

Wireless World

RADIO AND ELECTRONICS

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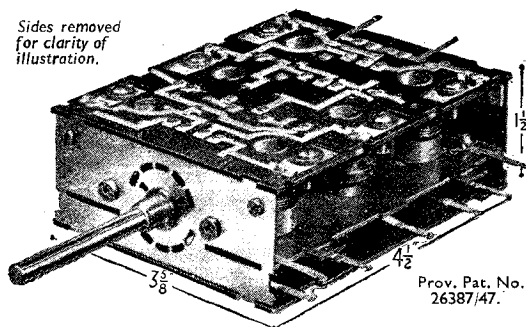
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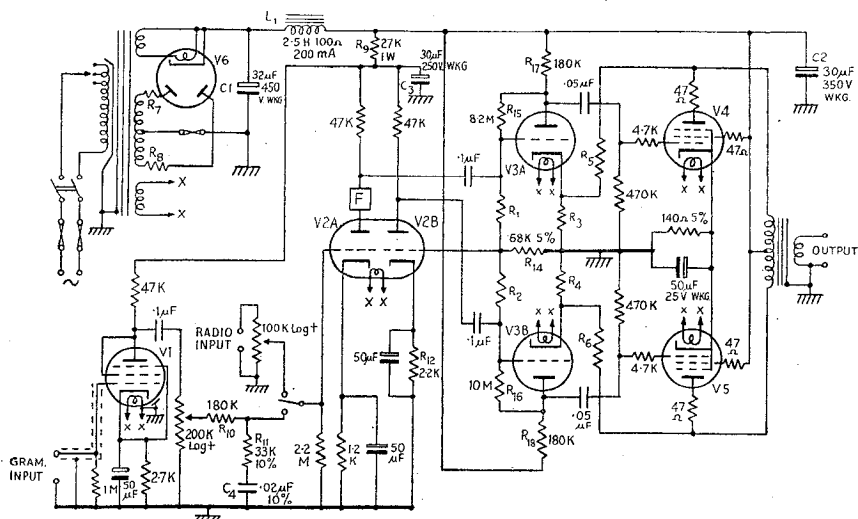
Valves and their applications

HIGH FIDELITY AUDIO AMPLIFIER USING EF37, EGC33 AND EL37

The introduction of wide frequency range gramophone recordings and pick-ups, together with the projected B.B.C. transmissions in the 90 Mc/s. band, means that if the extra fidelity so

made available is not to be wasted, considerable care has to be exercised in the design of the reproducing equipment. One of

and the paraphase valve V2B (ECC33). A low pass filter F. may be inserted in the anode circuit of V2A to reduce surface noise when gramophone records are being played. The two phases are then fed into the two halves of V3. (ECC33) which is inserted into the chain to facilitate the application of degenerative feedback to the two output valves V4 and V5. (2-EL37s). The feedback is direct coupled from each of the output valve anodes by the resistors R5 and R6 back into the cathode circuits of the driver valves R3 and R4. The resistors between the grids and anodes of V3 and in the cathode circuit of V2B are to maintain the correct D.C. operating conditions for these valves, whilst those in the grid screen and anode circuits of V4 and V5 are to stop parasitic oscillations. The power supply is derived from a 350-0-350V open circuit voltage H.T. winding on the mains transformer, the rectifier being a GZ32. Adequate smoothing is provided by the components C1, L1, C2, R9 and C3.



At less than 1% total distortion the full output is 18 watts for 0.3 volts at the grid of V2A or 0.12 volts at the gramophone input terminals. The hum level is more than 60 dB. below 18 W. The frequency response is 0.5 dB. below the 1 Kc/s level at 25 c/s and 12 Kc/s with an output transformer of reasonable design.

the most important items in this is the A.F. Power Amplifier and Gramophone Pre-amplifier.

In a large room or small hall, say between 2,000 and 5,000 cu. ft. in volume, it will be found that the mean level of the electrical input to a normal type loudspeaker is of the order of 50 mW. As the peak amplitudes are 20 to 25 dB. above the mean level it follows that the available power output from the amplifier should be about 15 watts.

It is also necessary that the non-linear distortion is kept to a low level, in particular the high order odd harmonic and inter-modulation products. It is not usually the presence of the higher frequency components which causes annoyance but the products of non-linear amplification, these are invariably present when a pentode output valve is working into an inductive load such as the speech coil of a loudspeaker.

The circuit of a suitable amplifier is shown in Fig. 1. It consists of a Gramophone pre-amplifier stage V1. (EF37) the output of which is fed into a volume control and then into a bass boost circuit R10, R11 and C4 for correcting the recording characteristic. Then follows a voltage amplifying stage V2A.

$R_2=1.22 R_1 \pm 2\% = 220K \pm 10\%$, $R_3=R_4 \pm 2\% = 3.9K \pm 10\%$
 $R_5=R_6 \pm 2\% = 220K \pm 10\%$ (1W), $R_7=R_8=100\Omega$ Total Eff.Res. Mains Transformer H.T. Secy. 350-0-350 v. off load. Output matching load, 4000 Ω anode to anode. Sensitivity :-Radio, 0.3 v. Gram., 0.12V.



Reprints of this report from the Mullard Laboratories, together with circuit notes and further performance data may be obtained free of charge from the address below.

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Wireless World

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RADIO AND ELECTRONICS

Comments of the Month

RADIO PRE-HISTORY

SEVERAL of the historians of radio have commented on the fact—strange to our generation—that some of the earliest radio pioneers dissipated their energies in unprofitable lines of work, and tended to ignore the possibilities of electromagnetic waves for communicating intelligence. Hertz summarily dismissed the whole idea of wireless telephony as quite impracticable: Popov's main early interest was in his "lightning recorder": Tesla gave much time to the still-unsolved problem of wireless transmission of power.

It now seems that Captain (later Admiral Sir Henry) Jackson, the father of wireless in the British Navy, at first failed to recognize its real significance for communications. At any rate, it would so appear from the Report for 1896 of H.M.S. *Vernon*, extracts from which have recently been made available to us by the Admiralty. Jackson suggested that radio emissions should be used for purposes of identification by torpedo boats: as a precursor, in fact, of wartime radar I.F.F. (identification of friend or foe). In fairness to his foresightedness, it should be noted that he quickly changed his views, and undoubtedly it was his persistence in the face of opposition that brought about such rapid development of Naval wireless communications.

SYMPATHETIC CIRCUITS

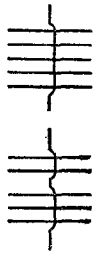
ELSEWHERE in this issue we print an article on the technique of drawing circuit diagrams by an author who has made that subject very much his own. *Wireless World* does not endorse all the detailed proposals made, but does find itself in complete and wholehearted agreement with the underlying principles for which Mr. Bainbridge-Bell stands. Clearly, he believes that a circuit diagram should be something more than a collec-

tion of graphical symbols, grouped more or less at random and joined together as neatly as may be by connections following the shortest path. We assume that, like us, he regards a well-drawn diagram as an aid to understanding the functioning of the circuit concerned, and not merely a graphical record. In view of the increasing complexity of modern circuitry, that is more than ever desirable, especially when the diagram is likely to be studied by those with an imperfect knowledge of all the details which it purports to show. Indeed, we would go so far as to say that a small collection of diagrams, drawn with understanding and sympathy towards the difficulties of the potential user, constitutes almost a textbook in miniature.

Dissected Diagrams

Some twenty years ago, when the inception of broadcasting greatly increased interest in the technical side of radio, this journal published a series of so-called "dissected diagrams" with the object of familiarizing new readers with the graphical and symbolical representation of circuits. What was thought at the time to be a rather trivial contribution soon proved to be almost embarrassingly popular, being obviously considered as an easy short-cut to knowledge, both theoretical and practical. Times have changed, but the advantages of applying the principles which Mr. Bainbridge-Bell discusses are greater than ever.

We differ from him in the detail of "bridge cross-overs," and think he weakens his case by admitting that they may be used for "double security." When a number of leads are to be crossed, the "flyover" is clear, but a better plan is to divide the leads into groups, according to their functions. This answers most objections. With not more than three wires in a group, it is easy to trace any particular one.



COMMUNICATION THEORY

Establishing Absolute Criteria of Performance

By THOMAS RODDAM

THE last months have been heavy with the rumblings of an approaching revolution. True, it is a technical revolution, but it resembles a political revolution in that the thoughts of a few philosophers will set in motion many men who have no understanding of their philosophy. Until recently we have been living under the bene-

units in either radio or wire transmission. The coder and decoder require some explanation. The message itself may be either speech, a picture, or a written message. First of all we shall consider a written message. A typical one would be:—
"Please buy me 1,000 Bongo



Fig. 1. Basic elements of a communication system.

ficient influence of Hartley's Law, which is the engineer's equivalent of "Everything is for the best in the best of all possible worlds." You can read all about Hartley's Law in two articles by "Cathode Ray" (*Wireless World*, June and July, 1947). Unfortunately the recent developments described by him can now be seen to have been steps away from a general communication theory, so that some of the conclusions reached apply only to the special problem of transmitting speech.

To develop the new theory in a simplified form it will first be necessary to treat Hartley's Law briefly in a rather different way from that adopted by "Cathode Ray" (*loc. cit.*). I shall start at the very beginning, because the new theory is the result of a more close examination of the fundamentals of communication, while Hartley's Law is obtained if you gloss over some of the elementary problems. I'm sorry that I shall have to break into mathematics at one point, but the important thing about the new theory is that it enables system performance to be calculated. If the reader wants to know what he has been spared in the way of "sums" he should refer to the *Bell System Technical Journal*, July, 1948.

To begin with, therefore, let us define a communication system. Fig. 1 shows a basic system. The transmitter, medium and receiver may be regarded as conventional

State Loan 3% shares at 94." In ordinary telegraphic practice no one would write this, of course, but would write:—
"Buy 1,000 Bongo 3% 94" . . (A) This change involves what are called the semantic aspects of communication, and is nothing to do with our problem.

If a teleprinter is used, this message comes out as a set of mark and space currents rather like those shown in Fig. 2. ("The actual teleprinter code has not been used here.) Each letter takes up five time units; a separate symbol is used to indicate that figures follow; each time unit is occupied by either a "mark" or a "space." The total coding operation therefore transforms the message (A) into a set of marks and "spaces" of electric current. The first coding, which derived (A) by leaving out some words, is outside our scope, as its efficiency depends on psychological factors. The message shown in Fig. 2 is a standard message type, and it was this sort of message which Hartley considered.

Hartley's treatment was made in the days before we were all aware of the nature of network pulse responses, and it will be a bit clearer if we look at it in post-radar terms. If we pass one mark signal through a band-limiting filter (a low-pass filter if

we consider the "video" circuits) we shall get out a rather distorted pulse, as shown in Fig. 3. We can go on narrowing the band (or lowering the cut-off frequency) until the tail of the pulse is so big that we cannot decide whether we have one pulse or two. The limit is somewhere between (d) and (e) in Fig. 3. If the mark is made longer, of course, we can reduce the cut-off in proportion, because we know that we get quite a reasonable pulse shape if we pass all frequencies up to $1/\tau$, where τ is the width of the pulse: anyone who doesn't agree with that can look up dozens of television and radar papers which discuss this point. Now the speed at which we send our message depends on how long each mark or space must be to pass through the filter, because we need to send a definite number of these marks (from now on I shall often write "mark" to mean *either* an "on" or an "off"). For a given message, therefore, if we double the bandwidth we can make each mark last half as long, and so send all the marks in half the time.

That was Hartley's way of deriving what we have come to know as Hartley's Law. Suppose that we now present ourselves with another piece of information: we measure the actual response of the system from transmitter input to receiver output. This we can



Fig. 2. Fragment of teleprinter message.

do either directly as the response to a pulse, or indirectly through amplitude and phase measure-

ments. For a very short pulse we shall arrive at the response shown in Fig. 4, which is quite a typical curve. By Hartley's method we should not be able to put another pulse into the system until the time corresponding to B was reached. At time X, however, the head of what Brillouin calls the first precursor of the response will have arrived.

After a prescribed time the voltage has risen to A, which gives an amplitude proportional to the input pulse. We can start measuring from the arrival of infinitesimal signals, because we assume that the system is free from noise. We now know the whole characteristic of the curve (b), for the shape is settled by the system response, and the scale factor is settled by our measurement at A.

We can then construct a local circuit to generate the second waveform shown in Fig. 5. This waveform is such that when added to the received waveform it cancels it exactly at all times later than C. The output then becomes that shown in the third line of

simply that we must know the response of the system and the size of the input pulse with increasing precision as we speed up the operation. And, of course, the network which generates the cancelling waveform becomes more complicated. We do know a bit about networks for this job, however, because in some ways the problem is the same as that of cancelling "permanent echoes" in a radar system. The one thing which has enabled us to take this additional step is that we are assuming that we can predict the future exactly. As soon as we introduce noise, we lose this power of exact prediction, and the solution found here is no longer valid: it will be more convenient

the effect of noise is, we code our message in a different way. Let us take the original message, and

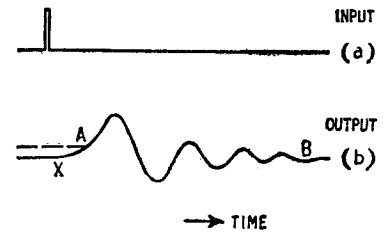


Fig. 4. Response of system to a very short pulse.

code it by numbering each letter:
 B U Y B O N G O
 2 21 25 2 15 14 7 15
 The message can then be sent in the form shown in Fig. 6, so long as the minimum level used is greater than that of noise. Now the amount of information in the message is dependent on the number of mark signals sent, and on the number of possible sizes of each mark signal. In fact, if we write for the "size" of the message, L for the number levels and n for the number of marks

$$M = L^n$$

n is now proportional to the product of bandwidth \times time, since noise prevents us using the trick we used before to get round the Hartley relationship. We can follow Gabor and write $n = \frac{1}{2}BT$, or we can absorb the $\frac{1}{2}$ into M by redefining the "size" of the message.

L depends on the signal-to-noise ratio, and is equal, in the limiting case, to $(1 + S)$, where S is the signal/noise ratio. If the receiving device works on peak voltages we take (peak signal)/(peak noise): if it works on energy we take (r.m.s. signal)/(r.m.s. noise). Finally, however, we have $M \propto (1 + S)^{BT}$

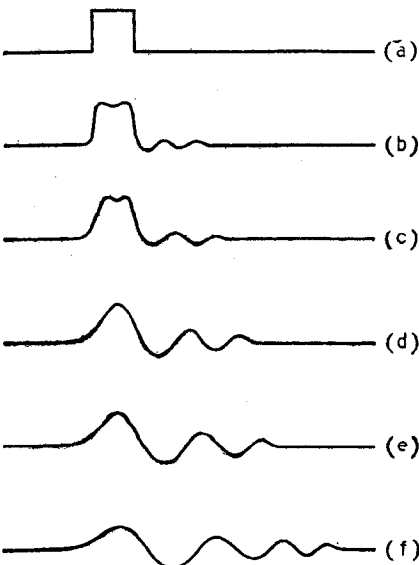


Fig. 5, and we can apply to the circuit a new pulse for which the new point A arrives at time C. The bandwidth in the actual channel has not been increased, but we now assume that we know exactly what the amplitude of the input pulse and the response of the channel are. The closer we make the matching of the compensating waveform, and the more sensitive we make the detector, the nearer together can X, A and C be brought, so that we can increase the number of pulses indefinitely. We can thus send as much information as we choose, in as short a time as we choose, using as narrow a bandwidth as we like. The sky, in fact, is the limit. The price we pay is

Fig. 3. Distortion of pulse in passing through a filter.

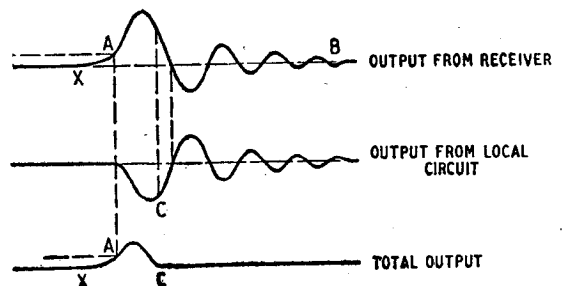
to discuss the effect of noise from a rather different standpoint, however.

We can see now why Hartley found it hard to get a numerical constant to equate to the product "bandwidth \times time": there just isn't one. Gabor has arrived at the value $1/2$, which depends on the application of the Hartley method to a transmission system having a Gaussian frequency response. The objections to this are, first that no physical system can have exactly a Gaussian response and secondly that anyway, such a system has its amplitude response defined over an infinite band, the "bandwidth" term being simply the bandwidth at half-amplitude.

The new theory does not stop at the point reached above, which is, in its own way, as limited as the Hartley treatment. The presence of noise must always be assumed in any real communication system, and looking at Fig. 4 again we can see that we cannot move A too near to X, or we shall not have enough signal to override the noise. To see what

EFFECT OF NARROWING BANDWIDTH

Fig. 5. Cancellation of pulse distortion.



Communication Theory—

First of all, we shall do a little mathematics using this expression. If we have two systems, with bandwidths B_1 and B_2 , we can obtain the same value of M for the same time if

$$(1 + S_1)^{B_1 T} = (1 + S_2)^{B_2 T}$$

where S_1 and S_2 are the two signal/noise ratios. If we take $B_2 = kB_1$, $(1 + S_1)^{B_1 T} = (1 + S_2)^{kB_1 T}$ so that $(1 + S_1) = (1 + S_2)^k$.

Suppose, for example, that the value of S_2 is 3. For an increase of bandwidth by a factor of 4, the signal-to-noise ratio S_1 is given by

$$(1 + S_1) = (1 + S_2)^4 = 4^4$$

$$S_1 = 255$$

The increase of bandwidth has raised the signal-to-noise ratio from 9.5db to 48db. If the bandwidth had been increased by a factor 5, the signal-to-noise ratio would have been increased to 61.6 db. This corresponds to the bandwidth increase used in ± 75 db deviation in f.m. broadcasting, which gives only about 18db improvement, the corresponding value of signal-to-noise ratio being then 28db.

For the benefit of those who suspect the mathematics, I shall show how we can move the message into a wider band, at the same time reducing the required signal-to-noise. Our original message was

BUY BONGO which we wrote as

$$2, 21, 25, 2, 15, 14, 7, 15$$

We can rewrite this in the scale of two, thus

$$00010, 10101, 11001, 00010, 01111, 01110, 00111, 01111.$$

In this, the digit abcde = $a \times 2^4 + b \times 2^3 + c \times 2^2 + d \times 2 + e$.

We could, if we liked, write it in the scale of 3, as

$$002, 210, 221, 002, 120, 112, 021, 120, \text{ in which}$$

$$abc = a \times 3^2 + b \times 3 + c.$$

In Fig. 6(b) the message is shown, coded in the scale of 3, and arranged to occupy the same time as in Fig. 6(a), which is a 26 step system. It is easily seen that the message requires three times the bandwidth, since three times the number of steps are to be transmitted. For the same peak amplitude of signal, however, the noise can be more than ten times as great.

So far we have only considered the transmission of telegraphic messages. We can apply this to

telephony by the technique called "pulse code modulation," which was described in a previous article. In simple terms, what pulse code modulation does is to send a string of messages which enable the receiver to plot the waveform of the speech to the desired accuracy. It is therefore possible to obtain these enormous improvements in signal-to-noise ratio for speech, or music, or television. Unlike

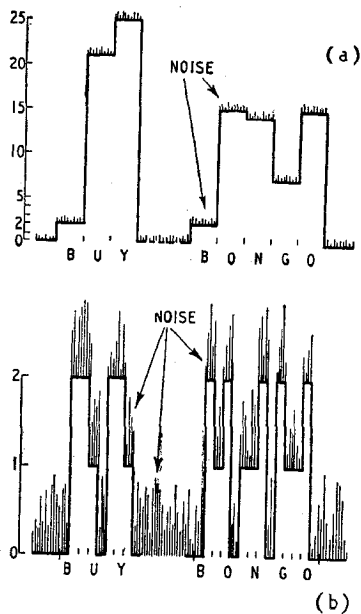


Fig. 6. Message coded (a) in a 26-step system, (b) in a scale of 3. The latter can tolerate a higher noise level.

the Vocoder, the system does not depend upon the special character of the signal, which must be speech to operate the vocoder.

If we want, as we often do, to reduce the bandwidth, we find the situation is rather unpleasant. Suppose that we wish to transmit a signal in $1/3$ of the normal bandwidth—that is, we want to cram speech into 1,000/s. We need a final speech-to-noise ratio of 40db, at least, so that we must have a value of S_1 given by $(1 + S_1) = (1 + S_2)^3 = 100^3 = 1,000,000$, or $S_1 = 120$ db. We need, then, a power increase of 80 db, so that instead of sending out, say, 1 milliwatt we shall need 100W. Clearly, there is not much value in reducing bandwidth at such an enormous cost in power.

The importance of the new theory lies in the fact that we

now have a way of judging modulation systems in terms of their efficiencies relative to an ideal system. In the past we have always had to express the performance in terms of another system, so that we have said, for example, that f.m., with such and such deviation ratio, gives an improvement of so many decibels over amplitude modulation. Now we can say that f.m., with such and such deviation, gives a signal to noise ratio so many decibels below ideal. We can also see just how much more we can hope to gain by the use of systems which approach the ideal more closely. It may not be profitable to make use of these systems: we have seen that band compression is incredibly extravagant in power, so that it will never be adopted. We can, however, get down to the job of finding the cheapest way of providing a given signal-to-noise ratio at the receiving end.

One consequence should be a reconsideration of the policy of adopting frequency modulation for local broadcasting. We want to provide high-quality programmes at a minimum cost to the whole nation. If we have a million listeners, it is worthwhile to spend an extra £100,000 at the transmitters if we can save 5/- in the cost of each receiver. I have, in the past, urged a closer study of the possibilities of pulse transmission, especially if several programmes are to be radiated. It is most important that a fuller study should be made of the whole problem, especially from the point of view of national economics, not merely to find the policy which involves the least expenditure of B.B.C. money. The money all comes from the same place in the end. It is not impossible that the answer may turn out to be very high level amplitude modulation, say, 500kW. That sounds like a lot of power, but if it only amounts to 1 watt per listener, it can be saved by eliminating only one valve from a receiver.

LOWER-POWERED "BUSINESS RADIO"?

When the organization of e.h.f. "private" radio-telephone services was recently discussed by the Radio Section of the I.E.E., it was suggested that in many instances the licensed power was unnecessarily high, and should be reduced in order to lessen interference.

PLANAR ELECTRODE VALVES FOR V.H.F.

Reducing Interelectrode Capacitance and Transit Time

(Contributed by the Research Staff, M.O. Valve Company)

DURING the past ten or fifteen years considerable progress has been made in improving the high-frequency performance of triodes and pentodes by reducing the inductance of the leads to the electrodes. One of the first attempts in this direction was the "acorn" valve, which was designed with a very small electrode system, the leads from which projected as radial pins passing through the all-glass envelope. It is interesting to note that the earliest forms of this type of valve employed planar electrodes¹ similar in some respects to those which will be mentioned later. However, this construction was abandoned in favour of a very small cylindrical electrode system when "acorns" were eventually produced and marketed. The "acorn" type of valve, while enabling a considerable improvement to be obtained in the effective amplification at very high frequencies, has proved to be a difficult manufacturing proposition and has been superseded by valves with conventional electrode systems, mounted on flat glass bases through which pass the lead-out wires, which themselves form the valve pins. Two forms of such designs are represented in present-day commercial products in the button seal pressed-base valves, commonly known as the miniature, and the ring seal moulded-base type. In all these valves the electrode lead-out wires themselves form the connecting pins and the necessity for an external base with separate pins has been obviated.

These glass-based valves represent a big step forward in valve design, and there seems little doubt that the majority of receiving valves in the future will be mounted on this form of base. Quite apart from the advantages of this construction for high-

frequency operation, it has led to a reduction in size and freedom from loose base troubles, which, under some conditions, occur with the cemented plastic base. Furthermore, with large-scale production the cost of manufacture of some forms of pressed glass base valves may be less than with earlier designs. Fig. 1 shows an "acorn" valve, a modern valve on a pressed-glass base and a valve mounted on the conventional glass "pinch," a feature which owes its origin to the electric lamp.

In a wide-band amplifier it is normal for the dynamic resistance of the circuits to be of a comparatively low order and several considerations arise in the design of a suitable valve for high gain combined with low noise in such amplifiers.

The gain of a single stage of a wide-band amplifier is proportional to the ratio of the mutual conductance (g_m) to the sum

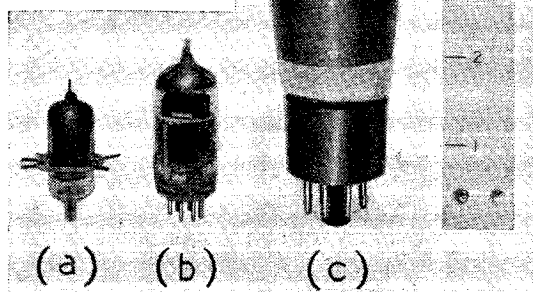


Fig. 1. Types of valve construction (a) "acorn," (b) pressed-glass base, (c) conventional "pinch" seal and moulded base.

of the input capacitance (C_i), the output capacitance (C_o) and the stray capacitances (C_s). It is important therefore to make this ratio as high as possible. In addition, for successful high-frequency operation the interelectrode capacitances should be kept small, in order to keep as much as possible of the circuit external to

the valve, and the electron transit time should be reduced to a minimum.

Now it can readily be shown that the requirements of high ratio of mutual conductance to capacitance and of low electron transit time require a high ratio of electron current density to grid-cathode spacing. The further requirement of low interelectrode capacitance necessitates a small cathode area. Thus the best performance is likely to be obtained with a valve having a small cathode area, small grid-cathode spacing and operating at a high current density.

The ultimate sensitivity of a high-gain amplifier depends on its signal-to-noise performance. If the gain of the first amplifier stage of a receiver is more than about

5 db then most of the noise output is contributed by the first stage. The amount of noise contributed by a valve is usually regarded as being equivalent to that generated in an imaginary resistance, R_n , in the grid circuit of the valve. R_n is known as the

"equivalent noise resistance" of the valve and is approximately inversely proportional to the mutual conductance. If R_1 is the dynamic resistance of the input circuit, then it can be shown that the signal-to-noise ratio is a function only of the ratio R_1/R_n and will increase as this ratio increases. Now R_1 cannot be increased in-

¹ "Vacuum Tubes of Small Dimensions for Use at Extremely High Frequencies," B. J. Thompson and G. M. Rose, *Proc. I.R.E.*, Vol. 21, p. 1707, 1933.

Planar Electrode Valves for VHF—definitely owing to the inherent losses in circuit components so that the only way to improve the signal-to-noise performance is by reducing R_s and this means increasing the mutual conductance of the valve.

For frequencies above a few hundred megacycles per second a greater decrease in lead inductance proves necessary than has been achieved in the conventional concentric cylindrical arrangement of electrodes, and this improvement has been achieved by making the electrodes integral with metal discs which pass through the envelope and which may be directly connected to cavity resonators if desired. Such valves have been described elsewhere.²

These valves are known as the disc-seal type and such are capable of operation at frequencies up to about 4,000 Mc/s. The valves employ planar electrodes which allow very small interelectrode spacings to be achieved, permitting a high mutual conductance from a small cathode area and a high ratio g_m/C_{g-k} .

An example is the Osram and Marconi disc-seal triode type DET 23 in which the mutual conductance is 7.0 mA/volt at an anode current of 10 mA, and the total input and output capacitances including the discs which pass through the envelope are

2.4 pF and 1.1 pF respectively, of which the discs themselves account for about 0.7 pF in each case. Thus:

C_{g-k} is 1.7 pF and C_{a-g} is 0.4 pF. This high ratio of mutual conductance to input capacitance is better

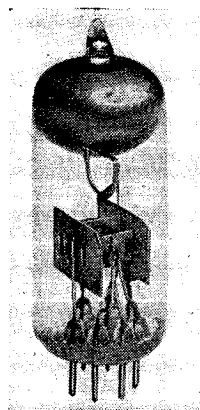


Fig. 3. Experimental parallel electrode triode (E1714) on pressed glass base.

than has hitherto been achieved with concentric electrode arrangements, and is due to the fact that the spacings are small only at the operating surfaces of the electrodes.

These disc-seal valves (illustrated in Fig. 2) which were designed primarily for ultra-high frequencies will be seen to satisfy the wide-band amplification requirements set out above. It therefore seemed desirable to employ a similar electrode arrangement in valves designed for more general use in the u.h.f. range, such as valves mounted on pressed glass bases with the pins forming the lead-in wires. Valves of this type are easier to use and less costly than the disc-seal valves.

A typical triode of this class is the experimental type E1714 and is illustrated in Fig. 3.

The very small grid-cathode spacing employed (0.003 in) necessitates the use of extremely fine and closely spaced wires for the grid, and the design of the grid (Fig. 4) is one of the principal features of valves of this type. In the conventional type of electrode system in which the grid wires are located on two separating rods the wires themselves must be sufficiently strong to carry the separate rods so that the whole structure is rigid enough for handling during the assembly of the valve electrodes without risk of distortion, and this sets a lower limit to the diameter of wire which can be employed. In planar electrode valves a departure from convention has been made, which enables rugged grids to be manufactured with wires as small as 0.0006 in.

The grid is in the form of a metal

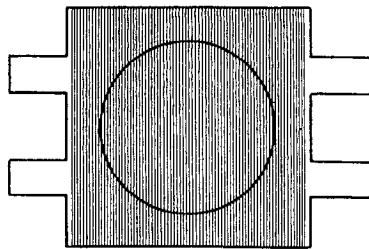


Fig. 4. Grid assembly of planar-electrode valve.

plate pierced by a circular aperture across which the grid wires are stretched, while the cathode and anode are the end surfaces of two short cylindrical members, supported from or integral with a relatively thick and therefore rigid plate. These plates and the grid frame are located in slotted mica bridges which serve to hold the electrodes in the correct relative positions. Stray capacitances between the electrodes are in this way reduced to a minimum, only the operating surfaces of the electrodes being in close proximity. The leads connecting the electrodes to the pins in the valve base are also well spaced and contribute little to the total capacitances. The electrode assembly for this type of valve is shown in Fig. 5.

The very small diameters of grid wire possible with this construction allow adequate grid dis-

² "Triodes for Very Short Waves," Bell, Gavin, James, Warren, *Journal I.E.E.*, Vol. 93, Part IIIA, p. 833, 1946.

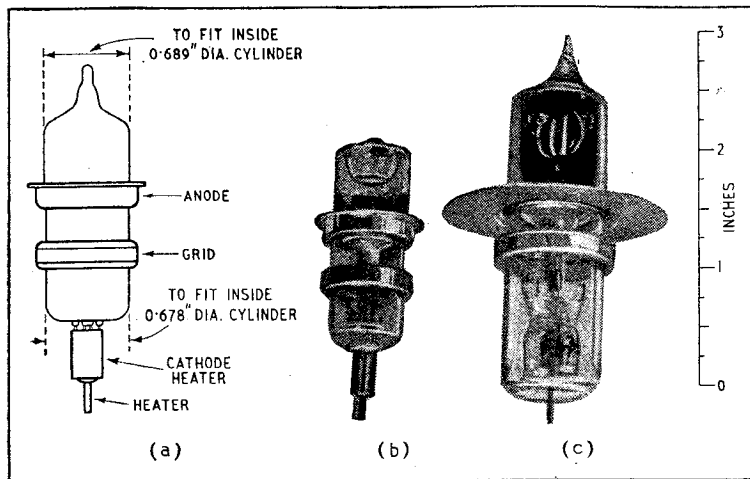


Fig. 2. Examples of disc-seal triodes (a) outline of DET23, (b) E1599, (c) E1368.

sipation for amplifiers and for low-power oscillators. Furthermore, the grid frame serves to

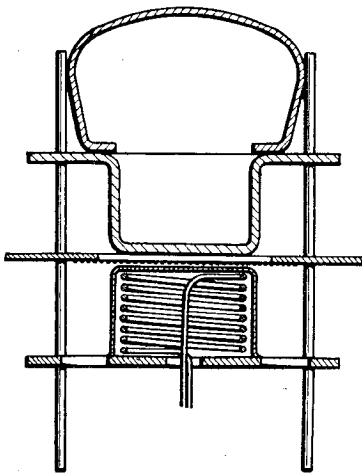


Fig. 5. Electrode assembly in the type E1714 triode.

radiate heat and thus minimizes the risk of primary grid emission.

The characteristics of the E1714 are as follows:—

- Filament voltage.. 6.3
- Filament current.. 0.5 amp
- Anode voltage .. 250 max
- Amplification factor 40
- Mutual conductance 8.0mA/V measured at anode voltage 150 and anode current 10 mA.

	Capacitances with cathode cold:	Capacitances with cathode hot (I _a = 10mA)
C _{g-k}	1.6 pF	2.9 pF
C _{g-all (except anode)}	2.6 pF	3.7 pF
C _{a-g}	0.9 pF	—
C _{a-all (except grid)}	1.1 pF	—

Equivalent noise resistance 500 ohms (I_a = 10mA).

These characteristics undoubtedly represent the best performance which has been obtained with a triode operating at frequencies of the order of 45 Mc/s, covering a bandwidth of 10/15 Mc/sec.

Coaxial / waveguide transformations matching 70-80-ohm lines can be made in a variety of forms, and standardized markings are used to distinguish power inputs and outputs. Among the components available are connectors, adaptors and bushes, loop-probe junctions, tuning plungers, matching stubs and crystal detector units. Measuring instruments include a bolometer in a bridge circuit covering 100 mW to a fraction of a milliwatt over a frequency range of 100 to 10,000 Mc/s, line attenuators using "Caslite" iron-dust cores, a piston attenuator for the non-dissipative "E" mode with a micrometer head calibrated directly in db, and coaxial-line wavemeters with ranges up to 20 and 40 cm

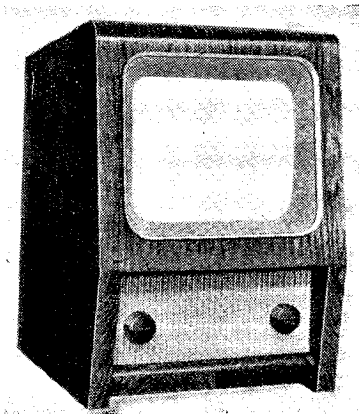
Television E.H.T. Supply

A n.e.h.t. supply unit with an output of 5-8 kV at 300 μA has just been produced by Haynes Radio, Queensway, Enfield, Middlesex. It is of the r.f. oscillator type. A 6V6 valve is used as a 100-kc/s oscillator and draws 28 mA at 300 V. Rectification is by an EY51 which

MANUFACTURERS' PRODUCTS

H.M.V. Transformerless Television Receiver

THE new 1807 table model is of the transformerless type and suitable for use on a.c. or d.c. sup-



H.M.V. Model 1807 television receiver.

plies of 220-250V. A 10-in tube, with an aluminized screen, is used and operated at 5.5 kV, the supply being obtained from the line fly-back. A permanent magnet is used for focusing and adjustment of focus is obtained by varying the e.h.t. supply by changing the fly-

back conditions. The picture size is 9in by 7in.

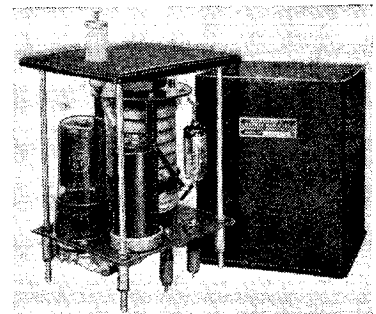
A metal rectifier is used to provide h.t., but a valve rectifier with its filament heated from the line-scan transformer is used in the e.h.t. circuit.

The receiver is of the straight type and of moderate sensitivity; for extreme range the addition of a pre-amplifier is recommended. The panel controls are Sound Volume and Picture Brightness, the on-off switch being combined with the latter. The set measures 19½in high by 19in deep by 13½in wide and weighs 30lb. The price is £37 16s plus £8 12s purchase tax.

Standardized E.H.F. Components

A NUMBER of coaxial line components and measuring instruments for centimetre and decimetre wavelengths with standardized interconnections has been introduced by the Plessey Company, Ilford, Essex.

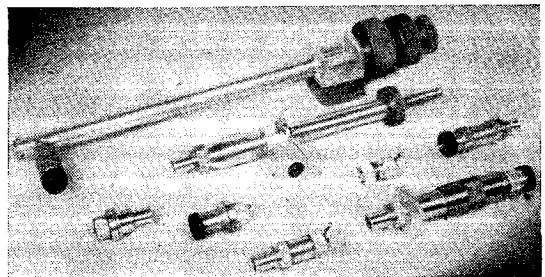
Representative components and instruments in the Plessey e.h.f. standardized range.



Haynes Radio R.F. E.H.T. unit, type 828.

has its filament heated from the r.f. coil.

The output is controllable below the maximum of 8 kV by reducing the h.t. voltage applied to the unit. The reservoir capacitor is of 0.001 μF only, so that a dangerous shock can hardly be obtained. The unit is completely screened and costs £5 8s.



SINGLE-VALVE FREQUENCY-

2.—Practical Details of Design and Use

By K. C. JOHNSON, B.A.

IN last month's issue it was shown that it is possible to obtain electronic frequency modulation of an oscillator if an unusual circuit is employed with the tuning coil and condenser connected in series in the cathode circuit of a pentode valve, and with a second mutually-inductive coil carrying the anode current. The effect of the second coil is then simply to change the effective inductance of the first, and so the resonant frequency of the tuned circuit, as the suppressor grid voltage of the valve is varied and the fraction of the total cathode current which flows to the anode is changed. By this arrangement it is possible to obtain frequency modulation over ranges of as much as 30 per cent, using either of the circuits shown in Fig. 1.

In this article it is intended to discuss the many practical details which arise in the design and use of these circuits as "wobblers" for receiver alignment.

The two-valve circuit shown in Fig. 1 (a) has several advantages over the single-valve version, for use in elaborate signal generators or as the local oscillator in panoramic superhet receivers, where the phase-inverter valve can conveniently be the triode section of a normal frequency changer. However, for a simple unit working on a fixed central frequency the single-valve version is more economical and can be made to give an almost equally good performance in range and constancy of amplitude.

Unlike reactance valve arrangements, these two-valve oscillators give practically constant amplitude over wide ranges without any difficulty, since the frequency-modulation mechanism would not be expected to affect the loop gain, and, moreover, there is a strong limiting action, since the peak oscillatory current cannot exceed the mean current through the valve. The single-valve circuit, however, is not quite so good in this respect, since there must inevitably be some change of gain with frequency due to resonance in the phase-inverter coil; but with careful coil design this need

only cause a fall of about 10 per cent in amplitude at the extreme ends of a range as great as 30 per cent in frequency.

For most ordinary purposes, such as the alignment of i.f. band-pass circuits in broadcast receivers, a coverage of 20kc/s at 1 Mc/s is adequate, so that the amplitude even of the single-valve circuit will be practically constant. The linearity, also, will be practically perfect, since the voltage swing on the suppressor grid need be no more than two volts, or one-fifteenth of the total grid-base. The most generally useful oscillator, then, will be designed to have a fairly wide frequency range, even if only a small fraction of it is actually required, so as to get linearity and constant amplitude.

Valves.—It would appear at first sight that the natural choice of a valve for use in this circuit would be one of the new "suppressor-slope" pentodes which are now available, but although these have the great advantage that their suppressor grids are made to close tolerances, they are not the best valves for the purpose. This is because the minimum screen current is much greater than in ordinary pentodes, so that the available range of current division is much less, and also because the high suppressor sensitivity means that the small, but inevitable, voltage swing on the cathode will affect the current distribution between screen and anode.

The valve chosen is the EF50, which has a suppressor grid with a moderate sensitivity and made to definite tolerances, but almost any r.f. pentode can be used if the suppressor connection is available. As already described the linearity of the valve is not important when only a small range is required, but the EF50 does in practice give quite reasonable linearity over the whole range of control.

It must be remembered that in these circuits the valve may easily

be run with the entire cathode current flowing to the screen grid and care must be taken that the h.t. supply voltage does not exceed about 180 volts. The cathode resistance is used to provide automatic bias for the suppressor grid in the usual manner, and the value for the EF50 is normally 2 k Ω , although it is convenient to use a 5 k Ω potentiometer so as to obtain a "d.c." frequency-shift control. This resistance is necessary also to carry the steady valve current and to avoid short-circuiting the tuned circuit; but it will be noticed that the tuned circuit behaves as an ideal bypass condenser at the oscillation frequency, so that the voltage swing at the cathode is actually extremely small and there is no need to put filters in the heater leads unless unusually good screening is required.

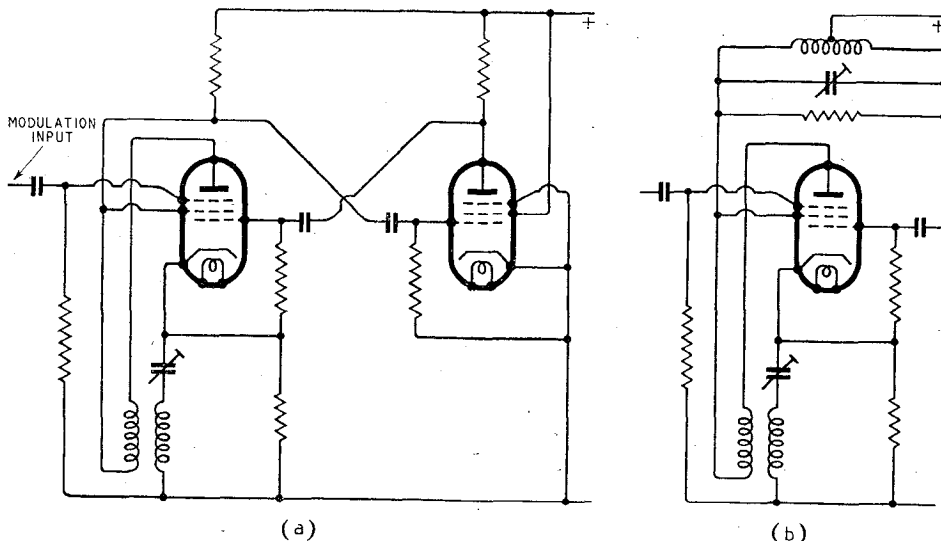
Tuning Coil.—The main tuning coil must be designed so that the frequency range available is as large as possible. This means that the mutual inductance between the two windings must be made negative and large so that it subtracts a maximum amount from the self-inductance. The self-capacitance of both coils must also be kept small so that there is no chance of the anode coil resonating even at the highest frequencies, and so that the maximum amount of the current in the cathode coil flows through the valve. The capacity between the windings must also be small, but this is not important if the tuning condenser is connected at the cathode end of the coil, so that the "dead" ends of the two windings come together. This has the additional advantage that the tuning condenser can then be used as the h.t. bypass and the windings of the coil need not be carefully insulated from each other. If, however, it is desired to use a variable tuning condenser with an earthed frame, the coil windings must be insulated and to avoid

MODULATED OSCILLATORS

capacity effects the connections will have to be reversed so that the mutual inductance is positive. The centre of the tuned circuit and the valve anode are the "hot" points where capacity must be avoided, but if the coil is so arranged that these voltages are in phase and roughly equal they can be close together in the winding without any serious effects. It will also be noticed that the valve anode impedances must be kept high to reduce damping effects, and this is assisted by bypassing the suppressor to earth at radio frequency with a small condenser.

The actual coil used for 1 Mc/s is wound on a $\frac{1}{2}$ in diameter former with an iron-dust core, and each winding is a layer of 100 turns of close-wound 34 s.w.g. enamelled wire, the second being wound directly on top of the first, spiralling in the same direction. If iron cores are not obtainable it is possible to use a similar design of air-cored coil with 120 turns of 38 s.w.g. enamelled in each layer, but this does not give such a good frequency coverage,

Fig. 1. (a) Frequency-modulated oscillator with valve phase-inverter. (b) Single-valve version with a damped auto-transformer phase-inverter.



since the iron increases the mutual-inductance in a greater proportion than the self-inductance and enables the self-capacitances to be reduced.

Phase Inverter Coil.—In the two-valve circuit the second valve serves simply to give a phase-inversion with a slight gain, and unless it is desired to have a variable tuning condenser or multi-range switching, the valve can quite satisfactorily be replaced by

a tuned and damped auto-transformer. This must be adequately damped, however, so that the phase-shift and amplification remain nearly constant over the frequency range, and this can only be achieved by using a good coil of high L/C ratio and shunting it with a low resistance. This coil must be fitted with either a variable iron core or a normal capacity trimmer, and this must be adjusted until the amplitude falls off equally at either end of the sweep range, or so that the total valve current is a minimum, but this adjustment is not critical.

For 1 Mc/s a wave-wound iron-cored coil of 75 turns of 34 s.w.g. tapped at 25 turns from the "anode" end on a $\frac{1}{2}$ in diameter former is suitable, and 1,000Ω is a satisfactory damping resistance for an EF50, though this would have to be increased for a valve

from the two-valve circuit can most conveniently be taken from a tapping on the anode load of the second valve, and the low impedance which is available makes the design of an attenuator comparatively easy. The single-valve circuit is not so convenient, however, and the output must be taken, at much higher impedance, from the "anode" end of the phase-inverter coil or from a tapping on it. In any case the oscillator unit must be placed in a screened box to avoid radiation and interference with other receivers, since owing to the limiting action of the valve, harmonics as well as the fundamental are generated and may be radiated strongly.

For use in routine bandwidth adjustment of broadcast receiver i.f. amplifiers the frequency-modulated output is taken at low impe-

dance to the frequency changer grid so as to avoid effects due to the preselector coils, and the 1 Mc/s signal is tuned-in in the usual way. The a.v.c. bias must be shorted and the signal at the diode-load volume control, or other suitable point, taken to the Y deflection of an oscilloscope. The X deflection of the oscilloscope and the modulation input of the wobbulator are then both connected to the 50 c/s mains, and

Output Circuits.—The output

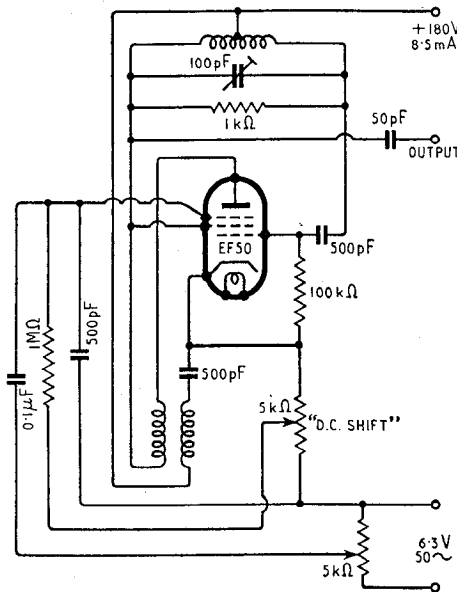
Single-valve Frequency-modulated Oscillators—

the set response curve will be obtained.

It will be noticed that the trace obtained on the oscilloscope is not quite the same in each direction. This is because it takes a definite time for the signal amplitude to build up in each tuned circuit, and unless the scanning is infinitely slow this will tend to make the second of two equal humps look higher. It can be shown that to obtain a "resolving power" of n c/s the rate of scan must not be greater than n^2 times per second, so that if a range of 20 kc/s is scanned 50 times each second it is only possible to distinguish two humps if they are more than about 1 kc/s apart. In practice this is more than sufficient for most purposes, but it is essential to use a sinusoidal scan and see "both sides of the picture" so as to be able to eliminate the distortion caused by the lag in building up the signal, which is far from negligible.

When the i.f. amplifier has been adjusted to any desired response characteristic the oscillator can be connected to the aerial terminal and the pre-selectors adjusted for maximum signal by trimming and padding in the usual way. One advantage of using a wobulator at 1 Mc/s, rather than at the i.f., is that it can be used without alteration for any medium-wave set, and another is that a very rapid estimate of the pass bandwidth can be obtained

Fig. 2. Completed circuit of the single-valve wobulator capable of a frequency deviation range of at least 30 per cent.



simply by tuning the receiver and watching its wavelength scale whilst the response curve moves its own width across the screen.

Practical Performance.—Fig. 2 shows the circuit of the single-valve wobulator unit with all the component values and the details of the arrangements for obtaining a sweep of variable width and variable central frequency. This

"d.c. shift" control is very convenient in practice, and it has the additional advantage that it makes it very easy to adjust the phase-inverter tuning by means of the current variations over the range. The two-valve equivalent of this circuit can be easily visualized, and it need only be said that the load in the first valve should be no greater than 50Ω , whilst an EF50 in the second stage will give sufficient amplification with an anode load of $1,000\Omega$. The single-valve circuit shown in Fig. 2 will give a frequency deviation range of at least 30 per cent with very nearly constant amplitude and reasonably good linearity.

There is no reason at all why this circuit should not be used for television receiver alignment at 45 Mc/s, but unless the experimenter possesses a tunable receiver for these frequencies it will be found to be almost impossible to check the operation of the oscillator. The author has, however,

coverage of 2 Mc/s at a central frequency of 11.25 Mc/s, but the amplitude variations of the fourth harmonic, which swept the whole television frequency band, could not be examined.

Clearly this is only one of the many interesting possibilities which this new principle offers and which remain to be developed. Some others which suggest themselves are simple wide-band panoramic or remotely controlled receivers working on either the superregenerative or synchrodyne principles, and single-valve portable f.m. transmitters, but there are many more possible applications and it would be impossible to discuss them fully in this article.

MOON ECHOES

New Method of Ionosphere Research

INVESTIGATIONS of the transmission characteristics of the F-region of the ionosphere, making use of radio echoes from the moon, are in progress in Australia; they are reported by Kerr, Shain and Higgins in the February 26th, 1949, issue of *Nature*. Arrangements have been made with the Postmaster-General's Department, by the Division of Radiophysics, Department of Scientific and Industrial Research, Australia, to have the use of transmitters VLC9 (50 kW, 17.8 Mc/s) and VLB5 (70 kW, 21.54 Mc/s) during periods when they are not in use for beamed transmissions to the U.S. and Canada.

As the aeriels are fixed, it is necessary to wait for the moon to pass through the beam before making observations, but it has been found possible to carry out experiments on about 20 days in the year.

The receiver is an R.C.A. Type AR88 used in conjunction with a rhombic aerial system and both aural and c.r. tube observations of the echoes are made. By using a pulse length of 2.2 sec, short-term fluctuations of the returned signal have been studied, and particular attention is being paid to a comparison of the observed maximum angle of incidence on the F_2 layer for penetration, with the angles calculated from current ionosphere theory. It appears that the transmission through the ionosphere in different directions follows different paths, and that this lack of reciprocity could arise from the effect of the earth's magnetic field.

It is expected that the new technique will prove superior to observations of solar noise for exploring the higher levels of the ionosphere.

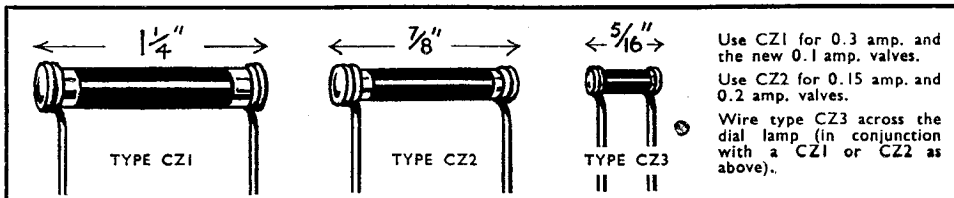
experimented with a circuit using a single EF50, a main coil with two layers each of 15 turns of close wound 30 s.w.g. on a $\frac{3}{16}$ in diameter air-cored former and a phase-inverter coil using 30 turns of the same wire on a similar former tapped at 10 turns. Tuning these coils with about 70 pF and 10 pF respectively it was found to be possible to get a

BRIMARIZE that "Switch-on" Surge!

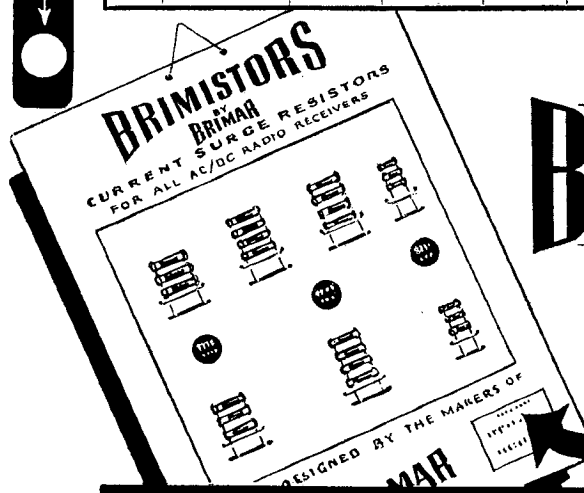
Eliminate the switch-on surge in AC/DC receivers with BRIMISTORS. Wired in series with the valve heaters, a BRIMISTOR holds the starting current at a low value, allowing it gradually to rise to its working level, so eliminating the rush of current which shortens the lives of dial lights and valves.

When cold, a BRIMISTOR has a high resistance and this falls to a very low value as it warms up. Generally this low resistance can be ignored and at the most calls for a slight reduction in the resistance of linecord or other voltage dropper.

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TYPE	COLD RESISTANCE	RESISTANCE WITH THE FOLLOWING CURRENTS FLOWING				LIST PRICE	
		0.1 amp.	0.15 amp.	0.2 amp.	0.3 amp.		
CZ1	3000 ohms	180 ohms	100 ohms	75 ohms	44 ohms	3/6	1. Owing to the high operating temperature (up to 250°C.), Brimistors must be spaced away from coils and waxed components. 2. They should be inserted in the "live" end of the heater chain—i.e., between mains resistance and rectifier valve heater. 3. At least 1/2" of wire must be left at each end before soldering to a tag.
CZ2	5500 "	170 "	90 "	66 "	38 "	.2/6	
CZ3	1500 "	100 "	50 "	35 "	Max. Current 0.2 amp.	1/6	



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