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EDITORIAL

Some Ohmic and Non-Ohmic Reflections

THE October Editorial, "Ohm and his Law," has led to some interesting and very varied correspondence. Our attention has been drawn to the suggestion in Hart's "Makers of Science" (1923) that it was Fourier's classical book on heat, published in 1822, that gave Ohm the idea on which his work was based. Now it would be quite impossible for anyone who studied Ohm's original publications to agree with this suggestion. In the introduction to the "Collected Works of Ohm," Lommel (1892) mentions the legend that had grown up in France, due originally to Pouillet, who believed that he was the first to establish Ohm's law experimentally in 1831. Not knowing of Ohm's earlier experimental work, he thought that Ohm had only established the law mathematically on hypothetical assumptions, based presumably on the work of Fourier. Lommel states that this baseless legend was not only found in French publications but even in some German text-books.

Ohm's original approach to the subject was purely experimental and, as the result of inaccurate observations, his first published statement of the law was logarithmic and quite wrong. As we pointed out, it was only when he repeated his experiments with thermo-junctions instead of batteries as the source of electromotive force that he was led to the correct statement of the law. Ohm is very generous in his references to any earlier work and gives credit to Davy, Barlow, Becquerel and to the "immortal Laplace." He was acquainted with Fourier's work and there is little doubt that it influenced the mathematical treatment, for in the introduction to his 1827 paper "Die galvanische Kette, mathematisch bearbeitet," he says, "The form and treatment of the differential equations so obtained are so similar to those given by Fourier and Poisson for the transfer of heat that one has every right to draw the conclusion that there is some internal connection between the two phenomena, and this "identity relationship" increases the further one pursues it." Again, at a later stage he points out that his views on the molecular action are in agreement with those propounded by Laplace and Fourier in the theory of heat.

Non-Ohmic and Non-Linear

With regard to the use of the term non-ohmic, one correspondent says that ohmic means "measured by the ohm," and that it is therefore impossible for a resistance to be non-ohmic. Now, although it is true that the Oxford Dictionary gives this meaning, it is surely obvious that, when so used it applies to the measurement and not to the thing measured. An ohmic measurement is a measurement in ohms, but a resistor is ohmic or non-ohmic irrespective of the units in which its resistance is measured. When a medical man refers to faradic currents, he does not imagine for a moment that they are measured by the farad. Measurement by the quart does not make a thing a quartic.

nor measurement by the ton a tonic. A kilogramme does not cease to be metric because it is not measured by the metre. The terms ohmic and non-ohmic, as used by electrical engineers at the present day, refer to properties of the material in relation to Ohm's law and would not be affected in any way if the unit of resistance were not called an ohm but a volt, as was suggested at the 1861 British Association meeting.

Similarly, although, as we pointed out, the terms "linear" and "non-linear" apply to its volt-ampere characteristic and not to the resistor itself, and although a dictionary may give as an alternative meaning "of, or belonging to a line," nobody in the scientific world has any doubt as to what is meant by a linear relationship between the voltage and the current. One correspondent, who has used the term "non-ohmic" in his publications, expresses his approval of "nomadic" as a great improvement in euphony, and hopes that it, or something like it, will pass into general use.

Ohm's Law. A Clear Statement

Our attention has been called to a book entitled "The Metallic State," by W. Hume-Rothery (1931) in which there is no vagueness as to what constitutes Ohm's law. On page 1 we read "Under all normal conditions the resistance of a piece of metal is independent of the strength of the current which is flowing through it, and it is this fact which is the basis of the well-known Ohm's Law, according to which the current strength is directly proportional to the difference of potential, and inversely proportional to the resistance. This law rests solely on an experimental basis, and its establishment is thus of great importance."

Whatever tricks the conductor may play one can always write $R = E/I$, but that is merely a definition of R and not a law. Ohm's law states that $E/I = R =$ a constant independent of I, and, as Hume-Rothery states on p. 4, Barlow showed that no deviation from Ohm's law could be detected for a gold film carrying a current of two million amperes per square centimetre. On the other hand, on p. 19 we read "In both these Group VI metals the conductivity varies considerably with the applied potential, so that Ohm's law breaks down." Surely it is reasonable to refer to such metals, to which Ohm's law does not apply, as non-ohmic, whatever units may be employed. We have given these quotations simply because they seemed to state the conditions very clearly.

We have referred above and in the October Editorial to Ohm's acknowledgment of the earlier work of Davy. On 5th July 1821 Davy read a Paper before the Royal Society entitled "Farther Researches on the Magnetic Phenomena Produced by Electricity; with some New Experiments on the Properties of Electrified Bodies in their Relation to Conducting Power and Temperature" (*Phil. Trans.* 1821). This was four years before Ohm's first publication on the subject, and on reading it one is impressed by the great difficulties under which they worked, the vagueness of the ideas and the crudities of the measurements, and one is also impressed by the great advances made by Ohm in 1826 and 1827, both in accuracy of experimental work and in analysis of the results.

We give some interesting quotations from Davy's paper. "The most remarkable general result . . . was that the conducting power of conducting bodies varied with the temperature, and was lower in some inverse ratio, as the temperature was higher. Whether the heat was occasioned by the electricity or applied to it from some other source, the effect was the same." "Their conducting powers appeared to be in the inverse ratio of their lengths. Operating in this way I ascertained that one inch of platinum [of 1/220 inch diameter] was equal to about 6 inches of silver, to $5\frac{1}{2}$ inches of copper, to 4 of gold, to 3.8 of lead, to about 9/10 of palladium and 8/10 of iron, all the metals being in a cooling fluid medium." "I found, as might be expected, that the conducting power of a wire [of a given material and length] was nearly directly as the mass. I found that when equal portions of wires of the same diameter but of different metals were connected together in the circuit of a powerful voltaic battery, the metals were heated in the following order: iron most, then palladium, then platinum, then tin, then zinc, then gold, then lead, then copper, and silver least of all. . . . It appeared that the generation of heat was nearly inversely as their conducting power."

Davy also drew attention to the enormous difference between the resistances of different classes of material and says "the conducting power of the best fluid conductors . . . must be some hundreds of thousand times less than those of the worst metallic conductors."

He then says " Mr. Children has ingeniously referred the heat produced by the passage of electricity through conductors to the resistance it meets with, and has supposed, what proves to be the fact, that the heat is in some inverse ratio to the conducting power. . . . But till the causes of heat and electricity are known, and of that peculiar constitution of matter which excites the one and transmits or propagates the other, our reasoning on the subject must be inconclusive."

These quotations give one some idea of the knowledge of the subject possessed by the leading scientists at the time that Ohm was beginning his experiments.

We mentioned in the Editorial that the ohm was a unit of liquid measure long before it had any electrical significance. We have to thank Mr. Thomas Carter of Newcastle-on-Tyne for drawing our attention to the following quotation from Longfellow's " Golden Legend,"

" It comes from Bacharach on the Rhine,
Is one of the three best kinds of wine,
And costs some hundred florins the ohm."

Whether, being measured " by the ohm," he would describe this wine as " ohmic," Mr. Carter does not say.

G. W. O. H.

VARIABLE A.F. PORTABLE OSCILLATOR*

By A. R. A. Rendall, Ph.D., M.I.E.E., and F. A. Peachey, A.M.I.E.E.

(Designs Department, Engineering Division, British Broadcasting Corporation)

SUMMARY.—Although this oscillator was designed primarily for the testing of lines used for broadcasting programmes, it would appear to be of general utility for normal test room and laboratory work. The designers do not claim to have developed any new principle but have combined features in such a way that very good results are obtained from a relatively light and small piece of apparatus of a reasonably low cost.

AN oscillator was required for testing inter-station programme channels to a degree of accuracy which would permit close adjustment of equalizing networks. In order that such single links when cascaded in a chain should, in aggregate, provide a good overall response it is often necessary to establish sending and receiving accuracy to $\pm$ 0.1 db. This alone decided that the accuracy of level setting of the output of this oscillator at specific levels should be of this order. An amplifier detector, which is the receiver counterpart of this sender, was described in *Wireless Engineer*, July 1946. For a similar reason and having in mind that such circuits may have quite steep frequency-response characteristics when unequalized, a reasonable precision in frequency accuracy was also required. An accuracy in this respect of $\pm 1\%$, or 1 c/s, whichever is the greater, was deemed necessary.

The oscillator was also required for amplitude-distortion measurements of the order of 1% (unwanted products due to distortion), so that its waveform over the bulk of the frequency range had to be considerably better than this. At the lower end of the audio range it is well known that a greater degree of distortion is tolerable from the aural point of view and thus (fortunately) this limit might be relaxed below, say, 100 c/s. Nevertheless laboratory work often demands good wave forms at low frequencies, so it was felt that the best wave form economically possible should be obtained. These " overload " and non-linear tests demanded that a test signal of at least 100 mW could be sent into 600 ohms.† It was therefore necessary to provide an output impedance within 3% of this value over the whole of the range with a reactive component of not more than 10%. As the oscillator might also be used to test lines not associated with the normal screened repeating coils, the output balance had to be preserved at all level settings.

The normal testing range required for such circuits is 50–10,000 c/s, but obviously an extension to this range might sometimes be

* MS accepted by the Editor, December 1946.

† The Corporation has standardized terminal conditions for such links at 600 ohms.

needed. These factors determined the design of the oscillator which is to be described.

The choice of type seemed to lie between a heterodyne oscillator and an RC oscillator.

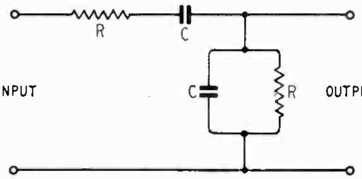


Fig. 1. *Discriminating network in positive feedback path.*

The latter was chosen as it could be made to meet the requirements in a rather lighter and consequently more portable form than the former. This choice particularly applies to portable gear in which thermal balancing of the heterodyne oscillators, and their proper shielding to avoid "pulling" (and consequent bad waveform at low frequencies), demand weight and complication which does not arise if a simple resistancec-apacitance network is employed.

Consideration of the circuit fundamentals, and the components readily to hand, disclosed at an early stage that the main factor which would influence frequency stability was the temperature coefficient of the discriminating-network components. For reasons which will be seen later the amplifier required a high degree of stabilization and a network with a relatively flat frequency-phase characteristic was admissible.

Because, therefore, of its low loss and its simplicity the network shown in Fig. 1 was chosen. With equal values in the series and shunt arms the loss of the network at its critical frequency (zero phase change where $R = 1/\omega C$) can be found as follows:—

$$\text{Impedance of shunt arm} = \frac{-jR^2}{R(1-j)} = \frac{R}{2} - j\frac{R}{2}$$

$$\text{Impedance of series arm} = R - jR$$

Provided the circuit feeding the network has a low impedance, the ratio of the voltage applied to the grid to the e.m.f. in the supply circuit =

$$\frac{\frac{R}{2} - j\frac{R}{2}}{\frac{3R}{2} - j\frac{3R}{2}} = \frac{1}{3}\angle 0$$

The loss is therefore 9.5 db and the phase angle 0.

For a condition of oscillation to be established at this critical frequency the gain of the amplifier needs be slightly in excess of this. This means that the design of a well stabilized amplifier is simple. It may be an amplifier of considerable initial gain reduced by heavy negative feedback. Such an amplifier is shown in Fig. 2. V_1 and V_2 are two pentode valves, resistance-capacitance coupled and with negative feedback supplied from a potential divider connected with its extremes across the output of V_2 and its

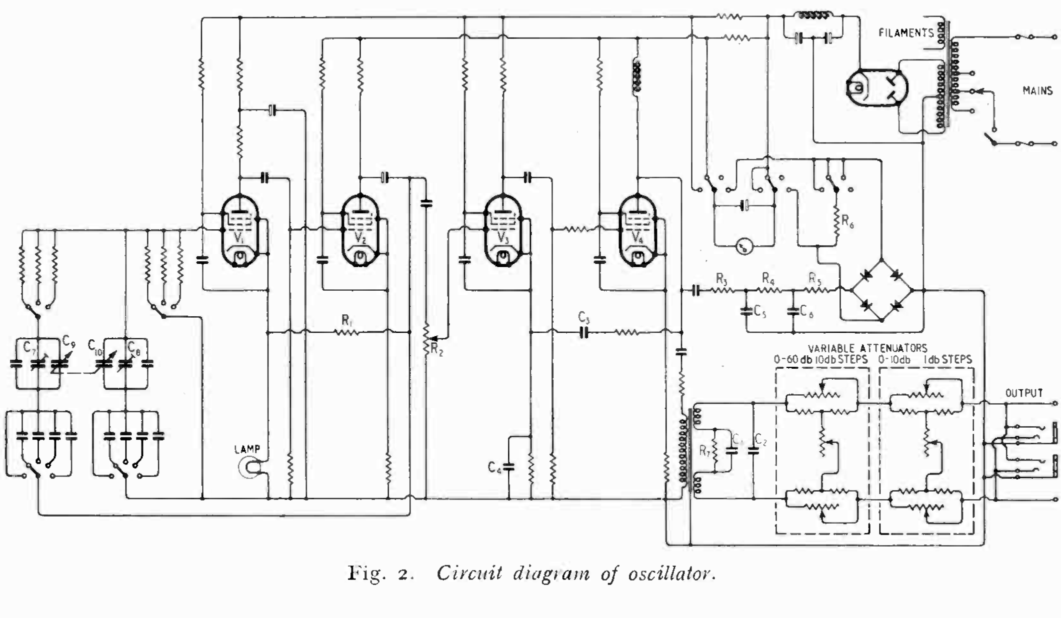


Fig. 2. *Circuit diagram of oscillator.*

tapped point to the cathode of V_1. Actually the potential divider comprises R_1 and a resistance lamp in series, the resistance lamp carrying both the feedback current and the small cathode current of V_1. This lamp plays an important part in the stabilization of the oscillating condition required when the positive feedback path (as described above) is connected between the output of V_2 and the input of V_1, and in practice is not quite so straightforward in its effect as might at first appear.

It is well known that in an oscillator with no automatic control (such as this lamp in the feedback path provides) the level of oscillation is limited almost entirely by the non-linear amplitude characteristic of the output stage. Under such conditions it is, of course, possible by the critical adjustment of a gain control in the oscillator path to arrange the balance between loss and gain to be such that reasonable purity of waveform is obtained. In a good amplifier the change in gain with amplitude is very small indeed until overload is approached and, although adjustment may be made to regulate the level with respect to this small change, the slightest change in circuit conditions will cause a big change to the amplitude of the signal and consequent change in harmonic content.

In a stable oscillator neither condition can be tolerated and the resistance—current characteristic of the lamp (Fig. 3) provides a potential divider in the feedback path which regulates the gain of the amplifier so that, in spite of small incidental changes in the circuit, the required level of oscillation is maintained within proper limits. In practice the output stage must be run at a level sufficiently high to work the lamp filament on its proper characteristic and yet not so high that purity of waveform is appreciably impaired. Another factor influences this part of the design. It is the hunting of the thermal gain control and the level adjustment of the circuit. As pointed out previously, if the amplifier has a characteristic which is slightly non-linear with amplitude, a very small change of circuit conditions in the oscillator path produces a large change in the level at which oscillation occurs. If such a change occurs the lamp slowly responds (because of its temperature lag) to the big change in output level.

An explanation of the hunting or bouncing in level caused by the thermal gain control appears to be as follows. When, for instance, the oscillator is first switched on, the lamp filament is cold and is delayed by its thermal lag in acquiring its stabilizing resistance value. Meantime the level of oscillation grows and with the amplifier at high gain becomes excessive being controlled only by the valve limitations. The filament of the lamp slowly warms and by virtue of its current—resistance characteristic affects the loss in the negative feedback path little at first but decreases it more rapidly later. This change affects the amplitude of oscillation at an increasing rate as its value is brought more within the linear scope of the valve.

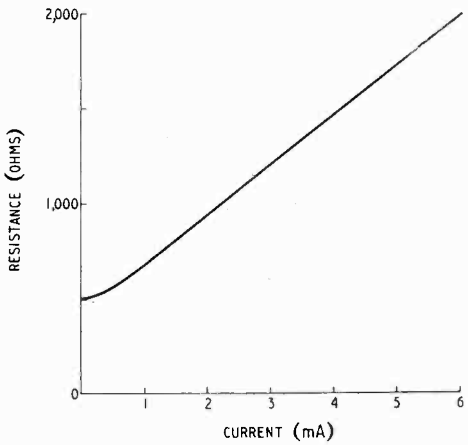
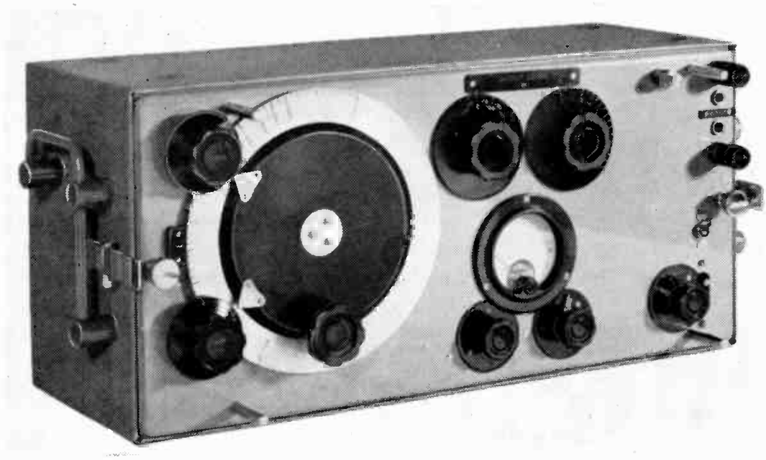


Fig. 3. *Typical current—resistance characteristic of a metal-filament lamp.*

Any change in the volume or frequency in an oscillating circuit must necessarily introduce frequency components other than the fundamental. A highly stable circuit can be regarded as offering resistance to the formation of such components, and in the extreme, oscillation could be supressed. In general the strength of oscillation is merely reduced whilst these conditions pertain. The thermal control would compensate for this by a decrease in resistance to a value below its normal stabilizing value. It lags in action however, and holds the negative feedback at a value rather above that demanded by the lower output conditions. As the negative feedback slowly decreases the volume rises in a corresponding manner and, being a smaller change than that first experienced, is accompanied by less distortion to the wave and so the final steady



Oscillator (with panel cover removed).

little signs of volume bounce at normal adjustment speeds. Under these conditions and with the components used, the thermal control gave little waveform distortion at the lowest frequencies or little effect on output volume due to the effect of ambient temperatures on the lamp.

balance point is effectively reached after a number of diminishing "bounces."

This effect is noticeable for the same reason if the frequency setting of the oscillator is changed suddenly and may become very disturbing indeed if it represents appreciable changes in output volume while the oscillator is being adjusted in a normal manner. It appears to be accentuated (as one might expect) if the oscillator is working at a level considerably below that at which the output valve effectively limits, and also if the normal balancing resistance value of the lamp occurs at a low filament temperature. This means that in the design, a compromise must be made between distortion of the wave on the one hand and disturbing bounces in volume on the other. If a highly stabilized amplifier is used in the oscillating circuit, it will be characterized by a level range almost free from distortion and then a sharp drop in gain after a certain critical level is reached. With this type of amplifier it is possible to adjust the power point of oscillation, so that the waveform is sufficiently pure, although the amplifier may not be working far from its limiting power, with a result that the tendency to bounce may be much reduced. It was found that by proportioning the power-handling capacity of the output stage and the lamp filament current at the stabilizing point, a satisfactory waveform could be obtained which showed

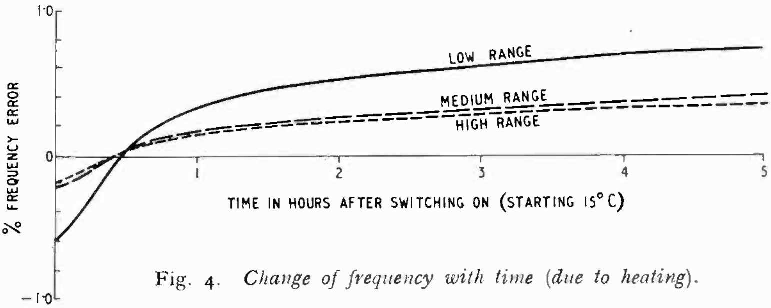


Fig. 4. *Change of frequency with time (due to heating).*

Now that the general functioning of the amplifier with its negative and positive feedback paths has been described we shall revert to the positive feedback discriminating network for further details of its design.

It is well known that the critical frequency (zero phase) of such a network is given when $R = 1/\omega C$; this is when:

$$f = \frac{10^6}{2\pi RC} \text{ where } R \text{ is expressed in ohms}$$

and C is expressed in microfarads

The ranges chosen for this oscillator were 20-200 c/s, 200-2,000 c/s, 2,000 to 20,000 c/s. The variable elements were air capacitors. To avoid excessive values of R, these capacitors were made as large as practicable and actually consisted of two 4-gang capacitors of total capacitances 0.002 μF (Fig. 2, $C_9 + C_{10}$) coupled together by gear wheels. One gear wheel is made in two face-to-face sections with springs between them so that the respective sets of gear teeth are forced apart between the teeth of the driving gear. This practically eliminates backlash.

The gears are of reasonable diameter with relatively closely pitched teeth so that the action is smooth and consistent. The resistances are mainly of the stabilized carbon type with small ordinary resistances for "topping up" purposes. These three sets covering the ranges described have nominal resistance values of 3.3 MΩ, 0.33 MΩ, and 33 kΩ. As the makers of these resistances suggest, the latter two values are more satisfactory from the temperature-coefficient point of view than the former. Fig. 4 shows the change in frequency which occurs as the oscillator warms up. The increased change on the lower-frequency range is practically entirely due to the temperature coefficient of the high-value resistors. Cooling of the unit will, of course, offset this, but as in this particular case the mains supply unit is self-contained, a certain temperature rise could not be avoided. The manufacturers expect to produce opposite coefficient compensating resistors in the near future, so that this point may be improved if it is deemed desirable.

A novel feature is provided by an increment and decrement switch providing

	Plus			Minus	
lower range	0.4	0.2	0	0.2	0.4 c/s
middle range	4	2	0	2	4 c/s
upper range	40	20	0	20	40 c/s

This is effected by switching in the appropriate value of fixed capacitor as shown in Fig. 5, it being observed that :

without C_A in circuit $f_2 = \dfrac{1}{2\pi RC}$

with C_A in circuit

$$f_1 = \frac{C + C_A}{2\pi RCC_A}$$

The increment in frequency is

$$f_1 - f_2 = \frac{1}{2\pi RC} \times \frac{C}{C_A}$$

$$= \frac{1}{2\pi RC_A}$$

From this it is seen that the increment value is independent of C.

This arrangement is not provided with any pretence that the oscillator may be set to absolute frequencies of this order, but as a useful and reasonably accurate facility for studying the performance of resonant circuits, narrow-band filters, etc.

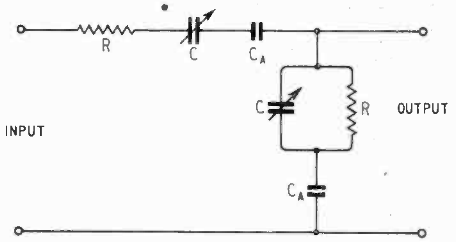
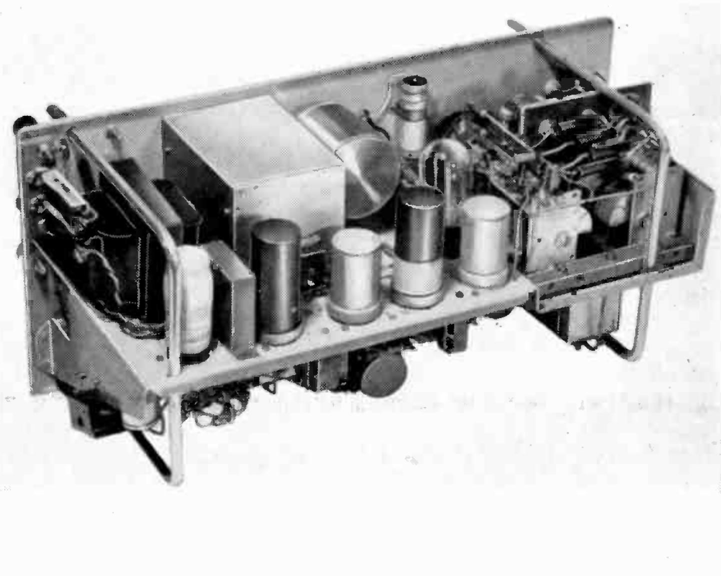


Fig. 5. *Showing addition of "increment" capacitor C_A to network.*

Trimming capacitors C_7 and C_8 (Fig. 2) are provided to equalize the minimum values of the series and shunt element main variable capacitors and so establish equal loss in the network at both ends of the scale. Small adjustments to cover lack of matching over the rest of the scale are made by appropriately bending the split plates of the variable air capacitors C_9 and C_{10}. These adjustments prove quite simple as they are small in scope and the latter process is unnecessary unless particularly constant output volume with frequency is required. The three ranges are brought within close level limits with respect to one another by adjustment of the relative shunt and series resistance trimmers.

The three controls affecting the frequency setting of the oscillator are grouped close together so that mistakes are minimized. The main frequency dial is calibrated at



Rear view of Oscillator.

salient points in the usual manner against standard frequency sources and subdivided by a method described in the Appendix.

The output amplifier is very similar to the oscillator amplifier, having two stages, resistance-capacitance coupled, and its gain reduced by heavy negative feedback. From the gain point of view one stage would be sufficient if it were not for the high degree of stabilization that has been sought. The arrangement has the added advantage that the output impedance of V_4 is made very low indeed and in comparison with the impedance of the transformer which it feeds, appears practically as a generator with no internal resistance. This means that the output level may be measured at this point. The voltage at this point is practically the e.m.f., and the reading is inappreciably affected by the load on the output terminals and the need for substitution of the output load by a 600-ohm load when the level is to be adjusted is thus avoided. The low impedance of the output stage is stepped down by the output transformer and built out on both the primary and secondary sides by resistances so that with a 600-ohm load on the secondary the valve looks into its optimum load impedance while the output appears as 600 ohms.

The leakage reactance of the transformer gives an impedance at the output terminals which is not within the limits required, so this is corrected to a large extent by the capacitors and resistors shown. The reactive component is partially corrected by C_1 in conjunction with R_2 (Fig. 2)—reducing the resistive component to a value below 600 ohms. This is then increased and most of the remaining reactive component neutralized by C_2. The resulting output impedance characteristic is shown in the Table. Any small deviation from 600 ohms is, of course, rendered negligible by the insertion of appreciable attenuation. The two attenuators have steps of 5 db and 1 db respectively, covering a total level change of 0 to 70 db. As the maximum output of the oscillator with no attenuation in circuit is +20 db (with respect to 1 mW in 600 ohms) this provides a range of level in steps of 1 db from +20 to −50. Intermediate levels to the nearest 0.1 db may be set by means of the output level meter and the gain control R_3. The meter has a suppressed zero so that changes of 0.1 db. are more discernible.

The attenuators are of the balanced bridged-T type. It is appreciated that these afford less attenuation to longitudinal currents than the π type but were chosen as the element and switch arrangements are simpler. Longitudinal currents become comparable with the transverse (and wanted) signal with large values of attenuation as the output transformer is not perfect although it is screened. These unwanted currents can, of course, be reduced to negligible quantities by making proper connections to the output circuit, such as by earthing or the use of further screened transformers. For all general purposes longitudinal currents, in this case, may be ignored at levels from +20 to −10.

Again it should be noticed that the level controls, together with the output meter, are grouped together to facilitate the use of the apparatus. The overall frequency response, Fig. 6, is required to be flat to ±0.25 db over the frequency range 20-20,000 c/s so that for most purposes no adjustment of the level control R_2 shall be necessary. A small amount of equalization is, therefore, introduced by capacitors C_3 and C_4 in the feedback path of the output amplifier. For closer adjustments (to within 0.1 db) by means of the level control R_2 it was found simpler to distort the measuring circuit slightly by capacitors $C_5 + C_6$ in conjunction with the building-out resistors R_3, R_4 and R_5 to compensate for the residual lack of linearity in frequency response.

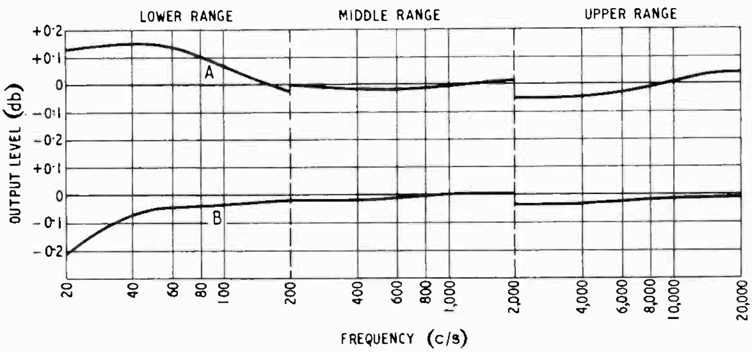


Fig. 6. *Output-level / frequency characteristic.*

The same meter is used, by means of a switch to read valve feeds. These are read in pairs to simplify switching, this being sufficient to enable an operator to discover a faulty valve and replace it in the field. To protect the bridge rectifier against damage while the meter is switched to the "measure feeds" position, a resistance R_6 is switched in as a substitute load for the meter.

Output impedance characteristic
(at output level of + 20 db.)

Frequency (c/s)	**Impedance** (Ω)
50	$600\,\underline{|6.0}$
100	$600\,\underline{|3.0}$
250	$599\,\underline{|1.0}$
500	$601\,\underline{|0.5}$
1,000	$603\,\underline{|0.25}$
2,000	$602\,\underline{|0.25}$
5,000	$603\,\underline{|0}$
10,000	$600\,\underline{|0.5}$
20,000	$620\,\underline{|4.4}$

The complete unit is assembled on a duralumin chassis and panel which slides into a duralumin box and the framework is so constructed that the unit, when out of the box, may be turned over on any face without damage to the components or wiring. The four long pillars on the front enable the unit to be rested on its face without damage to the front or possible interference with the calibrated dial. Lightness combined with strength is of paramount importance if gear of this kind is to be a useful and versatile tool in the field. The unit weighs 31 lb and drops into a canvas cover with comfortable grip handles.

Acknowledgment

The authors wish to express their gratitude to the British Broadcasting Corporation for permission to publish this article, and to several colleagues for their assistance in the design of component details.

APPENDIX

As the accurate calibration of the dial of a direct-reading instrument of this kind is not usually straightforward, the authors feel that a description of the method is of interest.

It has been previously pointed out that the three frequency ranges 20–200, 200–2,000, and 2,000–20,000 c/s have their corresponding resistance values so adjusted that 200 c/s on the one range is 2,000 c/s on the next, and so on. As the variable element—the capacitor—is common to all ranges and the phase response of the amplifier part of the oscillator is zero (or practically so) from 20–20,000 c/s, the calibration of one range, gives the calibration of the other two.

A reliable and precision source of at least one frequency is necessary. From this frequency (actually 1,000 c/s was available), frequencies of simple multiples were calibrated on the dial by the aid of a cathode-ray tube, using the well known methods. For instance clear figures were obtained on the tube for 2,000 c/s, 200 × $\frac{3}{4}$, 2,000 × $\frac{2}{3}$ . . . . . 1,000 c/s etc. These calibration points were marked in pencil on the plain metal dial against the index line. When as many points as practicable were marked in this way, the three screws which hold and locate the flat dial on its mounting boss were removed and the dial was rolled without slipping along the abscissa line of a piece of graph paper. As each pencil mark came against the paper a corresponding mark was made on the abscissa line. At the end of its "roll," the dial (without lifting it from the paper) was rolled back again to ensure that no slip had occurred. This was proved by coincidence of the first frequency marked from the dial with its corresponding mark on the dial.

The ordinate scale of the graph was then marked in corresponding range of frequencies, but with a linear scale, and a graph drawn of frequency against dial calibration.

The quite large graph so obtained was attached to a drawing board and by means of a T-square and right-angled square the actual frequencies and subdivisions *required* to be marked on the dial were projected from the curve on to the abscissa. (The ordinate scale being used to decide the point on the curve from which the projection should be made).

All but the end scale and a few wanted marks were rubbed off the dial and the dial again rolled along the abscissa and its calibration marked with a pencil at wanted points and subdivisions in the opposite sense to the marking of the graph described earlier. By rolling the dial back and observing and marking coincidence, a check was made that no slip had occurred.

The marks on the dial were then engraved over and labelled in a suitable manner.

This seems a practical and simple method of dealing with the problem as direct calibration at all marked points is usually very difficult unless inferior standards are countenanced. It ensures also that subdivisions are reasonably accurate.

NEW YEAR HONOURS

In the New Year Honours List the following were created Knights Bachelor:—J. D. Cockcroft, C.B.E., Ph.D., LL.D., M.Sc., F.R.S., director of the Atomic Energy Research Establishment at Didcot; V. Z. de Ferranti, M.C., chairman and managing director of Ferranti Ltd., and R. V. Southwell, F.R.S., rector of the Imperial College of Science and Technology.

C. G. Phillips, for services as assistant director of telecommunications at the Ministry of Civil Aviation, and F. J. Toone, managing director of Parmeko Ltd., were appointed Officers of the Order of the British Empire. A. E. Adams, chief designer of Scophony Ltd., and L. I. Farren, technical assistant in the G.E.C. Research Laboratory, were appointed Members of the Order of the British Empire.

INTERFERENCE MEASUREMENT*

Effect of Receiver Bandwidth

By G. L. Hamburger, Dipl.-Ing., M.Brit.I.R.E., A.M.I.E.E.

(Formerly of the British Electrical and Allied Industries Research Association.)

CONTENTS

1. Introduction

THIS paper deals with the experimental verification of certain concepts about the measurement of radio interference. An examination of the effects of bandwidth on various types of interference, such as fluctuation noise, single and repeated impulses and noise generated by d.c. motors, is carried out. The findings of the investigations have a bearing on the prediction and assessment of electrical interference at different bandwidths and make it sufficient to measure the interference at one standard bandwidth only.

2. Measuring Equipment

An amplifier was constructed for operation at a fixed mid-band frequency of 5 Mc/s. This comparatively high value was chosen because bandwidths up to about 0.5 Mc/s were envisaged. A well-balanced and symmetrical overall response curve of optimum shape was considered to be important in order to eliminate tedious corrections and also to comply with B.S.I. specifications ; hence the variation of bandwidth was effected by interchangeable filter sets rather than a single set of continuously variable filters, with which a satisfactory response curve is difficult to obtain.

The circuit diagram of this 5-Mc/s amplifier is shown in Fig. 1. The interchangeability of the filters, and the presence of three attenuators distributed at different stages, require the circuit to be tolerably free from spurious feedback since the shape of the response curve would otherwise depend on the attenuator settings. The scheme adopted, therefore, is one which avoids the presence

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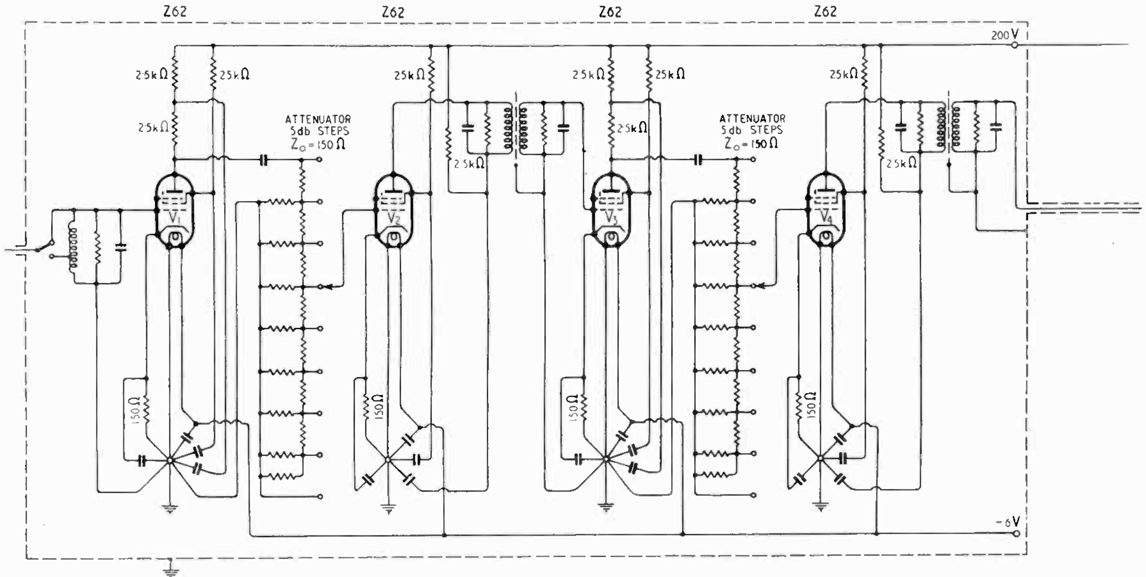
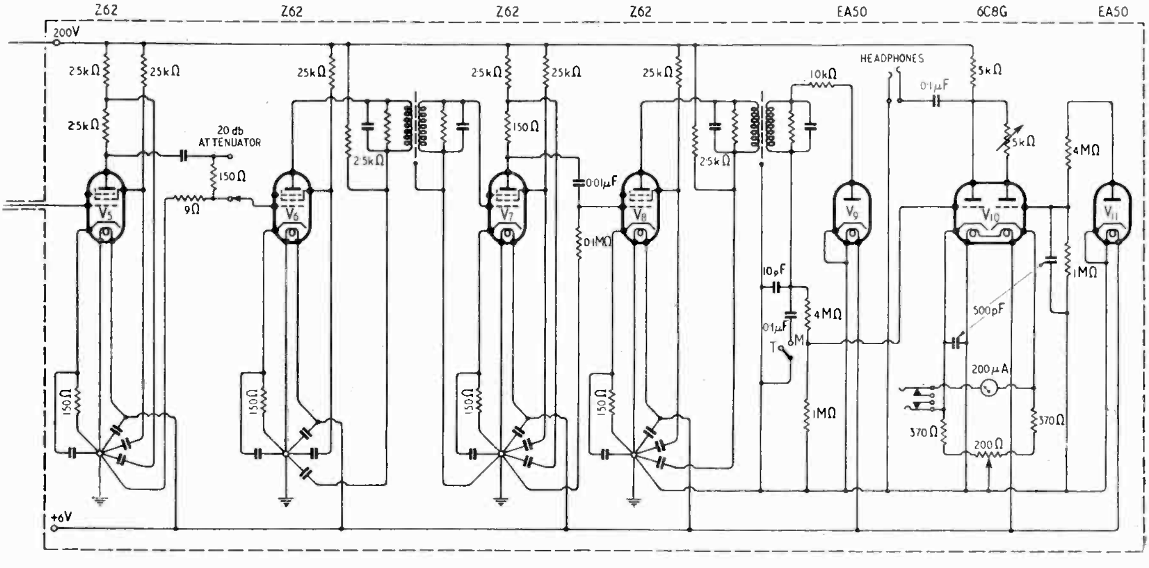


Fig. 1. *Circuit diagram of the measuring band-pass amplifier ; all decoupling*

of a tuned circuit in anode and grid circuits of the same amplifying pentode by the insertion of a low-impedance unity-gain stage between each of the four amplifying stages. Thus, the total number of stages is eight. Each individual stage is decoupled to one earth-point by means of 0·05-μF miniature capacitors. Measurements indicate that the series resonance of these capacitors is 5 Mc/s if the total lead length is about 1½in. The four decoupling capacitors of each stage are mounted directly on the octal valveholder so that the leads have the appropriate lengths, and they form, together with the quarter-watt decoupling resistors, a circuit unit grouped around the valveholder. Each earth-point is then connected straight to the chassis, a measure found to be essential for achieving stability. The input can be applied either to a tapping on the first single-tuned circuit or directly to the grid of the first valve. The first method gives a low input impedance and a high signal-to-noise ratio, and the tuned circuit is broad enough to serve all bandwidths. The first valve has an anode load formed by a 150-ohm ladder-type attenuator capable of a total attenuation of 35 db in 5-db steps, and with the high-slope pentode used the maximum gain of this first stage is about unity.

The next stage V_2 has a coupled filter as anode load and like the other amplifying stages V_4, V_6 and V_8 gives a gain of approximately 32 db. The unity stage V_3 is again provided with a 35-db ladder-type attenuator, whereas stage V_5 offers a choice of either zero or 20-db attenuation, and V_7 gives a fixed unity gain. Each filter is housed in a rectangular copper can which is divided into two compartments each containing one tuned circuit. The partition is formed by an electrostatic wire screen such that electric fields cannot penetrate, yet magnetic fields traverse it unimpaired, thus providing the coupling between the two tuned circuits. The filter connections are brought up to pins fitting valveholders so that the filters can be exchanged fairly rapidly.

The last band-pass filter feeds a diode peak voltmeter similar to that specified in B.S. No. 727 but with the discharge time constant increased to 500 milliseconds, as recommended in the proposed revision of B.S. No. 727 at present under consideration. The diode load resistance of 5 MΩ is tapped at 1 MΩ thus feeding only one-fifth of the rectified voltage to the d.c. amplifier, the reason being that the diode then operates at a higher voltage level (in the linear range) without overloading the d.c. amplifier. The latter is of the bridge type where one arm is represented by the operating triode fed by the rectified signal or noise, and the other by a dummy triode fed by a dummy diode without signal. This scheme helps to compensate for h.t. variations and to a certain extent also for l.t. variations by making both bridge arms identical for the zero signal condition.



and coupling capacitors are miniature 0.05 *μF,* 350 *V unless otherwise stated.*

To overcome differences in the parameters of the two triodes an additional variable, but pre-set, resistor is inserted in one of the anodes. This in effect modifies the characteristics of that triode so as to match the other. The cathode load resistances can be balanced by a 200-ohm potentiometer which acts as zero-setting control for zero signal. A

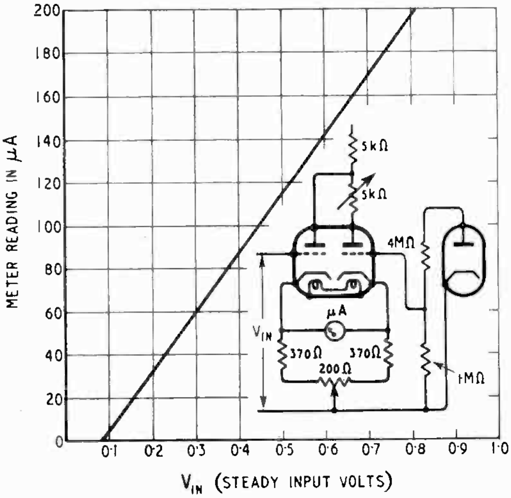


Fig. 2. *Calibration curve of the d.c. amplifier.*

200-μA meter reads the unbalance between the two cathodes and its deflection plotted against the d.c. input to the first triode is shown of Fig. 2. In fact, this arrangement gives a constant zero setting with h.t. variations of the order of ± 15 per cent. A switch permits the reduction of the discharge time constant from 500 milliseconds to 50 microseconds so as to reproduce the envelope of the noise or signal for listening and aligning purposes.

The four sets of filters were aligned by means of the calibrated response-curve tracer,[1] and the four response curves for the sets A, B, C and D are shown in Fig. 3. If they were reduced by ratios inversely proportional to their relevant bandwidth they would essentially merge into one single curve, which is only to be expected since the filter couplings are all set to the same value expressed in $1/Q$. A check was made whether the response curves were independent of the gain setting of the amplifier, and in fact the shape was found to be substantially constant.

3. Measurement of Smooth Noise

Fluctuation or smooth noise, like that which occurs across a resistance or in the anode circuit of a receiving valve, can conveniently be generated at higher levels by means of a temperature-limited diode. Temperature limitation implies current saturation and so precludes the use of oxide and thoriated-tungsten cathodes, since they have so high an emission that their destruction would usually be the result of an attempt to reach saturation. Therefore, a tungsten-filament diode CV172, specially designed for this type of operation, is used. The r.m.s. value of the noise current, i, is under saturation conditions, determined by the equation—

$$\overline{di^2} = 2eI\,df$$

where e is the electronic charge, I the average anode current and df the element of bandwidth considered.

This noise diode is mounted in a small brass box, plugging directly into the coaxial input socket of the amplifier. All battery leads are decoupled to make sure that no signals other than the diode noise can reach the first grid of the amplifier; its input switch is used in the "direct" position, since the input impedance in the "tapped" position is made to match a 75-ohm feeder, and would so shunt the diode load of 2,000 ohms that no appreciable noise voltage could

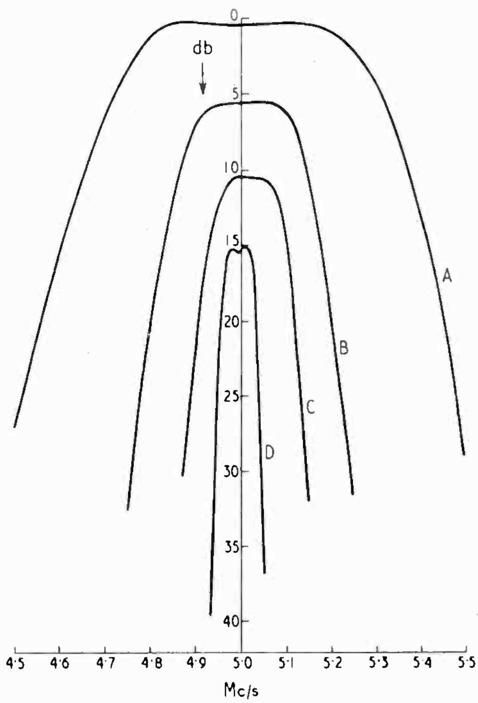


Fig. 3. *Overall response curves for the filters A, B, C and D.*

be developed at the grid of V_1. The circuit diagram of this arrangement is shown in Fig. 4.

It is evident from the above equation that the noise increases with the square root of the bandwidth provided it is measured on a root-mean-square basis. The actual noise voltage, dv, across the impedance Z for an element of bandwidth df is then—

$$dv^2 = Z^2 di^2 = 2eIZ^2\, df$$

In the case of an amplifier, Z can be interpreted as that impedance across which an output voltage is developed; Z will then follow the absolute values of the response curve so that the total r.m.s. value of the noise voltage in the output of an amplifier of gain g is given by—

$$v^2 = 2eIg\int_0^\infty Z^2\, df$$

which can be set equal to—

$$2eIg\, Z^2_{max}\Delta f_N$$

where Δf_N represents the equivalent bandwidths of the amplifier with reference to smooth noise. It follows from the equations that Δf_N can be obtained by squaring the overall response curve, determining its area, and constructing an equivalent rectangle of height Z^2_{max}, its width then being Δf_N.

This procedure has been carried out for all four response curves and the equivalent noise bandwidths Δf_N for the filter sets A to D are tabulated in Table 1, row 2. The attenuation in db corresponding to the noise bandwidth is shown in row 3, whereas the bandwidth corresponding to the standard figure of 6 db is tabulated in row 1 for the sake of comparison.

It is shown in Appendix I that a value of about 3 db attenuation on the total response curve can be expected on theoretical grounds for the equivalent noise bandwidth. The deviations between 3 db and the values actually quoted in row 3 can be explained by the fact that the filters are very slightly overcoupled instead of critically coupled; also the slightest asymmetry in the two humps makes itself felt in this connection.

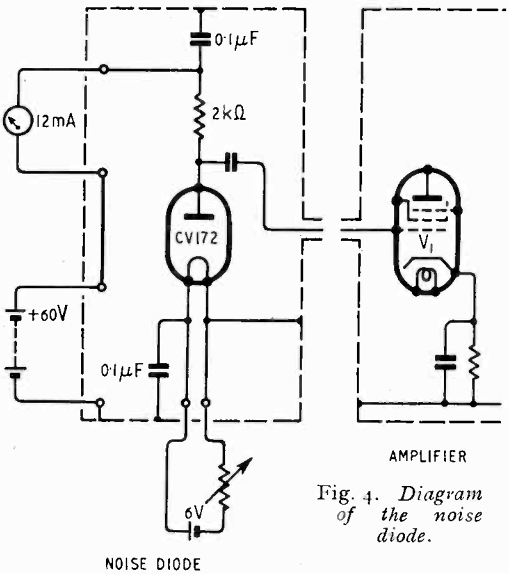


Fig. 4. *Diagram of the noise diode.*

The measurement was made by using the amplifier as a voltmeter. The noise diode, the quiescent current I_0 of which was kept strictly constant, was plugged in the input, and the amplifier attenuators set so that a convenient reading X was obtained. Then

TABLE 1

BANDWIDTH MEASUREMENTS WITH FLUCTUATION NOISE

		A	B	C	D
1	6-db bandwidth (kc/s)	618	285	185	73
2	Calculated equivalent noise bandwidth Δf_N (kc/s) ..	497·5	241	150	59·7
3	Attenuation measured in db on response curve to give Δf_N	2·3	3·3	3	2·7
4	Signal generator input in db above 1 μV to give same reading X as noise (average value of 12 measurements)	34·5	31·4	29·3	25·0
5	Measured response differences in db	3·1		2·1	4·3
6	Response differences in db calculated from the $\sqrt{\Delta f_N}$-law	3·1		2·1	4·0

the noise diode was replaced by the signal generator and its r.f. voltage set to give the previous reading X (see Fig. 5). This procedure was repeated for the rest of the available bandwidths. By thus going a dozen times through all four bandwidths an average value for the equivalent signal generator r.f. input was evaluated for each bandwidth and was tabulated in row 4 of Table I in decibels above one microvolt. The differences between these measured and averaged values for the different bandwidths represent the effect of bandwidth on the reading of fluctuation noise; they are tabulated in row 5, whereas the differences calculated from the equivalent noise bandwidths assuming the aforementioned expected square-root law are compiled in row 6. It can be seen that theoretical and practical values agree quite closely.

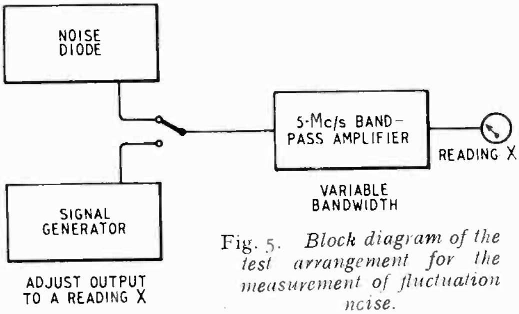


Fig. 5. *Block diagram of the test arrangement for the measurement of fluctuation noise.*

Two conclusions can be drawn from these results. First, the effect of bandwidth on fluctuation noise follows the expected square-root law, and, secondly, the results imply that the diode rectifier does indeed indicate the root-mean-square value of this type of noise, a fact established theoretically by R. E. Burgess[2].

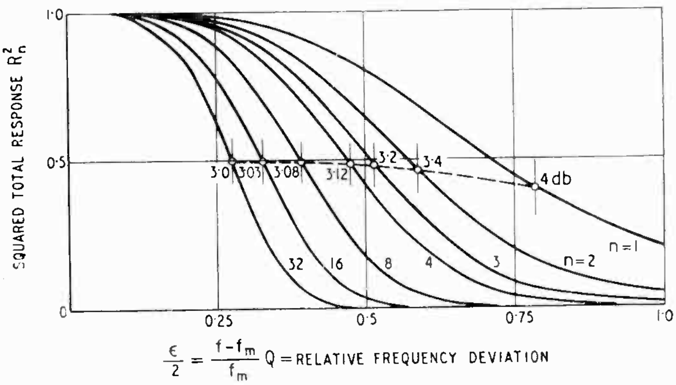


Fig. 6. *Squared-response curves for n critically-coupled filters with the same individual Q.*

4. Single Impulses

Electrical noise can be regarded in two ways. It can either be considered as a time integral of elementary discontinuous phenomena in the electrical state of a medium, the distribution of the occurrence and magnitude of these phenomena being characteristic of the noise source, or it can be considered as a spectrum of essentially continuous frequencies, the phase and amplitude distribution of which is again characteristic for the noise source observed and may be regarded as changing with time. Naturally both definitions describe the same electrical phenomena; they correspond respectively to a function and its representation as a Fourier series or integral.

When treating the subject of noise measurement it is, therefore, of great interest to investigate the effects on the measuring apparatus of one single "element" of noise; i.e., one single discontinuous phenomenon that can be described mathematically as Heaviside's unit step function. All noise may be regarded as the sum of such elements.

The Fourier integral permits in principle the calculation of the resulting transient in the output of a band-pass filter if a unit function is applied to it[3]. If an ideal band-pass filter is defined as a network passing an infinitely sharply defined band of frequencies between f_1 and f_2 with constant attenuation and delay but completely rejecting all other frequencies, the Fourier integral can be evaluated between the limits f_1 and f_2. If the pass-band is narrow compared with the mid-frequency, an approximation can be derived. The response is then given by—

$$R(t) = \frac{\Delta\omega}{\pi\omega_m}\left\{\frac{\sin \frac{1}{2}\Delta\omega t}{\frac{1}{2}\Delta\omega t}\right\} \cos \omega_m t$$

where—

$R(t)$ = response; i.e., the output transient as a function of time

ω_m = $\pi(f_1 + f_2)$; i.e., $\omega_m/2\pi$ = mid-band frequency

$\Delta\omega = 2\pi(f_1 - f_2)$, where $f_1 - f_2$ is the bandwidth of the ideal rectangular frequency response of the filter.

It has been shown[3] that a response such as that depicted in Fig. 8 is unrealizable from a circuit which obeys Kirchhoff's laws since constant delay for all frequencies within the band is impossible for a rectangular response curve, even if an infinite delay is permitted. However, a close approximation can be obtained with narrow bands as far as the main lobe of response is concerned, and the error can be confined to the earlier (if any) or later lobes.

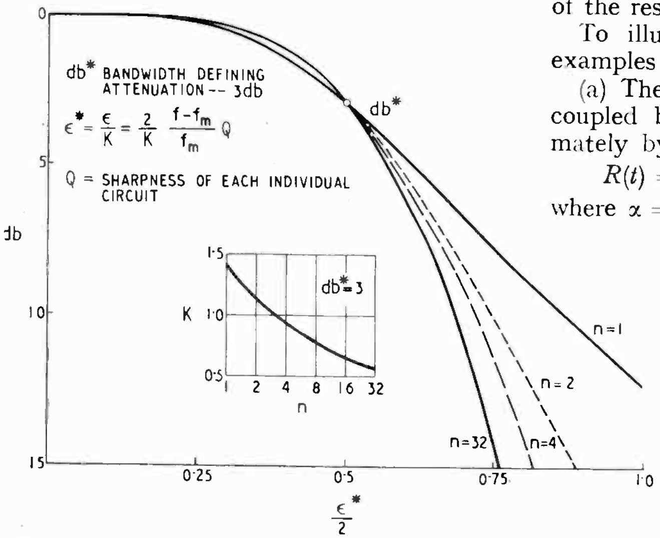
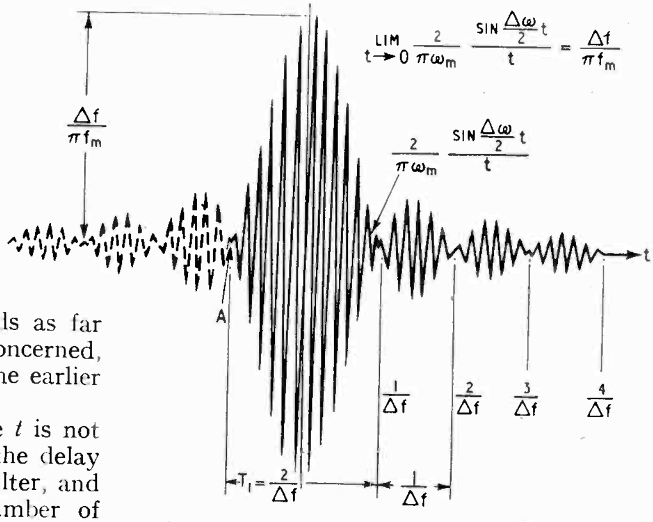


Fig. 7. *Total response curves of n critically coupled band-pass filters referred to the same bandwidth at 3-db attenuation.*

Fig. 8 (*right*). *Transient due to unit function applied to an idealized filter of rectangular band-pass response of width Δf.*



It has also shown[3] that the time t is not measured from the time zero, for the delay would then be infinite in an ideal filter, and this would require an infinite number of conventional elements for its realization[4]. However, if attention is paid to the way in which such filters are approximated as by the methods of Bode and Dietzold[6] or of Levy[5], it can be shown quite generally[3] that the output response cannot start before the zero of $(\sin \frac{1}{2}\Delta\omega t / \frac{1}{2}\Delta\omega t)$. It is therefore possible to put—

$$t = t' - 2\pi/\Delta\omega - t_0 = t - \Delta t$$

where t' is the time measured from the instant of application of the input step. $2\pi/\Delta\omega$ is the minimum delay which makes the transient start at least at A (see Fig. 8), t_0 is an additional delay as occasioned by the transient due to properties typical of the type of filter used, and Δt is the total delay between the input pulse and the maximum of the response.

To illustrate the meaning of t_0 a few examples may be quoted:

(a) The transient of one single critically coupled band-pass filter[3] is given approximately by

$$R(t) = \epsilon^{-\alpha t} \sin \alpha t \sin \omega_m t$$

where $\alpha = R/2L$, R is the series loss resistance of the coils, L their inductance, and ω_m the midband frequency. It differs from the response of the idealized filter, Fig. 8, most essentially in the durations of the lobes. These are identical for the main lobe and the following lobes in the former case, but the main lobe is twice as long as the other in the case of the idealized filter. The delay t_0 is zero for a single-coupled filter; i.e., the transient starts at the beginning of the main lobe. The maximum of the main lobe occurs in this case approximately at $t = \pi/2\alpha$ which follows from putting $\alpha t = \pi/2$. Therefore the total delay is $\Delta t = \pi/2\alpha$.

(b) The transient of a series of single-tuned circuits loosely coupled to each other[7] has been worked out to be—

$$R(t) = \frac{t^{n-1} \epsilon^{-at}}{(LC\beta)^n 2^{n-1} \Gamma(n)} \sqrt{\frac{\pi\beta t}{2}} J_{(n-\frac{1}{2})}(\beta t)$$

where

$$\alpha = \frac{R}{2L}$$

$$\beta = \sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}} \approx \omega_m, \text{ the mid-band frequency.}$$

$J_{(n-\frac{1}{2})}$ is the Bessel function of the order $(n - \frac{1}{2})$ and n the number of circuits in cascade. This response consists of only one lobe decaying exponentially and starting some time after $t = 0$. Employing the usual approximation for Bessel functions for larger arguments, the time at which the maximum of the response envelope occurs can be calculated as $\Delta t = n - 1/\alpha$. It is evident from this formula that the total delay Δt of the maximum of the response increases with the number n of stages.

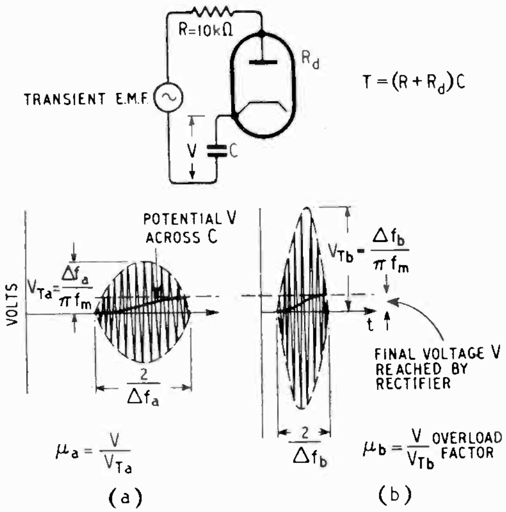


Fig. 9. *Rectification of impulse transients occurring at two different bandwidths Δf_a and $\Delta f_b = 2\,\Delta f_a$.*

(c) A similar behaviour can be expected from a cascade of critically coupled band-pass filters though there appear to be considerable difficulties in calculating the response. However it has been possible to show this effect experimentally with the present band-pass amplifier by feeding the stimulating pulse into successive stages. Fig. 11 (c) shows a 5-Mc/s transient of the filter set A if the pulse is fed into two, three and four successive stages. Two points are noteworthy: first, the delay time $t_0 = a$ (see also Fig. 13) measured from the instant of the applied pulse to the "beginning" of the main response lobe does indeed increase with the number of filters; secondly, it can be seen how the duration of the main lobe increases with the number of stages, which is only to be expected since the bandwidth becomes narrower by increasing the number of filter units.

Finally, it is instructive to realize what shape of static response curve can be obtained by increasing the number of stages of coupled filters. Fig. 7 shows the trend of shape variation to be expected if more and more filters are used but so adjusted that the total bandwidth, defined at 3-db attenuation, is kept the same. The diagram disparagingly indicates that it would be unwise to attempt the approach to a rectangular response curve by going on cascading coupled filters, for the improvement in shape by going from four to the absurd number of 32 filters is really negligible.

The essential task of this part of the investigation was, therefore, to find out the qualitative and quantitative facts about the transient response of the present band-pass amplifier and, if possible, to correlate the real and idealized cases.

An oscillographic method was employed for this purpose and in spite of its inherent difficulties at a frequency of 5 Mc/s, because a true picture of the transient, including delay time, could be obtained. The alternative of a diode rectifier incorporated in the amplifier was considered inadvisable for the following reasons:—It has been shown[3] that the main lobe of the transient of an ideal 9-kc/s band-pass filter, due to unit step, causes the capacitor of a diode rectifier having a charging time constant of 1 msec, to charge up to 6 per cent only of the peak value of the main lobe. It is also explained that "the voltage received by the voltmeter will roughly depend upon the product of envelope peak and duration of the main impulse"; i.e., on $\Delta f/\pi f_m \times 2/\Delta f = 2/\pi f_m$. This shows that the reading will, to a rough approximation, be independent of the bandwidth. This fact is illustrated in Fig. 9, as well as the circuit for which it is valid.

There are two difficulties in the use of the rectifier peak voltmeter for measuring single impulses. They have been described in connection with the assessment of discontinuous interference[3], and they are particularly serious for wider bandwidths.

The first difficulty lies in the unreasonable amount of applied r.f. signal required, as a simple calculation will demonstrate. To give a mid-scale reading on the meter about 0.5 V steady input to the d.c. amplifier is necessary (see Fig. 2). Since the latter is fed from a 1 : 5 tapping of the diode load (for reasons of rectification linearity) about 2.5 V would be required as total rectified voltage. As this is only 6 per cent of the transient crest value, it would require a maximum output of 40 V from the last r.f. pentode in the case of a 9-kc/s band. For 50-kc/s or even 500-kc/s bandwidth the available r.f. voltage would have to be 212 and 2,120 V respectively; surely an unreasonable demand for a measuring set. The ratio of r.f. voltage required to produce a convenient mid-scale meter reading to the rectified voltage produced across the diode load can be defined as the overload factor $1/\mu$, hence the unreasonable overload factor is the first obstacle in the way of measuring single clicks.

The second point to be considered is this: the transient e.m.f. is not produced in a zero impedance source as indicated in Fig. 9, but the above considerations only hold for zero internal impedance. In fact the transient e.m.f. appears across a tuned circuit which, since it is fed by an r.f. pentode of high anode resistance, can be considered as the source impedance. If now the capacitor is entirely discharged and the transient applied, the tuned circuit becomes shunted by 10,000 ohms during the positive half cycles of the transient, which in consequence causes the source voltage to drop to a degree determined by the ratio of resonance impedance to 10 kΩ. The actual reading will therefore be smaller than originally expected. Summarizing, it can be said that overload and filter damping considerations make the present arrangement unsuitable for direct impulse measurements. Indeed a single click applied to the 500-kc/s bandwidth filter set gave no readable indication on the meter.

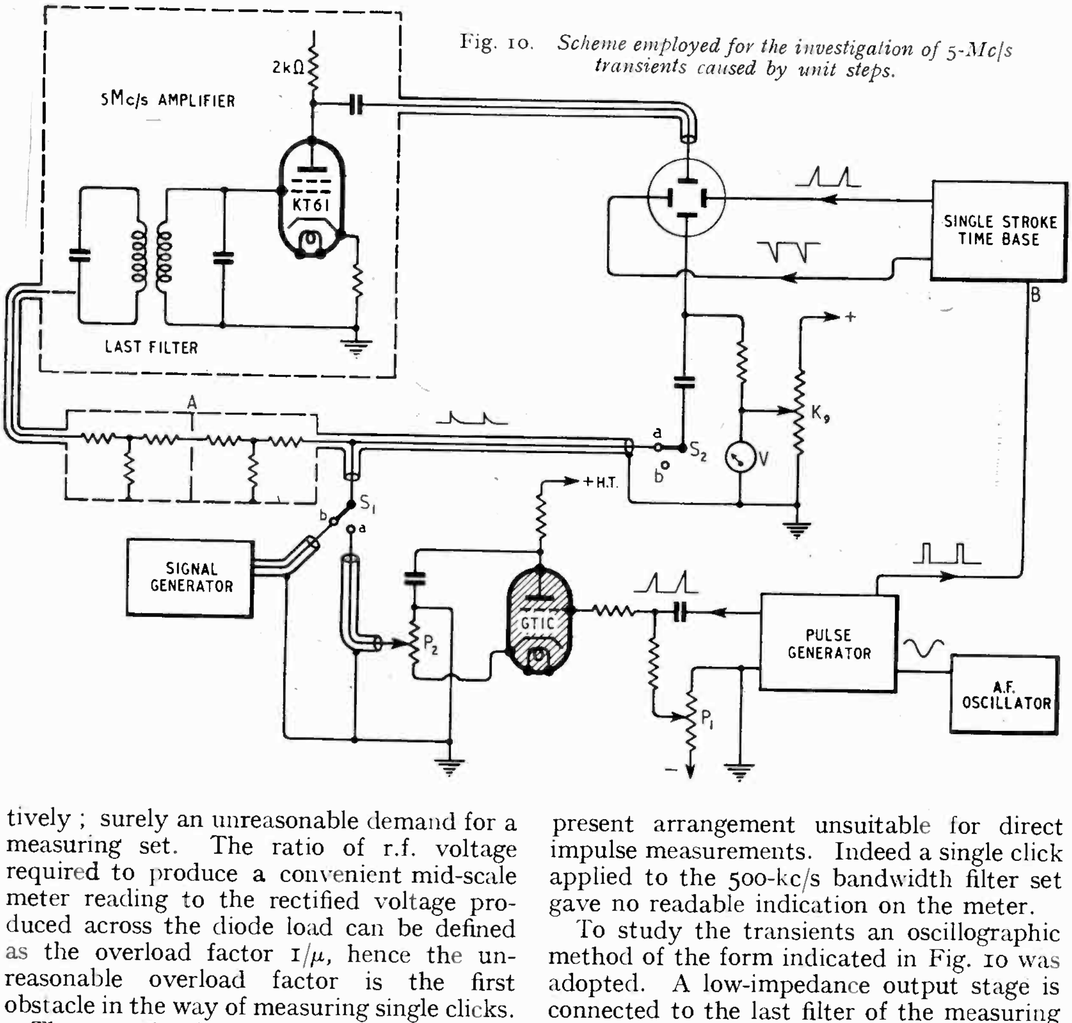


Fig. 10. *Scheme employed for the investigation of 5-Mc/s transients caused by unit steps.*

To study the transients an oscillographic method of the form indicated in Fig. 10 was adopted. A low-impedance output stage is connected to the last filter of the measuring amplifier in order to put a faithful replica of the transient on to the upper plate of the cathode-ray tube without detuning the filter. The actual unit step was produced by a thyratron discharge, and photographic measurements have revealed that the leading edge of the pulse thus produced has a duration of about 1/30 to 1/50 of a microsecond. Fig. 11 (a) is a record of that pulse; its trailing edge is formed by a discharge curve of about 0.2-microsecond time constant. It is shown in Appendix II that the error committed by considering this particular pulse as a true unit step in the 5-Mc/s region is only of the order of 0.1 db.

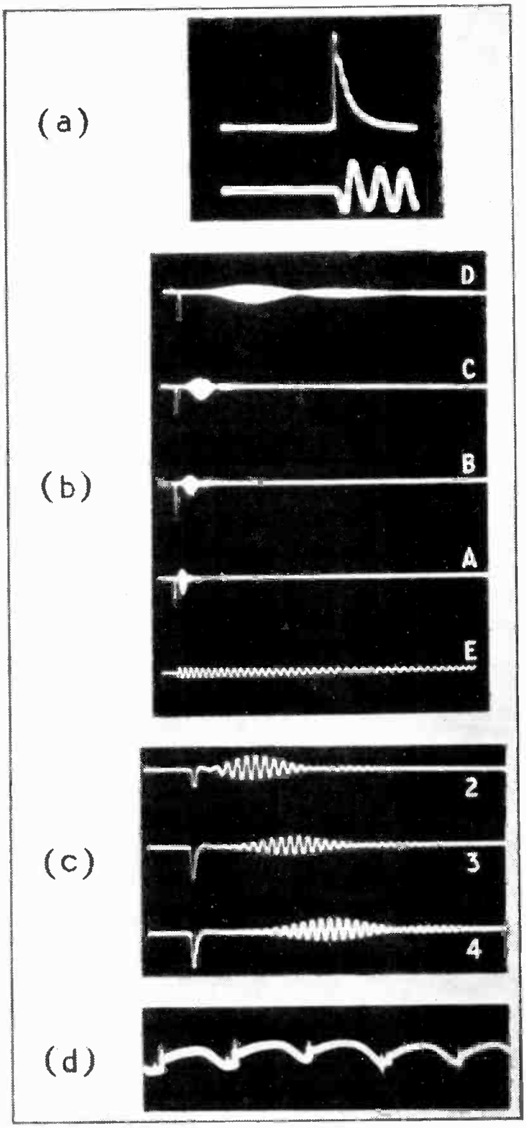


Fig. 11. *(a) Artificial impulse used for transient measurements and 5-Mc/s timing wave; (b) transients of filters A—D and single circuit E tuned to 0.5 Mc/s initiated by the artificial impulse and shown on a common time scale; (c) transients of band-pass amplifier with filter-set A when using 2, 3 or 4 filters; (d) oscillogram of the commutator voltage showing five commutations. The vertical lines represent the r.f. interference generated.*

To produce a steady picture of the transient a recurrent single-stroke system was adopted. An a.f. oscillator controls the number of recurrences of the transient per second and is synchronizing a pulse generator producing a rectangular pulse of a width variable between 5 and 200μsec as well as a slanting pulse of the same duration. The rectangular pulse is fed to the E.R.A. Universal Oscillograph[6] to trigger the single-stroke time base and switch on the brightness for the duration of the stroke. The other pulse, however, is fed to the grid of the thyratron and the negative bias setting P_1 controls the exact instant of ignition with respect to the start of the stroke, owing to the slope of the leading edge of that pulse. The "unit step" thus produced by the thyratron is fed to the lower Y-plate of the c.r. tube and via an attenuator to the input of the amplifier if the switches S_1 and S_2 are in position "a."

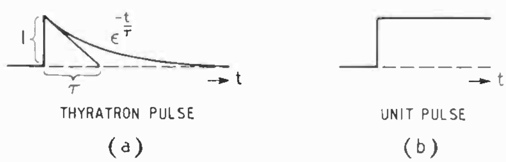


Fig. 12. *Pulses for transient investigation.*

Photographs of the transients thus obtained for the different filter sets A to D are reproduced in Fig. 11 (b). They show the initiating pulse and, after a certain delay time, the 5-Mc/s transient. The transient was measured in terms of time and amplitude, and then compared with calculated expectations. All data are computed in Table 2. The procedure of measurement will be explained with the aid of Figs. 10 and 13.

As a first step the height V_P of the pulse was measured with the voltage calibration control K_9 as described in a previous paper[8], and the crest amplitude V_T of the transient was determined (without necessarily knowing its absolute value). Then switches S_1 and S_2 were thrown over to "b" which causes the signal generator to take the place of the "unit steps" and disconnects the lower Y-plate. On the screen appeared, instead of the transient, a 5-Mc/s wavetrain, the constant amplitude of which was made identical with the previous crest value V_T by setting the signal generator output accordingly. Its output voltage V_{SG} had to be multiplied by $\sqrt{2}$ and a factor k in order to find out the real peak value at the point S_1. produced in an ideal filter for which the equation and diagram of Fig. 8 hold. At any rate it is interesting to compare both sides of the equation in practice, for $\Delta f/\pi f_m$ is a calculated quantity and $V_{SG}\sqrt{2}k/V_P$ a measured value. Both are tabulated in

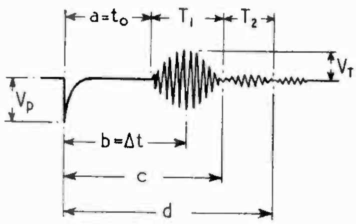
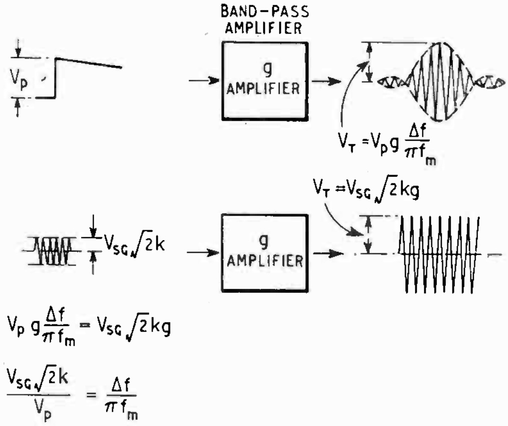


Fig. 13[1] *Measurement of transients due to single impulses by means of the scheme of Fig. 10.*



"k" takes into account the output impedance of the signal generator (10, 15 or 52.5 ohms according to its multiplier setting) and the input impedance (75 ohms) of the external attenuator A; k therefore assumes the values of 0.88, 0.83 and 0.588 respectively. The measurement procedure is illustrated in Fig. 13 and it follows quite simply that the quotient $V_{SG}\sqrt{2}k/V_P$ should be equal to $\Delta f/\pi f_m$ provided the transient is

Table 2 as well as their respective discrepancies expressed in decibels.

As a next step the bandwidth of the filters was determined from the duration T_1 of the main transient lobe. For the purpose the c.r. screen had to be calibrated in terms of microseconds, which was carried out by means of a timing wave [see Fig. 11 (b)]; viz., a damped wave-train of known frequency initiated by the impulse. The bandwidth

TABLE 2
SINGLE TRANSIENT MEASUREMENTS

Filter Set		A		B		C		D	
Δf Bandwidth (kc/s)	6 db		618		285		185		73
	$2/T_1$	645	587	273		143		66	
	Noise, Δf_N		497.5		241		150		59.7
Intervals (μsec)	$a = t_0$	1		1.8		4		10	
	b	2.65		5.1		11.2		25	
	c	4.1		9.13		18		40	
	d	6.25		14.7		27.2		62	
	T_1	3.1	3.4	7.33		14		30	
	T_2	2.15		5.54		8.2		22	
Measured quantities	V_{SG} (mV)	450	455	200	190	316	185	80	
	V_P (mV)	9600	8700	8200	9000	20,000	12,250	22,500	
	$V_{SG}k\sqrt{2} \times 10^3/V_P$	30	43.5	20.2	17.6	13.1	12.5	4.17	
Calculated $\frac{\Delta f}{\pi f_m} 10^3$	6 db					11.8		4.65	
	$2/T_1$	41	37.5	17.4	17.4	9.1	11.9	4.23	
	Noise, Δf_N	31.9		15.3	15.3	8.9	8.9	3.8	
Error in db	6 db					+0.5		−1	
	$2/T_1$	−2.6	+1.3	+1.3	0	+3.2	+0.5	−0.3	
	Noise, Δf_N	−0.5		+2.4	+1.2	+3.4	+2.9	+0.8	

could then be calculated as $\Delta f = 2/T_1$ (see Fig. 8). One must, however, bear in mind that "bandwidth" as far as real filters are concerned, is subject to definition; i.e., unless a definition is specified the mere statement of a bandwidth value for a real filter is futile. Three such definitions have been mentioned so far: (a) the bandwidth defined—rather arbitrarily—at 6-db attenuation on the response curve, (b) the equivalent-noise bandwidth corresponding to about 3-db attenuation, and (c) the equivalent bandwidth corresponding to a certain duration T_1 of the main transient lobe. All three definitions will in general give different bandwidth values for one individual filter set.

Regarding (c), the bandwidth relevant to the transient duration, it could tentatively be obtained by integrating the response curve—since the transient is a summation of all frequencies within the band—and by constructing a rectangle of equal area and maximum height, its width then being the equivalent bandwidth for the transient duration. Similar calculations have been carried out in the previous section and, in fact, the curve in Fig. 6 for $n = 2$ can be used straightaway as it represents the squared response of two filters, which amounts to the same as the unsquared response of four filters. With sufficient accuracy one can, therefore, take the noise bandwidth f_N obtained from the measured response curves (see Table 1) also as the bandwidth corresponding to the transient duration T_1.

Comparisons can be drawn from a perusal of Table 2 and the following observations can be made :—

(i) The transient of four coupled band-pass filters differs in certain particulars from the transient response of an idealized filter but behaves in fairly satisfactory accord with the theoretical discussion already given. No small lobes precede the main lobe but this is not to be expected since the practical filter characteristics differ widely from the hypothetical case of a rectangular response curve and linear phase shift.

(ii) There is a finite delay time (a) measured between the instant of the impulse and the crest of the transient [see Figs. 13 and 11 (c)] which corresponds to the delay Δt already discussed; it increases with the number of stages cascaded.

(iii) The relative shape of the transient envelope is independent of bandwidth and, in the present case, is only dependent on the number of filters cascaded (provided the relative coupling is always the same). This can be seen by forming proportions between the quantities a, b, c, d, T_1 and T_2 compiled in Table 2 (see Fig. 13) and by inspection of Fig. 11 (b). Actual differences are due to the limited accuracy of the measurement and slight discrepancies between the relative shapes of the response curves in Fig. 3.

(iv) The crest values of the real transients agree with those of the ideal transients within about plus or minus 3 db. Since quite a number of factors enter the rather complex measuring set-up of Fig. 10, these deviations may be regarded as inaccuracies of measurement, so that it can be stated that the crest values of real and idealized transients agree fairly well.

(v) The bandwidth corresponding to the duration of the main lobe agrees fairly well with the noise bandwidth Δf_N except in the case of filter A. The reason for this discrepancy of 20 to 30 per cent can be traced to the fact that the A filters, as the widest set, show the typical feature of a high L/C ratio which is needed for getting the necessary amplification. Hence the output stage KT61 of Fig. 10 appreciably detunes the last tuned circuit. In the other cases a low L/C ratio ensures relative independence of tuning from small capacitance variations.

(To be concluded)

DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH

It is announced that the Lord President of the Council has approved the promotion of R. L. Smith-Rose, D.Sc., Ph.D., M.I.E.E. to the new post of Director of Radio Research.

Dr. Smith-Rose has been Superintendent of the Radio Division of the National Physical Laboratory since 1939, and in his new post he will be in charge of the radio research work of the Department of Scientific and Industrial Research. It is intended eventually to establish a new station for this which will incorporate the work now being carried out in the Radio Division of the National Physical Laboratory, the Radio Research Station of D.S.I.R. at Slough and work now being done for D.S.I.R. at the Telecommunications Research Establishment at Malvern.

PHYSICAL SOCIETY'S EXHIBITION

The annual exhibition of Scientific Instruments and Apparatus will be held at Imperial College, South Kensington, London, S.W.7, on 6th—9th April inclusive. Admission is by ticket only and the exhibition is to be open from 10 a.m. until 8 p.m. (except between 1 and 2 p.m.) each day except the first, for which the hours will be 2—9 p.m.

TRANSMITTER-BLOCKER CELLS*

Cm-Wave Fixed-Tuned Types

By R. H. Kay, M.A., A.Inst.P.,† *and M. Surdin, Docteur-ès-Sciences*‡

(Admiralty Signal Establishment)

SUMMARY.—The transmitter-blocker cell (also referred to as the a.t.r.[1] cell, the transmitter disconnect switch) is a resonant-waveguide device positioned between the receiver branch and the magnetron in centimetric r.f. heads so as to present on reception a high impedance in series with the magnetron branch of the main guide run and prevent the undue reception loss which would otherwise be associated with certain values of magnetron cold impedance. The device may either be tunable, or fixed tuned for operation over a finite band.

In this paper the theoretical principles involved in the design of fixed-tuned t.b. cells are outlined, together with the experimental methods employed.

From equivalent circuits for the cell and associated waveguide systems, the impedance conditions for maximum bandwidth for a given edge-band resistive component are derived and methods are given for computing the loss over the band.

A section is included to cover the magnetron cold impedance parameters relevant to this paper.

Finally, possible lines of approach are outlined for increasing the bandwidth of systems employing t.b. cells.

MAIN SYMBOLS

Z, Z_0 = impedance, characteristic impedance.
y, y_0 = admittance, characteristic admittance.
R_T, X_T = effective series resistive, reactive components of the t.b. impedance.
R_M, X_M = effective series resistive, reactive components of the magnetron impedance.
f, f', etc. = frequency.
f_0 = resonant frequency.
λ = free-space wavelength.
λ_g = wavelength in the guide.
λ_{g0} = wavelength in the guide at a resonant frequency of the system.
ω, ω_0 = angular frequency, resonant angular frequency, $\omega_0 = 2\pi f_0$.
δ = bandwidth = $\dfrac{2\Delta\omega}{\omega_0}$ where $2\Delta\omega$ is the angular frequency separation between the extreme ends of the frequency band under consideration.
Q_0 = unloaded Q of a resonant system.
Q_L = loaded Q of a resonant system.
ρ = reflection coefficient.
ρ_0 = reflection coefficient at resonance.
j = the operator $\sqrt{-1}$
η = efficiency.

[Special modifications of these symbols are defined in the appropriate sections of the paper.]

* MS accepted by the Editor, November 1946.

† Now at the Clarendon Laboratory, Oxford.

‡ Now at the research laboratories of the French Atomic Energy Commission.

[1] "Anti-Transmitter Reception Switch," a term coined when these devices were first developed.

Introduction

IN centimetric r.f. heads which are designed for common transmission and reception on a single aerial system two self-acting electronic switches are employed as shown in Fig. 1. The first of these is the "t.r. cell" (the transmit-receive switch), consisting essentially of a section of waveguide sealed off with low-Q resonant glass windows transparent at the operating frequency and filled with gas at low pressure. The t.r. cell is an integral part of the feed to the receiver, branching from the main guide run. On transmission the gas immediately behind the window breaks down into a glow discharge,

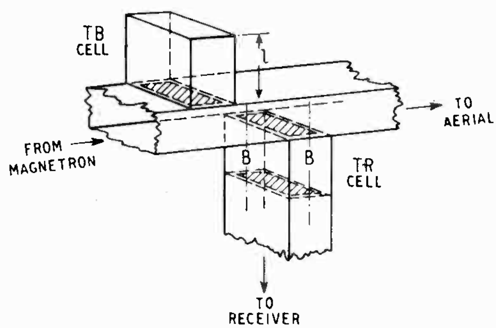


Fig. 1. *Arrangement of "Transmitter-Blocker" cell and "Transmit-Receive switch" cell on a waveguide.*

effectively short-circuiting the receiver branch and protecting the sensitive crystal from what would be certain burn-out if it were exposed directly to the high-power levels of transmission. The cell is positioned so that, on transmission, the magnetron power suffers little attenuation or reflection by the arc and is transferred to the aerial with negligible loss. On reception, the received power being minute and insufficient to initiate the glow discharge, the t.r. cell becomes once more transparent and behaves as an integral part of the feeder system to the mixer.

The second electronic switch, the "transmitter-blocker cell," consists essentially of a resonant waveguide stub, branching from the main guide run at a position between the t.r. cell and the magnetron. This section, short-circuited at the end remote from the main guide run, is sealed off with a low-Q resonant glass window positioned so as to be effectively coplanar with the wall of the main guide run, so that on transmission, when a glow discharge is set up across the window, the power radiated by the magnetron suffers little reflection and is attenuated only by an amount determined by the resistive loss in the discharge. By suitable choice of gas filling and window dimensions the latter can be made very small. On reception, the power level in the main guide run is insufficient to initiate the arc and the stub length "l" is adjusted to resonate the system so that a high impedance is introduced in series with the magnetron branch, the value of which is ideally unaffected by the "cold impedance" of the magnetron so that the conditions of power transfer to the receiver branch are isolated from the effects of variability of this latter quantity. The device may be tunable for particular frequencies (i.e., "l" may be capable of adjustment) or may be fixed-tuned for operation over a finite band in which case "l" is fixed so that the system resonates at or near the mid-band frequency. The distance t.b.–t.r. is chosen for maximum power transfer to the receiver.

The need for the t.b. cell is illustrated by the fact that, in the absence of such a device, the transfer of received power into the receiver branch is governed by the relation :
Reception efficiency

$$= \frac{\text{Power transfer to receiver}}{\text{Total available power}} = \frac{4}{(2+R)^2+X^2}$$

where R and X are the normalized impedance components of the magnetron branch *referred to the plane of the receiver* (BB in Fig. 1) and the receiver branch is taken to be matched $\left(\frac{z}{z_0} = 1\right)$; see Section 2. The loss is usually expressed in decibels below the value unity. Thus, assuming initially a purely reactive magnetron, the loss may be anything from zero to infinity, depending on the phase of the wave reflected from the magnetron referred to the receiver branch ; and for a magnetron presenting on reception a voltage standing-wave ratio (v.s.w.r.) of 20 : 1, the efficiency will vary from $4/(22)^2$, corresponding to a loss of 19 db, to $4/(2.05)^2$, corresponding to a loss of 0.2 db, depending on the phase.

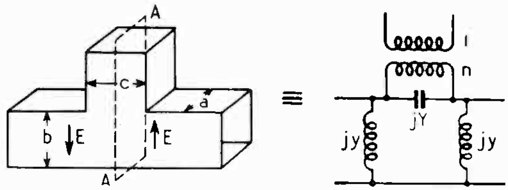


Fig. 2. *T-junction in the E-plane and its equivalent circuit.*

With a t.b. in circuit, however, when R and X now represent the impedance components of the t.b.-cum-magnetron branch referred to the plane of the receiver, the loss will be limited ; for we may consider the t.b. and magnetron impedances to be effectively in series and R and X are now made up of series components R_T, X_T and R_M, X_M, values appropriate to the t.b. and magnetron respectively. If the resistive component of the t.b. cell itself is made very large and the cell, series connected to the main guide, is positioned an odd number of quarter-wavelengths behind the t.r. cell, then R_T, the resistive component referred to the plane of the receiver, will be small and the efficiency will at the most be equal approximately to $4/(2+R_T)^2$ for a purely reactive magnetron, which for a t.b. of high-resistive component will be equivalent to a loss of only a fraction of 1 db.

Briefly, then, from this elementary introduction to the problem, the requirement will be to design a t.b. with a resistive component sufficiently high over the entire frequency band to ensure a negligible reception loss whatever the value of the magnetron impedance. In this paper we will elaborate the design principles, deriving equivalent circuits for the cell and associated waveguide systems (1.1 ; 1.2), expressions for the impedance conditions for maximizing the band for a given restricted edge-band loss (1.3) and also methods, for computing the loss at edge-band and over the band, for given values of t.b. impedance (2. ; 2.1 ; 2.3). One section (2.2) deals with the magnetron cold impedance characteristics relevant to this report and finally (3) possible methods of increasing the bandwidth are outlined. Relevant experimental methods are described throughout.

This paper is the result of investigations carried out in the 10,000-Mc/s frequency range in the design of a fixed-tuned transmitter blocker cell for Admiralty Signal Establishment.

1.1. *Equivalent Circuit of the T-Junction*

The behaviour of the T-junction proper will first be considered before introducing any window. (The series T-junction will be dealt with initially because its bandwidth is greater than that of a parallel junction.) The mode of attack used will be based on equivalent circuits. Although the choice of equivalent circuit is not unique, the one used here gives a good approximation to the experimental results and has the advantage of being particularly simple.

Frank and Chu[1] have shown that a T-junction in the E-plane may be represented by the equivalent circuit shown in Fig. 2, where

$$jy = \frac{1}{jf_2}, \; jY = j\frac{1 + f_1 f_2}{2f_2}, \; n = \frac{f_3}{2}, \quad (1.1.1)$$

The symbols y and Y represent admittances normalized with respect to the characteristic admittance of the waveguide ($Y_0 = Z_0 = 1$), and f_1, f_2, f_3 are real positive quantities and are functions of b/λ_g and c/λ_g only. For the case $b = c$, f_1 and f_3 increase and f_2 decreases as b/λ_g increases. "n" is the "transformation ratio," defined so that a susceptance jY_1

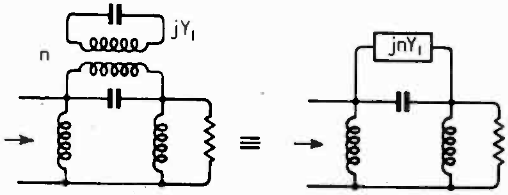


Fig. 3. *Other circuits equivalent to Fig. 2.*

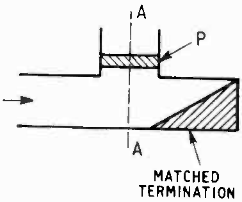
in the side arm of the T-junction appears as a susceptance jnY_1 in series with the termination of the main member when viewed from the input end of the main guide. Therefore the circuits of Fig. 3 are other equivalents.

Impedances and admittances are referred to the axis of symmetry AA of Fig. 2 (i.e., to the mid-section of the stub), and the quantities f_1, f_2, f_3 have been computed. Their values for a rectangular waveguide of 1in by ½in internal dimensions and at $\lambda = 3.1$ cm, and for $c = b$ (the only case considered in detail hereafter) are

$$f_1 = 0.88, f_2 = 9.2, f_3 = 3.1 \quad .. \quad (1.1.2.)$$

The mathematical relations between the parameters discussed above are for the completely general case extremely complex, but the values of y, Y and n may be determined experimentally in the following way. Having matched one of the main guide branches of the T-junction, measure the variation in the admittance at the plane of symmetry AA, looking in from the other main guide branch, as a piston P is moved in the stub (see Fig. 4).

Fig. 4. *Arrangement used in measuring the circuit parameters.*



When this admittance is plotted on the complex plane, a circle tangent to the B-axis will be obtained with centre at $\left(\frac{1 + y^2}{2}, y\right)$ and of diameter $(1 + y^2)$. This experiment determines the value of y.

If now the piston is placed in one of the main guide branches, the other branch being matched, and the admittance as viewed from the stub is plotted on the complex plane, another circle is obtained with its centre at $\left(\frac{1 + y^2}{2n}, \frac{Y}{n}\right)$ and of diameter $\left(\frac{1 + y^2)}{n}\right)$.

The value of y having been determined from the first experiment, the values of n and Y may be computed from the second.

A third experiment may be performed as a check by plotting the admittance as viewed from one branch of the main run, the stub being matched, and the piston now being moved in the other branch of the main guide run. A circle should then be obtained with its centre at $\left(\frac{n^2 + y^2}{2n}, y\right)$ and of diameter $\left(\frac{n^2 + y^2}{n}\right)$.

In practice it will be found that the check is not perfect and this may be attributed to the fact that the equivalent circuit does not fully represent the actual conditions at the junction. In fact, experiments have shown that y is smaller and Y larger than the values given by Frank and Chu; n is very nearly the same as predicted theoretically.

Since the experimentally determined value for y is even smaller than that predicted by Frank and Chu $\left(|y| = \frac{1}{f_2} = \frac{1}{9.2} = 0.11\right)$ there is reasonable justification in neglecting y in drawing the equivalent circuit for the t.b. in Section 1.2 below.

1.2. *The Equivalent Circuit of the Transmitter-Blocker*

Considering the resonant window as a parallel-resonant circuit across the guide, the length "l" of the stub, as a susceptance in parallel, and the loss in the stub, as a parallel resistance, Fig. 5 may be taken as the equivalent circuit of the complete t.b. when mounted in the guide. L is the lumped inductance of the window and stub, C is the capacitance of the window including Y, as defined above, and R is the resistive loss of the window and the stub. In Section 1.1 it has been shown that y may be neglected.

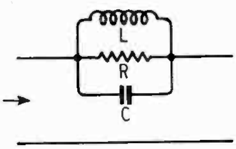


Fig. 5. *Equivalent circuit of the t.b. cell when mounted in the guide.*

To avoid the unnecessary repetition of Z_0 in the mathematics, normalized impedances and admittances ($Y_0 = Z_0 = 1$) will be used; and R_0 will be written for the value "R/Z_0" as a reminder that this is being done.

To compute Q_L, the loaded Q of the system, in terms of easily measurable quantities, consider Fig. 6, where the guide (of characteristic impedance $Z_0 = 1$) is fed by a matched generator and is matched also beyond the t.b. cell so that the LC circuit is in effect loaded by R_0 and 2 in parallel.

The input impedance of the circuit is clearly given by:

$$Y = \frac{1}{R_0} + j\left(\omega C - \frac{1}{\omega L}\right) \quad \ldots \quad (1.2.1.)$$

and the susceptance vanishes when

$$\omega = \omega_0 = \frac{1}{\sqrt{LC}},$$

where $\omega_0 = 2\pi f_0$ is the resonant angular frequency of the circuit and f_0 is the resonant frequency.

We have also the further relation that, at f_0,

$$\omega_0 L = 1/\omega_0 C = \sqrt{L/C} = Z_c \ldots \quad (1.2.2.)$$

The admittance of the circuit at any frequency can now be expressed in the form

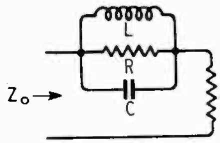
$$Y = \frac{1}{R_0} + \frac{j}{Z_c}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \quad \ldots \quad (1.2.3.)$$

$$\text{Now } \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} = \frac{2\Delta\omega}{\omega_0}\left(1 + \tfrac{1}{2}\frac{\Delta\omega}{\omega_0}\right)\left(1 + \frac{\Delta\omega}{\omega_0}\right)^{-1},$$

where $\Delta\omega = \omega - \omega_0$, and for small values of $\Delta\omega$ this expression reduces to $\frac{2\Delta\omega}{\omega_0}$, or writing

$$\delta = \frac{2\Delta\omega}{\omega_0} \text{ we have } Y = \frac{1}{R_0} + j\delta/Z_c \quad \ldots \quad (1.2.4.)$$

Fig. 6. *Equivalent circuit of the t.b. cell when the wave-guide is matched beyond the cell.*



The sharpness of the resonance curve depends on the ratio $Q_0 = \frac{R_0}{Z_c}$ for the unloaded circuit. In this case the circuit is loaded by R_0 and 2 in parallel and we may define the loaded Q by:

$$Q_L = \frac{1}{1/R_0 + \frac{1}{2}} \cdot 1/Z_c = \frac{2R_0}{R_0 + 2} \cdot 1/Z_c \quad \ldots \quad (1.2.5.)$$

so that:

$$Y = \frac{1}{R_0}\left(1 + jQ_L\delta\,\frac{R_0 + 2}{2}\right). \quad \ldots \quad (1.2.6.)$$

The reflection coefficient "ρ" is given by:

$$\rho = \frac{1/Y + 1 - 1}{1/Y + 1 + 1} = \frac{1}{1 + 2Y} \quad \ldots \quad (1.2.7.)$$

and substituting for Y we have:

$$1/\rho = 1 + 2Y = (1 + 2/R_0)(1 + jQ_L\delta),$$

$$\text{or} \quad 1/|\rho|^2 = (1 + 2/R_0)^2(1 + Q_L^2\delta^2) \quad (1.2.8.)$$

At resonance $\delta = 0$, $\rho = \rho_0$; at frequencies f_1 and f_2 such that $\delta Q_L = 1$, $\rho = \rho_1$, the power reflected (proportional to $|\rho|^2$] is half the power reflected at resonance.

To bring these relations into terms of the more directly measurable quantity, the field standing-wave ratio, $K = \frac{E_{max}}{E_{min}}$, we have:

$$|\rho| = \frac{K - 1}{K + 1}, \text{ and } \frac{1}{|\rho|^2} = S = 1 + \frac{4}{K + 1/K - 2} \quad \ldots \quad (1.2.9.)$$

At resonance :

$$1/|\rho|^2 = S_0 = (1 + 2/R_0)^2 \quad .. \ (1.2.10.)$$

At frequencies f_1 and f_2 where $\delta Q_L = 1$ we have :

$$S_1 = 1/|\rho_1|^2 = 2\ (1 + 2/R_0)^2 = 2S_0 \quad .. \ (1.2.11.)$$

This last formula suggests an experimental method for assessing the value of the loaded Q_L of the system, for we may measure the variation of the field standing-wave ratio with frequency in the guide, terminated by its characteristic impedance, fed by a matched generator and incorporating the t.b. as a series branch. We may then plot the parabola $S = 1 + \dfrac{4}{K + 1/K - 2}$ as a function of frequency. The next step would be to determine the value of $s = s_0$ at the resonant frequency and also for two frequencies, f_1 and f_2, at which $s = s_1 = s_2 = 2s_0$, when the value of the loaded Q will be given by $\delta Q_L = 1$; i.e., $Q_L = \dfrac{f_0}{|f_1 - f_2|}$ (see Fig. 7).

Now the unloaded Q is given by :

$$Q_0 = Q_L\left(1 + \frac{R_0}{2}\right)$$

and at resonance $K_0 = R_0 + 1$, so that :

$$Q_0 = Q_L \cdot \tfrac{1}{2}(1 + K_0) \quad .. \quad .. \ (1.2.12.)$$

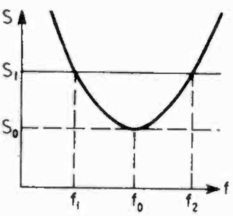


Fig. 7. *Curve of $S = 1 + 4/(K + 1/K - 2)$ plotted against frequency; K is the standing-wave ratio.*

This general method of determining the loaded Q of a parallel-resonant circuit in series with the guide,[2] although usually a convenient method, is not of easy application in this case where the Q_L values are of order 10. For this implies, since s is varying but slowly with frequency, that only a small portion of the parabola $s = \phi(f)$ can be plotted for quite an extensive set of measurements of K over a wide frequency range and that an equation of the form $s = a\ (f_0 - f)^2 + S_0$ must be fitted to what is available of the complete curve and values of f_1 and f_2 found by solving the equation

$$S_0 = a\ (f_0 - f)^2.$$

Fig. 8 shows a typical set of curves giving K and s as a function of λ (free-space wavelength). To the latter curve, the following parabola was fitted $s = 1.07 + 78(3.091 - \lambda)^2$ yielding a value $Q_L = 13.1$ and, for the unloaded Q, $Q_0 = 350$ (see also Appendix I).

The following method has been used by the authors for the determination of the Q values and has the advantage that the range of frequency over which measurements are necessary is reduced by a factor $\dfrac{Q_0}{Q_L}$ as compared with the previous methods and the further considerable merit that the quantities measured are those to which the useful bandwidth of a given t.b. cell can most directly be attributed. The latter will be discussed in a later paragraph.

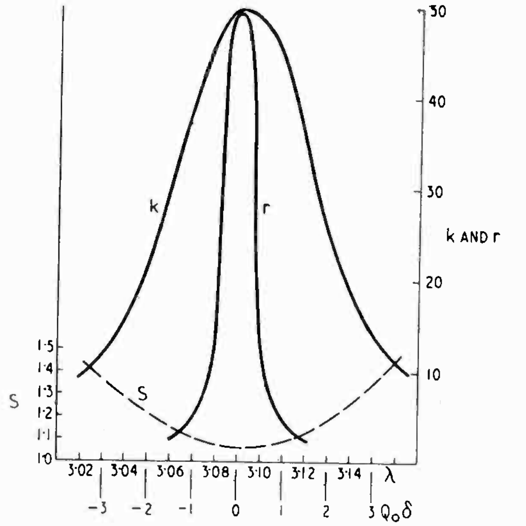


Fig. 8. *Curves of K, S and r as a function of free-space wavelength.*

Consider Fig. 9, which represents a system similar to that of Fig. 6 with the exception that the matched load in the main guide run beyond the t.b. cell has been replaced by a movable short-circuiting piston, P.

The impedance of the parallel LCR circuit is $Z = \dfrac{R_0}{1 + Q_0^2\delta^2} - j\dfrac{R_0 Q_0 \delta}{1 + Q_0^2\delta^2}$ and the impedance presented by the piston at the plane of the t.b. junction is $-j \tan \beta l$.

If now by displacing the piston we set it so that $j \tan \beta l - j\dfrac{R_0 Q_0 \delta}{1 + Q_0^2\delta^2} = 0$ then the field standing-wave ratio looking from the generator is given by $r = \dfrac{R_0}{1 + Q_0^2\delta^2}$. (It will be seen that the value obtained in this experiment for the v.s.w.r. at resonance, when $Q_0\delta = 0$, gives the value of R_0 directly.) Now if we plot the curve $r = \psi(f)$ by cancelling at each frequency the reactive component of the LCR circuit against the reactance of the length of line between the t.b. and the piston, which can be varied at will, we obtain a curve which should be a parabola near the vertex. If then two frequencies f' and f'' at which $r = r_1 = \frac{R_0}{2}$ are determined, then, from the equation above, it follows that $Q_0 = \frac{1}{\delta} = \frac{f_0}{|f' - f''|}$. The curve of r plotted as a function of λ is also given in Fig. 8 for the same system.

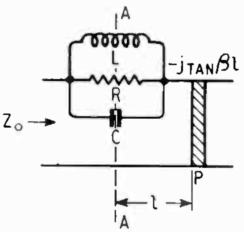


Fig. 9. *Circuit similar to that of Fig. 6, but with the matched load beyond the t.b. cell replaced by a short circuiting piston P.*

Although this method for determining Q has the advantage that a smaller range of frequencies is necessary to obtain adequate data, it has the drawback that at each frequency one must readjust the piston to cancel once more the reactive component of the cell. The experiment is, however, on the whole less laborious than alternative methods and is valuable in that the measurement of r is more directly related to the satisfactory performance of the t.b. than are the other circuit parameters, for reasons which will appear after a study of the next section.

1.3. *T.B.-cum-Magnetron Behaviour*

It now remains to consider the input impedance of the t.b.-cum-magnetron branch and to determine how use may be made of the parameters discussed above in designing a t.b. to operate most satisfactorily over a given band with given magnetrons.

The t.b. and magnetron may be considered to be effectively in series and it is therefore possible that, if the frequency band over which the fixed-tuned t.b. of finite Q is required to operate is large, its impedance at edge-band may become predominantly reactive and there will exist the possibility of a series resonance occurring between the effective reactance of the magnetron at the plane of the t.b. (AA, Fig. 2) and the reactance of the t.b. itself such that the net impedance at the t.b. plane may become resistive and fairly low.

The suggestion that it is the edge-band resistive component of the t.b. which will determine its useful bandwidth presents itself immediately and also that the ideal solution would be that of a t.b. with the highest possible value for R_0 and the lowest Q; i.e., a purely highly resistive t.b. with $Q = 1$. Qualitatively, then, the aim should be to design for a reasonably high resistive component over the band and to avoid any rapid falling off in r at frequencies on either side of the mid-band frequency. To determine quantitatively the optimum value for R_0 for operation in a specified frequency band, let us assume, as a first approach to the problem, that the magnetron may present a purely reactive impedance at its operating frequency of such a magnitude and sign as to cancel out the reactive component of the t.b. impedance. This is a more stringent condition than that normally prevailing in practice, but it is useful to consider initially the most unfavourable case possible. We impose the condition that at the edges of the band the t.b., whose reactive component has been completely cancelled by that due to the magnetron, shall still present a v.s.w.r. of magnitude say r_1. In other words, that, at the edges of the band, the resistive component of the t.b. shall be r_1. This will determine the maximum reception loss obtained when the reactances of t.b. and magnetron cancel at the plane AA.

It remains to determine the constants of the t.b. such that under these conditions its bandwidth shall be a maximum.

We have obtained above the relation:

$$r = R_0/(1 + Q_0{}^2\delta_0{}^2) \quad \ldots \quad \ldots \quad (1.3.1.)$$

This may be rewritten:

$$\delta^2 = \frac{R_0 - r}{rQ_0{}^2} = \frac{1}{Q_0{}^2}\left(\frac{R_0}{r} - 1\right)$$
$$= \frac{Z_c{}^2}{R_0{}^2}\left(\frac{R_0}{r} - 1\right) \quad \ldots \quad \ldots \quad (1.3.2)$$

The only condition imposed is that at edge band we have $r = r_1$, and hence

$$\delta_1{}^2 = \frac{Z_c{}^2}{R_0{}^2}\left(\frac{R_0}{r_1} - 1\right) \quad \ldots \quad \ldots \quad (1.3.3.)$$

It remains to determine the constants R_0 and $Q_0 \left(= \frac{R_0}{Z_c}\right)$ so as to maximize δ_1, the effective bandwidth under the conditions imposed.

It will be noticed immediately that δ_1 increases as R_0 increases and as Q_0 decreases. This result was already implied in the statement that the best t.b. would present on reception a pure high resistance.

Assuming for the moment Z_c to be fixed, in other words the geometrical structure to be fixed, we may write down $\frac{\partial(\delta_1^2)}{\partial R_0} = 0$ to determine the value of R_0 for which δ_1 is a maximum.

Write therefore :

$$\frac{\partial(\delta_1^2)}{\partial R_0} = Z_c^2\left(\frac{2}{R_0^3} - \frac{1}{r_1 . R_0^2}\right) = 0$$

whence :

$$R_0 = 2r_1 \quad \ldots \quad \ldots \quad \ldots \quad (1.3.4)$$

Thus, if we impose the condition that at the edges of the band the resistive component of the t.b. impedance shall be r_1, then for maximum bandwidth δ_1, the resistance at resonance should be $R_0 = 2r_1$.

Amplifying the implications of this result, we have seen that $r = \frac{R_0}{1 + Q_0^2\delta_0^2}$. Therefore, at edge-band, under the conditions just arrived at for maximum δ we have $r = r_1$, $R_0 = 2r_1$ and therefore, $Q_0\delta_1 = \pm 1$.

Hence the reactive component of the t.b. impedance designed as above is :—

$$X_T = -\frac{jR_0Q_0\delta_1}{1 + Q_0^2\delta_1^2} = \pm j\frac{2r_1}{2} = \pm jr_1.$$

Thus, for a purely reactive magnetron, the reactance of which may cancel that due to the t.b., we have that once the reception-loss permissible at the edges of the band has been decided upon (i.e., when the value of r_1 has been fixed), the maximum bandwidth is obtained if at resonance one has $R_0 = 2r_1$. The reactive component at edge band under these conditions will be $\pm jr_1$.

We have so far taken the worst possible case in which the magnetron is able to cancel completely the reactive component due to the t.b. It will be seen that, in a cell designed for maximum bandwidth as described above, this would imply a reactance at the plane of the t.b., due to a magnetron operating at edge-band, of $\pm jr_1$ in order to cancel completely the reactance due to the t.b. If, therefore, the magnetron impedances are such as to present at the plane of the t.b. reactances different from $\pm jr_1$, then the maximum loss appropriate to the value r_1 for the edge-band resistive component of the cell when operating with a purely reactive magnetron will never occur and the effective bandwidth will be greater than δ_1 predicted above. Moreover, r_1 and the resistive component of the magnetron impedance referred to the plane of the t.b. will be additive and the effective δ again increased. In actual practice, magnetrons are never purely reactive and the effect is thus in general to make the useful bandwidth of a fixed-tuned t.b. somewhat greater than that predicted above.

In Appendix II it is shown that with such magnetrons, in general, maximum loss does *not* occur when the reactance of the t.b. at edge-band is cancelled by that of the magnetron.

(To be concluded)

Book Reviews

Micro-Waves and Wave Guides

By H. M. Barlow, B.Sc., (Eng.), Ph.D., pp. 122, with 70 illustrations. Constable & Co., Ltd., 10, Orange St., London, W.C.2. Price 15s.

The aim of this book is to present to engineers the simpler properties of waveguides and associated apparatus in elementary yet precise terms, and to link them up with the known properties of ordinary transmission lines.

The first three chapters deal in general terms with wave propagation in rectangular and circular guides, and the properties of the wave types defined as superpositions of their plane-wave components. The fourth chapter derives Maxwell's equations directly in Cartesian and cylindrical co-ordinates without reference to vectors. The next three chapters deal mathematically with propagation in rectangular and circular guides and in the coaxial line. In the final chapter various parts of microwave apparatus and technique are described.

The main criticism of the book is that the relationship of the various modes of propagation is not made at all clear. In the early chapters the various wave types are treated generally, and the more important ones selected for mathematical treatment. For the rectangular guide this analysis deals only with the H_{0m} type, in particular the H_{01} type. The treatment postulates the simple form of electric field associated with these waves, and simplifies the field equations accordingly; but no indication is given of the procedure to be adopted for H_{nm} or E waves. Again, for circular guides only the E_{01} wave is dealt with mathematically, and, although graphs of the Bessel functions $J_0(x)$ and $J_1(x)$ are shown, no characteristic values are stated other than that appropriate to the E_{01} wave. Thus the reader is left with the impression that by suitable pre-selection these particular waves can be shown to exist, rather than that they are inevitable consequences of the general solutions of the wave equation. No indication is given of the complementary nature of E and H type solutions.

However, all the steps in the mathematical treatment are given, and it is clear and easy to follow.

Another failure to correlate is in the treatment of characteristic impedance. Two different definitions of waveguide impedance are given, and formula and curves deduced, but there is no discussion of their correspondence with the characteristic impedance of the coaxial cable (although this itself is treated), or of their use in transmission-line theory.

In short, the book gives sufficient to understand the wave types principally used, but has insufficient depth to allow an intelligent understanding of the principles of design of waveguide systems. The sections on microwave equipment can hardly be said to give more than a glimpse of the techniques involved.

The printing of the book is good on the whole but is marred by frequent use of *o* for o, particularly in inferiors, so that, for instance, H_{01} appears as H_{o1} and symbols have a tendency to alternate between roman and italic type.

H.R.L.L.

Radio Engineering (Volume 1)

By E. K. SANDEMAN, Ph.D., M.I.E.E. Pp. 775 + xxiv. Chapman and Hall, Ltd., 37, Essex Street, London, W.C.2. Price 45s.

The title of this book is conventional enough, but in most other respects it is quite unusual. The fact that it is written by a B.B.C. engineer, on a basis of matter intended for the training of B.B.C. maintenance engineers, and with the co-operation of many of the B.B.C. technical staff, is not only acknowledged in the Preface but is manifest from a perusal.

That fact imparts to the book its value and its limitations. Although of course much of the basic theory which occupies one third of the present Vol. 1 can be applied generally, a more appropriate title would have been "B.B.C. Broadcast Transmitter Engineering." Subject matter outside that field is hardly noticeable. The whole subject of receivers is relegated to Chapter XIX, in the as yet unpublished Vol. 2; while amplifiers, to the author, are r.f. amplifiers forming parts of broadcast transmitters, with the cursory exception of some devices admittedly used at a.f. between microphone and modulator. The introductory theory on Parallel Representation of Impedances is illustrated by a reference—ahead of the treatment of resonance and far ahead of valves and amplifiers—to "tuning the anode circuits of class C amplifiers." Another example of the author's preoccupation is his devoting one quarter of the first chapter to a list of B.B.C. call signs as they were in May 1944. Certain unfamiliar nomenclature—e.g., "tail" for "stub," and "Kraus aerial" for "folded dipole"—is presumably current in the B.B.C., but unfortunately the equivalents used in the outside world are not mentioned. The "inverted amplifier" is what we know as the earthed-grid amplifier, and not the inverted amplifier of Terman. Under the heading of "The Screen-Grid Valve or Tetrode" we read "This has not come into general use because of the peculiar form of its characteristics, although there are one or two examples of its use." The kinkless tetrode is not mentioned. The only example of a beat-frequency oscillator is the square-law-modulator type, presumably used by the B.B.C.; the much more usual and less tricky linear type is not hinted at. Nor is the fact that RC oscillators may be continuously variable. In a longish chapter on electrical effects and units, the m.k.s. system, now internationally established, is ignored; and there is no reference either in Vol. 1 nor in the Index (which covers Vol. 2) to v.h.f., time bases, or differentiating circuits. Radar and television are specifically excluded.

The aspects of radio excluded by the author are, in general, those on which an extensive literature already exists; so the present volume, instead of being just another radio textbook, deals authoritatively and very thoroughly with the design, operation and maintenance of broadcast transmitters,—an almost virgin field. The first eight chapters are devoted to basic principles, and a notable feature is the thorough treatment of complex impedance. Many examples are worked out to illustrate the use of the various notations. The remaining eight chapters cover the components of broadcast transmitters, including feeders and aerials. The treatment is for the practical engineer, not for the physicist. It makes only modest demands on the reader's mathematics, and provides him with many time-saving design charts. As the author points out, in a delightful touch, with reference to curves for obtaining f.m. spectra, "the fact that these are called Bessel functions does not make the curves any more difficult to read."

There is a full table of contents, but the only clues to the future Vol. 2 are in the Index, from which one can deduce, rather laboriously, that it will deal with such subjects as receivers, measuring instruments, equalizers, overmodulation limiters, negative feedback, networks and filters. One of the unconventional conventions is a self-imposed limitation on the use of full stops in the decimal section numbering, whereby what other books would denote, e.g., by "16.3.1.2" is given as "XVI.3.12," necessitating a note to point out the possible ambiguity.

Although the author claims to assume "complete ignorance on the part of the reader," and addresses himself to the "non-technical person," he has no inhibitions about the use of terms and ideas which have yet to be explained, later in the book, or in Vol. 2, or (so far as one can tell from the Index) not at all. Specific references to the literature are almost completely absent; which is a pity, as when derivations or limitations of data are not shown.

There are a few loose statements, such as "Work is another name for energy," and "if the value of the sine and cosine etc., of an angle is known, there are two angles which correspond," and the 1 000 A/in² rule for current-carrying capacity of copper conductors is unqualified by any reference to whether the conductor is a straight bare wire in the open air or an inner turn of a heavily-insulated coil; but these are not typical.

In spite of minor flaws and idiosyncrasies, and the confusion into which a reader would be thrown if he really did start completely from scratch, this book is an important contribution to radio engineering literature. It deals clearly and thoroughly with high-power transmitters, and deepens and widens a superficial knowledge of basic principles. The diagrams are worthy of special praise for their accuracy and clarity. In circuit diagrams where components values are helpful, they are shown where it is most convenient to have them—alongside the components.

M. G. S.

CORRESPONDENCE

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

Valve Operating Point with Cathode Load

SIR,—In the design of a d.c. amplifier for electro-encephalography[1] it was required to find the exact operating conditions of a cathode follower with a known voltage applied to the grid. The result can readily be obtained by use of a construction based on the static anode-current/anode-voltage curves, as shown in Fig. 1. Draw in a line of slope corres-

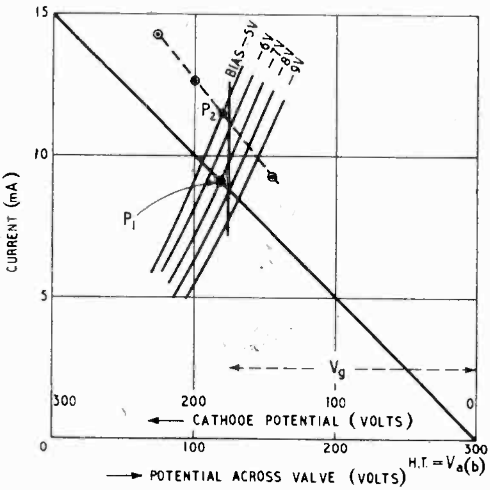


Fig. 1.

ponding to the cathode resistance R_k (20 000 ohms in example), cutting the voltage axis at the value of the h.t. applied to the valve and the current axis at $I_a = V_{a(b)}/R_k$. Where $V_{a(b)}$ is the h.t. or anode battery voltage. Mark off a voltage scale reading to the left from the h.t. voltage as zero, then the voltage read on this new scale below any point on the R_k-line represents the potential difference across the cathode resistance for the current at the point selected. Next draw in a line parallel to the current axis through the point V_g on the cathode-voltage axis equal to the potential applied to the grid (175 V). Now, the load line is the locus of the cathode potential. The required operating point must lie on the load line and it must also fulfil another condition ; viz, that the difference of potential between the cathode and grid should be equal to the bias value pertaining to the V_a–I_a curve passing through the point. In the example, the – 7-V bias I_a–V_a curve is found to cut the load line at 7 V to the left of the V_g level (at P_1), thus corresponding to the cathode being 7-V positive with respect to the grid, and satisfying the operating condition. The anode current at P_1 is 9.2 mA and the cathode potential is 182 V.

When the cathode follower is used to supply current to a non-ohmic circuit element it is necessary to take account of the current-voltage relation for the element in question ; this is done by drawing a new load line. For example, for the currents drawn by a stage of a d.c. amplifier the following figures were obtained :

Voltage applied (volts) ..	225	200	175	155
Current drawn by stage (mA)	3.0	2.6	2.3	2.0

These values of current are used to plot the dotted load line of Fig. 1 by adding the current drawn off by the external circuit at the above voltage (read on the cathode-voltage scale) to the respective current values read on the previous cathode-resistance load line. The new load line is then used as before to find the anode current corresponding to the bias $V_k - V_g$. By trial a point P_2 is found at which the – 5-V bias I_a–V_a curve intersects the new load line and for which the cathode potential is 5 volts positive with respect to the grid. At no other point on the load line will the bias value of the I_a–V_a curve intersecting it agree with the difference between the cathode potential and the grid potential. At P_2 the total current drawn is 11.5 mA and the cathode is at 180 V.

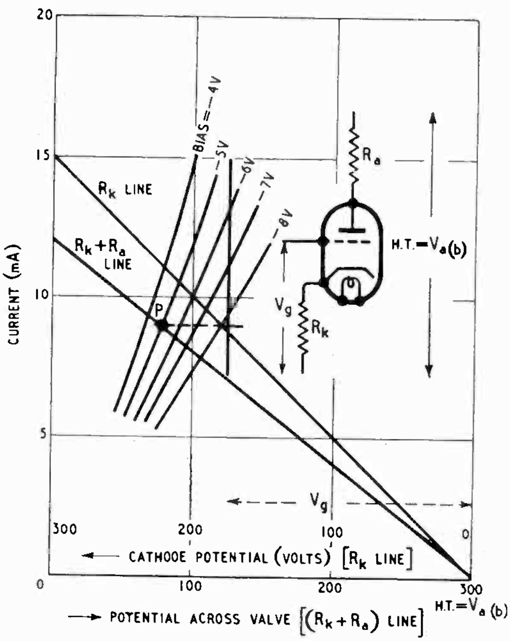


Fig. 2.

In the case of amplifying valves in circuits having a cathode resistance sufficiently high to reduce the potential across the valve by an appreciable amount as distinct from the drop in the anode load, it is possible to apply a similar method as shown in Fig. 2. The output stage of the amplifier described by Johnston [2] provides a case in which the method is of value. Draw two load lines, one corresponding to R_k and the other corresponding to the sum $R_k + R_a$ where R_a is the anode load, each cutting the voltage axis at $V_{a(b)}$ and cutting the current axis at $I_a = \frac{V_{a(b)}}{R_k}$ and $I_a = \frac{V_{a(b)}}{(R_k + R_a)}$ respectively. As before, make a scale of cathode voltage reading to the left starting from the h.t. voltage $V_{a(b)}$ as zero. Then the cathode potential on the scale under any point on the R_k–load line is the cathode potential for the current for that point, while the potential corresponding to the same current under the $(R_k + R_a)$-line, using the valve potential scale reading to the right, gives the potential across the valve. In the example, R_k is 20 kΩ, R_a is 5 kΩ, the h.t. value is 300 volts and the grid voltage is made 175 volts. The operating point on the $(R_k + R_a)$-line has to be found so that the bias value of the $(I_a - V_a)$ curve intersecting it at the point is equal to the voltage difference read between the R_k-load line and V_g at the same current level. At 9 mA in the example, the $-$ 5-V, $(I_a - V_a)$ curve intersects the $(R_k + R_a)$ line, and at 9 mA the cathode potential, read on the R_k-line, is 5 volts positive with respect to V_g so P is the required operating point. The potential across the valve is 78 volts, and the cathode is at + 180 V.

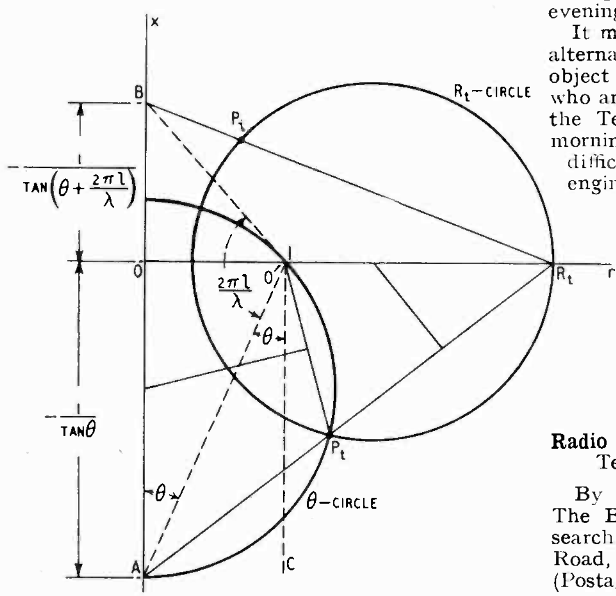
University College, London. E. J. HARRIS.

[1] Harris and Bishop, *Wireless Engineer*, December 1947, Vol. 24, p. 375.
[2] Johnston, *Wireless Engineer*, August, Sept., Oct. 1947, Vol. 24, pp. 231, 271, 292.

Transmission Line Calculations

SIR,—In the paper of W. C. Vaughan on "Transmission Line Calculations," it is stated that an accurate system of Z_i determination cannot be developed on the Cartesian Circle Diagram

I believe, however, that a construction analogous to the one shown in the paper can be extended also to the latter case and the work is as simple as that used for the Polar form.



All the θ-circles in fact intersect the r-axis at $r = 1$ (point O') and have their centre on the x-axis. If a point P_t is representative of an impedance Z_t connected to a line of length l, the corresponding θ-circle will have its centre at the intersection of the x-axis and the line bisecting P_tO' at right angles. The value R_t is then found by producing AP_t to meet the r-axis in R_t. As before the intersection of the r-axis with the line bisecting P_tR_t at right angles gives the centre of the R_t-circle.

It is clear that $\theta = O\hat{A}O' = A\hat{O}'C$; thus making $AO'B = \frac{2\pi l}{\lambda}$; the point B is the intersection of the $\left(\theta + \frac{2\pi l)}{\lambda}\right)$-circle with the x-axis. The meeting of BR_t with the R_t-circle gives the point P_i representative of the input impedance of the line of length l terminated in the impedance Z_t.

The point B can be found also by computation since $OB = -\frac{1}{\tan\left(\theta + \frac{2\pi l}{\lambda}\right)}$.

Milan, Italy. GIUS. CIONI.

Degrees for Ex-Servicemen

SIR,—With regard to degrees for Ex-Servicemen and those employed in the engineering industry during the War, I have once more been in contact with the Ministry of Education who are discussing with the London County Council and the Local Education Authorities in Essex, Herts., Middx., Surrey, Kent and Bucks., together with the Technical College Authorities, the question of the introduction of Saturday instruction. It is considered that lack of staff and the natural desire for the teaching staff to have a five-day week, may make it impracticable to arrange for Saturday training as an addition to evening work.

It may, however, be possible to arrange it as an alternative to evening work but even then the object can only be fully achieved if qualified persons who are engaged in industry are prepared to assist the Technical Colleges by teaching on Saturday mornings in order to avoid some of the staff difficulties. It is hoped, therefore, that senior engineers will see the importance of training the younger members of the industry, both in physics and engineering, and that they will offer their services to the Technical Colleges throughout the country and particularly in the counties mentioned above and in the London area.

Beaconsfield, Bucks. O. S. PUCKLE.

TECHNICAL REPORT

Radio Interference Tests on an Electrified Railway, Technical Report, Reference MT/92.

By S. F. PEARCE, B.Sc., A.Inst.P. Pp. 5. The British Electrical and Allied Industries Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 1s. 6d. (Postage 3d.)

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each.

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

586 138.—L.F. amplifier in which the cathode biasing resistance is not by-passed, the resulting negative reaction being offset by the use of a positive feed-back.
Marconi's W.T. Co. Ltd., assignees of W. R. Koch). Convention date (U.S.A.) 13th January, 1943.

586 284.—Low-frequency amplifier in which the significant signals from a photo-electric cell, or microphone, are separated from an accompanying noise voltage by means of triggered multivibrators.
Gesell. Zur Forderung &c. Technischen Hochschule. Convention date (Switzerland) 4th March, 1943.

586 502.—Loudspeaker in which a fabric formed with metallic threads or strands serves to centre and feed the moving-coil.
H. C. Willson and L. G. Aston. Application date 12th December, 1944.

586 531.—Circuit arrangement of a pair of tetrode or beam power-valves operating as a zero-bias class B amplifier.
Guided Radio Corpn. Convention date (U.S.A.) 15th July, 1943.

586 783.—D.C. amplifier wherein the interstage coupling includes a shunt valve which is arranged to produce a change of potential along a series resistance without any resulting loss in amplification.
T. H. Flowers and W. W. Chandler. Application date 17th Novmber, 1944.

586 861.—Guiding and supporting the head-carriage relatively to the lead-screw, in order to prevent uneven wear, when recording sound on an endless band.
H. H. Allen. Application date 19th September, 1944.

586 979.—Flexible support and pivoting means for the stylus of a moving-coil pick-up for a gramophone.
Philco Radio and Television Corpn. (assignees of L. J. Bobb). Convention date (U.S.A.) 11th September, 1943.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

586 410.—Mechanical modulator associated with a coaxial-line resonator for transmitting the fixed-tone frequency in a directional radio-beacon.
Standard Telephones and Cables Ltd. (Communicated by International Standard Electric Corpn.) Application date, 9th January, 1945.

586 750.—Single-channel delay network for measuring the phase-difference between two waves, particularly in direction-finding and blind-landing systems.
Standard Telephones and Cables Ltd. and C. W. Earp. Application date 11th May, 1942.

586 814.—Radio-beacon system in which the carrier-strength and signal-level are oppositely varied with changes in azimuth so as to produce a sharp "off-course" indication in a receiver fitted with a.v.c.
Standard Telephones and Cables Ltd. (assignees of W. D. McGuigan). Convention date (U.S.A.) 15th November, 1943.

586 833.—Radiolocation system in which the reflected signals are used to trigger the exploring pulses so that the repetition frequency serves to indicate the distance of the target.
Standard Telephones and Cables Ltd., P. K. Chatterjea and L. W. Houghton. Application date 20th March, 1942.

586 904.—Devices to be released from an aircraft for "scattering" the exploring pulses used in radiolocation and so frustrating any accurate determination of the position of the craft.
Marconi's W.T. Co. Ltd. and J. C. Barton. Application date 6th December, 1944.

587 009.—Parabolic reflector, with a quarter-wave radiator mounted to rotate about the focus, for scanning a field of observation, as in radio-location.
Western Electric Co. Inc. Convention date (U.S.A.) 14th August, 1943.

RECEIVING CIRCUITS AND APPARATUS

(*See also under Television*)

586 122.—Tuning and filtering arrangement for centimetre waves utilizing the variable reflection from a non-contact-making piston sliding in a resonant chamber.
Standard Telephones and Cables Ltd., J. H. Fremlin and T. A. J. Jaques. Application date 2nd October, 1942.

586 167.—Short-wave oscillation-generator comprising separate transmission-line resonators with a ganged tuning-control for maintaining a constant frequency-difference between them.
Hazeltine Corpn. (assignees of L. R. Malling). Convention date (U.S.A.) 25th September, 1943.

586 316.—Broadcast receiver with a clock-controlled relay for automatically tuning-in pre-selected stations or programmes.
Marconi's W.T. Co. Ltd. (assignees of K. J. Magnusson). Convention date (U.S.A.) 6th May, 1943.

586 339.—Beat-frequency generator which is stabilized against temperature-variations by magneto-strictive control of the main valve-oscillators.
H. Saint, R. B. Cobb and A. L. Slater. Application dates 11th November, 1944 and 3rd October, 1945.

586 534.—Receiving circuit in which two frequency-changing stages precede a super-regenerative detector valve, in order to secure a high degree of selectivity.

Marconi's W.T. Co. Ltd. (assignees of W. R. Koch). Convention date (U.S.A.) 8th September, 1943.

586 968.—Method of inter-coupling the sliding cores in a permeability-tuned superheterodyne receiver so as to ensure accurate "tracking" over a given frequency-range.

Marconi's W.T. Co. Ltd. (assignees of W. R. Koch). Convention date (U.S.A.) 11th June, 1943.

587 001.—Receiver for phase-modulated signals, comprising a frequency-modulation detector and discriminating networks for offsetting the normal accentuation of the higher audio-frequencies.

Marconi's W.T. Co. Ltd. (assignees of B. Trevor). Convention date (U.S.A.) 24th June, 1943.

TELEVISION CIRCUITS AND APPARATUS

For Transmission and Reception

586 154.—Television receiver in which a gain-control voltage is used to facilitate the separation of the synchronizing-pulses from the video signals.

Philco Radio and Television Corpn (assignees of A. R. Applegarth, Jr.). Convention date (U.S.A.) 31st July, 1942.

586 616.—Production of television mosaic-screens by a process which includes the step of printing the mosaic pattern in silver-oxide ink from an etched plate.

Marconi's W.T. Co. Ltd. and C. P. Fagan. Application date 27th October, 1944.

586 683.—Circuit arrangement comprising a series of cathode-follower valves for amplifying the output from a television-transmitter.

Marconi's W.T. Co. Ltd. (assignees of K. Schlesinger). Convention date (U.S.A.) 7th May, 1943.

TRANSMITTING CIRCUITS AND APPARATUS

(See also under Television)

585 746.—Transmitting unit for operation with the rhombic type of aerial, and particularly suitable as a military set for jamming undesired signals.

Standard Telephones and Cables Ltd. (assignees of G. T. Royden). Convention date (U.S.A.) 5th January, 1944.

585 995.—Two-wire transmission-line wound around a core of variable permeability, suitable say for feeding a push-pull amplifier from an unbalanced source of signals.

"Patel Hold" Patentverwertungs &c. A.G. Convention date (Switzerland) 8th October, 1942.

586 088.—Device for measuring standing-waves on a transmission line, comprising a pair of plane electrodes which are shunted by a diode and a variable load-resistance.

The British Broadcasting Corporation and J. L. Bliss. Application date 14th November, 1944.

586 458.—Waveguide constructed with a concertina-like section to permit the guide to be bent or flexed without loss of the transmitted energy.

The British Thomson-Houston Co. Ltd. Convention date (U.S.A.) 30th August, 1943.

586 584.—Two-stage valve-oscillator which is controlled through an impedance-inverting network from a piezo-electric crystal, suitable for frequency-modulation.

Marconi's W.T. Co. Ltd. (assignees of M. G. Crosby). Convention date (U.S.A.) 24th November, 1943.

586 619.—Construction and arrangement of tapered impedance-transformers or couplings for coaxial transmission-lines.

Standard Telephones and Cables Ltd., A. W. Gent and P. J. Wallis. Application date 16th November, 1944.

586 788.—Stabilizing the frequency of an oscillator of the velocity-modulation type by automatically controlling the voltage that controls the transit-time of the electrons.

Standard Telephones and Cables Ltd., C. N. Smyth and A. S. Grant. Application date 9th April, 1942.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

585 450.—Triode type of valve wherein an external disc is sealed to the grid in order to screen the input and output circuits, for centimetre waves.

Standard Telephones and Cables Ltd., F. D. Goodchild, and W. H. Wolsey. Application date 24th October, 1941.

585 451.—High-frequency amplifier with external metallic discs, sealed through the glass tube and connected to the grid and anode, to facilitate the direct connection of the valve to a transmission line.

Standard Telephones and Cables Ltd. and F. D. Goodchild. Application date 2nd December, 1941.

585 452.—Electrode assembly and mounting of a high-frequency amplifier in which the control-grid is formed with an external screening-disc (addition to No. 585 449).

Standard Telephones and Cables Ltd., F. D. Goodchild, and W. H. Wolsey. Application date 2nd December, 1941.

585 456.—Electrode spacing and mounting of a high-frequency amplifier in which the control-grid is associated with an external screening-disc (addition to No. 585 448).

Standard Telephones and Cables Ltd., S. G. Tomlin, and J. Foster. Application date 5th June, 1942.

585 659.—Velocity-modulation type of valve in which the resonator is made in two parts, which are capable of relative rotation to facilitate tuning.

R. A. R. Tricker and C. S. Wright. Application date 10th April, 1942.

585 752.—Arrangement for focusing and collecting the electron stream in a discharge tube of the velocity-modulation type.

Standard Telephones and Cables Ltd., J. H. Fremlin, and C. H Foulkes. Application date 14th January, 1944.

586 121.—Electrode construction and assembly of a discharge tube of the velocity-modulation type, particularly designed to avoid excessive local heating.

The British Thomson-Houston Co. Ltd., C. J. Milner, W. D. Sinclair, and L. Rushforth. Application date 30th October, 1941.

586 149.—Multi-stage valve in which the heat dissipated on the anode of one stage is utilized to liberate electrons from the cathode of the succeeding stage.

Standard Telephones and Cables Ltd. and A. J. Maddock. Application date 22nd May, 1942.