effective value of R. For the same values of R and C and feedback factor, the frequency of the shunt-resistance oscillator (Fig. 14) is one-third of that of the shunt-capacitance form (Fig. 13). In both oscillators the frequency depends on the feedback factor and can be controlled, for example, in the circuit of Fig. 14, by adjusting the value of the cathode resistor R_c.

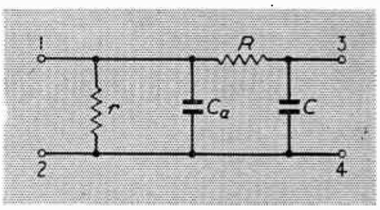

Fig. 6. *Shunt-capacitance type of coupling network*

Some discussion of the relative merits of the shunt-resistance and shunt-capacitance types of polyphase and single-phase cathode-coupled oscillators from the aspects of variable-frequency operation and the effects of stray capacitance can be found in references 2–5. In the following sections the main types of oscillator are compared in respect of frequency stability and harmonic generation.

3. Effect of Harmonics on Steady-State Frequency

It is well known that the frequency of an oscillator is affected by the presence of harmonic voltages at the grid and anode of the amplifier valve, and Groszkowski[6] has derived a very general formula for the frequency of a single-phase oscillator in terms of the magnitudes of these voltages. In this section it is shown how Groszkowski's method can be adapted to polyphase oscillators.

In Section 1 it was shown that, under linear conditions, the relation between consecutive grid voltages is

$$v_{2g} = v_{1g}\exp(-j\phi)$$

When the complex exponential time factor is restored to both sides the equation can be written

$$V\exp(j\omega t) = V\exp[j\omega(t - T)]$$

where V is the magnitude of the two voltages and $\omega T = \phi$. Under steady-state conditions V is constant and v_{1g} and v_{2g} are functions of time such that $v_{2g}(t) = v_{1g}(t - T)$. It is now assumed that, under steady-state conditions, a similar relation holds also when the voltages are non-sinusoidal. The frequency is then changed to ω_0 and the time delay to T_0, but the product

$$\omega_0 T_0 = \omega T = \phi = 2\pi h/n \quad \ldots \quad \ldots \quad (3.1)$$

remains constant. The relation between v_{1g} and v_{2g} is

$$v_{1g}(t) = v_{2g}(t + T_0) \quad \ldots \quad \ldots \quad \ldots \quad (3.2)$$

Let Fig. 3 represent one section of an n-phase oscillator. The alternating anode current in valve 1 can be expressed in the form

$$i_{1a} = \tfrac{1}{2}\sum_{-\infty}^{\infty} I_n \exp(jn\omega_0 t + j\theta_n) \quad \ldots \quad \ldots \quad (3.3)$$

with I_n real and $I_{-n} = I_n$, $I_0 = 0$, and $\theta_{-n} = \theta_n$. Since the grid current has been assumed zero the relations between v_{1a}, v_{1g} and i_{1a} are $v_{1a} = -Z_l i_{1a}$ and $v_{2g} = -Z_t i_{1a}$. Hence and from (3.3)

$$v_{1a} = -\tfrac{1}{2}\sum_{-\infty}^{\infty} Z_{ln} I_n \exp(jn\omega_0 t + j\theta_n)$$

$$v_{2g} = -\tfrac{1}{2}\sum_{-\infty}^{\infty} Z_{tn} I_n \exp(jn\omega_0 t + j\theta_n)$$

where Z_{ln} and Z_{tn} are the values of Z_l and Z_t at frequencies $n\omega_0$. From (3.1) and (3.2) it follows that

$$v_{1g} = -\tfrac{1}{2}\sum_{-\infty}^{\infty} Z_{tn} I_n \exp(jn\omega_0 t + jn\phi + j\theta_n)$$

If the amplification factor μ is constant, the anode

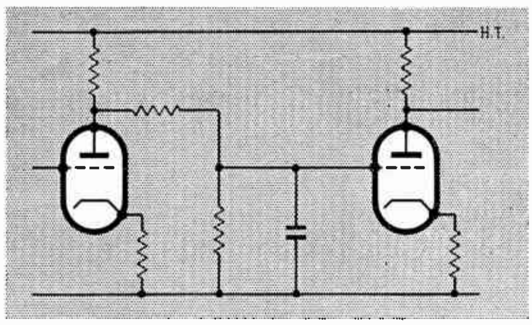

Fig. 7. *Practical form of Fig. 6*

current is a single-valued function of $v_g + v_a/\mu$, say $i_a = f(v_g + v_a/\mu)$. Since i_a is single-valued its integral with respect to $v_g + v_r/\mu$, taken over a complete cycle of oscillation, is zero.

Thus

$$\oint i_a\, d(v_g + v_a/\mu) = \oint i_a\,(dv_g/dt + dv_a/\mu dt)dt = 0$$

It is obvious that the constant components of the current and voltages contribute nothing to the integral. Hence the expressions for the alternating components i_{1a}, v_{1a} and v_{1g} may be substituted for i_a, v_a and v_g.

When this is done and the integrations carried out the result is

$$\sum_{-\infty}^{\infty} nI_n^2(Z_{tn}\exp(jn\phi) + Z_{in}/\mu) = 0$$

When the sign of n is changed Z_{tn} and Z_{in} change to their complex conjugate values. Hence if

$$Z_{tn} = R_{tn} + jX_{tn} \text{ and } Z_{in} = R_{in} + jX_{in}$$

the equation can be written

$$\sum_{1}^{\infty} nI_n^2 (R_{tn}\sin(n\phi) + X_{tn}\cos(n\phi) + X_{in}/\mu) = 0 \quad (3.4)$$

Since R_{tn}, X_{tn} and X_{in} are known functions of frequency the equation provides a means of calculating

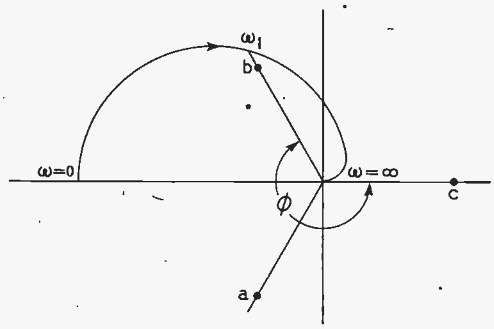

Fig. 8. *Locus of stage gain A for circuit of Fig. 6*

ω_0 when the magnitudes of the harmonic currents are known.

Usually it is more convenient to express the result in terms of the harmonic grid voltages. The relation between i_{1a} and v_{2g} may be written as

$$i_{1a} = -Y_t v_{2g} = -(G_t + jB_t)v_{2g}$$

where $Y_t = 1/Z_t$ is the open-circuit transfer admittance.

Hence if $v_{1g} = \sum_{1}^{\infty} V_n \cos(n\omega_0 t + \alpha_n)$ then

$I_n = |Y_{tn}|V_n$ and (3.4) becomes

$$\sum_{1}^{\infty} nV_n^2(G_{tn}\sin(n\phi) - B_{tn}\cos(n\phi) + |Y_{tn}|^2 X_{in}/\mu) = 0 \quad (3.5)$$

where G_{tn} and B_{tn} are the values of G_t and B_t at the frequencies $n\omega_0$.

In the absence of harmonics the oscillation frequency ω_1 is given by

$$G_{t1}\sin\phi - B_{t1}\cos\phi + |Y_{t1}|^2 X_{i1}/\mu = 0 \quad .. \quad (3.6)$$

These results are now applied to some of the circuit arrangements discussed in section 2. When the capacitances C_a and C_g in Fig. 4 are neglected, elementary

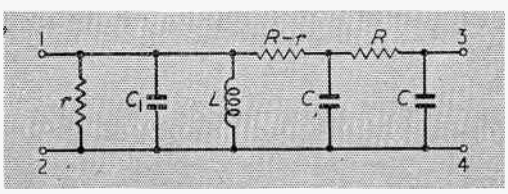

Fig. 9. *Coupling network for 4-phase oscillator*

analysis gives the transfer conductance and susceptance of the network at a frequency $n\omega_0$ as

$$\left.\begin{aligned} G_{tn} &= (r + R)/rR, \quad B_{tn} = -1/n\omega_0 CrR \\ \text{and} \quad |Y_{tn}|^2 X_{in} &= -1/n\omega_0 CR^2 \end{aligned}\right\} \quad (3.7)$$

In the absence of harmonics the frequency is given by (3.6) as

$$\omega_1 C(r + R) = \frac{(r/\mu R - \cos\phi)}{\sin\phi} \quad .. \quad .. \quad (3.8)$$

Hence and from (3.5) the formula for ω_0 is

$$\frac{\omega_0}{\omega_1} = \frac{\sum_{1}^{\infty} V_n^2(r/\mu R - \cos(n\phi))\sin\phi}{\sum_{1}^{\infty} nV_n^2(r/\mu R - \cos\phi)\sin(n\phi)} \quad .. \quad (3.9)$$

For 3-phase operation corresponding to point 'a' in Fig. 5 $\phi = 2\pi/3$. If also μ is large and the harmonic amplitudes small compared with the fundamental (3.9) can be expanded in the form

$$\omega_0/\omega_1 = 1 + 3(V_2/V_1)^2 - 2(V_3/V_1)^2 + \quad .. \quad (3.10)$$

Similarly for the network of Fig. 6 (with C_a neglected) the frequency without harmonics is

$$\omega_1 C(r + R) = \frac{\sin\phi}{r/\mu(r + R) + \cos\phi} \quad .. \quad (3.11)$$

and the frequency ratio is

$$\frac{\omega_0}{\omega_1} = \frac{\sum_{1}^{\infty} nV_n^2(r/\mu(r + R) + \cos\phi)\sin(n\phi)}{\sum_{1}^{\infty} n^2V_n^2(r/\mu(r + R) + \cos(n\phi))\sin\phi} \quad (3.12)$$

For 3-phase operation corresponding to point 'b' of Fig. 7 $\phi = 4\pi/3$. If also μ is large and the harmonic amplitudes are small (3.12) becomes

$$\omega_0/\omega_1 = 1 - 6(V_2/V_1)^2 + 18(V_3/V_1)^2 + \quad .. \quad (3.13)$$

From an inspection of these results it might be concluded that the first arrangement (shunt-resistance network) is more stable than the second (shunt-capacitance network). This conclusion, however, would not be justified, for the amplitudes of the harmonic voltages are not the same in the two cases. Also, the fact that the frequency is changed in different directions by the second and third harmonics makes it difficult to draw any general conclusions. In order to decide the point, an analysis taking explicit account of the nonlinear behaviour of the amplifier is required. The utility of the above formulae lies in the fact that it is often comparatively simple to calculate approximately the harmonic amplitudes from the nonlinear theory when the frequency shift is neglected.

Single-phase oscillators may be treated similarly. In this case $\phi = 0$ and (3.5) becomes Groszkowski's formula except for the difference in notation. When $\mu = \infty$ this takes the form

$$\sum_{1}^{\infty} nV_n^2 B_{tn} = 0$$

It is a straightforward matter to show that for the shunt-capacitance type of oscillator (Fig. 2) with arbitrary values of resistance and capacitance, the transfer susceptance is

$$rB_{tn} = n\omega_0(\omega_1^2 - n^2\omega_0^2)C_1C_2C_3(R_1 + r)R_2R_3$$

where

$$\omega_1^2 = \frac{(R_1+r)\ (C_1 + C_2+C_3) + R_2\,(C_2+C_3) + R_3C_3}{(R_1 + r)\ R_2R_3C_1C_2C_3}$$

The frequency is given by

$$(\omega_0/\omega_1)^2 = (\sum_1^{\infty} n^2 V_n^2)/(\sum_1^{\infty} n^4 V_n^2)$$

If the harmonic amplitudes are small

$$\omega_0/\omega_1 = 1 - 6(V_2/V_1)^2 - 36(V_3/V_1)^2 - \ldots \quad (3.14)$$

Similarly for the shunt-resistance type of circuit obtained by interchanging the positions of the resistors and capacitors in Fig. 2 the frequency is given by

$$(\omega_0/\omega_1)^2 = (\sum_1^{\infty} V_n^2/n^2)/(\sum_1^{\infty} V_n^2)$$

If the harmonic amplitudes are small

$$\omega_0/\omega_1 = 1 - (3/8)\ (V_2/V_1)^2 - (4/9\ (V_3/V_1)^2 \ldots \quad (3.15)$$

These formulae hold also for the cathode-follower-coupled oscillator of Fig. 12 and its derivatives, provided the cathode-follower stages operate linearly and all the harmonics are generated in the final amplifier.

Again no useful comparison can be made between the various types of single-phase oscillator or between single-phase and polyphase oscillators until the actual values of the harmonic voltages are known.

4. Nonlinear Analysis

The nonlinear analysis of phase-shift oscillators is greatly simplified by the fact that resistance-capacitance coupling networks have very poor frequency discrimination. Consequently, if the harmonic voltages are to be small the harmonic currents generated by the amplifier must also be small. On the assumption that the oscillation amplitude is limited by the nonlinear characteristic of the amplifier it follows that the small-signal gain of the amplifier can be only slightly greater than that required to initiate oscillation. It is, of course, possible to construct oscillators in which the gain is limited by other means, but these are excluded from the present discussion.

Since only comparative results are required, it is sufficient to consider the simplified case where $\mu = \infty$. The relation between anode current and grid voltage will be taken as

$$i_a = a_0 + a_1 v_g + a_2 v_g^2 + a_3 v_g^3 + \quad \ldots \quad \ldots \quad (4.1)$$

Although the oscillation amplitude is limited mainly by the odd-power terms the harmonics and modulation products arising from these terms may be small since the regeneration is near to the critical value. In these circumstances, and with the usual three-halves law amplifier characteristic, the relative magnitudes of the coefficients in (4.1) are such that only the term $a_2 v_g^2$ is effective in producing harmonics, and the only harmonic of importance is the second. The alternating grid voltage and anode current of valve 1 are then

$$v_{1g} = V_1 \cos(\omega_0 t) + V_2 \cos(2\omega_0 t + \theta) \quad \ldots \quad (4.2)$$

$$\begin{aligned} i_{1a} &= a_1 v_{1g} + a_2 v_{1g}^2 \\ &= a_1 V_1 \cos(\omega_0 t) + a_2 V_1 V_2 \cos(\omega_0 t + \theta) \\ &\quad + \tfrac{1}{2} a_2 V_1^2 \cos(2\omega_0 t) + a_1 V_2 \cos(2\omega_0 t + \theta) \quad (4.3) \end{aligned}$$

Harmonics higher than the second are neglected.

Since the expressions for the voltages and currents are real it is convenient to write the transfer impedance in its operational form, $Z_t(D)$, where $D = d/dt$. The grid voltage at the second amplifier in Fig. 3 is then

$$v_{2g} = -Z_t(D)\ i_{1a} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (4.4)$$

A 3-phase oscillator with coupling networks of the type shown in Fig. 4 is first considered. For the present purpose, the capacitances C_a and C_g may be neglected and there is also no loss of generality in supposing that $r \ll R$. (For a more exact analysis it is only necessary to replace R by $R + r$ in the equations which follow.) With this network $\phi = 2\pi/3$ and the transfer impedance is

$$Z_t(D) = r/(1 + 1/DCR) \quad \ldots \quad \ldots \quad \ldots \quad (4.5)$$

In the steady state the relation between consecutive grid voltages is given by (3.1) and (3.2). Hence and from (4.2),

$$v_{2g} = V_1 \cos(\omega_0 t - 2\pi/3) + V_2 \cos(2\omega_0 t - 4\pi/3 + \theta)$$

From (4.4) and (4.5)

$$\begin{aligned} -r i_{1a} = {} & V_1 \cos(\omega_0 t - 2\pi/3) + (V_1/\omega_0 CR) \sin(\omega_0 t - 2\pi/3) \\ & + V_2 \cos(2\omega_0 t - 4\pi/3 + \theta) + (V_2/2\omega_0 CR) \sin \\ & (2\omega_0 t - 4\pi/3 + \theta) \end{aligned}$$

i_{1a} is next eliminated between this expression and (4.3)

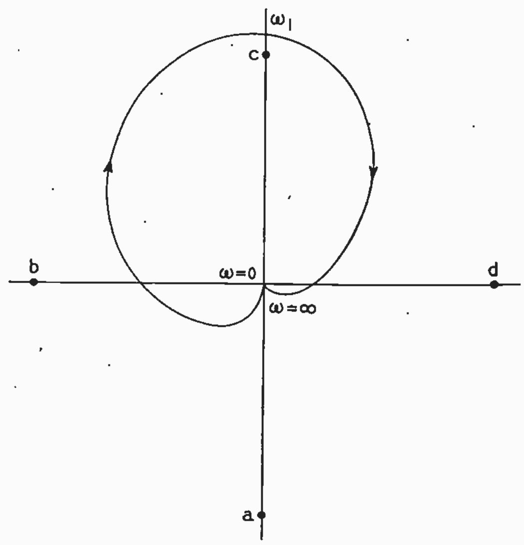

Fig. 10. *Locus of stage gain A for circuit of Fig. 9*

and the sine and cosine terms of frequency ω_0 and $2\omega_0$ are equated separately to give four equations:

$$\begin{aligned} &\tfrac{1}{2}3^{1/2} - \tfrac{1}{2}/\omega_0 CR = r a_2 V_2 \sin\theta \\ &\tfrac{1}{2} + \tfrac{1}{2}3^{1/2}/\omega_0 CR = r a_1 + r a_2 V_2 \cos\theta \\ &(1 - 2 r a_1 - \tfrac{1}{2}3^{1/2}/\omega_0 CR) \tan\theta = 3^{1/2} + \tfrac{1}{2}/\omega_0 CR \\ &(1 - \tfrac{1}{2}3^{1/2}/\omega_0 CR - 2 r a_1) \cos\theta + \\ &\qquad (3^{1/2} + \tfrac{1}{2}/\omega_0 CR) \sin\theta = r a_2 V_1^2/V_2 \end{aligned}$$

When the harmonic voltage is zero the first two equations give the frequency and the condition for critical regeneration as $\omega_1 = 1/3^{1/2}CR$ and $ra_1 = 2$. If it is assumed that ra_1 is only slightly greater than 2, these approximations can be inserted into the last two equations to give $\theta = 5\pi/6$ and

$$V_2 = r a_2 V_1^2/3^{3/2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (4.6)$$

These results may be substituted back into the first equation to obtain a more accurate value for the frequency. Thus

$\omega_1/\omega_0 = 1 - (ra_2V_1)^2/9$. Substituting for ra_2 according to (4.6) gives

$$\begin{aligned}\omega_0/\omega_1 &= 1 + (ra_2V_1)^2/9 + \ldots \\ &= 1 + 3(V_2/V_1)^2 + \ldots \qquad \ldots \quad \ldots \quad (4.7)\end{aligned}$$

This is in agreement with (3.10) under the assumed conditions.

A similar analysis can be carried out for an oscillator using the shunt-capacitance type of coupling network shown in Fig. 6. When C_a is neglected, the transfer impedance is $Z_t(D) = r/(1 + DCR)$, and for 3-phase operation $\phi = 4\pi/3$. The analysis proceeds on the lines above with the results:

$$\omega_1 = 3^{1/2}/CR,\ \theta = 7\pi/6,\ V_2 = ra_2V_1^2/3^{1/2}6 \qquad (4.8)$$

$$\omega_0/\omega_1 = 1-(ra_2V_1)^2/18 - \ldots = 1-6(V_2/V_1)^2- \qquad (4.9)$$

The last equation is in agreement with (3.15).

If the valve parameters a_1, a_2 and a_3 and the load resistance r are the same for both types of oscillator then the value of V_1 is also the same. A comparison of (4.6), (4.7), (4.8), and (4.9) then shows that in the shunt-capacitance oscillator the harmonic grid voltage is one-half and the frequency shift also one-half of the values for the shunt-resistance type. However, the output would usually be taken from the anodes rather than the grids of the amplifiers. The second-harmonic anode voltage in the shunt-capacitance type is greater than in the shunt-resistance type in the ratio $(13/7)^{1/2}$.

Single-phase oscillators may be treated similarly. In the cathode-follower circuit shown in Fig. 12, the relation between grid voltage and anode current in the amplifier stage is $\quad -i_ar_1 = (1 + DCR)^3v_g$ where r_1 is an equivalent load resistance, and R is assumed large compared with the cathode resistors in the coupling stages. The nonlinear relation between i_a and v_g is again taken as the parabolic equation (4.3). The results are:

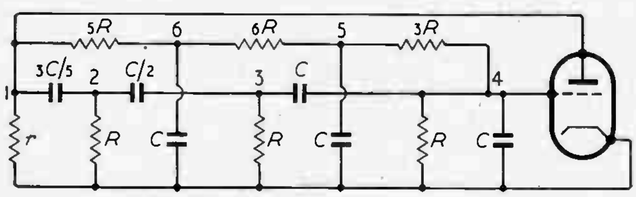

Fig. 11. *A.C. equivalent circuit of composite coupling network*

$$V_2 = ra_2V_1^2/18(21)^{1/2} \qquad \ldots \quad \ldots \quad \ldots \quad (4.10)$$

$$\omega_0/\omega_1 = 1-(ra_2V_1)^2/1134 = 1-6(V_2/V_1)^2 \qquad \ldots \quad (4.11)$$

Similarly, for the shunt-resistance form of circuit obtained by interchanging the positions of R and C in Fig. 12,

$$V_2 = 4ra_2V_1^2/9(39)^{1/2} \qquad \ldots \quad \ldots \quad \ldots \quad (4.12)$$

$$\omega_0/\omega_1 = 1-2(ra_2V_1)^2/1053 = 1-(3/8)(V_2/V_1)^2 \qquad (4.13)$$

Equations (4.11) and (4.13) are in agreement with (3.14) and (3.15).

For the same magnitude of fundamental grid voltage the frequency shift in the shunt-capacitance oscillator is about 0·5 and the second-harmonic grid voltage about

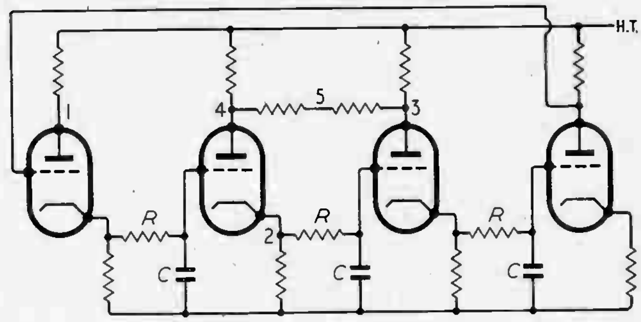

Fig. 12. *Cathode-follower-coupled oscillator*

0·17 of the corresponding magnitudes for the shunt-resistance type. In comparison with the corresponding forms of symmetrical 3-phase oscillators the frequency stability of the single-phase oscillators with respect to second harmonic is about 60 times better for the same magnitude of fundamental voltage at the amplifier grid.

Smiley[4] has described a very-low-frequency 3-phase oscillator of symmetrical form in which the oscillation amplitude is limited by symmetrical peak-clipping. The equivalent nonlinear amplifier characteristic could be crudely represented by the expression $i_a = a_1v_g + a_3v_g^3$ and the important harmonic would be the third. Analyses similar to those for the parabolic characteristic can be carried out, and although the results differ somewhat from those above, the main conclusion is upheld: namely, the great superiority of single-phase oscillator in respect of frequency stability.

5. Build-Up of Oscillation

A study of the way in which the oscillation builds up to the steady-state regime has both practical importance and theoretical interest. If the build-up rate, i.e., the fractional increase in amplitude per cycle of oscillation, is small, an oscillator for very-low frequencies may not be ready for use until some considerable time after it is switched on. From the theoretical standpoint the resistance-capacitance oscillator shows in a striking way the importance of the frequency change which accompanies amplitude change. This relationship is not peculiar to resistance-capacitance oscillators but exists in any oscillator where the frequency-response characteristic of the coupling network is asymmetrical with respect to the oscillation frequency. In resistance-capacitance oscillators the asymmetry is very marked.

In order to obtain a manageable solution certain simplifying assumptions are made; in particular, the effects of harmonic voltages are neglected. During the build-up period the oscillation is then a sinusoidal wave of slowly-varying amplitude and frequency. Depending on the initial conditions there may also exist other monotonic or oscillatory disturbances of exponentially decreasing amplitude. These, however, rapidly become negligible compared with the main oscillation and are therefore neglected. Attention is confined to 3-phase and single-phase oscillators.

Let the grid voltage in one section of a 3-phase oscillator be $v_{1g} = V\cos\theta$ where V and θ are functions of time. If μ is very large the anode current depends only on v_{1g}, thus $i_{1a} = f(v_{1g})$. If harmonics are neglected the current is $i_{1a} = I \cos \theta$, where, by Fourier's theorem,

$$I = F(V) = (2/\pi) \int_0^\pi f(V \cos x) \cos x \, dx \quad .. \quad (5.1)$$

The grid voltage in the next section of the oscillator is $v_{2g} = - Z_t(D) \, i_{1a}$ where $D = d/dt$. It was shown in section 1 that in a 3-phase oscillator the phase displacement between successive grid voltages is $2\pi/3$ or $4\pi/3$; i.e., $\pm 2\pi/3$. Hence $v_{2g} = V \cos (\theta \pm 2\pi/3)$. When the expressions for v_{2g} and i_{1a} derived above are substituted, an equation for V and θ is obtained.

$$V \cos (\theta \pm 2\pi/3) = -Z_t(D) \, [F(V) \cos \theta] \quad .. \quad (5.2)$$

In order to proceed, it is necessary to consider particular forms of the functions $Z_t(D)$ and $F(V)$. If the relation between anode current and grid voltage can be expressed by the first four terms of the power series (4.1) the function $F(V)$ defined by (5.1) is

$$F(V) = a_1 V + 3a_3 V^3/4 \quad .. \quad .. \quad .. \quad (5.3)$$

When the stray capacitances are neglected and when $r \ll R$ the network shown in Fig. 4 has a transfer impedance $Z_t(D) = r/(1 + 1/DCR)$. With this network the minus sign must be taken with the phase shift in the expression for v_{2g}. Equation (5.2) can then be written in the form

$$(1 + DCR) \, V \cos (\theta - 2\pi/3) = -rDCR \, ((a_1 V + 3a_3 V^3/4) \cos \theta) \quad (5.4)$$

In the steady state V and $d\theta/dt$ are constants with values V_1 and ω_1 which are easily shown to be $V_1^2 = -4 \, (ra_1 - 2)/3ra_3$ and $\omega_1 = 1/CR3^{1/2}$. The condition for the oscillation to build up from a small amplitude is that $ra_1 > 2$ and, for a final stable amplitude to be possible, a_3 must be negative. Let $\frac{1}{2}g$ be the fractional amount by which the gain exceeds the minimum value necessary for the oscillation to build up; i.e., let $ra_1 = 2(1 + \frac{1}{2}g)$ or $g = ra_1 - 2$.

When the operations of differentiation have been carried out in Equation (5.4) two relations between amplitude and frequency are obtained by equating separately the sine and cosine terms. These are:

$$\omega(3 + 2g - 2gV^2/V_1^2) = 3\omega_1 + 3^{1/2}V'/V \quad .. \quad (5.5)$$

$$V'(3 + 2g - 6gV^2/V_1^2) = 3^{1/2}V(\omega_1 - \omega) \quad .. \quad (5.6)$$

in which $V' = dV/dt$ and $\omega = d\theta/dt$, and g, ω_1 and V_1 have the values indicated above.

Equation (5.6) shows how the deviation of the instantaneous frequency ω from its final value ω_1 depends on the rate of change of amplitude. ω can be eliminated between (5.5) and (5.6) to obtain an equation for V which can be written

$$\left[\frac{1 + g + g^2/3}{V} + \frac{(1 - g)V}{V_1^2 - V^2} - \frac{g^2 V}{V_1^2}\right] \frac{dV}{dt} = \tfrac{1}{2}g\omega_1/3^{1/2}$$

The integral is

$$(1 + g + g^2/3) \log_e (V^2/V_1^2) - (1 - g) \log_e (1 - V^2/V_1^2) - g^2 V^2/V_1^2 = g\omega_1(t - t_0)/3^{1/2} \quad (5.7)$$

where t_0 is an arbitrary constant.

V/V_1 cannot be expressed as an explicit function of t, but it is not difficult to show that for values of g less than 3 (in practice, g would be less than 1) V/V_1 increases monotonically with t. The build-up time may be arbitrarily defined as the time taken for the amplitude to increase from 10% to 90% of its final value. This time T_b can be calculated from (5.7) and is given by

$$\omega_1 T_b = 10{\cdot}5/g + 4{\cdot}8 + 1{\cdot}2g$$

The number of cycles of oscillation over which the build-up takes place is $(1/2\pi) \int \omega dt$. From (5.6) and the foregoing expression for T_b this number is

$$n = 1{\cdot}67/g + 0{\cdot}15$$

When V/V_1 is small V can be expressed approximately as $V = V_0 \exp (k\omega_1 t)$ where

$$k = \frac{\frac{1}{2}g}{3^{1/2} \, (1 + g + g^2/3)}$$

From (5.6) the initial frequency is

$$\omega/\omega_1 = 1 - k(3 + 2g)/3^{1/2}$$

When the rate of increase of amplitude is small, ω is close to ω_1. It might therefore be assumed, as van der Mark and van der Pol assumed, that in such circumstances it would be permissible to put $\omega = \omega_1$. However, if this is done Equ. (5.5) becomes

$2\omega_1 g(1 - V^2/V_1^2) = 3^{1/2}V'/V$ of which the integral is $\log_e (V^2/V_1^2) - \log_e (1 - V^2/V_1^2) = 4 \, g\omega_1 \, (t - t_0)/3^{1/2}$.

When V/V_1 is small

$$V = V_0 \exp (2g\omega_1 t/3^{1/2})$$

By comparing this result with (5.8), it can be seen that the rate of amplitude build-up given by the van der Mark–van der Pol theory is about four times the true value. The physical explanation of this discrepancy is to be found in the frequency shift. During the build-up period the frequency is less than the steady-state value. Hence the attenuation in the coupling network (Fig. 4) is increased. The excess gain and so also the rate-of-

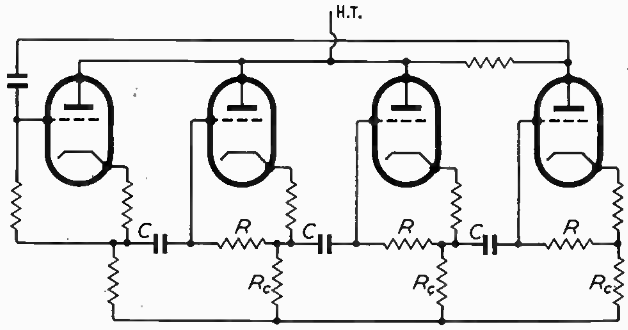

Fig. 14. *Single-phase oscillator with cathode feedback*

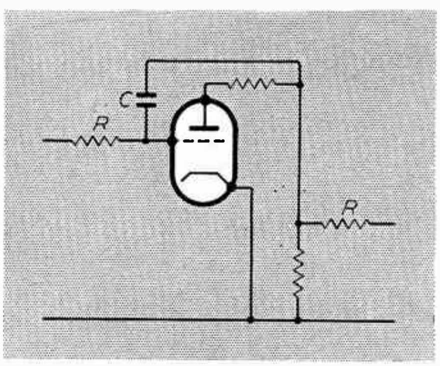

Fig. 13. *Section of polyphase oscillator with anode feedback*

amplitude increase is then less than if the frequency remained constant.

An analysis similar to that above can be carried out for the shunt-capacitance type of coupling network shown in Fig. 6. When stray capacitance is neglected and $r \ll R$ the operational equation is

$$(1 + DCR)V\cos(\theta + 2\pi/3) = -r(a_1V + 3a_3V^3/4)\cos\theta.$$

The equations corresponding to (5.5) and (5.6) are

$$\omega_1(3 + 2g - 2gV^2/V_1^2) = 3\omega + 3^{1/2}V'/V$$

and $$3^{1/2}V' = V(\omega - \omega_1)$$

where now $\omega_1 = 3^{1/2}/CR$, and the other symbols have the same meaning as before. The solution for V is

$$\log_e(V^2/V_1^2) - \log_e(1 - V^2/V_1^2) = g\omega_1(t - t_0)/3^{1/2}$$

The build-up time is given by

$$\omega_1 T_b = 10\cdot5/g$$

For small values of g this is the same as for the shunt-resistance type of coupling network.

Single-phase oscillators can be treated similarly. With the circuit shown in Fig. 12 and for small values of V/V_1

$$V = V_0 \exp(k\omega_1 t)$$

where $k = [(1 + \frac{1}{2}g)^{1/3} - 1]/3^{1/2}$ and $\frac{1}{2}g$ is the excess gain measured as a fraction of the critical value. When g is small $k = g/3^{1/2} \times 6$ and the initial rate of build-up is therefore one-third of that for a 3-phase oscillator. A similar result is found for the shunt-resistance type of oscillator.

In deriving the results of this section it has been assumed that the working point on the nonlinear valve characteristic remains fixed as the oscillation builds up. In practice, this working point might be varied by an automatic grid-bias control circuit, and a similar effect could be produced by thermally-sensitive control elements. The initial gain might then be high and would decrease as the oscillation built up. In such arrangements the build-up times would be somewhat greater than the above estimates.

6. Conclusions

In comparison with the single-phase oscillator the symmetrical type of polyphase oscillator is inferior in respect of frequency stability and purity of waveform. The existence of unwanted modes of oscillation in some forms of polyphase oscillator is a disadvantage which is almost entirely avoided in the single-phase oscillator. Single-phase oscillators are equally suitable for the production of polyphase supplies without restriction on the number of phases, whereas the symmetrical polyphase oscillator requires a rather special type of coupling network for even-phase operation.

In favour of the polyphase oscillator is the fact that the h.t. supply contains (ideally) no alternating component, but this is of little importance except at the very lowest frequencies. The build-up time is also much less than for the single-phase oscillator.

The build-up time for the common form of 3-phase oscillator has been shown to be much greater than was formerly supposed and the importance of the frequency change which takes place during the build-up process has been demonstrated.

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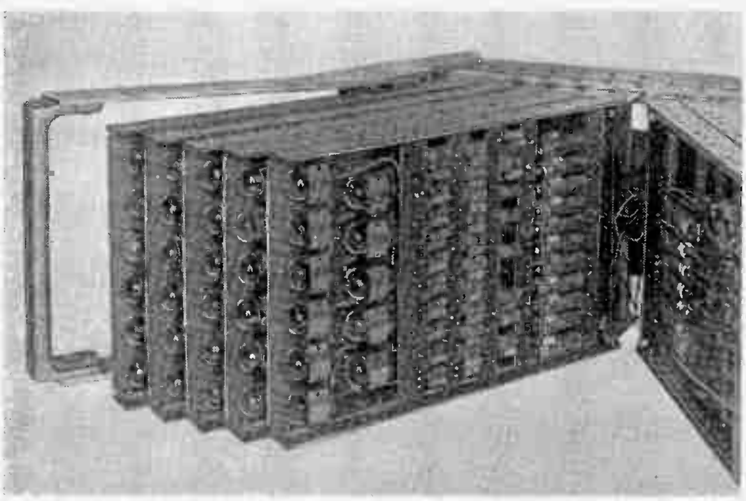
ELECTRONICS AND THE POST OFFICE

The photograph illustrates a constructional technique employed in the G.P.O.'s new Group Routing and Charging Equipment (GRACE), which performs automatically the functions of a trunk telephone operator in setting up a call. The part of GRACE shown is a register for storing dialled digits. Cold-cathode tubes form the memory elements. The unit shown consists of a number of chassis which can be opened like the leaves of a book to provide access for servicing and testing.



A somewhat similar equipment developed by Automatic Telephone & Electric Co. was recently put into service in London. This is an electronic director for routing calls between exchanges in a large city. A magnetic drum is employed for data storage. This equipment may, in time, replace electromechanical directors.

The Post Office is also investigating the possibilities of automatic letter sorting. A semi-automatic machine, which enables an operator to sort about 3,000 letters an hour into 144 pigeon-holes, compared with 1,500 into 48 pigeon-holes by manual methods, is already in use. In order to replace the operator by a machine, it will be necessary to code addresses in some way. A five-unit code, giving information about the town of destination and street of destination, has been proposed. The success of such a system depends on the degree of co-operation which can be obtained from the public, and this will be assessed before further work proceeds. An exchange of information between Britain, Canada and the U.S.A. has been arranged.

Numerical Extraction of Roots

Suppose we require to find the nth root of a given number N, and that a is an approximation, so that $\sqrt[n]{N} = a + x$, where x is small compared to a. Then a better approximation is

$$b = \frac{1}{n}\left[(n-1)\,a + \frac{N}{a^{n-1}}\right] \quad \ldots \quad \ldots \quad (1)$$

which can easily be evaluated with or without a calculating machine.

For example, if we start with $1{\cdot}4$ as an approximation to $\sqrt{2}$, so that in (1), $a = 1{\cdot}4$, $n = 2$, $N = 2$, we have

$$b_1 = \frac{1}{2}\left[1{\cdot}4 + \frac{2}{1{\cdot}4}\right]$$

$$= \frac{1}{2}\,(1{\cdot}4 + 1{\cdot}428571\ldots) = 1{\cdot}41429$$

which agrees with the value $1{\cdot}4142$ given in five-figure tables with an error of only 1 in the fifth significant figure. If we now put $1{\cdot}41429$ for a in (1), we find

$$b_2 = \frac{1}{2}\left[1{\cdot}41429 + \frac{2}{1{\cdot}41429}\right]$$

$$= \frac{1}{2}\,(1{\cdot}41429 + 1{\cdot}41413713)$$

$$= 1{\cdot}414213565$$

and $\quad b^2_2 = 2{\cdot}000000007$!

Again, supposing we require $\sqrt[6]{800}$, and start with the approximation $a_1 = 3$ $(3^6 = 729)$. We find from (1), with $n = 6$, $a = 3$, $N = 800$

$$b_1 = \frac{1}{6}\left[5 \times 3 + \frac{800}{243}\right] = \frac{1}{6}\,[18{\cdot}29218]$$

$$= 3{\cdot}048697$$

Returning to (1) with $a = 3{\cdot}048697$, we find

$$a^6 = 802{\cdot}944596$$

$$a^5 = 263{\cdot}373040$$

$$b_2 = \frac{1}{6}\,[15{\cdot}243485 + 3{\cdot}037517] = 3{\cdot}046834$$

and $b^6_2 = 800{\cdot}0051$

so that b^6_2 is in error by about 1 part in 160,000, and b_2 by about 1/6 of this or 1 in 10^6; our starting approximation was quite crude, and an approximation as good can be obtained by means of a table of higher powers of numbers up to 100. This is given, for example, in Attwood's 5-figure tables, p. 52.

To prove (1), we have

$$N^{1/n} = a + x$$

$$\therefore \frac{N}{a^{n-1}} = a\left[1 + \frac{x}{a}\right]^n = a + nx + \frac{1}{2}n(n-1)\frac{x^2}{a} + \ldots \quad (2)$$

$$\therefore \frac{1}{n}\left[(n-1)\,a + \frac{N}{a^{n-1}}\right] = (a+x) + \frac{n-1}{2}\cdot\frac{x^2}{a} + \ldots \quad (3)$$

so that the absolute error is about

$$\frac{1}{2}\,(n-1)\,\frac{x^2}{a}$$

and is an over-estimate for $n > 1$.

Note that all we are doing is to take the first three terms of a binomial series (2) in which the quantity x/a is small; we are not therefore concerned with whether n is an integer or not, or even whether n is positive or not. We could therefore obtain

$$N^{2/3} \text{ by putting } n = \frac{3}{2} \text{ in (1)}$$

$$N^{-1/4} \text{ by putting } n = -4 \text{ in (1)}$$

and so on. Clearly it is an advantage to start with a good approximation for a, but using a crude approximation only means that we have to apply (1) once, or at worst twice, more often to obtain a satisfactory approximation.

As a rough general rule, it may be stated that the approximation b given by (1) is correct to twice as many significant figures as was the original approximation a. For example, to calculate $\pi^{2/3}$, we have $\pi^2 = 9{\cdot}8696$, $2{\cdot}1^3 = 9{\cdot}261$, $2{\cdot}2^3 = 10{\cdot}648$, so $\pi^{2/3}$ or $\sqrt[3]{\pi^2}$ is about $2{\cdot}14$, and this value will do for a. Since $n = 1{\cdot}5$,

$$b_1 = \frac{2}{3}\left[\frac{2{\cdot}14}{2} + \frac{3{\cdot}14159265}{\sqrt{2{\cdot}14}}\right]$$

We can obtain $\sqrt{2{\cdot}14}$ by applying (1) again, with $n = 2$, $N = 2{\cdot}14$, and a can be obtained from tables or slide rule as say $1{\cdot}46$. Our second approximation to $\sqrt{2{\cdot}14}$ is thus $1{\cdot}462877$, the square of which is $2{\cdot}1400091$, and if $1{\cdot}462877$ is used for $\sqrt{2{\cdot}14}$ in b_1, we find $b_1 = 2{\cdot}145029$, which is correct to six places. Alternatively, we could have calculated $\pi^{1/3}$ from (1) with $N = \pi$, $n = 3$ and $a = 1{\cdot}47$ [since $1{\cdot}4^3 = 2{\cdot}744$ and $1{\cdot}5^3 = 3{\cdot}375$, $\pi^{1/3}$ or $\sqrt[3]{\pi}$ must be about $1{\cdot}47$] and squared the result, but in this case putting $n = 1{\cdot}5$ as above appears to be superior.

To obtain $e^{-1/4}$, we note that $1{\cdot}2^4 = 2{\cdot}0736$ and $1{\cdot}3^4 = 2{\cdot}8561$, so that $e^{1/4}$ is about $1{\cdot}28$, and we can take a to be $1/1{\cdot}28 = 0{\cdot}781$. This gives

$$b_1 = -\frac{1}{4}\left[-5 \times 0{\cdot}781 + 2{\cdot}71828183 \times (0{\cdot}781)^5\right]$$

and since $(0{\cdot}781)^5 = 0{\cdot}29057294$, $b_1 = 0{\cdot}778785$ whereas the correct value of $e^{-1/4}$ is $0{\cdot}778801$.

Colour Printing

ELECTRONICS APPLIED TO BLOCK-MAKING

The essential features of the commonest type of colour-printing processes are straightforward. The coloured original is photographed through optical filters to produce three colour-separation negatives, one for each of the three ink colours normally employed, yellow, red (magenta) and blue (cyan) and an additional negative (black) for the greys and blacks when the more commonly used four-colour process is involved. When the colour printing is confined to three printings only, no black is used and the darker areas of the original are strengthened by the use of a deeper shade of blue ink.

During the making of the colour-separation negatives, the image, after passing through the normal process camera lens and colour filter, which is situated immediately behind the lens, has also to pass through a 'screen' before reaching the sensitized negative plate.

This screen consists of two sheets of plate glass each ruled diagonally across the whole surface with very fine lines engraved in the glass and filled with a black pigment. The lines on each sheet are ruled at right angles to each other and placed face to face to form a complete lattice-work pattern. The screens are ruled mathematically correct to a specified number of lines and for colour work this is usually 120 or 133 lines per inch.

The effect of the interposing of the screen is to break up the image into a series of dots, and where there is plenty of highlight reflection passing through the screen that portion of the negative will be dark with only minute white spots to be seen. Alternatively, where little light is reflected from the dark portions of the image the negative will be almost clear with only very small black dots visible.

Intermediate tones will show dots of varying sizes in relation to the lights and shades of the image.

The negative after development and fixing is used to print down the screened image on to a metal surface, usually copper which has been coated with a solution sensitive to light. This print on metal has the image hardened on its surface by heat treatment to make it resistant to the acid etching process which follows. The etching bites into the metal surface where there are clear portions of the image leaving untouched only the dots in their varying sizes according to the shade densities.

The etch is carried to sufficient depth to enable the resulting half-tone printing block, when inked in the normal way on a printing machine, to reproduce on paper only the 'dots'.

Each of the colour-separated negatives is treated in this manner, and when it is explained that the 'screens' for colour work are arranged at different angles for each of the colours to be used it will be understood how the resulting superimposition of each printing can build up a colour-separated pattern on the printing paper which will reflect to some extent the colour values of the original picture.

Colour printing is a subtractive process. The illuminant consists of nominally white light. Colours are produced by removing parts of the spectrum, the function of the ink being to act as a filter which absorbs light rays selectively. The light is transmitted through the ink, reflected from the underlying white paper, and passes through the ink a second time before it reaches the eye. The inks available are band-pass filters. To produce yellow, it is necessary to transmit green and red and absorb blue. To produce magenta, red and blue must be transmitted but not green; and for cyan, blue and green must be transmitted, and red absorbed. The ideal transmission characteristics are shown in Fig. 1 alongside those of standard printing inks.

The departure of standard inks from the ideal is the root of nearly all the difficulties encountered in colour printing. Consider the process of making a set of colour-separation negatives. Let us suppose, for simplicity, that the original picture contains an area of magenta with exactly the same spectral characteristics as the ink which will be used in the final printing process. It is obvious that the magenta colour-separation negative must show colour in the appropriate area and nowhere else. To produce the negative, the original is photographed through a green filter. Reference to Fig. 1

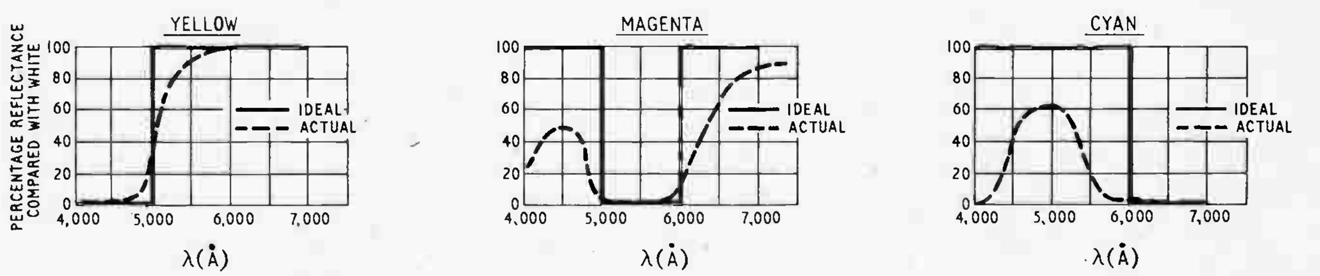


Fig. 1. *Ideal spectral responses of printing inks compared with actual responses (shown dashed)*

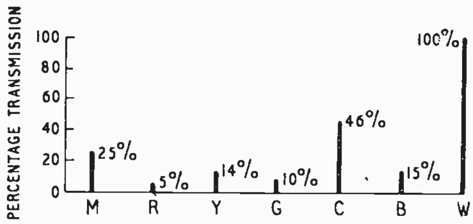


Fig. 2. *Response of filter used in making a yellow plate to various colours*

shows that, in fact, magenta ink absorbs nearly all the green light. The resulting magenta colour-separation negative will therefore show colour in the right places. Now consider what happens when the yellow negative is made. This should, of course, show no colour. A blue filter is employed, since yellow is the only ink colour with no blue in its make-up, this fact serving as a distinguishing feature. If the magenta ink were perfect, all the blue light would be reflected, and all would be well. In fact, as can be seen from Fig. 1, only about half of the blue light is reflected. This means that the yellow colour-separation photograph will register about 50% yellow in magenta areas, and yellow will eventually be printed in these areas, thereby degrading the magenta. Similarly, when the cyan colour-separation negative is made with the aid of a red filter, the imperfect reflection of red by the magenta will cause a certain amount of cyan to be registered in magenta areas. Thus, our hypothetical original, which had only magenta, will be reproduced as a mixture of magenta, yellow and cyan. The result will be a rather dull reddish colour bearing little resemblance to the original.

In order to obtain proper colour rendering, a process of colour correction is necessary. If the colour blocks are made by the traditional process, the necessary correction is performed by hand by a skilled craftsman. He starts with the uncorrected half-tone blocks and corrects them by etching away part of the surface in appropriate areas, so that, when the blocks are used for printing, the intensity of colour is reduced in proportion to the amount of surface material removed (because the area of contact between block and paper is thereby reduced). For example, the craftsman would compare the yellow block with the original and, where the original had magenta, he would etch away the surface. The process is, of course, complicated by the fact that real originals are not in pure colours. The craftsman must then decide from his experience what proportion of the primaries are present in a particular colour and 'fine-etch' his blocks accordingly. This is a tricky business and takes a lot of time. A set of colour blocks may take a week's work.

A new machine, the Finella Colour Klischograph, employs electronic methods of colour correction. Basically, these take the form of measuring the size of the unwanted responses, deriving appropriate a.g.c. voltages and using the latter to modify the printing block at the outset, instead of after it has been made. Correction is made possible in this particular machine by the fact that no photographic processes are involved. The small areas of the surface of a printing block which are removed to leave the 'dots' of the picture are cut away by a stylus instead of being etched selectively with the aid of a 'resist'. The colour correction can, therefore, be applied at any point on the picture by suitably modifying the depth of cut of the V-shaped stylus.

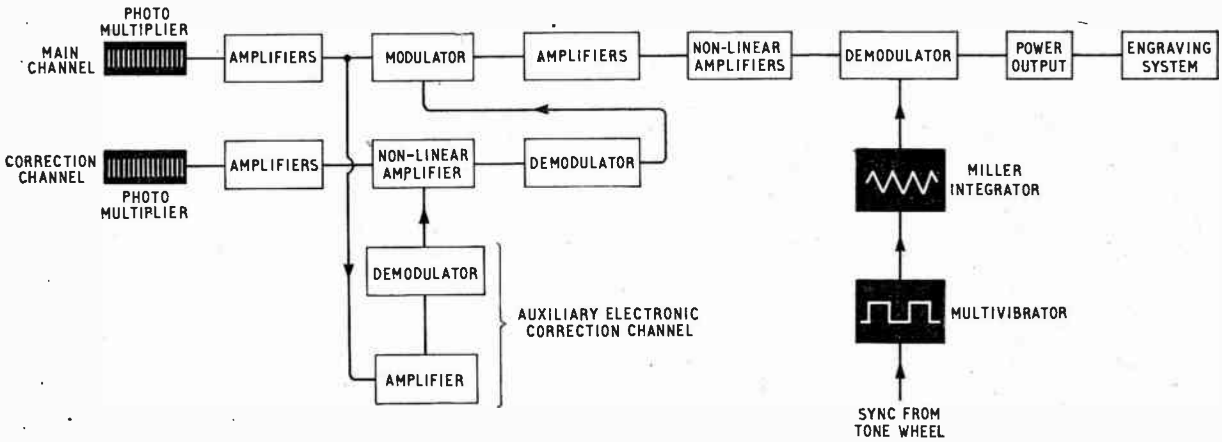


Fig. 3. *Block diagram of the electronic parts of the Colour Klischograph equipment*

It is convenient to call the peak response of any filter to white light 100, and the response to other colours some fraction of this. On this basis, the response of the blue filter employed when making a yellow printing block to various colours is as shown in Fig. 2. The responses shown are those which occur when a completely saturated colour, illuminated by white light, is viewed through the filter by a photocell. Violet, cyan and magenta should all give 100% responses when viewed through a blue filter. In practice, all three, and especially violet, give signals that are too low. If these outputs were passed to the engraving stylus, the depth of cut would be decreased and this would increase the intensity of the yellow printed at that point. Since cyan, magenta and violet contain *no* yellow, this is obviously undesirable.

We wish, therefore, to increase the amplification for magenta, cyan and violet, but not for white. A colour-sensitive amplifier is required and, to effect this, a second filter and photocell are employed.

Fig. 3 is a block diagram of the electronic part of the Colour Klischograph machine. The light source used in the scanning system is 'chopped' to avoid the need for d.c. amplification. Scanning is effected by moving the light source and picture in a series of lines, like a television raster, over the original. The system carrying the cutting stylus is mechanically coupled to the light source so that the position of the light over the original picture is followed by that of the stylus over the plate.

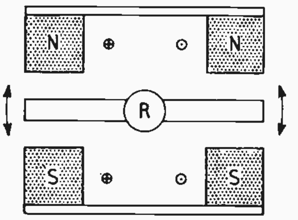
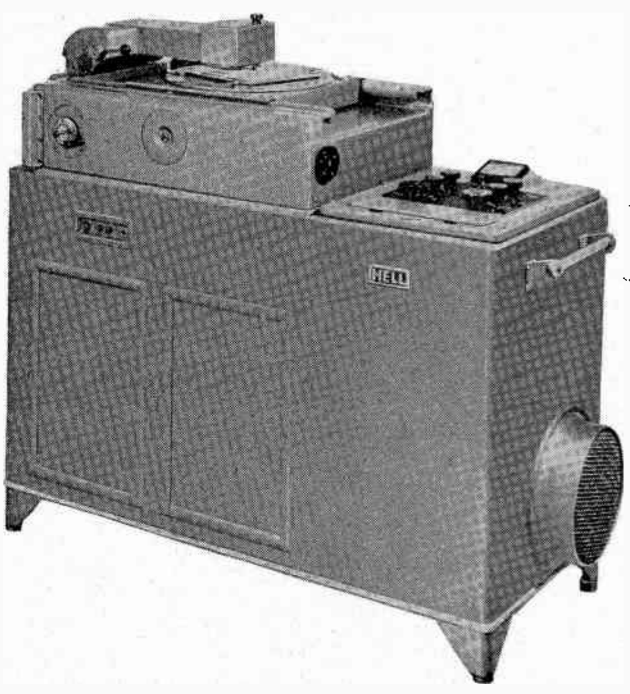


Fig. 4. *Electromagnetic torsion system for producing stylus movement. The rod R carries a soft-iron armature*

An electromagnetic-torsion system (Fig. 4) is used in the engraving process. A rod *R* is fixed at one end and supported by a bearing at the other. It twists as the coils are energized, and the movement is converted mechanically to a vertical stylus movement. The continuous up-and-down movement of the stylus, which gives the plate its dot structure, is imparted by adding a triangular waveform to the main channel output. It is essential that there should be a fixed number of dots per inch of scan. To obtain this, a tone-wheel pick-off on the table drive-shaft is employed to trigger the multivibrator which generates the dot frequency.



The Finella Colour Klischograph machine

The modulator consists of two pentodes in push-pull. The gain for small signals is a linear function of the d.c. bias applied to it from the second channel.

The filter, which looks at the same point on the original as the main filter, is chosen to see a low signal (compared with white) when looking at magenta, cyan and violet, but a reasonably high signal for yellow, green and red. A certain 'green' filter fulfils this requirement.

The light passing through this second filter (the 'correction filter') impinges on a photomultiplier. The resulting signal is amplified, rectified and passed to the modulator stage in the main channel. The nature of this is such that the higher the d.c. voltage from the second channel, the lower the amplification in the main channel, and vice versa.

Thus, in the above case, the amplification is a minimum for white but larger for violet, magenta and cyan. By proper choice of filter and non-linear characteristic of the second-channel amplifier, it is possible to arrange for the signal coming out of the modulator to be equal for white, magenta, cyan and violet, which is what is required.

The modulator is followed by a non-linear amplifier. This has the purpose of allowing for the non-linear relationship between depth of cut and tone value (reflectivity) and also introduces the necessary distortion of the curve required for the half-tone printing process.

The non-linear amplifiers used in the equipment are made up of three parts:

1. A linear amplifier.
2. A triode limiter-amplifier, which limits both positive and negative half-cycles.
3. A linear amplifier, which is biased so that it does not pass signals until some predetermined level is exceeded.

The non-linear relationship is different for each colour. This is necessary to ensure correct reproduction of neutral greys. Different proportions of the three colours have to be printed to obtain greys, because of the imperfect absorption of the inks.

All power supplies to the equipment are thoroughly stabilized. A.C. supplies are stabilized by saturable reactor systems and, in addition, d.c. stabilization is used in the h.t. supplies to photomultipliers (gas-filled stabilizers), picture lamps (barretters), and the main h.t. supply (series-parallel valve stabilizer). Gas-filled voltage stabilizers are used at various points in the circuit. In practice, the equipment gives reliable operation. A newer and more complex machine, the Multi-Vario Klischograph, has also been developed. This has additional facilities for enlarging or reducing the size of the colour blocks with respect to the original, working from transparencies, as well as opaque originals, either monochrome or in colour, and producing line engraving or line-and-tone combinations as well as half-tone blocks. It works about 50% faster than the original machine, and can produce large blocks up to about 17 in. by 12 in.

Underwater Acoustic Echo-Ranging

SPECTRUM OF REVERBERATION AND INFLUENCE OF PULSE DURATION AND RECEIVER BANDWIDTH ON SIGNAL/REVERBERATION RATIO

By J. W. R. Griffiths, B.Sc.,* and A. W. Pryor, Ph.D.†

SUMMARY. *It is shown as a result of some experiments at sea that reverberation from the sea bottom resulting from an acoustic pulse transmission has a power spectrum similar to that of the pulse.*

Using this information, a theory has been developed to determine the effect of pulse duration and receiver bandwidth on reverberation level, peak signal/reverberation ratio. A comparison is made with the corresponding effects when the background is white noise, and it is shown that in both cases optimum results are obtained when the product of bandwidth and pulse duration is approximately unity.

An underwater echo-ranging system is similar in principle to radar but, since electromagnetic waves are severely attenuated in sea water, it is necessary to use some other form of transmission; it has been found that acoustic waves of a frequency lying in the ultrasonic region are most suitable. Historically, acoustic echo-ranging devices preceded radar by many years, very crude systems being used for detecting enemy submarines during the first World War. During World War II, under the code name A.S.D.I.C. (*A*llied *S*ubmarine *D*evices *I*nvestigation *C*ommittee) acoustic echo-ranging systems played a big part in the war against the U-boats. (The United States used similar systems under the code name S.O.N.A.R.; i.e., *SO*und *NA*vigation and *R*anging). In this country, the name 'asdic' is now used in a much wider sense to cover most underwater echo-ranging systems in which the beam is aimed in a direction other than vertically downwards below the ship. When the beam is so directed; i.e., when determining the depth of water below the ship, the traditional name of echo-sounder is retained. A more detailed discussion of underwater echo-ranging can be found in a very comprehensive study of the subject by D. G. Tucker[4].

The background against which a wanted echo-signal has to be detected comprises two main parts, namely :—

(a) Noise; i.e., random fluctuation not dependent on the acoustic transmission, comprising circuit noise or sea noise caused by waves, currents, etc.

(b) Reverberation; i.e., acoustic energy which is derived from the acoustic transmission of the asdic set being scattered by the particles or gases suspended in the water or lying on the sea bed, or by the irregularities in the water-air surface.

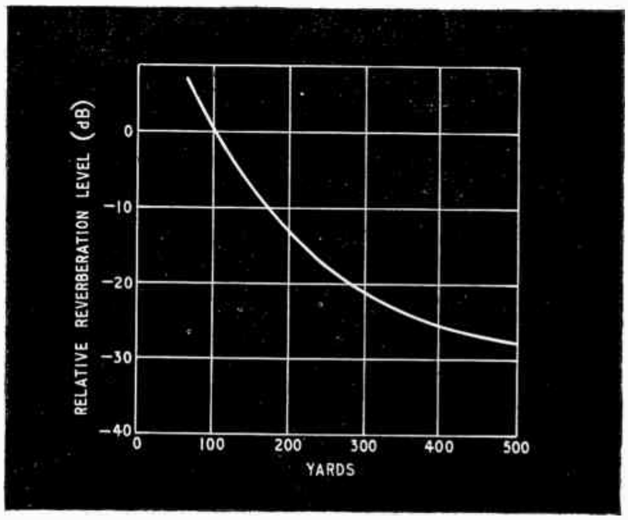
In the design of an asdic set, both factors have to be known, although, as the reverberation power is proportional to the transmitted power, it is often possible to make the reverberation predominate by increasing the latter. It is therefore important to know what is the spectrum of reverberation and how its shape affects detection. It is the subject of this paper to investigate these matters.

General Considerations of the Problem

In radar, the background is usually 'white' noise; i.e., noise having constant energy per c/s over the relevant frequency band. The effect of this type of noise on the detection of pulsed signals has been extensively investigated by many authors (e.g., Lawson and Uhlenbeck[1]) and it has been shown that there is an optimum bandwidth for the receiver, a fact which depends on the uniform energy spectrum of the noise.

Reverberation is formed from the addition of numerous individual reflections of the transmitted pulse; since

Fig. 1. *Typical curve of decay of reverberation energy with range*



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(*The authors were both in the Royal Naval Scientific Service when the experimental work discussed was carried out.*)

the path lengths will not be identical, the pulses are of random phase and consequently the instantaneous amplitude of the reverberations will be random. But as the reverberation is generated by pulses—the frequency spectrum of which is not uniform—there is every reason to suppose that the reverberation spectrum is also not uniform although, owing to the fact that the number of contributions to the total return is varying continuously, some spreading of the spectrum would be expected, together with some smoothing of the discontinuities in the power spectrum of the pulses. When the spectrum of the transmitted pulse is fairly narrow (i.e., for pulse durations of 100 msec or more) marked spreading has been reported[2], but the experiments discussed below, in which the maximum pulse duration was 1 msec, show that for small pulse durations the spreading is sufficiently small not to influence the shape of the spectrum significantly.

The reverberation spectrum is thus different from that of white noise and, hence, it becomes necessary to re-examine the problem of the effect of receiver bandwidth on the output signal/background ratio.

In the calculations that follow, the case considered is that where the signal echo consists of a single pulse identical in shape to that transmitted. This sort of echo will only be received from the simplest of targets (e.g., a smooth sphere) and, in general, the returning echo comprises a number of individual reflections from different parts of the target. These reflections are spaced in time to combine to form a pulse shape which can be quite different from that of the transmitted pulse. The effect of this on the calculations depends on how the signal/background ratio at the output of the receiver is defined.

Consider a single echo of rectangular shape in a background of noise. It is obvious that while the echo is being received the signal/background ratio—defined as the ratio of signal amplitude to r.m.s. background—is constant but, over the remainder of the scan, the signal/background ratio is zero. If the receiver bandwidth is insufficient to pass the pulse without envelope distortion then, at the output of the receiver, the signal/background ratio will increase to a maximum and then fall again to zero. Most works on the subject of the detection of signals in noise consider, in such cases, the signal/background ratio at the maximum as being the criterion of detection. If the ratio of the average signal power to average background power is used as a criterion (i.e., the distribution of the energy in time is ignored) then it is obvious that when the spectrum of the signal and the background are similar, as in the case of reverberation, variation of receiver bandwidth can have no effect.

It is shown in reference 5, in which the addition of a number of randomly phased sinusoids is considered, that the probability distribution of the total amplitude rapidly tends to a Gaussian form even when the number of contributions is relatively small. Hence, when the echo is complex (i.e., consisting of a number of individual reflections) the amplitude distribution tends to be similar to that of the background, and thus we are aware of the presence of the echo only because of an increase in the mean background power—the increase lasting over a period which will be determined by both the extent of the target and the transmitted pulse duration. A useful criterion of detection under these conditions is the ratio of the new mean—or better still the change in mean when the signal is present—to the r.m.s. background, the maximum value of this ratio being used.

Practical asdic sets produce echoes which will obviously lie between the two extremes; i.e., the single-echo return on the one hand and the very complex return on the other. High-definition asdic, such as might be used to distinguish large fishes, is likely to produce relatively simple echoes, whereas an asdic using a fairly long pulse will receive a very complex echo from, say, a shoal of fish. Consequently, the results obtained below must be assessed in relation to the particular requirements of performance of the asdic set considered.

Experimental Determination of the Spectrum of Reverberation

Owing to the spreading and attenuation of the acoustic energy both on the outward and on the return journey the reverberation energy, as measured at the receiver, from the more distant reflectors is smaller than for those nearby; hence measurements of reverberation must be made at a particular range. This is possible by choosing the returning energy at a definite time after transmission. Fig. 1 shows a typical curve of the decay of reverberation energy with range. For the purpose of these experiments, the reverberation energy from a strip of range about 10 yds long was selected by means of a 'range gate'. Since the reverberation is random, however, it was necessary to average the samples over a number of successive transmissions. This was achieved by using a detector with a fairly short charging time-constant (20 msec) and a long discharge time-constant (2 secs).

The asdic equipment used for the experiments operated at 100 kc/s which meant that, since a suitable

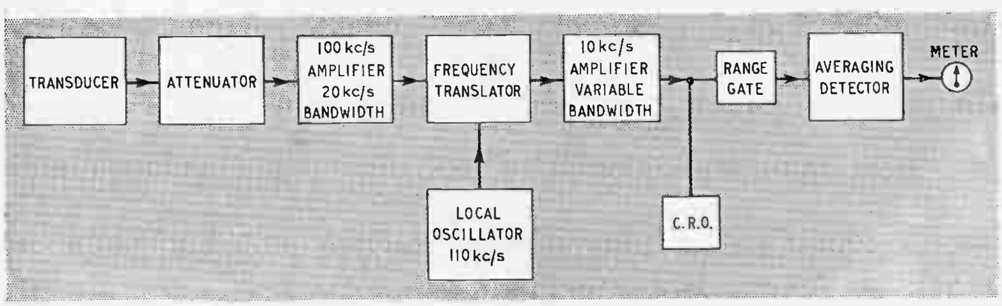

Fig. 2. *Block diagram of experimental equipment*

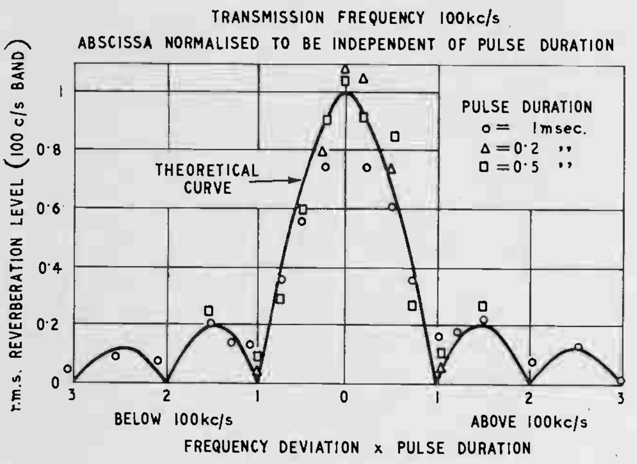

Fig. 3. *Frequency spectrum of reverberation as explored with* 100-*cycle pass-band*

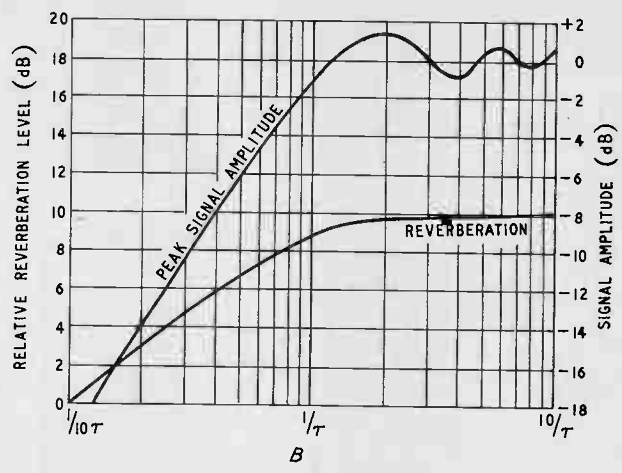

Fig. 4. *Curves showing effect of bandwidth on peak signal and reverberation level*

commercial frequency analyser was not available at this frequency, it was necessary to improvise using, as far as possible, available instruments. A block diagram of the arrangement is shown in Fig. 2. The electrical output from the transducer was applied via an attenuator and a 100-kc/s amplifier of 20-kc/s bandwidth to a linear frequency translator. The bandwidth of the 10-kc/s receiver following the translation was set at 100 c/s and the range gate set at a fairly close range, and adjusted so that no predominant single echoes were included. By varying the frequency of the local oscillator, it was possible to sweep the spectrum of the reverberation signal with the 100-c/s bandwidth of the 10-kc/s receiver. By adjusting the attenuator at each frequency the meter output was kept constant and the attenuator reading recorded.

The spectrum of a rectangular pulse of duration τ and carrier frequency f_0 has the well-known form shown by the full line in Fig. 3. Although the width of the spectrum is dependent on τ the shape is not [see Equ. (1) in Appendix 1], and it is possible to draw one curve representing all possible values of τ by using a scale of frequency deviation multiplied by pulse duration for the abscissa. Experimental results for three different pulse durations, namely 0·2 msec, 0·5 msec and 1 msec, are shown on the graph, from which it can be observed that the measured values suggest that the reverberation spectrum does not depart greatly from that of the pulse.

It may be noted that the cusps are not so evident on the results based on the 1-msec pulse duration, but this is probably due to the smoothing effect of the 100-c/s receiver bandwidth. For the 1-msec measurements, the zeros are only 1,000 c/s apart and, hence, the 100-c/s receiver bandwidth is significant in comparison whereas, with the 0·2-msec measurements, the zeros are 5,000 c/s apart and the smoothing is less noticeable.

Effect of Receiver Bandwidth and Pulse Duration on Signal/Background Ratio

Having examined the spectrum of reverberation and decided that it is of the same shape as that of the pulse, it is of interest to see how this affects detection of a single target* and, in particular, to calculate the effect on the output signal/reverberation ratio of passing a pulse signal, in a background of reverberation, through a receiver of limited bandwidth.

The bandwidth of the receiver, when very wide, will have negligible effect on either the pulse signal or the reverberation background; but, as the bandwidth is reduced, it will control the shape of the pulse and, hence, the maximum amplitude which the signal attains at the receiver output, and also will control the amount of reverberation energy which is allowed to pass.

If we assume that the receiver has a rectangular pass-band of width B c/s, then it can be shown that the maximum height which the pulse attains is:

$$\mathrm{Si}\left(\frac{\pi B\tau}{2}\right)$$

where τ = pulse duration (sec.)

and $$\mathrm{Si}(x) = \int_0^x \frac{\sin x}{x}\, dx.$$

In the Appendix to this paper an expression is derived relating the reverberation energy at the output of the receiver to the receiver bandwidth. Curves representing the maximum amplitude attained by the signal and the output of reverberation energy, plotted against bandwidth are shown in Fig. 4. The ordinate scales of both these curves are relative and have no particular relation to one another.

If we define the output signal/background ratio as the ratio of peak signal/r.m.s. background then, from the above results, we can determine the effect of bandwidth on this ratio. A curve of this is shown in Fig. 5 and, for comparison, a similar curve has been calculated for the peak signal/noise ratios. It may be noticed that in contrast to the case of white noise there is no defined optimum bandwidth for the reverberation-limited detection other than the small oscillations introduced by the oscillatory nature of $\mathrm{Si}(x)$ and probably of little practical significance. Thus, for reverberation-limited

* The echo from the target is assumed to be different from the transmitted pulse in amplitude only

detection, it is only necessary to make $B > 1/\tau$ in order not to introduce any deterioration. However, if B is made much larger than necessary more 'white noise', due either to the receiver itself or noise in the sea, is allowed to pass and, hence, detection may no longer be reverberation-limited. Taking this into consideration, it would appear that optimum results will still be obtained, as in the noise case, with the bandwidth kept in the region $1/\tau$ to $2/\tau$.

So far in the theoretical investigation, we have kept τ fixed and varied B. Now if we fix B and vary τ then another factor has to be taken into consideration. Reverberation energy is proportional to the transmitted energy; hence since the latter is proportional to the pulse duration so also, ipso facto, must be the reverberation energy. Taking this into account and assuming that the peak amplitude of the incoming target-signal is unaffected by the pulse duration, Table 1 can be drawn up to show how the background, bandwidth and pulse duration affect the signal/background ratio.

If both B and τ can be varied independently then obviously we obtain maximum signal/reverberation ratio by reducing τ as much as possible while, at the same time, maintaining the product $B\tau$ at a value of approximately unity. In practice, this process may be limited by the performance of the display. It is shown in reference 6 that, due to the physical size of the spot on a cathode-ray tube and the stylus size on a chemical recorder, the display acts like a low-pass filter and thus limits the effective bandwidth of the system. Hence once τ is reduced to the point where its product with this effective bandwidth is unity, then no further improvement of signal/background ratio results from subsequent reduction.

TABLE 1

The Effect of B and τ on Signal/Background Ratio

Background		$B\tau \ll 1$	$B\tau \gg 1$
Reverberation	Effect of τ	Nil	$\frac{S}{R} \propto 10 \log \frac{1}{\tau}$
	Effect of B	$\frac{S}{R} \propto 10 \log B$	Nil
White Noise	Effect of τ	$\frac{S}{N} \propto 20 \log \tau$	Nil
	Effect of B	$\frac{S}{N} \propto 10 \log B$	$\frac{S}{N} \propto 10 \log \frac{1}{B}$

Summary of Results and Conclusions

The practical results show that for short pulses the reverberation power-spectrum can be assumed to be of the same shape and extent as the power-spectrum of the pulse, and from this is has been possible to derive the effect of pulse duration and receiver bandwidth on the Signal/Reverberation ratios. The main conclusions are:

(a) For narrow bandwidths, pulse duration has little or no effect, and the S/R ratio is proportional to 10 log B dB.

(b) For wide bandwidths, the bandwidth has little or no effect, and the S/R ratio is inversely proportional to 10 log τ dB.

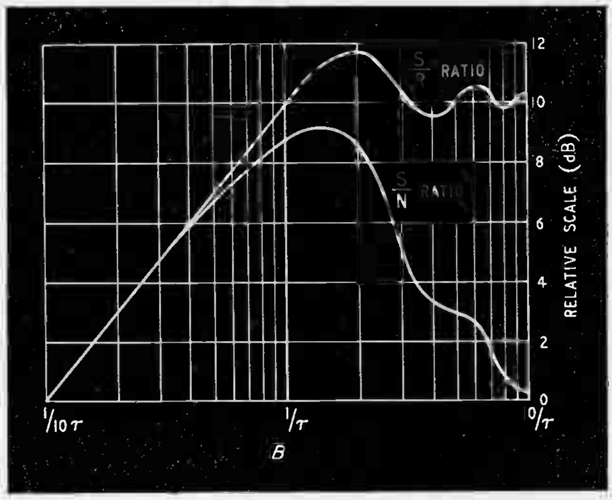

Fig. 5. *Comparison of effect of bandwidth on signal/reverberation and signal/noise ratios*

Thus, when detection is limited by reverberation, as short a pulse as possible should be used and the bandwidth of the receiver adjusted so that the product of the bandwidth and pulse duration is approximately unity.

Acknowledgement

The authors wish to acknowledge the permission of the Admiralty to publish this paper.

APPENDIX

The determination of the reverberation energy passed by a receiver with a rectangular band shape of width B c/s.

The energy spectrum of the pulse as shown in Fig. 3 is proportional to:

$$\left\{\frac{\tau \sin \pi (f-f_0) \tau}{\pi (f-f_0) \tau}\right\}^2 \qquad (1)$$

Thus the energy in any band can be found by integrating this expression between the appropriate limits. In particular, to find the energy E_B in the band $f_0 - B/2$ to $f_0 + B/2$ (i.e., a band of B c/s symmetrically placed around the carrier frequency f_0) then E_B is proportional to

$$\int_{f_0 - B/2}^{f_0 + B/2} \left\{\frac{\tau \sin \pi (f-f_0) \tau}{\pi (f-f_0) \tau}\right\}^2 df$$

Making the substitution $\pi (f-f_0) \tau = x$ we obtain

$$E_B = \frac{2K\tau}{\pi} \int_0^{\pi B\tau/2} \left(\frac{\sin x}{x}\right)^2 dx$$

where K is a factor of proportionality.

This expression can be integrated by parts to obtain

$$E_B = \frac{2K\tau}{\pi}\left\{\mathrm{Si}\,(\pi B\tau - \frac{\sin^2 (\pi B\tau)/2}{(\pi B\tau)/2}\right\} \qquad (2)$$

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Safety Ohmmeter

WHEATSTONE BRIDGE ENERGIZED BY SELENIUM PHOTOCELL

Fairey photo-electric safety ohmmeter. The selenium cell is concealed by a hinged cover which is lifted in use

The use of a selenium photocell as the power source for a resistance bridge eliminates the normal hazards involved in the testing of detonating circuits. Tests rely on a measurement of resistance, which necessarily implies the application of electrical energy to the circuit. To remove the possibility of accidental ignition, the energy applied during the test must be below the safety margin of the component under any condition of instrument failure or misuse. In conventional methods this is achieved by including a current or voltage limiting device with the power source. In the event of a breakdown of this device, dangerously high currents can be introduced into the firing circuit, with a consequent risk of ignition. This danger is overcome by the use of a selenium barrier-layer photocell, the output of which, under any condition of light saturation or failure, cannot exceed a short-circuit current of 10 mA or an open-circuit voltage of 0·7 V.

TABLE I

Range	Percentage Scale Reading				
	5%	10%	50%	100%	
0–10 Ω	9	5·5	3	2·7	Illumination intensities required to produce a detectable galvanometer current for a ± 10% incremental change in resistance setting (foot-candles).
0–100 Ω	2	1·3	0·8	0·7	
0–1 kΩ	2·5	2	1·7	1·6	
0–10 kΩ	13	12	12	12	
	Foot-candles				

Designed and developed by the Weapon Division of The Fairey Aviation Company, Ltd., Heston, Middlesex, the instrument utilizes a Wheatstone bridge, providing a range of resistance measurement from 0–10 kΩ in four switched ranges (0–10 Ω, 0–100 Ω, 0–1 kΩ, 0–10 kΩ). A 'transit' position on the switch is arranged to place a short-circuit across the meter terminals for galvanometer damping. The power source for the bridge is a sensitive infra-red selenium photocell, the characteristics of which are modified in manufacture to limit the short-circuit current of the cell at light saturation, and yet maintain a high sensitivity at low illumination intensities. Null indication is provided by a centre-zero galvanometer, the pole pieces of which are shaped for off-balance insensitivity. To achieve a high accuracy and stability, ± 0·1% wire-wound resistors, cast in an Araldite block, are used for the bridge ratio arms, and a wire-wound cam-corrected potentiometer of ± 0·1% linearity for the variable element. To compensate for the ± 5% overall resistance tolerance of this potentiometer, a 5-Ω variable wire-wound resistance is included in the fixed ratio arm. This is accurately adjusted on initial assembly and locked.

The fundamental bridge accuracy (determined by component tolerances) is of the order of ± 0·3%. The measurement accuracy, however, is dependent upon the sensitivity of measurement and the scale-reading accuracy. At low illumination intensities (i.e., below 50 foot-candles), and under the load conditions presented by the bridge at any setting, the photocell is essentially a constant-current generator. The current output, and therefore the sensitivity of the bridge, is directly proportional to the intensity of the light incident on the photocell. The illumination intensities required to produce a detectable galvonometer current for different range and scale settings are shown in Table 1. They are based on a ± 10% incremental change of the variable element and are an indication of the measurement sensitivity of the instrument.

Using normal bridge-balancing techniques, an improvement of at least 2 : 1 in bridge-setting accuracy can readily be achieved, since the figures quoted for measurement sensitivity relate to a detectable galvanometer deflection. Thus, the intensities given in Table 1 are sufficient to provide a measurement accuracy of approximately ± 5%.

Checking a firing circuit with a Fairey safety ohmmeter

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Measuring Earth Conductivity

SIR,—I have read with interest the article by M. Strohfeldt under the above title in your November issue, and am rather concerned about certain statements made regarding the Wenner or four-electrode method. In the section on "Comparison of Data" the author states: "The Wenner equation is derived on the assumption of homogeneous ground and neglects also the highly variable resistances in the electrode-to-earth contacts". Again, in his "Conclusions" he states: "Apart from the inaccuracy of the Wenner method when applied to non-homogeneous ground, it is not possible to allow for variations in the ground-to-electrode contacts".

The Wenner method has been used very successfully for many years in the resistivity method of geophysical surveying and there is a large amount of literature available on the subject. The simple formula given by the author applies only to homogeneous soil, and there are available other formulae which deal with various forms of non-homogeneity. By making measurements by the Wenner method and making use of the theoretical investigations it is possible to make deductions as to the nature of the underlying soil, resistivities, etc.

The author appears to have a wrong conception of the part played by the resistances of the electrodes themselves. The two outer electrodes are included in the main current path, and this combined resistance is one of the factors determining the value of this current. Since this current is measured and produces the voltage drop between the inner electrodes, the value of this combined resistance can have no influence at all on the final value of resistance. Whether the resistance of the two inner electrodes affect the result depends on the nature of the equipment used to measure the potential difference between them. If this is of the potentiometer type, drawing no current from the soil at balance, then the resistances of the electrodes have no effect on the measured potential difference and hence on the calculated resistance. If the measuring system is of the voltmeter type drawing current from the soil, then an error can be introduced if the resistances of the electrodes are not small enough to be negligible compared with the resistance of the voltmeter. It is, however, a simple matter to measure the resistances of the electrodes and to apply a correction. Instruments are available which will measure the resistance directly on a scale, and in which it is possible to correct easily for high electrode resistances. It is thus clear that electrode resistances are no problem, and that a final result can be obtained which is independent of them.

Evershed & Vignoles Ltd., G. F. TAGG
Chiswick, London, W.4.
20th November 1957.

New Books

An Introduction to Transistor Circuits

By E. H. COOKE-YARBOROUGH, M.A., M.I.E.E. Pp. 154 + xii. Oliver & Boyd, 39a Welbeck Street, London, W.1. Price 15s.

The first chapter, Transistor Action, gives an elementary account of the action of semiconducting devices and explains junction diodes and both junction and point-contact transistors. As is but natural in only 26 pages, the treatment is very far from exhaustive, and the main purpose of this chapter is to give the equations relating voltage and current in the transistor upon which much of the rest of the book depends.

Chapters 2 and 3 deal with Low-Frequency Amplification and The Performance of Transistor Amplifier Circuits. Chapter 4 covers Pulse Circuits, while Chapters 5 and 6 treat Some Circuit Applications of Transistors and The Application of Transistors in Automatic Computers.

It is a disconcerting peculiarity of the book that all p-n-p transistor circuits are drawn upside down. This is done so that the positive power-supply line can be at the top of the diagram and conventional direct currents flow downwards in the manner usual in valve circuits. The author claims that this assists "the building up of a mental picture of the flow of currents in a transistor circuit".

It has the effect of making quite simple circuits unrecognizable. Instead of being able to see the general form of a circuit at a glance, one has to examine it carefully to find out what it really represents. This is very necessary if one is not to be misled. The circuit of a direct-coupled amplifier with d.c. feedback on p. 61, for instance, *looks like* the transistor equivalent of three cathode followers. It is only the arrows of the transistor symbols which reveal it to be three cascade-connected common-emitter transistors.

It is not possible with p-n-p transistors to draw a diagram which is in all ways analogous to that of a valve circuit; either the diagram must be inverted or the directions of voltages and currents. It is debatable which should be done, as the author points out, but it is the reviewer's opinion that it is easier to cope with reversed voltages and currents than with an upside-down diagram.

The treatment involves only elementary mathematics and the general approach is quite different from the usual one. The circuit relations are developed in a way which gives a good insight into the important factors. Only small signal amplifiers are dealt with but, in the case of pulse circuits, conditions in which transistors are swung between cut-off and bottoming are treated.

No great attention is given to the practical problem of stabilizing the operating point against temperature changes, and power amplifiers are not discussed. As its title indicates, the book is an introduction. As such, it is a good one which merits study by any serious student of transistor circuitry. W.T.C.

Progress in Semiconductors (2)

Edited by ALAN F. GIBSON, B.Sc., Ph.D. Pp. 280 + vii. Heywood & Co. Ltd., Southampton Street, Strand, London, W.C.2. Price 63s.

This is the second volume of an annual collection of critical review articles under the general editorship of Dr. A. F. Gibson. The need for such a publication is evident because of the hundreds of papers which are published every year in this field. Indeed, the authors of one of the papers in the present collection feel it necessary to point out that their bibliography of 65 references is chosen to be representative rather than exhaustive. It is therefore useful to have a limited number of topics discussed each year by authorities in particular branches of the subject as an antidote to the ever-increasing specialization to which semiconductor physics is subjected.

The eight papers deal with the properties of semiconductor materials rather than devices constructed from them, which are mentioned only incidentally. Until recently, most work on the study of semiconductor materials was concerned with germanium and silicon. A search is now under way in some laboratories for new materials, partly in order to obtain optimum properties for certain applications and partly by way of scientific curiosity. A paper by Herman, Glicksman and Parmenter (of R.C.A.) describes experimental and theoretical work on the germanium-silicon alloy system. A significant advance has been made in this field and it has proved possible to interpret the observed variation with alloy composition of certain major electrical and optical properties. A paper by Cunnell and Saker (of S.E.R.L.) describes the properties of semiconducting compounds of elements from groups III and V of the periodic table (such as indium antimonide). Investigation of these compounds was initially stimulated by the hope that they would be useful in the transistor field because of their large energy gaps and high mobilities together with low melting points. However, there is at the moment no clear evidence of transistor action (i.e., power gain), although several of the compounds show good rectification. Consequently, such practical applications as have been found for these compounds have been based on the Hall effect or the magneto-resistive effect. This is partly because the materials, as prepared at the moment, are comparatively impure. A great deal of further effort will have to be given to purification before their future potentialities become clear.

Despite recent research on other materials, almost the whole of transistor production uses germanium. The demand for improved transistor characteristics and greater uniformity between samples has resulted in increased demands being made on the technology of producing germanium single crystals. A paper by Cressell and Powell (of M.W.T.) surveys the available methods of crystal growing and deals particularly with the horizontal zone levelling technique recently introduced for preparing crystals with uniform resistivity along their length. The authors describe their own successful results with this method. The best-known impurities in germanium are the elements in groups III and V which are often used for doping (such as antimony and indium). A paper by Dunlap (formerly G.E.) presents some recent results on the donor and acceptor action of these elements and discusses particularly other elements besides those in groups III and V, which have been shown to have electrical activity in germanium.

A paper by Crawford and Cleland (Oak Ridge Lab., U.S.A.) deals with the effects of atomic radiation on semiconductors. The sensitivity of the electrical properties of semiconductors to imperfections of the crystal lattice is well known. Fast particle bombardment has been used to investigate the defect sensitivity of semiconductors. Conversely, because of their great sensitivity to lattice defects, semiconductors have aided in the study of the nature of radiation damage. A paper by J. B. Gunn (Univ. of British Columbia) deals with the effects of high electric fields in semiconductors and with avalanche ionization in particular. This process is of considerable technical importance, as it limits the reverse bias voltage which can be applied to a p-n junction. Among the experimental results described is the emission of visible light from a region of avalanche multiplication in silicon.

A paper by Rose (R.C.A.) discusses lifetimes of free electrons and holes in solids, a subject of great importance in the broad fields of luminescence, photoconductivity and semiconductor devices. The final article by Curie (Univ. of Paris) deals with theories to explain the luminescence which certain phosphors, such as zinc sulphide, exhibit under the action of an electric field. J.E.F.

Electronic Voltage Stabilizers for Laboratories, Computers and Control Systems

By J. Miedzinski, B.Sc., A.M.I.E.E., and S. J. Zgorski. Pp. 19 + 5 of figures. The British Electrical & Allied Industries Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 12s. 6d.

Discusses the economics of the provision of multiple stabilized power packs for laboratories, etc., and proposes the solution of using a single source of high-voltage (600 V) d.c. as the basic h.t. supply. This is fed to individual stabilizers which derive l.t. power from the a.c. mains. The design and construction of the stabilizer used for one application are described.

Piezoelectricity

By members of the Staff of the Post Office Research Station. Pp. 369. H.M.S.O., York House, Kingsway, London, W.C.2. Price 75s.

A collection of research reports comprising theoretical papers and information required for the practical application of certain water-soluble crystals. Techniques for growing and fabricating crystals are also described.

Microwave Measurements

By Edward L. Ginzton. Pp. 515. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 90s.

Contains chapters on: Generation of Laboratory Signals; Detection of Microwave Power; Measurement of Microwave Power; Impedance Concepts at Microwave Frequencies; Measurement of Impedance; Representation and Measurement of Microwave Circuits; Measurement of Wavelength; Measurement of Frequency; Resonant Cavity Characteristics—Measurement of 'Q' and R_0/Q_0; and Measurement of Attenuation. The book is intended for use as a textbook in a first-year course in microwave measurements and covers the subject in general terms, not merely with respect to radar. The treatment is not highly mathematical.

Elecktronenröhren: Lehrbuch der drahtlosen Nachrichtentechnik, Vol. 3

By M. J. O. Strutt. Pp. 391. Springer Verlag, Reichpietschufer 20, Berlin, W. 35. Germany. Price DM 58.50.

Contents include: the physics of electron devices; manufacturing techniques; valve characteristics; electron optics, electron devices (photocells, transistors, gas-filled valves, etc.); cathode-ray tubes, (including those for colour television).

Index of Technical Articles

Published monthly by Iota Services Ltd., 38 Farringdon Street, London, E.C.4. Annual subscription £6. 6s. Articles are classified under 33 headings, which include "Electrical and Electronic Engineering" and "Process Control Systems".

Transistor A.F. Amplifiers

By D. D. Jones, M.Sc., D.I.C., and R. A. Hilbourne, B.Sc. Pp. 152. Published for *Wireless World* by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 21s.

Deals with the design of audio amplifiers with outputs up to 20 watts. Details of five circuits of proved performance are included. The authors are at the research laboratories of The General Electric Company.

Definitions and Formulas: Radio and Television Engineering

By A. T. Starr, M.A., Ph.D., M.I.E.E. Pp. 65. Sir Isaac Pitman & Sons Ltd., Pitman House, Parker Street, Kingsway, London, W.C.2. Price 2s.

***Wireless World* Guide to Broadcasting Stations 1957–58**

Published by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Pp. 80, Price 2s. 6d.

Tabulated information giving frequency, wavelength, and power of over 2,000 short-wave stations, and some 750 European medium and long-wave transmitters.

Basics of Phototubes and Photocells

By David Mark. Pp. 136. John F. Rider Publisher Inc., 116 West 14 Street, New York 11, N.Y. Price $2.90.

A non-mathematical account of photocells and their application.

CABMA Register 1957-58 of British Industrial Products for Canada

Published jointly by Kelly's Directories Ltd., and Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Pp. 624. Price 15s., post free.

BRITISH STANDARDS

Methods of Testing Vulcanized Rubber

B.S. 903. Pts. A4 and A9: 1957. Price, Part A4, 3s.; Part A9, 5s.

Universal Decimal Classification

Abridged English Edition. B.S. 1000A: 1957. Pp. 252. Second Edition, revised 1957. Price 42s. (bound), 35s. (unbound).

Contains a general introduction explaining the U.D.C. system, tables of auxiliaries, main tables and alphabetical index.

Memorandum on the Standard Conditions for use during the Testing and Pre-Conditioning of Electrical Insulating Materials

B.S. 2844: 1957. Price 3s.

The Reduction and Presentation of Experimental Results

B.S. 2846: 1957. By J. T. Richardson, B.Sc. Price 10s.

Flexible Insulated Sleeving for Electrical Purposes

B.S. 2848: 1957. Price 6s.

Nickel-Iron Transformer and Choke Laminations

B.S. 2857: 1957. Price 3s.

The above are obtainable from British Standards Institution, British Standards House, 2 Park Street, London, W.1.

NATIONAL BUREAU OF STANDARDS

Standard Samples and Reference Standards

N.B.S. Circular 552 (2nd Edition). Pp. 24. Price 25 cents.

Report of the International Commission on Radiological Units and Measurements (ICRU) 1956

N.B.S. Handbook 62. Pp. 48. Price 40 cents.

Worldwide Occurrence of Sporadic E

By Ernest K. Smith, Jr. N.B.S. Circular 582. Pp. 278. Price $3.35.

N.B.S. Annual Report 1956
Pp. 158. Price 60 cents.

Fractional Factorial Experiment Designs for Factors at Two Levels
N.B.S. Applied Mathematics Series 48. Pp. 85. Price 50 cents.

Metrology of Gage Blocks
N.B.S. Circular 581. Pp. 119. Price $1.50.

The above are obtainable from the Superintendent of Documents, U.S. Government Printing Office, Washington 25, D.C., U.S.A. (One-fourth of publication prices to be added for postage.)

MEETINGS

I.E.E.

22nd January. "Special Problems of Broadcasting in Sweden", by E. Esping.

27th January. "An Enquiry into the Specification of Transistors", informal talk by F. F. Roberts, B.Sc.(Eng.).

28th January. Symposium on "Long-Distance Propagation above 30 Mc/s". At 2.30, "Ionospheric Forward Scatter Propagation", to be followed at 5.30 by papers on "Tropospheric Propagation Beyond the Horizon". Both members and non-members will be required to register and to purchase a set of the papers if they wish to attend the symposium. Price of the set of papers to members is 25/- (no registration fee) and to non-members 35/- (registration fee 10/-).

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C. 2, and will commence at 5.30, *except where otherwise stated.*

Brit. I.R.E.

7th January. "Mass Production of Television Tuners", discussion to be opened by P. C. Ganderton.

29th January. "Ultra-High-Speed Oscillography", by I. Maddock, O.B.E., B.Sc.

These meetings will commence at 6.30 *and will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.*1.

The Television Society

9th January. "A French Portable Television Camera", by J. Polonsky, at 7 o'clock at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.

23rd January. "Crystal Valves", Fleming Memorial Lecture by T. R. Scott, D.F.C., B.Sc., to be held at 7 o'clock at the Royal Institution, Albemarle Street, London, W.1.

The Institute of Navigation

17th January. "The Influence of Atmospheric Conditions on Radar Performance", by J. R. Saxton, D.Sc., Ph.D. at 5.15 at the Royal Geographical Society, 1 Kensington Gore, London, S.W.7.

STANDARD-FREQUENCY TRANSMISSIONS

The National Physical Laboratory has just published a new edition of its pamphlet describing MSF—the Standard-Frequency Transmissions from the United Kingdom. These radio transmissions are on the air almost continuously from the Post Office Station at Rugby. They enable anyone needing a precise frequency to check his apparatus against a standard which is known to one part in ten thousand million.

The pamphlet announces that the MSF frequencies are now based on the resonant frequency of the caesium atom. Provisionally, the value of 9 192 631 830 c/s per second has been adopted for this. This frequency is a fundamental physical constant, free from the small corrections and uncertainties associated with astronomical time. The precise value of the frequency is based on the value of astronomical time for 1955. Some years must pass before the astronomical and atomic units can be compared to the full accuracy attained by the atomic clock, because astronomical time must be averaged over a long period to eliminate small errors and uncertainties of measurement. If any corrections prove necessary in years to come, they can readily be made and therefore will not restrict the accuracy already available.

The MSF frequencies and modulations are maintained to ± 5 parts in 10^9 but are monitored to ± 1 part in 10^{10}. The results published in *Electronic & Radio Engineer* are given only to 1 part in 10^9 to allow for slight transmission and propagation errors, but even more precise average values can be used when steps are taken to eliminate these errors. The modulation programme of the MSF transmissions is as follows:—

Minutes past each hour				Modulation
0–5		30–35	45–50	1000 c/s.
5–10	20–25	35–40	50–55	1-c/s pulses, 60th pulse increased in duration.
10–14	25–29	40–44	55–59	Unmodulated.
14–15	29–30	44–45	59–60	Speech announcement.

Transmissions made on 2·5, 5, and 10 Mc/s are continuous except for a break between 15 and 20 minutes past each hour, and there is also a transmission on 60 kc/s between 14·29 and 15·30 hours G.M.T. Temporarily 19·59–21·00 G.M.T. The carrier frequency of the 16 kc/s GBR transmitter is now also controlled by the MSF standard.

A pamphlet entitled "MSF—Standard Frequency Transmissions from the United Kingdom" is available free on application to: The Director, National Physical Laboratory, Teddington, Middx.

(Communication from the National Physical Laboratory)
Deviations from nominal frequency for November* 1957

Date 1957 November	MSF 60 kc/s 2030 G.M.T. Parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^8
1	+ 3	+ 1
2	+ 3	+ 1
3	+ 3	+ 1
4	+ 3	+ 2
5	+ 3	+ 2
6	+ 3	+ 2
7	+ 4	+ 2
8	+ 4	+ 2
9	+ 4	+ 2
10	+ 4	+ 2
11	+ 4	+ 1
12	− 5	+ 1
13	− 5	+ 2
14	− 5	+ 2
15	− 5	+ 3
16	− 5	+ 3
17	− 4	+ 3
18	N.M.	+ 4
19	− 4	+ 4
20	− 4	+ 4
21	− 4	+ 5
22	− 4	0
23	− 4	0
24	− 4	0
25	− 4	0
26	− 4	0
27	− 3	− 1
28	N.M.	− 1
29	− 3	− 1
30	− 3	− 1

* *Nominal frequency is defined to be that frequency corresponding to a value of* 9 192 631 830 *c/s for the N.P.L. caesium resonator. N.M. = Not Measured.*

New Products

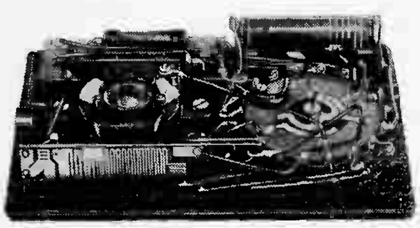
Pocket-Size Multi-Range Meter

Avo have produced a new multi-range meter, the Multiminor. Two models are available: Model 1 for use in temperate climates and Model 2 for use under adverse climatic conditions.

The makers state that all components which could be moulded have been made on precision moulding presses, the panclimatic version using a recently developed moulding powder which endows the finished product with very good electrical insulation under conditions of extreme humidity, and resists the growth of fungus. The indicating movement is constructed in a dustproof casing, and all wiring is termite proof.

The instrument is fitted with a high-grade rotary switch, similar to that incorporated in the Avometer.

A feature of the instrument is the use of printed resistors. It is claimed that this is the first time that such resistors have been used in any mass-produced apparatus. In one instance, it has been possible to combine a printed resistor with an auxiliary switch plate as an integral part of the selector-switch mechanism, another printed resistor forming the universal meter shunt. High-stability carbon resistors have also been employed wherever necessary.



Voltage and Current Ranges

D.C. VOLTAGE	A.C. VOLTAGE	D.C. CURRENT
Sensitivity (10,000 Ω/V) First indication 2 mV	Sensitivity (1,000 Ω/V) First indication 200 mV	First indication 2 μA
0 – 100 mV	0 – 10 V	0 – 100 μA
0 – 2·5 V	0 – 25 V	0 – 1 mA
0 – 10 V	0 – 100 V	0 – 10 mA
0 – 25 V	0 – 250 V	0 – 100 mA
0 – 100 V	0 – 1,000 V	0 – 1 A
0 – 250 V		
0 – 1,000 V		

Two resistance ranges are provided (0–20 kΩ and 0–2 MΩ), both of which employ an internal $1\frac{1}{2}$-V cell of international availability (Type U12). An ohms adjuster control provides compensation for the deterioration of the cell with age and use.

The accuracy is given as d.c., 3% of full scale and a.c., 4% of full scale. Approximate weight: 1 lb. (0·45 kg); overall size: $5\frac{5}{8}$ in. by $3\frac{5}{8}$ in. by $1\frac{3}{8}$ in. (14·3 cm by 9·2 cm by 3·5 cm).

Avo Ltd.,
92–96 *Vauxhall Bridge Road, London, S.W.*1.

Directly-Heated Subminiature Valves

Hivac Ltd. are marketing an augmented range of directly-heated subminiature valves, many of which are exact equivalents of American types used in portable radio communication equipment.

The range includes XFY14, output pentode (U.S. equivalent 5672); XFR1 r.f. amplifier pentode (U.S. equivalent 1AD4); XFR3 r.f. oscillator triode (U.S. equivalent 5676) and the XR4 r.f. power amplifier (U.S. equivalent 6397). All the valves, with the exception of the XR4, are 38·1 mm long, 10·1 mm wide and 7·6 mm in thickness. The filament voltages are 1·25. The XR4 is 40·64 mm long and has a diameter of 10·16 mm. The r.f. amplifier pentodes have metallized screening.

Hivac Ltd.,
Stonefield Way, Victoria Road, South Ruislip, Middx.

U.H.F. Attenuator Pads

Type A83 attenuators are coaxially-constructed pads with rod and disc-type resistors in T networks. Five units are available in steps of 10 dB and each unit is fitted with Plessey minor coaxial fittings so that they may be joined together in the same line.

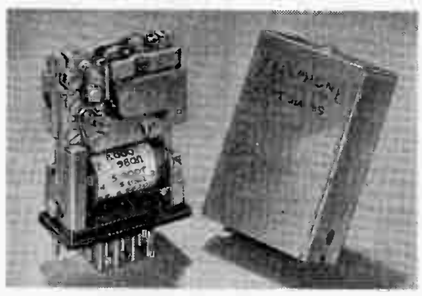
These attenuators are said to be capable of handling up to 0·5 W of sine-wave power. In sharply-pulsed conditions, their rating is limited by the peak voltage of the signal. The limit of the voltage is of the order of 100 V for pulses of a few microseconds duration. The attenuators are claimed to be suitable for use from d.c. to 3,000 Mc/s, and performance figures are given below:

Accuracy: ± 1 dB up to 1,000 Mc/s.
Input/Output Impedance: 75 ohms.
V.S.W.R.:
10-dB attenuator pad, less than 1·2 d.c. to 1,000 Mc/s. Increases above 1,000 Mc/s to not more than 1·7 at 3,000 Mc/s.
20–50-dB attenuator pads, less than 1·2 d.c. to 3,000 Mc/s.

Advance Components Ltd.,
Roebuck Road, Hainault, Ilford, Essex.

Polarized Relays

An improved Carpenter polarized relay, Type 51, is available in both centre-stable (Type 51M) and each-side-stable (Type 51A) forms. In the latter, the armature remains in contact with the side to which it was last operated until the relay is energized in the reverse direction, when a snap-over action takes place. The sensitivity of the 51A is quoted as 60 μW, and that of the 51M as 10 μW. Both relays are suitable for use in aircraft.



Telephone Manufacturing Co. Ltd.,
*Dulwich, London, S.E.*21.

Instrument Cathode-Ray Tubes

Two new instrument cathode-ray tubes have been introduced by Electronic Tubes Ltd. They are a 5 in.-diameter precision oscillograph tube with a two-stage distributed post-deflection accelerator (Type 5BKP1), and an inexpensive $2\frac{3}{4}$-in. tube of unusually high sensitivity (Type 3AFP1).

Some details from the makers' specification are given below:

5*BKP*1 (5-*in. circular screen*)

With a p.d.a. ratio of 5·5: 1, the maximum pattern distortion and maximum deviation from deflection linearity are 2%. Y-plate sensitivity for an overall voltage of 10 kV is 0·8 mm volt. The screen uses medium-persistence P1 (green) phosphor and has a metal backing. Heater ratings: 6·3 V, 0·55 A; capacitances x_1 to x_2, 2·3 pF; y_1 to y_2, 1·7 pF.

Typical operating conditions: V_{a1}, 1,400 V, V_g, −45 to −90 V; V_{a2}, 440 to 560 V; V_{a3}, 1,800 V; S_x, 26·5 V/cm; S_y, 12·5 V/cm; V_{a4}, 4,000 V; V_{a5}, 10,000 V.

3*AFP*1 ($2\frac{3}{4}$-*in. circular screen*)

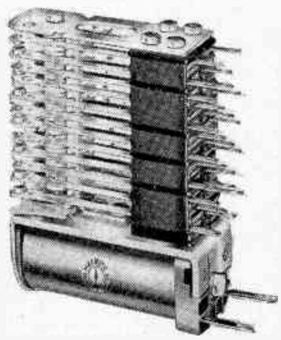
This tube is suitable for symmetrical or asymmetrical operation. Heater is 6·3 V, 0·55 A.

Typical operation conditions: V_{a1+a3}, 1,000 V; V_{a2}, 210–320 V; S_x, 20 V/cm; S_y, 11·5 V/cm; V_g, −28 to −65 V.
Electronic Tubes Ltd.,
Kingsmead Works, High Wycombe, Bucks.

New D.C. Relay

A compact relay, designated Series 285, has been designed by Magnetic Devices Ltd. It has a lifter comb and buffer block between stacks of contacts, and the clamped ends of the contact springs are encapsulated to ensure good insulation. The makers state that up to ten sets of light-duty change-over contacts, or up to eight sets of heavy-duty contacts, can be used, and that contact bounce is negligible. The sensitivity is such that with a 12,750-Ω coil and two sets of change-over contacts the operating current is 3 mA.

The Series 285 relay is available with



standard coil windings to suit a variety of common operating voltages. Coil resistances are up to 25,000 Ω.
Magnetic Devices Ltd.,
Exning Road, Newmarket, Suffolk.

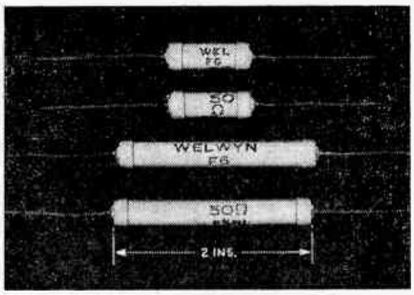
Small Ships' Radio

A t.r.f. type of radio receiver, the 'Heron' employing transistors, has been designed for use in such vessels as inshore fishing boats, coasters and yachts. It is also suitable for light aircraft. The receiver, which weighs 5½ lb. and measures 8 × 4⅞ × 2⅞ in., is said to operate for 500 hours from four Mallory hearing-aid cells when headphone reception is used, and half as long with loudspeaker reception. Frequency coverage is 150–255 kc/s, 250–420 kc/s, and 570–1,550 kc/s. Six transistors are used. There are two r.f. stages, the second of which can be made to oscillate for the purpose of c.w. reception. The receiver is hermetically sealed.

A simple d.f. device employing a ferrite loop aerial is also available. There are two versions to cover the frequency ranges 250–350 kc/s (shipping) and 190–420 kc/s (aircraft).
Brookes & Gatehouse Ltd.,
5 *Captains' Row, Lymington, Hants.*

Power Oxide Resistors

Medium-power rod-type resistors, in which the resistive element is a metal oxide bonded to a porcelain rod, are now available.



Advantages claimed are low price, resistance to moisture and mechanical damage, low reactance, and a reliability as good as that of comparable wire-wound resistors. Operation up to 250°C is permissible, provided that the power dissipation is suitably reduced above 40°C. The photograph shows 3-watt and 6-watt resistors, but 8-watt types are also made.

Points from the manufacturers' specification are given below:

Normal Tolerance: ± 5%.

Stability: Average resistance change less than 2% after 12 months storage. Less than 5% change after 1,000 hours' operation at full rated power.

Temperature Coefficient: Less than 0·035% per °C.

The maximum value of resistance is 10 kΩ–20 kΩ according to type, and the minimum is 10 Ω to 25 Ω. The inductance and capacitance of these resistors are stated to be negligible up to 10 Mc/s and, in general, the high-frequency characteristics are expected to be similar to those of high-stability carbon resistors with spiral tracks.
Welwyn Electrical Laboratories Ltd.,
Bedlington, Northumberland.

New Plessey Selector Unit

A new lightweight selector unit, known as type 9008, has been developed by The Plessey Company in collaboration with the Royal Aircraft Establishment. This is described as a remotely-controlled device for rotating a shaft to any one of twelve predetermined angular positions contained within one 320°-turn, and locking it in the selected position within 6 minutes of arc. It is currently being employed in u.h.f. transmitter/receivers for the selection of 12 pre-set channels by remote switching.

Primary movement of the unit, which has been granted R.A.E approval, is provided by an electric motor operated from a 48-volt d.c. supply, which may vary between ±6 volts. A relay is used to control the operation of the motor and to reverse its rotation. Selection of the required pre-set position is provided by means of a single-pole 12-position switch at the remote operating position.

Each of the angular positions can be set individually without affecting the remaining eleven settings, and facilities are provided at the rear of the unit for limiting the arc in which the twelve angular positions are contained. Accurate settings can be obtained by fine adjustments being made to the output or tuning shaft with a geared-down control knob. When not in use, this control knob can be securely housed in the front panel.

The unit is said to be capable of driving an external frictional load of 2¼ lb.
The Plessey Co. Ltd.,
Ilford, Essex.

Foamed Materials

Isocyanate compounds which can be made into flexible or rigid 'foams' are available from I.C.I. These are said to have good electrical properties, and to be resistant to mechanical shock. Among the uses suggested for isocyanate-castor oil compounds is the 'potting' of components.

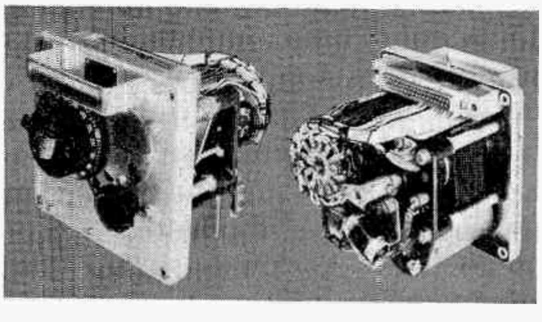
Rigid polyurethane foam compounds can be prepared by mixing two chemicals. The mixture can be poured into a cavity, where it will foam, set hard, and stick. The density of the final product can be controlled and can be as little as 2–3 lb. per cubic foot.

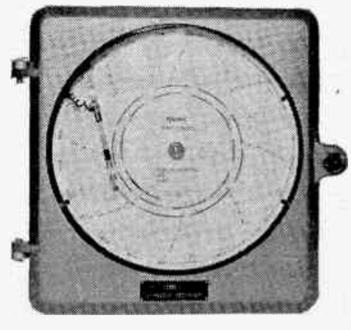
The new compounds can be applied as coatings to 'difficult' materials such as aluminium and nylon.
Imperial Chemical Industries Ltd.,
*Imperial Chemical House, Millbank, London, S.W.*1.

Operation Recorder

This instrument has been developed for recording the timing of operations and events, such as the on/off times of machines or electrical circuits. Twenty-four hour and seven-day charts are available, the periods of revolution being one hour and twenty-four hours, respectively.

A pair of contacts must be provided which close when it is desired to make a trace. An 'inkless pen' with a heated stylus is used in conjunction with heat-sensitive paper to produce a black line 1/16 in. wide. During off periods, enough power is supplied to produce a faint thin line; this enables the trace to be followed when no operations are recorded.
Fielden Electronics Ltd.,
Wythenshawe, Manchester, 22.





Abstracts and References

COMPILED BY THE RADIO RESEARCH ORGANIZATION OF THE DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH AND PUBLISHED BY ARRANGEMENT WITH THAT DEPARTMENT

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses. Copies of articles or journals referred to are not available from Electronic & Radio Engineer. *Application must be made to the individual publisher concerned.*

U.D.C. NUMBERS. Extensions and changes in U.D.C. numbers published in P.E. Notes, up to and including P.E. Note 609, will be introduced in Abstracts and References where applicable, notably the subdivisions of 621.372.8 "Waveguides", published in P.E. Note 594. U.D.C. publications are obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W.1, England.

JOURNAL REFERENCES. References to Russian publications will henceforward be based on the Russian titles, which will be abbreviated on the principles of The World List of Scientific Periodicals. The main changes are as follows: *Dokl. Ak. Nauk. S.S.S.R.* (formerly *C.R. Acad. Sci. U.R.S.S.*). *Izv. Ak. Nauk. S.S.S.R.* (formerly *Bull. Acad. Sci. U.R.S.S.*).

ACOUSTICS AND AUDIO FREQUENCIES

534.2-13 : 536.3 **1**

Effect of Heat Radiation on Sound Propagation in Gases.—P. W. Smith, Jr. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 693–698.)

534.21-8-14 **2**

Ultrasonic Propagation Measurement in Sea Water up to 400 kc/s.—T. Hashimoto & Y. Kikuchi. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 702–707.) Methods of measurement for horizontal and for vertical propagation are described. Measured values of absorption were 10–20 dB/km at 28 kc/s increasing to 120 dB/km at 400 kc/s.

534.23 **3**

Sound Transmission through Thin Cylindrical Shells.—P. W. Smith, Jr. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 721–729.) An analysis of the impedance presented by a thin cylindrical elastic shell to a pressure or normal stress as a function of the axial wavelength and the angular dependence of the forces.

534.23 : 537.227 **4**

Vibrations of Ferroelectric Cylindrical Shells with Transverse Isotropy: Part 1—Radially Polarized Case.—J. F. Haskins & J. L. Walsh. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 729–734.) Expressions for the coupled mechanical vibrations and electrical admittance of ferroelectric tubes having transverse isotropy are derived and the results supported by experimental data.

534.232 **5**

Experimental Investigation of Conical Horns Used with Underwater Sound Transducers.—W. R. Owens & C. M. McKinney. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 744–748.) A simple conical horn lined with pressure-release material is very effective in increasing the axial sensitivity of transducers and improving directivity. Design curves are given.

534.232 : 534.24 **6**

Effect of a Reflecting Plane on the Power Output of Sound Sources.—U. Ingard & G. L. Lamb, Jr. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 743–744.) The effect of a rigid reflecting plane on the output from a sound source above it is expressed as a 'power amplification factor'. This is calculated for a number of elementary sound sources and power output is determined as a function of source height.

534.414 **7**

Acoustic Impedance of a Helmholtz Resonator at Very High Amplitude.—D. A. Bies & O. B. Wilson, Jr. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 711–714.) A Helmholtz resonator terminating a 10-in.-diameter tube has been investigated for sound pressure levels 100–170 dB. With increasing sound pressure the same general rise in acoustic resistance and resonance frequency was observed for two different mountings.

534.612 **8**

Acoustical Radiation Pressure due to Incident Plane Progressive Waves on Spherical Objects.—G. Maidanik. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 738–742.)

534.75 **9**

On the Fusion of Sounds Reaching Different Sense Organs.—D. E. Broadbent & P. Ladefoged. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 708–710.) Experiments with synthetically produced speech show that fusion occurs when the first formant is presented to one ear and the second formant to the other, but not if the formants are given different fundamental frequencies.

534.78 **10**
Constant-Ratio Rule for Confusion Matrices in Speech Communication.—F. R. Clarke. (*J. acoust. Soc. Amer.*, June 1957, Vol. 29, No. 6, pp. 715–720.) "Three experiments are reported which give support to an empirical rule which may be used for predicting the entries in a closed confusion matrix for any subset of items drawn from a master set of items with a known confusion matrix."

534.78 **11**
The Intelligibility of Synthetic Speech.—O. Warns. (*Frequenz*, June 1957, Vol. 11, No. 6, pp. 169–175.) The results of subjective tests on the intelligibility of synthetic logatoms are analysed. For the German language 700–1 100 interchangeable speech elements would be adequate, requiring an information rate of 40 bits/s.

621.395.625.3 : 621.397.5 **12**
Video Tape Recorder Design.—(See 294.)

AERIALS AND TRANSMISSION LINES

621.315.212 **13**
The Characteristic Impedance of Coaxial Cables.—F. Raggi. (*Note Recensioni Notiz.*, March/April 1957, Vol. 6, No. 2, pp. 174–188.) Note summarizing the factors which determine the characteristic impedance of coaxial cables.

621.315.212 : 621.397.24 **14**
A New 4-Mc/s Coaxial Line Equipment—C.E.L. No. 6A.—M. E. Collier & W. G. Simpson. (*P.O. elect. Engrs' J.*, April 1957, Vol. 50, Part 1, pp. 24–34.) The equipment described is suitable for the transmission of either 16 telephony supergroups (960 circuits) or a 405-line television signal.

621.315.212 : 621.397.24 **15**
A 4-Mc/s Coaxial Line Equipment—C.E.L. No. 4A.—E. Davis. (*P.O. elect. Engrs' J.*, July 1957, Vol. 50, Part 2, pp. 92–97.) A description of equipment suitable for the transmission of either 17 telephony supergroups (1 020 circuits) or a television signal of 3-Mc/s video bandwidth. See also 14 above.

621.372.2 **16**
Return Loss: Part 1.—T. Roddam. (*Wireless World*, Nov. 1957, Vol. 63, No. 11, pp. 521–524.) The use of the concept of return loss instead of standing wave ratios is recommended in dealing with reflections of short pulses on long loss-free incorrectly terminated lines.

621.372.8 : 621.372.5 **17**
Noisy and Noise-Free Two-Port Networks Treated by the Isometric Circle Method.—E. F. Bolinder. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1412–1413.)

621.372.823 **18**
Circular Electric Wave Transmission through Serpentine Bends.—H. G. Unger. (*Bell Syst. tech. J.*, Sept. 1957, Vol. 36, No. 5, pp. 1279–1291.) An analysis of the propagation in a circular waveguide with equally spaced discrete supports and deformed elastically into a serpentine bend under its own weight.

621.372.831.4 **19**
Slots in an Imperfectly Conducting Waveguide.—B. Chatterjee. (*Indian J. Phys.*, May 1957, Vol. 40, No. 5, pp. 278–282.) The shunt conductance of a slot initially increases with departure from perfect conductivity of the guide wall but decreases rapidly in almost direct proportion to further decreases in conductivity.

621.372.852.21 **20**
Theory of Curved Circular Waveguide containing an Inhomogeneous Dielectric.—S. P. Morgan. (*Bell Syst. tech. J.*, Sept. 1957, Vol. 36, No. 5, pp. 1209–1251.) General equations are derived for all modes in a curved circular waveguide containing an inhomogeneous dielectric. The results are applied to the problem of preventing mode conversion from TE_{01} to TM_{11}. Design equations are given for various compensators with criteria for keeping the power level of spurious modes at a minimum.

621.372.853.1 **21**
Circular Electric Wave Transmission in a Dielectric-Coated Waveguide.—H. G. Unger. (*Bell Syst. tech. J.*, Sept. 1957, Vol. 36, No. 5, pp. 1253–1278.) The mode conversion from the TE_{01} wave to the TM_{11} wave in a curved circular waveguide may be reduced by applying a thin uniform dielectric coating to the inner wall. Design curves are given for both uniform bends of small radius and for large bending radii which may occur in a normally straight guide.

621.372.853.1 **22**
Normal Mode Bends for Circular Electric Waves.—H. G. Unger. (*Bell Syst. tech. J.*, Sept. 1957, Vol. 36, No. 5, pp. 1292–1307.) The TE_{01}-TM_{11} degeneracy is removed by a thin dielectric coating, and a further reduction in the mode conversion at a bend is obtained by tapering the curvature along the guide. Details are given of the bend loss for $\lambda = 5{\cdot}4$ mm. See also 21 above.

621.396.674 : 621.396.93 **23**
A Versatile Multiport Biconical Antenna.—R. C. Honey & E. M. T. Jones. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1374–1383.) This aerial can be used as a wide-band direction finder, as it excites in the feeding waveguide modes whose ratios are azimuth dependent. It also has uses as a multiplexer.

621.396.677 **24**
Bi-directional F.M. Aerials.—H. B. Dent. (*Wireless World*, Nov. 1957, Vol. 63, No. 11, pp. 534–536.) Description of stacked horizontal half-wave dipoles which are simple to construct and erect.

621.396.677.43 **25**
Rhombic Aerials.—F. J. Norman & J. F. Ward. (*Electronic Radio Engr*, Nov. 1957, Vol. 34, No. 11, pp. 398–403.) The design of aperiodic rhombic aerials is examined and a method derived whereby a set of charts of open scale yields all the aerial parameters of practical significance for high-frequency operation. The angle of fire and the gain with respect to a dipole in free space are displayed for an adequate range of aerial side lengths and included angles. Corrections are given for the height above ground and a simple method to find the shape of the main lobe is suggested.

621.396.677.8 : 523.16 **26**
The Mullard Radio Astronomy Observatory, Cambridge.—Ryle. (See 107.)

621.396.677.833 : 523.16 **27**
World's Largest Radio Telescope.—(See 108.)

621.396.677.85 **28**
Calculated Intensity and Phase Distribution in the Image Space of a Microwave Lens.—G. W. Farnell. (*Canad. J. Phys.*, June 1957, Vol. 35, No. 6, pp. 777–783.) Intensity and constant-phase contours and an energy-flow diagram are shown for a $3{\cdot}2$-cm-λ lens with exit pupil of 25 cm radius distant 105 cm from the image point. Calculations are based on scalar diffraction theory. Results of measurements show close agreement.

621.396.677.85 **29**
Study of Optical Diffraction Images at Microwave Frequencies.—M. P. Bachynski & G. Bekefi. (*J. opt. Soc. Amer.*, May 1957, Vol. 47, No. 5, pp. 428–438.) "Experimental investigations of the intensity distribution in the region of the focus of microwave lenses are described. The circularly symmetric lenses, made from polystyrene plastic, were illuminated by a linearly polarized beam of radiation of $1{\cdot}25$-cm wavelength. Most of the measurements are presented in the form of intensity contours in planes both containing the principal ray and lying perpendicular to it. Images formed by nearly perfect optical systems and by systems suffering from various third-order monochromatic aberrations were examined. In most cases the results are compared with calculations made from the scalar-diffraction theory of optical systems. In general, good agreement between theory and experiment is found."

AUTOMATIC COMPUTERS

681.142 **30**
Controlling the Digital Computer.—R. W. Hamming. (*Sci. Mon.*, Oct. 1957, Vol. 85, No. 4, pp. 169–175.) A non-mathematical discussion of the programming and application of computers for scientific purposes.

681.142 **31**
An Experimental 50-Mc/s Arithmetic Unit.—R. M. Walker, D. E. Rosenheim, P. A. Lewis & A. G. Anderson. (*IBM J. Res. Developm.*, July 1957, Vol. 1, No. 3, pp. 257–278.) Detailed description of a unit which performs a repetitive multiplication program and checks the results for errors.

681.142 **32**
The Design of the Ferranti Pegasus Computer: Part 2.—G. Emery. (*Electronic Engng*, Sept. 1957, Vol. 29, No. 355, pp. 420–425.) The detailed engineering aspects of the design problem are discussed. Part 1: 3753 of 1957 (Braunholtz).

681.142 **33**
A 32 000-Word Magnetic-Core Memory.—E. Foss & R. S. Partridge. (*IBM J. Res. Developm.*, April 1957, Vol. 1, No. 2, pp. 102–109.) The storage matrix consists of $128 \times 256 \times 36$ ferrite cores and has a read-write cycle of 12 μs. Its construction is described, together with the driver and sense amplifier circuits.

681.142 **34**
A Positive-Integer Arithmetic for Data Processing.—R. W. Murphy. (*IBM J. Res. Developm.*, April 1957, Vol. 1, No. 2, pp. 158–170.)

681.142 **35**
Irredundant Disjunctive and Conjunctive Forms of a Boolean Function.—M. J. Ghazala (Gazale). (*IBM J. Res. Developm.*, April 1957, Vol. 1, No. 2, pp. 171–176.) A thorough algebraic method is described for determining the complete set of irredundant normal and conjunctive forms. The procedure is mechanical and the method can be applied for programming computers.

681.142 **36**
Electronic Computers.—W. Taeger. (*Frequenz*, June 1957, Vol. 11, No. 6, pp. 185–191.) The application of analogue computers to the solution of control-system problems is discussed.

681.142 **37**
Approximating Nonlinear Functions by Shunt-Loading Tapped Potentiometers in Analogue Computing Machines.—D. W. C. Shen. (*Electronic Engng*, Sept. 1957, Vol. 29, No. 355, pp. 434–439.)

681.142 : 621.314.7 **38**
The Multipurpose Bias Device: Part 1—The Commutator Transistor.—B. Dunham. (*IBM J. Res. Developm.*, April 1957, Vol. 1, No. 2, pp. 116–129.) A study of the application of the Rutz commutator transistor to three-input, one-output logical problems for handling by automatic computer systems. The Rutz commutator is a single-emitter amplitude-sensitive device with three separate input wires applied to the emitter. Its design will be detailed in a subsequent paper.

681.142 : 621.314.7 **39**
Two-Collector Transistor for Binary Full Addition.—Rutz. (See 315.)

681.142 : 621.395.625.3 **40**
A Mathematical Model for Determining the Probabilities of Undetected Errors in Magnetic Tape Systems.—M. Schatzoff & W. B. Harding. (*IBM J. Res. Developm.*, April 1957, Vol. 1, No. 2, pp. 177–184.)

681.142 : 621.395.625.3 **41**
The Magnetic Tape Store for Pegasus.—T. G. H. Braunholtz & D. Hogg. (*Electronic Engng*, Oct. 1957, Vol. 29, No. 356, pp. 484–489.)

681.142 : 621.395.625.3 **42**
The Recording of Digital Information on Magnetic Drums.—D. G. N. Hunter & D. S. Ridler. (*Electronic Engng*, Oct. 1957, Vol. 29, No. 356, pp. 490–496.) "Methods of representing binary digital information on magnetic drums are discussed with regard to their reliability, cost and technical merits."

CIRCUITS AND CIRCUIT ELEMENTS

621.3 : 061.3 **43**
International Components Symposium.—(*Electronic Radio Engr*, Nov. 1957, Vol. 34, No. 11, pp. 428–431.) Summary of papers read at the Royal Radar Establishment, Malvern, in September 1957. Reliability, efficient use of space, liquid cooling techniques and optimum component shapes for automatic assembly are discussed.

621.3.011.3 **44**
Calculation of Inductance of Toroids with Rectangular Cross Section and Few Turns.—R. F. Schwartz. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1416–1417.)

621.3.049.7 : 621.39.03 : 061.3 **45**
'Solid Circuits.'—(*Wireless World*, Nov. 1957, Vol. 63, No. 11, pp. 516–517.) Discussion of the International Symposium on Electronic Components held at Malvern, England. A 'solid circuit' forming the equivalent of several transistors connected by resistors and capacitors is obtained by depositing films of conductive, resistive and dielectric materials on a small block of semiconductor material. The solution of problems regarding the operation of components at high temperatures is mentioned.

621.3.049.75 **46**
Printed Circuits in Receiver Construction.—W. I. Flack. (*J. Telev. Soc.*, Jan./March 1957, Vol. 8, No. 5, pp. 176–185.) Outline of development of the printed circuit, and description of modern production techniques.

621.3.049.75 **47**
Pro & Con on Seven Different Methods of Printed Wiring.—A. E. Stones. (*Electronic Ind. Tele-Tech*, March 1957, Vol. 16, No. 3, pp. 64–66..158.) The basic techniques for each method are described and compared. A table of relevant U.S. patents is included.

621.314.6 **48**
A General Circuit Theorem on Rectification.—J. W. Gewartowski. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, p. 1410.) Nonlinear capacitors and inductors cannot effect rectification unless nonlinear resistance is present in the circuit.

621.318.424 **49**
New Method of Evaluating Ferro-inductors.—W. J. Polydoroff. (*Electronic Ind. Tele-Tech*, April 1957, Vol. 16, No. 4, pp. 78–79..150.) Formulae for calculating the insertion loss and the resistance increase due to the iron core, are derived. The practical factors governing optimum coil design are investigated.

621.318.57 : 621.387 **50**
Cold-Cathode Voltage-Transfer Circuits.—J. H. Beesley. (*G.E.C. Telecommun.*, Feb. 1957, No. 23, pp. 6–19.) Practical techniques used in the application of cold-cathode triodes are given in detail. A number of circuit diagrams is included.

621.319.4 : 537.311.33 **51**
Charge and Discharge of a Nonlinear Condenser through a Linear Nondissipative Inductance.—A. Mozumder. (*Z. angew. Math. Phys.*, 25th July 1957, Vol. 8, No. 4, pp. 261–280.) Macdonald (*J. Chem. Phys.*, Aug. 1954, Vol. 22, No. 8, pp. 1317–1322) has shown that for a capacitor with a semiconductor as dielectric the capacitance is proportional to $\frac{\sinh \alpha V}{\alpha V}$ where V is the voltage across the capacitor and α is a constant. A circuit is analysed in which such a capacitor is connected in series with a lossless inductance and a direct-voltage source, and a comparison is made with an exponential capacitor ($C \propto \exp|\alpha V|$). Applications to the generation of rectangular and triangular waveforms are indicated. See also 958 of 1955 (Macdonald & Brackman).

621.319.42 : 621.3.049.75 **52**
Evaluating Base Materials for Printed Capacitors.—J. J. Logan. (*Electronic Ind. Tele-Tech*, Aug. 1957, Vol. 16, No. 8, pp. 72–73..160.) Definition of terms specifying the losses of base materials used for printed circuits, which affect the size and Q factor of printed capacitors.

621.372.412.002.2 : 549.514.51 : 621.793.14 **53**
A Vacuum Plant for the Frequency Calibration of Quartz Vibrators.—J. Green, L. Holland & B. D. Power. (*Electronic Engng*, Sept. 1957, Vol. 29, No. 355, pp. 440–444.) See also 503 of 1956 (Awender et al.).

621.372.5 **54**
A Network Theorem.—E. Green. (*Marconi Rev.*, 2nd Quarter 1957, Vol. 20, No. 125, pp. 35–38.) The impedance level of half a symmetrical network can be changed without altering the frequency response; the use of this theorem for the solution of network problems is discussed and illustrated by an example.

621.372.5 : 621.372.8 **55**

Noisy and Noise-Free Two-Port Networks Treated by the Isometric Circle Method.—E. F. Bolinder. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1412–1413.)

621.372.5 : 621.375.2.018.75 **56**

On the Design of Four-Terminal Interstages for Pulse Applications.—A. K. Choudhury & N. B. Chakrabarti. (*Indian J. Phys.*, April 1957, Vol. 40, No. 4, pp. 193–210.) The limits set by the input and output capacitance on the time response, and the conditions for no overshoot are determined. Methods for achieving the necessary pole distributions are described and circuit networks suggested.

621.372.52 : 621.314.7 **57**

The Mathematical Treatment of Transistor Feedback Circuits.—W. Glaser. (*NachrTech.*, April 1957, Vol. 7, No. 4, pp. 159–162.) Formulae are tabulated giving in convenient form the parameters of transistor networks for four types of feedback circuit.

621.372.54.029.64 **58**

Design and Development of Strip-Line Filters.—E. H. Bradley. (*Trans. Inst. Radio Engrs*, April 1956, Vol. MTT-4, No. 2, pp. 86–93. Abstract, *Proc. Inst. Radio Engrs*, July 1956, Vol. 44, No. 7, p. 956.)

621.372.543.2 : 538.652 **59**

Concentric-Shear-Mode 455-kc/s Electromechanical Filter.—R. W. George. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 186–194.) An experimental filter of simplified design is described which consists of four magnetostrictive ferrite disk resonators.

621.372.55 : 621.396.4 : 621.376.3 **60**

A Broad-Band Variable Group-Delay Equalizer.—R. Hamer & R. G. Wilkinson. (*P.O. elect. Engrs' J.*, July 1957, Vol. 50, Part 2, pp. 120–123.) The equalizer has been developed for use in f.m. microwave systems. It is inserted in the 70-Mc/s i.f. signal path in long radio-relay systems. The performance is adequate for the transmission of frequency-division-multiplex telephony of at least 600 channels, or subcarrier colour-television transmission.

621.373.029.64 : 538.569.4 **61**

Comments on Frequency Pulling of Maser Oscillators.—C. H. Townes. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, pp. 920–921.) The physical origin is illustrated of the theoretically predicted reduction in frequency-pulling with increasing number of beam molecules, and reasons are given for its failure to occur in practice.

621.373.029.64 : 538.569.4 :538.221 **62**

A Solid-State Microwave Amplifier and Oscillator using Ferrites.—M. T. Weiss. (*Phys. Rev.*, 1st July 1957, Vol. 107, No. 1, p. 317.) Experimental results obtained with an oscillator based on the proposal by Suhl (3076 of 1957). The active element is a piece of magnetized ferrite at room temperature.

621.373.421 **63**

Sensitivity of a Self-Excited Oscillator.—I. S. Shpigel', M. D. Raĭzer & E. A. Myae. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 387–390.) Theoretical and experimental investigation of the characteristics of a valve feedback oscillator suitable for nuclear resonance detection.

621.373.5 : 621.396.621 **64**

Crystal Oscillators in Communication Receivers.—A. G. Manke. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. PGVC-7, pp. 10–15. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1957, Vol. 45, No. 2, p. 258.)

621.373.52 : 621.373.431.1 **65**

Unijunction Transistor forms Flip-Flop.—E. Keonjian & J. J. Suran. (*Electronics*, 1st Sept. 1957, Vol. 30, No. 9, pp. 165–167.) The construction and performance are described of a multivibrator circuit for astable or monostable operation (see also 2878 of 1955). Its active element is the double-base diode, a 3-terminal *p-n* junction device [see 3562 of 1956 (Suran)].

621.373.52.018.756 **66**

Pulse Generator uses Junction Transistors.—E. J. Fuller. (*Electronics*, 1st Sept. 1957, Vol. 30, No. 9, pp. 176–179.) A transistor circuit is described giving pulses of 1–10 μs duration at repetition rates of 50–5 000 pulses/s with internal delays of 1–100 μs.

621.373.52.018.78 **67**

Distortion in Transistor Amplifiers.—P. Tharma. (*Mullard tech. Commun.*, March 1957, Vol. 3, No. 22, pp. 62–64.) An analysis of the distortion in low-level grounded-emitter stages due to input-circuit nonlinearity and to variations of collector-to-base current gain.

621.373.52.029.4 **68**

Low-Frequency Transistor Oscillators.—L. G. Cripps. (*Mullard tech. Commun.*, March 1957, Vol. 3, No. 22, pp. 44–58; *Electronic Applic. Bull.*, May 1957, Vol. 17, No. 3, pp. 113–128.) Theoretical treatment of a junction-transistor oscillator for frequencies of a few kc/s. Frequency and amplitude stability, conversion efficiency and the conditions governing starting, are considered. A practical circuit design is derived and experimental results are given.

621.374.32 **69**

Build-Up of Large Signals with Elimination of Reflections in Magnetostrictive Storage Lines.—D. Maeder. (*Z. angew. Math. Phys.*, 25th July 1957, Vol. 8, No. 4, pp. 326–327.) Greater reliability and increased storage capacity are obtained by substituting twelve energizing coils for the usual one. With this arrangement, a steel-wire delay line can be used. See also 2385 of 1957.

621.374.33 : 621.314.7 **70**

A Transistor Gating Matrix for a Simulated Warfare Computer.—W. H. MacWilliams, Jr. (*Bell Lab. Rec.*, March 1957, Vol. 35, No. 3, pp. 94–99.) Brief description of a circuit, devised in 1949, which used a matrix of 4 × 10 point-contact transistors. Amplification of signals was also achieved.

621.374.5 **71**

Signal-Enhanced Delay Line.—T. I. Humphreys. (*Electronic Ind. Tele-Tech*, Aug. 1957, Vol. 16, No. 8, pp. 70–71 . . 167.) To maintain pulse shape the high-frequency response of the delay line described is improved by selective amplification.

621.374.5 : 621.396.962.3 **72**

Pulse Compression.—R. Krönert. (*NachrTech.*, April 1957, Vol. 7, No. 4, pp. 148–152, 162.) A method is described for compressing frequency-modulated r.f. pulses, i.e. shortening them and increasing their amplitude, by passing them through a delay-distortion network consisting of a chain of lattice quadripoles. The theory outlined is based on work by Cauer, published posthumously. Its practical application in the field of radar should result in higher resolving power for a given signal/noise ratio.

621.375 : 621.372.51 **73**

Minimizing Mismatch Loss.—H. E. Hollmann. (*Electronic Ind. Tele-Tech*, Aug. 1957, Vol. 16, No. 8, pp. 74–75 . . 165.) Simple formulae and curves are given.

621.375.2 + 621.375.4 **74**

Upper Limits of Output Power in Vacuum-Tube and Transistor A.C. Amplifiers.—L. M. Vallese. (*Commun. & Electronics*, March 1957, No. 29, pp. 87–92.) Simple analysis, and comparison based on idealized characteristics.

621.375.2 **75**

Stacked Valve Circuits.—J. B. Earnshaw. (*Electronic Radio Engr*, Nov. 1957, Vol. 34, No. 11, pp. 404–406.) An analysis by means of feedback theorems and modified equivalent circuits. Application of the method to the cascode amplifier and the cathode-follower circuits is given in detail.

621.375.2.029.3 **76**

Single-Ended Push-Pull Output Stages.—(*Electronic Applic. Bull.*, May 1957, Vol. 17, No. 3, pp. 81–106.) The development of low-d.c.-resistance pentodes and high-resistance moving-coil loudspeakers has made possible the construction of transformerless output stages several types of which are described. The paper is based on the work of W. Aschermann and J. Rodrigues de Miranda (see 3727 of 1957).

621.375.2.029.4 **77**

A Low-Frequency Selective Amplifier.—C. K. Battye. (*J. sci. Instrum.*, July 1957, Vol. 34, No. 7, pp. 263–265.) The frequency is variable in steps from 1 to 1 000 c/s. The selectivity can be set up to $Q = 100$. The amplifier is d.c. coupled; a twin-T network shunted by a capacitor acts as a selective network in the negative feedback line.

621.375.2.029.55 **78**

PT-15 P.A. Unit.—J. N. Walker. (*Short Wave Mag.*, Feb. 1957, Vol. 14, No. 12, pp. 624–630.) Description of a band-switching r.f. amplifier for 80, 40 and 20 m, which uses a pair of valves Type PT 15 in parallel.

621.375.221.029.62 : 621.396.4 : 621.376.3 **79**

A Broad-Band Intermediate-Frequency Amplifier for Use in Frequency-Modulation Microwave Radio-Relay Systems.—R. Hamer & C. H. Gibbs. (*P.O. elect. Engrs' J.*, July 1957, Vol. 50, Part 2, pp. 124–126.) A 70-Mc/s amplifier is described which is suitable for the transmission of 600-channel frequency-division-multiplex telephony or subcarrier colour-television signals in f.m. microwave radio-relay systems. A test-tone/noise ratio of 70 dB is not significantly degraded after transmission through five amplifiers in tandem.

621.375.227.029.3 **80**

400-Watt Audio Amplifier.—G. R. Woodville. (*Wireless World*, Nov. 1957, Vol. 63, No. 11, pp. 543–546.) The circuit described includes small output valves in parallel push-pull connection.

621.375.3 **81**

Self-Balancing Magnetic Amplifiers of the Differential-Feedback Type.—W. A. Geyger. (*Commun. & Electronics*, March 1957, No. 29, pp. 39–46.) A 60-c/s magnetic amplifier for extending the range of a moving-coil ink recorder is described.

621.375.4.029.3 : 629.13 **82**

The Use of Transistors in Airborne Audio Equipment.—V. P. Holec. (*Trans. Inst. Radio Engrs*, July/Aug. 1956, Vol. AU-4, No. 4, pp. 90–93. Abstract, *Proc. Inst. Radio Engrs*, Nov. 1956, Vol. 44, No. 11, p. 1641.)

621.375.43.029.3 : 621.314.7 **83**

Negative-Feedback Transistor Amplifiers.—H. R. Lowry. (*Radio Telev. News*, May 1957, Vol. 57, No. 5, pp. 55–58 .. 109.) Design data are given for high-fidelity a.f. amplifiers.

621.375.9.029.64 : 538.569.4 **84**

Microwave Amplification by MASER Techniques.—W. V. Smith. (*IBM J. Res. Developm.*, July 1957, Vol. 1, No. 3, pp. 232–238.) Elementary analysis of maser operation, including its application to wide-band, short-transit-time amplification.

621.376 : 621.316.7.078 **85**

Demodulator-Limiter for Control System Signals.—N. L. Johanson. (*Electronics*, 1st Sept. 1957, Vol. 30, No. 9, p. 155.) Description of transistor circuit for automatic control systems which acts as limiter during modulation or demodulation.

621.376.239 **86**

Magnetically Keyed, Phase-Sensitive Demodulators.—R. B. Mark, W. X. Johnson & P. R. Johannessen. (*Commun. & Electronics*, March 1957, No. 29, pp. 1–6.) The operation and performance of a pulse-shaping circuit using magnetic reactors is described. The properties of three types of phase-sensitive demodulator are compared. Diode demodulators have linear characteristics, magnetic amplifier demodulators have high input impedance, and transistor demodulators combine good linearity with low noise level.

621.376.4 : 621.314.7 **87**

Transistorized Phase Discriminators.—A. N. De Sautels. (*Commun. & Electronics*, March 1957, No. 29, pp. 19–26.) Half-wave and full-wave phase discriminators are described with two- or three-terminal d.c. outputs. An analysis of the operating characteristics of transistors in discriminator circuits is given; temperature stability and power limitations are discussed.

GENERAL PHYSICS

535.33-1 **88**

Interferometry for the Far Infrared.—J. Strong. (*J. opt. Soc. Amer.*, May 1957, Vol. 47, No. 5, pp. 354–357.)

535.61-1 **89**

Infrared Radiation and its Detection.—(*Electronic Radio Engr*, Nov. 1957, Vol. 34, No. 11, pp. 412–415.) A comparison of the sensitivity and time constant of available detectors of infrared radiation. Thermal detectors of thermocouple, bolometer and pneumatic type and photoelectric detectors are discussed.

537.311.1 : 536.21 : 537.311.33 **90**

The Lorenz Number.—P. J. Price. (*IBM J. Res. Developm.*, April 1957, Vol. 1, No. 2, pp. 147–157.) The theory of the Lorenz number of a conducting crystal is developed for the 'band' and 'one-electron' models of the electron assembly. With certain approximations, the Lorenz number is found to be equal to the mean square fluctuation of the thermoelectric power.

537.311.1+536.212.2] : 539.13 **91**

Some Scattering Problems in Conduction Theory.—P. G. Klemens. (*Canad. J. Phys.*, April 1957, Vol. 35, No. 4, pp. 441–450.) An expression derived for the scattering of electrons by inhomogeneous slowly-varying strain fields is applied to scattering by dislocations and stacking faults.

537.5 **92**

Free-Path Formulae for the Coefficient of Diffusion and Velocity of Drift of Electrons in Gases.—L. G. H. Huxley. (*Aust. J. Phys.*, March 1957, Vol. 10, No. 1, pp. 118–129. Correction, *ibid.*, June 1957, Vol. 10, No. 2, p. 329.)

537.5 : 538.6 **93**

Type of Plasma Oscillations in a Magnetic Field.—S. I. Braginski. (*Dokl. Ak. Nauk S.S.S.R.*, 21st July 1957, Vol. 115, No. 3, pp. 475–478.) Mathematical analysis based on a model of ideal electron and ion gases.

537.5 : 538.63 **94**

Free-Path Formulae for the Electronic Conductivity of a Weakly Ionized Gas in the Presence of a Uniform and Constant Magnetic Field and a Sinusoidal Electric Field.—L. G. H. Huxley. (*Aust. J. Phys.*, June 1957, Vol. 10, No. 2, pp. 240–245.) "A general free path formula is given for the drift velocity of electrons in a weakly ionized gas in a sinusoidal electric field. Most special cases of interest, including the magnetic deflection of an electron stream in a gas, are readily derivable from the general formula. The results find application in microwave and ionospheric studies of the motion of electrons in gases as well as in experiments on the magnetic deflection of an electron stream." See also 92 above.

537.525.8 **95**

Oscillations in Direct-Current Glow Discharges.—A. M. Pilon. (*Phys. Rev.*, 1st July 1957, Vol. 107, No. 1, pp. 25–27.) Experimental results are given concerning the stable oscillations with the same period which occur in the current and in the light emitted by the positive column.

538.244.2 **96**

Magnetization in the Presence of Magnetic Viscosity.—M. I. Rozovski. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 355–359.) A mathematical expression is derived for solving problems of magnetization of magnetically viscous samples, in particular those of cylindrical shape. See also 3697 of 1956 (Tikhonov & Samarski).

538.566 : 535.42 **97**

Diffraction of Electromagnetic Waves by Ribbon and Slit: Part 1.—Y. Nomura & S. Katsura. (*J. phys. Soc. Japan*, Feb. 1957, Vol. 12, No. 2, pp. 190–200.) Rigorous solutions are presented for a wave incident on a plane normal to the edge of the ribbon or slit, but with arbitrary angles of incidence and polarization. For diffraction by circular plate and hole, see 3235 of 1955.

538.566 : 535.422]+534.26 **98**

Diffraction of a Spherical-Wave Pulse by a Half-Plane Screen.—J. R. Wait. (*Canad. J. Phys.*, May 1957, Vol. 35, No. 5, pp. 693–696.) Fourier-integral treatment for a point source.

538.569.4 **99**

Distortions of Magnetic Resonances by Additional and Excessive Oscillatory Fields.—H. R. Lewis, A. Pery, W. Quinn & N. F. Ramsey. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 446–449.) A theoretical explanation is given of the shift of the resonance frequency which occurs as a result of the presence of an extraneous r.f. electric field. See also 1386 of 1956 (Ramsey).

538.569.4 : 539.15.098 **100**

The Dependence of the Amplitude of the First Harmonic of Nuclear-Magnetic-Resonance Absorption on the Amount of Detuning.—I. S. Shpigel', M. D. Raĭzer & E. A. Myae. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 351–354.)

538.569.4 : 546.171.1 **101**

Equivalence between the Line Width and the Rate of Energy Relaxation in the Inversion Spectrum of Ammonia.—K. Tomita. (*Progr. theor. Phys.*, March 1957, Vol. 17, No. 3, pp. 513–515.)

538.6 **102**

On the Equilibrium Configurations of Oblate Fluid Spheroids under the Influence of a Magnetic Field.—S. P. Talwar. (*Proc. nat. Inst. Sci., India*, Part A, 26th Sept. 1956, Vol. 22, No. 5, pp. 316–323.)

538.6 **103**

Toroidal Oscillations of a Spherical Mass of Viscous Conducting Fluid in a Uniform Magnetic Field.—K. Stewartson. (*Z. angew. Math. Phys.*, 25th July 1957, Vol. 8, No. 4, pp. 290–297.) See also 3352 of 1956.

539.1 : 541.124 **104**

Theoretical Treatment of the Kinetics of Diffusion-Limited Reactions.—T. R. Waite. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 463–470.) The diffusion-limited reaction $A + B \rightarrow AB$ has been solved for a random initial distribution of A and B particles.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 : 551.510.535 **105**

Radio-Star Ridges.—M. Dagg. (*J. atmos. terr. Phys.*, 1957, Vol. 11, No. 2, pp. 118–127.) The distinctive amplitude variations known as 'ridges' can probably be explained in terms of divergent-lens effects in the E region. See also 2058 of 1956 (Wild & Roberts).

523.16 : 551.510.535 **106**

A Consideration of Radio-Star Scintillations as Caused by Interstellar Particles Entering the Ionosphere: Part 1—Daily and Seasonal Variations of the Scintillation of a Radio Star. Part 2—The Accretion of Interstellar Particles as a Cause of Radio-Star Scintillations.—G. A. Harrower. (*Canad. J. Phys.*, May 1957, Vol. 35, No. 5, pp. 512–535.) An analysis is made of measurements of scintillations of the r.f. source in Cassiopeia recorded at Ottawa during 1954 at a frequency of 50 Mc/s. The data show certain daily maxima occurring at solar times dependent on the date of the year and grouped unsymmetrically in a way which suggests that they are due to the infall of interstellar particles from outside the solar system. The velocities of certain of these particles are derived by simple applications of vector addition employing the known velocity of the earth. See 2116 of 1957.

523.16 : 621.396.677.8 **107**

The Mullard Radio Astronomy Observatory, Cambridge.—M. Ryle. (*Nature, Lond.*, 20th July 1957, Vol. 180, No. 4577, pp. 110–112.) The observatory was formally opened in July 1957. Two instruments then under construction were a radio-star interferometer for $1 \cdot 7$ m λ, and a pencil-beam system for investigating continuous galactic radiation at $7 \cdot 9$ m λ. The aperture synthesis technique applied in the two systems is described. See also 3749 of 1957.

523.16 : 621.396.677.833 **108**

World's Largest Radio Telescope.—(*Overseas Engr*, Aug. 1957, Vol. 31, No. 356, pp. 15–18.) An illustrated description of the design and construction of the 250-ft fully steerable paraboloid of the Jodrell Bank instrument, with details of the driving system and drive control equipment.

523.5 **109**

The Distribution of the Orbits of Sporadic Meteors.—A. A. Weiss. (*Aust. J. Phys.*, March 1957, Vol. 10, No. 1, pp. 77–102.) Radio observations at Adelaide during 1952–1956 are analysed. A distribution of orbits is derived which is consistent with radar and visual observations in the northern hemisphere.

523.5 : 621.396.11.029.62 **110**

Variations in the Intrinsic Strength of the 1956 Quadrantid Meteor Shower.—C. O. Hines & E. L. Vogan. (*Canad. J. Phys.*, June 1957, Vol. 35, No. 6, pp. 703–711.) Analysis of the occurrence rate of meteoric signals detected on a v.h.f. forward-scatter east-west path in Canada. See also 3392 of 1955 (Forsyth and Vogan).

523.5 : 621.396.11.029.62 :551.510.535 **111**

A Theory of Long-Duration Meteor-Echoes based on Atmospheric Turbulence with Experimental Confirmation.—L. A. Manning & V. R. Eshleman. (*J. geophys. Res.*, Sept. 1957, Vol. 62, No. 3, pp. 367–371.) Comments on 1417 of 1957 (Booker & Cohen). The experimental evidence is examined with the conclusion that the theory does not accurately represent the properties of meteoric echoes.

523.5 : 621.396.96 **112**

Meteor Activity in the Southern Hemisphere.—A. A. Weiss. (*Aust. J. Phys.*, June 1957, Vol. 10, No. 2, pp. 299–309.) Activity and radiants of meteor showers are discussed and compared with similar data for the northern hemisphere. See also 109 above and 2938 of 1955.

523.74 : 551.510.535 **113**

Ionospheric Indices of Solar Activity.—W. J. G. Beynon & G. M. Brown. (*J. atmos. terr. Phys.*, 1957, Vol. 11, No. 2, pp. 128–131.) It is suggested that a simple index based on E-layer measurements at a single station is adequate, and that the additional effort required to provide the index I_{F_2}, based on F_2-layer measurements at three stations, is not justified. See also 1397 of 1956 (Minnis).

523.75 **114**

Observation of a New Type of Flare.—R. J. Bray, R. E. Loughhead, V. R. Burgess & M. K. McCabe. (*Aust. J. Phys.*, June 1957, Vol. 10, No. 2, pp. 319–323.) A mass of very bright material was ejected from a flare about 28° from the sun's limb. Except for the velocity, the mass possessed all the properties of a small flare.

550.372 **115**

The Effective Electrical Constants of Soil at Low Frequencies.—J. R. Wait. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1411–1412.) A possible explanation is presented for the large apparent dielectric constants ($\approx 10^3$) which occur at frequencies of the order of 15 kc/s.

550.389.2 **116**

I.G.Y. Rocket Program.—N. W. Spencer. (*Sci. Mon.*, Sept. 1957, Vol. 85, No. 3, pp. 130–142.) An outline of the work of the I.G.Y. subcommittee of the U.S. Technical Panel for Rocketry. A table is given listing organizations participating, and their objectives.

551.510.52 : 551.510.535 **117**

A Possible Troposphere-Ionosphere Relationship.—S. J. Bauer. (*J. geophys. Res.*, Sept. 1957, Vol. 62, No. 3, pp. 425–430.) A relation is suggested between F_2-layer characteristics and the passage of meteorological fronts through the troposphere. A hypothesis for dynamical coupling between the two regions is outlined.

551.510.53 **118**

Electric Field Measurements in the Stratosphere.—C. G. Stergis, G. C. Rein & T. Kangas. (*J. atmos. terr. Phys.*, 1957, Vol. 11, No. 2, pp. 77–82.) Above the exchange layer the field decreases monotonically up to the limit (90 000 ft) of measurements made in 1955 and 1956. The measured and calculated values are in good agreement.

551.510.53 : 551.594.21 **119**

Electric Field Measurements above Thunderstorms.—C. G. Stergis, G. C. Rein & T. Kangas. (*J. atmos. terr. Phys.*, 1957, Vol. 11, No. 2, pp. 83–90.) In measurements at heights of 70 000 to 90 000 ft above central Florida a positive upward-flowing current of about $1 \cdot 3$ A was found.

551.510.535 **120**

Models of the Lower Ionosphere as may be Inferred from Absorption Results.—P. Bandyopadhyay. (*Indian J. Phys.*, June 1957, Vol. 31, No. 6, pp. 297–308.) Values of deviative and nondeviative absorption and their variations with $\cos \chi$ are calculated for the D-region models of Nertney (2306 of 1953), Piggott (unpublished) and Mitra (see 2087 of 1954) and for the E-region model of Jones (1661 of 1955). Satisfactory agreement with experimental results is found for Mitra's model of the D region but not for Jones's model of the E region.

551.510.535 **121**

A Study of 'Spread-F' Ionospheric Echoes at Night at Brisbane: Part 3—Frequency Spreading.—D. G. Singleton. (*Aust. J. Phys.*, March 1957, Vol. 10, No. 1, pp. 60–76.) Diurnal and seasonal variations in spreading of the penetration frequency are compared with similar world-wide data. The observations are interpreted in terms of scattering from clouds of enhanced ionization near the F_2-layer maximum and a seasonal vertical movement of these clouds is suggested the extent of which increases with latitude. Parts 1 & 2: 119 of 1957 (McNicol et al.).

551.510.535 **122**
Severe Ionospheric Storm.—(*Wireless World*, Nov. 1957, Vol. 63, No. 11, p. 553.) Notes on the disturbances of 29th August–6th September 1957.

551.510.535 : 523.75 :621.396.11.029.62 **123**
Disturbances in the Lower Ionosphere Observed at V.H.F. following the Solar Flare of 23 February 1956 with Particular Reference to Auroral-Zone Absorption.—D. K. Bailey. (*J. geophys. Res.*, Sept. 1957, Vol. 62, No. 3, pp. 431–463.) The effects of the flare on ionospheric-scatter links in the band 30–40 Mc/s at high latitudes are reported. An abrupt increase in oblique-incidence signal intensity, occurring almost simultaneously with the arrival of cosmic rays, was followed during the next three days by abnormally high levels at night and low levels during the day. Background cosmic noise measurements showed increased absorption at night and greatly increased absorption during daylight. The penetration into the D-region of moderately heavy solar atomic ions is suggested as an explanation of the absorption phenomena.

551.510.535 : 523.78 **124**
Solar Eclipse of 30th June 1954 and its Effect upon the Ionosphere.—S. N. Mitra. (*Indian J. Phys.*, June 1957, Vol. 31, No. 6, pp. 309–323.) The investigations at Delhi reported include vertical-incidence ionospheric measurements, the recording of variations of s.w. and m.f. signals, and the recording of solar noise on 204 Mc/s. Evidence of three distinct 'corpuscular eclipses' was obtained both on the ionization density of the F-layer and on the absorption in the nondeviating region. These were found to occur 2, 4, and 6½ h before the optical eclipse. Theoretical arguments and past observations indicate that these corpuscular emissions may have originated from the so-called M-region of the sun.

551.510.535 : 523.78 : 621.396.11 **125**
Ionospheric Research.—A. P. Dale. (*R.S.G.B. Bull.*, May 1957, Vol. 32, No. 11, pp. 499–501.) Note on signal-strength changes during the annular solar eclipse on 25th December 1954 and the control days for transmitters both in South Africa and in other countries.

551.510.535 : 550.385 **126**
Disturbances in the F_1 and E Regions of the Ionosphere Associated with Geomagnetic Storms.—T. Sato. (*J. Geomag. Geoelect.*, March 1957, Vol. 9, No. 1, pp. 57–60.) The E and F_1 layer variations which occur during geomagnetic disturbances are examined for different phases of the solar cycle and for a number of latitudes. The data are discussed in relation to the electron drifts associated with geomagnetic disturbances. See also 127 below.

551.510.535 : 550.385 **127**
Disturbances in the Ionospheric F_2 Region associated with Geomagnetic Storms: Part 2—Middle Latitudes.—T. Sato. (*J. Geomag. Geoelect.*, March 1957, Vol. 9, No. 1, pp. 1–22.) Disturbances at middle latitudes are divided into negative and positive types, characteristic of high and low latitudes respectively. These results are discussed in terms of vertical drift of electrons and the disturbance daily variations in the geomagnetic field. The bearing of the *Sq* dynamo current on seasonal and latitude variations of disturbances is discussed. Part 1: 3139 of 1957.

551.510.535 : 621.396.812.3 **128**
Ionospheric Irregularities Causing Random Fading of Very Low Frequencies.—Bowhill. (See 251.)

551.510.535 : 621.396.9 :523.3 **129**
Ionospheric Studies by the Lunar Technique.—B. C. Blevis. (*Nature, Lond.*, 20th July 1957, Vol. 180, No. 4577, pp. 138–139.) Equipment operating near Ottawa comprises a 10-kW 488-Mc/s transmitter, keyed 'on' for 2 sec and 'off' for 4 sec, feeding a 28-ft paraboloid via a horn, and orthogonal dipoles with a similar reflector feeding a dual-conversion receiver (i.f. 32 and 4 Mc/s). Synchronous detection is used to give an a.f. output signal equal in frequency to the Doppler shift of the signal reflected from the moon. Preliminary results indicate the presence of fading due to libration, long- and short-period Faraday rotation and long-term fading without Faraday rotation.

551.594.21/.25 **130**
The Electrification of Precipitation and Thunderstorms.—R. Gunn. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1331–1358.) A review article. Cosmic rays, radioactivity and other agencies produce ion pairs which are transferred to cloud droplets or ice crystals and establish a droplet charge distribution greatly influenced by differences in conductivity for positive and negative ions. Rain formed from these droplets may be electrified, usually both positive and negative drops being present. Processes are considered which charge this rain and may eventually lead to a lightning stroke. The conditions necessary to establish gross free charge distributions by regeneration are specified and shown to be met under frequently occurring meteorological situations. 47 references.

551.594.5 **131**
Rotational Temperatures Measured in Aurorae at Churchill, Manitoba.—R. Montalbetti. (*Canad. J. Phys.*, Aug. 1957, Vol. 35, No. 8, pp. 831–836.) The measurements indicate there is no latitude effect. A vertical temperature gradient of 6°K/km is deduced. See also 3871 of 1957 (Montalbetti & Jones).

551.594.6 : 523.745 **132**
A Study of the Audio-Frequency Radio Phenomenon Known as 'Dawn Chorus'.—G. McK. Allcock. (*Aust. J. Phys.*, June 1957, Vol. 10, No. 2, pp. 286–298.) Experimental evidence supports the hypothesis that 'dawn chorus' signals are generated by the entry into the ionosphere of clouds of positively charged particles of solar origin. Propagation appears to take place in the extraordinary magneto-ionic mode along the earth's magnetic lines of force.

LOCATION AND AIDS TO NAVIGATION

621.396.932 **133**
Electronic Clock-Coder for Marine Radio Beacons.—A. C. MacKellar & A. J. B. Baty. (*Brit. Commun. Electronics*, Aug. 1957, Vol. 4, No. 8, pp. 476–477.) Crystal control is used to give more reliable service.

621.396.932.2 **134**
The TALBE: a V.H.F. C.W. Radio Aid for Air/Sea Rescue.—W. Kiryluk. (*J. Brit. Instn Radio Engrs*, Sept. 1957, Vol. 17, No. 9, pp. 489–500.) Describes the development and construction of a 'talk-and-listen beacon', together with prior work on an unmodulated dinghy transmitter.

621.396.932.2 **135**
An Introduction to Radio Aids to Air/Sea Rescue.—G. W. Hosie. (*J. Brit. Instn Radio Engrs*, Sept, 1957, Vol. 17, No. 9, pp. 481–488.) Developments since 1940 are surveyed, and the m.f. c.w. and v.h.f. radar beacons in general use during the war are discussed, together with post-war equipment. A crystal-controlled v.h.f. beacon of low duty cycle is suggested, and also a crash location beacon.

621.396.962.3 : 621.374.5 **136**
Pulse Compression.—Krönert. (See 72.)

621.396.965.8 **137**
Optimizing the Dynamic Parameters of a Track-While-Scan System.—J. Sklansky. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 163–185.)

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 546.48.241.1 **138**
Preparation and Photoconductive Properties of Cadmium Telluride Films.—G. G. Kretschmar & L. E. Schilberg. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, pp. 865–867.) A thin film gave a high dark resistance, falling by 10 to 200 times with 2 ft-candle illumination. A film prepared in indium vapour gave a dark resistance of 1–10 MΩ falling to a few hundred kΩ.

535.37 **139**
Critical Comment on a Method for Determining Electron Trap Depths.—C. H. Haake. (*J. opt. Soc. Amer.*, July 1957, Vol. 47, No. 7, pp. 649–652.) "A critical analysis of Garlick and Gibson's method [3421 of 1948] based on a mathematical determination of the applicable temperature range and on experimental evidence shows that the trap depths obtained are lower than those found by other methods. Especially the frequency-of-escape constants calculated subsequently contrast strongly to those determined by separate measurements."

535.37 **140**

Energy Transport by Cascade and Resonance Processes in Doubly Activated Phosphors.—E. W. Claffy & C. C. Klick. (*J. electrochem. Soc.*, July 1957, Vol. 104, No. 7, pp. 445–447.)

535.376 **141**

Review of Articles on Luminescence for 1955–1956.—G. R. Fonda. (*J. electrochem. Soc.*, Aug. 1957, Vol. 104, No. 8, pp. 524–530.) Review covering articles published in English, German and French. 156 references.

535.376 : 546.472.21 **142**

Electroluminescence of ZnS Single Crystals.—G. F. Neumark. (*Sylvania Technologist*, April 1957, Vol. 10, No. 2, pp. 29–34.) Electroluminescence in ZnS depends on the number and type of traps in the crystal. Impact ionization appears to be the predominant mechanism for the excitation. 31 references.

537.226/.228.1 : 546.431.824-31 : 534.232 **143**

An Investigation of some Barium Titanate Compositions for Transducer Applications.—D. Schofield & R. F. Brown. (*Canad. J. Phys.*, May 1957, Vol. 35, No. 5, pp. 594–607.) The addition of Co to 5% Ca, 95% Ba titanate compositions produces a large reduction in the dielectric loss in high electric fields without significant effect on the piezoelectric characteristics. Preliminary results were given earlier (3160 of 1957).

537.226/.227 : 534.1.087 **144**

The Use of Ferroelectric Ceramics for Vibration Analysis.—R. C. Kell, G. A. Luck & L. A. Thomas. (*J. sci. Instrum.*, July 1957, Vol. 34, No. 7, pp. 271–274.)

537.227 : 546.431.824-31 **145**

Preparation of $BaTiO_3$ Single Crystals in Coal Gas Atmosphere.—K. Kawabe & S. Sawada. (*J. phys. Soc. Japan*, Feb. 1957, Vol. 12, No. 2, p. 218.)

537.227 : 546.431.824-31 **146**

Heating Effects in Single Crystal Barium Titanate.—D. S. Campbell. (*J. Electronics Control*, Sept. 1957, Vol. 3, No. 3, pp. 330–338.) The coercive field increases with frequency up to a certain limiting frequency above which it falls due to heating effects. This gives a criterion for the maximum frequency at which crystals can be expected to operate in a matrix storage system.

537.227 : 546.431.824-31 **147**

Quasi-static Hysteresis in Barium-Titanate Single Crystals.—K. Zen'iti, K. Husimi & K. Kataoka. (*J. phys. Soc. Japan*, April 1957, Vol. 12, No. 4, p. 432.)

537.227 : 546.431.824-31 **148**

On the 180°-Type Domain Wall of $BaTiO_3$ Crystal.—W. Kinase & H. Takahasi. (*J. phys. Soc. Japan*, May 1957, Vol. 12, No. 5, pp. 464–476.) At room temperature the domain width is about 10^{-4} cm and its energy is $1 \cdot 4$ erg/cm^2. The structure of the domains and their reversing process are also discussed.

537.228.5 : 546.36 **149**

Stark Effect on Caesium-133 Hyperfine Structure.—R. D. Haun, Jr, & J. R. Zacharias. (*Phys. Rev.*, 1st July 1957, Vol. 107, No. 1, pp. 107–109.) The change of the hyperfine-structure separation energy caused by an electric field has been measured in ^{133}Cs by the atomic-beam magnetic-resonance method and is compared with other experimental and theoretical data.

537.311.33 **150**

Generation of an E.M.F. in Semiconductors with Nonequilibrium Current Carrier Concentrations.—J. Tauc. (*Rev. mod. Phys.*, July 1957, Vol. 29, No. 3, pp. 308–324.) A general formulation of the problem is presented. Solutions are obtained for two special cases for which the impurity-concentration change occurs over a distance either long or short compared with the diffusion length of current carriers; these are bulk and barrier-layer photovoltaic effects respectively. Practical implications are discussed. 102 references.

537.311.33 **151**

Statistics of the Charge Distribution for a Localized Flaw in a Semiconductor.—W. Shockley & J. T. Last. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 392–396.) The charge situation is described by a set of energy levels which are independent of the Fermi level but are temperature-dependent. The charge of the flaw increases by one electron unit of negative charge as the Fermi level is raised successively above each level of the set and, conversely, as it is lowered.

537.311.33 **152**

Scattering of Current Carriers in Semiconductors with Ionic Type of Bond.—T. A. Kontorova. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 269–274.) The scattering of carriers in the optical and acoustic modes is investigated and their free path is calculated. Approximate formulae for the carrier mobility are derived. The dependence of mobility on temperature is examined.

537.311.33 **153**

High-Frequency Relaxation Processes in the Field-Effect Experiment.—C. G. B. Garrett. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 478–487.) A theoretical treatment is given of two dispersion phenomena in the semiconductor field-effect experiment [see e.g. 3532 of 1951 (Montgomery)]: (*a*) dispersion arising from the finite time required to generate minority carriers, and (*b*) relaxation of the fast surface states.

537.311.33 : 538.632 **154**

On the Electrical Properties of Degenerate Semiconductors.—E. Haga. (*J. phys. Soc. Japan*, Feb. 1957, Vol. 12, No. 2, p. 217.) Theoretical derivation of quantitative data on resistivity and the Hall coefficient.

537.311.33 : 538.632 **155**

Resistivity and Hall Coefficient of Semiconductors.—K. Shogenji & S. Uchiyama. (*J. phys. Soc. Japan*, April 1957, Vol. 12, No. 4, p. 434.) Calculations, assuming scattering of current carriers by acoustic and optical modes of lattice vibration, but neglecting scattering by impurity centres.

537.311.33 : 546.23 : 621.314.6 **156**

The Surface Potential of Selenium and its Relation to the Rectification and the Photovoltaic Effects.—U. Yoshida & A. Suzuki. (*J. phys. Soc. Japan*, May 1957, Vol. 12, No. 5, pp. 459–463.) The changes with time of the surface potential and the rectification ratio after heat treatment are reduced by coating with a thin film of synthetic resin.

537.311.33 : 546.24 **157**

Electrical Properties of Tellurium at the Melting Point and in the Liquid State.—A. S. Epstein, H. Fritzsche & K. Lark-Horovitz. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 412–419.) The experimental results show that semiconductor properties persist in the liquid state, but as the temperature is raised a gradual transition to metallic behaviour occurs.

537.311.33 : 546.27 **158**

Electrical Properties of Boron Single Crystals.—W. C. Shaw, D. E. Hudson & G. C. Danielson. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 419–427.) The electrical resistivity, Hall coefficient and thermoelectric power of microscopic crystals have been measured as functions of temperature. A value of $1 \cdot 55 \pm 0 \cdot 05$ eV is deduced for the energy gap.

537.311.33 : [546.28 + 546.289 **159**

Expansions in Reactor Irradiated Germanium and Silicon.—M. C. Wittels. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, p. 921.) The lattice expansions observed were greater in Ge than Si, as expected from theory. The retained damage in Ge was dependent on the irradiation temperature.

537.311.33 : 546.28 **160**

Donor Electron Spin Relaxation in Silicon.—E. Abrahams. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 491–496.) The effects of spin-orbit coupling on the relaxation time for donor electron spins are considered.

537.311.33 : 546.28 **161**

Damage to Silicon Produced by Bombardment with Helium Ions.—U. F. Gianola. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, pp. 868-873.) Helium ions at 3×10^4 eV were used. The surface layer was found to be converted to a quasi-stable amorphous form, which becomes partially crystalline on annealing. The amorphous material is soluble in aqueous hydrofluoric acid and has about 1/10th the thickness of the affected Si.

537.311.33 : 546.28 **162**

Oxygen Content of Silicon Single Crystals.—W. Kaiser & P. H. Keck. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, pp. 882–887.) The oxygen content of pulled Si crystals is correlated with the infrared absorption at 9 μ. Methods of altering the oxygen concentration are discussed.

537.311.33 : 546.289 163

Study of the Minority-Carrier Mobility in Germanium.—M. Shtenbek & P. I. Baranski. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 221–232.) Report of investigation on *n*-type single-crystal Ge. The transit time, the mobility and the coefficients of longitudinal and transverse diffusion of minority carriers are calculated.

537.311.33 : 546.289 164

Experimental Investigation of the Correlation between Peltier Effect and Thermo-e.m.f. in Germanium.—M. Shtenbek & P. I. Baranski. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 233–237.) This correlation is investigated in the impurity and intrinsic conduction bands for samples of *n*- and *p*-type Ge of differing specific resistance.

537.311.33 : 546.289 165

Nature of Recombination Centres in Germanium Originated by Low-Temperature Heat Treatment.—T. V. Mashovets & S. M. Ryvkin. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 238–241.) See also 2794 of 1956.

537.311.33 : 546.289 166

Probability of Capture by Carrier Recombination at Frenkel-Type Defects in *n*-Type Germanium.—L. S. Smirnov & V. S. Vasilov. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 427–429.) Bombardment of Ge crystals with electrons of energy over 500 keV produces, at room temperature, stable dislocations of crystal structure, called Frenkel defects, corresponding to several levels in the forbidden zone, which act as effective recombination centres of holes and electrons.

537.311.33 : 546.289 167

Electrical Properties of Nickel-Doped Germanium at Low Temperatures.—Y. Kanai & R. Nii. (*J. phys. Soc. Japan*, Feb. 1957, Vol. 12, No. 2, pp. 125–133.) Anomalous behaviour was found at temperatures in the range 120–170°K, the resistivity increasing as the temperature was reduced. The electrical properties could be explained by use of the model proposed by Hung (*Phys. Rev.*, 15th Aug. 1950, Vol. 79, No. 4, pp. 727–728.)

537.311.33 : 546.289 168

Peripheral Inhomogeneities of the Alloyed Germanium *p-n* Junction.—M. Kikuchi. (*J. phys. Soc. Japan*, Feb. 1957, Vol. 12, No. 2, pp. 133–139.) The local photocurrent exhibits peaks, the positions of which depend on the bias voltage. A model for the junction is proposed.

537.311.33 : 546.289 169

Observation of the Drift of Germanium Surface by Field-Effect Experiment.—M. Kikuchi. (*J. phys. Soc. Japan*, April 1957, Vol. 12, No. 4, p. 436.) The surface conductance and photoconductance were measured at various time intervals after the etching of *n*-type Ge.

537.311.33 : 546.289 170

Alloying Properties of Germanium Free of Edge Dislocations.—C. W. Mueller. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 205–212.) The main factors involved in alloying In on dislocation-free Ge are considered. The controlled alloying method described gives extremely flat and uniform junctions.

537.311.33 : 546.289 171

The Dissolution of Germanium by Molten Indium.—B. Goldstein. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 213–220.) Experimental results of an investigation of dissolution as a function of crystal axis orientation, temperature and amount of Ge already contained in the molten In solvent.

537.311.33 : 546.289 172

Preferential Diffusion of Sb along Small-Angle Boundaries in Ge and the Dependence of this Effect on the Direction of the Dislocation Lines in the Boundary.—F. Karstensen. (*J. Electronics Control*, Sept. 1957, Vol. 3, No. 3, pp. 305–307.) Experiment shows that preferential diffusion occurs if the grain boundary contains dislocation lines parallel to the direction of diffusion, but does not occur if these are perpendicular to it.

537.311.33 : 546.289 173

Diffusion of Antimony out of Germanium and Some Properties of the Antimony-Germanium System.—R. C. Miller & F. M. Smits. (*Phys. Rev.*, 1st July 1957, Vol. 107, No. 1, pp. 65–70.) Experimental Sb distributions are shown to be consistent with a rate limitation or potential barrier at the Ge/ambient interface. Introduced external rate limitations provide data for calculating partition and sticking coefficients for the system and estimating the binding energy of.Sb in Ge. See also 1478 of 1957.

537.311.33 : 546.289 174

Distribution and Cross-Sections of Fast States on Germanium Surfaces in Different Gaseous Ambients.—A. Many & D. Gerlich. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 404–411.) Simultaneous measurements of surface recombination velocity and trapped charge density in the fast states as a function of surface potential, show that the energy distribution of the fast states consists of four discrete sets of levels. Only one of these sets is significant in the recombination process.

537.311.33 : 546.289 175

The Depth of Surface Damage Produced by Lapping Germanium Monocrystals.—D. Baker & H. Yemm. (*Brit. J. appl. Phys.*, July 1957, Vol. 8, No. 7, pp. 302–303.) The depth of damage is estimated from the variation of the effective diffusion distance of minority carriers as the disturbed layer is removed by etching.

537.311.33 : 546.289 : 534.13-8 176

Ultrasonic Attenuation by Free Carriers in Germanium.—G. Weinreich. (*Phys. Rev.*, 1st July 1957, Vol. 107, No. 1, pp. 317–318.) The close relationship with the acoustoelectric effect is noted and the amount of attenuation is predicted from the observed magnitude of this effect. See also 2499 of 1957 (Blatt).

537.311.33 : 546.289 : 537.32 177

Magnetic Field Dependence of the Seebeck Effect in Germanium.—M. C. Steele. (*Phys. Rev.*, 1st July 1957, Vol. 107, No. 1, pp. 81–83.) The thermoelectric power of *n*-type Ge single crystals increases with magnetic field at temperatures from 78° to 278°K. The thermal conductivity is constant for magnetic fields between zero and 11 500 oersteds.

537.311.33 : 546.289 : 538.6 178

Investigation of Thermomagnetic Effects in *p*-Type Germanium.—I. V. Mochan, Yu. N. Obraztsov & T. V. Krylova. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 242–259.) The transverse and longitudinal Nernst-Ettingshausen and Hall effects are measured on samples of *p*-type Ge of specific resistance 69 Ω.cm at room temperature, in the impurity conduction band. In the temperature range 80°–240°K and a field of 5 000 G the tranverse Nernst–Ettingshausen effect appears to be due to hole transfer by phonons.

537.311.33 : 546.289 : 538.63 179

On the Measurements of the Galvanomagnetic Effect in Semiconductors at Microwave Frequencies.—T. Fukuroi & M. Date. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A.*, June 1957, Vol. 9, No. 3, pp. 190–195.) The resonant cavity method is used to make measurements at 9 800 Mc/s at temperatures down to 1·5°K. The results obtained for the magnetoresistance of pure Ge are compared with those of Fritzche (153 of 1956).

537.311.33 : 546.289 : 538.632 180

Galvanomagnetic Theory for *n*-Type Germanium and Silicon: Hall Theory and General Behaviour of Magnetoresistance.—L. Gold & L. M. Roth. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 358–364.) The magnetoresistance $\Delta\rho/\rho$ and Hall coefficient R_H for *n*-type Ge and Si are calculated from the resistivity tensor. The angular dependence of $\Delta\rho/\rho$ is compared with experimental data. The field dependence of $\Delta\rho/\rho$ and of R_H are considered for various current and field orientations.

537.311.33 : 546.289 : 539.1 181

Diffusion-Limited Annealing of Radiation Damage in Germanium.—T. R. Waite. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 471–478.) The theory given in 104 above has been applied to experimental data [see e.g. 1469 of 1954 (Brown et al.)]; reasonable agreement was found.

537.311.33 : 546.682.19 : 535.215 182

Photoelectromagnetic Effect in Indium Arsenide.—J. R. Dixon. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 374–378.) Experimental studies on *n*- and *p*-type single crystals at room temperature are described. The dependence of the photoelectromagnetic short-circuit current on magnetic induction is consistent with the theory of Kurnick & Zitter (2429 of 1956). Bulk lifetimes are found to be about 6×10^{-8} for *n*-type, and 5×10^{-10} for *p*-type material; estimates of surface recombination velocity for *n*-type material are 0–10^3 cm/s for an etched surface, and 10^5 cm/s for a ground surface.

537.311.33 : 621.3.049.7 **183**

Technique for Connecting Electrical Leads to Semiconductors.—O. L. Anderson, H. Christensen & P. Andreatch. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, p. 923.) Two techniques are described which give adhesion between certain soft metals and semiconductors at temperatures well below the eutectic point.

537.311.33 : 621.314.63 **184**

Minority Carrier Lifetime in *p-n* Junction Devices.—M. Byczkowski & J. R. Madigan. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, pp. 878–881.) Minority-carrier lifetime is conveniently determined by studying the switching time in junction diodes. Some theoretical and experimental results are given.

537.533 : 546.883 **185**

Electrostatic Emission from Tantalum Single Crystals.—N. A. Gorbatyĭ, L. V. Reshetnikova, E. P. Sytaya & G. N. Shuppe. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 296–298.) Field-emission patterns for Ta single crystals appear identical to those obtained for W and Mo.

538.22 **186**

Effective Gyromagnetic Ratio for Triangular Ferrimagnetic States.—A. Eskowitz & R. K. Wangsness. (*Phys. Rev.*, 15th July 1957, Vol. 107, No. 2, pp. 379–380.)

538.22 **187**

Magnetic Properties of Perovskites Containing Strontium: Part 1—Strontium-Rich Ferrites and Cobaltites.—H. Watanabe. (*J. phys. Soc. Japan*, May 1957, Vol. 12, No. 5, pp. 515–522.) Magnetic susceptibilities and electrical conductivities for samples sintered in vacuo differ from those sintered in oxygen. An interpretation of experimental results is given.

538.22 : 538.569.4 : 546.763-31 **188**

Antiferromagnetic Resonance in Cr_2O_3.—E. S. Dayhoff. (*Phys. Rev.*, 1st July 1957, Vol. 107, No. 1, pp. 84–91.) Theory and experimental data are given for the antiferromagnetic resonance spectrum of single-crystal Cr_2O_3 for 12–2·8 mm λ and magnetic field intensities from zero to 30 000 oersteds.

538.22 : 538.613 **189**

Magnetography—the Microscopy of Magnetism.—F. G. Foster. (*Bell Lab. Rec.*, May 1957, Vol. 35, No. 5, pp. 175–178.) The Kerr magneto-optical effect is applied to the microscopy of magnetic domain structure.

538.221 **190**

Transverse Impedance Transformation for Ferromagnetic Media.—F. R. Morgenthaler. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, p. 1407.) The transformation relates to media magnetized transversely, and is valid for TE_{m0} modes with the magnetization in the E-field direction.

538.221 **191**

Domain Wall Orientations in Silicon-Iron Crystals.—C. D. Graham, Jr, & P. W. Neurath. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, pp. 888–891.) The stable orientation of a 180° ferromagnetic domain wall in a cubic crystal is calculated and the results are compared with experimentally determined orientations.

538.221 : 538.122 **192**

Experimental Investigation of the Magnetic Field Distribution due to Artificial Surface Defects in Ferromagnetic Substances.—N. N. Zatsepin. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 368–373.)

538.221 : 539.23 **193**

Magnetization Reversal in Thin Films at Low Fields.—R. L. Conger & F. C. Essig. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, pp. 855–858.) Statistical theory is presented for reversing fields less than the anisotropy field under the assumption that reversal takes place by domain wall motion; experimental data supporting the theory are given. See also 1513 of 1957.

538.221 : [621.318.124 + 621.318.134 **194**

Dielectric Spectroscopy of Ferromagnetic Semiconductors.—P. A. Miles, W. B. Westphal & A. von Hippel. (*Rev. mod. Phys.*, July 1957, Vol. 29, No. 3, pp. 279–307.) Methods of measuring the electric and magnetic spectra, within the range from d.c. to optical frequencies, are described and results discussed. In addition to a general review some new experimental data including the r.f. spectra of a single-crystal Ni ferrite, are presented.

538.221 :[621.318.124+621.318.134 **195**

On the Origin of the Magnetic Anisotropy Energy of Ferrites.—K. Yosida & M. Tachiki. (*Progr. theor. Phys.*, March 1957, Vol. 17, No. 3, pp. 331–359.) The anisotropy energy for Ni ferrites comes mainly from Fe^{3+} ions, for magnetite from Fe^{2+} and Fe^{3+} ions and for Mn ferrites from Fe^{3+} and Mn^{2+} ions. The large anisotropy energy of Co ferrites arises from the pseudo-quadrupole and anisotropic exchange reactors among Co and Fe ions.

538.221 : [621.318.124+621.318.134 **196**

A Theory of the Uniaxial Anisotropy Induced by Magnetic Annealing in Ferrites.—S. Taniguchi. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, June 1957, Vol. 9, No. 3, pp. 196–214.)

538.221 : 621.318.134 **197**

Magnetic and Magnetostrictive Properties of Magnesium-Nickel Ferrites.—P. O. Hoffmann. (*J. Amer. ceram. Soc.*, 1st July 1957, Vol. 40, No. 7, pp. 250–252.) A study of mixed Ni-Mg ferrites of various compositions shows them to be inferior to Ni ferrite.

538.221 : 621.318.134 **198**

Magnetic Properties and Associated Microstructure of Zinc-Bearing Square-Loop Ferrites.—G. G. Palmer, R. W. Johnston & R. E. Schultz. (*J. Amer. ceram. Soc.*, 1st Aug. 1957, Vol. 40, No. 8, pp. 256–262.) A series of Mn-Mg ferrites in which Zn is substituted for Mg has been prepared and tested in a successful attempt to obtain a material having a lower coercive force and higher magnetic saturation than standard rectangular-loop ferrites used extensively in digital computers.

538.221 : 621.318.134 : 538.6 **199**

Rotation of the Plane of Polarization in a Longitudinal Magnetic Field (Faraday Effect) in the Millimetre Wavelength Range.—D. I. Mash. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 360–363.) Results of tests on ferrites are summarized and discussed.

538.632 **200**

Temperature Dependence of the Hall Coefficients in some Silver Palladium Alloys.—F. E. Allison & E. M. Pugh. (*Phys. Rev.*, 1st July 1957, Vol. 107, No. 1, pp. 103–105.)

666.1.037.5 **201**

Fundamentals of Glass-to-Metal Bonding: Part 3—Temperature and Pressure Dependence of Wettability of Metals by Glass.—R. M. Fulrath, S. P. Mitoff & J. A. Pask. (*J. Amer. ceram. Soc.*, 1st Aug. 1957, Vol. 40, No. 8, pp. 269–274.) Part 2 : 3576 of 1957 (Mitoff).

MATHEMATICS

512.25 **202**

A Least-Squares Solution of Linear Equations with Coefficients Subject to a Special Type of Error.—J. K. Mackenzie. (*Aust. J. Phys.*, March 1957, Vol. 10, No. 1, pp. 103–109.)

MEASUREMENTS AND TEST GEAR

529.786+531.711 **203**

The Units of Time and Length.—L. Essen. (*Nature, Lond.*, 20th July 1957, Vol. 180, No. 4577, pp. 137–138.) Difficulties in an ideal system using an atomic frequency standard to define the unit of time are discussed. A practical solution would be to define time in terms of the caesium line, and length in terms of the radiation of a suitable light source. A system in which time interval is based on an atomic unit and epoch on astronomical measurements has been satisfactory during two years of operation at the National Physical Laboratory.

529.786 **204**

The Caesium Resonator as a Standard of Frequency and Time.—L. Essen & J. V. L. Parry. (*Phil. Trans. A.*, 8th Aug. 1957, Vol. 250, No. 973, pp. 45–69.) The construction, operation and testing of the standard are described. The present experimental model can be used to define frequency with a standard deviation of $\pm$ 1 part in 10^{10}. The design of a resonator with sharper resonance less than 100 c/s wide appears possible and should improve the definition. See also 2471 of 1956.

529.786 : 538.569.4 **205**

Recent Developments in Measurement of Time.—C. H. Townes. (*Nuovo Cim.*, 1957, Vol. 5, Supplement No. 1, pp. 222–229. In English.) The use of the molecular oscillator described in 100 of 1955 (Gordon et al.) as a frequency standard is discussed.

621.3.018.41(083.74) + 529.786] : 538.569.4 **206**

Precise Frequency of the 3,3 Inversion Line of Ammonia.—K. Shimoda. (*J. phys. Soc. Japan*, May 1957, Vol. 12, No. 5, p. 558.)

621.3.018.41(083.74) : 621.385.029.6 **207**

A New Microwave Frequency Standard by Quenching Oscillator Control.—N. Sawazaki & T. Honma. (*Trans. Inst. Radio Engrs*, April 1956, Vol. MTT-4, No. 2, pp. 116–121. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 44, No. 7, pp. 956–957.)

621.317.1.087 : 519.24 **208**

The Automatic Classification and Storage of Measurement Results.—H. J. Vogt & E. Zimmer. (*Elektronik*, July 1957, Vol. 6, No. 7, pp. 191–197.) Description of methods and electronic equipment for use in the statistical analysis of manufacturing processes, with particular reference to applications in the textile industry.

621.317.3 : 537.311.33 **209**

Direct-Reading Minority-Carrier-Lifetime Measuring Apparatus.—A. R. Engler & C. J. Kevane. (*Rev. sci. Instrum.*, July 1957, Vol. 28, No. 7, pp. 548–551.) The decay of optically-injected excess minority carriers is compared with voltage decay in a *RC* circuit, and lifetimes ($>10\ \mu s$) are read directly from a calibrated dial.

621.317.3 : 538.63 **210**

Geometrical Effects in Transverse Magnetoresistance Measurements.—J. R. Drabble & R. Wolfe. (*J. Electronics Control*, Sept. 1957, Vol. 3, No. 3, pp. 259–266.) "The effect of finite length-to-width ratio and of probe positions on measurements of resistivity in a transverse magnetic field is investigated for various Hall angles. Wick's method (3161 of 1954) is extended to include the case of measurements made between probes symmetrically placed at various distances from the ends. Graphs are included which give the fractional change of resistance for various probe positions and Hall angles for specimens of any length-to-width ratio greater than two."

621.317.3 : 538.632 **211**

Experimental Use of Hall-Effect Linear Detector for Measurements.—V. N. Bogomolov & V. D. Vasil'ev. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 260–261.) The circuit described incorporates an *n*-type Ge plate as the Hall detector; an experimental application is briefly described.

621.317.3 : 621.314.63 **212**

Measurements of Crystal Impedances at Low Levels.—H. N. Dawirs & E. K. Damon. (*Trans. Inst. Radio Engrs*, April 1956, Vol. MTT-4, No. 2, pp. 94–96. Abstract, *Proc. Inst. Radio Engrs*, July 1956, Vol. 44, No. 7, p. 956.)

621.317.3 : 621.314.7 **213**

Transistor Tests Predict High-Frequency Performances.—R. L. Pritchard. (*Electronic Ind. Tele-Tech*, March 1957, Vol. 16, No. 3, pp. 62–63 . . 144.) Methods are described for measuring the alpha cut-off frequency, the ohmic base resistance and the collector-base capacitance, from which parameters the high-frequency performance may be determined.

621.317.3 : 621.396.822 : 621.396.62.029.6 **214**

Absolute Measurement of Receiver Noise Figures at U.H.F.—E. Maxwell & B. J. Leon. (*Trans. Inst. Radio Engrs*, April 1956, Vol. MTT-4, No. 2, pp. 81–85. Abstract, *Proc. Inst. Radio Engrs*, July 1956, Vol. 44, No. 7, p. 956.)

621.317.331 : 621.316.993 **215**

Measuring Earth Conductivity.—M. Strohfeldt. (*Electronic Radio Engr*, Nov. 1957, Vol. 34, No. 11, pp. 425–427.) Measurements at an 'ideal' site by radio methods gave a much lower value than those obtained by electrode methods. Some correlation exists between measurements by the two methods when applied to the variation of conductivity with distance.

621.317.335.3.029.64 **216**

A Rod Method of Measuring Dielectric Constants. Application to $(NH_4)_2HI_3O_6$ at 3-cm Wavelength.—H. Gränicher & W. Schurter. (*Z. angew. Math. Phys.*, 25th Sept. 1957, Vol. 8, No. 5, pp. 382–400.) A modified version of the method described by Le Bot & Le Montagner (2067 of 1953) was adopted to measure the complex permittivity of rod specimens. The experimental results are discussed.

621.317.34.029.63/.64 **217**

Automatic Microwave Transmission Measuring Equipment.—J. B. Linker, Jr, & H. H. Grimm. (*Rev. sci. Instrum.*, July 1957, Vol. 28, No. 7, pp. 559–563.) The phase of a 200-kc/s heterodyne is compared with that of a phase-coherent 200-kc/s reference source by a null-seeking servo phase-discriminator, and loss is measured by a commercial ratio meter. A travelling-wave valve produces the phase-coherent 200-kc/s frequency offset which is passed through a waveguide containing the ferrite or other material to be measured.

621.317.343.019.6 **218**

V.H.F. Line Measurements.—F. J. Charman. (*Electronic Engng*, Oct. 1957, Vol. 29, No. 356, pp. 499–503.) A method is described of using a single loop directional coupler by which the impedance or admittance of an unknown load may be obtained graphically. The method is compared with that of the impedometer [155 of 1950 (Parzen)]; the parameters measured by the new device are not directly applicable to a Smith chart.

621.317.351 : 621.397 **219**

Equipment for the Oscillographic Display of the Phase and Amplitude Characteristics of Video and I.F. Networks.—W. Kroebel & L. A. Wegner. (*Rundfunktech. Mitt.*, April 1957, Vol. 1, No. 2, pp. 37–44.) In the equipment described parameters representing phase shift and amplitude distortion are displayed on two c.r. tubes as a function of modulation frequency in the range 0·1–6·0 Mc/s, together with an electronically produced graticule. Special methods of phase measurement are described and some test results are reproduced.

621.317.353.1 **220**

A New Method of Measuring Non-linear Distortion.—O. Henkler. (*Nachr-Tech.*, April 1957, Vol. 7, No. 4, pp. 145–147.) The 'combination-factor' method outlined is more suitable, particularly for application to carrier-frequency systems, than the conventional distortion-factor and difference-tone methods. Two test signals of equal amplitude are used, with a frequency difference depending on the application. The 'combination factor' obtained is the ratio of the amplitude of a certain intermodulation product to that of one of the test signals.

631.317.382 **221**

A Wide-Band Level-Measuring Set.—R. C. Bolt. (*A.T.E. J.*, April 1957, Vol. 13, No. 2, pp. 151–156.) The instrument is suitable for power measurements at various input impedances, in the frequency range 200 c/s–612 kc/s at levels from −50 to + 40 dB relative to 1 mW.

621.317.444 : 621.385 **222**

Rectilinearity of Electron-Beam Focusing Fields from Transverse Component Determinations.—P. P. Cioffi. (*Commun. & Electronics*, March 1957, No. 29, pp. 15–19.) An accurate method of measuring the transverse component of the magnetic focusing field using a small search coil is described. The longitudinal field component produced by errors in positioning of the coil is eliminated. For a description of the recording fluxmeter used, see 3174 of 1952.

621.317.733.025 **223**

Component Testing Bridge.—C. D. Lindsay. (*Wireless World*, Nov. 1957, Vol. 63, No. 11, pp. 549–550.) Design of an a.c. comparator working at 50 c/s.

621.317.755 **224**

Calibrated D.C. Oscilloscope.—B. Pearce. (*Wireless World*, Nov. 1957, Vol. 63, No. 11, pp. 539–542.) Design details are given.

621.317.755 : 621.397.5 **225**

Oscillographs for Television Work.—O. H. Davie. (*J. Telev. Soc.*, April/June 1957, Vol. 8, No. 6, pp. 225–233.) A description of the design and application of ancillary equipment required when using a general-purpose oscilloscope for television measurements.

621.317.79.029.64 **226**

A Focused Spectrometer for Microwave Measurements.—E. G. A. Goodall & J. A. C. Jackson. (*Marconi Rev.*, 2nd Quarter 1957, Vol. 20, No. 125, pp. 51–59.) Two variants of spectrometer are described for X-band frequencies. Measurements of transmission and reflection characteristics for incident angles of 15° to 70° on single sandwich-type radome samples are presented.

621.317.79.029.64 : 538.569.4 **227**

A Frequency-Modulated Microwave Spectrometer for Electron-Resonance Measurements.—A. C. Rose-Innes. (*J. sci. Instrum.*, July 1957, Vol. 34, No. 7, pp. 276–278.) The instrument is designed for use in the 3-cm band. Frequency modulation allows spectra to be recorded with simple equipment; details are given of a special technique for increasing the sensitivity. The instrument is especially suitable for measurements at low temperatures and for recording wide lines.

621.317.794 **228**

Comparison of Subtraction-Type and Multiplier-Type Radiometers.—J. Galejs. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1420–1422.)

621.317.794.029.64 **229**

A Calibration Procedure for Microwave Radiometers.—R. N. Whitehurst, F. H. Mitchell & J. Copeland. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1410–1411.) A procedure allowing for the aerial side lobes in estimating radiation from a stellar source. Three measurements are made : one with the fixed aerial pointing to the west of the source, one towards it, and one with the forward lobes absorbed. See also 2739 of 1957.

621.317.799 : 621.385.832 **230**

An Electrostatic Cathode-Ray-Tube Tester.—R. B. Hale. (*Brit. Commun. Electronics*, Aug. 1957, Vol. 4, No. 8, p. 496.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.373.52 : 612 **231**

A Simple Physiological Stimulator, using a Transistor Oscillating Circuit. W. T. Catton, L. Molyneux & B. Schofield. (*Electronic Engng*, Oct. 1957, Vol. 29, No. 356, pp. 496–498.)

621.373.52.029.3 : 53.087.25 : 616 **232**

Monitoring of Low-Frequency Phenomena.—C. F. Rothe & M. W. Street. (*Science*, 12th July 1957, Vol. 126, No. 3263, pp. 77–78.) A transistor regenerative oscillator has been adapted to convert sub-audio frequencies into a.f. for monitoring phenomena such as the electrical activity of the heart in a way more convenient than visual observation of instruments.

621.387 : 654.924.56 **233**

The Ionization Fire-Alarm. Electronics in the Service of Fire Protection. —H. P. Halm. (*Elektronik*, July 1957, Vol. 6, No. 7, pp. 205–207.) The apparatus described uses an ionization chamber in which the presence of combustion gases causes a change in the normal ionization produced by a radioactive source. Some recent refinements are discussed.

621.387.4 : 621.374.3 **234**

Some Electronic Instruments used in Nuclear Spectroscopic Investigations.—B. Åström. (*Ark. Fys.*, 1957, Vol. 12, Part 3, pp. 215–236.) A description is given of two complete sets of equipment : (*a*) an automatic scintillation spectrometer comprising scintillation detector and single-channel pulse-height analyser, (*b*) an α-particle spectrometer comprising ionization chamber and 50-channel pulse-height analyser of stacked discriminator type. To facilitate maintenance and improve cooling, all valves are mounted horizontally and concentrated in a ventilated space.

621.398 : 551.508.822 **235**

Telemetering System is Balloon Borne.—E. K. Novak. (*Electronics*, 1st Sept. 1957, Vol. 30, No. 9, pp. 158–164.) Description of f.m./a.m. telemetry system for a weather radiosonde. It provides a 1 680-Mc/s carrier with three simultaneous channels one of which is time-multiplexed into 12 additional channels with sampling rate of one sample per minute; the maximum range is approximately 360 miles.

PROPAGATION OF WAVES

621.396.11 **236**

A New Solution of the Problem of Propagation over a Flat Earth.—G. Boudouris. (*Nuovo Cim.*, 1957, Vol. 5, Supplement No. 1, pp. 71–91. In French.) Simplifications of classical theory are introduced in fixing the boundary conditions for the problem. This 'method of approximate boundary conditions' is more convenient for practical applications.

621.396.11 **237**

Electromagnetic Fields due to Current Flowing Parallel to Interface of Two Different Media.—K. Horiuchi. (*J. phys. Soc. Japan*, Feb. 1957, Vol. 12, No. 2, pp. 170–176.) Integral equations are established for various configurations. The results have applications in problems of wave propagation over an inhomogeneous earth, or through a stratified atmosphere.

621.396.11 **238**

Fading of Radio Waves Scattered by Dielectric Turbulence.—R. A. Silverman. (*J. appl. Phys.*, Aug. 1957, Vol. 28, No. 8, p. 922.) Note of corrections to 3274 of 1957.

621.396.11 **239**

Short-Range Echoes Observed on Ionospheric Recorders.—R. L. Dowden. (*J. atmos. terr. Phys.*, 1957, Vol. 11, No. 2, pp. 111–117.) Echoes from the 20–100-km region observed at ionospheric stations near the sea are found to arrive at very low angles of elevation with vertical polarization. They are due to coherent back-scatter from sea waves of length $\lambda/2$.

621.396.11 : 551.510.52 **240**

A New Technique for the Study of Scatter Propagation in the Troposphere.—J. H. Chapman, W. J. Heikkila & J. E. Hogarth. (*Canad. J. Phys.*, Aug. 1957, Vol. 35, No. 8, pp. 823–830.) Fluctuations of received field strength may be considered to have the characteristics of noise. Measurements on 500-Mc/s near-optical links show the 'propagation noise' power varies as $1/f^2$ for $0{\cdot}1$ c/s $< f <$ 10 c/s, f being the difference between carrier and sideband frequencies.

621.396.11 : 551.510.535 **241**

Refractive Corrections to Scatter Propagation.—A. D. Wheelon. (*J. geophys. Res.*, Sept. 1957, Vol. 62, No. 3, pp. 343–349.) An examination of the role of the mean electron density of the ionosphere in scatter propagation. Correlations between the field strength of F-layer scattered signals and the maximum usable frequency for classical propagation along the path are studied.

621.396.11 : 551.510.535 **242**

A Method for Obtaining L.F. Oblique-Incidence Reflection Coefficients and its Application to 135·6-kc/s Data in the Alaskan Area.—J. E. Bickel. (*J. geophys. Res.*, Sept. 1957, Vol. 62, No. 3, pp. 373–381.)

621.396.11 : 551.510.535 **243**

The Relation of Forward Scattering of Very High Frequency Radio Waves to Partial Reflection of Medium Frequency Waves at Vertical Incidence.—J. B. Gregory. (*J. geophys. Res.*, Sept. 1957, Vol. 62, No. 3, pp. 383–388.) Some characteristics of v.h.f. forward scatter, such as temporal variations and the region of the ionosphere responsible, are shown to be similar to those of vertical-incidence reflection at 1·75 Mc/s. It is considered that the two phenomena have a common origin.

621.396.11 : 551.510.535 **244**

Studies of Transequatorial Ionospheric Propagation by the Scatter-Sounding Method.—O. G. Villard, Jr, S. Stein & K. C. Yeh. (*J. geophys. Res.*, Sept. 1957, Vol. 62, No. 3, pp. 399–412.) Long-delay echoes on a h.f. radar are interpreted as ground back-scatter propagated by two or more successive reflections from the F region without intermediate ground reflection. Ionospheric tilts could lead to this type of propagation, which occurs regularly in equatorial regions.

621.396.11 : 551.510.535 **245**

Self-Distortion of Radio Waves in the Ionosphere, near the Gyro Frequency.—F. H. Hibberd. (*J. atmos. terr. Phys.*, 1957, Vol. 11, No. 2, pp. 102–110.) The theory described in 2198 of 1956 is extended to include the effect of the geomagnetic field near the gyro frequency. The reduction in modulation depth varies slowly with frequency and shows no resonance-like variation. The reduction is proportional to radiated power and decreases rapidly as the modulation frequency is increased.

621.396.11 : 551.510.535 **246**

Ionospheric Demodulation of Radio Waves at Vertical Incidence.—G. J.

Aitchison. (*Aust. J. Phys.*, March 1957, Vol. 10, No. 1, pp. 204–207.) Agreement between observations and theory indicates that demodulation occurs in the E layer, affecting the wave on both upward and downward paths. The actual mechanism of the demodulation requires further investigation. See also 3070 of 1955 (Aitchison & Goodwin).

621.396.11 : 551.510.535 : 523.78 **247**

Ionospheric Research.—Dale. (See 125.)

621.396.11.029.6 **248**

Radio Propagation above 40 Mc/s over Irregular Terrain.—J. J. Egli. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1383–1391.) The available statistical wave-propagation data on terrain effects as a function of frequency, aerial height, polarization and distance are analysed and expressed by empirical formulae and in the form of nomographs and correction curves.

621.396.11.029.6 : 551.510.535 **249**

Delayed Signals in Ionospheric Forward-Scatter Communication.—G. W. Luscombe. (*Nature, Lond.*, 20th July 1957, Vol. 180, No. 4577, p. 138.) Long-delay multipath effects observed at Slough in the reception of 37-Mc/s pulsed transmissions from Gibraltar, indicate that the signal with the largest delay was due to round-the-world propagation and that ground back-scatter was not involved. See 1217 of 1957 (Crow et al.).

621.396.11.029.62 : 651.510.535 : 523.75 **250**

Disturbances in the Lower Ionosphere Observed at V.H.F. following the Solar Flare of 23 February 1956 with Particular Reference to Auroral-Zone Absorption.—Bailey. (See 123.)

621.396.812.3 : 551.510.535 **251**

Ionospheric Irregularities Causing Random Fading of Very Low Frequencies.—S. A. Bowhill. (*J. atmos. terr. Phys.*, 1957, Vol. 11, No. 2, pp. 91–101.) An investigation, based on measurements at 75 and 150 kc/s, into the two component sources of random fading of signals reflected from the ionosphere: (*a*) scattering in the reflecting layer, (*b*) diffraction during propagation down to the ground.

RECEPTION

621.376 **252**

Intermodulation Products for v-Law Biased Wave Rectifier for Multiple-Frequency Input.—E. Feuerstein. (*Quart. appl. Math.*, July 1957, Vol. 15, No. 2, pp. 183–192.) The intermodulation products obtained by passing the sum of $N+1$ sinusoids of amplitudes $P_1, \ldots P_N$ through a rectifier of characteristic

$$I = \begin{cases} 0 & V < B \\ \alpha\,(V-B)^{\nu} & V > B \qquad \nu > 0 \end{cases}$$

has been expressed in terms of contour integrals involving products of Bessel functions [see e.g. 2169 of 1945 (Rice) and 3506 of 1947 (Bennett)]. These integrals are rewritten as improper integrals on the real line plus constant terms. These integrals converge fast enough in many cases to be useful in numerical integration.

621.376.23 : 621.396.822 **253**

Optimum vs. Correlation Methods in Tracking Random Signals in Background Noise.—R. C. Davis. (*Quart. appl. Math.*, July 1957, Vol. 15, No. 2, pp. 123–138.) A comparison of the use of statistical methods of analysis in estimating unknown parameters of probability distribution with the use of correlation methods previously applied to the problem of obtaining maximum signal/noise ratio.

621.376.332 **254**

An Improved F.M. Discriminator.—V. B. Hulme. (*Electronic Engng*, Sept. 1957, Vol. 29, No. 355, pp. 416–419.) Details are given of a constant-area type circuit which provides a relation linear to 0·1% between input frequency and output voltage over the range 100 kc/s–1 Mc/s.

621.396.62.029.6 : 621.396.822 : 621.317.3 **255**

Absolute Measurement of Receiver Noise Figures at U.H.F.—E. Maxwell & B. J. Leon. (*Trans. Inst. Radio Engrs*, April 1956, Vol. MTT-4, No. 2, pp. 81–85. Abstract, *Proc. Inst. Radio Engrs*, July 1956, Vol. 44, No. 7, p. 956.)

621.396.621 **256**

Receiver Selectivity.—B. J. Rogers. (*R.S.G.B. Bull.*, April 1957, Vol. 32, No. 10, pp. 444–448.) Some modern methods of improving the selectivity of communication receivers are described.

621.396.621 **257**

F.M./A.M. 'Second Set'.—G. D. Browne. (*Mullard tech. Commun.*, March 1957, Vol. 3, No. 22, pp. 38–43.) Design suggestions for a simple and economical receiver in which all valves are operative in both systems of reception.

621.396.621 : 621.314.7 **258**

Design Considerations in the First Stage of Transistor Receivers.—L. A. Freedman. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 145–162.) Noise performance of transistor r.f. stages using capacitive aerials and mixer stages with loop aerials is discussed with examples. Comparisons are made between transistor stages and corresponding valve stages.

621.396.621 : 621.314.7 **259**

Circuit Techniques associated with Transistor Broadcast Receivers.—J. N. Barry. (*Electronic Engng*, Sept. & Oct. 1957, Vol. 29, Nos. 355 & 356, pp. 408–415 & 478–483.) A discussion of transistor circuit problems and their solution. The various stages of the receiver are considered and the economic aspects of using transistors are discussed with reference to the performance of six commercial-type portable and car receivers.

621.396.621 : 621.314.7 **260**

The Thunderbird — a New Transistorized Portable Radio.—T. Vanacore. (*Sylvania Technologist*, April 1957, Vol. 10, No. 2, pp. 35–37.)

621.396.621 : 621.376 **261**

A New Method of Demodulation for Phase, Frequency and Amplitude.—P. Kundu. (*Indian J. Phys.*, April 1957, Vol. 40, No. 4, pp. 231–234.) Limiting and differentiating a modulated sinusoidal wave produces pulses with leading edges separated by an amount dependent upon the instantaneous phase of the modulated wave, and upon its amplitude if the limiting level is above the zero axis. The pulses are converted to variable-amplitude sawtooth waves and passed through a low-pass filter to reproduce the modulating signal. See also 551 of 1956.

621.396.621 : 621.396.822 **262**

Receiver Detects Signals below Noise Level.—W. L. Blair. (*Electronics*, 1st Sept. 1957, Vol. 30, No. 9, pp. 168–171.) The integrating receiver described uses a modified form of Dicke's technique (475 of 1947) with a comparison switching rate of 500 c/s and a counter display.

621.396.621.029.63/.64 **263**

Design for a Broadband Microwave Receiver.—B. Rosen & R. Saul. (*Electronic Ind. Tele-Tech*, March 1957, Vol. 16, No. 3, pp. 81–82 . . 170.) Brief description of a receiver covering the frequency range 950–11 260 Mc/s in four bands. It is suitable for a.m., f.m. and p.m. reception and for use as a microwave power meter.

621.396.8 : 621.3.018.41(083.74) **264**

Doppler Shift of the Received Frequency from the Standard Station Reflected by the Ionosphere.—I. Takahashi, T. Ogawa, M. Yamano, A. Hirai & M. Takiuchi. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, p. 1408.) Equipment for recording frequency continuously is described, and results of Doppler-shift measurements made on waves reflected from the E region on 4 Mc/s over a 400-km path are given.

621.396.8 : 621.396.931.029.62/.63 **265**

Propagation Tests of Frequencies for V.H.F. Mobile Radio.—E. Shimizu, T. Morinaga, T. Kawano, S. Sato & M. Hirasaki. (*Rep. elect. Commun. Lab., Japan*, Jan. 1957, Vol. 5, No. 1, pp. 13–16.) Measurements are given of signal and noise received at a mobile station from fixed transmitters in Tokyo operating at 60, 200 and 470 Mc/s. Less fluctuation was observed than in comparable U.S. measurements, and this is attributable to the wooden houses normal in Japan.

621.396.812.3 : 621.396.666.029.64 **266**

Some Results with Frequency Diversity in a Microwave Radio System.—F. H. Willis. (*Commun. & Electronics*, March 1957, No. 29, pp. 63–67.) The results of tests with 240-Mc/s frequency separation and an appreciable switching differential are discussed.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 : 621.372.012 **267**

Signal-Flow Graphs and Random Signals.—L. A. Zadeh : W. H. Huggins.

(*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1413–1414.) Comment on 1227 of 1957 and author's reply.

621.394.3 : 621.394.828 **268**
Distortion Correction of Teleprinter Signals by Electronic Means.—E. K. Aschmoneit. (*Elektronik*, July & Aug. 1957, Vol. 6, Nos. 7 & 8, pp. 199–203 &244–245.) The principles and circuits of some electronic distortion-correcting and regenerative repeaters are described.

621.396.41 : 621.396.324 : 621.314.7 **269**
Transistorized Multiplex Radio-Teletypewriter.—P. G. Wray. (*Electronics*, 1st Sept. 1957, Vol. 30, No. 9, pp. 150–154.) Lightweight four-channel time-division multiplex equipment for use on board ship is described. Ring counter and digital synchronizer circuits using transistors give an operating speed of 100 words/min.

621.396.41.029.63 **270**
Ultra-high-Frequency Radio Equipment for 60 Speech Circuits.—(*G.E.C. Telecommun.*, Aug. 1956, No. 22, pp. 34–49.) The Type-SPO 5500 equipment described uses f.m. and operates in the band 1 700–2 300 Mc/s. Up to three r.f. channels each carrying 60 speech channels can be arranged in multiplex.

621.394.441 **271**
An Electronic Error-Correcting Multiplex Telegraph System.—(*P.O. elect. Engrs' J.*, April 1957, Vol. 50, Part 1, p. 44.)

621.396.5.029.6 : 621.396.822 **272**
Noise Considerations on Toll Telephone Microwave Radio Systems.—T. A. Combellick & M. E. Ferguson. (*Commun. & Electronics*, March 1957, No. 29, pp. 67–70.) The noise level requirements for microwave radio equipment to be used in multichannel telephone systems are discussed. The magnitude of and relation between the noise generated in the transmitter, in the receiver, or by intermodulation, are examined.

621.396.71.029.55 **273**
The Rugby 'B' High-Frequency Transmitting Station.—A. Cook & L. L. Hall. (*P.O. elect. Engrs' J.*, April 1957, Vol. 50, Part 1, pp. 15–23.) The layout and technical facilities of the station are described. There are 28 h.f. transmitters of 30-kW power rating in operation and some 70 aerials on a site of 700 acres. See also 2538 of 1956 (Booth & MacLarty).

621.396.74 : 621.397.743 **274**
Frequency Transposers as Very-Low-Power Transmitters.—A. Kolarz. (*Rundfunktech. Mitt.*, April 1957, Vol. 1, No. 2, pp. 53–57.) A 50-mW relay transmitter for providing a local television service is described. The transposition from received to transmitted frequency is made without the use of an i.f. See also 1927 of 1957.

621.396.74.029.62 : 621.376.3 **275**
Investigation of Coverage Facilities with Frequency-Offset U.S.W. F.M. Operation.—E. Belger, E. Paulsen & I. Dahrendorf. (*Rundfunktech. Mitt.*, April 1957, Vol. 1, No. 2, pp. 58–64.) The limitations and advantages of this method of obtaining regional coverage are discussed with reference to experimental results. See also 1869 of 1956 (Belger & von Rautenfeld).

621.396.931 **276**
Radio Communication in Railroad Tunnels.—H. S. Winbigler. (*Bell Lab. Rec.*, Feb. 1957, Vol. 35, No. 2, pp. 57–60.) Radiation from a continuous length of twin transmission line suspended from the tunnel wall is received satisfactorily up to 5 000 ft from the transmitter, using frequencies in the 159–162-Mc/s band.

621.396.932 **277**
Ships and Coast Stations.—N. P. Spooner. (*Short Wave Mag.*, Feb. 1957, Vol. 14, No. 12, pp. 636–639.) Details of systems, procedure, operating frequencies and general organization.

621.396.945 **278**
The Guided Radio Telephone.—W. H. Hill. (*A.T.E. J.*, April 1957, Vol. 13, No. 2, pp. 113–118.) This communication system uses currents induced in metallic conductors by the transmitter. A satisfactory portable transmitter/receiver is described for use in coal mines.

SUBSIDIARY APPARATUS

621.3-71 **279**
Removal of Heat from Sealed Miniature Equipment.—F. P. Newell. (*Brit. Commun. Electronics*, Aug. 1957, Vol. 4, No. 8, pp. 468–475.) Experimental investigation has shown that the most satisfactory way of dissipating the heat generated internally is by providing thermal conduction paths from the heat sources to the front panel.

621.311.6 : 537.311.33 : 535.215 **280**
Utilizing the Sun's Energy.—(*Elect. J.*, 1st Nov. 1957, Vol. 159, No. 4142, pp. 1256–1257.) Report on progress in the development of solar batteries of the Si *p-n*-junction type, with description of experimental installations in England.

621.311.6 : 621.314.7 **281**
A Stabilized D.C. Power Supply using Transistors.—T. H. Brown & W. L. Stephenson. (*Electronic Engng*, Sept. 1957, Vol. 29, No. 355, pp. 425–428.) A mains-operated unit is described which provides an output voltage variable from 0 to 30 V at currents of up to 1 A.

621.314.63 : 546.289 **282**
Germanium Rectifiers as Electronic Components.—J. T. Cataldo. (*Electronic Ind. Tele-Tech*, Aug. 1957, Vol. 16, No. 8, pp. 61–63 . . 168.) Discussion of advantages and applications. Cooling methods and overload characteristics are considered.

621.316.722.1 : 621.314.7 **283**
A Low-Voltage Stabilizer employing Junction Transistors and a Silicon Junction Reference Diode.—D. Aspinall. (*Electronic Engng*, Sept. 1957, Vol. 29, No. 355, pp. 450–454.)

621.316.722.1 : 621.314.7 **284**
Voltage Regulator uses Multivibrators.—W. A. Scism. (*Electronics*, 1st Sept. 1957, Vol. 30, No. 9, pp. 184–186.) Voltage changes vary the frequency of an astable transistor multivibrator and the average value of the load voltage is thus kept constant.

621.396.662.029.5 **285**
Multichannel Drives for Transmitters and Receivers in the H.F. Band.—D. J. Fewings. (*Marconi Rev.*, 2nd Quarter 1957, Vol. 20, No. 125, pp. 60–76.) Methods of tuning accurately to a prescribed frequency are described, as applied to (*a*) a diversity oscillator with manual tuning, (*b*) an automatically tuned airborne receiver/transmitter drive.

621.396.664 : 621.376 **286**
Monitoring the Modulator.—P. M. Carment. (*Short Wave Mag.*, Feb. 1957, Vol. 14, No. 12, pp. 631–632.) A circuit for monitoring the output of a high-power modulator on a pair of headphones.

TELEVISION AND PHOTOTELEGRAPHY

621.397 : 628.93 **287**
Techniques of Television Lighting.—D. Thayer. (*J. Soc. Mot. Pict. Telev. Engrs*, April 1957, Vol. 66, No. 4, Part 1, pp. 212–216.)

621.397.2 : 621.372.55 **288**
Error-Predicting D.C.-Restoring Circuits for Television Signals.—E.L.C. White. (*Electronic Engng*, Oct. 1957, Vol. 29, No. 356, pp. 472–477.) The performance of direct and negative-feedback types of d.c. restoring circuit is analysed. By a combination of slope equalization and error prediction the low-frequency cut-off of the circuit prior to d.c. restoration may be raised by a large factor. Some experimental results are given.

621.397.24 **289**
Performance of the A2A Video Transmission System.—R. W. Edmonds. (*Bell Lab. Rec.*, March 1957, Vol. 35, No. 3, pp. 85–88.) Design considerations and performance details of a broad-band local wire transmission system for monochrome and colour television.

621.397.24 : 621.315.212 **290**
A New 4-Mc/s Coaxial Line Equipment—C.E.L. No. 6A.—Collier & Simpson. (See 14.)

621.397.24 : 621.315.212 **291**
A 4-Mc/s Coaxial Line Equipment—C.E.L. No. 4A.—E. Davis. (See 15.)

621.397.242 **292**

Television Terminals for the L3 System.—J. J. Jansen. (*Bell Lab. Rec.*, May 1957, Vol. 35, No. 5, pp. 179–183.) 600 telephone channels and a 4·2-Mc/s television channel may be transmitted by the coaxial cable system the terminal equipment of which is described.

621.397.5 :535.623 **293**

Some Alternatives to the N.T.S.C. Colour Television System.—E. L. C. White. (*J. Telev. Soc.*, Jan./March 1957, Vol. 8, No. 5, pp. 191–206.) The N.T.S.C. system is compared with the two-subcarrier, the French, L.E.P. '3 double-message', and the Valensi systems. For a three-gun picture tube, the first system seems most satisfactory. The signal requirements of single-gun tubes, such as the chromatron and 'apple' tubes, are then discussed; here a symmetrical ratio system may be better.

621.397.5 : 621.395.625.3 **294**

Video Tape Recorder Design.—(*J. Soc. Mot. Pict. Telev. Engrs*, April 1957, Vol. 66, No. 4, Part 1, pp. 177–188.) Three papers, describing equipment incorporating a system of four magnetic recording heads on a revolving drum for recording transverse tracks on tape travelling at 15 in./sec. See also 4018 of 1957 (Snyder).

Comprehensive Description of the Ampex Video Tape Recorder.—C. P. Ginsburg (pp. 177–182).

The Modulation System of the Ampex Video Tape Recorder.—C. E. Anderson (pp. 182–184).

Rotary-Head Switching in the Ampex Video Tape Recorder.—R. M. Dolby (pp. 184-188).

621.397.5(083.74) **295**

Comparison of Four Television Standards.—R. D. A. Maurice. (*Electronic Radio Engr*, Nov. 1957, Vol. 34, No. 11, pp. 416–421.) "The resolutions of four C.C.I.R. standard television systems are compared, some account being taken of the effects of the asymmetric sideband reception. The extent to which some of the distortions may be due to nonlinearity of the phase-frequency characteristic is briefly mentioned."

621.397.611 : 778.5 **296**

The Limits of Optical Compensation by Polygonal Prisms in relation to Freedom from Flicker, Registration and Relative Aperture.—H. Grabke. (*Rundfunktech. Mitt.*, April 1957, Vol. 1, No. 2, pp. 65–72.) Theoretical treatment of the problem based on earlier experiments with film scanning equipment (1227 of 1956). The design of a 'polygon-ring scanner' is discussed and data for calculating tolerances and colour correction of the system are given. The manufacture of such a scanner with 40–60 faces appears feasible; it should produce satisfactory pictures comparable to those obtained by other flying-spot scanners.

621.397.621.2 **297**

Television Frame Pulse Separator.—H. D. Kitchin. (*Wireless World*, Nov. 1957, Vol. 63, No. 11, pp. 554–559.) A single-pulse circuit for accurate interlacing is described.

621.397.621.2 **298**

The Return of Electrostatic Focusing.—R. R. Pearce. (*J. Telev. Soc.*, April/June 1957, Vol. 8, No. 6, pp. 237–248.) A survey of the developments in post-war television tubes and techniques with a detailed discussion of the characteristics and relative merits of modern electrostatic guns.

621.397.82 **299**

The Visibility of Sinusoidal Interference in Television Images.—H. Grosskopf & R. Suhrmann. (*Rundfunktech. Mitt.*, April 1957, Vol. 1, No. 2, pp. 45–52.) The nature of interference patterns is reviewed and the effects of interference due to frequencies which are specially related to line or frame frequency are examined. Curves for evaluating some of these effects are given and the requisite 'safe' frequency spacing between adjacent transmitters is compared with that recommended by C.C.I.R.

TRANSMISSION

621.396.61.029.6 : 621.396.662 **300**

Transmitter Tuned by Distortion Indicator.—C. R. Ellis, K. Owen & G. R. Weatherup. (*Electronics*, 1st Sept. 1957, Vol. 30, No. 9, pp. 180–183.) The system described gives harmonic distortion of less than 3% r.m.s. over the range 200 c/s–20 kc/s when the transmitter is operated at 225–400 Mc/s, with 80% a.m. of a 1-kW carrier.

621.396.61.072.9 **301**

A New Device for the Automatic Radio-Frequency Synchronization of Common-Frequency Broadcasting in the Medium-Wave Band.—A. Karaminkov. (*NachrTech.*, April 1957, Vol. 7, No. 4, pp. 163–166.) In the system proposed synchronization is carried out by radio transmission during breaks in the program. During these interruptions of about 12-s duration which are intentionally introduced at intervals of about 1 h, the carrier of the local transmitter is stopped so that the distant master transmitter can be received for the automatic frequency correction of the local carrier. Using quartz oscillators with long-term stability 1 part in 10^6 an hourly stability within about 1 part in 10^8 can be achieved.

VALVES AND THERMIONICS

621.314.63 : 621.317.3 **302**

Measurements of Crystal Impedances at Low Levels.—H. N. Dawirs & E. K. Damon. (*Trans. Inst. Radio Engrs*, April 1956, Vol. MTT-4, No. 2, pp. 94–96. Abstract, *Proc. Inst. Radio Engrs*, July 1956, Vol. 44, No. 7, p. 956.)

621.314.63 : 621.372.632 **303**

The Frequency Dependence of Noise Temperature Ratio in Microwave Mixer Crystals.—M. E. Sprinks, G. T. G. Robinson & B. G. Bosch. (*Brit. J. appl. Phys.*, July 1957, Vol. 8, No. 7, pp. 275–277.) A discussion, with experimental results, of the contribution of the noise temperature ratio of the mixer to the overall noise factor of the receiver. The latter has an intermediate frequency in the range 10–60 Mc/s.

621.314.7 **304**

Some Aspects of Transistor Progress.—H. W. Loeb. (*A.T.E. J.*, April 1957, Vol. 13, No. 2, pp. 119–134.) Reprint. See 295 of 1957.

621.314.7 **305**

Heat Sinks for Power Transistors.—O. J. Edwards. (*Mullard tech. Commun.*, March 1957, Vol. 3, No. 22, pp. 59–61.) Sinks of mild steel and aluminium are examined and the effects of thickness, area, and surface treatment are discussed.

621.314.7 **306**

Design Theory for Depletion-Layer Transistors.—W. W. Gärtner. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1392–1400.) This new type of high-frequency transistor uses maximum attainable carrier velocities in solids by injecting electrons or holes into the high electric fields prevailing in properly-designed depletion layers of reverse-biased *p-n* junctions. Low- and high-frequency performance, small-signal behaviour, power gain and stability are discussed in a particular theoretical design. The short transit times should be important for solid-state microwave amplification.

621.314.7 **307**

Design for an Improved High-Frequency Transistor.—C. Thornton, J. Roschen & T. Miles. (*Electronic Ind. Tele-Tech*, July 1957, Vol. 16, No. 7, pp. 47–49 . . 124.) The construction and properties of the Type L-6100 silicon surface-alloy transistor are described. It has a base width of 7×10^{-5} in. and a maximum oscillation frequency of about 16 Mc/s.

621.314.7 **308**

The Junction Transistor as a Charge-Controlled Device.—R. Beaufoy & J. J. Sparkes. (*A.T.E. J.*, Oct. 1957, Vol. 13, No. 4, pp. 310–327.) Transistor operation is described in terms of the charge in the base region so that the transistor performance is defined in the form of time constants (pC/μA). The analysis was developed for large-signal operation but yields complete expressions for small-signal admittance parameters. Methods of measurement are outlined.

621.314.7 **309**

Flow Line Analysis.—R. B. Hurley. (*Electronic Ind. Tele-Tech*, April 1957, Vol. 16, No. 4, pp. 52–54 . . 140.) A pictorial representation of the expressions derived by Ebers & Moll (884 of 1955) to analyse transistor behaviour in various types of switching circuit.

621.314.7 : 546.289 **310**

Influence of Surface Oxidation on Alpha$_{cb}$ of Germanium *p-n-p* Transistors.—J. T. Wallmark. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 255–271.) Oxidation of the Ge surface is shown to be a major factor influencing surface recombination and thereby the slow decay of the current transfer ratio α_{cb} in *p-n-p* transistors. An outline of the underlying processes is given.

621.314.7 : 621.317.3 **311**

Transistor Tests Predict High-Frequency Performance.—Pritchard. (See 213.)

621.314.7 : 621.318.57 **312**

Determination of Transient Response of a Drift Transistor using the Diffusion Equation.—H. B. von Horn & W. Y. Stevens. (*IBM J. Res. Developm.*, April 1957, Vol. 1, No. 2, pp. 189–191.) The one-dimensional diffusion equation is solved to give the time required for the collector current to reach a steady value in response to a step function of emitter current.

621.314.7 : 621.373.51 **313**

Unique Properties of the Four-Layer Diode.—W. Shockley. (*Electronic Ind. Tele-Tech*, Aug. 1957, Vol. 16, No. 8, pp. 58–60 .. 165.) Description of a new bistable two-terminal semiconductor device and details of some of its applications. See also 3899 of 1956 (Moll et al.).

621.314.7 : 681.142 **314**

The Multipurpose Bias Device: Part 1—The Commutator Transistor.—Dunham. (See 38.)

621.314.7 : 681.142 **315**

Two-Collector Transistor for Binary Full Addition.—R. F. Rutz. (*IBM J. Res. Developm.*, July 1957, Vol. 1, No. 3, pp. 212–222.) Two versions of a two-collector transistor are described, one using point contacts, and the other using *p-n* 'hook' junctions as collectors. The function of such a transistor as a binary adder, and its circuit characteristics, are discussed. See also 38 above.

621.314.7 (083.7) **316**

Semiconductor Symbols.—P. M. Thompson & J. Bateson. (*Wireless World*, Nov. 1957, Vol. 63, No. 11, pp. 525–528.) A logical system for diodes, transistors and other junction devices is proposed.

621.383.27 **317**

Photomultiplier Tubes.—J. Sharpe. (*Brit. Commun. Electronics*, Aug. 1957, Vol. 4, No. 8, pp. 484–492.) The design and applications of the various types of photomultiplier tube are described. A table of representative tubes available in the U.K. is given.

621.383.27 : 621.387.464 **318**

Improved Performance of Photomultipliers in Scintillation Counters.—A. Ashmore, B. Collinge & S. K. Sen. (*Indian J. Phys.*, May 1957, Vol. 40, No. 5, pp. 261–264.) A cathode follower reduces the effective capacitance to earth of the collecting electrode of a photomultiplier, enabling bigger voltage pulses to be obtained for the same degree of nonlinearity in the photomultiplier.

621.383.4 : 535.371.07 **319**

An Improved High-Gain Panel Light Amplifier.—B. Kazan. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1358–1364.) A grooved photoconductor light-amplifying picture panel is described, whose gain is more than 10 times greater than any previous amplifier and whose threshold for input light is reduced. See also 933 of 1956 (Kazan & Nicoll).

621.385 : 621.317.321 **320**

A Survey of Methods Used to Determine Contact Potentials in Receiving Tubes.—E. R. Schrader. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 243–254.)

621.385 : 621.317.444 **321**

Rectilinearity of Electron - Beam Focusing Fields from Transverse Component Determinations.—Cioffi. (See 222.)

621.385.029.6 **322**

Investigation of a Special Type of Reflex Klystron.—W. N. Shevchik, S. A. Suslov & Yu. D. Zharkov. (*Zh. tekh. Fiz.*, Feb. 1957, Vol. 27, No. 2, pp. 377–386.) A theoretical and experimental investigation of a reflex klystron operating with a large transit angle ($> 2\pi$) in the interaction space is described. Good agreement is found between the theoretical evaluation of the valve efficiency and the experimental results.

621.385.029.6 **323**

On the Nonlinear Behaviour of Electron-Beam Devices.—F. Paschke. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 221–242.) The nonlinear space-charge-wave equation is derived and reduced by third-order successive approximation to three simultaneous linear differential equations which are solved for the case of a velocity-modulated electron beam. Nonlinear phenomena in travelling-wave valves can be treated similarly.

621.385.029.6 **324**

Electron Beam Focusing in Three-Anode Guns for Travelling-Wave Tubes.—T. S. Chen & L. Kovach. (*J. Electronics Control*, Sept. 1957, Vol. 3, No. 3, pp. 287–304.) The electron trajectory in the accelerating region is solved and beam rippling in the drift region discussed. The principle of compensated three-anode guns is developed in the form of design formulae and curves. Experiments confirm that compensated guns result in a reduction of the focusing magnetic field for parallel-flow beams.

621.385.029.6 **325**

The Large - Signal Behaviour of Crossed-Field Travelling-Wave Devices.—J. Feinstein & G. S. Kino. (*Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, pp. 1364–1373.) The limits of validity of the conventional small-signal theory are obtained theoretically, and the analysis is extended to follow the interaction to power saturation. Forward-wave and backward-wave amplifiers and a backward-wave oscillator are considered. In the amplifier valves the gain is inherently low, and suggestions for improving this deficiency are examined.

621.385.029.6 **326**

An Experimental High-Power Pulsed Travelling-Wave Tube.—J. F. Gittins, N. H. Rock & A. B. J. Sullivan. (*J. Electronics Control*, Sept. 1957, Vol. 3, No. 3, pp. 267–286.) A high-impedance slow-wave structure with good thermal dissipation and suitable for a 100-kV electron beam is used. A saturated efficiency of 34%, power output of 3 MW and a gain of 35 dB are obtained.

621.385.029.6 : 621.372 **327**

Interdigital and other Slow-Wave Structures.—J. C. Walling. (*J. Electronics Control*, Sept. 1957, Vol. 3, No. 3, pp. 239–258.) A method of calculating the properties of structures consisting of arrays of parallel conductors is developed and a table of numerical values is presented. This is applied to obtain dispersion characteristics, input and beam-coupling impedances for the Karp structure, interdigital line and tape helix.

621.385.029.6 : 621.376 **328**

The Modulation of Travelling-Wave Tubes.—G. F. Steele. (*Electronic Engng*, Sept. 1957, Vol. 29, No. 355, pp. 429–433.) "The simple theory of travelling-wave tube modulation is discussed and some expressions derived for the cases of phase and amplitude modulation. Details are given of the experimental procedure used to measure phase shift, and the sideband power resulting from a 60-Mc/s modulation."

621.385.029.64 **329**

Velocity-Jump Amplification at 10 000 Mc/s.—B. V. Dore. (*Canad. J. Phys.*, June 1957, Vol. 35, No. 6, pp. 742–752.) Description of a v.m. valve based on an earlier design [2068 of 1951 (Field et al.)]. Theory previously developed for space-charge-wave amplifiers is applied to its operation and conclusions are drawn regarding the operation of coupling helices and the effect of small gaps in the amplifying section of the valve. See also 3721 of 1954 (Agdur).

621.385.832 : 621.317.799 **330**

An Electrostatic Cathode-Ray-Tube Tester.—R. B. Hale. (*Brit. Commun. Electronics*, Aug. 1957, Vol. 4, No. 8, p. 496.)

MISCELLANEOUS

621.39 (091) **331**

Fifty Years of the Institution of Post Office Electrical Engineers.—(*P.O. elect. Engrs' J.*, Oct. 1956, Vol. 49, Part 3, pp. 147–274.) A number of papers surveying the history and activities of the Institution and detailing its achievements in all fields of telecommunications.

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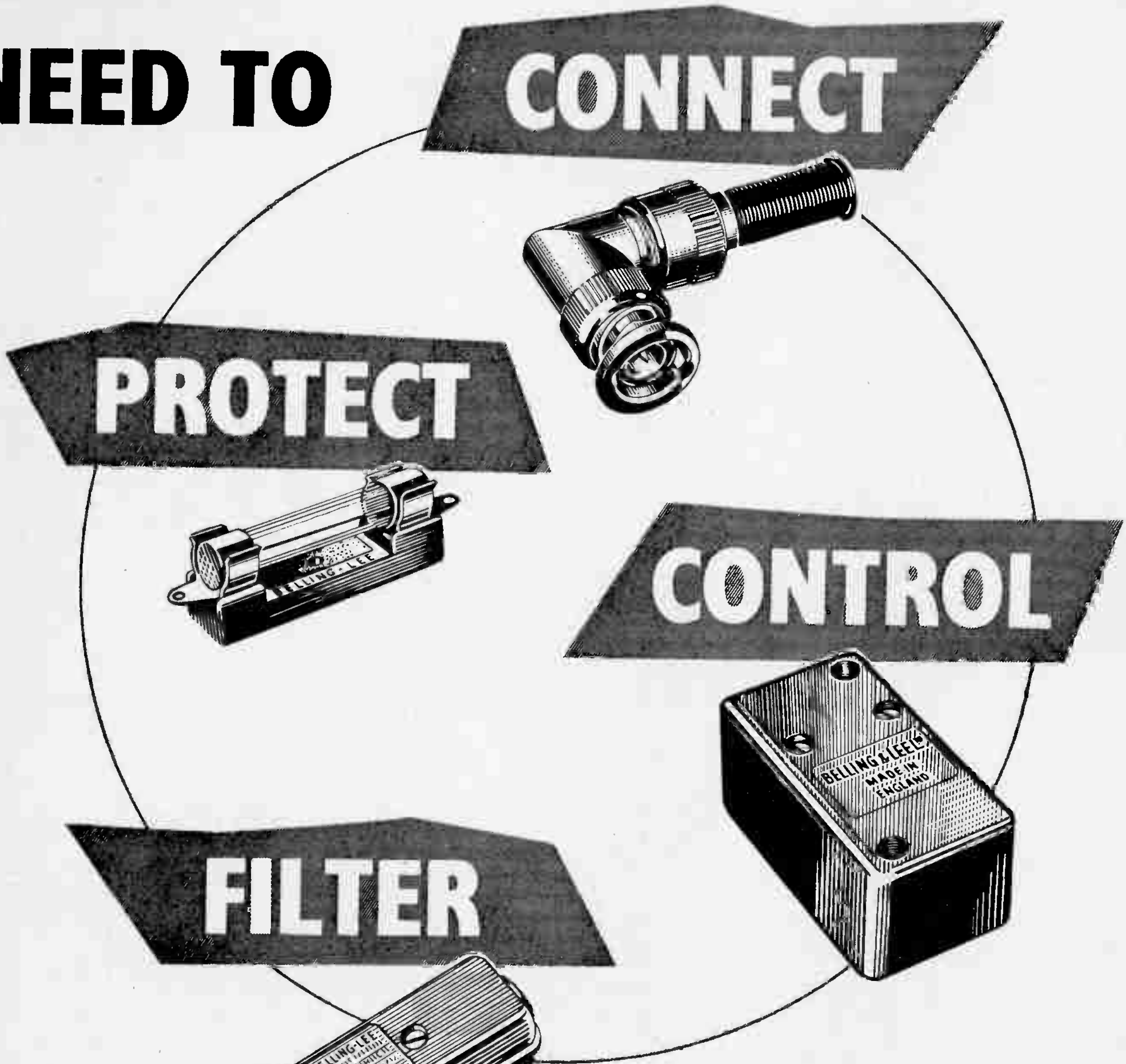

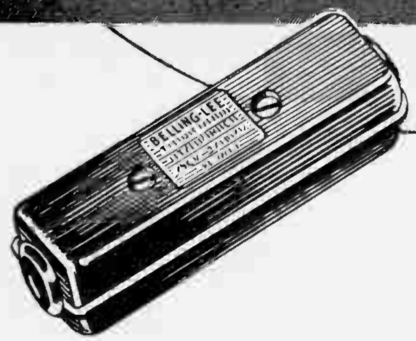

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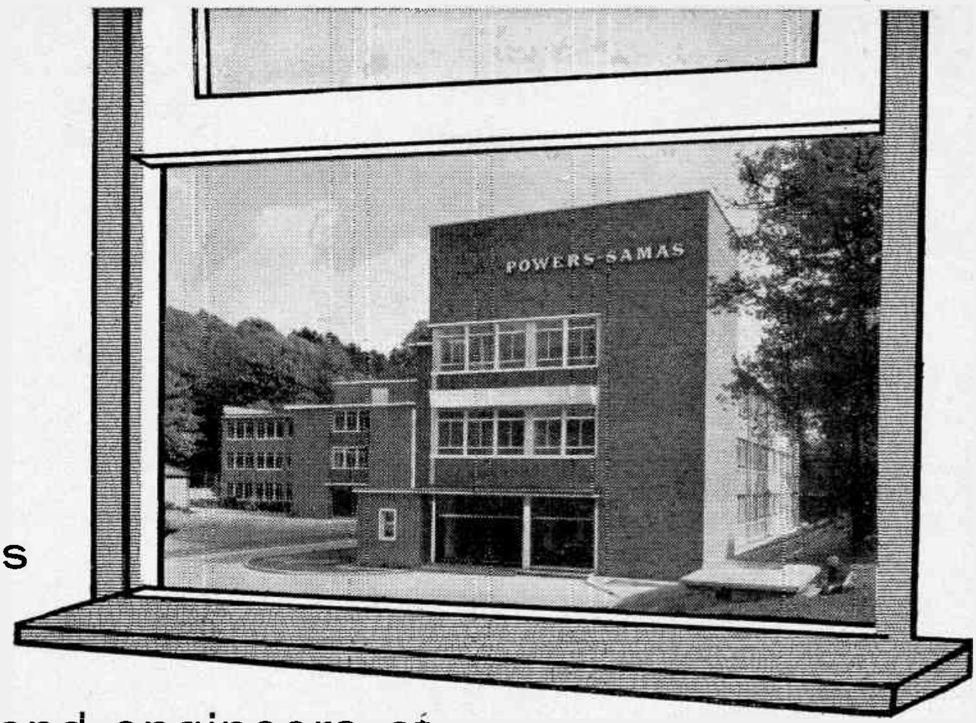

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