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## Commentary

**T**HE demand for coal has risen rapidly during the past years and one cause has been the increasing demand for electrical energy. Figures recently published by the British Electricity Authority show just how phenomenal has been the demand for electrical energy in this country.

From a mere 5 600 million units in 1925 the annual demand rose to about 30 600 million units in 1945—a five-fold increase in twenty years. The growth has continued in the post-war years and at the end of last year the demand was about double the 1945 figure—some 60 000 million units. Yet the end is not in sight, for it is estimated that in 20 years' time the demand will be some  $3\frac{1}{2}$  times the present level.

Practically all this energy is produced by the coal fired power stations and unless some alternative source of energy can be found, and found quickly, the prospects of Britain's survival as an industrial power would indeed be very bleak.

Not a moment too soon, therefore, comes the announcement of the Government's plans to build twelve nuclear power stations during the next ten years, with a total capacity of  $1\frac{1}{2}$ -2 million kilowatts, and at a cost of some £300 million.

As is already known, an experimental power station is in the course of construction at Calder Hall in Cumberland, at a cost of £15-£20 million, to produce electrical power on a large scale from this new source, and this station will form the basis of the design of the future nuclear stations.

During the next ten years two types of reactors are likely to be brought into use on a commercial scale. The first—an improved version of the Calder Hall (gas cooled) type—could be designed and brought into operation within the next four years. This type will burn only a small proportion of the natural uranium as fuel but will produce, as well as heat, the valuable by-product of plutonium.

The second type of reactor will almost certainly be liquid cooled and will be more efficient in that it will produce more heat from the same amount of fuel than the gas cooled type, for the same capital cost.

It is proposed that constructional work will start on two stations of the Calder Hall gas cooled reactor type in 1957 and that they will be in operation during 1960-61. These will be followed by two more stations using improved gas cooled reactors in 1958-9 to be completed in 1963 and the four stations will have an output of from 400 000 to 800 000 kilowatts.

The second stage of the construction of the remaining eight stations, probably with liquid cooled reactors, will start in 1960 and, when completed in 1965, will have an installed capacity of well over 1 million kilowatts.

The stations will be built by private industry for the Electricity Authorities who will own and operate them, but the Atomic Energy Authority, as the only body with the necessary

experience, will be responsible for giving technical advice on the nuclear plant, since British industry as yet has no experience of nuclear technology.

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Two men, each of whom has, in his own way, made outstanding contributions to the advancement and understanding of our science, have died recently.

One—an American—was a brilliant mathematical physicist while the other—British—was of lesser education yet possessed of an extraordinary inventive mind.

George Ashley Campbell who died in November of last year, joined the American Bell Telephone Company in 1897 at a time when the problems of telephone transmission were little understood. He had just returned from four years of study at various universities in Europe and his first task was to find some method of reducing the attenuation of telephone lines.

Familiar with the work of Heaviside, Campbell was able to develop the theory of loading, but there followed one of those coincidences, which all too frequently happen, in which the same result was arrived at simultaneously by another investigator—in this case Professor Pupin, at Columbia University. Pupin was adjudged the originator on the basis of earlier conception but Campbell pursued his studies on loaded lines and eventually arrived at the electric wave filter which was perhaps his greatest contribution. It opened the way to the carrier current system of telephony and from this has developed the multi-channel radio telephone and carrier systems of today.

Donald Macadie who died in February of this year had none of the brilliant academic background of Campbell. Born in 1871 in Scotland, he came to London at an early age and joined the Edison Bell Phonograph Company, later transferring to the National Telephone Company. When the latter firm was absorbed by the Post Office he was given an appointment at the Birmingham Post Office Factory and he remained with the Post Office until his retirement in 1935.

Macadie had a remarkable inventive ability and he produced a number of ingenious devices which are now in common use throughout the world. He was largely responsible for the design of a machine used to sort out the millions of Postal Orders returned after cancellation and, in addition, he developed an automatic coil winding machine of great flexibility which formed the basis of the present day designs.

Yet his fame probably rests on what is by comparison a much simpler device—namely the first combined multi-range ammeter voltmeter and ohmmeter which subsequently became known as the Avometer.

# A High-Speed Revolution Counter

By E. L. Harrington\*

*The instrument described herein has been designed for measuring the rotational speed of gas turbine engines to within  $\pm 1$  rev/min at 20 000 rev/min. It operates by counting a frequency proportional to speed over a one second sampling period, and at sampling intervals of three seconds. An unusual feature of the equipment, apart from accuracy, is the flexibility of the control unit.*

THE progress of engineering has always been largely dependent on new instrumentation techniques. With the arrival of the turbine in aero engines, there has been a call for more accurate and faster response instruments with wide ranges of measurement. On the particular problem of rotary speed measurement many papers have been published<sup>1,2,3</sup>. For research gas turbines and compressors running at high speeds, where the characteristic of thrust against speed lies between a cubic or quintic curve it is important to know the speed accurately.

## Consideration of Techniques

A specification for the instrument to meet all research requirements would read as shown in Table 1, columns (a) and (b).

TABLE 1.

(a) PERFORMANCE PARAMETER	(b) DESIRED	(c) PROPOSED
Reliability ..	Absolute	Good
Accuracy ..	Absolute	$\pm 1$
Range ..	0 to 20 000 rev/min	1 000 to 20 000 rev/min
Response ..	Instantaneous	Within 3sec max
Answer ..	Direct reading	Direct reading
Display ..	(In duplicate if required)	(In duplicate if required)
Recording ..	Photographic	Photographic With camera synchronizing contacts
Simplicity of operation ..	Fully automatic	Fully automatic
Simplicity of technique ..	Absolute	Not very simple
Simplicity of construction and mainten- ance ..	Absolute	Fairly good
Ground/Flight operation ..	Both	Either

The decision on measuring techniques was settled by the accuracy requirement. Counting is the only practicable method which approaches absolute accuracy for measuring a continuously variable number. Having decided on a counting technique using Dekatrons<sup>4,5</sup>, the remainder of the proposed specification was largely predetermined.

Some consideration remains for the automatic operation of the equipment in order to relieve a flight test observer of added responsibility, and on the ground to avoid inept operation. Laboratory experiments showed that it was pos-

sible to read a random 5-digit number displayed on Dekatrons in 2 seconds, but impossible to do this in 1 second. Since the number to be displayed would not be random, but invariably have a pattern such as 12579, 12581, 12582, etc., with probably only the last two digits varying, a 1.8sec reading or display time was allowed. This with a 0.2sec reset period, and 1sec counting period, gives a 3sec duty cycle for the instrument, and accounts for the "3sec maximum" response time in Table 1, column (c).

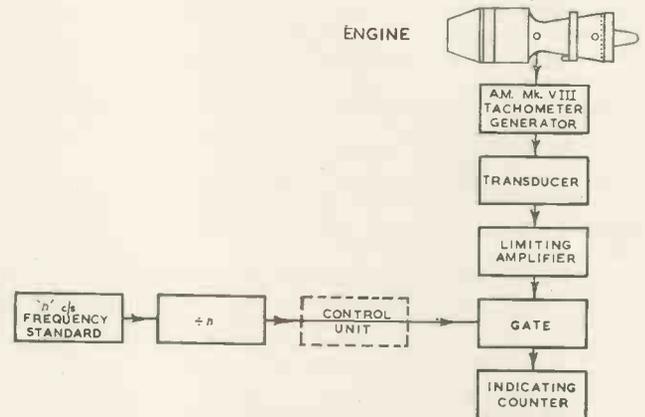


Fig. 1. Basic frequency counter

## The Functional Diagram

The simplest form of revolution counter is shown in Fig. 1. It employs a timing standard of  $nc/s$  followed by a scale of  $n$  divider which produces 1sec timing pulses. These timing pulses open and close a gate, thus controlling the flow of signal pulses from a source or transducer device, to an indicating counter. There is no provision for resetting the indicator, nor for adjusting the duty cycle. To perform these two functions, a control unit is inserted between the scale of  $n$  divider and the gate.

A particular control unit function is defined by Fig. 2. The unit receives two trains of 1p/s signals on separate channels, the pulses on channel *A* being delayed by 0.2sec from those on channel *B*. Pulses on channel *A* operate a three-state circuit and are also fed to switch the gate. The three-state device and a following time delay together block off one pulse out of every three applied to the gate; it is arranged for the gate to be open for one second and shut for two seconds. It should be noted that the exact value of the time delay is not at all critical, and hence a slugged relay will suffice. Reference should be made at this point to the timing sequence diagram in Fig. 3.

\* The Bristol Aeroplane Co. Ltd.

The pulses on channel *B* are fed to a circuit which is required to provide the reset signal to the display unit at 0.2sec before a count period. The reset circuit is held quiescent until the one-second period before the count period. It thus ignores the triggering pulses on channel *B*. When the three-state circuit is in the condition immediately preceding the count period, it prepares the resetting circuit for action. This action is triggered by the channel *B* pulse, and is only possible when a channel *B* pulse occurs during the "reset" condition of the three-state circuit. In the foregoing it has been assumed that the pulses in channels *A* and *B* are perfectly synchronous and lack coincidence by 0.2sec. This is so, and is simply achieved, as shown later.

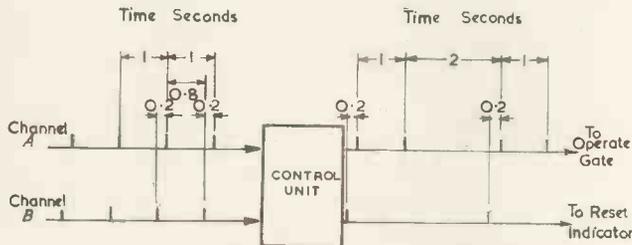


Fig. 2. Definition of control unit function

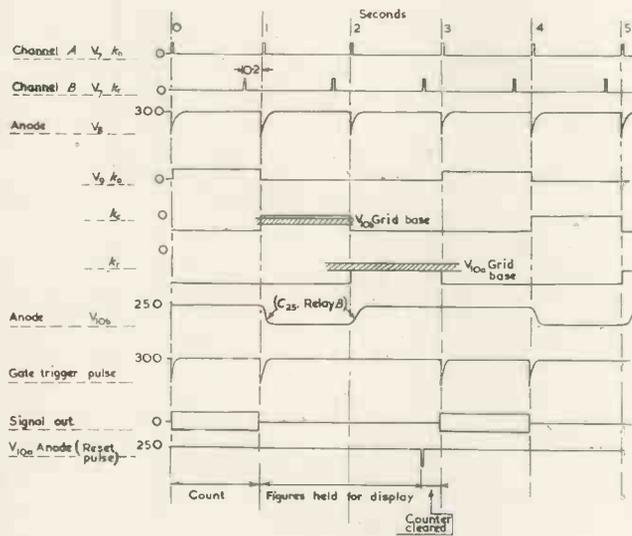


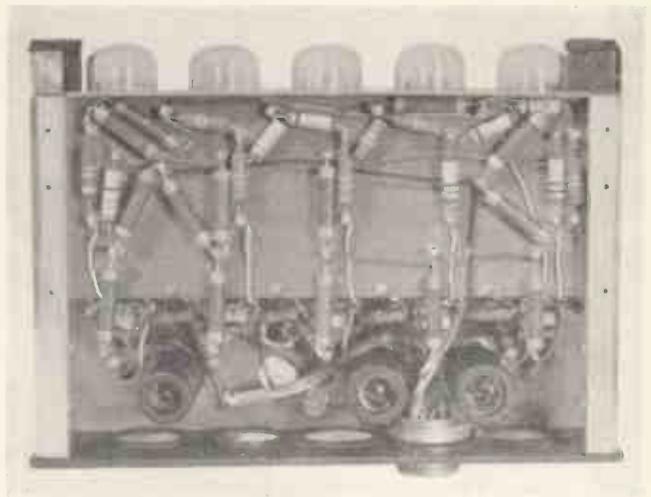
Fig. 3. Timing sequence

### Detailed Circuit Techniques

The circuit diagram is shown in Fig. 4.

A 1kc/s crystal was chosen as a frequency standard. This permits direct scale-of-ten dividing to derive the 1sec timing pulses. The crystal is not temperature controlled, but has been cut to give an accuracy of 1 part in 20 000 at 20°C. By suitable padding with negative temperature coefficient capacitors it is estimated that the error due to the crystal can be made negligible over a temperature range of +40°C to -60°C.

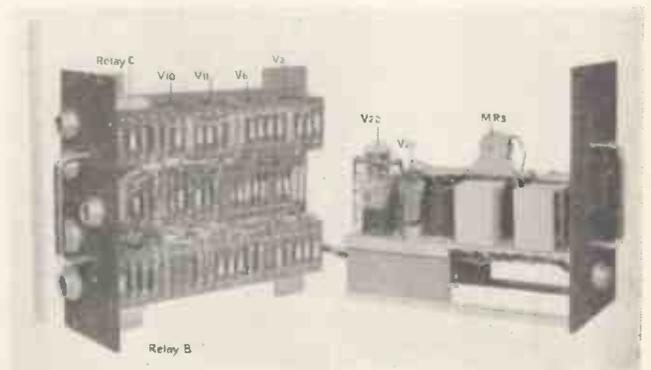
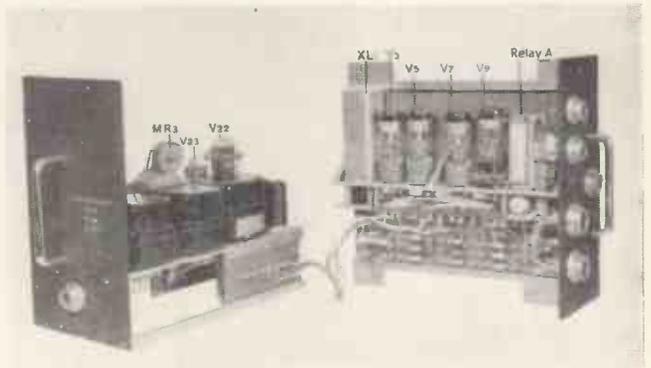
The last stage ( $V_7$ ) of the scale-of-1000 divider uses a valve type GC10/4B, which is capable of giving an output at the 8<sup>th</sup> and 10<sup>th</sup> input pulses. Thus the 8<sup>th</sup> output cathode  $k_7$  provides the channel *B* pulse 0.2sec before the 10<sup>th</sup> cathode  $k_0$  provides the channel *A* pulse.



The indicator unit showing the counter tubes at the top

A valve type GS12C is used as the three-state circuit  $V_9$ . This valve has 12 output cathodes, every 3<sup>rd</sup> one being commoned, to form three groups of four cathodes. The  $k_0$  cathode circuit,  $R_{40}$ , simply forms a waiting period while the count is in progress. The next,  $k_0$ , cathode circuit in time (and position),  $R_{03}$ , raises the grid potential of  $V_{10b}$  causing it to conduct heavily and energize relay *B*. This relay removes the trigger pulse from  $V_8$  to the gate pair  $V_{12}, V_{13}$ , and since the relay has a delayed release, the gate misses the pulse which it should receive at the instant  $V_{10b}$  is de-energized. The gate is left shut.  $V_{12}$  cut off, and the

Two views of the control and power units





GS12C ( $V_9$ ) third cathode circuit,  $k_1$ , is now energized,  $R_{50}$ ,  $R_{51}$ . The 3<sup>rd</sup> cathode of  $V_9$  now raises the grid of  $V_{10a}$  to the point of conduction.  $V_{10a}$  does not conduct until it receives the  $B$  channel pulse, 0.8sec after it received the preparation signal from  $V_9$ . This time is also 0.2sec before the preparation signal disappears. Thus  $V_{10a}$  is caused to conduct heavily for a time of much less than 200msec, and in doing so energizes the operating winding of relay  $A$ . One pair of relay  $A$  changeover contacts remove an earth from the guide bias circuits and common cathodes of all the GC10D indicating Dekatrons, and take these electrodes to a positive potential. The glow discharge can then only be maintained on the output cathodes of the indicator Dekatrons.

The function of relay  $B$  is not only to block off every third-gate operating pulse. It is also the "camera" relay, one pair of contacts, not shown, being brought out to a socket and providing a camera synchronizing signal at the correct time in the duty cycle.

Two modifications were introduced to the basic design in the light of customer experience. One of these was a test switch,  $S_3$ , which removes the signal produced by the transducer and supplies in its place the 1kc/s frequency standard to the gate circuits. The other modification is a "hold" switch,  $S_2$ , which in series with a pair of relay  $B$  normally open contacts, short-circuits the output of  $V_5$  in the second division stage of the control unit divider. Thus when  $S_2$  is operated and the "camera" relay made the cyclic controlling sequence will stop with the latest result displayed on the indicator.

### Starting Sequence

The foregoing operation sequence will only be performed if it starts in the correct manner, i.e. if the sequence starts on the "wrong foot" then the gate may be open for two seconds instead of one. To fulfil the requirements of a correct sequence, two conditions are necessary, and sufficient, viz, the gate must be shut and the GS12C ( $V_9$ ) three-state circuit must be in a corresponding position. The gate is held shut when its power supply is energized, by asymmetry of the grid resistors,  $R_{70}$ ,  $R_{71}$ , the former being made lower to ensure that  $V_{12}$  is cut off. The start condition of  $V_9$  is that the glow initially occurs on cathode  $k_6$ ,  $R_{53}$ , since this cathode is highly negative. When the glow of  $V_9$  invests  $k_6$ ,  $V_{10b}$  conducts and holds off any random disturbances to the gate pair.

It is worth noting at this stage that the control unit is of such a form that the three-state circuit,  $V_9$ , may in fact have any number of states within the valve limitations, i.e. three to twelve. Also the duration of the states of  $V_9$  may be simply altered by changing the cathode connexions on the valve itself. A proposed development of this instrument in order to read 100 000 rev/min will have a six-second counting state, a three-second display state, and a one-second reset state. This will be achieved by plugging into the socket now receiving  $V_9$  (GS12C) an appropriately connected valve type GS10C.

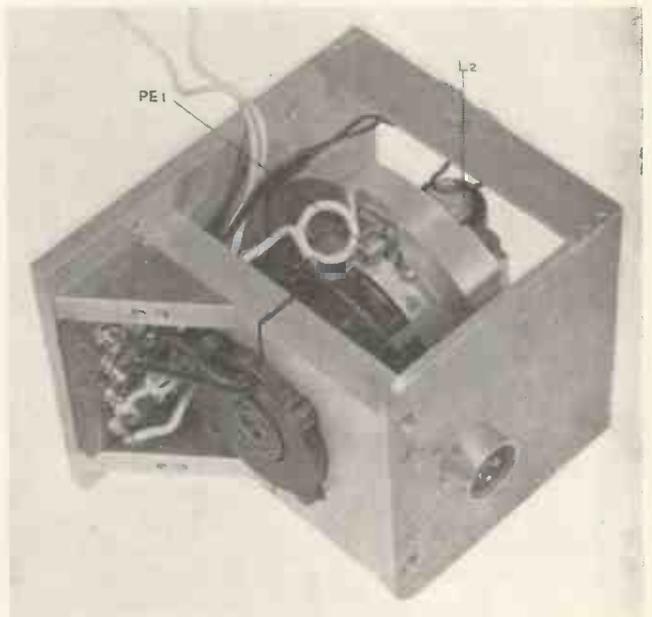
### The Transducer

In order to measure a speed of  $n$  rev/min to an accuracy of  $\pm 1$  and use a sampling period of one second, it is necessary to generate  $nc/s$  at  $n$  rev/min. Hence the transducer or tachometer generator must generate 20kc/s at an engine speed of 20 000 rev/min.

Aero-engines are normally fitted with a three-phase tachometer generator which runs at either  $\frac{1}{2}$  or  $\frac{1}{4}$  engine speed from

an accessory gearbox. These generators drive sets of tachometer indicators which are used for approximate indication of speed. The tachometer indicator consists in part of a synchronous-induction motor, i.e. a three-phase motor with a permanent magnet rotor and a copper cage. Such a construction gives the good starting characteristic of an induction machine, and a synchronous running characteristic.

To form a suitable transducer, such a motor unit was taken, and fitted with a disk having 240 peripheral slots. The factor 240 is comprised of 4 (for the engine gearbox) times 60 (pulses per revolution). The slots are each about 0.018in wide, and light from an ordinary 2 watt dial light is focused on to them by a single



*The transducer assembly*

lens. A germanium photocell receives the gated light, and produces a sinusoidal signal of about 2V peak-to-peak at all conditions. Some modulation of the signal envelope is inevitable, and is composed of about 10 per cent at 100c/s and some 5 per cent at disk frequency, the latter being due to irregular cutting of the slots. Since the signal is finally limited this modulation is of little consequence. The major difficulty anticipated was that of hunting of the motor when attempting to following a sudden change in engine speed. Damping is present in the form of eddy currents in the machine rotor bars and windage of the slotted disk. Tests have shown that no overshoot results from a step change in speed, and the total recovery time for a sudden change of 1 000 motor rev/min (4 000 engine rev/min) is about 100msec. Hunting does occur at low engine speeds, less than 3 000 engine rev/min where the synchronizing torque between motor and generator is also low. It is proposed to attempt viscous damping of future transducer motors if necessary, but the worst error introduced by hunting seems to be about  $\pm 2$  rev/min.

### Power Supplies

Two major decisions, i.e. the function of the control unit, and the use of Dekatrons, leave little choice in circuit design. The sole remaining problem is the power supply. All lines except the 6.3V a.c. must be stabilized, and the

supplies decided on were:

- +390V Dekatrons
- +310V Trigger Tubes
- +240V General Thermionic Valves
- 75V General Bias

Considerable efforts were made to simplify the power supplies. The bias supply is essential and no advantage is gained from changing its value. It was decided to series valve stabilize the 240V line for general thermionic valves, and to use this voltage as a reference level, adding to it 150V for the other line. This arrangement gives all the desired supply voltages with a minimum number of valves. A metal rectifier was chosen for the bias supply, since it has no warming-up period, and thus the bias is available before other voltages. This condition is necessary to fulfil the starting arrangements of the timing sequence, as shown previously. It was also necessary to use an indirectly heated

to attach a test lead to any point in the circuit, and a soldering iron can be manoeuvred inside the equipment with reasonable care.

Construction has been carried out in general to R.S.C.S. standards, material supplies permitting. The proportion of high stability and/or close tolerance components is 10 per cent, but production experience would be required to determine whether this allows a suitable safety margin. Valve interchangeability is considered satisfactory, the equipment performs reliably with all thermionic valves +20 per cent and -40 per cent of normal rating. A problem of parallax error on the indicator unit was considerably reduced by placing the engraved number panel in the same plane as the dekatron glow discharge, and designing the case so that when mounted in a panel the number plate was also in the plane of the panel.

#### Acknowledgments

The Author wishes to thank the Directors of the Bristol Aeroplane Company for permission to publish this article. The assistance of various colleagues, notably Messrs. R. F. Miller and D. Mills is also gratefully acknowledged.

#### APPENDIX

##### NOTES ON TEST TECHNIQUES

The accuracy of the instrument tested as a frequency meter devolves on the crystal frequency and the minimum resolving time of the counter. If the crystal is accurately set up then two simple tests are sufficient to determine the performance of the equipment.

##### (a) Gate Timing Accuracy

It is considered necessary to check the accuracy of the timing circuit in view of the relatively slow rate of change of voltage available from trigger tubes<sup>6</sup>. The error to be considered is the randomness in time of the gate opening and closing events. A slight unbalance deliberately inserted in the gate valves will cause an output pulse due to change of state under no-signal conditions. This pulse is differentiated and applied to one beam of a double beam c.r.t. The other beam is excited by the crystal at 1000c/s, the time-base being locked to it with a 1000 $\mu$ sec sweep. With a reasonable sweep length, say 6in, a randomness of 10 $\mu$ sec in gate timing is detectable, and this order of sensitivity is sufficient for our purposes.

##### (b) Counting Accuracy

The obvious method of testing the counter, that of making it count its own frequency standard, is not very satisfactory since the timing standard is 1kc/s and the maximum signal frequency is 20kc/s. It was, therefore, decided to generate 20kc/s and 1kc/s synchronous signals. These were fed to the indicating and timing circuits respectively through phase shifting networks with a 360° coverage at 20kc/s. It was found that over a wide phase range the counter would indicate 20kc/s, that the minimum resolving time of the counter was less than 50 $\mu$ sec, and that the upper frequency limit was correspondingly above 20kc/s.

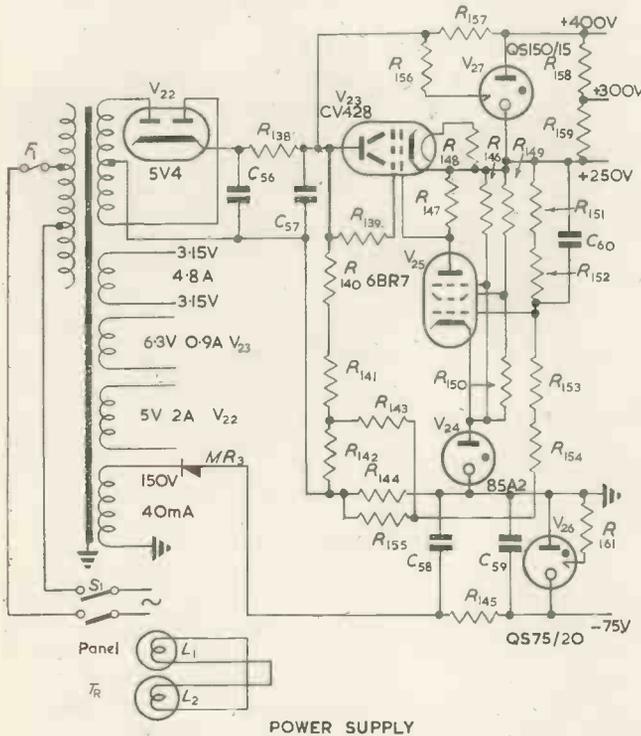


Fig. 5. The power supply unit

main rectifier valve  $V_{22}$  and use the thermal time-constant of this valve to delay the application of d.c. to the power supply stabilizer. The d.c. output of the stabilizer will not then tend to overshoot on switching on and cause the trigger tubes to fire and upset the starting conditions.

The circuit diagram is shown in Fig. 5. For a flight power supply, three single-phase 400c/s transformers are used, with an appropriate change in filter and reservoir capacitors.

#### Construction

The complete equipment is housed in two cases, an indicator for remote viewing, and a case containing the power supply, transducer and controller. These latter units are assembled on a front panel which is withdrawn from a case for servicing. Each unit is detachable and interchangeable for replacement purposes, this condition applying to all units for flight or ground operation. It is possible

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# Some Aspects of V.H.F. Sound Broadcasting and F.M. Broadcast Stations

(With Reference to the Performance of the BBC's Wrotham V.H.F. Station)

(Part 2)

By P. A. T. Bevan\*, M.I.E.E., Sen.M.I.R.E.

## F.M. MODULATORS

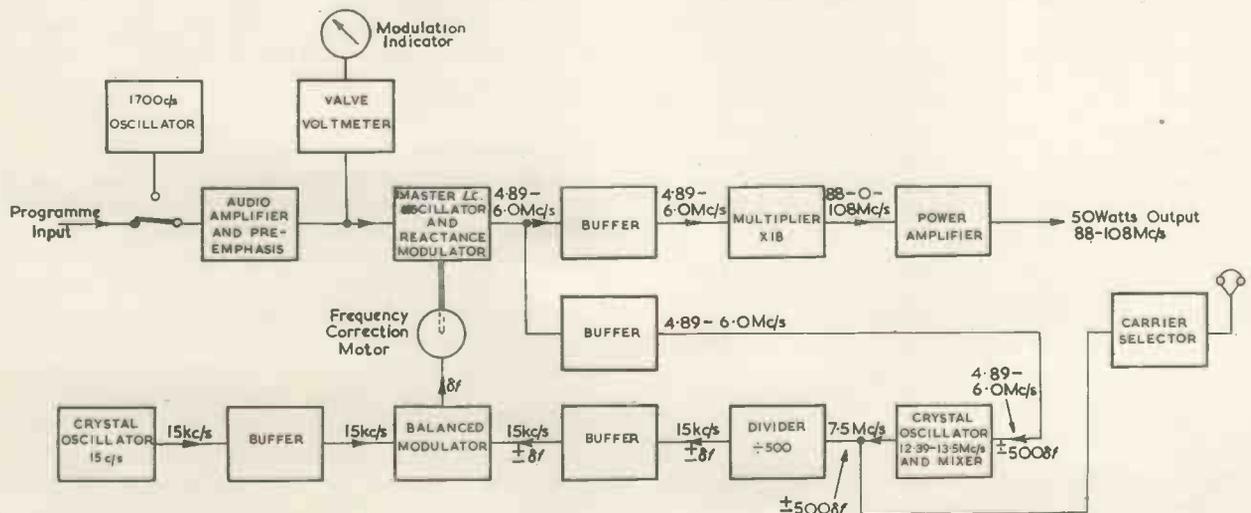
The performance of an f.m. transmitter is chiefly dependent on that of its f.m. modulator. Until recently the most generally used method of effecting frequency modulation at the transmitter was the so-called "direct" method, most variations of which use a frequency modulated LC oscillator whose centre frequency is then accurately stabilized in relation to the fixed frequency of a crystal controlled oscillator. The choice of system and detailed design of all f.m. modulators is considerably influenced by the practical requirement that the modulator must be capable of being set to operate at any assigned frequency within the band without making radical changes to its circuit. A good example is the modulator developed by Standard Telephones and Cables for use on Band 2 and which will be used at certain of the BBC's new stations. A block schematic of this modulator is shown in Fig. 5. It consists of a push-pull LC oscillator, tuneable over the range of 4.86 and 5.55Mc/s, which is frequency modulated by a push-pull reactance valve modulator driven from the audio input amplifier. The frequency modulated output of the oscillator is passed to two buffer amplifiers. One amplifier drives a standard frequency multiplier which multiplies by 18 to raise the frequency to the range 87.5 to 100Mc/s, while the other amplifier provides the output for the centre frequency stabilizing chain. For this chain the frequency of the LC oscillator is first brought to a standard value of 7.5Mc/s by mixing the output of the amplifier with that of a crystal oscillator

operating at a frequency in the range of 12.36 to 13.05Mc/s, the selected frequency depending on the assigned frequency of the transmitter concerned. This frequency is then divided by 500 in a standard frequency divider to give an output frequency of 15kc/s. This output is passed to a balanced modulator and compared with the fixed frequency of a 15kc/s reference crystal oscillator of high stability.

The balanced modulator gives a two-phase output at any difference frequency which exists due to drift of the modulated LC oscillator, which operates a small, very slow speed synchronous motor mechanically coupled through reduction gears to a trimming capacitor in the anode tuned circuit of the LC oscillator. This capacitor is thus moved by an appropriate amount in the right sense to correct the centre frequency of the modulated LC oscillator. The system is capable of maintaining the centre frequency correct to within 500c/s in 100Mc/s. Failure of the automatic control circuit does not pull the LC oscillator off frequency and the setting-up procedure is simple.

Automatic frequency control circuits are a complication, and systems have been devised to eliminate them. One of these is the FMQ (Frequency Modulated Quartz) system<sup>10</sup>, developed by Marconi's Wireless Telegraph Company Ltd. originally for Wrotham and which will also be used at some of the new stations. Previous attempts to modulate the frequency of a conventional crystal oscillator in a sufficiently linear manner had not been particularly successful, partly because a normal type of crystal does not

Fig. 5. Block schematic of "direct f.m." modulator



\* Formerly BBC Engineering Division, now Chief Engineer, I.T.A.

behave as a simple resonator but has a number of secondary resonances which may occur within the f.m. spectrum. In the FMQ system the introduction of a quarter-wave coupling network between the reactance valve modulator and the crystal enables linear frequency modulation to be obtained, and a type of crystal has been developed in which secondary resonances have been practically eliminated. When the equivalent network of this crystal is put into a form suitable for modulation, by transforming its series resonant impedance into a parallel resonant impedance, the resonant circuit so obtained has a high Q value and a stability far greater (about 100 times) than that of an equivalent LC circuit. The frequency versus susceptance characteristic of the circuit is very linear, and provided a balanced modulator of good linearity, stable mutual conductance and low noise level is used, the system is capable of providing deviations in excess of 125kc/s with a centre frequency stability of

not needed. A practical difficulty, however, is that the crystal oscillator must have a comparatively low frequency (about 100kc/s). This is because the system uses steep-fronted waves whose rise times would otherwise be very short. The equivalent deviation at low modulating frequencies is thus rather small, and large frequency multiplication is required to obtain 75kc/s deviation at the final centre frequency. Frequency mixers are needed to give additional deviation for the lower audio frequencies and, because of the greater frequency multiplication involved, difficulty might be experienced in keeping f.m. noise sufficiently low. Most of the modulators of this type have so far been designed for narrow-band f.m. sound for u.h.f. television service, where very high fidelity and low noise are perhaps not so important.

#### F.M. TRANSMITTER PERFORMANCE

The Wrotham installation permitted an extension of the field trials relating to propagation, comparison of modu-

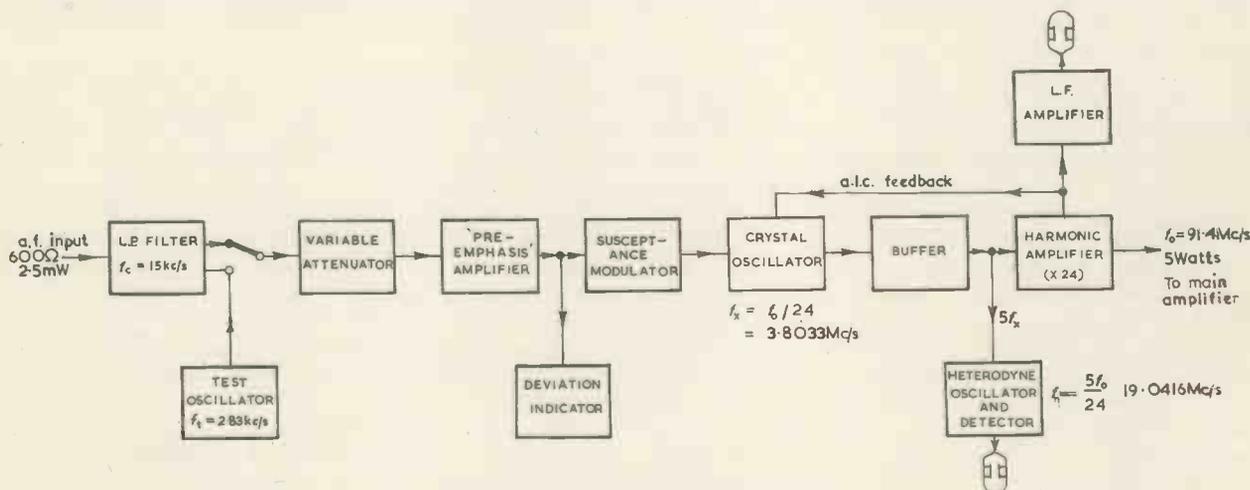


Fig. 6. Block schematic "FMQ" modulator

better than  $\pm 10$  parts in  $10^6$ , together with the low noise and distortion appropriate for a Band 2 high fidelity sound broadcasting service. A block schematic of this system is given in Fig. 6.

Another practicable system is one incorporating a method of phase modulation whereby the timing of a pulse is varied in accordance with the instantaneous amplitude of the audio frequency modulating signal, the phase modulated pulse then being converted into an equivalent frequency modulated output. This is the principle of the so-called "Serrasoid" modulator<sup>11</sup>, in which a linear sawtooth is generated, having a repetition rate controlled by a stable crystal oscillator. In the unmodulated condition the sawtooth is clipped halfway up, and the pulse is produced at the instant when the rising front of the sawtooth changes to the horizontal. The a.f. modulating signal is then arranged to control the point on the sawtooth at which clipping occurs, and the timing of the generated pulse thus follows the instantaneous value of the a.f. signal. The maximum phase deviation of the pulse is limited by the length of the linear part of the sawtooth, which can in practice occupy about  $300^\circ$ .

In common with all phase modulated systems, this method has the advantage that the centre frequency stability is that of the crystal oscillator, which can be made very high, so that automatic frequency control is

lating systems and listening tests, and also provided the background for the development of prototype transmitting equipment together with a v.h.f. high gain aerial system and their trial under service conditions. Both the f.m. and a.m. transmitters for Wrotham were designed to provide the highest performance commercially attainable, so that the full potentialities of a v.h.f. service on either f.m. or a.m. could be determined. The performance of the original f.m. transmitter is given in Table 2 and can be regarded as reasonably typical of the future transmitters of the BBC's v.h.f. service.

#### F.M. TRANSMITTER MONITORING

A high-grade radio monitoring receiver with suitable instrumentation for checking the centre frequency shift during modulation, the frequency deviation and the audio frequency quality of the radiated signal is an essential item of station equipment. It is not convenient to extract the actual centre frequency component of an f.m. signal for monitoring purposes, because this component varies in amplitude with the index of modulation and for certain values of modulation index is zero. In practice, the "mean frequency" of the signal is used, this being defined as the continuous average of the instantaneous frequency taken over a time which is long compared with the lowest modulation frequency. In an undistorted f.m. wave this is

the same as the frequency of the centre component.

The receiver is basically a limiter followed by a phase discriminator, the output current of which corresponds, in both sense and amplitude, to the instantaneous frequency deviation of the applied f.m. signal from its assigned centre frequency, the current being zero at the assigned value. The zero current point may be adjusted by reference to the frequency of a calibrating crystal oscillator of high stability. Indication of the mean frequency shift is obtained by averaging the discriminator output current over a time which is long compared with the lowest modulating frequency. Any residual direct current component which exists is a measure of the shift of the mean frequency of the signal from its proper value, and may be used to operate a centre-zero d.c. meter calibrated to read in kc/s shift.

Indication of the frequency deviation is derived from the alternating current component of the discriminator output current which has the same form as the modulating signal. Its peak value is thus proportional to the peak frequency deviation and may be used to operate a peak-reading valve-voltmeter via a linear amplifier of stable gain. The audio frequency output signal is also derived from the a.c. component of the discriminator output and is raised to a convenient level by an amplifier which incorporates the appropriate de-emphasis network. Deviation indication circuits have reasonable stability, but it is desirable that the monitor should include equipment whereby an absolute check can be made, particularly to ensure that the maximum permissible deviation is not being exceeded. The Crosby method<sup>12</sup> is a convenient one and involves apparatus for the separation and amplitude indication of the actual centre frequency component  $F$

of the f.m. signal radiated by the transmitter and the application of a single a.f. modulating tone of known frequency  $f$  to the transmitter modulator. The well-known series of Bessel zero amplitude levels of the centre frequency component are sequentially indicated as the amplitude of the modulating tone is progressively increased and as each zero corresponds to a known deviation ratio  $B = \Delta F/f$  the absolute value of the frequency deviation is given by the product of the deviation ratio and the frequency of the modulating tone. Zero indications up to about the 6<sup>th</sup> are usually sufficient in practice. A really comprehensive monitor will also include means for measuring the noise and distortion of the f.m. signal, and such a monitor is in fact being constructed to BBC specification.

#### V.H.F. Aerial-Transmission Line Systems

The aerial developed by the BBC Research Laboratory, originally for Wrotham<sup>13</sup> and which was constructed by Marconi's Wireless Telegraph Company Ltd., is an excellent example of aerial engineering for operation at frequencies in Band 2. It is a high gain omnidirectional radiator of the slotted cylinder type giving horizontal polarization. A photograph of the aerial is given in Fig. 7. The type of construction was chosen mainly because it was a structural requirement that the aerial, when used at other stations, should also act as a support for a Band 1 television aerial<sup>14</sup>. Where this does not apply, and particularly at stations where the total transmitter power is lower, a simpler form of aerial may be used, such as stacked rings of dipoles or the batwing type of radiator in common use in America in 6 stack or even 12 stack form. The Wrotham type aerial<sup>15</sup> is rated to accept the simultaneous trans-

TABLE 2  
Measured Performance of 25kW f.m. Transmitter

FREQUENCY RANGE .. .. .	85 to 100Mc/s
CENTRE FREQUENCY STABILITY .. .. .	Better than $\pm 20$ parts in $10^6$ for a temperature rise of $20^\circ \text{C}$ and for supply voltage fluctuation of up to 6 per cent
CHANGE OF CENTRE FREQUENCY WITH MODULATION .. .. .	Less than $\pm 10$ parts in $10^6$ for 100kc/s deviation
POWER OUTPUT .. .. .	25kW
OUTPUT IMPEDANCE .. .. .	$51.5\Omega$ unbalanced
AUDIO INPUT LEVEL .. .. .	10dBm into $600\Omega$ for 75kc/s deviation
FREQUENCY DEVIATION EQUIVALENT TO 100 PER CENT MODULATION .. .. .	75kc/s; maximum deviation 100kc/s
PRE-EMPHASIS .. .. .	0, 50, 75 and $100\mu\text{sec}$ pre-emphasis networks are provided and can be selected at will
A.F. RESPONSE .. .. .	Within $\pm 0.5\text{dB}$ relative to 400c/s over frequency range 30c/s to 8kc/s, falling to approximately 3dB down at 15kc/s and 10dB down at 20kc/s
A.F. DISTORTION .. .. .	At 75kc/s deviation, total harmonic distortion is less than 1.5 per cent for audio frequencies from 30 to 60c/s and less than 1 per cent for audio frequencies from 60c/s to 15kc/s
F.M. NOISE LEVEL .. .. .	When measured through a weighted network, the noise level is better than $-70\text{dB}$ relative to 75kc/s deviation at 400c/s audio frequency
A.M. NOISE LEVEL .. .. .	Better than $-50\text{dB}$ relative to 100 per cent modulation at 400c/s
A.M. DUE TO F.M. DEVIATION OF CARRIER UP TO 100KC/S .. .. .	Does not exceed 1 per cent, plus an amount corresponding to the permitted reflexion coefficient of the aerial; total 6 per cent approximately
POWER SUPPLY REQUIREMENTS .. .. .	415 volts 3 phase 50c/s 90 kVA; 0.9 power factor $\epsilon = 31\%$

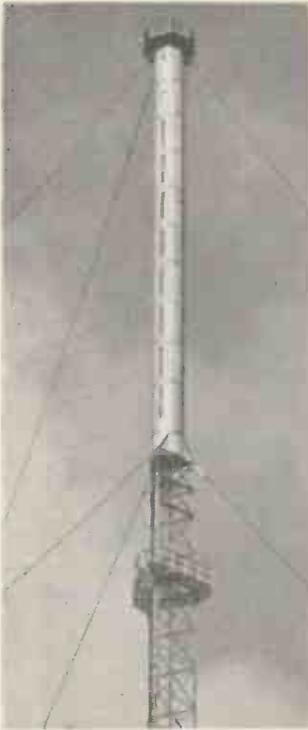


Fig. 7. The high gain v.h.f. slot aerial at Wrotham

missions from up to three 25kW f.m. transmissions in the Band 87.5 to 95Mc/s. The array of 32 co-phased vertical slots is arranged in 8 stacks of 4 spaced equally round the surface of a vertical steel cylinder 6ft 6in in diameter and 110ft high, which is supported by a triangular section lattice steel mast 360ft high. Each slot is approximately 8ft long, corresponding to  $0.75\lambda$  at mid-band, and 12in wide; the vertical spacing between slots is 10ft 6in, corresponding to  $\lambda$  at midband. The slots are fed at equal power and in identical phase by the successive bifurcation of the single  $51.5\Omega$  5in coaxial main transmission line<sup>14,16</sup> which enables the aerial to be fairly easily converted into the "split" form if required for the system of transmitter parallel operation mentioned later in this article. The standing wave

ratio of the line alone is about 1.03 over the band and the total attenuation approximately 0.65dB for the 730ft run. The overall standing wave ratio for the complete aerial system is about 1.1, but two sets of slot adjustments are necessary to maintain it over the band. One set of adjustments covers the range 87.5 to 93Mc/s, and a second set 89 to 95Mc/s. The aerial has a theoretical mean gain of about 9.5dB over a half-wave dipole, the deviation from the mean being about  $\pm 0.5$ dB. The gain is slightly reduced by small inequalities in the amplitude and phases of the slot-driving voltages, and when the transmission line and combining filter losses are subtracted the effective mean gain of the system is about 8dB. The calculated vertical radiation pattern is shown in Fig. 8.

At Wrotham the frequency spacing between the f.m. and a.m. transmissions was 2.4Mc/s, and to permit them to be fed into the common aerial a frequency selective "notch" type combining filter constructed from coaxial copper tube transmission line was used. This filter, which has been fully described in another article<sup>18</sup>, gives an insertion loss between transmitters of more than 60dB at the "stop" frequencies, while for the "pass" frequencies the loss is about 0.25dB. The reflexion coefficient is less than 2 per cent. At future stations where three f.m. transmitters will feed a common aerial, this type of filter is probably not the most convenient, a more practical approach being the use of two filters of the Maxwell Bridge type<sup>14</sup>, which are physically much more compact. These filters would be con-

nected so that one combines the outputs of two transmitters and the other the output of the first filter with that of the third transmitter (see Fig. 9).

At Wrotham no attempt was made to "fill in" the several minima which occur in the vertical radiation pattern to avoid points of low field strength close in to the aerial. It might, however, be desirable to do this (by changing by  $\pm 90^\circ$  the phase of the currents in one of the stacks of slots near the centre of the aerial) in future cases where the aerial may be located within a closely built-up area. Difficulties can also arise because of the narrowness of the main beam of the vertical radiation pattern of large aperture v.h.f. aerials with conventional symmetrical radiation patterns. The presence of the sharp zero in such patterns will result in a region of low field strength perhaps a mile or so from the transmitter and, in consequence, the relative strength of time-delayed signals reflected from objects lying in adjacent regions of much higher field strength could be large and cause a distorted combined signal at receiving aerials situated in the region of low field.

In some cases the aerial may not be required to radiate maximum field exactly along the horizontal, but arranged so that a greater proportion of the available power is directed below the optical horizon than beyond it. This can be done by "beam tilting," i.e. deflecting the axis of the main beam slightly downwards by progressively advancing the phase of the currents in the stacks of radiating elements with increasing height. This, however, is only a partial solution, and a more elegant technique is one whereby the amplitude and phase distribution of the radiator currents is based on the Fourier transform of an idealized radiation pattern<sup>17</sup>. The ratio of power radiated below the horizontal to that above it can then be made as high as 3:1. A possible idealized pattern is an asymmetric one having zero radiation above the horizon, constant radiation between  $0^\circ$  and twice the tilt angle, the radiation thereafter decreasing as  $\text{cosec } \theta$  where  $\theta$  is the angle of depression. Mechanical tilting of the array is sometimes used, but this tilts the beam up in one direction and down in the opposite direction. The angle of tilt required is usually very small (less than  $1^\circ$ ) and so can be upset by high winds. Combined mechanical and electrical tilting might, however, prove useful in cases where long range is required in some directions and short in others. These techniques are perhaps not justified for Band 2 aerials with apertures less than about 8, but for v.h.f. Band 3 and u.h.f. Bands 4 and 5, where aerials having apertures of up to 16 and 24 respectively and

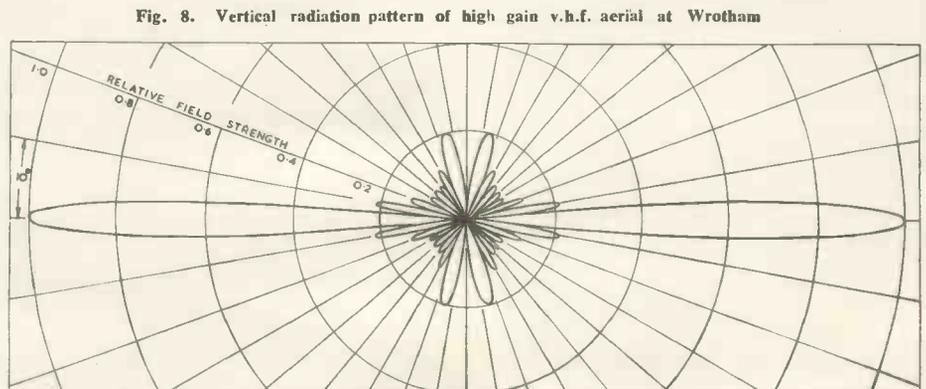


Fig. 8. Vertical radiation pattern of high gain v.h.f. aerial at Wrotham

hence much sharper radiation patterns are practicable, beam tilting can be of value, particularly if the aerial is very high, say 1500ft or more above the surrounding countryside.

With large aperture aerials for f.m., the design should permit a low reflexion coefficient to be realized at the operating frequencies concerned, and it is considered that 5 per cent as obtained with the Wrotham aerial should be aimed at in order to minimize the distortion which reflected time-delayed signals introduce into the radiated f.m. signal. A reflected signal, resulting from a mismatch between the aerial and its transmission line, will be partially reflected at the transmitter and radiated from the aerial delayed in time from the main f.m. signal. The resultant radiated signal then contains both amplitude

will transmit three separate programmes (Home, Light and Third) simultaneously radiated from a common aerial system. The transmitters for each transmission actually consist of a pair of identical transmitters effectively operated in parallel. Each pair of transmitters is driven from a common f.m. modulator via phasing units, and the paralleling of their outputs is by means of the "split" aerial system<sup>14</sup>. With this arrangement there are two completely independent transmission chains, each consisting of three transmitters, two combining filters and a single transmission line feeding one half of the aerial. The two halves of the aerial are designed so that the mutual coupling between them is very low, while the centre frequencies of each pair of transmitters is automatically maintained in phase by means of a phase monitor

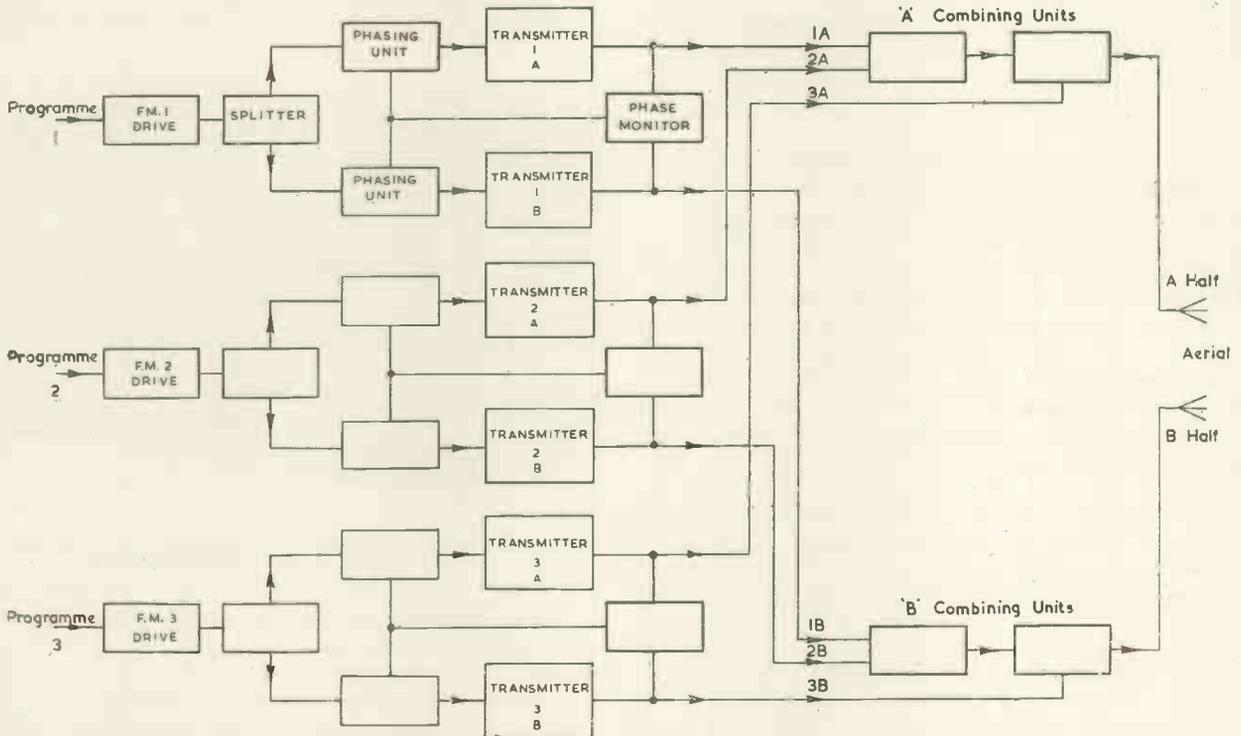


Fig. 9. Block schematic of complete f.m. transmitting station for 3 programme working and parallel operation of transmitters

modulated components and harmonic distortion terms<sup>18, 19</sup>. The magnitudes of these components vary cyclically between zero and a maximum with changes of phase-angle between the main and reflected signals and are, therefore, critically dependent on differences in the reflected and main signal-path lengths and on the magnitude and angle of the respective reflexion coefficients at the aerial and transmitter. The depth of modulation produced by the a.m. components cannot, however, exceed the reflexion coefficient of the aerial and in any case can be assumed to be removed by limiting in the receiver. In practice the harmonic distortion terms rarely reach their maximum amplitudes, so that the distortion introduced is small, although not necessarily negligible, in comparison with the distortion introduced by non-linearities in the phase/frequency characteristics of the transmitter and receiver circuits.

#### Parallel Operation of f.m. Transmitters

In the majority of cases the BBC's new f.m. stations

connected between the outputs of the pair. The advantages of this arrangement are that one transmission chain acts as a complete spare to the other, and the use of transmitter paralleling units is avoided.

A point of importance in regard to such parallel operation is that, taking any pair of transmitters, some difference both in amplitude and phase characteristics of the two transmission chains is inevitable. Small amplitude differences are not important, but any differential phase shift over the deviation frequency range of  $\pm 75\text{kc/s}$  occurring between the two chains may be so. If the differential phase shift is linear, some amplitude modulation will occur in the radiated signal. If the differential phase shift is non-linear, there will also be a phase difference between the distortion components in the two chains. Investigations have shown that the latter point is not serious, and the main concern need only be with the effects of the linear differential phase shift.

By far the greatest differential phase shift is due to the transmitters themselves, because the number of tuned

circuits they incorporate is considerable (the 25kW Wrotham transmitters have 15 tuned circuits) and under practical conditions their tuning will not be identical. For example, the total phase shift through one Wrotham transmitter was found to be about  $\pm 50^\circ$  for a deviation frequency range of  $\pm 75\text{kc/s}$ , while the differential phase shift occurring between the outputs of a pair of these transmitters when driven from a common f.m. modulator was about  $\pm 15^\circ$  for the same deviation and was virtually linear. The new transmitters, however, will have fewer power amplifiers in cascade and will be specifically designed for parallel operation, so that the differential phase shift between a pair is expected to be reduced to about  $\pm 10^\circ$ . The combining units are expected to contribute an additional  $\pm 5^\circ$ , while the differential phase shift occurring between each transmission line and half aerial should be negligible. It is therefore concluded that the total differential phase shift, over the deviation frequency of  $\pm 75\text{kc/s}$ , through the system is unlikely to exceed  $\pm 15^\circ$ , which corresponds to less than 2 per cent amplitude modulation due to this cause.

So far as the combining units, transmission line and half aerial are concerned it is necessary to consider not only differential phase shift, but also any fixed phase difference they may introduce between the two centre frequencies (because the phase of centre frequencies is monitored at the transmitter outputs). The combining units and transmission lines are not likely to introduce a phase difference exceeding  $\pm 10^\circ$  in the worst case, i.e. when the temperature difference between them is a maximum, while for the aerial the phase difference of the centre frequencies in the two halves is unlikely to exceed  $\pm 10^\circ$  in view of the accurate matching conditions imposed. Thus the total phase difference between the centre frequencies in the aerial should not exceed  $20^\circ$ , which is too small to have any significant effect on the polar diagram. Fig. 9 shows a block schematic of a complete f.m. transmitting station arrangement based on the scheme of operation discussed above.

#### Unattended Operation of f.m. Transmitters

The use of pairs of transmitters effectively operated in parallel for each transmission ensures a high degree of immunity from breakdown, which is all the more essential in the case of the BBC's new f.m. stations because it is intended that the transmitters shall normally be arranged for remote-controlled unattended operation. The unattended operation of transmitters introduces several new problems in transmitter engineering. In particular it means that the conventional sequential procedures involved in switching a transmitter on and off and the system of fault protection must be fully automatic. It also implies the use of a comprehensive overall programme monitoring system, automatic in action, whose task is to compare the incoming programme signal with the f.m. output signal, and operate appropriate alarms and trips if any serious discrepancy occurs. Space precludes a description of the programme monitoring system, but one of the forms of transmitter automatic control which shows good promise is mentioned below. The principle of this system is the provision of a special moving-coil automatic voltage regulator which controls the voltage of the input a.c. power supply to the complete transmitter equipment. On switching on the transmitter, the regulator energizes all supplies to the power amplifiers and

auxiliaries gradually and simultaneously, the supply voltage rising from approximately zero up to the normal working value in approximately one minute; it then automatically stabilizes the voltage at the correct value. In the event of an overload fault or incorrect operation in any of the transmitter circuits, the appropriate protective relay operates a fast release mechanism which causes the regulator to reduce the input supply voltage to approximately zero in about three seconds. An automatic reclosing device then comes into operation which restores the voltage to normal. If the fault is still present, the regulator again releases and remains at its zero voltage position. Normal switching off operates the fast release mechanism without bringing in the reclosing device. This system has the merit of simplicity because in addition to removing the need for sequential interlocking relays, contactors and motor starters, all a.c.

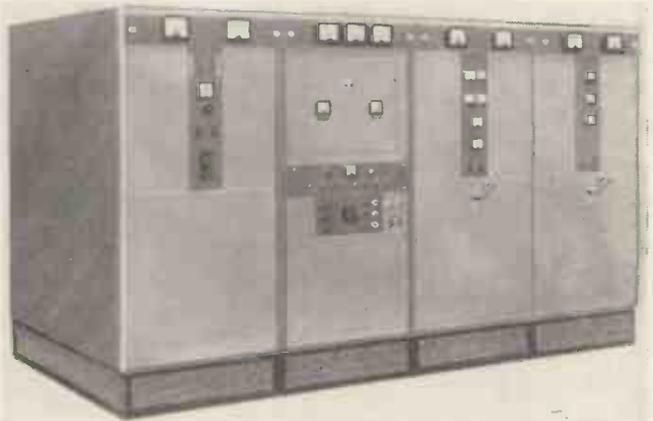


Fig. 10. A modern v.h.f. 10kW f.m. Broadcast transmitter designed for remote-controlled unattended parallel operation

sub-circuit distribution contactors and fuses are also eliminated since these circuits may be connected directly to the control regulator output terminals via current transformers, in the secondaries of which are standard overload protective relays which operate the release mechanism of the regulator. A single on-off switch controls the whole transmitter. A photograph of a modern 10kW f.m. transmitter designed on this basis is shown in Fig. 10.

#### Conclusions

The BBC has already received government approval for the construction of the first 10 high power f.m. stations. It is hoped that most of these will be in operation by the end of 1956 but, in the authors' opinion, the fulfilment of this programme will inevitably be determined by the ability of the transmission side of the radio industry to deal with it, bearing in mind the demands also being made on this side of the industry by the introduction of independent television on Band 3. Much of the equipment required for the Band 2 sound and Band 3 television services is similar, namely, v.h.f. transmitters of 10 to 20kW output, high gain aerial systems, low loss transmission lines and high masts.

Very high frequency sound broadcasting, in this country, is nevertheless about to emerge with great vigour from the experimental stage and take practical shape. Its adoption

should place sound broadcasting on a far better footing than it is today because it provides a means of escape from the worsening congestion in the low and medium frequency bands and at the same time confers the benefits of a much higher quality service. The v.h.f. scheme envisaged by the BBC is being planned to this end, and priority will be given to the areas which are in greatest need of an improved service. Eventually a three programme v.h.f. service to approximately 98 per cent of the population should be achieved. The decision that the system of modulation should be f.m. is based on the fullest possible appraisal by the radio industry in general of all the relevant theoretical, practical and economic considerations. The high standard of technology required for the development of f.m. transmitting plant and high grade receivers and adaptors for existing m.f. receivers is available in this country, and much of the plant for the new stations is already in works production. V.H.F. broadcasting, coupled with f.m., indeed seems to be "Sound Radio's Second Chance."

#### Acknowledgments

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## A Very High Speed Camera

A SEQUENTIAL image convertor (type SP.2), manufactured by Winston Electronics Ltd., has recently been flown to America for use at the Aberdeen Proving Grounds, one of America's major weapon-testing centres. The camera formed part of the freight on the first flight of the new transatlantic freight service inaugurated by Airwork Ltd.

The camera employs a Mullard image convertor tube (type ME1201AA), which is basically a cathode-ray tube fitted with a photo-cathode instead of a heated filament.

An optical image is focused by an optical system on to the photo-cathode. The light distribution in the optical image is converted into an exactly similar distribution of emitted electrons. The electrons are focused by a magnetic field and accelerated by the action of the grid and anode. On striking the luminescent screen the energy of the electron is converted into light and an exact reproduction of the original is obtained.

In order to use the image convertor as a high speed cine-camera the tube is normally cut-off by the action of the grid and it may be turned "on" for a length of time as short as 0.01  $\mu$ sec by applying suitable pulses to the grid.

Further the beam of electrons may be deflected by the action of an external magnetic field. A small square picture can thus be deflected into several distinct positions while the grid is at the same time pulsed switching the i.c.t. on.

The SP.2 image convertor does this giving two rows of three pictures each 20  $\times$  20mm size. Seven shutter speeds are available; the rate of scanning is locked to the shutter action in such a manner that the interval between the shutter pulses is always five times the length of the shutter pulse.

The following table gives the available shutter speeds with the respective frame rate and maximum repetition frequency (of a complete sequence).

SHUTTER SPEED ( $\mu$ sec.)	FRAME RATE (c/s)	REPETITION FREQUENCY (c/s)
0.1	2 $\times$ 106	500
0.2	1 $\times$ 106	500
0.5	4 $\times$ 105	500
1	2 $\times$ 105	500
2	1 $\times$ 105	250
5	4 $\times$ 104	100
10	2 $\times$ 104	50

The action of a complete cycle is initiated by a trigger pulse which must be roughly +100V high. The trigger action can be delayed by 0.1 to 1  $\mu$ sec or longer if required.

The optical image is focused on the photo-cathode by means of a Gregorian telescope system.

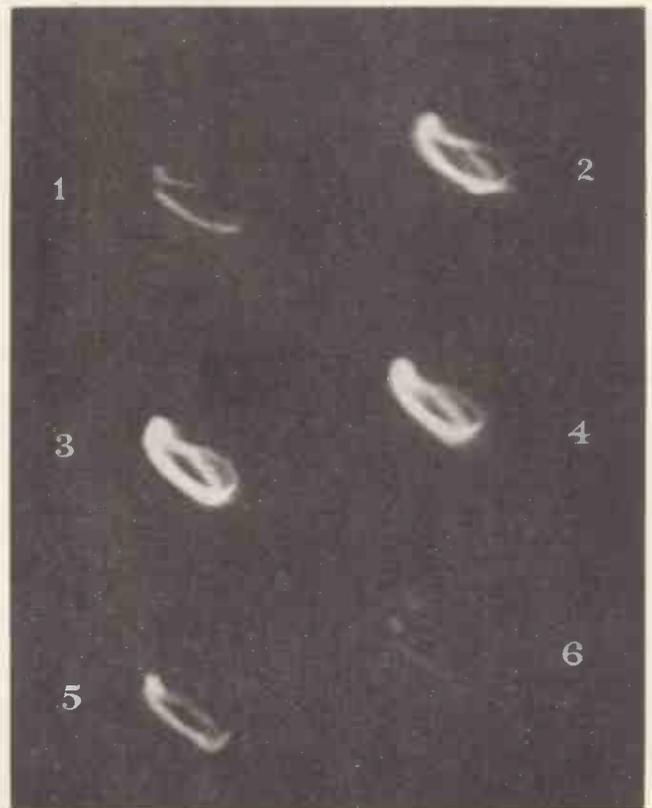
The resulting picture on the luminescent screen can be photographed on a 4in  $\times$  5in film.

Due to the very small amount of light being allowed to get through to the photographic plate the objects must either be intensely self-luminous such as spark discharges, explosions, etc., or they must be illuminated by a highly luminous source such as gas-filled flash tubes e.g. LSD.2. These light sources must of course be synchronized with the image convertor equipment.

At speeds larger than 1  $\mu$ sec the resolution of the complete system is approx 200 lines/cm and at higher speeds 60 lines/cm (these figures do not take into account the grain of the film). The maximum shake of the 2cm square picture is 0.03cm.

The complete equipment consists of two steel racks 6ft high on a 2  $\times$  2ft base.

Development of light in LSD2 flash tube. 0.1  $\mu$ sec. exposures at 0.5  $\mu$ sec intervals



# Some Recent Developments in American Electronic Musical Instruments

By Alan Douglas

*Since the inception of commercial musical instruments, the basic forms and types of generating systems have remained unaltered. No new method of tone generation has appeared. Development has therefore taken place to exploit already existing resources still further, with the principal objects of providing better tonal synthesis and increased flexibility of control. This article reviews the latest American practice in these matters.*

WE will take the Hammond organ first, as it is better known than any other instrument and has strongly marked characteristics of its own. It is not proposed to describe the complete instrument<sup>1</sup> or its theory<sup>2</sup> again, but to introduce the latest modifications.

The weakest element of the organ has always been the pedal section. Now, a complete series of harmonics is made available by the multiple pedal key switches, Fig. 1. The tone sources for these are derived from the main manual generators. Thus there is a wide range of synthetic sound to balance corresponding manual combinations.

A more interesting addition on the RT models is the

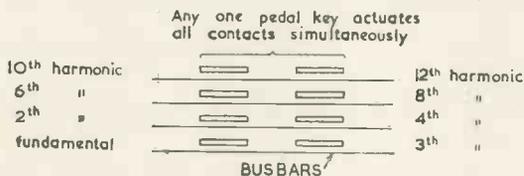


Fig. 1. Hammond pedal key contacts

solo pedal generator. This not only provides tones down to 32ft pitch, but enables a pedal solo to be taken on the right foot while the normal bass is maintained by the left foot—a situation capable of great exploitation. Fig. 2 is the complete circuit of this apparatus, which in essence is a continuously-tuned LC oscillator. Thus only one note at a time can be played. The normal use of the pedals is not invalidated, since the switching arrangements ensure that only the upper note will speak on the solo circuit, which can be disconnected at will. The interesting feature is that tone generation is at 1ft pitch, which is then divided down to 2, 4, 8, 16 and 32ft pitch. An enormous range of tonal combinations is thus available to the player, some of which can be preset on toe pistons. The rate of attack is not adjustable, but a good general compromise has been effected. This unit has its own power supply and expression pedal. No special loudspeakers have been evolved for this circuit, but the level of the 32ft tones is low and so they are unsuitable for use alone.

On certain models a main generator having complex tone wheels can be obtained. These supply odd harmonics and their judicious use does much to offset the lack of certain intervals which, due to tempered scale interference, are not suitable for the additive method of tonal synthesis, and which were not available on previous models.

The reverberation device has been improved in design and response, and is shown diagrammatically in Fig. 3. It will be recalled<sup>1</sup> that part of the final signal is passed directly to the main amplifier and part through this unit to the same amplifier. There is thus a delay which intro-

duces very complex and insoluble phase relationships but, since the frequencies comprising the signal are subjected to repeated reflexions through the springs, results in an apparent sustaining of the general sound spectrum. The initial delay is about 66msec, the second spring adds another 133msec, and so on; but there are innumerable secondary reflexions and these are in part adjustable by controlling the quantity of oil in the tubes, effectively

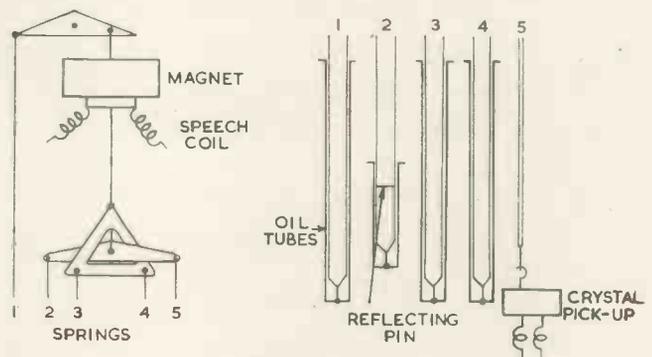


Fig. 3. Reverberation device

lengthening or shortening the springs, and so varying the propagation time.

In effect, when well set up, the reverberation device gives continuously overlapping phase shifts having five major time/frequency components and a very large number of smaller phase shifts. The effect is theoretically good but aesthetically poor in small rooms. True reverberation is a function of all the active areas of the room, whereas in this application it is confined to the radiating surface of the loudspeaker cones.

A number of minor changes have been made in the vibrato line scanner<sup>1</sup>, expression controls and the generator filter circuits, but these do not substantially alter the general tonal structure of the sole example of an additive music generator having individual tone sources for each note. Incidentally, there are 14 different models of the Hammond organ in current production, besides the Spinet<sup>3</sup>, Chord<sup>4</sup> and Solovox<sup>5</sup> instruments.

There has been a marked tendency in the United States to neglect the electro-mechanical tone generator and to concentrate on the valve instrument. In view of the ready availability of cheap valves and the profusion of well-engineered communication elements, this is not surprising. At the same time, it is by no means easy to make a successful rotating generator, and the manufacturing and assembly costs are high. A new organ to be introduced by the Baldwin company uses rotating photo-electric



generators and a few models are undergoing field tests; as it is not yet on the market, no details can be disclosed. This also has a reverberation system relying on springs, as an optional fitment.

The only remaining example of other electro-mechanical instruments is the vibrating reed organ made by Wurlitzer. It is largely due to their long experience with pipe organs and the suitability of their plant that they continue to produce this instrument.

Earlier models<sup>6</sup> were based on a development of the Hoschke patents<sup>7</sup>, and used electrically-operated valves as in a pipe organ to admit wind to the reeds as required. To meet the competition of instruments with a much quicker rate of speech, the pallet valves have been abandoned in the current series and the reeds vibrate continuously. The signal is then keyed through suitable transient filters.

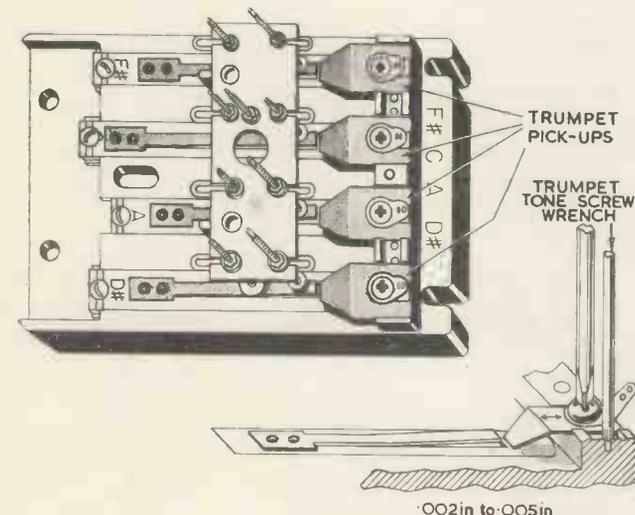


Fig. 4. Reed with multiple pick up

This time we will select a small model as exemplifying the resources of the method to the full. The type 44 is a domestic instrument using the extension principle. Briefly, this means that all pitches of any one tone colour are extracted from a continuous row of sound sources instead of having overlapping rows of different pitches as in many pipe organs. Thus a 4ft pitch is taken from the 8ft pitch reeds but one octave higher, and so on. The loss in combination is not so great as might be expected, being roughly 8 to 10 per cent. Most small pipe organs use this principle now, which results in marked economy in cost and space. The method is fully explained in the literature<sup>8</sup>. One thing extension cannot give in a physical instrument is a change of tone quality, and this is an outstanding advantage of electrical systems; from the same sound source, various contrasting effects are possible.

The type 44 Wurlitzer has only 73 reeds for the entire instrument, yet from these, flutes, strings and reed tone from 8ft to 1½ft pitch and covering 61 notes are compounded.

Fig. 4 shows a reed with its multiple pick-ups. Since such "free" reeds are in fact bars clamped at one end, many overtones are developed along the blade. By careful

disposition of the capacitive pick-ups, and by changes in their surface areas, different frequency bands can be emphasized. Further, by polarizing these pick-ups at different levels, the relative strength of the tones may be greatly varied.

The many pick-ups, with their transient suppression circuits (which are printed on phenolic cards), are collected at the multiple contacts under each playing key (Fig. 5). Here, by the busbar system shown, any number of these tone sources may be combined to produce a composite signal. So comparatively few reeds require only a modest supply of wind to maintain them in vibration, as distinct

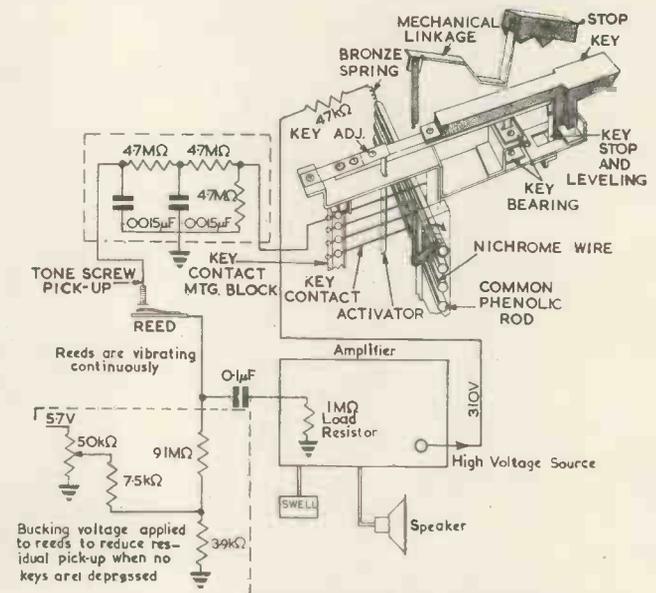


Fig. 5. Typical keying circuit to the tone screw pick-ups. M-44 Wurlitzer organ.

REED CHEST

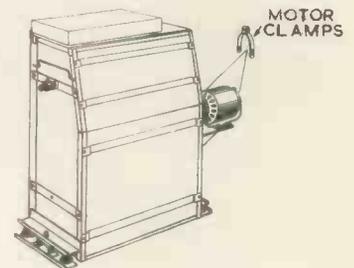
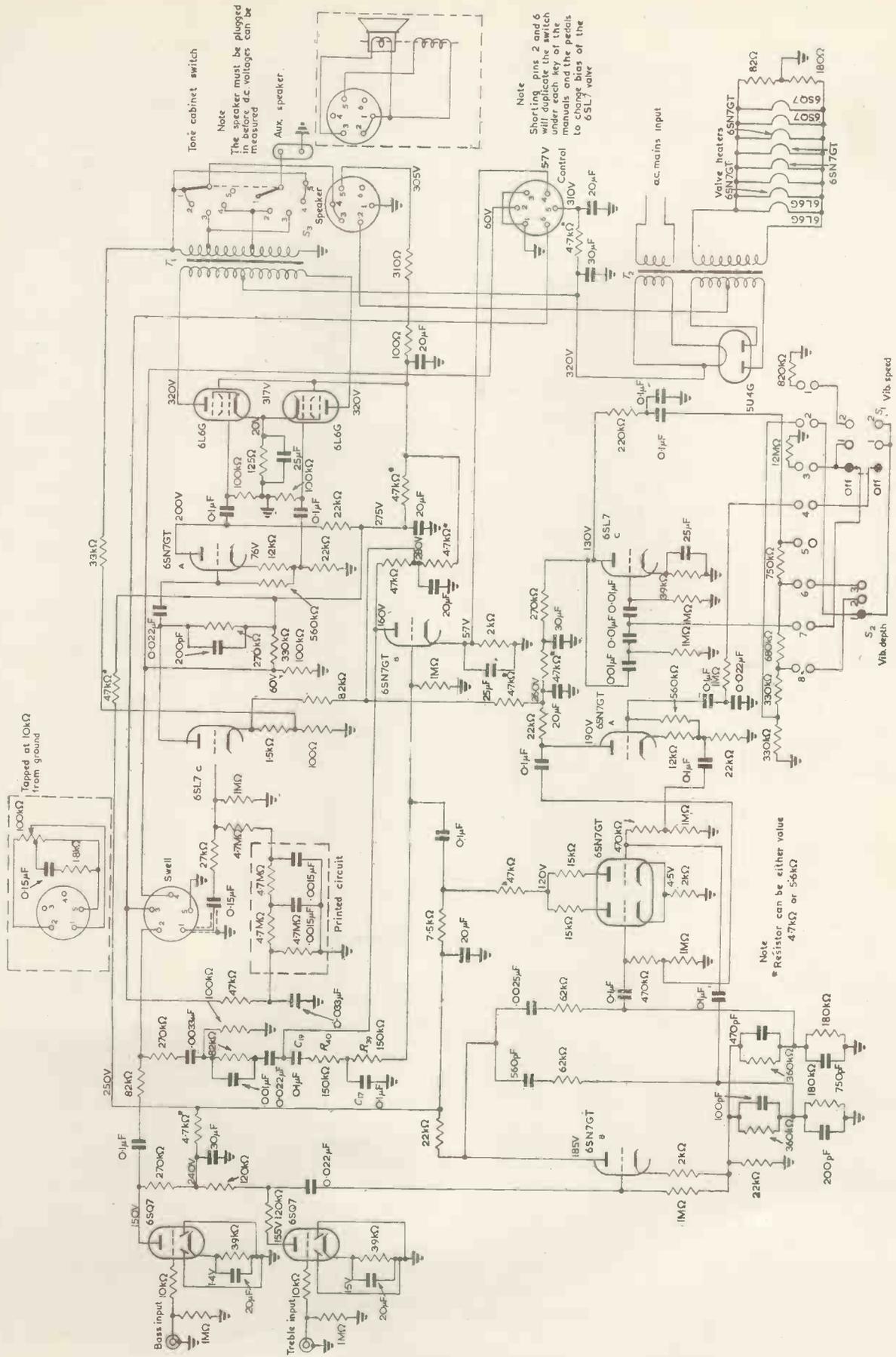


Fig. 6. Arrangement of reed chest and fan

from starting them; a motor of 1/12 h.p. drives an impeller submerged in the reed chest (Fig. 6). This contains all the reeds mounted on bakelite blocks as in Fig. 4. Each pick-up is adjustable by a special spanner.

The busbars are rotated to make contact with the key wires by the stop tablets, by a simple mechanical coupling. No "tone-forming" is carried out in the amplifier, which is divided into treble and bass input channels. The main purpose of this is to apply vibrato to the treble only. This is accomplished by the rather elaborate circuit of Fig. 7, which shows the complete amplifier. Note that the vibrato frequency is removed by the filter  $R_{39} C_{17}$ , so that each 7th or 8th signal is shifted in phase during the excursion of the vibrato oscillation. This produces a



Tone cabinet switch

Note  
The speaker must be plugged in before dc voltages can be measured

Aux. speaker.

Note  
Shorting pins 2 and 6 will duplicate the switch under each key of the manuals and the pedals to change bias of the 65L7 valve

Note  
Resistor can be either value 47kΩ or 56kΩ

Fig. 7. Wurlitzer amplifier and vibrato circuit

type of undulation which is particularly pleasing to the ear.

The mechanical excellence of the construction is typical of the workmanship for which Wurlitzer pipe organs have always been noted.

Perhaps the Conn organ is more generally interesting, since it depends on valve oscillators. In the original instrument<sup>9</sup>, 132 valves were employed. The generating circuits<sup>2</sup> are substantially unaltered, but the writer was

they act as couplers to make available the different pitches. The valuable "unison off" is thus obtainable. It will be noticed that the tones formed by Fig. 8 are appropriate to the small church organ, and one might feel that a more liquid flute or tibia would be an asset for the home. It is quite easily formed from the waveforms available.

Apart from the Baldwin organs<sup>11</sup>, the most interesting application of frequency division is to be found in the

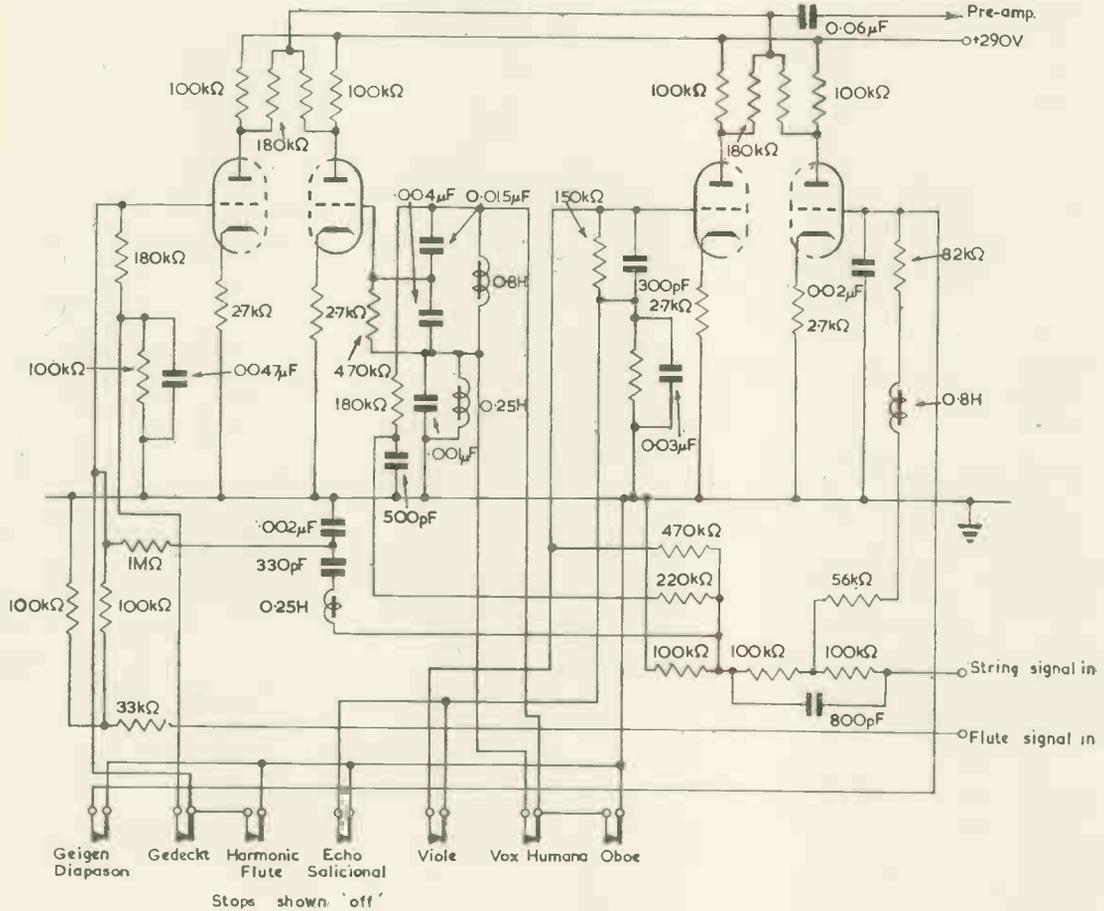


Fig. 8. Conn tone circuit, swell manual

always of the opinion that the tonal possibilities were not fully exploited. In the new instruments, a much better synthesis is achieved and the circuits are partly shown in Fig. 8. As has been explained<sup>10</sup>, it is impossible to add individual synthesized tones if these are derived from one waveform only, or if the filters or tone-forming elements are substantially common to one waveform. The Conn oscillators provide a sinusoidal output as well as a separate pulse. It is from this latter that most of the complex tones are formed. Very wisely, no attempt is made to form tones outside the capabilities of these waveforms; for instance, a sawtooth wave would permit many more tones than are possible from a quite triangular pulse.

The use of isolating valves is a sound way to avoid intermodulation, and as there are separate groups of oscillators for each keyboard and the pedals, a good chorus effect is possible. This company also favours the use of multiple contacts under the keys, but in this case

Minshall range of instruments. Phase shift oscillators feed a chain of cascaded dividers with the configuration of Fig. 9, which work as follows.

The first peak of the input signal voltage reduces the anode voltage as the valve conducts, and this is fed back to the control grid delayed in time by the constants of the feedback resistor and grid capacitor. So when the second peak of the input signal arrives, the grid potential is more negative (due to the fed-back signal) than when the first peak arrived. The input amplitude is not now high enough to overcome this extra bias and the second peak therefore causes no conduction.

The anode capacitor is thus not discharged, but in fact continues to charge up. By the time the third input peak arrives, the bias has returned to its original value and the anode capacitor is discharged. The resulting voltage drop is fed back to the grid, arresting the fourth input peak. The valve therefore conducts on alternate input peaks

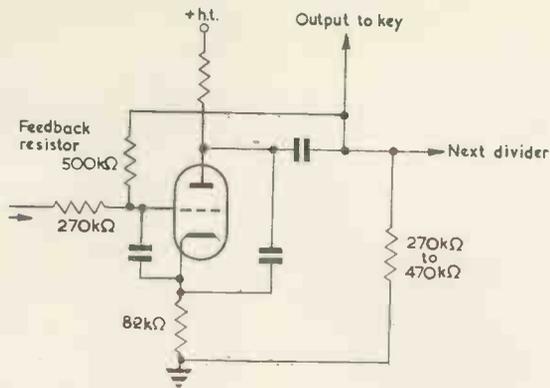


Fig. 9. Minshall frequency divider

only, and so divides the frequency by two. The circuit is to some extent frequency dependent, but will function over about half an octave with the same component values. The waveforms are shown in Fig. 10.

An attempt has been made to obtain the synthesis by simpler or more effective circuits than in the past; it might be added that the patent situation has a lot to do with this. In the Minshall system the tone switches or "stops" are shunt connected. Fig. 11 shows a typical circuit for one keyboard. The valve is connected as an amplifier with a reactive phase shift network between anode and grid. The impedance between input and earth will then vary markedly with frequency according to the proportions of the circuit constants.

Over some band of frequencies the phase of the feedback circuit reinforces the input voltage, and a peak in the response is produced. At low frequencies the phase of the feedback signal opposes the grid signal, producing a low impedance from here to earth. The response then falls off towards the bass end until the reactance of the input coupling capacitor becomes significant, when it will begin to rise again. The circuit acts like a resonant filter, peaking at one frequency and having attenuation at another. Control over these pass- and stop-bands is effected by the auxiliary networks marked with the tone-colours. The response of the networks is shown in Fig. 12. It will be noticed that the "diapason" uses the waveform as generated without modification. Such a music generator requires octave and sub-octave couplers to exploit the system fully.

In general, it may be said that no new oscillator circuits have appeared for musical instruments and no new loudspeakers have been developed, although the 27in Altec-Lansing is used by the Allen Organ Company for large installations. The introduction of two overlapping keyboards as in the Hammond "Spinet" is becoming

common, and for the smaller organs a fixed arrangement of 13 or 18 pedal keys is more or less universal, to give an elementary bass. The better-class instruments seek to develop the potential tonal content more accurately regardless of cost or complication, while the less expensive ones have tried to find still simpler and cheaper circuit artifices.

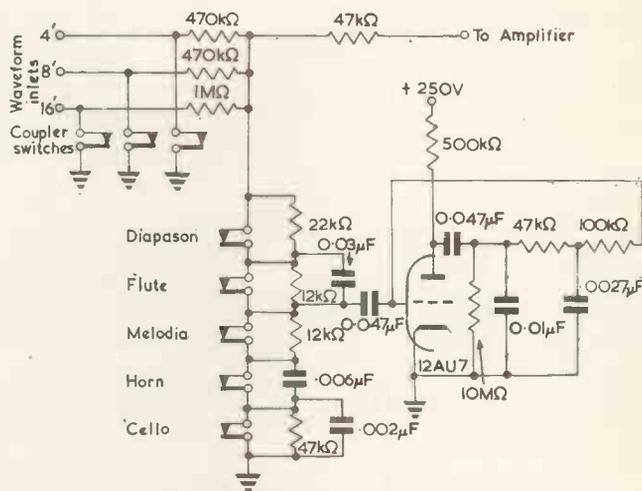


Fig. 11. Minshall tone forming circuits

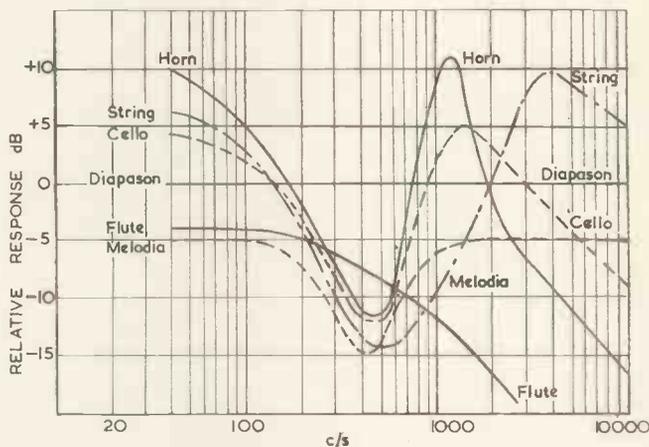


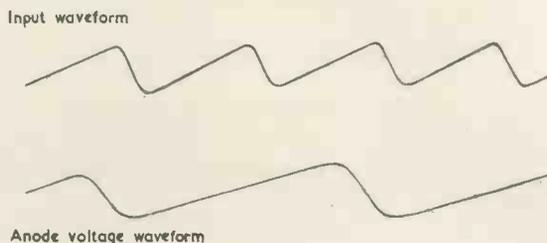
Fig. 12. Effect of tone filters on response

The impact of the electrical music generator on the domestic market has been remarkable. At one time, of course, the reed organ was to be found in nearly every home; hence the term *American organ*. Now its modern counterpart is well on the way to recapture the market. The magnitude of this can be gauged from the sales of one company alone in 1953—\$10 000 000.

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Fig. 10. Divider wave forms



# A High-Speed Waveform-Sampling Circuit

By G. D. Bergman\*, B.Sc., and D. M. MacKay\*, B.Sc., Ph.D.

*This article describes a new method of sampling a rapidly changing voltage waveform at a high sampling frequency. The circuit employs negative feedback in such a way that sampling pulses of specially small duration and amplitude can be used. Two practical forms of circuit are described which give a stepped output, and a theoretical discussion of the important design parameters is given.*

**P**ULSED sampling circuits can be designed to give outputs of two basic types, either a series of pulses modulated in height or duration to represent the amplitude of the sampled signal, or a stepwise modulated voltage which remains steady at each sampled value during the interval until the arrival of the next sampling pulse (Fig. 1).

This article is concerned with circuits of the second type, required to give a stepped output.

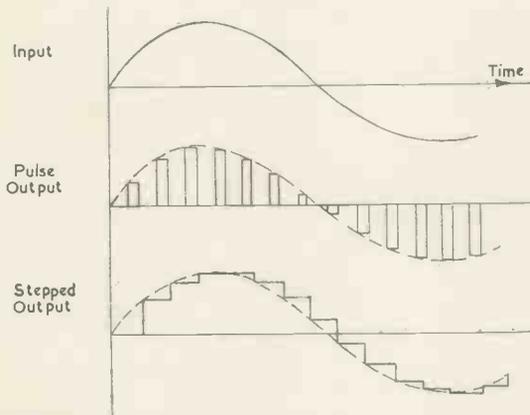


Fig. 1. Two types of output for pulsed sampling circuits

A conventional sampling circuit of this type is shown in essentials in Fig. 2. Short positive and negative sampling pulses  $P$  and  $Q$  respectively render diodes  $D_{1-4}$  conducting in order to allow the storage capacitor  $C$  to charge or discharge to the potential of the cathode of  $V_1$ , which follows the input.

When  $P$  and  $Q$  return to suitable resting levels  $C$  is left isolated and the output from the cathode of  $V_2$  remains steady in the absence of leakage.

Circuits of this type suffer from two disadvantages that limit their range of application:

- (a) The sampling pulses  $P$  and  $Q$  (of opposite polarity) must have amplitudes greater than the peak-to-peak input voltage range and must return to resting-levels below minimum and above maximum input respectively (see Fig. 3). At high sampling rates, where very short and rapidly-rising pulses must be used, this large voltage-swing demands a pulse generator of undesirably high power.
- (b) The heater-cathode leakage resistance ( $R_e$ ) of diode  $D_3$  sets an upper limit of  $CR_e$  to the time-constant of the

storage circuit unless the heater can be tied to the cathode and isolated. The output tends in consequence to "droop" unless the pulse-to-pulse interval is short compared with  $CR_e$ . To offset this by increasing the

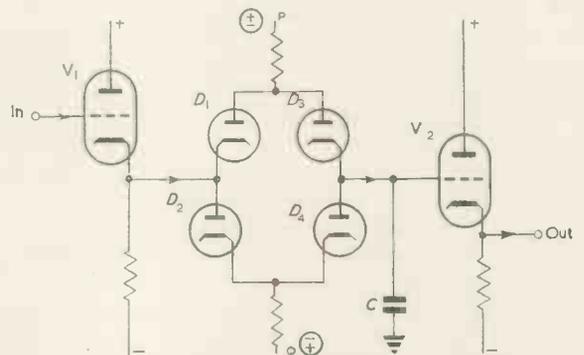


Fig. 2. A conventional pulsed sampling circuit to give a stepped output

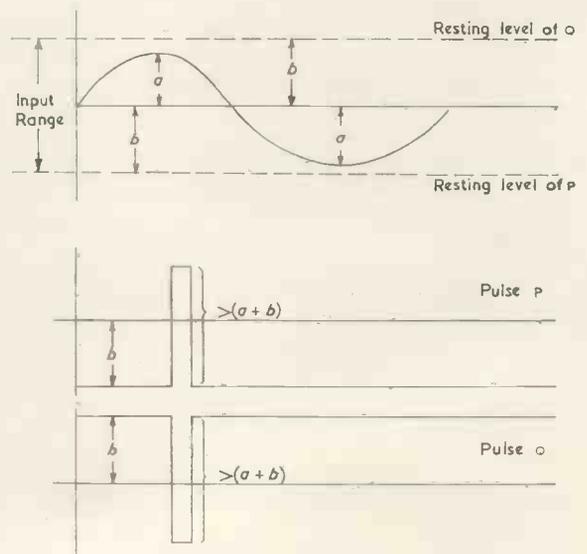


Fig. 3. Voltages required to operate conventional circuit of Fig. 2

value of  $C$  would increase proportionately the minimum time required to charge or discharge the capacitor, unless again a higher-powered sampling pulse were used.

In an effort to reduce these demands to a minimum, a

\* Dept. of Physics, King's College, University of London.

sampling circuit has been devised which requires much smaller switching-pulse amplitudes and can tolerate heater-cathode leakages greater by several orders of magnitude<sup>1</sup>.

### Basic Principles

The basic principle of this circuit is illustrated in Fig. 4.

A cathode-follower  $V_1$  receives the input  $V_{in}$  via a resistor  $R_1$ , and feedback from the output  $V_{out}$  via resistor  $R_2$ . A diode-switch circuit (represented by  $S$ ) leads from the cathode-follower to the input of a sign-reversing amplifier  $A$  of gain  $-\alpha$ , between whose input and output terminals the storage capacitor  $C$  is connected.

The two essential features are:

- (a) Subtraction of the output from the input to the diode-switch so as to bring the input level at  $Y$  more or less to zero when the diode-switch  $S$  is closed. This greatly reduces the pulse-voltages needed to operate  $S$ .

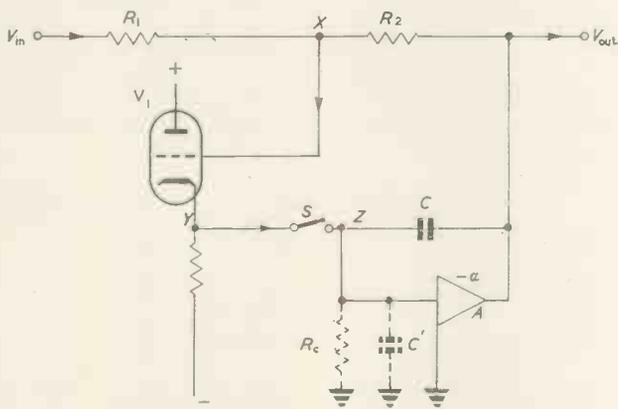


Fig. 4. Simplified diagram of new sampling circuit

- (b) Connexion of the capacitor  $C$  between input and output of amplifier  $A$ , so as to amplify the effective time-constant of the input circuit without increasing the charging current required.

The basic operation of the circuit may be followed in two stages.

- (1) If  $S$  is closed, the circuit functions as in a conventional seesaw circuit so as to bring the output voltage to a value approximately  $-(R_2/R_1)$  times the instantaneous input voltage. During this phase the capacitor  $C$  is charged to a voltage approximately equal to the output voltage, the necessary current flowing via  $S$ .
- (2) If  $S$  is now opened, the output remains clamped at the sample value by virtue of the negative feedback exerted through the charged capacitor  $C$ . Feedback through  $R_2$  ceases and the grid and cathode of  $V_1$  are free to follow changes in the input attenuated in the ratio  $R_2/(R_1 + R_2)$ .

If the leakage resistance from the diode switch  $S$  to earth is  $R_e$ , the effective leakage time-constant, owing to the well-known Miller effect, is now  $(\alpha + 1)CR_e$ , which can in theory be made as large as desired by making  $\alpha$  sufficiently large.

The output impedance while  $S$  is closed will be approximately the reciprocal of the loop transconductance,

$$Z_{out} = (1 + R_2/R_1)/g_m'$$

where  $g_m'$  is the transconductance between the point  $x$  and the output of  $A$ .

With  $S$  open, the output impedance is approximately:

$$Z_{out} = 1/g_m$$

Where  $g_m$  is the transconductance from the point  $z$  to the output of  $A$ . If leakage through  $R_e$  is not neglected, this expression must be divided by the attenuation operator for the potential divider  $(C, R_e)$ , so that in operational notation:

$$Z_{out} = (1 + 1/R_e C p)/g_m$$

More exact expressions for these quantities are derived in Appendix 1.

The input impedance changes from approximately  $R_1$  when  $S$  is closed, to  $(R_1 + R_2)$  in series with the steady output voltage when  $S$  is open. Except when the amplitudes of successive samples are the same, there will be an abrupt change in the input current to  $R_1$  when  $S$

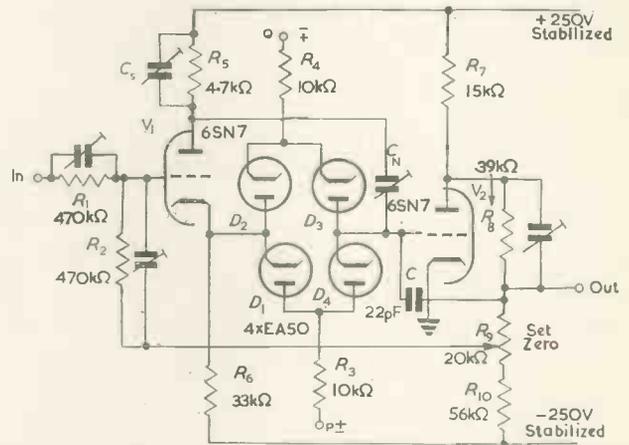


Fig. 5. A practical sampling circuit using the minimum number of valves

closes. Since, however, the impedance of  $R_1$  and  $R_2$  can normally be high, the discontinuity in load current will seldom have serious enough effects on the preceding stage to justify the use of an isolating cathode-follower ahead of  $R_1$ .

### Practical Circuit

A practical circuit using the minimum number of valves is shown in Fig. 5.

The diode switch  $D_{1-4}$  is supplied with pulses of the order of 20V in amplitude and 0.5μsec in duration. The grid of  $V_2$  need never change in potential by more than 5V, so that it is sufficient to prevent leakages if  $p$  and  $q$  are biased to some 5V below and above the mean grid voltage respectively.

While  $p$  and  $q$  are "resting" the cathode voltage of  $V_1$  may change, owing to a change in the input voltage, sufficiently to render  $D_1$  or  $D_2$  conducting.

Thus the value of resistors  $R_3$  and  $R_4$  must be high enough to avoid overloading the cathode-follower  $V_1$  under the worst conditions. At the same time  $R_3$  and  $R_4$  must be small enough to allow an adequate charging current to flow to or from  $C$  in a fraction of the duration of the sampling pulse.

If the maximum expected voltage change across  $C$  in one sampling operation is  $\Delta V$ , the mean current required in time  $\Delta t$  is  $C\Delta V/\Delta t$ . The pulse height above (or below) grid potential must then exceed  $C\Delta V/\Delta t$  times  $R_3$  or  $R_4$ .

For  $C = 20\text{pF}$ ,  $\Delta V = 25\text{V}$ ,  $\Delta t = 0.5\mu\text{sec}$  and  $R_3 = R_4 = 10\text{k}\Omega$  this gives a minimum pulse-height of  $10\text{V}$  above or below grid potential.

With the h.t. voltages shown, the input range of the circuit of Fig. 5 was limited (by the output stage) to some  $\pm 20\text{V}$ , but this does not, of course, represent a limitation of the sampling circuit. With smaller signals, or signals changing in smaller steps, the pulse voltage could be still further reduced.

Since the signal may be conducted by  $D_1$  or  $D_2$ , while  $P$  and  $Q$  are resting, an appreciable signal may be transferred to the grid of  $V_2$  via the self-capacitance of  $D_3$  or  $D_4$ , which may not be negligible compared with the storage capacitor  $C$ .

To offset this effect the neutralizing capacitor  $C_n$  is connected to the grid of  $V_1$ , and supplied with a signal in anti-phase with the cathode voltage. The shunt capacitor  $C_s$  helps to equalize the effects of stray capacitance at the

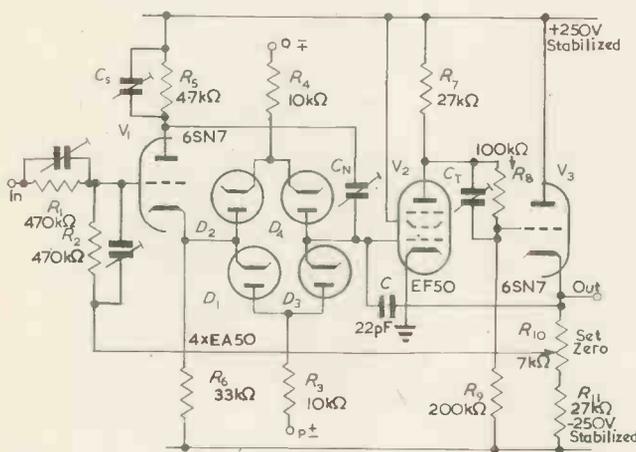


Fig. 6. A practical circuit with lower output impedance

cathode of  $V_1$ , and the arrangement results in an appreciable reduction of the stray signals in the output. It has been found especially valuable when the waveform sampled was a sawtooth, used in decoding position-modulated pulses.

If  $(R_1 + R_2)$  is of comparable order of magnitude with the output impedance an appreciable fraction of the input will be transferred to the output without change of sign during the resting periods. This fraction can, in particular cases, be made large enough to compensate for the strays across the diode switch mentioned in the previous paragraph. In some cases compensation may be excessive. It can then be neutralized by augmenting the stray capacitance across the diode switch. Since this adds to the capacitance between the grid of  $V_2$  and earth it will have the undesirable effect of increasing the output impedance during the resting periods. (See Appendix 1.)

The input and feedback resistors  $R_1$  and  $R_2$  are shunted by small trimming capacitors of the order of  $20\text{pF}$  to maintain a good high frequency response. The feedback through  $R_2$  must, of course, operate within a fraction of the duration of the sampling pulses.

The zero-level of the output is preset by adjusting the potential of the point from which feedback is taken to  $R_2$ .

The simple circuit of Fig. 5, designed for maximum economy, has a measured output impedance of  $7\text{k}\Omega$  and a leakage time-constant of the order of  $0.5\text{msec}$  with a

$22\text{pF}$  capacitor  $C$  and a leakage resistance  $R_0$  of  $2\text{M}\Omega$ .

It can obviously be much improved by replacing the triode  $V_2$  by a pentode with a cathode-follower output stage.

A satisfactory circuit incorporating these refinements is shown in Fig. 6.

It differs from Fig. 5, mainly in the range of output voltages it can handle, and in its output impedance.  $V_1$  fulfils the same function as  $V_1$  in Fig. 5, while  $V_2$  and  $V_3$  form the amplifier replacing  $V_2$  in Fig. 5. The d.c. coupling from  $V_2$  to  $V_3$  via  $R_8$  and  $R_9$  is frequency-compensated by the trimming capacitor  $C_T$ .

This circuit has a measured output impedance of about  $17\Omega$ , and was capable of handling input signals between  $\pm 50\text{V}$ , with the h.t. voltages shown.

#### APPENDIX

The circuit of Fig. 4 can be represented by the equivalent circuit shown in Fig. 7.

The cathode-follower is replaced by a voltage generator which has an output resistance  $R'$  and a gain  $+\beta$ . The amplifier  $A$  is replaced by a current generator of transconductance  $G_m$  and output resistance  $R''$ .  $C'$  represents the stray capacitance from the point  $Z$  to earth.

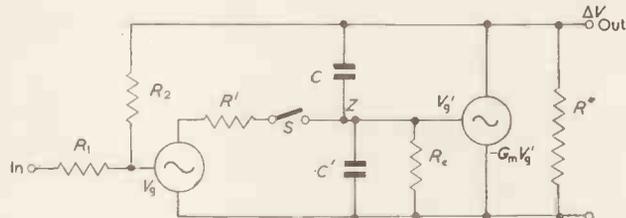


Fig. 7. Equivalent circuit of Fig. 4

The output impedance can be calculated for the two cases:

- (1)  $S$  closed, (2)  $S$  open.

#### CASE (1) $S$ CLOSED

Suppose the input is held at a constant potential and a voltage  $\Delta V$  is applied to the output terminal. The input voltage to the cathode-follower will change by an amount:

$$\Delta V_g = \frac{\Delta V R_1}{R_1 + R_2}$$

Since  $R'$  is very small compared with  $R_0$ , and small compared with the reactance of  $C$  and  $C'$  at the frequencies for which this equivalent circuit is applicable, the change in voltage at the grid of the amplifier  $A$  can be written:

$$\Delta V_g' = \frac{R_1}{(R_1 + R_2)} \Delta V \beta$$

The change  $\Delta V$  thus produces a change in current  $\Delta i$  at the output.

If the current in  $R''$  is neglected for the moment:

$$\Delta i = G_m \frac{R_1}{(R_1 + R_2)} \beta \Delta V$$

This corresponds to an output resistance:

$$\Delta V / \Delta i = (1 + R_2 / R_1) 1 / \beta G_m$$

This is in parallel with  $R''$  so the output resistance will be:

$$Z_{\text{out}} = \frac{R''(1 + R_2 / R_1) 1 / \beta G_m}{R'' + (1 + R_2 / R_1) 1 / \beta G_m}$$

CASE (2) S OPEN

Again  $\Delta V$  is applied to the output terminal. The voltage change at Z (in Transform notation) will be given by:

$$\begin{aligned} \overline{\Delta V}_z' &= \frac{\overline{\Delta V} R_o}{1 + pR_oC'} \\ &= \frac{\overline{\Delta V} R_o}{1/pC + \frac{R_o}{1 + pR_oC'}} \\ &= \frac{\overline{\Delta V} pR_oC}{1 + pR_o(C + C')} \end{aligned}$$

Neglecting  $R''$  the current change  $\Delta i$  produced by the voltage change  $\Delta V$  will be given by:

$$\begin{aligned} \overline{\Delta i} &= G_m \frac{\overline{\Delta V} pR_oC}{1 + pR_o(C + C')} \\ \therefore \overline{\Delta V}/\overline{\Delta i} &= \left( \frac{C' + C}{C} + \frac{1}{pR_oC} \right) 1/G_m \end{aligned}$$

The output impedance thus consists of a resistance  $\left( \frac{C' + C}{C} \right) 1/G_m$  in series with a capacitance  $R_oCG_m$ .

In parallel with this combination will be the output resistance of the amplifier  $R''$ .

If  $R_o$  is sufficiently large the capacitive part of the output impedance can be ignored. Then:

$$Z_{out} = \frac{R'' \left( \frac{C' + C}{C} \right) 1/G_m}{R'' + \left( \frac{C' + C}{C} \right) 1/G_m}$$

Approximate values for the parameters  $\beta$ ,  $G_m$ , etc. are given in Table 1 for both practical circuits.

TABLE 1

Approximate values of design parameters for both practical circuits

PARAMETER	TRIODE CIRCUIT (FIG. 5)	PENTODE CIRCUIT (FIG. 6)
$\beta$ —gain of input cathode-follower ..	0.9	0.9
$G_m$ —transconductance of amplifier A	0.3mA/V	125mA/V
$R''$ —output resistance of amplifier A	27k $\Omega$	400 $\Omega$
$R_o$ —leakage resistance from point z	Values vary for different diodes but are normally greater than 2M $\Omega$	
$\alpha$ —gain of amplifier A .. ..	8	50
C—storage capacitance .. ..	22pF	22pF
C'—stray capacitance from point z to earth .. ..	~10pF	~10pF
$Z_{out}$ calculated for S closed ..	6k $\Omega$	17 $\Omega$
$Z_{out}$ calculated for S open (neglecting $R_o$ ) .. ..	5k $\Omega$	12 $\Omega$
$Z_{out}$ measured (corresponding to $Z_{out}$ for S closed) .. ..	7k $\Omega$	17 $\Omega$
Leakage time constant (when $R_o = 2M\Omega$ ) .. ..	0.5msec	2msec

REFERENCE

1. MACKEY, D. M. British Patent Application 14482/54.

## A New Low Cost Marine Radar

OVER the past five years Decca Radar have introduced a series of high performance marine radar sets designed to meet the requirements of a wide range of shipping.

The majority of ships in the medium and large tonnage groups have now been fitted, but there is an increasing demand from owners of small ships for a set which will give them the benefits of radar navigation. It is primarily to meet this demand that Decca have developed their new equipment, the Decca 212, which has been designed to meet the rigorous and well tested demands of the British Ministry of Transport Marine Radar Specification.

It consists of three simple radar units, comprising a scanner and its associated turning motor, a radar frequency unit in which is included the magnetron, klystron and mixer, and a display unit with a 9in fixed focus cathode-ray tube.

The scanner unit has been designed for low top weight, and the total weight is of the order of 75lb. This low weight unit



River navigation on the one-mile range.

produces a narrow beam of 1.7° for high picture quality, with a very low side lobe level for good performance in narrow waters. The r.f. unit provides an output of 10kW with dual pulsing for solid paints at long range and crisp pictures at short range. It can be mounted below deck in smaller vessels or on larger vessels in a special waterproof box at the base of the mast.

A brief specification of the Decca 212 is given below:—  
FREQUENCY BAND. Within the Marine Band 9 320 to 9 480Mc/s.

AERIAL SYSTEM

Reflector: A single parabolic cylinder aerial of approx. 4ft.  
Rotation: Continuously rotating 24 rev/min.  
Horizontal Beam: 1.75° at half power points.  
Vertical Beam: 20° at half power points.  
Side Lobes: Better than 22dB within ± 10° of the main beam. Better than 30dB outside ± 10° of the main beam.

PULSE CHARACTERISTICS

Peak Power: 10 kW (approx.).  
Duration: 0.1 or 0.2 $\mu$ sec-pulse width. Selection automatically controlled from the range and anti-clutter switches.  
Recurrence Frequency: 1 000 per second.  
Pulse Generator: Triggered hard valve modulator driving a magnetron.

RANGE

Discrimination: 25 yards on shorter ranges.  
Minimum: 30 yards or less.  
Calibration Ring accuracy within 1 per cent of indicated range.

Range Scales and Calibration Rings.

Range scale	Calibration rings	Range scale	Calibration rings
1/2 mile	1/5th mile	8 miles	2 miles
1 mile	1/5th mile	15 miles	2 miles
3 miles	1/2 mile	30 miles	5 miles

BEARING

Accuracy: Within 1°.

# The Operation and Loading Characteristics of Valve Oscillators for Dielectric Heating

(Part 2)

By V. L. Atkins\*

## EFFECTIVE INTERNAL IMPEDANCE OF AN OSCILLATOR

When a valve is operated over the essentially linear portion of its characteristic curve such as obtained under class-A conditions the output current can be expressed as follows:

$$I_a = \frac{\mu V_g}{R_L + r_a} \dots \dots \dots (49)$$

Under these conditions the valve can be regarded as acting as a generator having an internal impedance value of

$$r_a (\text{=} \mu/g_m) \text{ delivering an output voltage of value } \frac{\mu V_g R_L}{R_L + r_a}$$

into an external load of value  $R_L$ . With class-C operation, however, the valve behaves in a more complex manner, the value of the internal impedance becoming a function of the angle of current flow  $2\theta_a$ .

The relation can be obtained by substituting for the value of  $V_g'$  as given by equation (22) in equation (34) and solving for  $I_a$ :

$$I_a = g_m/\pi \left( V_B + v_{g(\max)} - \frac{V_o - v_{a(\min)}}{\mu} \right) (\theta_a - \sin \theta_a \cos \theta_a) \dots \dots \dots (50)$$

but:

$$V_B + v_{g(\max)} = V_g \dots \dots \dots (51)$$

and:

$$V_o - v_{a(\min)} = I_a R_L \dots \dots \dots (52)$$

from which:

$$I_a = \frac{\mu V_g}{R_L + f_3(\theta_a) r_a} \dots \dots \dots (53)$$

where:

$$f_3(\theta_a) = \frac{\pi}{\theta_a - \sin \theta_a \cos \theta_a} \dots \dots \dots (54)$$

Thus the apparent internal resistance of a valve operating at an angle  $2\theta_a$  is equal to  $f_3(\theta_a)r_a$ . Values of  $f_3(\theta_a)$  are plotted in Fig. 3.

## OPERATING EFFICIENCY

The anode efficiency can now be computed in terms of  $R_L$  and  $f_3(\theta_a)r_a$ .

Anode efficiency =

$$\begin{aligned} \eta &= P_{\text{out}}/P_{\text{in}} = \frac{I_a^2 R_L}{2V_o I_o} = \frac{I_a R_L}{2V_o} \cdot I_a/I_o \dots \dots (55) \\ &= \frac{\mu V_g R_L}{V_o(R_L + f_3(\theta_a)r_a)} \cdot f_1(\theta_a) \times 50 \text{ per cent} \dots \dots \dots (56) \end{aligned}$$

Difficulty is usually experienced when endeavouring to establish accurately the performance of a power oscillator under varying conditions, as a change in a single parameter can affect most of the other parameters of the oscillating system. Furthermore, the operating characteristics

of a valve are not linear over the full working range as required to obtain the maximum power output. Fortunately, however, various assumptions can be made which greatly simplify the working computations, and allow operating conditions sufficiently accurate for most practical purposes to be evolved.

The anode efficiency can be written as follows:

$$\text{Anode efficiency} = V_a/V_o \eta_a$$

where

$$\eta_a = 50 \cdot f_1(\theta_a) \text{ per cent} \dots \dots \dots (57)$$

This shows that the maximum theoretical efficiency that can be obtained for a given operating angle  $2\theta_a$  is equal to  $\eta_a$ . This is known as the asymptotic efficiency, values of which are plotted in Fig. 5. The actual operating efficiency can be readily assessed by multiplying the asymptotic efficiency by the ratio of  $V_a/V_o$  or the factor  $(1 - \tau)$ .

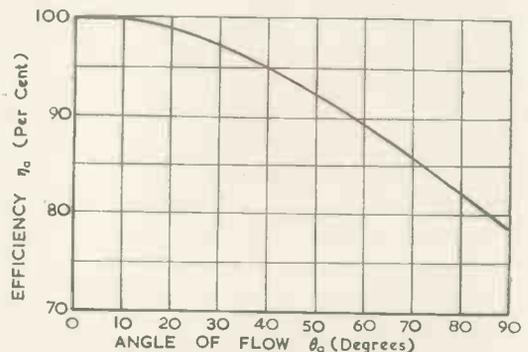


Fig. 5. Theoretical maximum efficiency of an oscillator as governed by the operating angle

## MAXIMUM EFFICIENCY OPERATING CONDITIONS

In the designing of a power oscillator it is necessary to fix the operating values of the minimum anode voltage  $v_{a(\min)}$  and the maximum grid voltage  $v_{g(\max)}$  which occur at the same instant. At this point the anode and grid currents reach peak value and the values of  $v_{a(\min)}$  and  $v_{g(\max)}$  are chosen so that the total peak current drawn equals or is slightly less than the maximum permissible emission of the cathode. Usually the value of  $v_{a(\min)}$  which should be slightly positive with respect to  $v_{g(\max)}$ , is arranged to lie between 5 per cent and 10 per cent of the h.t. supply voltage  $V_o$  and  $v_{g(\max)}$  between 2 per cent and 8 per cent of  $V_o$ . Saturation occurs when  $v_{g(\max)}$  equals  $v_{a(\min)}$  and on no account is the value of  $v_{g(\max)}$  allowed to exceed  $v_{a(\min)}$ , otherwise the grid current becomes excessive resulting in overheating of the grid electrode. If the ratio of  $v_{a(\min)}/V_o$  is made large both output power and efficiency are reduced, while the grid loss is increased. On the other hand the value of  $v_{g(\max)}$  cannot be made too small, otherwise the drive becomes insufficient for maximum conditions. The problem then remains to determine the values

\* Metropolitan-Vickers Electric Co., Ltd.

of  $v_{a(\min)}$ ,  $v_{g(\max)}$ , the load impedance and operating angle which will deliver maximum output for a given loss as governed by the valve design and construction. By taking the limiting conditions where  $v_{a(\min)}$  equals  $v_{g(\max)}$  as a reference level then a solution to the problem can be evolved as follows

At saturation:

$$v_{g(\max)} = v_{a(\min)} \dots \dots \dots (58)$$

from which:

$$V_g = V_o - I_a R_L + V_B \dots \dots \dots (59)$$

substituting for  $V_B$  as derived from equations (21) and (22)

$$V_g = \frac{V_o(1 + 1/\mu) - I_a R_L(1 + \cos \theta_a/\mu)}{1 - \cos \theta_a} \dots \dots \dots (60)$$

substituting this expression in equation (53) and solving for  $I_a$ :

$$I_a = \frac{V_o(1 + \mu)}{R_L(1 + \mu) + r_a(1 - \cos \theta_a)f_3(\theta_a)} \dots \dots \dots (61)$$

calling  $\frac{1 + \mu}{r_a} = \rho$  and  $f_3(\theta_a)(1 - \cos \theta_a) = f_4(\theta_a)$  then:

$$I_a = \frac{\rho V_o}{\rho R_L + f_4(\theta_a)} \dots \dots \dots (62)$$

the function  $f_4(\theta_a)$  is plotted in Fig 3.

The power output can be expressed thus:

$$P_{out} = \left( \frac{\rho V_o}{\rho R_L + f_4(\theta_a)} \right)^2 R_L/2 \dots \dots \dots (63)$$

and the power input as:

$$P_{in} = \frac{\rho V_o^2}{f_1(\theta_a)(\rho R_L + f_4(\theta_a))} \dots \dots \dots (64)$$

where

$$f_1(\theta_a) = \frac{\theta_a - \sin \theta_a \cos \theta_a}{\sin \theta_a - \theta_a \cos \theta_a}$$

therefore the power loss becomes:

$$P_a = V_o^2 \rho \left( \frac{1}{f_1(\theta_a)(\rho R_L + f_4(\theta_a))} - \frac{\rho R_L}{2(\rho R_L + f_4(\theta_a))^2} \right) \dots \dots \dots (65)$$

$$= V_o^2 \rho \Psi(\theta_a, \rho R_L) \dots \dots \dots (66)$$

The efficiency at saturation can be written as:

$$\eta_{(sat)} = \frac{f_1(\theta_a) I_a R_L}{2V_o} \dots \dots \dots (67)$$

$$= \frac{f_1(\theta_a) \rho R_L}{2(\rho R_L + f_4(\theta_a))} \dots \dots \dots (68)$$

$$= f(\theta_a, \rho R_L) \dots \dots \dots (69)$$

The function  $\Psi$  is a constant which can be applied to all valves and can be computed once the anode voltage  $V_o$  and power loss  $P_a$  have been selected, i.e.:

$$\Psi = P_a/V_o^2 \rho \dots \dots \dots (70)$$

By plotting values of  $\Psi$  and  $\eta_{(sat)}$  as functions of  $\theta_a$  and  $\rho R_L$  as shown in Fig. (6) and projecting vertical lines from the intersections obtained on the  $\Psi$  family of curves for a constant value of  $\Psi$ , to the corresponding values of  $\theta_a$  in the  $\eta_{(sat)}$  family of curves, the values of  $\theta_a$  and  $\rho R_L$  which give maximum efficiency conditions can be ascertained. Determining the value of  $v_{a(\min)}$  ( $= v_{g(\max)}$ ) at the optimum conditions is simply obtained from the relation:

$$v_{a(\min)} = V_o - I_a R_L \dots \dots \dots (71)$$

which written in terms of  $\rho R_L$  and  $\theta_a$  becomes:

$$v_{a(\min)} = V_o \left( 1 + \frac{\rho R_L}{\rho R_L + f_1(\theta_a)} \right) \dots \dots \dots (72)$$

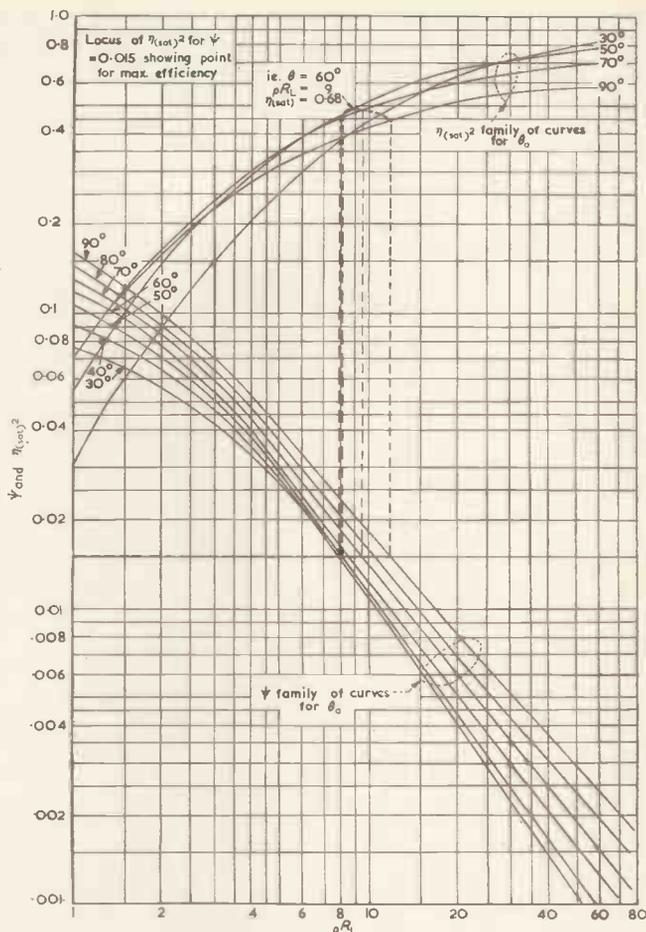


Fig. 6. Universal characteristics for designing power oscillators  $\Psi$  being a function of the anode voltage and the maximum permissible anode dissipation,  $\rho$  being a function of the amplification factor and internal impedance of the valve

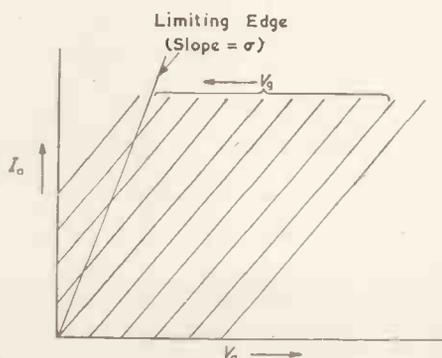


Fig. 7. "Ideal" set of  $I_a/V_a$  characteristic curves for triode valve showing the limiting edge where  $V_a = V_g$ .

Examination of the curves drawn in Fig. 6 shows that the efficiency at optimum conditions is greater the smaller the value of  $\Psi$ . This requires the value of  $V_o$  to be the maximum possible and shows that for a given loss the greater the amplification factor ( $\mu$ ) and smaller the  $r_a$  of the valve, greater will be the power output and efficiency.

Over the normal working range of a valve the grid family of  $I_a/V_a$  characteristic curves can be taken to be linear and it can be assumed for practical purposes that the line on which they terminate where  $v_a = v_g$  will also be straight

passing through the point of origin. This is illustrated in Fig. 7. Calling the slope of this limiting edge  $\sigma$  then the peak current drawn can be expressed as follows:

$$i_{a(pk)} = \sigma V_{a(min)} \dots \dots \dots (73)$$

Now at saturation the fundamental alternating current component is:

$$I_a = \frac{\rho V_o}{\rho R_L + f_4(\theta_a)} \dots \dots \dots (74)$$

In terms of the peak current the expression becomes:

$$i_{a(pk)} = \frac{\rho V_o}{\rho R_L + f_4(\theta_a)} \cdot f_2(\theta_a)/f_1(\theta_a) \dots \dots \dots (75)$$

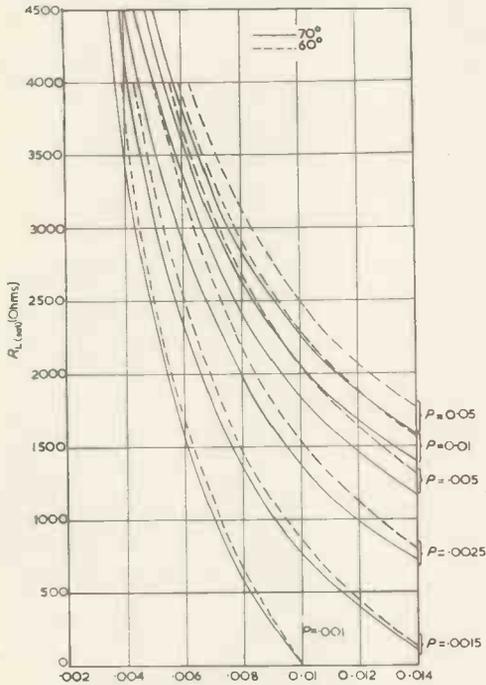


Fig. 8. Curves showing load at saturation for  $\tau = 0.1$  assuming ideal linear characteristics of Fig. 7

Substituting for equation (73) and evaluating in terms of the load  $R_L$ :

$$R_L = [V_o/\sigma v_{a(min)} \cdot f_2(\theta_a)/f_1(\theta_a)] - f_4(\theta_a)/\rho \dots (76)$$

Under normal operating conditions the ratio of  $V_o/v_{a(min)}$ ,  $(1/r)$  approximates to 10, thus the load at saturation can be expressed as follows:

$$R_{L(sat)} = [10/\sigma \times f_2(\theta_a)/f_1(\theta_a)] - f_4(\theta_a)/\rho \text{ approx. } \dots (77) \\ = f_4(\theta_a) (10/\sigma - 1/\rho)$$

For a valve of known constants  $\sigma$  and  $\rho$ , the relation gives for a specific operating angle the minimum value of load into which the valve can work before reaching saturation. Normally the load is designed to be slightly higher. Curves giving values of  $R_{L(sat)}$  at operating angles of  $60^\circ$  and  $70^\circ$  for different values of  $\sigma$  and  $\rho$  are shown in Fig. 8.

#### LOADING CHARACTERISTICS OF A VALVE OSCILLATOR

As shown later an inherent feature of a self-excited oscillator employing self-bias is that under normal operating conditions the value of  $V_a$  which almost reaches the value of the d.c. supply voltage  $V_o$ , only slightly reduces when the load is varied from no-load to full load conditions. Thus in the equation of anode efficiency (equation (56)) the part in parenthesis will tend to approach unity and the value of  $V_g$  which represents the proportion of the anode oscillation

fed back to the grid, can be assumed as remaining substantially constant. With  $\mu$ ,  $V_o$  and  $r_a$  being constants,  $f_3(\theta_a)$  becomes a function of  $R_L$  which thus determines the operating angle  $2\theta_a$ . As the value of the operating angle is controlled by the amount of the grid bias voltage  $V_B$  ( $V_g$  remaining constant in this case) then it follows that equilibrium conditions of the oscillating system are obtained when the grid bias voltage  $V_B$  reaches a value such that

$\frac{\mu V_g R_L}{V_o(R_L + f_3(\theta_a)r_a)}$  approximates to unity. Equating the expression to unity and solving for the value of  $\mu V_g/V_o$

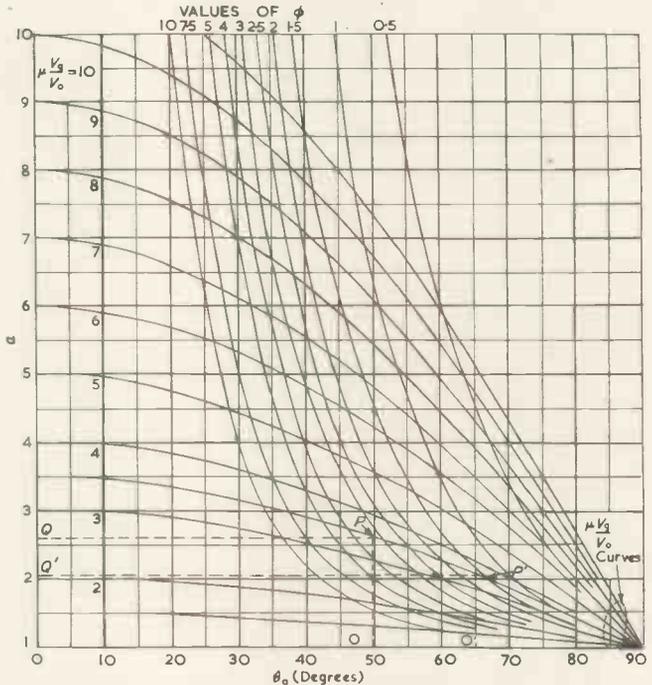


Fig. 9. Functions of  $\alpha$  for analysis of the loading characteristics of an oscillator. The dotted lines refer to cross co-ordination with Fig. 11

$$\mu V_g/V_o = \frac{R_L + f_3(\theta_a)r_a}{R_L} \dots \dots \dots (78)$$

and calling the ratio  $R_L/r_a$ ,  $\phi$

$$\mu V_g/V_o = (\alpha + \tau) = 1 + f_3(\theta_a)/\phi \dots \dots \dots (79)$$

From equations (21) and (22)  $\mu V_B/V_o$  can be expressed in terms of  $\mu V_g/V_o$  and  $\theta_a$  and again taking the value of  $V_a/V_o$  to be unity, then:

$$\mu V_B/V_o = (\mu V_g/V_o - 1) \cos \theta_a + 1 \dots \dots \dots (80)$$

Substituting for  $\mu V_g/V_o$  in equation (79) a further relation for  $\mu V_B/V_o$  is obtained, viz:

$$\mu V_B/V_o \frac{f_3(\theta_a) \cos \theta_a}{\phi} + 1 \dots \dots \dots (81)$$

In applying these relations it must be appreciated that they are based on the assumed "ideal" conditions where  $V_a$  becomes equal to  $V_o$ , whereas in practice this is not the case. Furthermore the relation cannot hold for all values of  $V_g$  and  $\phi$  owing to the limiting conditions reached at saturation or where insufficient grid current is drawn to provide the necessary bias required for class-C operation. The value of  $V_g$  is also limited by the maximum grid power loss. The relations do, however, form a basis upon which the loading characteristics for a particular valve may be evolved. For

having determined the value of  $V_g$ , the values of  $\phi$  and  $\mu V_B/V_o$  ( $\alpha$ ) for a given operating angle can be easily assessed. The plotting of a series of curves based on equations (80) and (81) as shown in Fig. 9 will indicate how for a constant value of  $\mu V_g/V_o$  the values of  $\alpha$  and  $\theta_a$  vary with respect to  $\phi$ . On the assumption that the amplitude of the grid voltage ( $V_g$ ) remains fairly constant for a reasonable "falling off" in load from maximum value, and providing the correct drive conditions are operating, then a useful relation between load and grid bias voltage is obtained. Thus a quantitative indication of a changing load is provided, once the operating conditions for a specific load are established and changes in the grid current are then observed with a change in load.

In designing a power oscillator it is necessary to establish the optimum value of anode load required to deliver the desired amount of power for specific operating conditions. The optimum value as chosen is based on obtaining a reasonable compromise between efficiency and power output. In practice efficiencies of between 60 per cent and 80 per cent are usually demanded to obtain maximum output as limited by the anode dissipation and peak filament emission of the valve. In order to obtain optimum loading conditions, the impedance (resistance) presented to the valve by the tank circuit must be equal to  $V_a/I_a$ , where  $I_a$  is the maximum value as permitted by the selected operating angle. Designating the optimum values by the suffix (opt) then:

$$R_{L(\text{opt})} = V_a/I_{a(\text{opt})} = Q_{(\text{opt})} L_{(\text{opt})} \omega \dots (82)$$

and the optimum value of the tank inductance will be:

$$L_{(\text{opt})} = R_{L(\text{opt})}/Q_{(\text{opt})} \omega \dots (83)$$

Under optimum loading conditions the peak filament emission which is determined by the values of  $v_{a(\text{min})}$  and  $v_{g(\text{max})}$ , reaches its maximum permissible value. The optimum value of  $R_{L(\text{opt})}$  can be expressed in terms of  $i_{a(\text{pk})}$  as follows:

$$R_{L(\text{opt})} = V_a/i_{a(\text{pk})} \cdot \left( \frac{1 - \cos \theta_a}{\theta_a - \sin \theta_a \cos \theta_a} \right) \dots (84)$$

at optimum conditions  $V_a \approx V_o$ .

$$\therefore R_{L(\text{opt})} = V_o/i_{a(\text{pk})} \cdot \left( \frac{1 - \cos \theta}{\theta - \sin \theta \cos \theta} \right) \dots (85)$$

$$= V_o/i_{a(\text{pk})} \cdot f_a(\theta_a) \text{ (approx.)} \dots (86)$$

Values of  $R_{L(\text{opt})}$  for varying values of  $V_o/i_{a(\text{pk})}$  and operating angle  $\theta_a$  are plotted in Fig. 10.

Departure from the optimum value results in either a reduction of power output or overloading of the valve when the load becomes excessive. The behaviour of a power oscillator under varying load conditions calls for special consideration when designing equipment for high frequency heating, as in most applications the load varies considerably.

The value of the a.c. voltage  $V_a$ , ( $V_o - v_{a(\text{min})}$ ) developed across the tank circuit depends upon the values of the fundamental a.c. current component  $I_a$  and the load impedance  $R_L$ . Under light load conditions the value of  $V_a$  will be at its maximum value, while the value of  $I_a$  will be at its minimum value. As the value of the load increases (i.e.  $R_L$  reduces) towards optimum value,  $V_a$  will tend to become less, while  $I_a$  will increase in order to supply the extra power to the tank circuit. The increase in the value of  $I_a$  is brought about by an increase in the angle of current flow  $2\theta_a$  which also increases the average input power. This arises from the fact that as the value of  $V_a$  reduces and the grid drive becomes less, the grid voltage  $v_{g(\text{max})}$

becomes relatively less positive with respect to the cathode, causing the value of the grid current to fall. This in turn reduces the value of the grid bias voltage  $V_B$  which causes the value of the grid voltage  $v_{g(\text{max})}$  and the operating angle  $\theta_a$  to increase. The action is self-adjusting, the bias level changing until new conditions of equilibrium are reached, where the increase in  $I_a$  obtained is sufficient to supply the increase in power required. With the anode current being thus sensitive to changes in the value of  $v_{a(\text{min})}$ , the amplitude of oscillation tends to remain substantially constant for reasonable changes in load. Thus the oscillator can be regarded as being virtually a constant output voltage generator, having an internal e.m.f. of  $\mu V_g$  volts. In practice the amplitude of  $V_a$  varies from approximately 95 per cent to 90 per cent of the d.c. supply voltage  $V_o$  when the

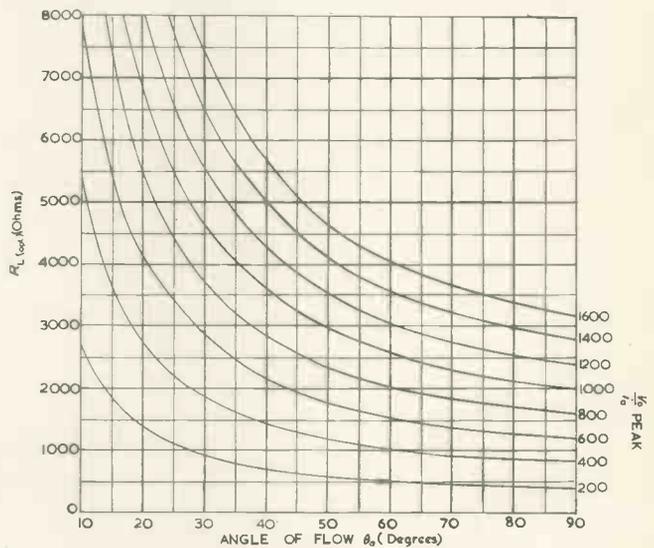


Fig. 10. Curves giving approximate value of optimum load for different value  $\theta_a$  and  $V_o/i_{a(\text{pk})}$

load is varied from no-load to full-load conditions. The value of  $V_a$  can never reach the value of  $V_o$  as this would give rise to conditions of no anode current flow and the grid current becoming dangerously excessive. On the other hand if the valve is loaded beyond its optimum rated output, saturation of the filament emission is reached and the value of  $I_a$  can no longer increase, with the result that the value of  $V_a$  falls and so increases the power dissipated at the anode resulting in the valve becoming overheated.

Although variation of load between no-load and full-load conditions causes only a slight reduction in the value of  $V_a$ , the change which occurs to both power output and anode efficiency is considerable. The effects which are readily observed when the load increases are the increase in the d.c. input current and the reduction in the grid current flow. What is not so obvious is the increase in the angle of flow of  $2\theta_a$ , which, apart from increasing the power input, changes the operating efficiency. As  $2\theta_a$  and  $I_o$  increase, the ratio of  $I_a/I_o$  also increases and thus the proportion of the increase in power input which is dissipated in the valve becomes greater, while the proportion which is delivered to the load becomes correspondingly less. If the load remains constant an increase in the grid excitation voltage  $V_g$  will cause the bias voltage  $V_B$  and the maximum grid voltage  $v_{g(\text{max})}$  to increase and the oscillation will assume a new point of equilibrium at a

reduced angle of flow. The grid loss will increase, but the anode efficiency will improve slightly, owing to the fact that with the amplitude of the output voltage  $V_a$  tending to remain constant, the fundamental alternating current component  $I_a$  will also remain fairly constant. Thus with the reduced angle of flow the direct current component  $I_o$  will be reduced and as a result so will the power input. The value of  $V_{g(max)}$  is, of course, limited to the point when the valve becomes overdriven, where apart from causing over dissipation of power in the grid circuit, the anode current waveform becomes considerably distorted and the algebraic analysis as used to determine the operating conditions of the valve can no longer be applied.

**DETERMINATION OF VALVE OPERATING CONDITIONS**

The effect which variable loading has on the operating performance of a power oscillator needs to be examined closely, and is one of the major problems associated with the design of h.f. heating equipment. An approximate method showing how the performance of an oscillator reacts to a changing load can be evolved fairly readily if it is assumed that little change occurs to both output voltage and frequency for reasonable variations in load. For under such conditions the power dissipated in the tank circuit will be inversely proportional to the value of the load resistance  $R_L$ , which includes the circuit losses. Thus if the optimum load resistance is termed  $R_{L(opt)}$  and the actual loading resistance is termed  $R_{L(act)}$ , then:

Power dissipated in load/optimum power available  

$$= R_{L(opt)}/R_{L(act)} \dots \dots \dots (87)$$

If  $I_{a(opt)}$  and  $I_{a(act)}$  represent the optimum and actual values of the fundamental a.c. currents respectively,

$$R_{L(opt)}/R_{L(act)} = I_{a(opt)} \dots \dots \dots (88)$$

Applying this relation to equation (53) and assuming a constant value of grid voltage:

$$I_{a(opt)} (R_{L(opt)} + f_3(\theta_{a(opt)}) r_a) = I_{a(act)} (R_{L(act)} + f_3(\theta_{a(act)}) r) \dots \dots \dots (89)$$

with:

$$I_{a(opt)} R_{L(opt)} = I_{a(act)} R_{L(act)} = V_a \text{ (assumed constant)}$$

$$I_{a(act)}/I_{a(opt)} = f_3(\theta_{a(opt)})/f_3(\theta_{a(act)}) \dots \dots \dots (90)$$

where  $f_3(\theta_{a(opt)})$  and  $f_3(\theta_{a(act)})$  are the functions of the optimum and actual operating angles  $\theta_{a(opt)}$  and  $\theta_{a(act)}$  respectively. In terms of the direct current components the relation can be written as:

$$I_{o(opt)}/I_{o(act)} = \psi_{(opt)}/\psi_{(act)} = f(\theta_{a(opt)})/f(\theta_{a(act)}) \dots \dots (91)$$

where  $\psi = \sin \theta_a - \theta_a \cos \theta_a$ .

From this relation a series of curves can be plotted as shown in Fig. 11 giving values of  $\theta_{(act)}$  as against  $\theta_{(opt)}$  for different values of  $I_{o(act)}/I_{o(opt)}$ .

To a certain limited extent this set of curves can be used in conjunction with those of Fig. 9 to provide a means of establishing approximately the operating conditions of a valve oscillator when the value of the grid voltage  $V_g$ , the load  $R_L$  and the flow angle are not known; factors which in practice are not always easily measured.

In order to derive the operating conditions with the help of these graphs all that is required is the values of the anode and grid currents for two different loading conditions of the valve with constant grid drive. Although it is not essential that one of the loaded conditions be the optimum value, satisfactory results will only be obtained provided the oscillator is working within the limiting conditions as previously specified. It is usually fairly easy to vary the loading

of a power oscillator, in fact in general the reverse is the case, it being difficult to maintain constant loading owing to changes in the characteristics of a load during a heating cycle. To find the co-ordinates on the graph of Fig. 9 which will indicate the operating conditions, it is first necessary to determine the values of  $\alpha (= \mu V_B/V_o)$  for at least two different loading conditions with the value of  $V_g$  remaining constant. Knowing the values of the grid bias resistor,  $\mu$  and  $V_o$ , the values of  $\alpha$  are easily assessed from the respective grid currents. Having established the two  $\alpha$  ordinates it remains for the appropriate  $\theta_a$  abscissae to be determined in order to fix the operating points. The ratio of the corresponding anode currents gives the ratio of the functions of  $\psi$ , from which, with the aid of the

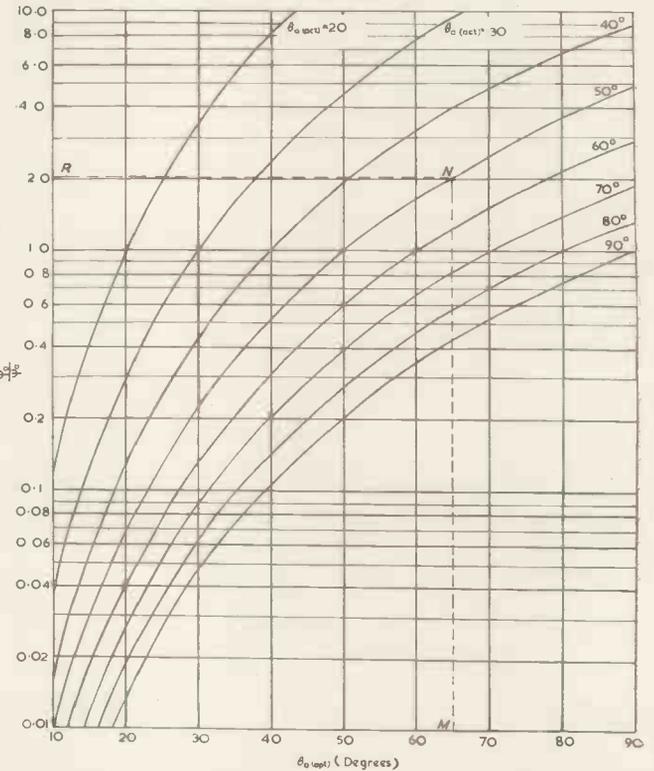


Fig. 11. Values of  $\psi_o/\psi_a$  giving the ratio of the d.c. currents obtained with a change in the angle of flow from  $\theta_{a(opt)}$  to  $\theta_{a(act)}$  (or  $\theta_{(act)}$  to  $\theta_{(opt)}$ ). This family of curves used in conjunction with those of Fig. 9 allow the operating conditions of a valve oscillator to be followed with variation in load. An example of the method is illustrated by the dotted lines

curves of Fig. 11, the two operating angles which satisfy the relations of both graphs may be evolved. The co-ordinates of the two operating points must lie on the same  $\mu V_g/V_o$  curve of the graph of Fig. 9,  $\mu V_g/V_o$  being assumed to remain constant, while the corresponding  $\theta_a$  co-ordinates on the graph of Fig. 11 must intersect on the correct  $\psi_{(opt)}/\psi_{(act)}$  ordinate. The method is illustrated by the dotted lines  $OP, O'P'$  and  $QP, Q'P'$  shown in Fig. 9 which represent respectively the two sets of the  $\mu V_B/V_o$  and  $\theta_a$  co-ordinates, the values of  $\theta_a$  corresponding to  $\theta_{a(act)}$  and  $\theta_{a(opt)}$  values as fixed by the co-ordinates  $MN, RN$ , as shown in Fig. 11.  $P, P'$  (Fig. 9) denotes the two operating points which lie on the constant  $\mu V_g/V_o$  line from which the value of  $V_g$  may be determined. The intersections made with the  $\phi$  series of curves allow the two different values of the load  $R_L$  to be found. The use of these graphs in the

manner as outlined can greatly assist the initial setting-up of a power oscillator. Furthermore, as reasonable changes in load (this includes circuit losses) may be ascertained, the graphs will be of particular use in dealing with the many applications met in practice where the load varies with change of temperature or moisture content etc. Obviously the problem is greatly simplified if the value of  $V_g$  is known, for then it is simply a matter of reading off values of  $\theta_a$  and  $\phi$  along the appropriate  $\mu V_g/V_o$  line for the various values of  $a$  obtained. Thus the anode current ratio curves of Fig. 11 would not be required.

Based on the assumption that the value of the alternating current component  $I_a$  and the anode voltage swing  $V_a$  remain substantially constant for slight changes in the grid drive voltage  $V_g$ , the relation between changes in the d.c. anode current with variation of grid drive for a constant load can be easily shown to be:

$$I_{o(opt)} / I_{o(act)} = f_1(\theta_{a(act)}) / f_1(\theta_{a(opt)}) \dots (92)$$

From this relationship a similar set of curves to those of Fig. 11 may be derived and so enable the operating points to be determined in the same manner as the previous case except that the curves of constant value will be those of the factor  $\phi$ . As previously explained the above treatment is restricted in operation to within certain limits of load variation. In practice this can be regarded as varying from light load conditions where the grid current drawn does not exceed the maximum permissible value to full (optimum) load conditions which consists of the working load plus the circuit losses. It is not essential to consider

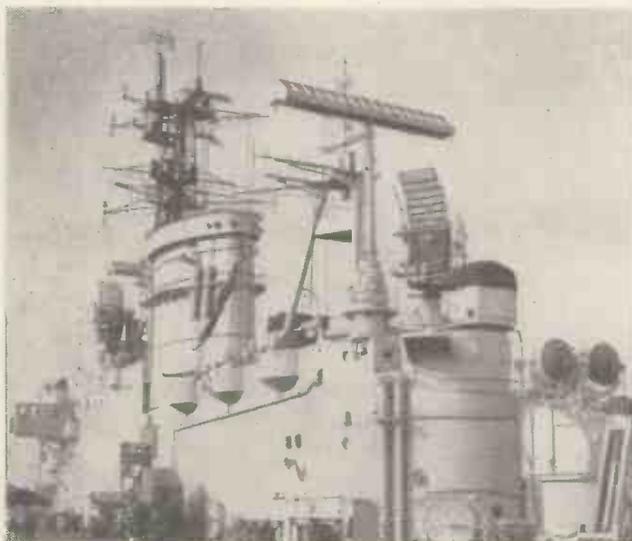
the behaviour of the operating conditions when the load is in excess of the optimum value as in general the equipment is so designed that it is fully protected from becoming overloaded.

A simple method which increases the amount of variable loading a valve oscillator can safely handle is obtained by arranging the grid bias resistor to have a non-linear characteristic possessing a positive coefficient. As previously mentioned, the maximum loading point is limited to the optimum value while the minimum loading point is limited to the maximum value of grid current. If the characteristic of the grid resistor is such that an increase in current increases the resistance and thereby increases the bias voltage, the minimum value of load, which can be imposed for the maximum value of grid current will be increased. It also means that for slight changes in load the grid current will tend to remain constant. In practice normal tungsten filament lamps are used operating at approximately half of their normal brightness.

Having now dealt with the operating characteristics of the valve oscillator under varying load conditions, it remains to investigate the problems associated with the transfer of power from the tank circuit to the load, and show how mismatching of the load impedance, linked with the circuit transfer efficiency, governs the overall performance of the equipment. This, together with typical power oscillator and work circuits, methods and control of power loading form the subject of a further article to be published at a later date.

## THE AERIAL ARRAY OF A MODERN BATTLESHIP

*The importance of radio and radar in modern naval warfare is exemplified by the accompanying photographs which show part of the aerial installation on the fighting control tower of H.M.S. Ark Royal the new 36 800 ton aircraft-carrier which was commissioned at Birkenhead in February.*



# Transistor Waveform Generators

By F. Butler, B.Sc., M.I.E.E., M.Brit.I.R.E.

*The characteristics of currently available transistors are so variable that the precise design of oscillator and pulse generator circuits is often impracticable. Many circuits can best be devised by making use of the analogy between transistor amplifiers and their valve-operated counterparts.*

*Descriptions are given of several experimental oscillators which have been developed by making use of these principles. They include tuned-circuit, crystal controlled and blocking oscillators.*

A GREAT deal of published work has appeared on the physics of point-contact and junction transistors and equivalent circuits have been derived from which the performance of idealized linear amplifiers may be predicted. The design problem is complicated because of wide variations in the characteristics of available transistors, nominally of the same type, and it is commonly found that individual adjustment of component values is required to meet a given specification. Because of this lack of uniformity in characteristics there are many transistor circuits which can best be devised by making intelligent use of a rather loose correspondence which exists between these and certain well-known valve circuits. It is the object of this article to describe a few useful oscillator and pulse generator circuits, mostly developed by cut-and-try methods aided by simple reasoning from the principles just mentioned.

## Transistor Operation

Point-contact transistors require that the collector element is biased negatively with respect to the base. In normal operation the emitter is positively biased and the emitter current should preferably be supplied from a high impedance (constant current) source. As in the case of valves, automatic biasing schemes may be devised so that only a single power supply is required.

Junction transistors are assembled from specimens of n- and p-type germanium, either in the sequence npn or pnp. Operationally the only difference is in the requisite polarity of the supply source. The pnp units call for the same polarity as the point-contact types while the npn units require that the batteries are reversed. Junction transistors will operate successfully at much lower power levels than the point-contact types.

Transistor units of all types have a very low thermal inertia and are easily damaged by momentary current surges. During development work a current-limiting resistor should be connected in the power supply leads and care must be taken to avoid the accidental discharge of a large capacitor through the transistor. High ambient temperatures adversely affect the performance but in normal practice no trouble need be experienced if a few elementary precautions are taken to safeguard the units.

## Transistor Amplifier Characteristics

Fig. 1 shows two amplifier circuits suitable for use with junction transistors. The earthed-base circuit in Fig. 1 (a) resembles an earthed-grid valve amplifier. The input impedance is low (of the order 200 to 500 $\Omega$ ), the output

impedance is very high and there is no phase change between the input and output voltages. The earthed emitter circuit shown in Fig. 1 (b) is similar to an earthed cathode valve amplifier. The input impedance is considerably higher than in the previous case, the power gain is greater and the output voltage developed across the load resistance  $R$  is reversed in phase as compared with the driving voltage. In most cases this circuit is preferable to the earthed base alternative. The third possible circuit, employ-

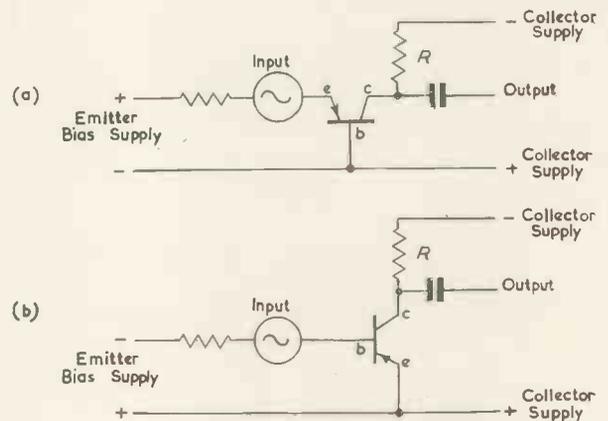


Fig. 1. Pnp junction transistor amplifiers  
(a) Earthed base circuit (b) Earthed emitter circuit

ing an earthed collector, is not of particular interest in the present connexion.

The point-contact transistor is quite satisfactory as an earthed base amplifier, except that the signal-to-noise ratio is rather poor, particularly at low frequencies. Under proper conditions the earthed emitter circuit can also be used successfully but there is a risk of instability due to the fact that in the point-contact transistor there is a current gain greater than unity between emitter and collector circuits. Stability can always be secured, at the expense of power gain, by increasing the external emitter and collector circuit resistances or by reducing the base circuit resistance which effectively reduces the positive feedback.

The power gain of junction transistor amplifiers falls off at frequencies higher than about 20kc/s but in other respects the performance is superior to that of the point contact type. In particular the noise output is low and both amplifiers and oscillators will function with high efficiency at exceedingly low power levels. A typical pnp audio frequency amplifier, deriving collector power and emitter bias from a single supply, is shown in Fig. 2.

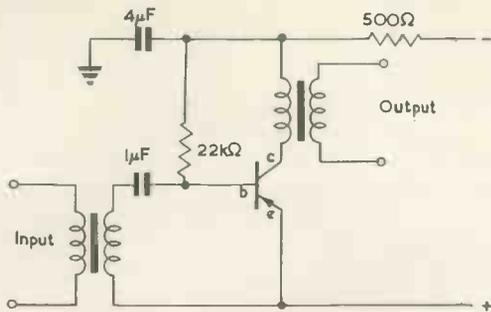


Fig. 2. Earthed emitter amplifier employing pnp transistor

### Transistor Oscillators

Provided that the power gain is large enough any tuned amplifier can be converted into an oscillator by coupling the output back to the input circuit. It is of course essential to arrange that the phase of the feedback voltage is in the correct sense to sustain oscillation. In a valve oscillator the a.c. conditions are satisfied by connecting the grid and anode leads to the opposite ends of the main tuned circuit while the cathode tap is taken to an intermediate point. The ratio of the voltage drop between anode and cathode to that between grid and cathode is equal to the stage gain of the valve and its associated circuit. All tuned-circuit oscillators of the three-terminal type satisfy these conditions and practical arrangements merely require the provision of d.c. feed and grid bias. Similar arguments apply even when a separate reaction or coupled winding is used instead of a tapped tuned circuit.

Transistor oscillators can be developed from similar considerations. Correct a.c. conditions are secured when the collector and base are connected to opposite ends of the tuned circuit while the emitter lead is taken to an intermediate point. A single earth connexion may be made to any point of the system without affecting its operation. In any type of oscillator the waveform is most nearly sinusoidal when the regeneration is set so that oscillations

are barely sustained. The stability of frequency is greatest when the tuned circuit is very loosely coupled to the maintaining system, (valve or transistor).

The oscillator circuits so far discussed are of the conventional three-terminal variety but two-terminal negative resistance valve oscillators can be devised, employing dynatron or transitron circuits. A two-terminal point-contact transistor oscillator of this type is also known. It makes use of the negative resistance which exists between the base and the other electrodes when low (or zero) resistances are externally connected in the emitter and collector circuits.

Three tuned-circuit oscillators employing pnp junction transistors are shown in Fig. 3. Component sizes are given so that the circuits may be adapted for use at other frequencies by making proportional changes in all the element values. The best arrangement is that given in Fig. 3(c), in which the transistor is very loosely coupled to the primary tuned circuit. The observed waveform is extremely good and the power efficiency very high. This circuit works satisfactorily from a single dry cell, under which conditions the total current drain is less than 1mA.

Many similar circuits can be devised, all making use of the same basic principles and all giving a comparable performance.

### Crystal Controlled Oscillators

It has already been remarked that the performance of junction transistor amplifiers falls off at relatively low radio frequencies, but there is no particular difficulty in designing circuits with sufficient gain to ensure reliable working of quartz oscillators up to about 100kc/s, using crystals of high activity. There is little or no published work on transistor drive circuits for use with low frequency flexural mode quartz vibrators using bars or ring crystals operating in the range 1 to 20kc/s. The theory of these vibrators and the design of suitable valve maintaining circuits have already been described in detail<sup>1</sup>. Such arrangements represent the most economical method of generating a series of precisely known audio frequencies without the necessity for many complex stages of frequency division. For example, a 1000c/s ring crystal followed by one conventional decimal divider stage and two simple cold-cathode decade counters could be devised to generate a whole range of low frequencies and to provide one-second time marker pulses. Ring and bar vibrators can only be operated at a low power level without the risk of crystal fracture and this power limitation also applies to transistor operation. It is therefore of interest to develop suitable transistor drive circuits, together with frequency dividers which will also function at extremely low power levels. The electrode arrangement of a three-terminal quartz flexural vibrator is shown pictorially in Fig. 4(a) and schematically in Fig. 4(b). At its resonant frequency the system is represented approximately by a phase-reversing transformer of unity ratio, having a fairly high resistance in series with one of its windings. The equivalent circuit is shown in Fig. 4(c).

A suitable circuit for maintaining the bar in continuous vibration is shown in Fig. 5. It employs a point-contact transistor of which the emitter is earthed. Assuming that the circuit is working, there is an alternating voltage developed across the collector load resistor. This is applied to the crystal causing an attenuated voltage of reversed phase to appear across the output electrodes and also

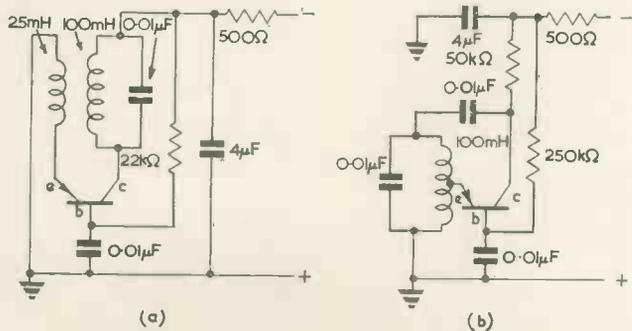
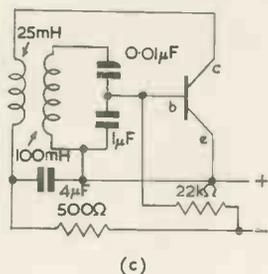


Fig. 3. Pnp junction transistor oscillators

- (a) Series feed earthed base oscillator (5 000c/s)
- (b) Shunt feed earthed base oscillator (5 250c/s)
- (c) Earthed emitter oscillator (5 000c/s)



across the main tuned circuit. A secondary winding provides the transistor driving voltage without further phase change. Self bias is provided by the RC combination in the base lead. By careful selection of the collector load resistance it is possible to raise the base circuit input impedance but there is a risk of instability if this becomes negative. In this case uncontrolled oscillations occur, the frequency of which depends on the LC product of the tuned circuit, and is unaffected by the presence of the

current and voltage unless the remaining network elements are of abnormally high impedance.

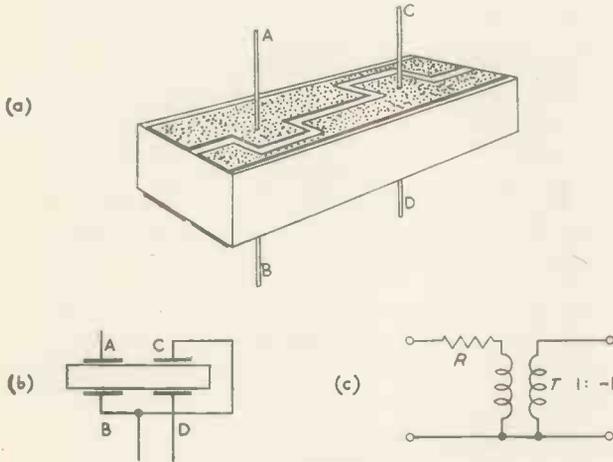


Fig. 4. Low frequency quartz vibrators  
 (a) Electrode arrangement of 3-terminal quartz flexural vibrator  
 A and D: Free terminals  
 B and C: Cross connected  
 (b) Schematic diagram of 3-terminal vibrator  
 (c) Simplified equivalent circuit at resonance

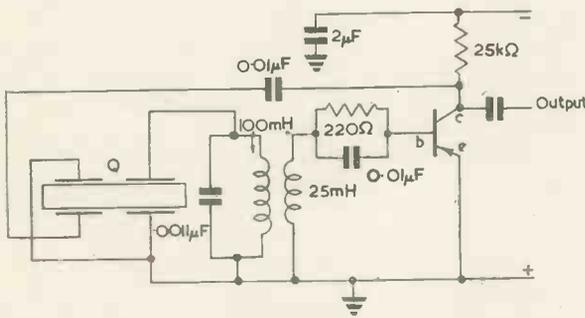


Fig. 5. Point-contact transistor oscillator with 15kc/s quartz vibrator

crystal. In practice, a fair amount of regeneration can be applied before instability is observed. Circuits employing positive feedback are not strictly designable and they are rightly suspect. Unfortunately the margin of reserve power gain of currently available transistors is so low that it is not always possible to avoid these subterfuges.

A circuit employing a junction transistor is shown in Fig. 6. It is somewhat difficult to set up because of the complex impedance matching and tuning adjustments which have to be made but its performance is otherwise superior to that of the arrangement shown in Fig. 5. If the two free electrodes of a three-terminal vibrator are coupled together the resulting two-terminal element can be used in any conventional crystal oscillator circuit. At its series resonant frequency the vibrator is electrically equivalent to a resistance of several thousand ohms. When used as a circuit element in a network it is evident that the vibrator will cause severe attenuation of output

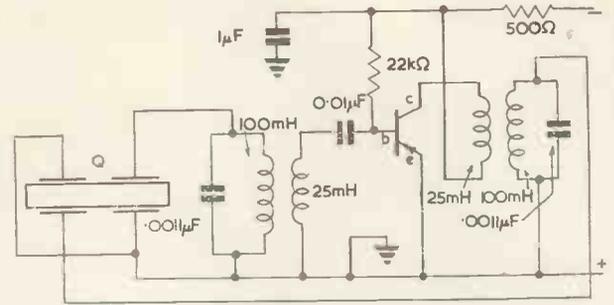


Fig. 6. Junction transistor oscillator with 15kc/s quartz vibrator

### Blocking Oscillators

Point contact transistors which give a current multiplication factor of two or more have been employed in a wide variety of counter and divider circuits<sup>2,3</sup>. Blocking oscillators employing junction transistors have received much less attention. Experience shows that they can be employed in simple circuits to give reliable operation with moderately high division ratios and at very low power levels. They are thus well suited for operation in conjunction with crystal controlled transistor oscillators.

One of the simplest circuits is given in Fig. 7. As shown, the arrangement is a self-running oscillator which produces a train of recurrent pulses the repetition frequency of which is primarily settled by the RC product in the base circuit. The pulses can readily be locked to an external signal at a multiple or sub-multiple frequency. Division ratios of about 15 are practicable, but for the highest stability the factor should be 5 or less. The damping diode connected across the primary winding of the transformer serves to prevent ringing and ensures that only a single pulse is produced during each cycle of operation. Trigger-

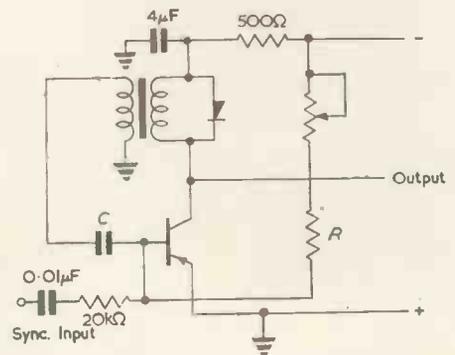


Fig. 7. Blocking oscillator using junction transistor

ing is easily achieved with either pulse or sine wave inputs. The transformer characteristics are in no way critical. The primary (collector circuit) winding should have an inductance of one or two henries for use at a frequency of 1kc/s. The secondary (emitter circuit) winding should be of rather lower inductance, preferably corresponding to a turns ratio of about 2:1. The winding resistances are unimportant.

The free-running frequency may be adjusted in steps

by switching different capacitors into the base circuit. Fine adjustment is made by means of the continuously variable resistor. Provision is made for injecting a locking signal into the emitter circuit. The component values shown are suitable for coupling the synchronizing terminal direct to the output terminal of the oscillator shown in Fig. 5.

The circuit of Fig. 7 works well at 5kc/s when drawing a total current of 200 $\mu$ A from a 2V collector supply. When the blocking oscillator is triggered at a sub-multiple frequency the output pulses appear in groups separated by definite spaces. It is easy to vary the spacing between pulse groups and to change the number of pulses in each group by altering the fundamental frequency of the blocking oscillator and by varying the sub-multiple locking frequency. This circuit is useful as a synchronous gating signal generator for periodically selecting any specified number of pulses from a stream of impulses such as might be employed in an electronic computer or communication system.

### Conclusion

From the limited amount of experimental work

described above, it appears that it should be possible to produce commercially a small battery-operated secondary standard of frequency consisting of a crystal oscillator and two or three blocking oscillator dividers, making available a wide range of precisely known audio frequencies. The range could be extended downwards by adding one or more cold-cathode counter tubes without reducing the reliability of the equipment, although at the expense of introducing circuit complications.

The performance of transistors is adversely affected by high temperatures and this disability tends to restrict their field of application. The power gain of currently available types is rather less than is desirable. It would be possible to simplify a number of the circuits described above if they could be improved in this respect.

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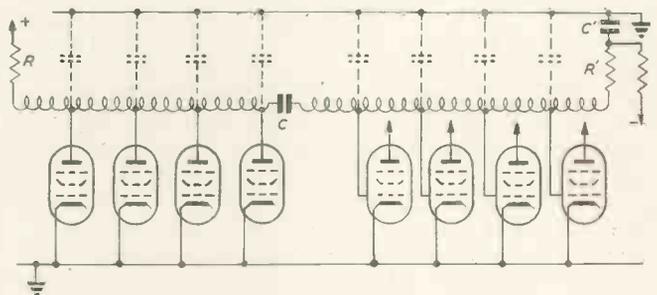
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## A Coupling Circuit for Distributed Amplifiers\*

ALTHOUGH distributed amplifiers were first proposed by W. S. Percival as long ago as 1935 it is only recently that they have been put to any considerable use. These circuits, as is well known, possess the virtue of providing a high degree of amplification over very wide bandwidths. As an illustration of this valuable property they can for instance be used with effect over frequency ranges extending from a few cycles per second up to say 200Mc/s.

In applications of this kind there is inevitably some problem in how to couple one such amplifier to another; and it does not require much imagination to realize that in so extensive a range a coupling arranged to operate satisfactorily for the high frequencies of the band may well be inadequate for the lower frequencies; whereas another circuit which correctly transfers the lower frequencies may not transfer the high frequency components. Simple capacitance couplings tend to fail in both respects. As a result practical coupling circuits have been of rather special kind, and several proposals have been based on the well known constant resistance circuit. A circuit of a very much simpler form has, however, been put forward by Yu, Kallmann and Christaldi† and this is illustrated in the diagram. On the left of the diagram there is shown the anode line of one distributed amplifier, which is coupled to drive the grid line of a further distributed amplifier shown on the right of the diagram. As will be seen the two lines are connected simply by a capacitor  $C$  and as is normal the uncoupled end of the anode line is terminated in a resistor  $R$  equal to the characteristic impedance of the line. The remote end of the grid line is closed by a resistor  $R'$  connected in series with a capacitor  $C'$ . The junction of these two elements is connected through a leak resistor to a bias source so that the correct bias is applied to the control grids of the valves of the driven amplifier and this resistor is made of large value compared with that of the resistor  $R$ . Further, the characteristic impedances of the two lines are arranged to be equal and in order to give a flat frequency response over the whole range the magnitudes of the resistor  $R$  and  $R'$  are chosen to be identical; likewise the capacitances of the capacitors  $C$  and  $C'$ . For high frequencies the reactance of the capacitors  $C$  and  $C'$  is very small and the anode line is connected in effect directly to the following grid line, the latter line being terminated correctly

in its characteristic impedance. For low frequencies such that the reactance of the capacitors  $C$  and  $C'$  cannot be regarded as negligible, the inductances of the lines act virtually as direct connexions and the shunt capacitances may be neglected. The arrangement then reduces to a driving amplifier of internal impedance equal to the characteristic impedance of the line in series with the reactance of the capacitor  $C$ , this amplifier operating into a load of identical impedance. The transfer between the two circuits is thus unity. In the cross-over range between the high and low frequency bands the response remains uniform.



Coupling of distributed amplifiers.

By an interesting departure which entails only a modification of circuit parameters, the arrangement may be made to function with a greater gain than is possible in the form just described. This follows from the fact that the total stray capacitance associated with the anodes of the valves driving the anode line tends in practice to be less than the total stray capacitances associated with the control grids of the control grid line. The anode line may therefore be operated with a higher characteristic impedance giving greater amplification. In these circumstances of course the simple choice of values advocated by Yu, Kallmann and Christaldi cannot be used but a flat response is obtained if the time-constant of the capacitor  $C$  and the resistor  $R$  is made equal to the time-constant of the capacitor  $C'$  and the resistor  $R'$ . It will be realized that operating the circuit with the two lines of unequal impedance must give rise to a reflexion at the higher frequencies from the input of the grid line back along the anode line. This reflexion, however, is of no consequence since it becomes absorbed by the terminating resistor of the anode line.

\* Communication from E.M.I. Ltd.

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# A Cascode Amplifier Degenerative Stabilizer

By V. H. Attree\*, B.Sc.

*A modified cascode amplifier has a gain of over a thousand times. A degenerative stabilizer using the cascode has a much better performance than a conventional stabilizer having a pentode shunt amplifier. The circuit is simple and does not require compensation from the unstabilized side.*

THE performance of a degenerative stabilizer is largely dependent on the gain available in the shunt amplifier. For instance, the output impedance  $Z_o$  of such a stabilizer is given by the approximate expression  $Z_o = 1/g_m\alpha$ , where  $g_m$  is the mutual conductance of the series valve and  $\alpha$  the effective shunt amplifier gain. In a typical circuit<sup>1</sup> using a pentode shunt amplifier  $g_m = 5\text{mA/V}$  and  $\alpha = 100$  giving an output impedance  $Z_o$  of  $2\Omega$ . This value of output impedance is about the best that can be achieved with a single pentode shunt amplifier. With stabilizers of this type a high gain cannot be obtained with the anode load of the shunt valve returned to the stabilized side of the supply, and it is usual to run the shunt valve from the unstabilized side. This gives rise to undesirable feedback at low frequencies and causes the effective gain to be much smaller, especially at frequencies around 10 to 15c/s, where the impedance of the unstabilized side is at a maximum. This reduction in gain increases the stabilizer output impedance at low frequencies; for instance, in the circuit previously mentioned<sup>1</sup> the impedance rises from  $2\Omega$  at 100c/s to a maximum of  $30\Omega$  at 12.5c/s. A not less important effect of running the shunt valve from the unstabilized side is to increase the effect of mains voltage changes on the output.

There are two main ways of overcoming the reduction in effective gain. The first is to provide a separate neon stabilizer for the anode of the shunt valve; this introduces another source of noise into the circuit and may give rise to striking-voltage difficulties. The second method, sometimes known as "forward feedback", is to inject a fraction of the unstabilized voltage into the shunt amplifier circuit. The use of compensation by forward feedback was first described by Miller<sup>2</sup> and a number of different circuit arrangements have been given in the literature<sup>1-7</sup>. The fraction of the unstabilized voltage required for compensation is approximately equal to the reciprocal of the shunt amplifier gain (i.e.  $1/\alpha$ ). Unfortunately the gain is a function both of the frequency and the load current, so that exact compensation is only possible at one frequency (which can be d.c.) and at one load current. A practical difficulty with forward feedback is that the optimum setting for the feedback potentiometer can be found only by modulating at a low frequency either the load or the power supply voltage: adjusting the control for minimum hum at the output, as is the usual practice, does not necessarily give the correct setting.

A very effective way out of these difficulties is to use a modified cascode shunt amplifier. This will give a gain of over 1 000 and will run directly from the stabilized side of the supply. Thus, not only is the performance much better, but forward feedback is no longer necessary, which eliminates one variable control.

The cascode arrangement is no more complex than the pentode and, in fact, from the component point of view,

the only difference is that a double-triode valve is required in place of the pentode, the number of resistors being the same.

## Cascode Amplifier

The basic cascode amplifier is shown in Fig. 1. The input signal is fed to  $V_1$  grid and the output is taken from  $V_2$  anode. The circuit was first given by Hunt and Hickman<sup>8</sup> and subsequently it has been discussed by a number of other writers<sup>9-15</sup>. The complete gain equation for the circuit is given below and for the present we will use the

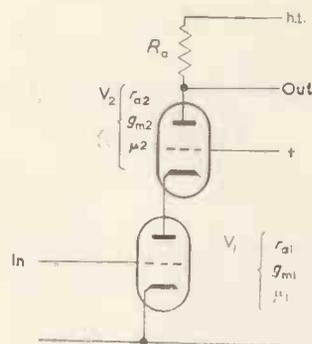


Fig. 1. Basic cascode circuit

approximation  $A = g_{m1}R_a$ , where  $A$  is the gain,  $g_{m1}$  the mutual conductance of  $V_1$ , and  $R_a$  the anode load. From this expression it would appear that the gain of the simple cascode would increase linearly with the value of the anode load. However, in a practical amplifier the supply voltage is fixed and an increase in the load results in a decrease in the anode current at which the cascode can be run. The reduced anode current results in a smaller value for the mutual conductance  $g_{m1}$ , as the slope of a triode is a function of the anode current. This point is illustrated by some measurements on a 12AX7 valve given in Fig. 2. The measurements were taken by a bridge method with a very small grid signal ( $<1\text{mV}$ ) and at an anode potential of 70V. Similar curves, but with somewhat lower values of  $g_{m1}$ , are obtained with other double triodes, such as the 12AT7 and 12AU7, operating in the low-current region. Both scales are logarithmic and a linear relation between anode current and mutual conductance would be represented by a line of unit gradient. It will be seen that for a range of currents of 5 to  $500\mu\text{A}$  the gradient is approximately unity and the relation may therefore be taken as being linear. Thus we may replace  $g_{m1}$  by  $Ki_a$ , where  $K$  is a constant and  $i_a$  the anode current, and the simplified expression for the gain of the cascode becomes  $Ki_aR_a$ . When the cascode is used as a shunt amplifier in a voltage stabilizer the important property is the gain obtainable for a particular voltage drop across the anode load: on this

\* Dept. of Electrical Engineering, College of Technology, Manchester

basis the cascode gain  $Ki_a R_a$  is independent of the load resistance.

The gain limitation due to the change of valve parameters with anode current may be overcome by running the lower valve  $V_1$  at a much higher anode current than the upper valve  $V_2$ . In this way the mutual conductance  $g_{m1}$  can be maintained at a high value irrespective of the value of  $R_a$  in  $V_2$ . The additional current for  $V_1$  is obtained by a resistor connected between the h.t. line and  $V_1$  anode. The impedance at the junction of  $V_1$  anode and  $V_2$  cathode is relatively low and is virtually unaffected by the added resistor.

**The Modified Cascode**

Although the very simple treatment given above leads to conclusions that are essentially correct, it is as well to examine the circuit in a little more detail in order to evaluate the magnitude of the approximations involved.

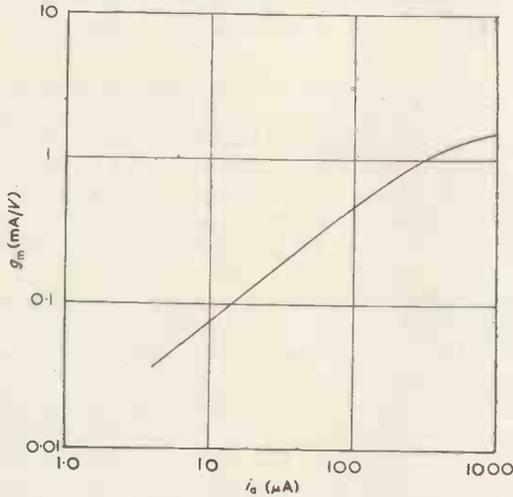


Fig. 2. Relation between mutual conductance and anode current for 12AX7 at  $V_a = 70V$

The exact expression for the gain of the cascode in Fig. 1 is:

$$A = \frac{\mu_1(\mu_2 + 1)R_a}{R_a + r_{a2} + r_{a1}(\mu_2 + 1)} \dots \dots \dots (1)$$

for equal valves this reduces to:

$$A = \frac{\mu(\mu + 1)R_a}{R_a + r_a(\mu + 2)} \dots \dots \dots (2)$$

However, in the modified cascode  $r_{a1}$  is not equal to  $r_{a2}$  and it is necessary to proceed as follows: For a triode valve  $\mu$  is substantially independent of anode current and both  $g_m$  and  $r_a$  are functions of current. The product  $g_m r_a$ , which is, of course, equal to  $\mu$ , remains constant. The valves  $V_1$  and  $V_2$  are both halves of the same double triode, hence  $\mu_1 = \mu_2$ . Further, if we take a 12AX7 valve  $\mu$  is 100 and we may write  $\mu$  for  $(\mu + 1)$  in equation (1) with no serious error. Hence equation (1) becomes

$$A \approx \frac{\mu^2 R_a}{R_a + r_{a2} + \mu r_{a1}} \dots \dots \dots (3)$$

dividing by  $\mu r_{a1}$

$$A \approx \frac{g_{m1} R_a}{1 + (R_a + r_{a2})/\mu r_{a1}} \dots \dots \dots (4)$$

If the lower valve  $V_1$  operates under nearly constant con-

ditions with an anode current of  $700\mu A$  its mutual conductance is  $1.4mA/V$  (Fig. 2) and  $r_{a1}$  is  $70k\Omega$ . The term  $\mu r_{a1}$  in the denominator is therefore  $7M\Omega$ . Thus if the gain is to be given by the numerator of equation (4) then  $(R_a + r_{a2}) \ll 7M\Omega$ . If  $R_a$  is  $1M\Omega$  and the anode potential drop is  $20V$  then the anode current is  $20\mu A$  giving  $g_{m2} = 0.1mA/V$  (Fig. 2) and  $r_{a2} = 1M\Omega$ . Hence  $(R_a + r_{a2})$  is  $2M\Omega$ , and the actual gain is 20 per cent less than  $g_{m1} R_a$ .

**Application to a Degenerative Stabilizer**

In a practical stabilizer circuit the gain is reduced by the effect of the reference neon in the cathode circuit of

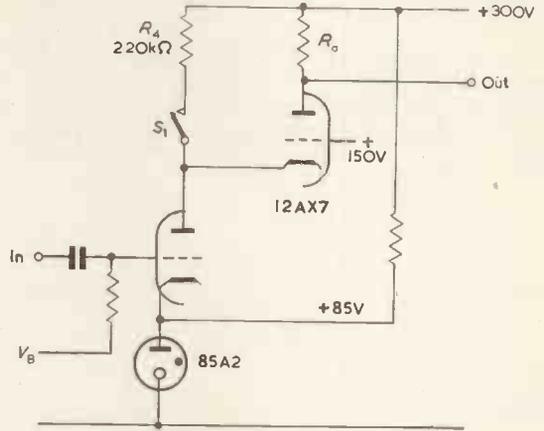


Fig. 3. Modified cascode circuit

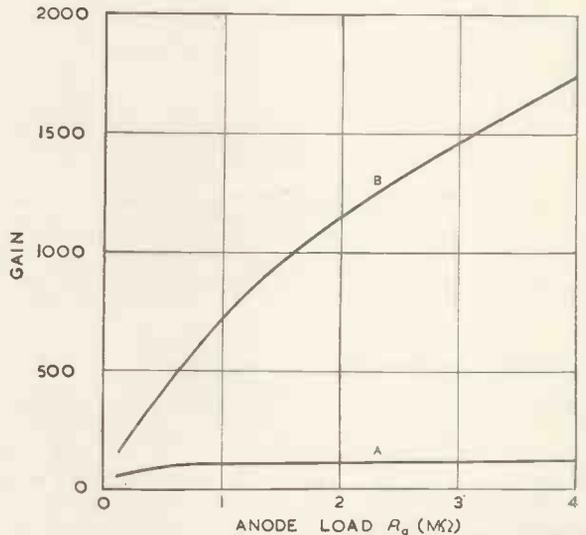


Fig. 4. Relation between anode load and gain of modified cascode with 20V d.c. drop on load  
Curve A,  $S_1$  open; curve B,  $S_1$  closed.

the lower valve. In order to measure the gain realized under working conditions the arrangement of Fig. 3 is used. The switch  $S_1$  permits the  $220k\Omega$  resistor  $R_a$  to be brought into circuit. The additional current supplied via  $R_a$  to the lower valve is  $670\mu A$ . The bias voltage  $V_B$  is adjusted to give the desired potential drop across  $R_a$  and gain measurements are made at a frequency of  $500c/s$ . The curves of Fig. 4 are made for a fixed potential drop of  $20V$  across the anode load. With  $S_1$  open and for anode loads of  $200k\Omega$  to  $4M\Omega$  the gain remains very nearly constant at

about 100 times (curve A), and with  $S_1$  closed the gain increases with load until at  $4M\Omega$  it is 1 700 times (curve B). The second set of curves, shown in Fig. 5, illustrate the relation between gain and potential drop across a fixed anode load of  $2M\Omega$ . With  $S_1$  open (curve c) the gain increases linearly with potential drop, as expected from the simple theory, while with  $S_1$  closed (curve D) the gain is 1 000 for a 10V p.d. improving to 1 500 at 100V p.d.

From these measurements it is evident that, even when taking the effect of the neon reference tube into account, the gain of the modified cascode will be at least 1 000 times for a potential drop of 10 to 100V across the load. This corresponds to anode currents of 5 to  $50\mu A$  in the upper valve.

A high gain at low anode currents can also be obtained by the so-called "starvation" circuits<sup>16,17</sup>, but these have less gain than the cascode and are reported to have poor stability<sup>18</sup>. Starvation circuits require much higher anode

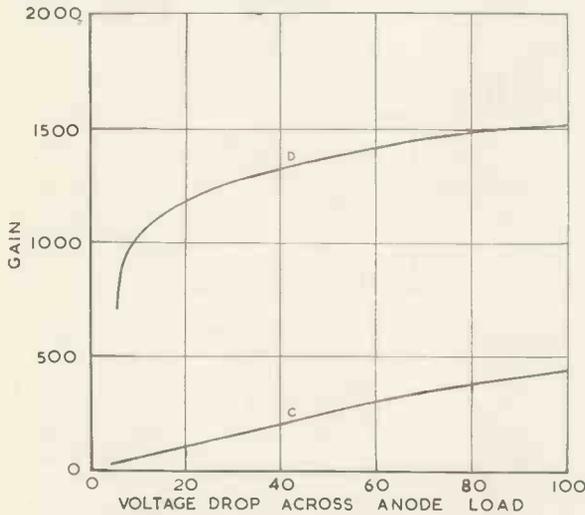


Fig. 5. Relation between voltage drop on anode load and gain of modified cascode with  $2M\Omega$  load

Curve c,  $S_1$  open; curve d,  $S_1$  closed.

loads than the cascode and have therefore an inferior frequency response.

### Cascode-Amplifier Degenerative Stabilizer

The circuit of a complete stabilizer is shown in Fig. 6. The gain reduction due to the neon impedance is avoided at medium and high frequencies by the  $0.1\mu F$  decoupling capacitor  $C_1$ . The  $100\Omega$  resistor  $R_6$  is necessary to ensure that there is no risk of relaxation oscillations between the capacitor and possible negative resistance of the neon tube. Unfortunately the added resistance increases the impedance in the cathode circuit of the lower triode at very low frequencies and d.c. The capacitor  $C_1$  also helps to reduce the amount of noise from the neon tube. The cascode anode load  $R_5$  is  $2.2M\Omega$  so that the gain characteristics are essentially as given by curve D in Fig. 5. The total capacitance at the grid of the 12E1 is  $32pF$  and the frequency response of the cascode is  $3dB$  down at  $2.5kc/s$ . A low output impedance at higher frequencies is maintained by the large capacitor  $C_3$  shunted across the output. Because of the high amplifier gain the cascode grid is susceptible to hum pick-up. Hum troubles are largely avoided by making the impedance of the voltage divider  $R_{7-10}$  fairly low and by locating the control potentiometer  $R_9$  close to the

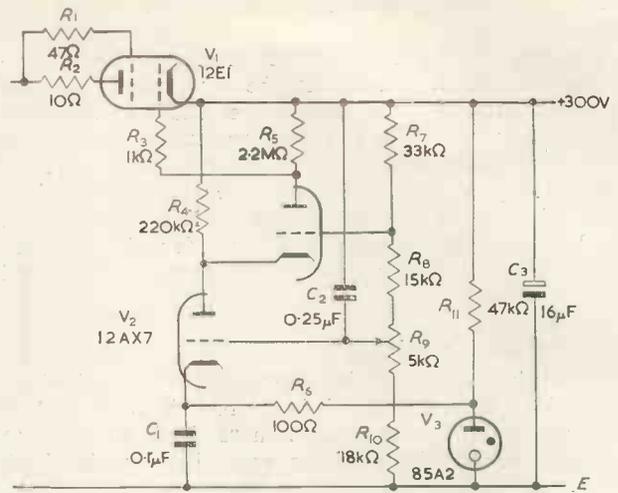
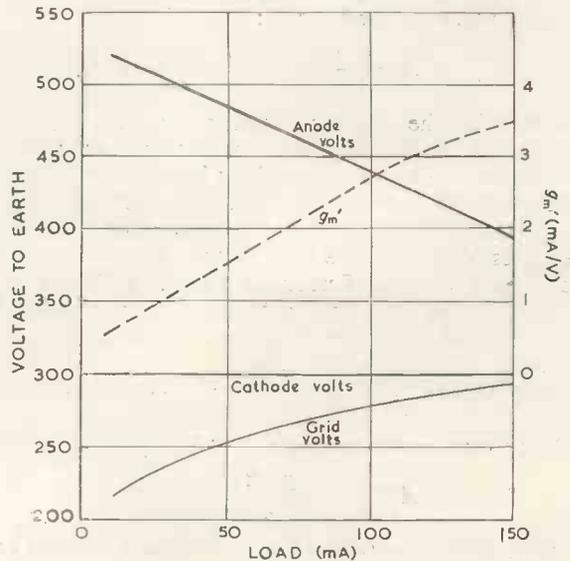


Fig. 6. Cascode-amplifier degenerative stabilizer

valve grid pin. The  $0.25\mu F$  by-pass capacitor improves the action of the stabilizer by eliminating the attenuation of the voltage divider at medium and high frequencies. With careful layout the total hum and noise level at the output is 100 to  $200\mu V$  r.m.s.; the exact value depends mainly on stray voltages induced from heater and mains wiring rather than on the hum level at the unstabilized side which is 2 to 5V.

As in the case of the shunt stage the valve parameters of the series valve change very considerably over the working range. The actual voltages on the 12E1 series valve  $V_1$ , are shown in Fig. 7. The voltage at the anode falls from 525V at 10mA load to 390V at 150mA and the corresponding grid voltage change is from 210V to 295V corresponding to a potential drop across the anode load of the cascode of 90V down to 5V. These are typical values for a stabilizer, although with a smaller series valve, such as the EL81, the grid base will be shorter. The effective value of the mutual conductance  $g_m'$  is found by differentiating the grid voltage curve with respect to load current. The value of  $g_m'$  changes from  $0.5mA/V$  at 10mA load to  $3.5mA/V$  at 150mA load.

Fig. 7. Characteristics of 12E1 series valve in Fig. 6



The large range of effective slope shows why forward feedback in a conventional stabilizer can be made to give exact compensation only over a narrow range of load current.

The output impedance of the stabilizer has been measured at several values of load current and for the frequency range 0 to 50kc/s. The impedance is in general about  $0.2\Omega$  increasing to  $0.5\Omega$  at d.c. where the capacitors  $C_1$  and  $C_2$  are ineffective.

The impedance rises at frequencies around 30kc/s and as the reasons for this are somewhat complex the matter is discussed in the next paragraph.

### High-frequency Performance of Cascode Stabilizer

In degenerative stabilizers having a large shunt amplifier gain high-frequency instability troubles are often encountered. The difficulty arises because there is more than one effective time-constant controlling the rate of gain cut-off at high frequency. In the case of the cascode the high-frequency cut-off is dominated by the time-constant at the grid of the 12E1 series valve and oscillation troubles are absent. However, there is sufficient phase-change elsewhere in the circuit to cause a rise in the output impedance at a frequency of 30 to 50kc/s. The magnitude of this impedance rise may be reduced either by a phase-advance network to the cascode grid or by introducing a loss component in the output capacitor ( $C_3$  in Fig. 6). The first method has the disadvantage that it entails a reduction in the gain at low frequencies and increases the danger of hum pick-up on the grid. The second method is much more convenient as the unwanted rise in output impedance can be almost completely suppressed by including a resistance of the order of  $1\Omega$  in series with the output capacitor.

It has been found that the necessary component is already present in electrolytic capacitors. The loss resistance of a number of different types of  $16\mu\text{F}$  electrolytic capacitors has been determined on a Q bridge. The measurements were made at 50kc/s with an accuracy of about  $0.1\Omega$ . In every case the loss is between  $0.7$  and  $1.5\Omega$  and magnitude of the rise in output impedance, obtained with the capacitor on the stabilizer output, does not exceed  $1.7\Omega$ .

### Balanced Cascode Circuit

One of the objections to the circuit of Fig. 6 is that heater voltage changes in the 12AX7,  $V_2$ ; alter the contact potential in the lower valve of the cascode and hence cause a change in the output voltage. For a  $\pm 10$  per cent change in the mains supply the output alters by  $\pm 0.4\text{V}$  which corresponds to a change in contact potential of  $110\text{mV}$  at the cascode grid. The difficulty may be overcome by using the circuit of Fig. 8. The neon potential is smoothed by the long time-constant filter  $R_{12}$  and  $C_1$  and appears on the grid of  $V_{3b}$ , while the voltage divider  $R_{7-10}$  feeds the grid of  $V_{3a}$ . In this way variations in contact potential are balanced out in the long-tailed pair  $V_{3a}$ ,  $V_{3b}$ . The upper valve of the cascode is  $V_{2b}$  and the triode section  $V_{2a}$  is used as a cathode-follower to drive the grid of the series valve. This improves the frequency response and is particularly useful when two or more series valves are used in parallel connexion in order to increase the available load current. The anode supply for the cathode-follower is taken from the unstabilized side of the supply; voltage variations at this point have negligible effect on the performance as the gain up to the cathode-follower grid is very high—of the order of 1 000 times.

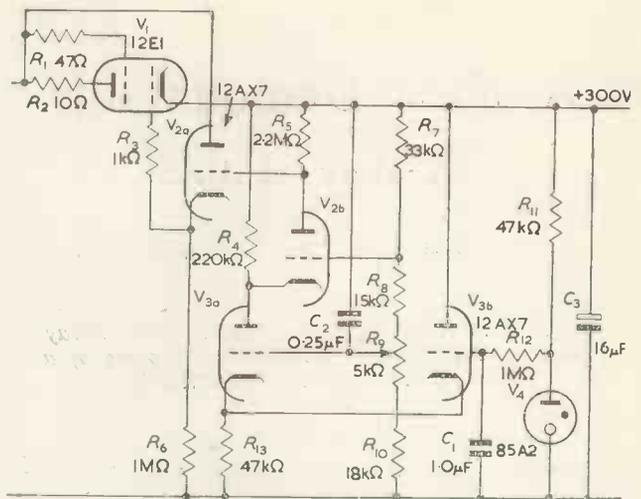


Fig. 8. Degenerative stabilizer with balanced cascode amplifier

With the balanced circuit a  $\pm 10$  per cent change in the mains supply alters the h.t. line by  $\pm 70\text{mV}$ ; of this output change  $\pm 30\text{mV}$  is due to the change in the voltage at the anode of the series valve and  $\pm 40\text{mV}$  is due to residual unbalance in the long-tailed pair  $V_{3a}$ ,  $V_{3b}$ . Contact potential variations in  $V_{2a}$ ,  $V_{2b}$  and the series valve  $V_1$  have practically no effect.

### Conclusion

In spite of the high gain the cascode presents no special difficulties of layout or stability and, at the time of writing some twelve power units have been constructed in this laboratory, mostly with the circuit of Fig. 6. These units all had hum levels of better than  $500\mu\text{V}$  over the full load current range, and output impedances of about  $0.2\Omega$  at 500c/s.

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# A New Demountable X-Ray Tube

By J. A. Darbyshire\*, M.Sc., Ph.D., F.Inst.P.

*The construction of a new demountable X-ray tube having a water cooled target capable of a dissipation of 500W is described.*

SOME time ago a small sealed-off X-ray tube commercially known as the B100 was introduced by Ferranti Ltd.

This tube is shown in Fig. 1 and its electrical characteristics are as follows:

Filament voltage	2.5V
Filament current	5.5A
Target voltage (max)	40kV
Screen voltage	0 to 1.4kV
Spot size (approx.)	
(at 40kV and 100 $\mu$ A)	0.6mm
Max. target dissipation (continuous)	25W
Max. target dissipation (for 20 seconds)	500W

In response to requests for a tube of higher power the demountable type of X-ray tube shown in Fig. 2 has been developed recently.

This tube is designed so that it can be very easily dismantled and it is possible to clean or renew the target and fit a new filament and window in a very short time.

The main constructional features are as follows:

The end plates are made of brass and the vacuum joints between the end plates and the main glass of the tube are made by means of "O" rings.

These "O" rings are compressed between the two-component flanges which comprise the end plates of the system.

A similar technique is used to attach the window to the side tube. This window is made of polythene and is therefore very transparent to low voltage X-rays. A window of beryllium or aluminium or any other material could easily be fitted if desired.

The filament is constructed from tungsten wire and it is surrounded by a shield which can be used as a modulating electrode, and to some extent, as a focusing electrode. The target is water cooled and is capable of a dissipation of 500W.

\* Ferranti, Ltd.

For X-ray diffraction the target is designed to be end-on to the beam and the X-ray output can be taken from the window on the side tube.

If required, this diffraction tube can be provided with two windows, one on either side of the tube. For radiographic work the target is inclined at an angle of 67° to the beam so that a large uniform field of radiation is available.

The target is usually made of copper, but the end plate can be removed very easily and a target of another material such as nickel, cobalt or chromium can be inserted when it is desirable to use characteristic radiation of different wavelengths.

The tube is evacuated by means of any conventional high vacuum system. For example, a rotary oil pump type 1S50

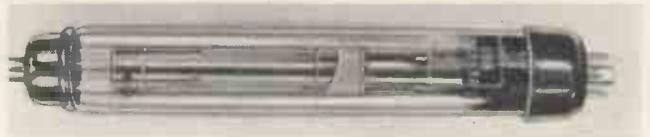


Fig. 1. The B100 X-ray tube

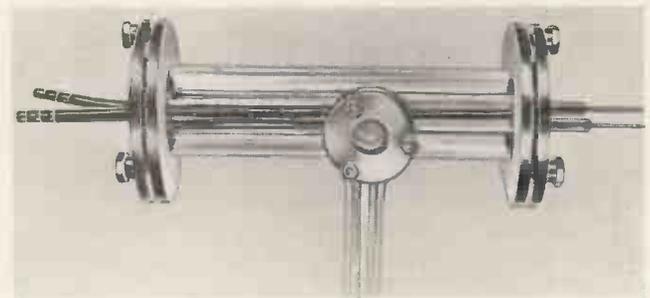


Fig. 2. The higher power sealed off X-ray tube

followed by a silicone diffusion pump of type 102 would be quite satisfactory.

The tube can be operated at any anode voltage up to 45kV and at any beam current such that the target dissipation of 500W is not exceeded. The beam current remains very steady during the operation of the tube when it is even as high as 15mA.

It has been found that very satisfactory X-ray photographs can be obtained when the target voltage is no higher than 25kV.

Fig. 3 is an illustration of a photograph of pure Al<sub>2</sub>O<sub>3</sub> taken using a copper target. The tube in this instance was operated from a half-wave rectified supply of 25kV at a beam current of 15mA and the exposure was three hours.

These X-ray tubes have been developed in the Physical Laboratory at Moston by Mr. R. D. Phillips and Mr. D. M. Taylor and Mr. E. Thomas.

Fig. 3. Reproduction of photograph of pure Al<sub>2</sub>O<sub>3</sub> using a copper target

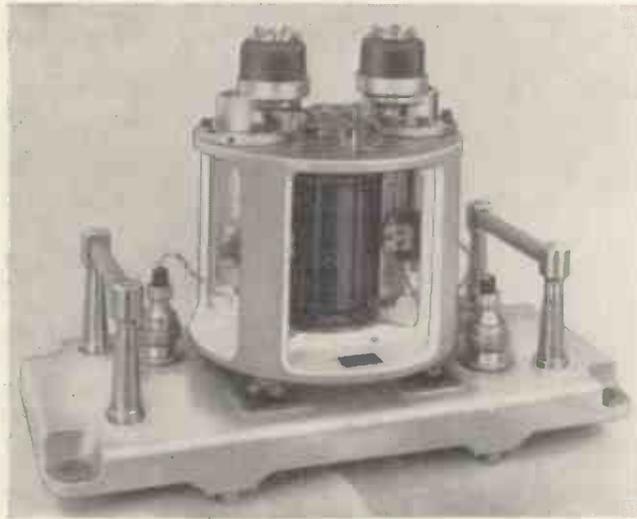


# DEUCE

(A New Digital Computer)

**T**HE English Electric Company have now developed a new digital electronic computing machine which is in commercial production.

This machine is the Digital Electronic Universal Calculating Engine (DEUCE). It is the outcome of years of intensive development of ideas first put forward shortly after the war at the National Physical Laboratory at Teddington. Sir George Nelson, Director of the English Electric Company, was approached at that time by Sir Charles Darwin of the N.P.L. with a request that the English Electric Company should assist the scientists at N.P.L. with the development of an Automatic Computing Engine, a machine of vast potential whose logical

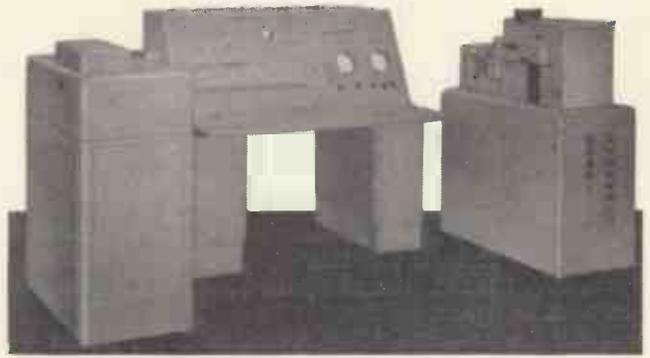


*The magnetic recording drum*

design had already been worked out at N.P.L. The company provided a team of engineers and craftsmen to work at Teddington and as a result a Pilot Model ACE was brought into successful operation in the Mathematics Division at N.P.L. in February 1952. This pilot model proved to be an outstanding success in solving problems for many branches of Government and industry. The experience gained with this early machine was of great help in determining the next stages of development. The English Electric Company decided to produce a fully engineered version of this machine embodying all the improvements which were suggested by the extensive operational experience at the N.P.L.

In the DEUCE, numbers and instructions are represented by sequences of pulses at intervals of 1 microsecond. A train (or "word") of 32 such pulses is used to represent either a number (equivalent to 9 decimal digits with sign) or an instruction. Two forms of storage are provided. The first is intimately associated with the computing circuits and includes 12 mercury delay lines each storing 32 words together with 10 shorter lines of various lengths which are used as accumulating registers, etc. The second store of large capacity (8 192 words—or more than a quarter of a million digits) is provided by a magnetic recording drum. Information can be written on or read from 256 tracks by means of two sets of 16 heads, either of which can be moved into any of 16 positions by a high speed shifting mechanism giving access to any track in 25 milliseconds.

Input and output of information is on 80 column punched



*The control desk and input and output mechanisms*

cards of standard type. Use of a card punch instead of a printer on the output enables the maximum speed of the computer to be utilized; subsequent sorting and printing can then be carried out on standard machinery without occupying valuable computer time.

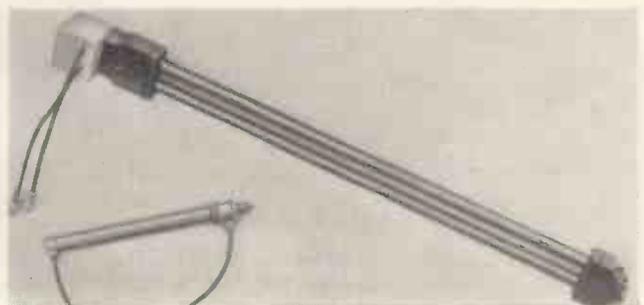
Each input card can carry 12 instructions or up to 32 decimal digits. The latter can consist, for example, of 3 9-digit numbers (each with sign) or several smaller numbers. Input of information in sterling or other non-decimal form is also possible. The computer can convert these numbers into binary form and send them to its storage system when cards are being read at the maximum speed of 200 per minute. The reverse process is carried out at the output, the speed in this case being 100 cards per minute, corresponding to about 50 decimal digits per second.

The physical form of the machine demonstrates its smallness compared with other machines of its scope and speed of operation. Although this overall size is conveniently small, this has not been achieved by close packing of components. A generous amount of space has been used and the components are easily accessible for maintenance and fault tracing. The main electronic units and the magnetic store are housed in nine vertical cubicles each containing about eight chassis of convenient size; together with the control desk it occupies a floor area of 14ft by 4ft 6in. Access is by doors to the components on one side of the chassis and by a central corridor to the valves which are all on the inward facing side of each chassis. The total number of valves used is about 1 300, all of standard miniature types and universally available. The mercury delay lines (other than short lines of one, two or four word lengths each) and contained in a thermostatically controlled drum, 3ft in diameter and 3ft 6in high, located a convenient distance away from the main electronic units.

A computing centre has already been set up in the Nelson Research Laboratories of the English Electric Company at Stafford and a further centre will shortly be established in London. A DEUCE computer will be permanently located at each of these centres and a team of mathematicians and operators will be available to offer a full computing service to organizations not requiring a machine for their own full time use.

A training service will be available based on the two computing centres for initiating scientists, mathematicians and machine operators in the actual use of machines supplied to their organization.

*The mercury delay lines*



# Notes from North America

## U.S. Defence Contracts

New regulations, which will make it easier for overseas countries to compete for U.S. Defence Department contracts, have been announced by the Secretary for Defence, Mr. Charles Wilson.

Under the so-called "Buy American Act", passed by Congress during the Depression, U.S. concerns were awarded the contracts even if their bids were as high as 25 per cent above foreign offers. Under the new order, dated February 21, foreign bidders will now be awarded contracts if the lowest acceptable foreign bid is 6 per cent or more below the lowest acceptable American bid.

Another easing of restrictions on foreign firms is contained in a re-definition of "foreign goods". Previously raw materials or manufactured items have been considered foreign goods if 25 per cent or more of the cost of materials went to foreign suppliers. Under the new regulations goods will be considered foreign only if 50 per cent or more of the cost of materials go to non-American sources of supply.

The new regulations continue to give the Secretary of Defence authority to refuse a contract to a foreign bidder if it would be detrimental to the national security or the national interest.

Two new exceptions have been added, based on provisions of the Presidential order issued last December. In certain cases the Secretary of Defence may favour a domestic bidder if it is necessary to do so in order to give a fair proportion of defence contracts to small business concerns, or to place defence contracts in areas of the United States where there is serious unemployment.

## Dielectric Potentiometer

A new voltage divider, which it is claimed is free from phase and frequency distortion over a range of frequencies from 20c/s. to 10Mc/s, is announced by the Technology Instrument Corporation of Acton, Mass.

The dielectric potentiometer is a three-terminal network employing a liquid as the variable element. The effective resistance-capacitance shunt and series elements of this network result from the inherent resistivity and dielectric constant of a lossy liquid dielectric immersing the terminal electrode structures. The output electrode is movable with respect to the input electrode. Since the resistance and capacitance components vary in exactly inverse manner with movement of the output electrode, a constant  $RC$  product is maintained in each of the ratio arms.

The stopless movement through the liquid dielectric provides a smoothly continuous voltage division. The constant  $RC$  product of the two ratio arms assures the same voltage division for any and all frequencies within the band range of the device. Similarly there is no shift of phase as the voltage division ratio is changed. Attenuation range is approximately 25dB.

## Computer Conference

The 1955 annual general meeting of the Association for Computing Machinery—the only one to be held during the

year—will take place at the Moore School of Electrical Engineering, University of Pennsylvania, September 14-16, inclusive. As in the past, this meeting is intended to serve as a place for the reporting of new ideas and developments in the applications of computing machinery.

## Cesium Atomic Frequency Standard

A new atomic frequency standard has recently been developed by Dr. J. R. Zacharias, of the Laboratory for Nuclear Science at the M.I.T.

The atomic clock built primarily for research is expected to have practical uses such as more accurate methods of navigation and studies in astronomy, geophysics and radio communications.

Time keeping in the device is controlled by the oscillation of electrons in the cesium atom.

One one-hundredth of a gram of cesium is enclosed in a metal crucible. Under heat, about that of boiling water, the cesium shoots a stream of atoms through a hole in the crucible. The atoms strike a detector screen and their frequency is reported through complex apparatus.

The cesium emits atoms at a rate of about one million per second, but the loss is only about one-millionth of a gram per day.

While for scientific work, clock hands would not provide sufficiently precise timing, the device can be connected with a conventional but finely calibrated clock face for use in such operations as navigation.

Standard time, with one second equal to  $1/86400$ th of the mean solar day and generally measured by the frequency of the oscillations of a crystal, is accurate to 1 part in  $10^9$ . The Atomic Frequency Standard is accurate to 1 part in  $10^{10}$ . Dr. Zacharias expects, through further development, to obtain accuracy of one part in ten to the twelfth power.

Associated with Dr. Zacharias in the work have been James G. Yates, an English electrical engineer who was a visiting professor at M.I.T. and is now at Trinity College, and Robert D. Haun, a research assistant in the Department of Physics at M.I.T.

*Dr. Zacharias with the cesium clock*



# Short News Items

The Television Society has awarded the following Premiums, of £5 each, for outstanding papers read before the London meetings in 1954. The *Electronic Engineering* Premium to Dr. G. N. Patchett, of Bradford Technical College, for his paper on "Problems of Inter-lacing". The Mullard Premium to Messrs. G. B. Townsend, E. Ribchester, and D. Bauer (G.E.C.) for their paper on "An Investigation of the 625 line C.C.I.R. System". The E.M.I. Premium to Messrs. R. J. Boddy and C. D. Gardner (E.M.I. Research Department) for their paper on "An Industrial Television Channel". A special Premium for the Society to Mr. G. G. Gouriet (BBC Research Department) for his paper on "Colour Television".

The Annual Dinner of the Institution of Electrical Engineers was held on Thursday, 24 February, when the President, Mr. J. Eccles, was in the Chair. The principal guest was The Rt. Hon. Sir Raymond Evershed, Master of the Rolls, who proposed the toast of "The Institution", to which the President replied. The toast of "Our Guests" was proposed by Mr. Harold Bishop, immediate past-president, and the reply was given by The Rt. Hon. Lord Citrine, Chairman, British Electricity Authority.

The Tenth Annual Exhibition organized by the Northern Division of the Institution of Electronics will be held at the College of Technology, Sackville Street, Manchester, from 14-20 July. The sessions will be from 2 p.m. to 10 p.m. on 14 July, 10 a.m. to 10 p.m. on 15, 18, 19 and 20 July, and 10 a.m. to 6 p.m. on 16 July. The exhibition, which is organized for its educational value, will consist of a Manufacturers' Section, a Scientific and Industrial Research Section, and will include a programme of lectures and film shows on subjects allied to electronics. Lecture and film show programmes (price 4½d. post free) will be available in June. Advance catalogues (price 2s. post free) will be on sale early in July. Admission to the exhibition, lectures and film shows will be by tickets obtainable, free of charge, from the Honorary Exhibition Organizing Secretary, Mr. W. Birtwistle, 78 Shaw Road, Rochdale, Lancs. A stamped addressed envelope should be enclosed.

Rocke International Ltd are organizing jointly with Bruel & Kjaer Laboratories Ltd a private exhibition of the latest American and Continental electronic instruments and components, to be held in London from 25 April to 6 May at 59 Union Street, London, S.E.1, between 9 a.m. and 7.30 p.m. Tickets are available on application to the address mentioned.

A Conference on the production and utilization of electromagnetically enriched isotopes will be held at the Atomic Energy Research Establishment, Harwell, from 13-16 September this year. Several leading members of the team responsible for the work on stable isotopes at Oak Ridge, U.S.A., are attending, while provisional acceptances have also been received from leaders of the separator groups in France, Holland and Denmark. Although a detailed programme has not yet been prepared, sessions will be included on design and operation of large mass separators; ion source and collector problems especially with high beam current; separation of radioactive materials; chemical aspects of the production work; preparation of isotopic targets; developments in stable isotopes mass-analysis; utilization of electromagnetically enriched isotopes. Those wishing to attend should write immediately to Dr. M. L. Smith, A.E.R.E. Harwell, Didcot, Berks, indicating also whether they will make a written contribution. Abstracts of papers must be submitted before 30 April and complete manuscripts before 30 June.

La Société des Editions Radio, 9 Rue Jacob, Paris, have produced a new bimonthly journal called *Electronique Industrielle*. It is hoped that this journal will provide useful information for those interested in the electronics field. The price of this new journal is 300 Francs an issue.

The Scientific Advisory Board of the journal *Research* has decided for the third successive year to organize, with the co-operation of *The Sunday Times*, a "Science in Industry" essay competition. The purpose of this competition is to encourage scientists to take greater interest in the problem of presenting the results of scientific research work to people less highly qualified than themselves.

The Electrical Industries Benevolent Association are celebrating their Golden Jubilee by a luncheon in the Grand Hall at the Connaught Rooms on Tuesday, 19 April. The principal guest is Sir John Maud, K.C.B., C.B.E., Permanent Secretary to the Ministry of Fuel and Power.

Marconi's Wireless Telegraph Co Ltd announce that Associated-Rediffusion Ltd, through their technical advisers, Central Rediffusion Services Ltd, have placed a major contract for the provision and planning of most of their studio and outside broadcast equipment with Marconi's. The initial order includes the complete installations for

three three-camera studios, one two-camera studio and two complete three-camera television outside broadcast vehicles. In addition, Marconi's are to supply one set of master control switching and central apparatus equipment, and the equipment for a further television outside broadcast unit, less vehicle. This large order is being completed so that Associated-Rediffusion can be on the air by 1 September.

Work has already been started by Marconi's on the temporary transmitter at Croydon, following upon the ITA announcement of their plans for a television station in the London area.

The Central Office of Information have made an announcement concerning a new free loan section of their film library. The following films are among those available. "The Electron Microscope"; "Electric Induction Heating"; "Ammeters and Voltmeters"; "Capacitors"; "Effects of Inductors and Capacitors on A.C. and Power Factor". Other titles will be added to the new free section in the near future.

The Imperial College of Science and Technology are holding a course of eight lectures by Professor W. D. Wright on "The Science of Colour applied to Colour Television". They will be held on Tuesdays and Thursdays at 4.30 p.m. commencing on Tuesday, 26 April. Application for admission should be made to the Registrar of Imperial College, Prince Consort Road, London, S.W.7. The fee for the course is £2 2s.

The Press Office of the United Kingdom Atomic Energy Authority has moved from Bedford Chambers, Covent Garden. The address of the Chief Press Officer, Mr. Stanley White, is now St. Giles Court, 1-13 St. Giles High Street, London, W.C.2.

**Addendum.** In connexion with the article "Submerged Repeaters for the Aberdeen-Bergen North Sea Cable", which appeared on p. 126 of the March issue, Submarine Cables Ltd have now submitted the following clarification. The repeaters in question were built and tested in the factory of Standard Telephones & Cables Ltd at North Woolwich, as stated, but they were provided with, and sealed into, pressure-resisting housings at the Ocean Works (Erith) of Submarine Cables Ltd. The sealed housings were there subjected to a test pressure of 1½ tons/in<sup>2</sup>, which in view of the relatively shallow water in which they were to be laid was amply sufficient. Test facilities at Ocean Works include for the testing up to 4 tons/in<sup>2</sup> of repeater housings for submersion in ocean depths of the order of 2 000 fathoms.

# R.E.C.M.F. EXHIBITION PREVIEW

A description, compiled from information supplied by the manufacturers, of selected exhibits to be shown at the Exhibition of the Radio and Electronic Component Manufacturers Federation, in Grand Hall, Grosvenor House, Park Lane, London, W.1., from 19 to 21 April.

(Figures in parenthesis refer to Stand Numbers).

## Advance (30)

**I**NCLUDED among a number of new productions on this stand will be the video signal generator type R1 (illustrated below). A feature of this generator, which employs an RC phase shift oscillator, is the wide frequency range covered in each of the two ranges, which are 30c/s to 3Mc/s and 3Mc/s to 10Mc/s, with a generous overlap between ranges. The effective dial length is 27in. Output is from a cathode-follower at an impedance of  $75\Omega$  via a three stage decade attenuator, maximum e.m.f. being IV. A high



impedance output (max. 10V) is also provided. The output is monitored by a crystal voltmeter.

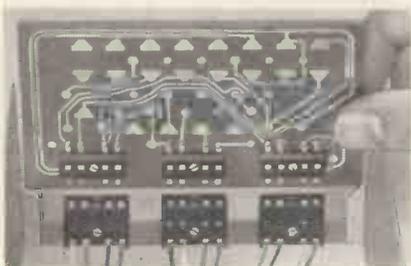
**Advance Components Ltd.,**  
Marlowe Road,  
Walthamstow,  
London, E.17.

## Belling-Lee (16)

**A** NUMBER of new components will be shown on this stand, including aerials and aerial adaptors for Band 3 television, and instrument fuse-links (type L.1 055) manufactured by a new technique which bonds caps, glass and filament into one unit.

In addition a range of plugs and sockets (illustrated below) designed specifically for use with printed circuits, will be shown.

**Belling & Lee Ltd.,**  
Great Cambridge Road,  
Enfield, Middlesex.



## B.I.C.C. (57)

**I**N addition to their normal ranges of cables, winding wires and couplers, B.I.C.C. will, this year, be showing a new range of television download cables insulated with cellular polythene.

**British Insulated Callenders Cables Ltd.,**  
21 Bloomsbury Street,  
London, W.C.1.

## Bulgin (21)

**T**HE vast range of Bulgin switches, plugs and sockets, lampholders, etc., has been still further increased and a number of new items will be shown.

The range of fittings for Acro micro-sensitive switches has been increased and types operating from pressures as low as 0.75oz and 0.01in movement are available. A new sub-miniature type will also be shown.

**A. F. Bulgin & Co. Ltd.,**  
Bye Pass Road,  
Barking, Essex.

## Dawe Instruments (73)

**O**NE of the latest instruments on display on this stand will be the type 1107 Visigauge.

The Visigauge measures thickness from one side by determining the fundamental natural frequency of vibration in the thickness direction.

The natural frequency of thickness vibration is essentially independent of the other physical dimensions of the material.



*NOTE.—As this issue will have gone to press almost a month before the opening of the exhibition, this preview is necessarily incomplete as a number of exhibitors have not yet finalized or released details of their exhibits.*

A vibrating quartz crystal is placed in contact with one side of the material under test so that an ultrasonic wave is transmitted into the material. This wave travels in a narrow beam through the material and is reflected by the opposite surface. At certain frequencies, when the transmitted and reflected waves are in phase, there will be a relatively large increase in the amplitude of the wave in the material. This is a resonance condition occurring at the fundamental frequency and harmonics of the fundamental frequency.

The fundamental resonance frequency is inversely proportional to twice the thickness and directly proportional to the velocity of sound in the material. The velocity of sound is known for most materials and thus the thickness can be determined.

The thickness is indicated on a c.r.t., a calibrated scale for a given material being fitted in front of the tube. Five crystals cover the range of 0.015in to 0.3in for steel.

**Dawe Instruments Ltd.,**  
99 Uxbridge Road,  
London, W.5.

## De La Rue (83)

**I**NCLUDED in this company's exhibits will be a new grade of laminate known as Delaron Copper Clad.

The laminate, faced with copper foil, is now being used in the manufacture of printed circuits, with the circuit drawing transferred to the copper clad by litho, silk screen or photo-engraving techniques. Surplus copper foil, not protected by the acid-resisting inks used in the printing of the circuit, is then etched away, leaving a permanent electrical conductor serving as a connecting medium for many kinds of electrical components.

De La Rue produce three grades, the base laminates conforming to R.C.S.1 000 and B.S.S.1 137 specifications.

**Thomas De La Rue & Co. Ltd.,**  
Imperial House,  
84/6 Regent Street,  
London, W.1.

## Electrothermal Engineering (120)

**T**HE Precistor range of wire wound high stability resistors has been extended to include resistance values up to  $3M\Omega$ . These are available, tapped or untapped, in any value as required. In addition to the high stability types a range of standard wire wound resistors has been introduced.

Of interest too is the Electrothermal Valve Retainer and Valve Data Manual, which details the correct types of retainer for any given valve.

**Electrothermal Engineering Ltd.,**  
270 Neville Road,  
London, E.17.

### Enthoven (58)

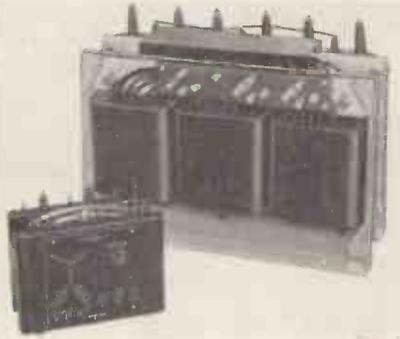
NEW developments shown by Enthoven this year will be aluminium cored solder and dip soldering materials for printed circuit techniques.

It is claimed that the aluminium cored solder solves all problems in the tinning and soldering of light-gauge aluminium and aluminium alloys and enables copper wire to be joined with ease to aluminium surfaces.

Enthoven Solders Ltd.,  
Enthoven House,  
89 Upper Thames Street,  
London, E.C.4.

### Ferranti (44)

A NEW range of "Pentland" series A-resin cast 3-phase transformers (illustrated below) will be shown in addition to a full display of single-phase transformers, chokes and delay networks employing the resin cast technique. T.R. cells, which will be displayed include 3cm tuneable and broad band types together with a range of



10cm cells. An integrated microwave test bench designed for use in the 3cm band and incorporating the milled block technique of waveguide manufacture will be shown.

There will be a display of 2in, 2½in and 3½in hermetically sealed instruments for the latest Service requirements. Included in these are instruments with American fixing centres providing interchangeability on N.A.T.O. equipment. Moving iron voltmeters with a frequency range of 50 to 2 000c/s will be shown.

Ferranti Ltd.,  
Hollingwood,  
Lancashire.

### G.E.C. (135)

A WIDE range of new valves will be shown. Included among these is the Osram DA42, which is designed for use in pairs in public address amplifiers where output powers of 175W are required. A triode with an indirectly heated cathode, it can be used to replace the DA41 in existing equipment.

The heater voltage and current are 7.5V and 1.2A respectively. In a typical push-pull class-B amplifier, the anode voltage for the pair is 1 000V for normal speech and music. Continuous 100 per cent tone modulation will, however, result in excessive dissipation and for such applications the normal anode-anode load of 10kΩ should be increased to 12kΩ, which will cause a reduction of the power output to 150W.

Other new valves displayed include: the KT55 power tetrode which is designed for

use as an audio frequency amplifier; the DH719 triple diode triode and the B719 r.f. double triode, for use in f.m.-a.m. radio receivers; the X719 triode heptode, which is primarily a frequency changer; the W719, a remote cut-off, high slope r.f. pentode for use in f.m. radio receivers; the HN309 triode pentode which has a high  $\mu$  triode; and the WD709 double diode pentode, the pentode section being suitable for use as an r.f. or i.f. amplifier.

Also introduced are three new semi-conductor devices. These are a medium power germanium pn junction rectifier, a high frequency point contact transistor and a junction transistor.

The General Electric Co. Ltd.,  
Magnet House,  
Kingsway,  
London, W.C.2.

### Gresham Transformers (70)

FOR some time Gresham Transformers Ltd. have held Approval Certificates for C-core hermetically sealed types and for



open types with standard laminations. Now their C-core open type has been granted Type Approval to Humidity Class H.1 and their "Leocast" (resin moulded) type with standard laminations has been granted limited approval to humidity class H.1.

The Leocast range offers particular advantages in physical and dielectric strength combined with outstanding resistance to humidity and temperature variation.

Illustrated above are, from left to right, in pairs, C-core open type; Leocast (resin moulded) with standard laminations, hermetically sealed C-core, open type with standard laminations.

Gresham Transformers Ltd.,  
Twickenham Road,  
Hanworth, Middx.

### Hunts (23)

INCLUDED in the new types of capacitors to be shown is the Midget Supermold seal type W96 (illustrated below). These metallized paper capacitors are enclosed in a new cast resin housing which is proof against humidity, mechanical shock and heat contact. They are available in a wide range of values up to 0.1  $\mu$ F.

Also to be shown are ranges of capacitors developed for use in all types of printed circuit.

A. H. Hunt (Capacitors) Ltd.,  
Bendon Valley, Garratt Lane,  
London, S.W.18.



### Imhof (25)

A WIDE range of cabinets, racks (including totally enclosed types), control consoles and fittings, will be shown.

The control console can be supplied in either double or single pedestal (the single with tubular leg, either left- or right-hand). The pedestal frames are drilled front and back to take standard 19in panels and an inset frame is provided mounted either 19in from the back or front so that standard chassis runners and telescopic runners may be fitted. Consoles are available with two types of superstructure, consisting of a full length 3-bank unit taking either 10½in or 15½in standard 19in rack panels inclined at 13° to the vertical. The console top is of grey linette formica.

Alfred Imhof Ltd.,  
112 New Oxford Street,  
London, W.C.1.

### London Electric Wire Company (61)

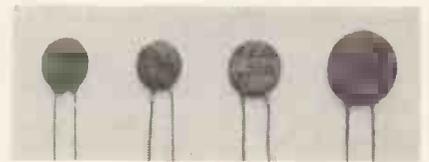
THE ever extending range of insulated electrical winding wires now includes self-fluxing enamelled wires including "Lewcosol" and "Lewtinex." Lewcosol wire can be soldered rapidly with an iron without prior removal of the enamel film. It has good all-round properties, complying in most respects with British Standard 184. Lewtinex is a modified type of Lewmex, which retains all the excellent properties of Lewmex and, in addition, can be dip-tinned in molten solder at 475°C.

Heat-resisting insulated wire including types covered with p.t.f.e. and terylene, will be shown.

The London Electric Wire Co. & Smiths Ltd.,  
24 Queen Anne's Gate,  
London, S.W.1.

### London Electrical Manufacturing Co. (36)

A WIDE range of silvered mica and silvered ceramic capacitors will be shown.



The range of F.E.C. silvered ceramic capacitors has been extended and includes tubular types from 1pF to 0.01  $\mu$ F, and disk types (illustrated above) from 470pF to 0.01  $\mu$ F, and the new lead through pattern which is designed to have a small physical size, exceptionally low inductance and low initial cost.

London Electrical Manufacturing Co. Ltd.,  
Beavor Lane,  
Hammer Smith,  
London, W.6.

### Marconi Instruments (103)

THIS company will be exhibiting a number of telecommunications test instruments including several new models.

In the field of frequency measurement, the company is showing the e.h.f. wavemeter TF1034 and the portable frequency meter TF1026 (series). The wavemeter is a double-cavity instrument designed for the measurement of frequencies in the range 32 000 to 35 500Mc/s. One tunable cavity forms a transmission-type wave-

meter of relatively low Q and is fitted with couplings for insertion in a waveguide circuit. The second cavity is coupled to the first and forms a supplementary wavemeter of the absorption type; independently variable, and of high Q, this second cavity can be tuned to produce a sharp reaction within the pass-band of the first cavity.

The latest addition to the range of voltage measuring equipment is the valve voltmeter TF1041 (illustrated below). This instrument measures up to 300V and 700Mc/s unbalanced, in six ranges; d.c. up to 1 000V balanced or unbalanced, in seven ranges. There are seven other ranges for the measurement of resistances between 0.2Ω and 500MΩ. A feature of the instrument is the large meter for fast and precise reading.

**Marconi Instruments Ltd.,  
St. Albans,  
Hertfordshire.**



**Multicore (69)**

A NEW type of Ersin Multicore solder which will be shown, containing five cores of the extra fast non-corrosive type 362 flux, is known as Savbit Alloy and contains approximately 2 per cent copper and is claimed to increase the useful life of a soldering bit by 10 times. This special alloy is already in production and can be supplied in nominal tin content alloys of 60, 50 and 40 per cent. There is no appreciable difference in melting points between Savbit alloys and conventional tin-lead alloys.

A solder containing silver for use on silver-plated ceramics and a complete fluxing and soldering process for use with printed circuits, will be shown.

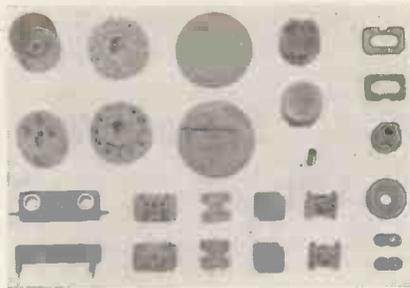
**Multicore Solders Ltd.,  
Multicore Works,  
Maylands Avenue,  
Hemel Hempstead,  
Hertfordshire.**

**Mycalex (3)**

A NEW glass-bonded insulating material known as Mycalon will be shown.

This material was evolved in conjunction with the Ministry of Supply, S.R.D.E., Christchurch, to meet service requirements under tropical and arctic conditions and, generally speaking, components requiring to be graded as H.1 can include Mycalon.

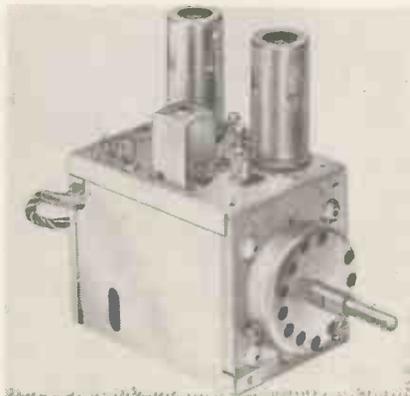
After subjection to six humidity cycles in the standard chamber to RCS11, Mycalon recovers to its full resistivity value within one hour and, upon removal from the testing chamber, the immediate figures are not lower than  $10^{11}\Omega$ , their full value when



dry being greater than  $10^{14}\Omega$ . The permittivity and power factor are lower, resulting in a loss factor of approximately half that of the existing grade on Injection Moulded Mycalex.

Mycalon can be injection or transfer moulded into any required shape, and to the finest tolerances desired because it does not shrink or distort after moulding. Illustrated above are some typical Mycalon components.

**Mycalex and T.I.M. Ltd.,  
Ashcroft Road,  
Cirencester,  
Gloucestershire.**



**N.S.F. (49)**

IN addition to a wide range of switches the N.S.F.-Oak switch type cascode television tuner will be shown (illustrated above). This is available in two versions, type 12K13 covering five channels in Band 1 and eight channels in Band 3, and type 12K14 with an additional switching position which will enable it easily to be converted to u.h.f. working.

**N.S.F. Ltd.,  
9 Stratford Place,  
London, W.1.**

**Parmeko (27)**

A SPECIAL display this year will be Amade of the Saturn Series Transducers



for Magnetic Amplifier Applications. Illustrated below left are three typical transducers from the standard range developed by Parmeko in collaboration with the Royal Aircraft Establishment. This range complies with the Joint Services Specification RCL(PROV) 217 and comprises auto-excited transducers of the push-pull input and single- and three-phase output types.

Also to be shown is a range of transformers designed specifically for use with printed circuits and a range meeting American service specifications.

**Parmeko Ltd.,  
Percy Road,  
Aylestone Park,  
Leicester.**

**Pye (96)**

EIGHT members of the Pye group of Companies will be exhibiting at this year's exhibition.

W. G. Pye & Co. Ltd. are showing an audio frequency oscillator with a frequency range of 20 to 20 000c/s and an output continuously variable from 0 to 20V peak-to-peak, electrostatic voltmeters covering ranges from 1 to 18kV, and a d.c. microvoltmeter which has a full scale range of 20μV and adjustment to reduce sensitivity by factors of 1/10, 1/100 and 1/1 000.

Magnetic Devices Ltd. are showing a range of relays for both a.c. and d.c. supplies and a special relay for r.f. switching. Prominent among the new products are a sensitive type of relay for electronic circuit applications and a hermetically sealed plug-in miniature relay.

Power Controls Ltd. display a range of multi-way plugs and sockets particularly suited to rack-mounted equipment and miniature hermetically sealed coaxial plugs and sockets for Uniradio 43 cables.

Labgear (Cambridge) Ltd. are showing a range of wire-wound resistors, two types of television receiver filter, for reducing interference to television from nearby radio transmitters operating at frequencies lower than 30Mc/s or higher than 70Mc/s, a standing wave ratio indicator, and a crystal oven, with a temperature differential of 1.5°C, which operates from a 6.3V supply.

Cathodeon Ltd. exhibit a wide range of glass-to-metal seals, aerial changeover relays and thermal delay switches.

Pye Industrial Electronics Ltd. are making their first appearance at this exhibition and are showing a number of types of junction transistors which are intended primarily for hearing aid and low-powered audio frequency application and are all hermetically sealed.

Cathodeon Crystals Ltd. are exhibiting a range of quartz crystals for frequency control. All are hermetically sealed and in general adhere to requirements of the inter-services specification.

Pye Ltd., Switch Division, add a further sixteen types to their already considerable range of industrial and aircraft micro-switches.

**Pye Ltd.,  
Cambridge.**

**Salford (34)**

NEW products displayed will include a range of precision potentiometers with metal housings. Available in both linear and functional card winding types, these units employ an improved type of wiper assembly, in which the wiper, being horizontal, gives a more even tracking over the resistance winding. On the twin wiper

arm type, the second arm is now capable of independent movement through 230°, whereas the standard units in the past were limited to a 90° or 180° movement.

An improved method of ganging is also used and this makes for a more compact assembly as well as permitting the housings to be rotated individually for calibration purposes. A servo mounting spigot is mounted concentrically with the housing and a rotary clamping plate can be fitted for applications where the body of the potentiometer has to be rotated before clamping. A high grade bakelite former is used in the general-purpose units. Special formers are available for applications involving ambient temperature variations as wide as from -20°C to +85°C.

(Illustrated below)

Salford Electrical Instruments Ltd.,  
Magnet House,  
Kingsway,  
London, W.C.2.



#### Stability Radio (51)

STABILITY Radio Components are showing beside their silvered mica capacitors a range of ceramic tubular capacitors rated at 100°C. These are inexpensive, robust and fully insulated components which cannot be damaged by prolonged contact with the soldering iron. Capacitance values are from 0.5pF to 0.005μF.

A new introduction for f.m. and Band 3 receivers is the "top coupling" range from 0.5pF to 1.5pF, which is supplied at an economic price with a tolerance of +0.1F.

Stability Radio Components Ltd.,  
Commerce Estate,  
Woodford Avenue,  
London, E.18.

#### S.T.C. (62)

A RANGE of selenium and germanium Rectifiers, Unistors, Thermistors, photocells, tantalum capacitors and quartz crystals will be shown.

Overtone crystals are now available in production quantities up to 55Mc/s. This frequency will shortly be extended, and to aid equipment designers, laboratory samples can be supplied at frequencies as high as 200Mc/s. The crystal oscillates in the third, fifth or higher overtone mode, thus the frequency of vibration is increased above the fundamental frequency approximately by the order of overtone employed. The circuit must be tuned in some manner to discriminate between the fundamental frequency and the desired overtone. The units are robust, stable and exhibit favourable electrical characteristics.

Standard Telephones & Cables Ltd.,  
Connaught House,  
Aldwych,  
London, W.C.2.

#### T.C.C. (54)

AMONG the new components shown by T.C.C. will be a range of low leakage high quality electrolytic capacitors known as Superlytics; these are of particular interest to designers of low frequency amplifiers, for grid coupling purposes.

The outstanding feature of these capacitors is their exceptionally low leakage which is such that for the first time it is possible to think in terms of insulation resistance of an electrolytic capacitor. Measurements taken have resulted in IR figures which are comparable with the standard paper dielectric capacitors. This characteristic results in a very long working life, and renders them suitable for operation at temperatures up to +85°C without voltage derating.

Also on display will be a range of sub-miniature electrolytic capacitors type CE58. These components measure only ½in by ⅝in and are intended for use in transistor circuits.

In order to comply with RCS132 moulded Plimoseal silvered mica capacitors are introduced. This method of protection not only prevents the ingress of moisture, it also improves the insulation resistance and power factor, without impairing the inherent stability of the capacitor unit.

In addition, various types of capacitors conforming to American specifications, a wide range of printed circuits and capacitors for use therein, will be shown.

The Telegraph Condenser Co. Ltd.,  
North Acton,  
London, W.3.

#### Steatite and Porcelain (38)

INCLUDED in the materials on this stand will be Faradex H, a high permittivity ceramic (K=3200) and Frequentite S, a material of extremely low dielectric loss which, being non-porous, is particularly suited for vacuum applications such as u.h.f. valve envelopes, magnetron windows, etc. Process improvements such as metalizing and machining of ceramics, will be shown.

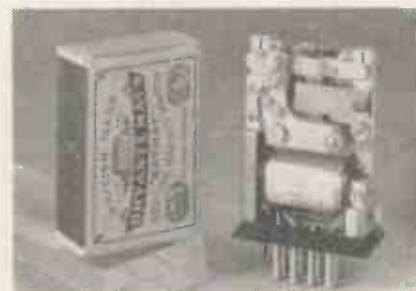
Steatite and Porcelain Products Ltd.,  
Stourport-on-Severn,  
Worcestershire.

#### T.M.C. (33)

"HIQUOR" powder cores will be shown in a large range of sizes with permeabilities from 15 to 125. These cores have a very low loss factor at frequencies up to 1Mc/s.

A range of metallized paper, polystyrene and American type capacitors will be exhibited.

Continuous development in the design of Carpenter polarized relays (illustrated below) is resulting in improved performance and stability over an extended temperature



range. One of the main advantages of the relays is their high degree of sensitivity and working models will show them operating from a thermo-couple junction, and a photo-electric cell.

Telephone Manufacturing Co. Ltd.,  
Hollingsworth Works,  
Martell Road,  
London, S.E.21.

#### Vactite (7)

THIS company will be exhibiting their Vacrom (nickel-chrome) and Eureka (cupro-nickel) resistance wires and tapes, pure nickel wires and tapes, molybdenum rods, wires and tapes and special wires for the radio valve industry.

Vactite have made a special study of superfine resistance wires and welcome inquiries for sizes in the 'triple O' range.

Vactite Wire Co. Ltd.,  
24 Queen Anne's Gate,  
London, S.W.1.



#### Whiteley (14)

THE range of Stentorian loudspeakers in sizes from 2½ to 18 in will be shown together with a wide variety of television and radio components including components potted in Polythene, Araldite and Marco resin. A C-core transformer moulded in Araldite and fitted with waterproof plugs and sockets is illustrated above.

Whiteley Electrical Radio Co. Ltd.,  
Radio Works,  
Victoria Street,  
Mansfield, Notts.

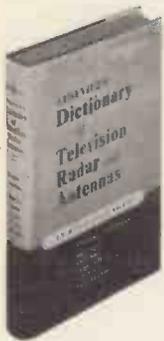
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**BOOK REVIEWS**

**Electronics: Electrical Engineering Approach**

By George F. Corcoran and Henry W. Price. 459 pp., 120 figs. Demy 8vo. John Wiley & Sons, Inc. Chapman & Hall Ltd. 1954. Price 56s.

THIS book would be more correctly named "Electronic Circuits" since the space devoted to the physics of the subject is relatively small, as the authors themselves admit.

A very brief summary of basic electrical theory and a glance at the electronic structure of the atom lead to a treatment of electron-movement with examples based on the cathode-ray oscillograph. A qualitative description of the effects of space charge in the vacuum diode introduces a fairly extensive section on rectifier circuits and other applications of the diode. The internal aspects of the triode are dismissed rather briefly but a full account of equivalent circuits for triodes is given. Tetrodes and pentodes are next introduced and a detailed chapter follows dealing with various types of class A amplifier. The treatment relies almost exclusively on the equivalent circuit with the associated complex algebra, graphical methods being avoided. A few special cases such as the cathode-follower, reactance-valve circuits and paraphase circuits are followed by a section on systematic methods of analysing complicated equivalent circuits. A note on noise is included.

Load lines are first used in the chapter on class A and class B power amplifiers. Load ellipses and composite characteristics (for push-pull stages) are also described. In the well-written chapter on feedback all the important points are quantitatively treated with the exception of the Nyquist criterion which is, perhaps wisely, stated without proof. A return to physics (excitation, ionization, etc.) heralds a treatment of the voltage stabilizer. A few words on gasfilled rectifiers and thyratrons lead to a second section on rectifier circuits.

Perhaps the chief attraction in the book is the very detailed chapter on germanium diodes and transistors. A simple but adequate account of the physics of germanium is first given and then the principles and practice of these relatively new devices is painstakingly explained. The last chapter, on oscillators, is somewhat surprisingly separated by over 100 pages from that on feedback and thus illustrates the slightly haphazard plan of the book as a whole. We find, for example, smoothing circuits treated partly under the vacuum diode and partly under gas-filled rectifiers, while paraphase amplifiers also appear in two instalments.

Perhaps the least satisfactory part of this book is that which purports to deal with fundamentals. Apart from being condensed to an absurd degree it exhibits that vagueness which usually accompanies the m.k.s. system of units in dealing with basic matters. For example, field strength ( $H$ ) is defined as rate of variation of ampere-turns with distances, a meaning-

less definition resulting from the fashionable abhorrence of the magnetic pole. It is symptomatic that Gauss's theorem does not qualify for a place among the fundamentals but creeps in apologetically as a footnote.

One cannot hope at this date for a change in the unmethodical American practice of calling a bi-phase half-wave rectifier circuit a "full-wave" circuit, but it is disappointing to find a paraphase amplifier called an "inverter," especially in view of the fact that the device properly called an inverter is not mentioned.

The chief characteristic of this book is a most praiseworthy thoroughness in matters of practical application. With numerous worked examples the authors show clearly how to tackle a variety of real-life problems. Many pages of examples for the student are also provided. Indeed, this thoroughness is partly responsible for the rather limited scope of the book. One might expect to find something about modulation, pulse generation, sweep circuits and other familiar matters in a volume of this size, but they do not appear.

The printing and binding are so good as to compensate in some degree for the high price, but intending purchasers would be well advised to compare the scope of this book with that of similar books by British authors, who are compelled by circumstances to practice the art of condensation.

K. A. MACFADYEN.

**Elektromagnetische Wellen (Electromagnetic Waves)**

By Karl Wagner. 267 pp., 185 figs. Demy 8vo. Verlag Birkhäuser. Basel. 1953. Price Sfr.33.30 bound, Sfr.29.10 paper cover.

THIS book, carrying the sub-title "An introduction to the theory as basis to its application in electrical transmission technique," is the outcome of lecture courses which the late Professor Wagner held in Sweden, Italy and Spain between 1947 and 1951. It may be considered as the swan song of a scientist and engineer whose many contributions to the development of electrical communication and power engineering are well known to all those who have followed with interest, since the publication of his first book *Electromagnetic Transients in Overhead Lines and Cables* in 1908, the brilliant career of one of the foremost German authorities in these fields within the last half century.

It is not a book for the beginner and a sound knowledge of Maxwell's theory, of vector analysis, of the theory of functions, of quadrupole theory and the like is required for its full understanding. It is perhaps a sign of the high standard of technical education on the Continent that in the author's opinion the mathematical background expected from the reader scarcely exceeds what is required for the intermediate examination of an engineering student.

A brief account of the main contents of the book will show its wide scope. The first chapter, dealing with plane waves in free space, shows after a brief historical introduction how waves are propagated. The wave equations are derived for constant and periodically varying fields. Linear and elliptical polarization and phase and group velocities are discussed.

The second chapter is devoted to reflexion and diffraction of electromagnetic waves. Various cases are discussed, e.g. that of vertical incidence, the case in which the reflected wave is extinguished, the total reflexion and the case when the refraction ratio is complex.

In dealing with waves conducted by wires in the third chapter, which is of particular interest for the power engineer, the telegraph equation is derived and the theory of travelling waves is given together with some examples of such waves, e.g. those caused by lightning or by switching operations. The effect of protective capacitors and chokes and the overvoltages caused by interrupting a short-circuit are discussed.

In dealing with periodical waves on wires the spiral diagram illustrating amount and phase of voltage and current at any point of the line is explained. Propagation-, phase- and attenuation-constants and characteristic impedance are calculated. The line is then treated as a quadrupole and it is shown how its characteristic values may be determined by no-load and short-circuit test. Examples of "short" and "long" lines are given and the consideration of standing waves illustrates the use of Lecher wires. A comparatively large section deals with the application of the theory to aerials. The skin effect in cylindrical conductors is briefly dealt with and inhomogeneous conductors, i.e. those in which the line parameters vary from point to point are given due consideration.

In the fourth chapter bunched conductors are dealt with and the mutual interference between a power and a telegraph line is investigated as an example.

The fifth chapter is devoted to waves in hollow conductors. E-waves and H-waves in circular, rectangular and square conductors and their attenuation, hollow conductors used as resonators and their properties considered as quadrupoles are discussed.

The comparatively brief sixth chapter deals with electric and magnetic dipoles, e.g. those used as loop aerials.

The concluding seventh chapter is a comprehensive survey of electric waves in the ionosphere. Their field intensity, attenuation, polarization, limiting frequencies, reflexion and refraction. Graphs show the daily and annual changes of the limiting frequencies in the various regions of the ionosphere. The effects of eclipses and of the number of sunspots and of other ionospheric disturbances are described and a brief final section deals with the transmission of electromagnetic waves through the ionosphere.

A name and a subject index are provided. The book is well produced and should prove helpful not only to the advanced student but also to practical research workers both in power and communication engineering.

R. NEUMANN:

### Telecommunications

By A. T. Starr. 455 pp., 65 figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1954. Price 35s.

THIS book is said to be written to cover the syllabus for the subject of telecommunications in the London University Degree. No doubt a student would like to think that for 35s. he could buy all the knowledge needed for one subject of his university degree, but most university teachers would prefer that he should acquire the habit of using specialized treatises and original papers for reference: the greatest fault of Dr. Starr's book is that it appears to have no references whatsoever to the rest of the literature. Moreover, in so encyclopaedic a book there are naturally some topics which are less fully covered than others, and here a pointer to other sources is essential. Without taking advantage of one's own special interests, one can cite "Radio Position and Direction Finding," which is given four and a half pages with the barest mention of the screened loop and no suggestion that the e.m.f. is calculable from the magnetic component of the incident wave. Again, those of us who dabble only in radio might be forgiven heretical views on the neper, but a book which covers also line communication should adhere to the definition that the neper is a logarithmic unit of current ratio, not of power like the decibel.

The book includes mathematic appendices on Heaviside's operational calculus, the Laplace transform and vector fields, but no guide to matrix notation which slips unannounced into pages 166-168. But despite all its faults, it must be agreed that this book has achieved something in presenting in a single volume a comprehensive survey of the field of telecommunication. It includes clear descriptions and drawings of relays, switches and teleprinter mechanisms which are so rarely described in books, and the longest chapter in the book is on "Lines and Transmission Networks," a subject in which Dr. Starr is an acknowledged authority. Among the topics covered in other chapters are acoustics, television, electromagnetic waves and microwave technique with a 6-page appendix on "Waveguide Theory" so that any student who is well versed in all the topics discussed in this book should be equipped to face both examinations and real-life problems in telecommunications.

D. A. BELL.

### Kempe's Engineers' Year Book 1955

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THE preface to the 1955 Kempe states that last year's edition was so completely revised that this year attention has been primarily confined to detail. Only minor changes, therefore, occur in the edition under review and the most significant in the chapter on Electronic Engineering is no more than a new paragraph under the sub-heading of Colour Television.

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# LETTERS TO THE EDITOR

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

## A Versatile Pulse Shaper

DEAR SIR,—With reference to the article by Mr. Kaufer in the February issue (p. 78), this circuit was tried in the laboratory several years ago. It was found that for precise operation the blocking oscillator could be superior to the thyatron due to the firing time delay with the latter type of pulse former. There was an appreciable time delay between the top of the leading edge of the trigger pulse and the output pulse. A graph shows this for a sample 2D21 (CV797). There were also considerable differences between other samples of this valve and firing delays up to 1.5Mc/s could be obtained. Although an output pulse could not be obtained by applying

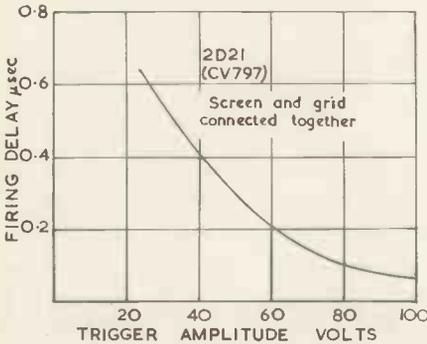


Fig. 1. Firing delay/trigger amplitude for 2D21

a trigger pulse to the screen only, it was possible to "gate" the valve with the screen providing the gating waveform was slightly in advance of the main trigger pulse.

Yours faithfully,  
K. E. WOOD,  
Leatherhead,  
Surrey.

## High-Speed Magnetic Amplifiers

DEAR SIR,—Mr. A. E. Maine in his article on high-speed magnetic amplifiers (page 514 of the December issue) describes two methods of coupling Ramey-type amplifiers. Another method is to use a pnp junction transistor, arranged as shown in Fig. 1, as the coupling device.

The transistor  $X_2$  normally has a high impedance between base and collector so that the re-setting voltage for core 2 is absorbed by the transistor. When core 1 fires it delivers a large emitter current to the transistor  $X_2$  decreasing its collector-to-base resistance so that the re-setting voltage is applied to the second core. In this way the degree of re-setting of core 2 depends on the firing angle of core 1, which in its turn can be controlled either with a voltage signal or

another transistor used as a controlling element. The sequence of waveforms in Fig. 2 shows how the re-setting voltage to core 2 is absorbed by the transistor to an extent depending on the firing angle of core 1. The small voltage pulse before the main collector voltage wave is due to  $Z_1$  being largely capacitive, and the incomplete voltage waveforms and the incomplete current waveform of Fig. 2(a) are attributed to the poor rectangularity of core material used. The supply frequency is 2 500c/s.

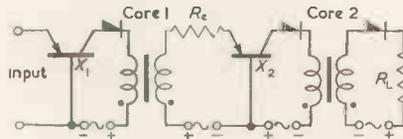


Fig. 1. Transistor coupled Ramey amplifier

When the re-setting of the first core is voltage controlled the coupling arrangement retains the linear relation between the mean input and output voltages, except in as much that the output voltage decreases with increasing voltage input. This property reduces the biasing requirements when the first stage is controlled by a valve or transistor.

The waveforms of Fig. 2 show that high collector currents and voltages do

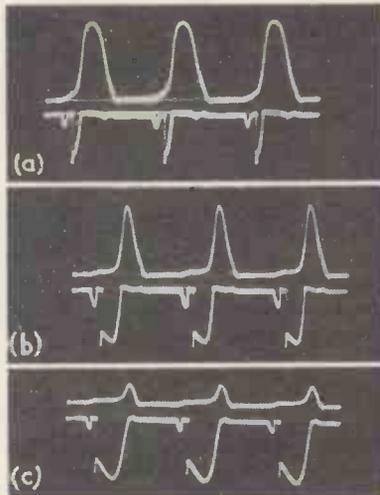


Fig. 2. Waveforms showing variations of the emitter current (top) and collector voltage (bottom) of the coupling transistor  $X_2$  as the emitter current of the controlling transistor  $X_1$  is increased from zero

not occur simultaneously. This enables transistors with low power ratings to couple circuits controlling much larger powers.

Yours faithfully,  
R. E. WRIGHT,  
The General Electric Co. Ltd,  
Stanmore.

## An A.C. Voltage Stabilizer

DEAR SIR,—In my article published in the April 1954 issue it was pointed out that the r.m.s. output voltage of the stabilizer could be affected by changes in the peak factor of the voltage waveform, since it is the peak value which is controlled. Under normal circumstances this characteristic will not cause any trouble. Changes in the peak factor of supply lines are very uncommon, and the waveforms of auxiliary generating sets, which are often not truly sinusoidal, rarely change sufficiently to result in significant changes in stabilizer output voltage. However, the distortion frequently present in the waveforms of small a.c. generators could cause trouble should a stabilizer be expected to hold the output voltage constant irrespective of whether the source is the normal supply line, or (in the event of a power failure) an auxiliary generating set.

The distortion present in the voltage waveforms of such generators is usually due to the presence of a small percentage of the third harmonic, plus very small amounts (normally insignificant) of the second and fifth harmonics. If, as is usually the case, this third harmonic is in phase with the fundamental, the r.m.s. voltage for a given peak value is proportionately higher. Thus it can be seen, for example, that the introduction of 5 per cent of the third harmonic (a not uncommon amount) would result in the output voltage of one of the above stabilizers rising by nearly 12V.

Neglecting the other harmonics, and letting  $k$  represent the percentage of the third present (in phase), it can be shown that the r.m.s. and average values of the voltage ( $V_r$  and  $V_a$  respectively) of such a distorted wave, in terms of the actual peak value ( $V_p$ ), are given by the following expressions (for  $k$  not greater than 11.1 per cent):

$$V_r = \frac{\sqrt{10,000 + k^2}}{\sqrt{2(100 - k)}} V_p \quad (1)$$

$$\text{and } V_a = \frac{300 + k}{1.5\pi(100 - k)} V_p \quad (2)$$

$$\text{From equations (1) and (2)} \\ V_r = \frac{1.5\pi\sqrt{10,000 + k^2}}{\sqrt{2(300 + k)}} V_a \quad (3)$$

By giving  $k$  in equation (1) and (3) various values from 0 to 10, and putting the results so obtained in terms of  $V_r = 100$  for  $k = 0$ , the following table is obtained.

This table shows how, if the peak value is held constant, the r.m.s. value rises as the percentage of third harmonic increases, whereas if the average value is controlled, the r.m.s. value falls. Therefore, for any particular amount of third harmonic it is possible, by making the voltage control unit of the stabilizer operate at a voltage appropriately disposed between the peak and average values, to hold the r.m.s. output voltage

k	V <sub>r</sub> for constant V <sub>p</sub> (from equation 1)		V <sub>r</sub> for constant V <sub>a</sub> (from equation 3)	
	PERCENTAGE 3RD HARMONIC	RELATIVE VALUE	PERCENTAGE RISE	PERCENTAGE FALL
0	100.0	.0	100.0	0.0
1	101.0	1.0	99.7	0.3
2	102.1	2.1	99.3	0.7
3	103.1	3.1	99.0	1.0
4	104.2	4.2	98.8	1.2
5	105.4	5.4	98.5	1.5
6	106.6	6.6	98.2	1.8
7	107.8	7.8	98.0	2.0
8	109.0	9.0	97.7	2.3
9	110.3	10.3	97.5	2.5
10	111.7	11.7	97.3	2.7

constant irrespective of whether the harmonic is present or not.

Perhaps this principle may best be illustrated by an example. With  $C_1^*$  of nominal value  $16\mu\text{F}$ , the d.c. voltage across the bridge circuit of the sensing unit is about 290V, being the peak value of the incoming 230V r.m.s. less a voltage drop in the rectifier  $MR_1$ . With  $C_1=0$ , this voltage is about 190V, being the average value of the 230V r.m.s. less a voltage drop due to the regulation characteristics of the system. If it is desired that the voltage across  $C_1$ , and hence across the bridge circuit of the voltage sensing unit, should be the same irrespective of whether the incoming a.c. is truly sinusoidal or contains, say, 4 per cent of the third harmonic, then this voltage should be

$$190 + \left( \frac{1.2}{4.2 + 1.2} \right) (290 - 190) = 212\text{V.}$$

To achieve this effect  $C_1$  should have a value of  $0.5\mu\text{F}$ . (As a matter of interest, the control point of a stabilizer modified to give the correct r.m.s. output voltage when there is 4 per cent of harmonic present will not vary by more than  $\pm 1$  per cent for a range of harmonic content from 0 to about 8 per cent.)

Consequent upon such an alteration being made to the voltage controlling unit, several other minor ones become necessary. The ratio of the divider system  $R_3, R_4, VR_1, R_5, R_6$  needs to be changed;  $R_1$  and  $R_2$  both need to be reduced; an additional  $0.5\text{ M}\Omega$  resistor needs to be connected between  $C_1$  (+) and the ignition electrode of V to ensure reliable ignition of V (since the voltage across it now drops below its operating point twice per cycle); and additional biasing of the moving-coil relay A is necessary to prevent chattering of its contacts (because of vibration caused by the 100c/s component now present). Typical values for the particular case quoted above are  $R_3=5\text{k}\Omega$ ,  $R_4=10\text{k}\Omega$ ,  $R_5=15\text{k}\Omega$ ,  $R_6$  and  $R_6=330\Omega$ .

Yours faithfully,

R. G. ACKLAND,

Defence Standards Laboratories,  
Victoria,  
Australia.

\* Ref. Fig. 3, *Electronic Engng.*, 26, 144 (1954).

### Computers and Computors

DEAR SIR,—I presume Mr. Puckle would have his computer operated by an "operator," and would refer to you, Sir, as an "editor" so as to leave the word "editor" for a device used to check the punched tapes fed to his machine.

It is not strictly true that the words "computer" and "computor" are different—rather are they alternative spellings of the same. Nevertheless, I agree that there is a case for a convention of the type suggested, and in fact this appears to be already fairly well established, in the opposite sense to that proposed by Mr. Puckle.

Yours faithfully,

B. M. ADKINS,

Radlett,  
Herts.

DEAR SIR,—I was sorry to read Mr. Puckle's letter in the February issue, claiming that there are two words "computer" and "computor." There is, in fact, only one accepted word (see the recognized English dictionaries), viz., "computer," which covers both the man and the machine, and the large majority of people consider "computor" merely a spelling mistake.

It is quite legitimate for a profession to coin a new word if confusion can arise between two different senses of the word as it is generally used. Any such step, however, should be taken only with agreement throughout the profession, because the English language covers the whole of the earth and divergences in spelling or meaning in different localities are obviously undesirable. No such agreement on "computor" has been given or even sought.

I submit that no confusion arises in practice between the man and the machine, the most convincing proof being that even those who make a difference in spelling usually make no difference in pronunciation. If a new word were desirable, etymology would suggest "computator," not "computor"; the latter is etymologically shocking and is otherwise unsatisfactory because it would still leave confusion in the spoken word. A proposal was made a quarter of a century ago to reserve the termination "or" for the machine in a general way, and this is responsible for such currency as "computor" and "convertor" still have. The initiative, however, received very little

support even in this country and none in the U.S.A. It was pointed out that there would be just as much reason for changing the spelling to "windscreen-wipor" for fear there should be confusion between the attachment to the motor-car and the garage attendant who cleans a wind-screen.

I hope Mr. Puckle will find it possible to drop "computor" which does not really do his job.

Yours faithfully,

E. H. ULLRICH,  
Richmond, Surrey.

### Correspondent replies :

DEAR SIR,—I much regret that I am unable to agree either with Mr. Ullrich or Mr. Adkins.

To take Mr. Adkins' letter first. He agrees that there is a case for a convention of the type suggested in my letter (February, 1955, issue) but he says that it is already established in the opposite sense. I maintain that since the two words "computer" and "computor" already exist there is no need for convention. Does Mr. Adkins suggest that "computor" is used to describe the user of the machine? It is precisely because I find that people are beginning to use the wrong spelling that I wrote my letter to you.

Mr. Ullrich's letter suggests that I have coined the word "computor." I thank him for the compliment but I must disclaim the honour. The only word I have coined which has come into general use is "sync" and, even then, I spelled it "synch." I still consider that the "h" is necessary to harden the "c," but at least I have got 80 per cent agreement! Mr. Ullrich's humorous effort, "wind-screen wipor" made me chuckle but does not affect the issue. The English language has many words whose spelling appears to be wayward but, nevertheless, it seems to me to be a pity to debate it by wrong spelling or by the wrong use of words. It does not appear to me to be of any importance that a word is not accepted in the U.S.A., for their language has quite naturally gone along its own path; differences need not cause any confusion. An English "frame" and an American one in television engineering have quite different meanings yet we are not confused. Nor is it necessary for us to follow their lead when we have an existing word in our language. It certainly won't upset our friendship if we fail to copy them, in fact they will respect us all the more for having minds of our own.

I particularly dislike Mr. Ullrich's suggestion to use "computator" which would introduce another horror like "commentator." In this case, the correct word is "commentor."

I hope that Messrs. Ullrich and Adkins will reconsider the matter and eventually follow the lead given by THE TIMES and ELECTRONIC ENGINEERING.

Yours faithfully,

O. S. PUCKLE,  
Beaconsfield,  
Bucks.

# Meetings this Month

## THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 13 April. Time: 6.30 p.m.  
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.  
Discussion: The BBC V.H.F./F.M. Sound Broadcasting Service.  
Opened by: K. R. Sturley.  
Date: 27 April. (Time and place as above).  
Lecture: Suppressed Aerials for the Aircraft H.F. Band.  
By: K. J. Coppin.

### North Eastern Section

Date: 13 April. Time: 6 p.m.  
Held at: Neville Hall, Westgate Road, Newcastle-upon-Tyne.  
Lecture: The Manchester University Electronic Computer.  
By: E. T. Warburton.  
Followed by Annual General Meeting.

### Merseyside Section

Date: 21 April. Time: 7 p.m.  
Held at: Riversdale Technical College, Riversdale Road, Liverpool 19.  
Annual General Meeting followed by technical films.

### South Wales Section

Date: 27 April. Time: 6.30 p.m.  
Held at: Cardiff College of Technology, Cathays Park, Cardiff.  
Lecture: Some Technical Problems in Sound and Television Broadcasting.  
By: K. R. Sturley.

## THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Date: 21 April.  
The Forty-Sixth Kelvin Lecture: Transistor Physics.  
By: W. Shockley.

### Measurements and Radio Sections joint meeting

Date: 5 April.  
Lecture: High Speed Electronic Analogue Computing Techniques.  
By: D. M. MacKay.

### Measurements Section

Date: 19 April.  
Discussion: The Measurement of Impulse Voltages and Currents with special reference to the testing of Surge Diverters.  
By: R. Davis.

### Radio Section

Date: 20 April.  
Lecture: A Study of the Long-Term Emission Behaviour of an Oxide Cathode Valve.

**North Eastern Radio and Measurements Group**  
Date: 4 April.  
Held at: King's College, Newcastle-upon-Tyne.  
Annual General Meeting followed by  
Lecture: Thermionic Valves of Improved Quality for Government and Industrial Purposes.  
By: E. G. Rowe, P. Welch and W. W. Wright.

### North Midland Centre

Date: 5 April. Time: 7 p.m.  
Held at: The Station Hotel, York.  
Lecture: The Possibilities of a Cross-Channel Power Link Between the British and French Supply Systems.  
By: D. P. Sayers, M. E. Laborde and F. J. Lane.

### South Midland Radio Group

Date: 25 April. Time: 6 p.m.  
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.  
Annual General Meeting followed by  
Lecture: A Transatlantic Telephone Cable.  
By: M. J. Kelly, Sir W. Gordon Radley, G. W. Gilman and R. J. Halsey.

### Southern Centre

Date: 1 April. Time: 6.30 p.m.  
Held at: South Dorset Technical College, Weymouth.  
Lecture: Cold Cathode Valves.  
By: J. R. Acton.  
Date: 13 April. Time: 6.30 p.m.  
Lecture: Modern Developments in Atomic Energy.  
By: T. E. Allibone.

## THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 6 April. Time: 5 p.m.  
Held at: The Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.  
Lecture: Regional Problems in the Provision and Maintenance of V.H.F., U.H.F. and S.H.F. Radio Links.  
By: F. Moxon.

## THE TELEVISION SOCIETY

Date: 1 April. Time: 7 p.m.  
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.  
Lecture: A Flying-spot (Mechau) Telcine System.  
By: J. L. Bliss.  
Date: 21 April. (Time and place as above).  
Lecture: Progress in Colour Television.  
By: L. C. Jesty.

## PUBLICATIONS RECEIVED

**THE COURTAULD ATOMIC MODELS** is a booklet describing what are believed to be the first models designed to incorporate an elastically-distortable bond of great use in the investigation of steric hindrances, developed in the Maidenhead laboratory of Courtaulds Ltd.

**ERMA CATALOGUE** of cable lugs, solderless terminals, crimping tools and equipment is a recent publication of Erma Ltd., Hong Kong Works, Exhibition Grounds, Wembley, Middlesex.

**GREEN SEAL TRANSFORMERS** is an illustrated booklet describing a series of tests carried out on transformers using new silicone insulation. This publication is obtainable free on application from Brentford Transformers Ltd., Kidbrooke Park Road, Kidbrooke, London, S.E.3.

**STANDARDS AND MEASURING EQUIPMENT FOR ELECTRICAL RESEARCH AND TELECOMMUNICATION ENGINEERING** is the new 1955 edition of the catalogue produced by H. W. Sullivan Ltd. It deals in a comprehensive manner with standards of resistance, capacitance and inductance of both fixed value and decade or continuously variable form. H. W. Sullivan Ltd., Leo Street, Peckham, London, S.E.15.

**SOLDERLESS WIRING CONNECTION FOR LIGHT AND HEAVY DUTY** is a booklet produced by the Plessey Company who pioneered and developed, primarily for the aircraft industry, the principle of solderless wiring connexion which ensures a reliable joint regardless of the skill of the operator, and which eliminates all troubles due to heat and the corrosive action of fluxes. Since then its use has been adapted to a much wider field of industrial application. The Plessey Co. Ltd., Ilford, Essex.

**POLARIMETERS, HILGER PHOTOELECTRIC TRISTIMULUS COLORIMETER, SOME APPLICATIONS OF THE DIRECT READING ATTACHMENT TO THE MEDIUM QUARTZ SPECTROGRAPH, NEW HILGER SPEKKER PHOTOELECTRIC FLUORIMETER**, are four recent publications of Hilger & Watts Ltd., Hilger Division, 98 St. Pancras Way, London, N.W.1.

**TEN YEARS OF SEMI-CONDUCTING MATERIALS AND TRANSISTORS** is a bibliography produced by the Research Department of Pye Ltd., Radio Works, Cambridge.

**TELEVISION FIELD SERVICE MANUAL WITH TUBE LOCATIONS, TELEVISION MANUFACTURERS' RECEIVER TROUBLE CURES VOLUME 6 and ADVANCED TELEVISION SERVICING TECHNIQUES** are recent publications applicable to American practice. John F. Rider Publisher, Inc., 480 Canal Street, New York, 13, U.S.A.

**FROM SCHOOL TO PROFESSIONAL ENGINEERING** is a booklet which aims to inform all those in public, secondary grammar and secondary technical schools of the opportunities which the Metropolitan-Vickers Electrical Company provides for training step by step with facilities for study in electrical and mechanical engineering, physics, chemistry and metallurgy. Metropolitan-Vickers Electrical Co. Ltd., Trafford Park, Manchester, 17.

**CANADIAN MARCONI MAGAZINE**, the first issue of which appeared in January, will be published five times yearly to inform readers about progress in the science and industry of electronics. Comments and suggestions are invited. Correspondence should be addressed to the Editor at Canadian Marconi Company Ltd., 2442 Trenton Avenue, Montreal 16, Quebec.

**G.E.C. AIRPORT ELECTRIFICATION** is a booklet in which the main emphasis has been placed on airport lighting, the various control systems used for this vital aspect of airport operation, on telecommunication and equipment peculiar to airports. A two runway airport, similar in size and layout to those in common use all over the world, has been taken as a basis for the illustration of various proposals, but such proposals are adaptable for any other layout. The General Electric Co. Ltd., Magnet House, Kingsway, London, W.C.2.

**SPERRY SYNCHROS** describes the range of instrument synchros manufactured by the Sperry Gyroscope Co. Ltd. It includes data sheets relating to the various models of synchros available from this Company and an explanation of how these units can be applied. The Sperry Gyroscope Co. Ltd., Great West Road, Brentford, Middlesex.

**THE LONDON AND HOME COUNTIES BULLETIN OF SPECIAL COURSES IN HIGHER TECHNOLOGY** for the Spring and Summer terms, 1955, is now available. Copies may be obtained from the Secretary, Regional Advisory Council, Tavistock House South, Tavistock Square, London, W.C.1, price 1s. 6d., post free.

**A BIBLIOGRAPHY OF GUIDES FOR TECHNICAL AUTHORS** has been reprinted from *Aslib Proceedings*, (Vol. 6, No. 4). It is restricted to books which are both readily available at the present time and which, in the experience of the compiler, have been found to offer most practical help to the technical author. Association of Special Libraries and Information Bureaux, 4 Palace Gate, London, W.8.

**INTERNATIONAL JOURNAL OF ABSTRACTS ON STATISTICAL METHODS IN INDUSTRY** No. 2 has recently been published by the International Statistical Institute, 2 Oostduinlaan, The Hague. Price of single copies 7s. 3d., annual subscription for three issues 18s.

**ROHDE & SCHWARZ MITTEILUNGEN** is the house journal of this German company which manufactures communications and laboratory instruments. The sole agents in this country are Aveley Electric Ltd., 44 Tottenham Court Road, London, W.1.

**SILVER PLATING** is a new booklet on the subject produced by Johnson Matthey & Co. Ltd. It is not intended to be an exhaustive treatise on the electrodeposition of silver, but deals in a practical manner with the various operations involved in silver plating, indicating suitable methods for most requirements and giving warnings of possible difficulties. The operation of silver plating may be sub-divided into three main sections—preparation, plating and finishing—and these are considered in the booklet under the appropriate headings. A further section deals with plating troubles, and useful appendices are included. The booklet, the reference number of which is 1851, is obtainable from the company free on request. Johnson, Matthey & Co. Ltd., 73-83 Hatton Garden, London, E.C.1.