A THESIS

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INVESTIGATIONS ON ULTRA-HIGH FREQUENCY

ELECTROMAGNETIC OSCILLATIONS.

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SUMMARY.

The Thesis opens with an introduction to electrical resonators and some instruments which are special to the centimeter region of wavelengths. The excitation of resonators by the principle of the velocity modulation of electrons is then given and it is followed by a brief mention of velocity modulated oscillators and magnetron oscillators. Little specific mention is made in these sections of the author's work. A reference is however made to an early split anode 3 cm. magnetron and also to a tunable velocity modulated oscillator. These were made by him before such tubes had passed the experimental stage. The development of magnetrons required some simple but reliable method of measuring pulsed high tension power. Some satisfactory methods developed by the author are described. To prevent there being any possible doubt, these power measurements were checked by careful measurement of the heat developed in the magnetron. To examine the electric fields within resonators and feeder lines small test probes and loops are necessary. These probes and loops require a consideration of the various types of rectifiers, detector instruments and high frequency matching sections. All are discussed both from their

simplicity of design and from their performance. For the design of 3 cm. magnetrons, whose output system was to be matched to the standard fitting used in the Services, an accurate standing wave measuring instrument was built. The errors occuring in these instruments are examined and shown to be negligible only under certain conditions of operation. The general properties of wave guides and coaxial cables are given prior to an account of the author's experiments on the measurements of magnetron output powers. Mention of preliminary work on power measurement is made and is followed by a full description of a rugged glass contained helical absorber suitable for low or for high power routine measurements. Full details of the calibration of the absorbers is given. There are also details of a calorimeter for coaxial output valves and also for wave guide output valves. These instruments have given a trouble free performance for nearly two years and are eminently suited for use in factories for testing the quality of the valves produced. The increase in the output powers from magnetrons soon necessitated a complete redesign in the output system. The author's experiments on what was, as far as he is aware, the first 3 cm. wave guide output valve are included.

<u>Note</u>. In this thesis the main work of the author is to be found in sections 4, 6, 7, 9 and 10. In all these sections the work is entirely his own. He wishes however to acknowledge the very considerable help and kindness he has received from all members of the Laboratory. The author's rather sudden departure to the U.S.A. has enabled him to glance only briefly at the final type-written copies but he hopes that they do not contain any serious error.

Ralph H. V. M. Dawton.

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INTRODUCTION.

The first systematic study of electromagnetic radiation was made by Hertz in 1887. The wavelengths he used were of the order of a few metres, although waves as short as 24 cm. were obtained. After the work of Hertz most attention was concentrated on wavelengths of several hundreds of metres because such wavelengths are eminently suitable for communication purposes. In the last decade, however, the production and utilisation of very much shorter electric waves have developed enormously; they offer a special advantage in that a concentrated beam of radiation may be produced. The properties of these short waves can therefore be usefully applied to navigational purposes, and some pre-war work, see for example (12), had been published on the production of wavelengths of about 50 cm. Still shorter wavelengths were required. The main difficulty has been that it was impossible to produce powers of more than milliwatts, or at the most, a few watts (13) and (15). The techniques, which are required to produce these short wavelengths, are very different indeed from those used for longer wavelengths. An important advance in producing higher powers was made just previous to the war by using the principle of velocity modulation (7), (40); and since the war the development of centimeter radiation has been essential for radio location. Some of the most important advances have been made in the Physics Department at the University of Birmingham. It is in these laboratories that the author has been engaged from the beginning and, in fact, he is the only member of the laboratory who has had previous experience in centimeter wavelengths.

This thesis gives a detailed account of the work with which the author has been associated at Birmingham University, and can be considered as consisting of four main parts. These four parts deal with:

(1) The general techniques used, and the apparatus required for the centimeter region - section 1.

(2) The brief mention of various types of oscillators, including some which the author constructed - sections 2 and 3.

(3) The development of an instrument for the measurement of standing wave ratios - sections 6 and 7.

(4) The development of radio frequency power measuring devices, and the first experiments on high power magnetrons fitted with a wave guide output - sections 9 and 10.

The whole of the thesis is written in a form suitable for guiding an experimenter who is beginning work in the centimeter region of wavelength but who, at the same time, has some knowledge of radio in the longer wavelength region. To complete the subjects discussed, and at the same time to avoid the dissertation becoming too long, the two sections on oscillators are rather brief. In published literature there is little mention of the solution of impedance problems by circle diagrams, although Bartlett in his book 'The Theory of Electrical Artificial lines and filters' (5), makes use of them. His treatment of the circle diagram is however, more general, and therefore more complex, than is required for most circuit work. It is with the intention of encouraging the application of this method of solving impedance problems, that part of section 5 of this thesis is given to circle diagrams, although since the author commenced this section, T.R.E. (Malvern) have circulated in a form suitable for the solution of most radio problems, a report on the theory of circle diagrams (50), and J.C.Slater has published his book, 'Microwave Transmission' (17) which includes much subject matter on circle diagrams.

SECTION 1.

AN INTRODUCTION TO APPARATUS SPECIAL TO CENTIMETER WAVELENGTHS.

1.1. Introductory.

This opening section is an introduction to the essential apparatus required for work with electromagnetic waves in the centimeter region of wavelengths. It deals very briefly with the usual forms of hollow resonators, wave guides, electromagnetic field testing units such as probes, and the designs of wavemeters. By this introduction the author feels that the explanation of an oscillator, for instance, will be simplified: it will not be necessary to side track from the electronics of the discussion to explain the property of resonators, such as the rhumbatron, which forms part of the oscillator. In the same way if one wishes to explain the Hartley oscillator, one needs to explain first the properties of "L-C" circuits before proceeding to the details of the oscillator. The form that these elements take in this short wave region is however very different from that of ordinary radio circuits.

The technique for centimeter waves is nevertheless essentially the same as for the longer waves, e.g., one uses matching transformers, loading coils and condensers; although a newcomer to this field will note the absence of the familiar coils of wire and flat plate condensers, etc. As a result

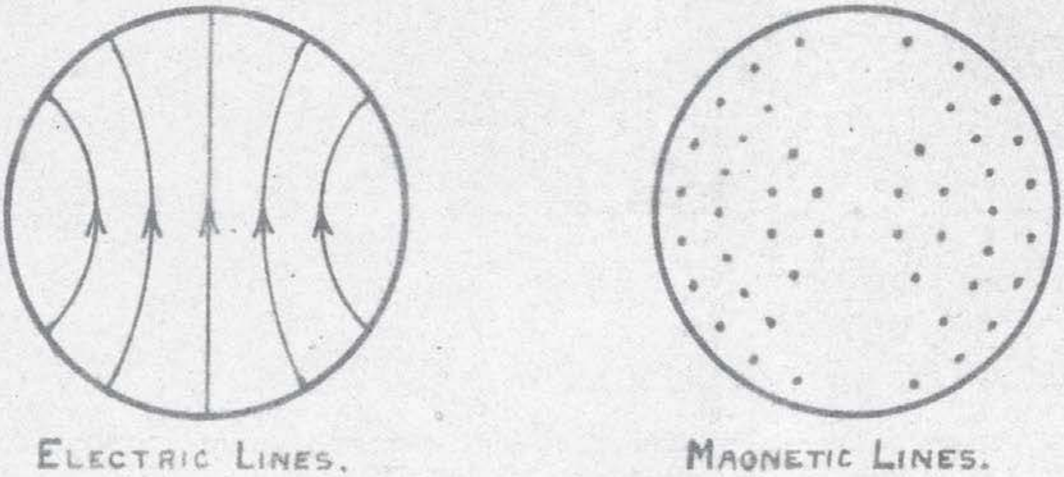


Figure 1.

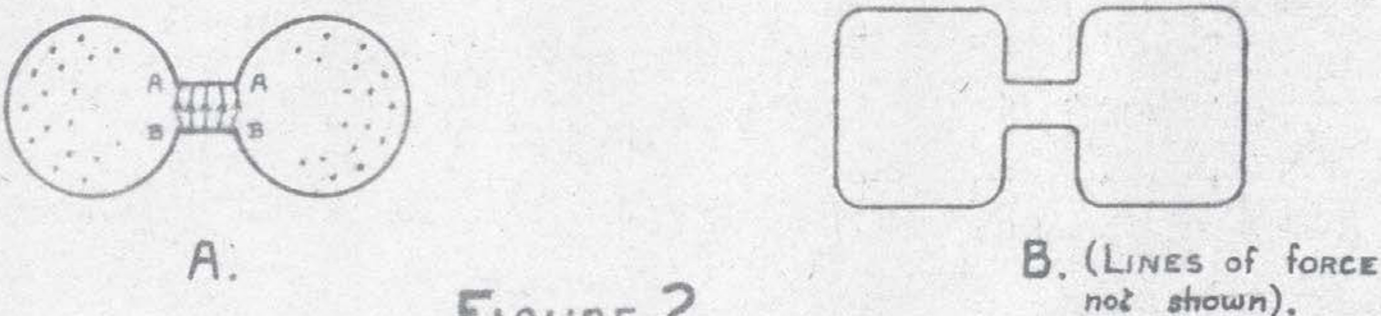


Figure 2.

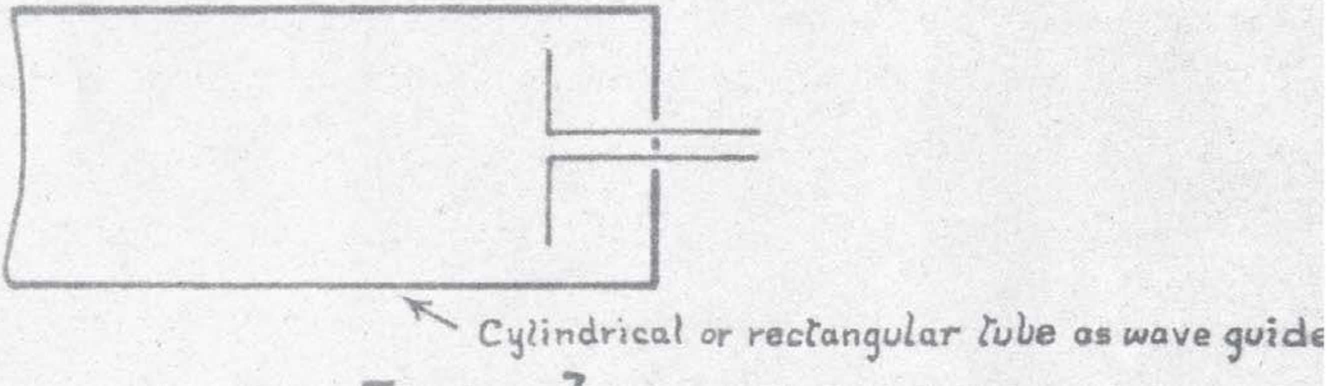


Figure 3.

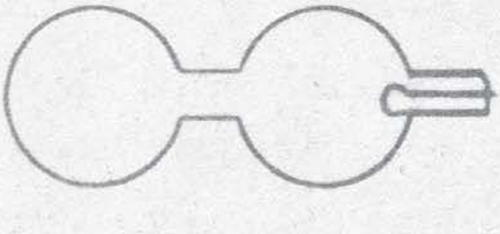


Figure 4.

he may feel that there is an unusual simplicity about centimeter wave apparatus. However such a newcomer is apt to overlook the fact that the addition of a few millimeters of wire to a feeder may alter an impedance by a hundred fold. The use of hollow boxes as resonators and the propagation of waves in metal tubes (somewhat similar to a coaxial feeder which has no central conductor) will be almost entirely new to him. He may have heard of them before, but if so, the interest was purely academic - Lord Rayleigh gave the mathematical bases of wave propagation in tubes in 1892, (14). We will now consider in turn some of the instruments mentioned in the first paragraph.

1.2. Hollow Resonators and Wave Guides.

Any closed (or nearly closed) metal container can be made to resonate electrically if the frequency employed is high enough. For example a hollow sphere can be excited; the electric and the magnetic fields for the fundamental mode of such a resonator are shown in figure 1. The mechanism by which resonance can be maintained will be explained later. It should be noted, figure 1, that the field is confined within the metal container and so there is no radiation into space. This is an important fact because at the frequencies involved the loss by radiation can be so great that efficient resonating properties are impossible, i.e. the "Q" ($\frac{\omega L}{R}$) has become low in value. At frequencies higher than the fundamental a great multitude of other resonant "modes" are formed. If the metal is clean and conducts well on its surface all modes are very sharp; that is the Q is high, being for most shapes of resonators of the order of 2000 although values up to 200,000 may be obtained. The lowest frequency at which resonance can occur gives a wavelength of the order of the diameter of the container. A very efficient and useful resonator shape is shown in figure 2A and is referred to as a Rhumbatron. A more practical form is given in figure 2B. The exact properties of these closed resonators are in general difficult to calculate (9), but the principle of their operation can be understood as follows. The resonator, as do all electrical resonators, consists essentially of an inductance and a capacity. The inductance is provided by the curved walls which have electrical properties similar to a single loop of wire open between A and B. The capacity is mainly, but not entirely, provided by the two surfaces AA and BB which form a parallel plate condenser and which are connected to the inductance just mentioned. Oscillating currents flow between A and B via the walls. The currents produce a magnetic field (which differs from that of a loop of wire in that it can occupy only the inside of the conducting enclosure). This magnetic field forms closed circles whose centres lie on the axis of revolution of the rhumbatron and hence it is only as a section that these lines appear in figure 2A. The electric field occurs mainly between AA and BB, and is shown in the same figure. Reference will frequently be made to these resonators and before concluding these preliminary remarks on them the following three important properties should be noted:-

(1) The electric field is located almost entirely between AA, BB.

(2) The distance between AA and BB can be small compared with the wavelength of the electrical oscillations, say $1/20\lambda$. (The advantage of this will be made clear in section 2).

(3) The magnetic lines of force form circles concentric with the principal axis of the rhumbatron.

Very brief mention will now be made of hollow wave guides. These are usually referred to simply as wave guides, and are often used for the transmission of power. They are metal tubes along which electric waves are propagated. They offer the advantage that relatively low losses are incurred. An account of wave guides is given in section 8: a simple example of an antenna feeding a wave guide is given in figure 3.

Wave guides and hollow resonators can have their electromagnetic fields examined by suitably positioned test probes or loops. Thus in a rhumbatron the loop is placed as in figure 4. Care must be taken that the test units do not noticeably distort the field of the guide or resonator. The author's examination of this important condition and the precautions necessary to be observed are given in section 6 and 7. A similar probe or loop arrangement can be used to excite the resonator or guide, except that now power is fed into the probe or loop from an oscillator of suitable frequency.

1.3. The measurement of Wavelength - Wavemeters.

The method used to determine the wavelength of a radiation depends very much on the accuracy required and the power available. For example, with an oscillator delivering a few watts, a rough indication can be obtained by running a neon lamp along a pair of lecher wires. Sometimes by sliding a bridge along the system the voltage nodes and anti-nodes can be detected by watching the reaction on some circuit, such as the anode circuit, of the oscillator. In such determinations, when too little attention is given to the change of conditions of the oscillator, the positions of the minima are not always mid-way between the maxima. Best results will then be obtained by measuring from one minimum to another (and not from a minimum to a maximum); but obviously measurements under such conditions are not advisable. For precise measurements it is essential not to disturb the oscillator. With lecher wires, loosely coupled to the main circuit, accurate results can be obtained by placing a sensitive probe (to be discussed later) near the wires. The response from this probe will go through a succession of maxima and minima if a short circuit bridge is moved along the wires. By

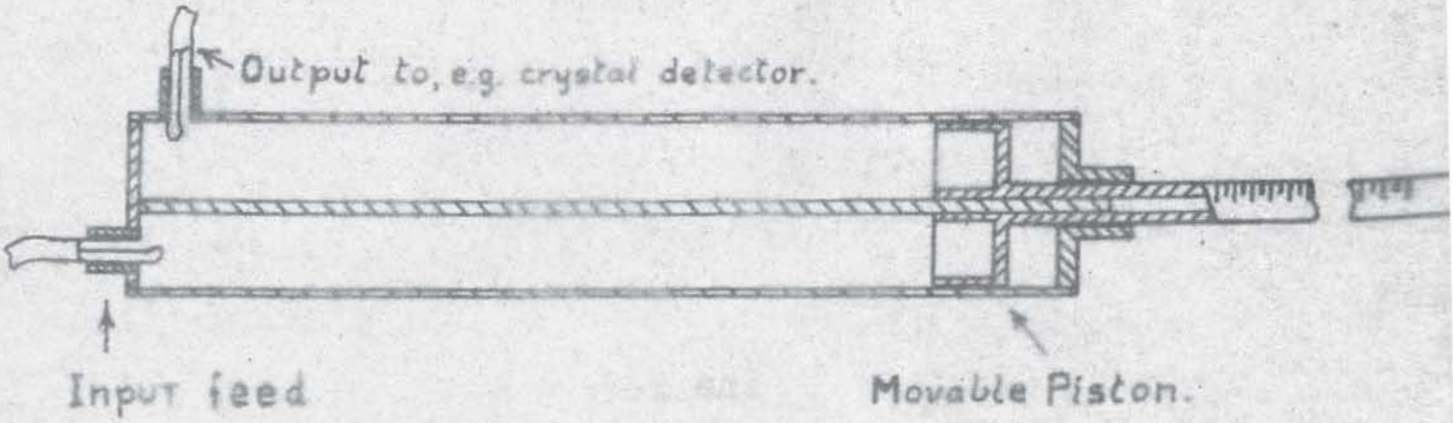


FIGURE 5 COAXIAL WAVEMETER.

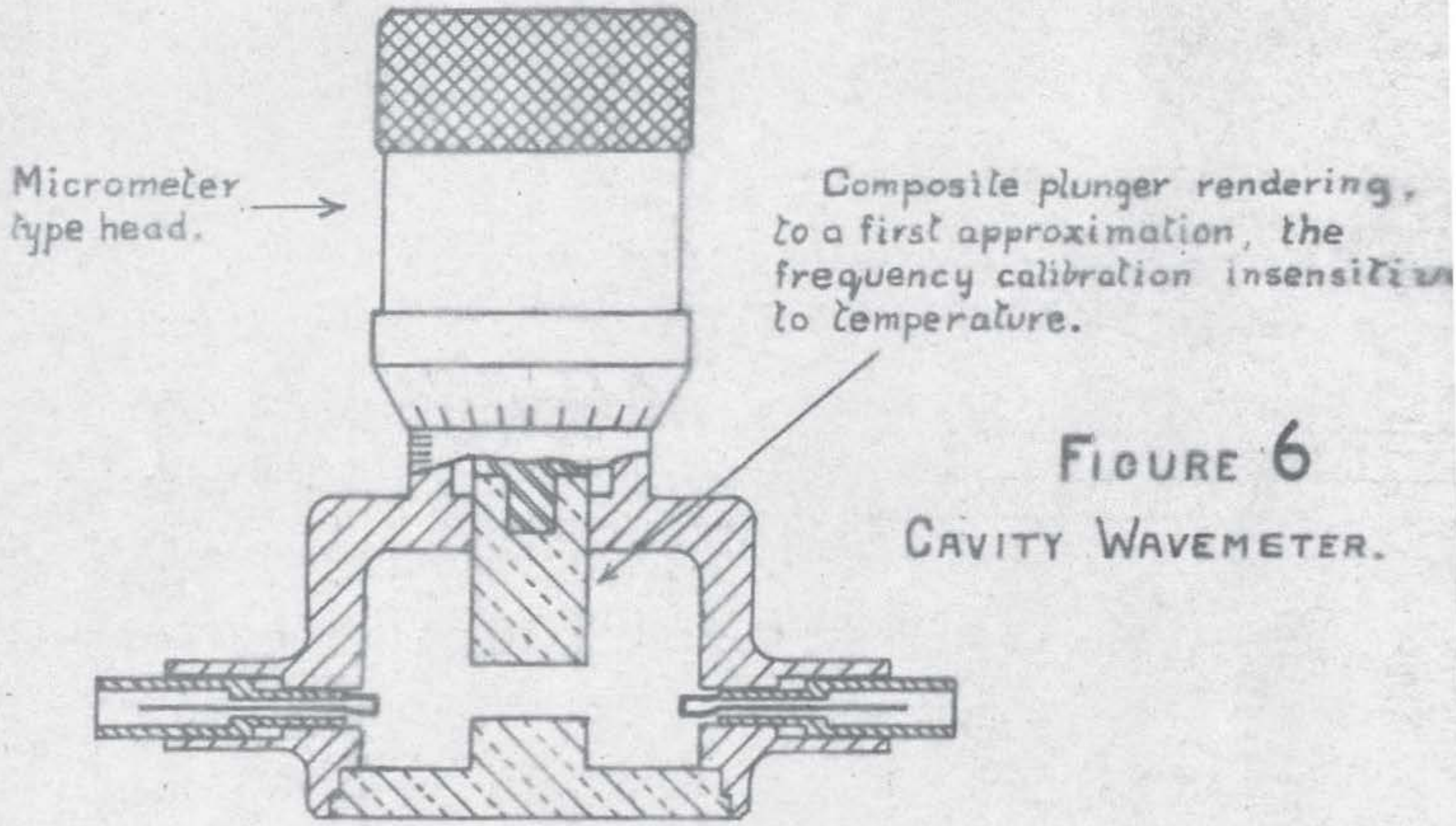


FIGURE 6
CAVITY WAVEMETER.

NOTE The insertion of the loops is variable, but there is a standard insertion corresponding to that required by the calibration chart.

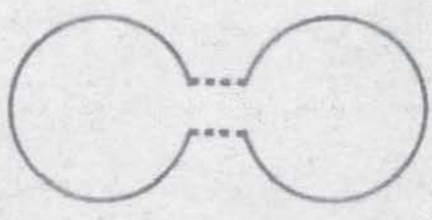


FIGURE 7.

In practice the grid is made finer than is indicated here, so as to cause less obstruction to the beam.

observation over a succession of say 20 λ a considerable accuracy is obtained.

A more practical form of wavemeter can be made by using a coaxial line in a similar manner. This method has the additional advantage of being non-radiative. As a wavemeter it is less easy to construct but when well made can yield very good results. The general arrangement is given in figure 5. The reference (38) also deals with this type of wavemeter.

Measurement of wavelength, by the method just explained, suffers from the disadvantage that a single observation does not give the result: at least two observations are necessary and these often several wavelengths apart. A further difficulty is that if more than one wavelength is present (and this often occurs with magnetron oscillators) the resulting series of maxima and minima are very confusing. To overcome these difficulties resonant cavity meters are used. A type first produced by Sayers of this laboratory is shown in figure 6. The resonant cavity has some resemblance to the rhumbatron, and its resonant period is controlled by the moving plunger. To determine a wavelength the input loop is loosely coupled to the radio frequency source and the thimble turned until a peak deflection occurs in the indicator system (section 6). The thimble scale now enables the wavelength to be read from the calibration chart which, with a suitable design of cavity shape, can be made sensibly linear. The relative accuracy is of the order of 1 part in several thousand while the absolute accuracy depends on the calibration and the faithfulness of the instrument in maintaining the calibration; e.g., the wear of the micrometer screw will produce errors. At their best the instruments have an absolute accuracy of 1 part in a few thousand. They are calibrated by harmonics from a quartz crystal.

<u>A Wavemeter to determine small frequency changes</u>.

To determine changes of frequency of the order of a Mc./sec. in a wavelength of 3 cm., and instrument capable of discerning changes of 1 part in 10^4 is required. For this one uses special high Q meters. These are similar in design to the resonant cavity instrument just described but employ a resonator excited in a mode to give the highest Q possible for practical working. An account of this wavemeter can be found from reference 39 in the list at the end of this work.

SECTION 2.

VELOCITY MODULATION OSCILLATORS.

Summary.

In this section it is explained how a rhumbatron can produce a "bunching" of an electron beam, and how this beam can then give high frequency energy to a second rhumbatron. Some oscillators using this principle are discussed.

2.1. One use for the Rhumbatron.

In the first section there was, among other things, an outline of the principal properties of the rhumbatron. It was mentioned in particular that the high frequency field of the resonator was concentrated between two surfaces and that these surfaces were not very distant from each other. We shall now see how this property is utilised in the design of oscillators where interaction between the field and an electron beam takes place. It is necessary however to understand first the principle of velocity modulation of an electron beam.

2.2. Velocity modulation and Bunching.

If a high frequency field acts along the direction of movement of electrons, some electrons will be accelerated and some retarded according to the phase of the field during the transit of the electrons. As a consequence of this a uniform beam of electrons becomes a beam in which the electrons are not all of a uniform velocity, (for some electrons will be moving faster than previously and some slower than previously): the beam has been "velocity modulated". (In actual fact the electron takes a finite time, say 1/5 of a period, to cross the high frequency field and so it is not correct to speak of the phase of the field during the electron time of transit but more exactly a range of phases must be considered. The general principle however is still the same and we will neglect the extra complication).

The result of this velocity modulation is that further along the beam the faster electrons catch up with, and the slower ones fall back upon, the electrons whose speed had not been altered. In other words clusters or bunches of electrons are formed and the beam becomes intermittent. (This last sentence, while bringing out the nature of the effect, is an exaggeration for more correctly the beam remains continuous but becomes density modulated). Still further along the beam the electrons continue to rearrange themselves and the bunching becomes destroyed, but this is not here of primary importance: the important conclusion is that velocity modulation at one point of the beam produces bunching at another point. From this simple explanation the smallest velocity modulation will produce bunching if the beam is allowed to continue along its path for a sufficient distance. This implies that an amount of energy however small will bunch the beam. In practice however de-bunching forces (due to mutual repulsion of the electrons) have to be considered. Nevertheless given efficient resonators the energy required to produced a bunched beam is very small.

Velocity Modulation by a Rhumbatron.

In section I it was explained that the high frequency electric field in a rhumbatron occurred between two surfaces whose separation was small compared with the wavelength corresponding to the resonant frequency. By replacing these surfaces by grids (figure 7) the performance of the rhumbatron is not materially altered, but an electron beam travelling parallel to the grids can now be passed through the resonator. This direction of travel of the beam is suitable for the field of the resonator to influence the electron velocities in a manner as explained in the previous section. By this means a rhumbatron can be used to bunch an electron beam; such a rhumbatron is usually referred to as a buncher.

2.3. Interchange of energy between electrons and Rhumbatrons.

Electrons crossing a rhumbatron when the phase of the high frequency field is such as to accelerate them, gain energy from the field. Conversely, if the transit is performed when the

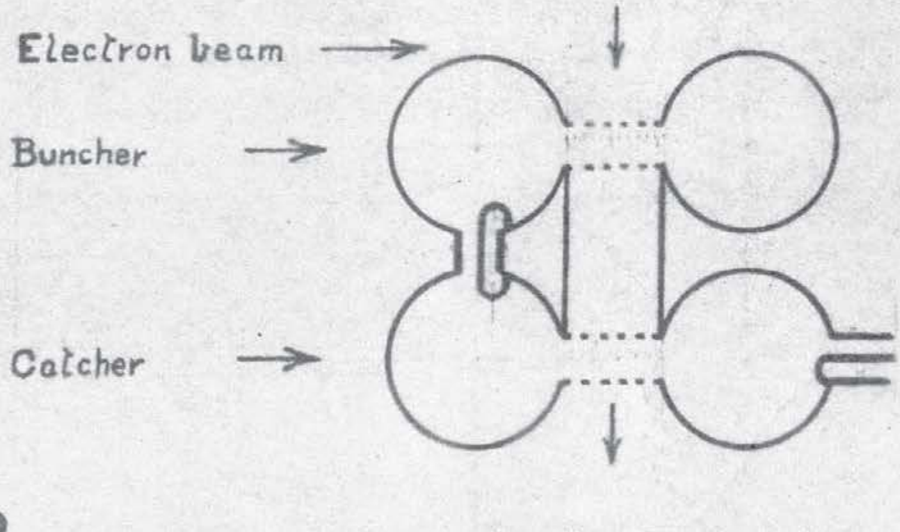


FIGURE 8 Diagramatic drawing of a Klystron.

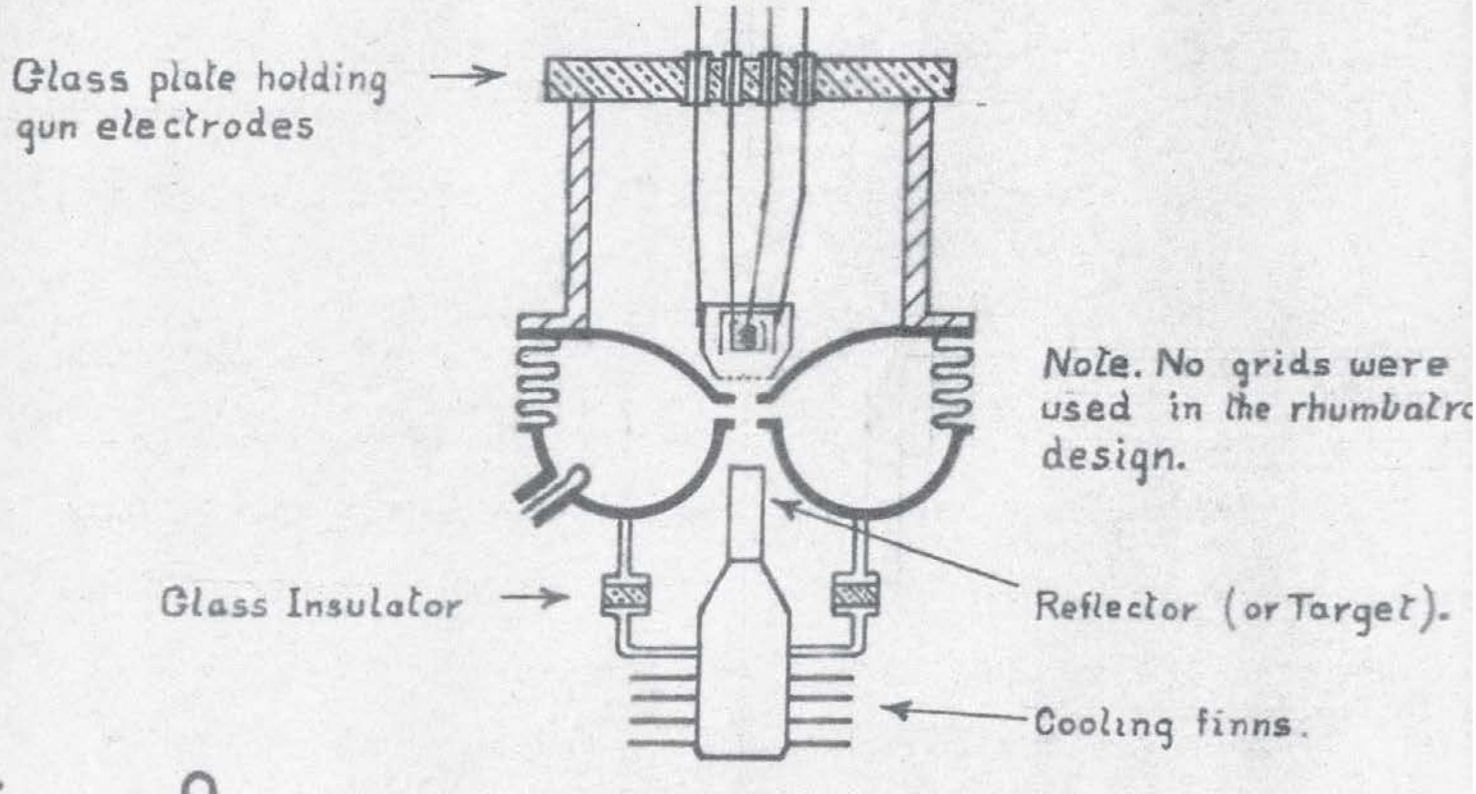


FIGURE 9 One of the author's early tunable oscillators of a demountable design. Wavelength 7 to 9·5 c.m.

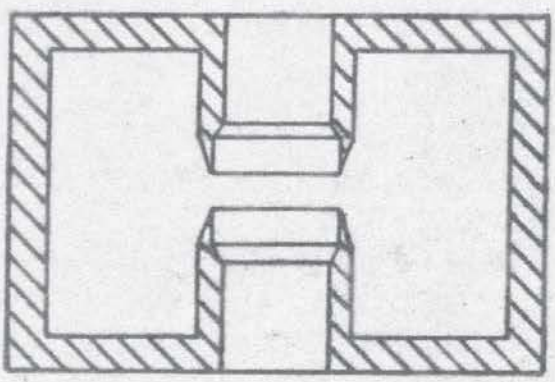


FIGURE 10 This shows the knife edges and the recesses to prevent the secondary emission of electrons at the rhumbatron gap.

field retards the electrons, energy is given to the field and the high frequency oscillation tends to be maintained. Thus, if the bunches of electrons are timed to cross another rhumbatron when its high frequency field opposes them, energy can be transferred from the beam to the high frequency field of the rhumbatron. In practice the beam is never perfectly bunched and some electrons may abstract energy from the field but these electrons should be very much in the minority. The gain in high frequency energy is obtained from the reduction in the mean velocity of the electron beam, i.e. in effect, from the potential originally used to produce the beam.

It is evident that power oscillations in a Rhumbatron can be maintained by passing through it a modulated electron beam. The modulation of this beam is usually produced by a buncher rhumbatron which is excited by a feed back of energy from the power oscillator. This principle is that of all velocity modulated tubes. An example of such a tube is the klystron, a diagrammatic drawing of which is given in figure 8.

2.4. Development of Velocity Modulated Oscillators.

These oscillators have the characteristics of all velocity modulation devices in that there is a number of well defined beam voltages at which oscillations occur. This property is due to the fact that for the efficient transfer of radio frequency power to the catcher, the transit time of an unmodulated electron from the buncher to the catcher is given by $t = 2\pi(n \pm \frac{1}{4})/\omega$. In this n is an integer and ω the angular frequency of oscillation. The minus sign corresponds to the "in phase" mode, i.e. when the buncher and the catcher voltages are simultaneously either accelerating or retarding the electrons, and the positive sign to the "anti phase" mode. Because of the coupling between the resonators these two modes of oscillation differ slightly in frequency.

The development of the Klystron followed two distinct branches; (1) the low powered continuous wave oscillator whose wavelength can be tuned. It is suitable as a local oscillator for receivers.(2) The high power Klystron transmitter.

<u>The low power Klystron.</u>

In the latter part of 1940 the Sperry Klystron was produced in America (41). These were tunable oscillators and operated at about 1,000 volts H.T. supply, with a maximum current of about 40 mA. Oscillation could however be obtained with currents of the order of 5 mA. The power output under suitable conditions was as high as a watt and the corresponding efficiency about 2.5%. Tuning was possible over a range from 9 cm. to 11.5 cm. This was attained by distorting the resonators to change the spacing between the grids, and hence the capacity of the resonator. The frequency stability of these tubes was satisfactory. Two causes of frequency drift are possible:-

(a) Thermal drift. This is caused by heat expansion of the resonators. In the Sperry tubes it amounted to about 2 or 3 Mc./sec. in the first three minutes from switching on. After this time the frequency change became very much less.

(b) Frequency pulling. This is mainly caused by a change in the H.T. voltage supply. For these Sperry oscillators the pulling was of the order of ¼ Mc/sec. for a change of 40 V. in the high tension voltage.

Another form of Klystron oscillator is the single rhumbatron type which utilizes the reflex principle for obtaining feedback. The electron beam is velocity modulated as it passes through the rhumbatron, and after this first passage it is acted on by a retarding electric field. This reflects the beam, now bunched, back through the resonator. The mechanism of these oscillators is more complex than with the double rhumbatron klystron tubes, because, in the reflection tube, the point at which the electrons turn back is a function of their speeds, i.e., of the velocity modulation they have received during their first transit through the rhumbatron. Greater power can be obtained by replacing the reflector by a secondary emitter, for example an aluminum electrode at 60 V. positive with respect to the cathode. This electrode then yields a secondary emission and this emission returns to the resonator as a bunched beam. These reflex oscillators are often more efficient than the double resonator klystron. This, combined with the simplicity of tuning, makes reflex tubes preferable as local oscillator sources for superheterodyne receivers.

Much attention was given to the development of these reflex oscillators by Rollin and Sutton in England. Tuning was not by distorting the rhumbatron as was the practice for the klystrons, but by altering the depth of a metal plunger inserted into the side of the rhumbatron. The Author early in 1940 constructed a demountable reflex oscillator tuned by compressing or extending the rhumbatron, which was made from metal bellows, figure 9. This oscillator was used as a short range transmitter for experiments by Moon and the Author on a 9 cm. super-regenerative receiver, (26).

2.5. Development of High Power Klystrons.

At the outset of the war Sayers began development work on the klystron and in 1941 published a report on experiments with a Birmingham ½ k.W. 6.7cm. Gun laying set. A novel feature was that the klystron provided both the transmitted pulse and the local oscillator frequency. This was effected by altering the high tension voltage to change the oscillator from the "In phase" mode to the "Anti Phase" mode, (28).

In all klystron oscillators the factor which limits the output power is mainly the difficulty of obtaining a sufficiently intense electron beam to the second rhumbatron. Another difficulty discovered by Sayers, and closely connected with that just mentioned, is that secondary emission of electrons near the rhumbatron gaps exerts a heavy damping on the electric oscillations. Much of this damping can be averted by the rhumbatron gaps being recessed and formed between "Chisel Edges" see figure 10. In the latter part of 1941 British Thomson-Houston Co.Ltd. undertook to produce a sealed-off klystron from one of the Sayers demountable klystrons. No fundamental change in design was necessary apart from a slight change in the electron gun. This was necessitated by the degree of vacuum in the sealed-off oscillators differing from that of the demountable type. These klystrons were not tunable but three designs were produced which gave wavelengths of 7.0 cm. 9.1 cm. and 10.7 cm. respectively; the tolerances being ±2%. Efficiencies were of the order of 10% - 20%. A continuous output of 300 W and a pulsed output up to 2 k.W. were obtained with lives of 300 to 500 hours.

SECTION 3.

THE MULTI- RESONATOR MAGNETRON.

Summary.

A brief survey of the development of the multi-resonator magnetron at Birmingham University.

3.1. The multi-resonator magnetron.

The pre-war magnetron, generally known as the split-anode magnetron, is very widely used to produce deci-metre waves and its properties are so generally well known that a detailed discussion is not necessary. In the centimeter region, however, the practical limit to electron speeds required a considerable reduction in the physical size of the magnetron. This resulted in a big decrease of the maximum allowable anode dissipation and therefore in the output power obtainable. For example, immediately before the war the author, then at the Royal Institution of Great Britain, was asked to begin work similar to that of Cleeton and Williams (6). The author's valve was made tunable by enclosing within its envelope a lecher system whose length could be varied by external control. On account of the small anode size, 2 mm. in diameter and 5 mm. long, the output was only about 1/5 watt. Wavelengths over a range of 2 cm. to 10 cm. were obtained.

The reduction in size is not the only difficulty, fc at wavelengths less than 15 cm. the conduction losses in

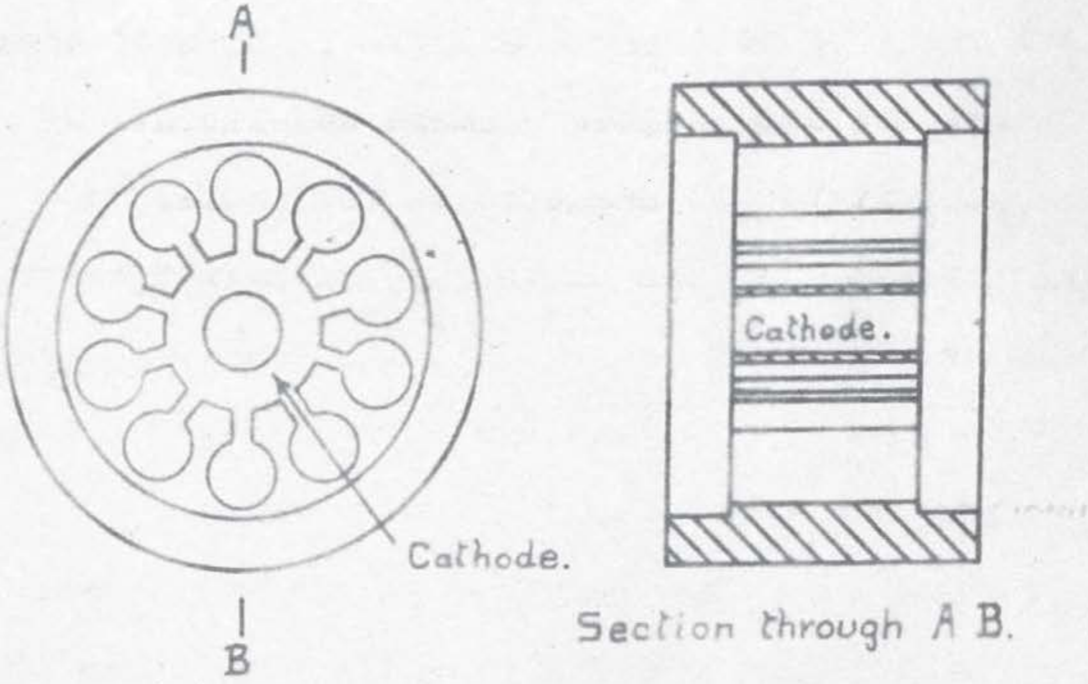


FIGURE 11 Multi-resonator Anode with Cathode.

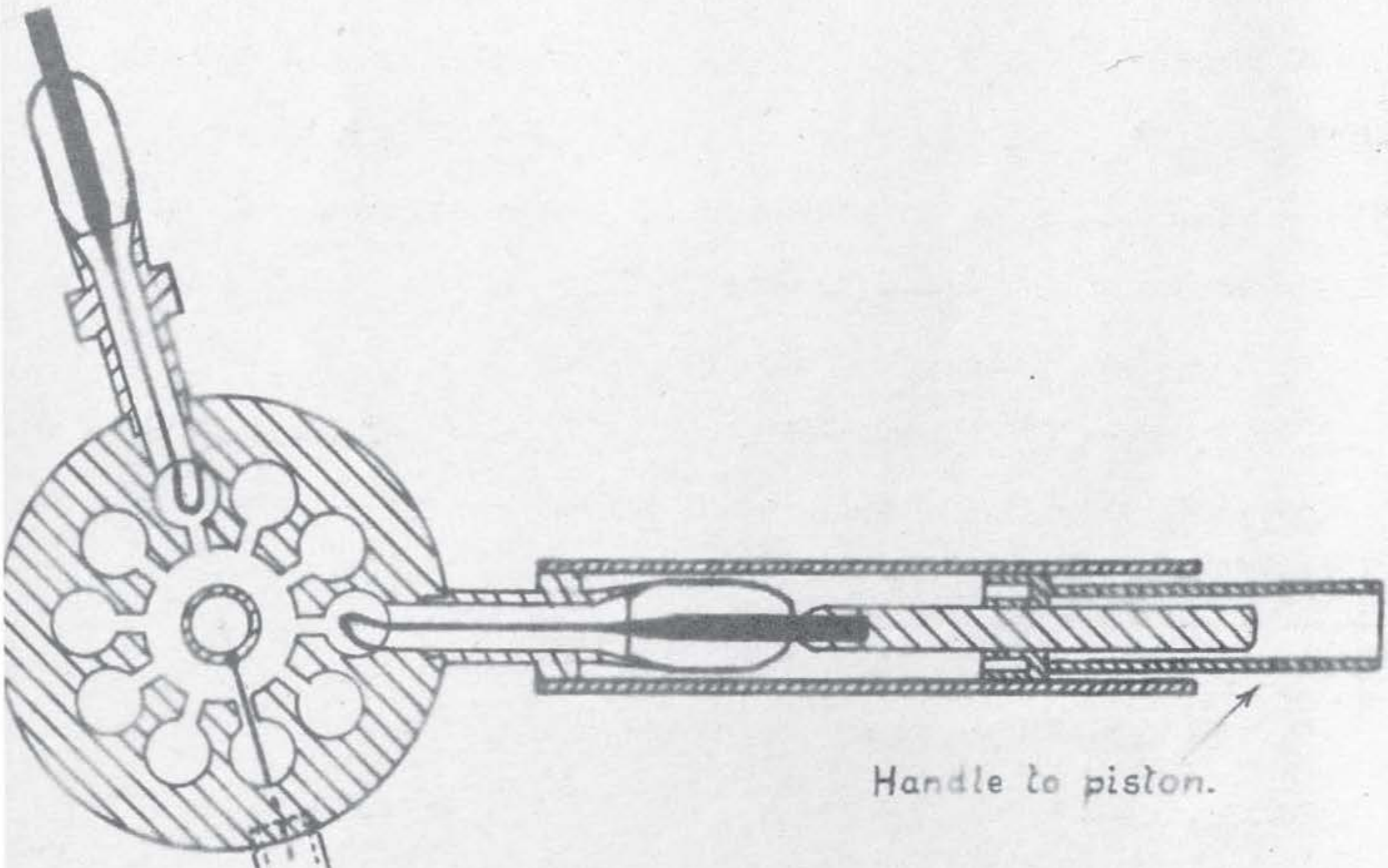


FIGURE 12 A COAXIAL LINE, USED as a variable reactance, is employed to change the frequency of the magnetron.

coaxial lines begin to be serious and the problem of losses by radiation becomes more acute. These considerations limiting the performance of magnetrons, led Professor Oliphant and his staff at the University of Birmingham to explore the possibility of using a multi-resonator anode enclosed within a metal envelope. Early in 1940 this new type was produced by Randall and Boot with extremely satisfactory results. The magnetron consisted, in essence, of a multi-resonator anode surrounding the cathode as in figure 11.

The exact operation of these oscillators is not fully understood, but Hartree, Slater, Stoner and others have published many papers (25) (34) (35), some in very great detail, on the mechanism of these valves. Of the many results from these theoretical papers, two, which are very important, are:

(1) The resonators give rise to a wave which is rotating with a phase velocity say ϕ per segment. This velocity is subject to the condition that, if N is the number of segments, $N\phi$ is a multiple of 2π.

(2) The electrons in the cathode space charge rotate with various angular velocities. For the successful operation of magnetrons the angular velocity of a certain layer of electrons in the space charge cloud bears a definite relation to the angular velocity of the anode potential wave, mentioned under (1).

In some designs of magnetrons it has been shown by Sayers and Sixsmith (32), that the space charge cloud is rotating much slower than the anode wave. In this case the space charge, instead of moving in step with the anode wave, is moving in step with a spatial harmonic of the wave. All these effects are extremely complex and are not considered in further detail at this stage.

From the practical consideration the relatively massive anode and heavy electron current (sometimes over a hundred amperes) make the magnetron capable of delivering very high peak powers. This resulted in its extremely rapid development and its general preferment to the Klystron when high power was required. At the time of writing peak powers of over a megawatt are possible. The high efficiencies now obtainable, often over 50%, were first produced by Sayers who connected segments of similar phase by copper conductors. This procedure is known as strapping, or as mode locking, and the next two paragraphs will be devoted to its behaviour in practical application.

3.2. The Strapping of multi-segment resonators.

The travelling wave of the resonator need not be limited to any one number of repeats round the anode. Thus a system of 12 resonators can have any number of repeats up to six. To each of these modes of oscillation corresponds a particular resonance frequency. In the power oscillator several of these frequencies can be excited. Their separation is only of the order of 2%, and as a result frequency instability can occur, due to changes of external loading, or, due to changes in the electronics of the valve. From the electrical stand point the straps behave like reactances which, to an approximation, have but little effect on the required mode (for the voltage phase at each end of the straps is the same) and a comparatively powerful effect on other modes. Thus in a 9 cm. unstrapped valve the separation of the unwanted frequency from the working frequency is a few per cent whereas with a strapped valve, the separation can be better than 25%. The strapping does not prohibit modes. It increases their separation thus tending to prevent a change from one mode to another. The conductivity of the straps is also advantageous in providing more equal loading of the resonators

On 9 cm. valves, strapping was an immediate success but on 3 cm. operation the great improvements expected were not at first obtained. Sayers and the author, while developing an early 3 cm. magnetron (29) attempted to strap it but received little success. The author then made some resonance experiments, i.e. he determined the various frequencies of the resonator system. The results showed that very little increase in separation occured with strapping. The reason was that, measured in terms of wavelength the strap length was about three times the length used on 9 cm. valves. When the 3 cm. magnetron was reduced in size, the expected improvement in performance was obtained.

3.3. Cathodes for Magnetrons.

The problem of cathode life is of extreme importance. The chief cause of short life is the high temperature of the cathode. This temperature is produced by the back bombardment of electrons. Important properties of the cathode are:-

(1) The design should be such that the heat is conducted or radiated away without too great a rise of temperature.

(2) The thermionic and the secondary emission should be high.

(3) The cathode materials should not be subject to appreciable vaporisation or flaking.

Further mention of cathodes will not be made in this thesis except for mention (27) of the publication from Birmingham University of a report "Cathodes for magnetrons" by J.T.Randall, H.Tomlinson, W.T.Cowhig. This report deals with the general property of cathodes and gives results for radiator-type cathodes, cathodes formed by Alkaline earth compounds on molybdenum, thorium oxide cathodes etc.

3.4. Tunable multi-resonator magnetrons.

The principal deficiency of magnetrons has been the complete inflexibility of their operating frequency. Within the last year however some fairly successful attempts to tune 9 cm. valves have been made. Electrically the

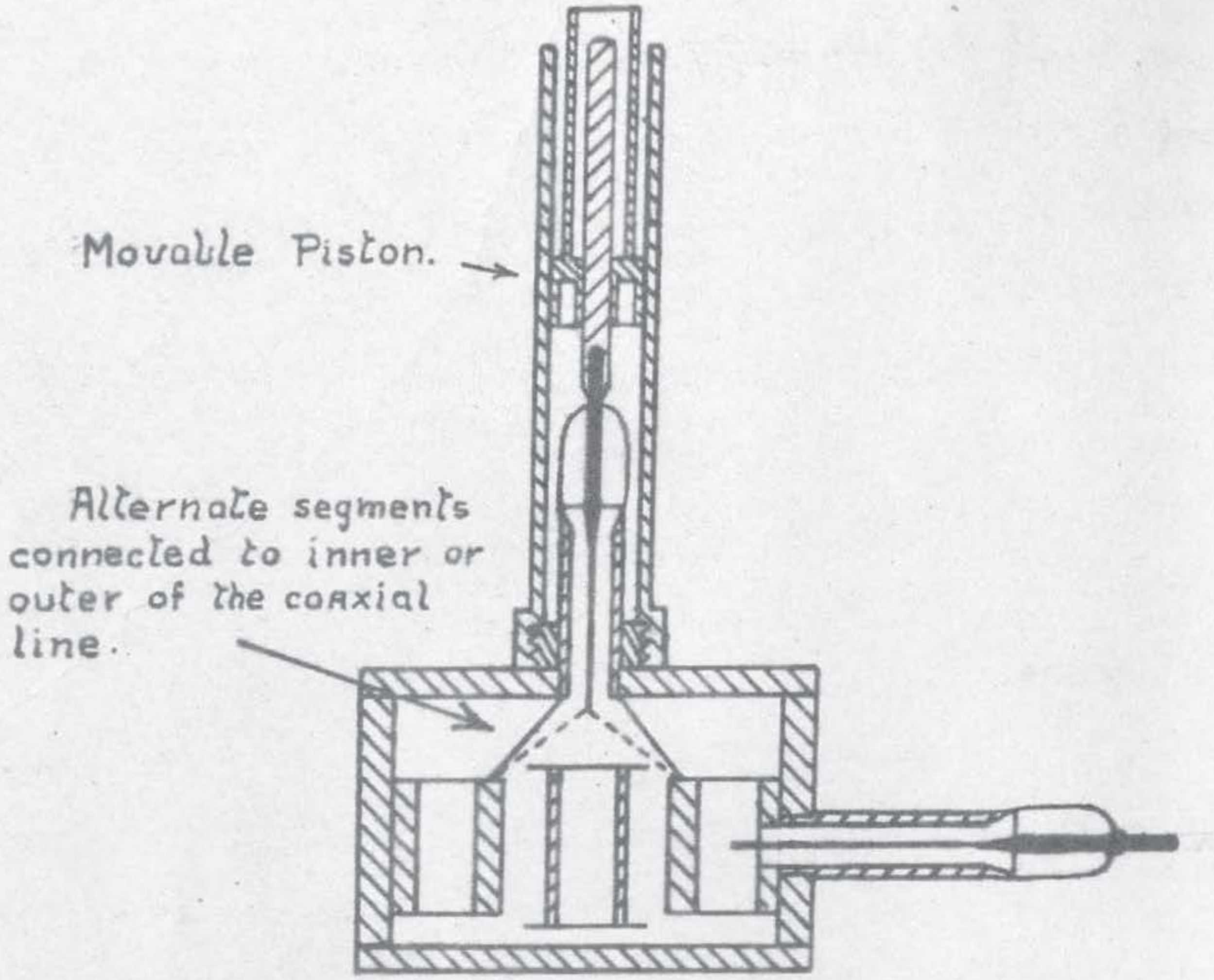
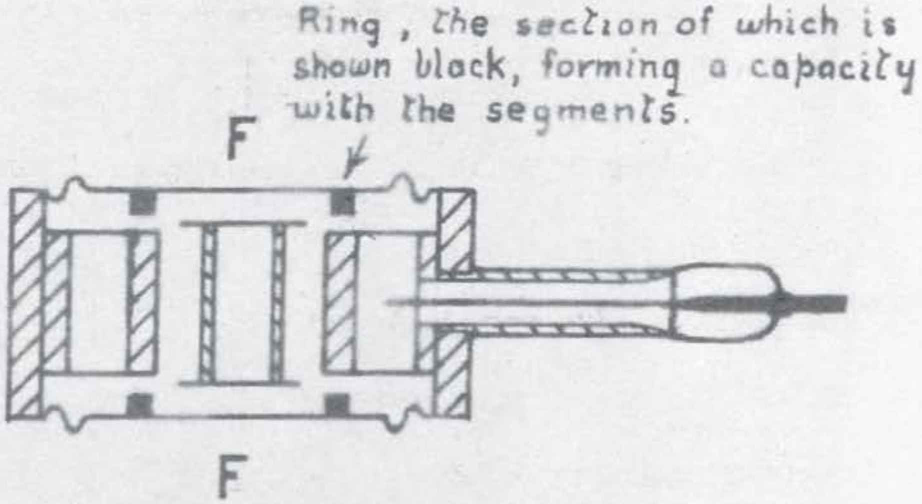


FIGURE 13 TUNABLE MAGNETRONS.



F represents a screw operated flexible diaphragm. The screw mechanism is not shown.

FIGURE 14 TUNABLE MAGNETRONS.

magnetron can be considered as a system of coupled L-C circuits which must be operated on to change the wavelength. In general there are two ways to alter the resonant frequency of a circuit:

(a) To couple into the circuit another circuit which can be tuned.

(b) To alter the reactance of one or more of the component parts of the system of L-C circuits comprising the magnetron.

A first example of (a) is the coaxial tuner coupled directly to the cavity figure 12. By changing the position of the plunger in the tuner the wavelength can be changed by about 1%. Another method is to couple the reactance direct to the straps. To do this the strapping can be modified by connecting the odd segments to the outer of the coaxial and the even to the inner, as in figure 13. This method produces wavelength changes by as much as 10%. See reference (43).

An example of the second method (b) is the end plate tuning. It is somewhat similar to that employed in earlier days by the author to tune a cylindrical klystron. It consists of a conducting plane (or preferably a ring) whose distance from the ends of the segments can be altered, see figure 14. The chief effect is the variation of capacity between the segments and the end plate and this causes an alteration in the resonant frequency.

Output feeds from multi-resonator magnetrons.

The output feed fitted to all magnetrons has been almost invariably the coaxial line. It had the advantages of simplicity of attachment. Further by suitable stubs, reasonably good matching with the external load could be obtained without undue trouble. There are however 3 cm. valves and 9 cm. valves in this laboratory of such power that the full output can never be approached because of voltage arcs and heating within the coaxial feeder. This had been foreseen and the possibility of a waveguide output feed was attractive as a cure. The author's experiments on what became as far as he is aware, the first 3 cm. valve with this type of output, are given in a later section.

SECTION 4.

THE MEASUREMENT OF INPUT POWER TO MAGNETRONS.

Summary.

Some methods of measuring the pulsed power supplied to magnetrons are described and the accuracy of the methods discussed. The agreement between the results is to within 2%.

4.1. The necessity for reliable measurements.

The very promising performance of magnetrons led them to be favoured for many operational uses in the Services. Development work was therefore concentrated on them at the University of Birmingham and at other establishments. For a successful comparison of characteristics and operating data, accurate measurement of the input pulsed power to the magnetron is required. It was found possible to obtain measurements to a few per-cent and this section describes the methods used by the author to ensure that a testing unit employed for the development of 3 cm. valves yielded reliable and accurate results. The disagreements, at times amounting to 10% between different testing establishments stressed the necessity for this work.

The problem is to measure to a few percent the

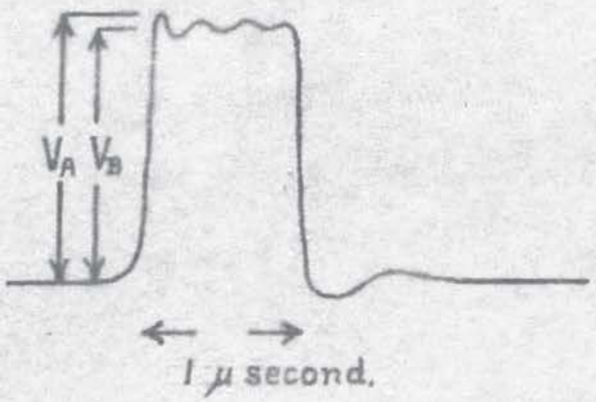


FIGURE 15.

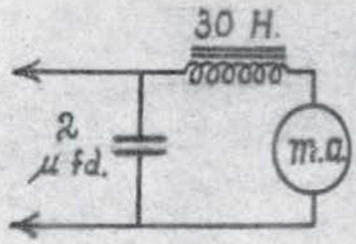


FIGURE 16.

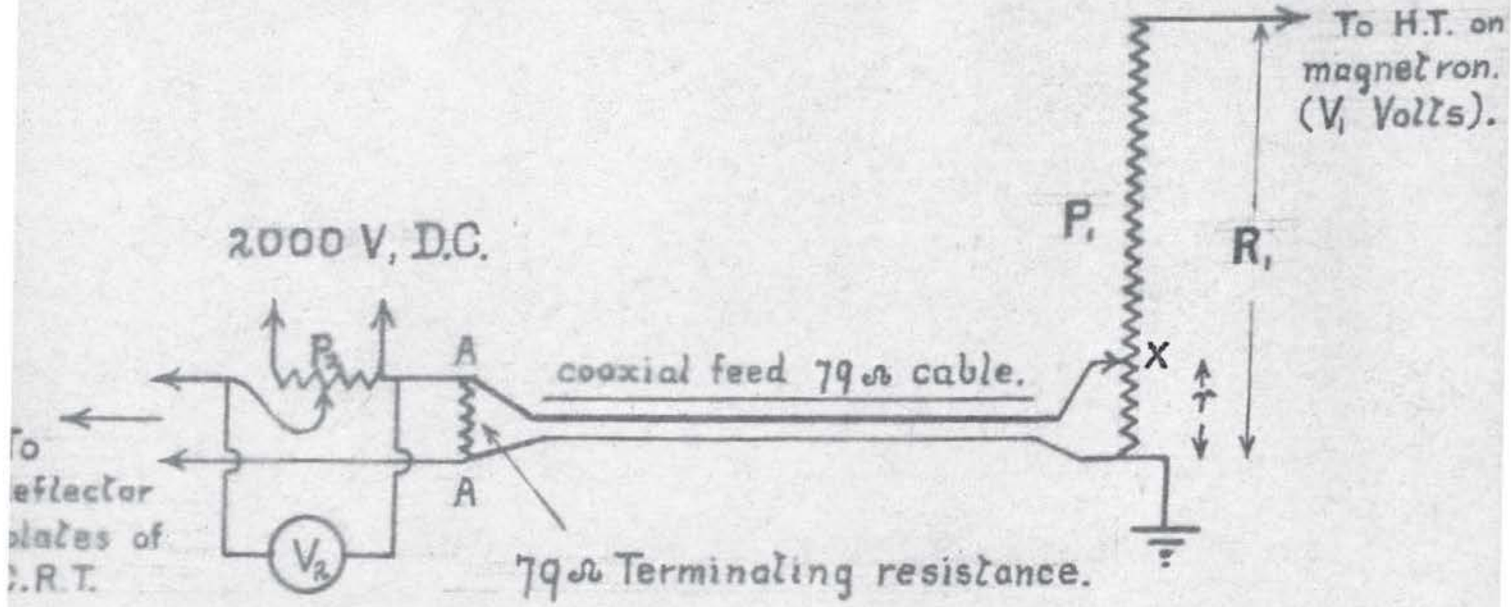


FIGURE 17.

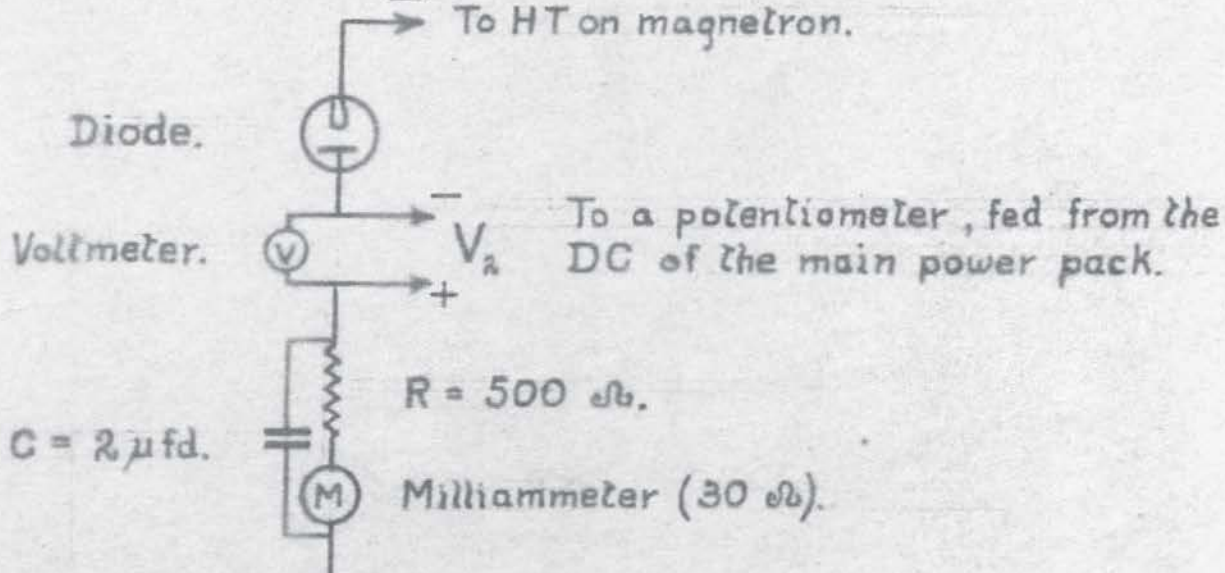


FIGURE 18.

supply to a magnetron operating at about 15,000 V., 20 A. ; this power being supplied in approximately square pulses of 1μ second duration and having a repetition rate of 500 per second. Both the average power and the peak power are required.

A typical shape of the voltage pulse, given on a cathode ray tube, is shown in figure 15. The current could also be registered on the same tube but was more generally recorded on a moving coil meter, which registered the mean value. Since this mean current was about 1/2000 of the peak current, a smoothing circuit consisting of a condenser and an inductance was incorporated to protect the meter (see figure 16). It should be observed that because the voltage pulse is essentially square, the average wattage is the product of the pulse voltage and the average current, but if the R.M.S. current should be used it must be multiplied by the R.M.S. voltage in computing the power.

A number of difficulties are encountered in these measurements; for example, if we attempt a measurement by a diode peak voltmeter, it may record the voltage V_A of figure 15 whereas some voltage such as V_B would be more nearly correct. Errors may also occur due to stray reactances, and to resistances having noticable reactive properties at the frequencies involved.

Voltages read by the cathode ray tube have the great advantage that their pulse shape can be observed, an important asset when examining magnetrons for pecularities in their operation, as, for example, mode jumps. Because the voltage pulse was substantially flat a comparison between measurements of voltage with the cathode ray tube, and measurements of voltage with a diode voltmeter was the object of the first experiments.

4.2. Measurement of voltages by a cathode ray tube.

The circuit used is shown in figure 17. A low reactance potentiometer P_1 of resistance R_1 is connected between the magnetron high tension voltage V_1 and earth, and a portion of this voltage is taken by the tapping point X and fed to the cathode ray tube via a properly terminated 79 Ω coaxial cable. The deflected trace on the tube is then returned to its zero position by a potentiometer P_2. The voltage indicated by the voltmeter V_2 then equals but opposes the voltage at AA due to the pulse. If V_1 is the magnetron pulse voltage then the voltage V_2 at AA is:-

$$V_2 = V_1 \frac{79 . r}{79 + r} \left(\frac{1}{R_1 - \frac{r^2}{79+r}} \right) .$$

and hence

$$V_1 = V_2 \frac{79 + r}{79 r} \left(R_1 - \frac{r^2}{79 + r} \right) .$$

R_1 could be varied, and values between 700 Ω and 1500 Ω were used; r_1 could have values between 30 Ω and 50 Ω

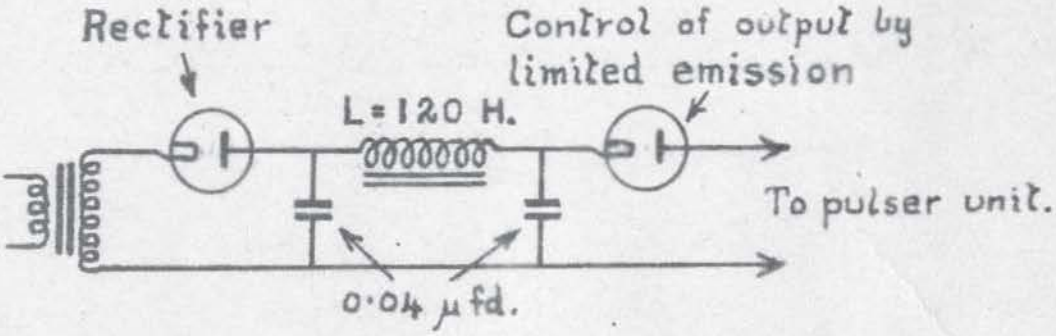


FIGURE 19.

Kilo volts by Cathode Ray Tube	Kilo volts by Peak Volt-meter.	The difference expressed as a percentage
10·0	10·2	2·0%
10·0	10·1	1·0%
11·2	11·4	1·8%
11·3	11·4	0·9%
14·7	14·6	−0·7%

FIGURE 20.

Kilo volts by Cathode Ray Tube.	Average current m.a.	Watts from (1) and (2).	Water flow cc/sec for 30 divisions rise.	Watts from water flow i.e. (4) x 14·0.	Ratio of (5)/(3).
9·03	4·25	38·4	2·77	38·8	1·01
9·26	4·20	38·9	2·69	37·7	0·97
9·63	8·62	83·0	5·77	80·7	0·974
9·67	8·60	83·2	5·80	81·2	0·976

FIGURE 21.

4.3. Measurement of voltage by a diode voltmeter.

The circuit is shown in figure 18. When V_2 is less than the high tension voltage supplied to the magnetron, pulses of current are sent through the diode. R and C provide a smoothing circuit to protect the milliammeter and it is shown later that the error introduced by their presence is negligible. The voltage V_2 was taken from the same magnetron power unit and could exceed the magnetron voltage by virtue of the controlling diode between the power pack and magnetron figure 19. If the power unit has an appreciable ripple the correct value of V_2 is that at the instant of the pulse. This ripple would be caused by discharges into the pulser line after every discharge through the magnetron. Each charge for 15,000 volts is equivalent to 20 A. for a μ second, that is 20.10^{-6} coulombs, and a consideration of the power pack components indicate that the ripple volts should then not exceed $\pm$ 150 V. Measurements by an auxiliary cathode ray tube suggested that the ripple was rather less than this. Hence the value for V_2, as given by a voltmeter (which indicates the mean voltage), should be to within 1% of the true value.

Another possible error may occur through the milliammeter having a finite sensitivity which results in the possibilities of voltages developed across R, and the diode, at the apparent point of balance. A simple calculation, however, shows that, with the smallest detectable current being about $\frac{1}{40}$ mA., the combined peak voltages across R and the milliammeter is well under a volt and therefore insignificant compared with the voltages in question. The emission from the diode was checked. For $\frac{1}{20}$ A. emission, which corresponds to $\frac{1}{40}$ mA. mean current, the voltage drop across the diode was considerably less than 100 V. The error is of the same sign as that introduced by the power pack ripple but even then the combined error is less than 1.5%. Transit time effects in the diode are of no account for the periods and the voltages involved.

The results of measurements taken simultaneously from the cathode ray tube and from the peak voltmeter are given in figure 20, and the agreement may be considered satisfactory. In routine measurements the cathode ray tube method was employed, the pulse voltage thus obtained being multiplied by the mean current to yield a value for the mean input power.

The input power can also be found from the temperature rise of the cooling water, provided that no high frequency power is radiated by the magnetron. This is a valuable means of showing that the routine method suffers from no unsuspected errors. Measurements by these two methods will now be described.

4.4. Calorimetric measurements of power.

Lagging was placed round the magnetron and a thermocouple unit placed on the input and the output water leads to indicate the rise in temperature. During the experiment this rise was kept constant by varying the water flow. The rate of flow would then be proportional to the heat removed by the water. The loss of heat by radiation and conduction from the lagged magnetron is constant and could be determined but owing to the temperature rise being less than $3\frac{1}{2}^{\circ}C$ this correction may be safely omitted. For example, the radiation loss from any part of the magnetron which happened to be exposed would not exceed the black body radiation loss of $2.4\ 10^{-3}$W. per square centimeter, and the conduction loss through the cotton wool lagging is less than 0.02 W. over the entire surface. That these losses were negligible was shown by the fact that when about half the lagging was removed no detectable change in the water temperature was observed. To avoid loss by radio frequency radiation, the output feed of the magnetron was wrapped with metal foil in good contact with the magnetron output stem. Precautions against losses from the cathode leads are given, in some detail, later.

The calibration of the calorimeter system by the cathode heater watts.

The cathode heater, rated at 5 to 6 volts had its input wattage measured by an ammeter and a voltmeter which were probably correct to 1% and certainly correct to 2%. At 5.95 V. the current was 2.80 A. corresponding to 16.7 W. The standard reading of the thermocouple meter recording the water heating was decided upon as 30, this corresponding to a deflection large enough for accurate re-setting and being equivalent to 3.3°C. rise in temperature. For the 16.7 W. input 1.19 c.c. of water per second were required. Hence the calibration of the system is 16.7 ÷ 1.19 or 14.0 W. input for each c.c. of water per second.

4.4.(continued) The determination of the input high tension power.

Before measurements could be made with certainty it was necessary to investigate whether any appreciable power is radiated down the cathode leads. It is impossible to surround these with foil as was done with the magnetron output feed, because these leads were at high voltage with respect to the anode of the magnetron. With 15,000 V. on the magnetron and 8 mA. mean current input, a small neon lamp indicated some loss of power, probably a watt or so, from the cathode leads. By lowering the voltage to 12,000 V. this loss was appreciably reduced being probably ½ a watt or so, while at 10,000 V. none whatsoever was detected. Some idea of the sensitivity of the neon lamp was obtained by a separate experiment which indicated that a radio frequency output of $\frac{1}{2}$ a watt mean power from a magnetron was ample to light the lamp. It is therefore safe to say that at 10,000 V. the loss of power by radiation would be less than $\frac{1}{2}$ a watt and since the mean input power varied from 37 W. to 81 W. the error involved would be less than 1.4% for the lower input wattage and less than 0.6% for the higher input wattage.

Results.

The high tension voltage and the average current were measured for four different powers, the water flow in each case being adjusted to give the standard deflection on the thermocouple meter. The results are given in figure 21 and indicate an agreement between the two methods which is well within the experimental error.

Conclusion.

Measurements of power and of voltage have been made using methods which differed very considerably in technique. The agreement obtained suggests that the calibration of the pulsing unit employed should not differ by more than 2% from the correct value.

SECTION 5.

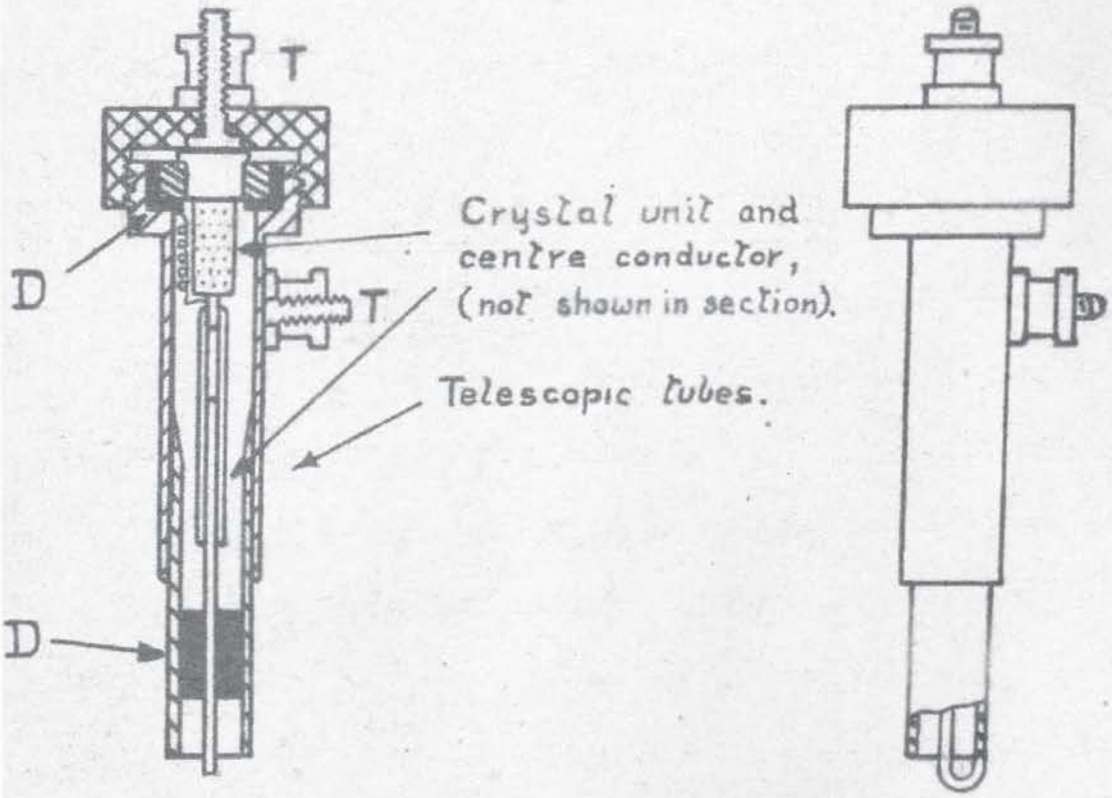
A GENERAL SURVEY IN PREPARATION FOR SECTIONS 6 and 7.

SUMMARY.

This section, while giving the necessity for the measurement of electric fields, has, as its main purpose, the explanation of a geometrical solution of electrical impedance problems. The results from the second of the two examples which conclude the section, are required in section 6.

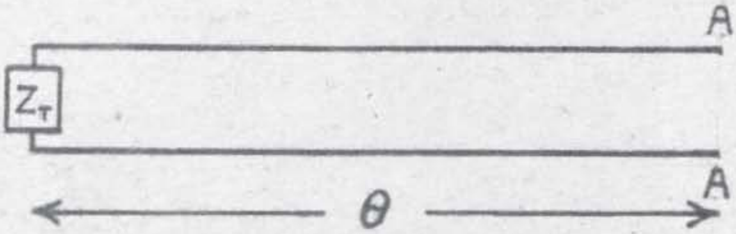
5.1. The necessity for good test probes and loops.

The chief properties of electric resonators have been described in sections 1 and 2. These properties for resonators of simple geometric shapes, e.g. the sphere, can be calculated; but for resonators of shapes most often met with in practical design, e.g. the multi-resonator magnetron, such properties are not amenable to calculation. For this reason they must be found by experimental determination and for this work, small field exploring units are required. These units are also vital for examining the standing wave ratio in transmission systems. It is of first order importance that the size



D. Distrene insulation.
T Output terminals.

FIGURE 22.



$\theta = \frac{2\pi l}{\lambda}$ where l is the length of the line in cm. if λ is in cm.

FIGURE 23.

of these test units and the power required to operate them, cause no appreciable distortion of the fields under observation. For example the resonance of the wavemeter shown in figure 6 must not be noticeably affected by the sizes of the input and the output units - in this case loops.

These test units consist either of a small probe or a small loop. Figure 22 shows one of each, together with a matching section of the author's design. That an efficient matching section is at times necessary was shown clearly by the author when first attempting accurate standing wave measurements on a feeder from a 3 cm. oscillator of some 100 mW. power. Before more is said about these units it is desirable to explain a geometrical construction for the solution of impedance problems. The proof can be found in reference (50).

5.2. A Geometrical construction (circle diagram).

The essentials of the problem are:- Given a length of transmission cable of characteristic impedance Z_0, and terminated by an impedance Z_T, what is (1) the impedance, Z_A, at AA figure 23 and what is (2) the standing wave ratio on the cable. For most problems it is adequate to assume the line is lossless, in which case the solution is to be found in the formula:-

$$Z_A = Z_0 \frac{Z_T \cos\theta + i Z_0 \sin\theta}{Z_0 \cos\theta + i Z_T \sin\theta}.$$

If $\theta = \pi$ or $\pi/2$ (i.e. a $\lambda/2$ or $\lambda/4$ section of cable) the formula simplifies to give $Z_A = Z_T$ and $Z_A = Z_o^2/Z_T$ respectively. Unfortunately to determine several values of Z_A between these points, and even one value of the standing wave ratio, (when Z_T is complex) requires no small amount of work. If it were possible to construct a graph of the expression, we could obtain at a glance the values of Z_A for all values of Z_T and of θ. This is in fact what the circle diagram does. We shall further see that for any one value of Z_T only two points, (the simple ones already found will do) are required to determine not only the full range of Z_A but also the standing wave ratio on the cable. With very little more trouble the values (as distinct from the range) of Z_A for any value of θ can be read from the graph. In this manner the entire solution to the problem can be obtained.

The graph is constructed as follows. Two axes are taken, one resistive and the other reactive as indicated in figure 24. All values of impedances will lie to the right of the reactive axis, where the resistive element is positive. Moreover, as θ is varied, all values of Z_A for any one value of Z_T will lie on a circle having its centre on OX. This circle can immediately be drawn because we know two values of Z_A. An example will make this clear. Suppose, as before, a line of variable length is terminated

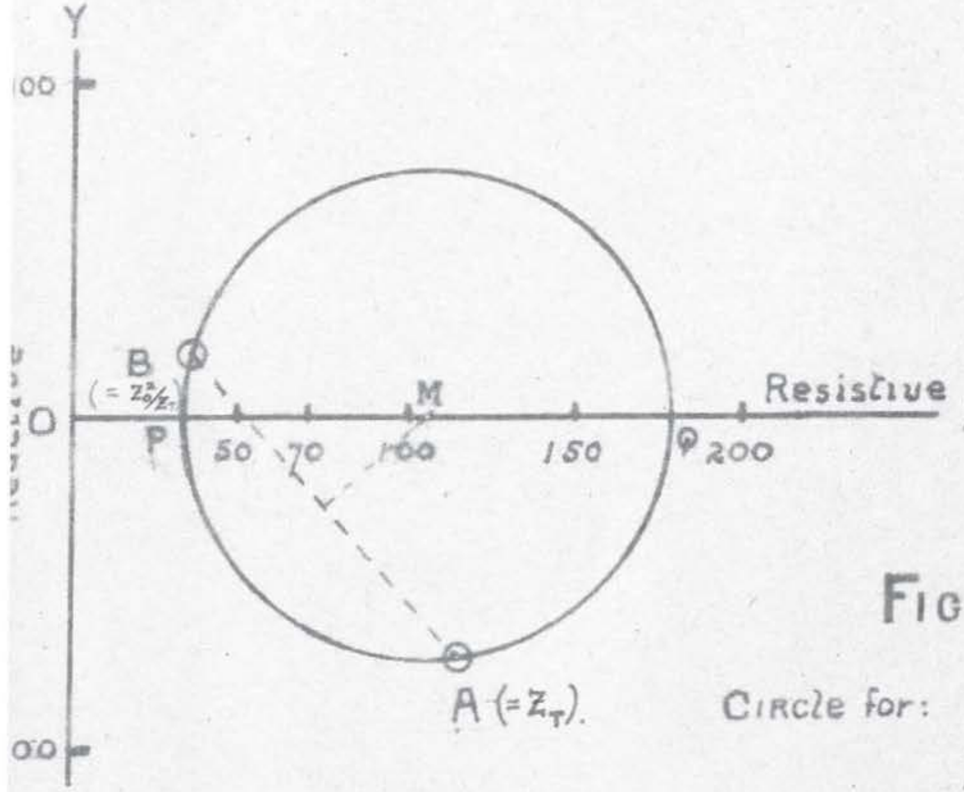


FIGURE 24.

Circle for: $Z_o = 70$ ohms.
$Z_T = 117 - i\,73$ ohms

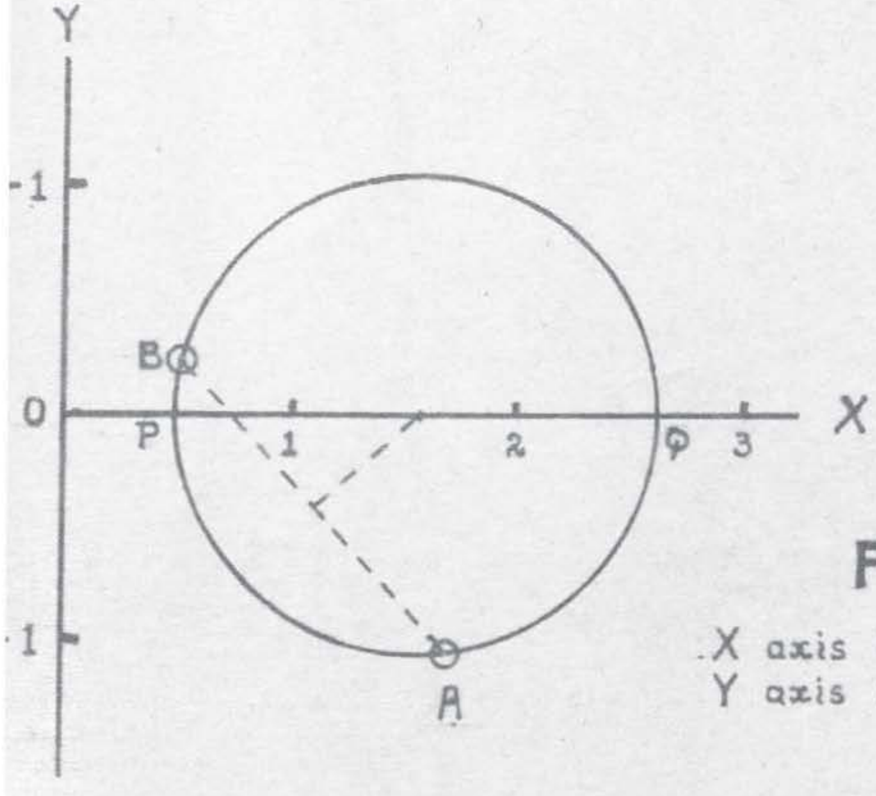


FIGURE 25.

X axis is RESISTANCE $\div Z_o$.
Y axis is REACTANCE $\div Z_o$.

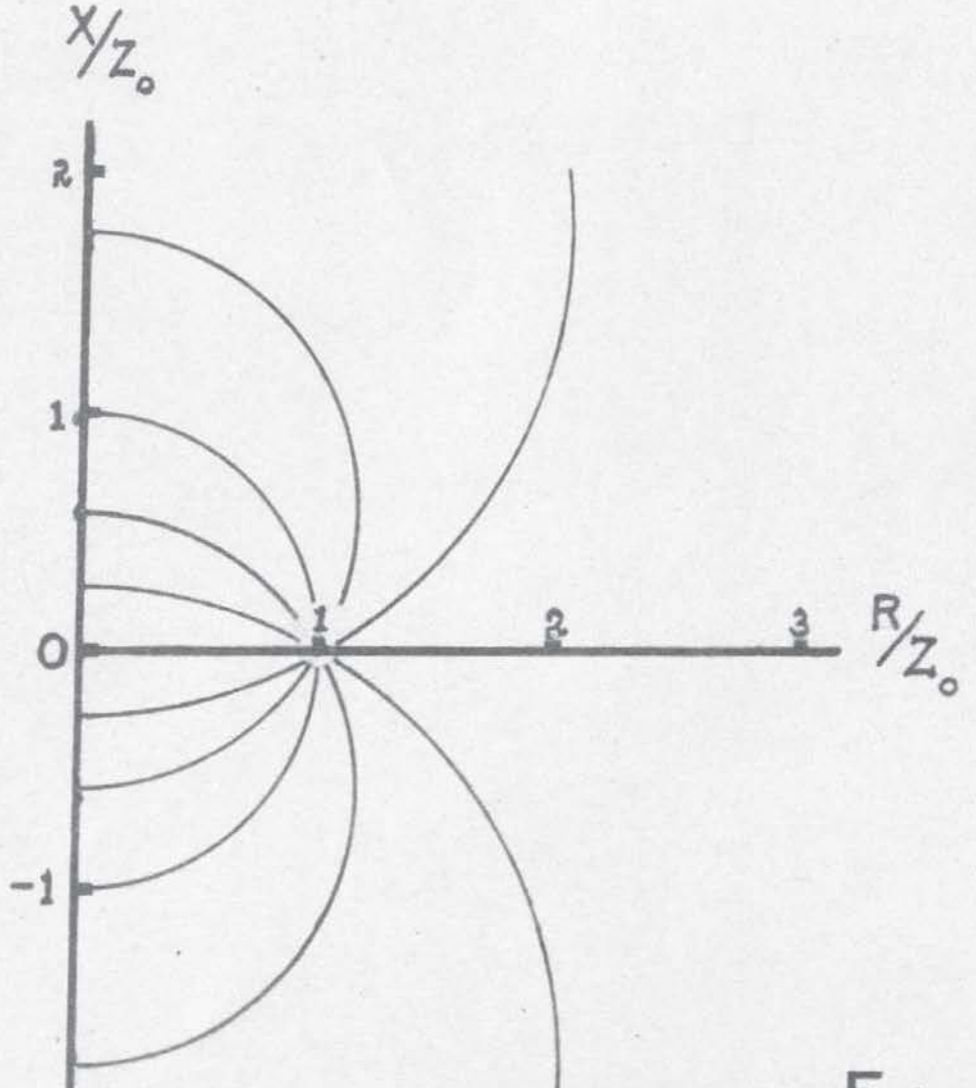


FIGURE 26.

A set of $\underline{v}$ circles 15° apart from each other. NOTE. One v circle coincides with the R/Z_0 axis.

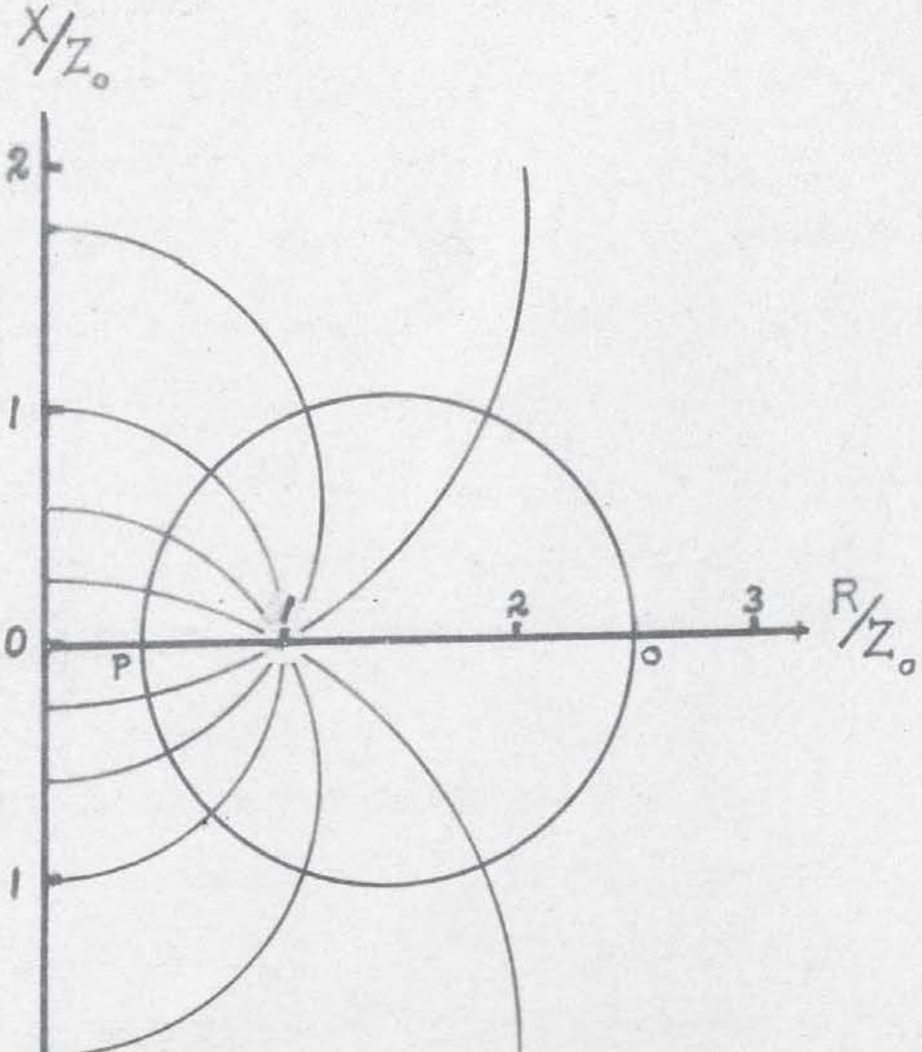


FIGURE 27.

by Z_T (complex). Plot the two points Z_T and Z_o^2/Z_T as shown by A and B of figure 24. Since the circle has its centre on OX it can be drawn by using the self explanatory construction of the figure; M is the centre. This circle gives all possible values of Z_A, and further the voltage standing wave ratio, which is fixed for any one circle, is given by $\sqrt{\frac{OQ}{OP}}$. There is only one more relation to be found, namely how Z_A varies round the circle with θ . This is given by another set of circles which have their centres on the reactance axis. These circles are called v circles to avoid confusion with the first set of circles which are called u circles. In practice it is not in general necessary to construct these v circles because, by a small change in figure 24, this second set of circles becomes fixed for all problems. They then have universal application. For this to be true we must plot not (as has just been done for simplicity) the actual resistances and reactances, but their ratios with respect to Z_o. Hence our axes become as in figure 25 and points A and B instead of being Z_T and Z_o^2/Z_T become Z_T/Z_o and Z_o/Z_T. Figure 26 gives the universal set of circles. When these are superimposed on the u circle, as in figure 27, the variation of Z_A, in terms of the cable length θ, is obtained. The v circles are drawn for values of θ differing by 15°, i.e. a motion along a u circle from one v circle to another corresponds to 15 electrical degrees of cable. If θ is measured from the point where the termination Z_T is across the end of the cable, an increase of θ causes the u circles to be traversed clockwise. It may be observed that any v circle cuts a u circle at two points 90° different in θ. The voltage standing wave ratio, for all points on the circle, is given as before by $\sqrt{\frac{OQ}{OP}}$.

The construction just given is usually referred to as the circle diagram solution.

It is important to observe the following extension of the circle diagram to admittances:

The equation used for impedances (see the beginning of 5.2.) was,

$$Z_A = Z_o\left(\frac{Z_T \cos\theta + i\, Z_o \sin\theta}{Z_o \cos\theta + i\, Z_T \sin\theta}\right).$$

This equation can be rewritten as:

$$\frac{1}{Z_A} = \frac{1}{Z_o}\left(\frac{Z_o \cos\theta + i\, Z_T \sin\theta}{Z_T \cos\theta + i\, Z_o \sin\theta}\right).$$

By writing $A_A = \frac{1}{Z_A}$; $A_T = \frac{1}{Z_T}$; $A_o = \frac{1}{Z_o}$ and multiplying each side of the equation by $A_T A_o$ the following equation in admittances is obtained:-

$$A_A = A_o\left(\frac{A_T \cos\theta + i A_o \sin\theta}{A_o \cos\theta + i A_T \sin\theta}\right).$$

This equation is exactly the same as that for impedances and hence the circle diagram solution can be applied to

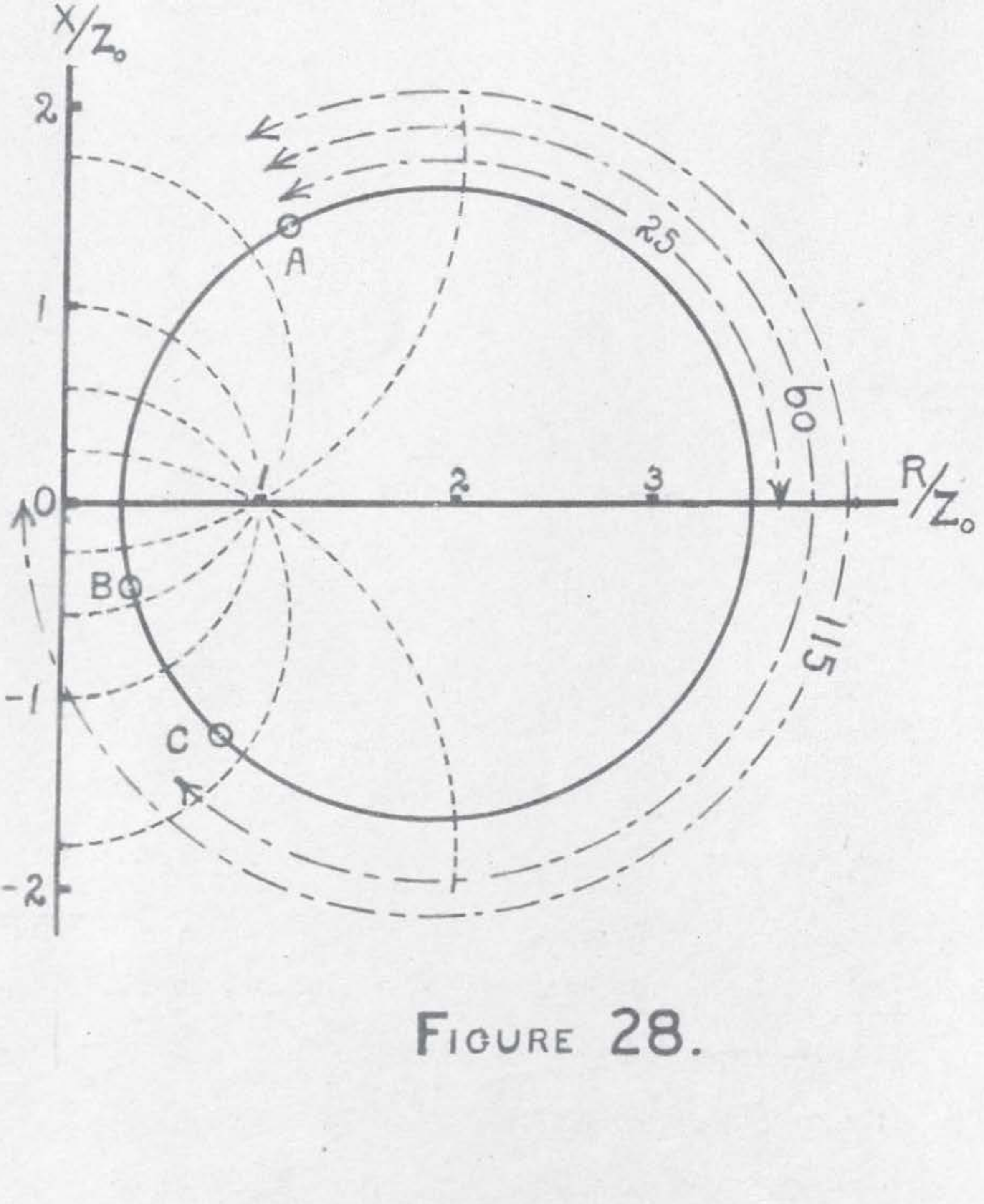


FIGURE 28.

admittances as well as to impedances. The advantage of this will be made clear in example 2.

Example 1.

Suppose we require the impedance at the terminals of a line 960° long (i.e. 5 x 180° + 60°) of characteristic impedance $Z_0 = 70$ ohms, terminated by an impedance Z_T of (80 + 100i) ohms.

Solution.

The two known values of Z_A expressed as a ratio of Z_0 are:-

$$Z_T/Z_0 \quad \text{and} \quad Z_0/Z_T .$$

$$\text{i.e.} \quad \frac{80+100i}{70} \quad \text{and} \quad \frac{70}{80+100i} .$$

or (1.143 + 1.43i) and (0.341 - 0.426i). By the means of these two points the circle can be constructed figure 28, points A and B. With application of the v circles it is clear that a 60° change in cable length from A (the point of termination), gives the value of indicated by C. Hence Z_A = (34.4-31i) ohms. Furthermore a glance at the diagram gives the following information:-

(1) The standing wave ratio is 3.6 : 1 in volts.

(2) As θ is varied the resistive value of Z_A lies between 20 ohms and 250 ohms, and Z_A is purely resistive at these extremes for which the cable has been increased by approximately 25° and 115° respectively over the original length.

(3) The reactance falls to zero twice for every 180° range of θ , but is always within the range of

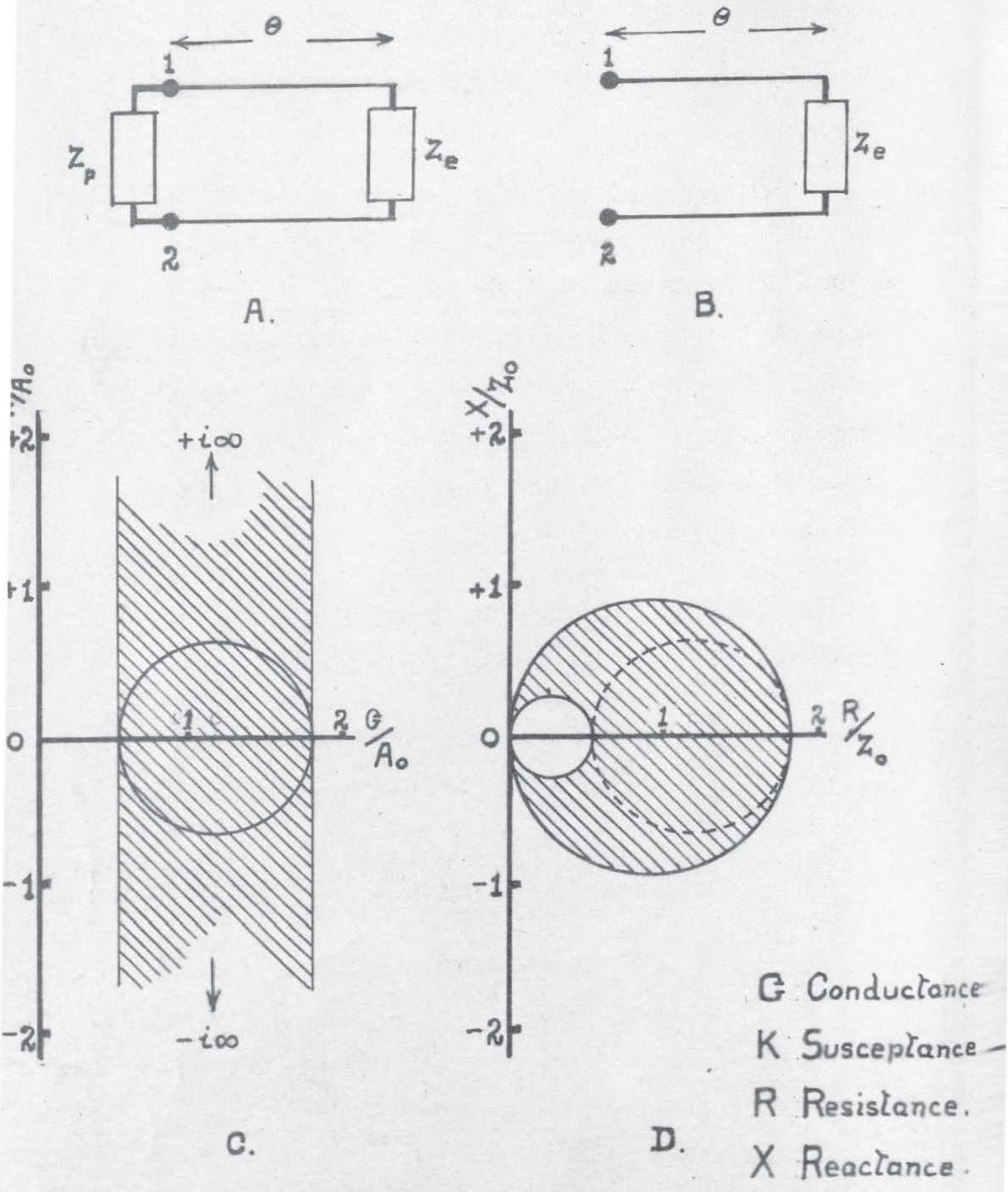

θ
1
2
Zp
Ze
A.
B.
+i∞
-i∞
+2
+1
0
-1
-2
1
2
G/A0
X/Z0
R/Z0
C.
D.
G Conductance
K Susceptance
R Resistance.
X Reactance.

FIGURE 29.

+115.51 and -115.51 ohms.

(4) can never be purely reactive.

Example 2.

In section 6, a circuit, such as is shown in figure 29A, had to be considered. Z_e is complex and assumed to be constant. We wished to examine the values of the impedances at points 1, 2, when the pure reactance Z_p and the length of line θ are varied independently. We also wished to compare these values with those obtained when the circuit to the left of points 1,2, is removed.

The solution is best obtained by working with admittances instead of impedances. Let:-

$A_{12} = \frac{1}{Z_{12}}$ where Z_{12} is the impedance across points 1,2.

$A_o = \frac{1}{Z_o}$ where Z_o is the characteristic impedances of the line.

$A_c = \frac{1}{Z_c}$ where Z_c is the impedance at points 1,2, when Z_p is removed, i.e. the impedance of the circuit shown in figure 29B.

$A_p = \frac{1}{Z_p}$ and varies from $-i\infty$ to $+i\infty$ in value.

Since A_c is the admittance looking to the right of 1,2; then as θ varies all values of A_c lie on a circle. Let this be the circle of figure 29C.

Further:

$A_{12} = A_p + A_c$. But since A_p varies from $-i\infty$ to $+i\infty$ (and independently of A_c) it follows that the range of A_p and A_c together is given by the shaded strip in figure 29C. This gives the range of values of A_{12}.

To obtain the answer in terms of impedances we require the reciprocal of A_{12}; i.e. we must invert all points on the shaded strip about the point O, the inversion constant being +1. This gives the shaded part of figure 29D.

The dotted circle of figure 29D gives the locus of Z_c, i.e. the range of impedances when the circuit to the left of 1,2, is removed.

SECTION 6.

THE MEASUREMENT OF ELECTROMAGNETIC FIELDS BY PROBES AND BY LOOPS.

Summary.

The general properties of probes and of loops for electromagnetic field measurements are given. Detailed information on matching sections, crystal rectifiers, and valve amplifiers, all of which have been developed for the efficient operation of probe systems, is also included.

6.1. Small probes and loops.

In high accuracy measurements of electromagnetic fields by test probes or loops there are two conflicting requirements. The first is that no appreciable distortion should occur in the field which is being measured, and the second that the measuring unit should give sufficient output to operate the recording instrument, which may be a microammeter or a cathode ray tube. The power required to operate a microammeter is about 10^{-8} W. This is therefore very small compared with the high frequency power available, which, in the case of a 3 cm. wavelength continuous wave oscillator, is of the order of 100 mW. It follows from this that distortions of the field by the measuring unit are due either to unsound mechanical construction or to

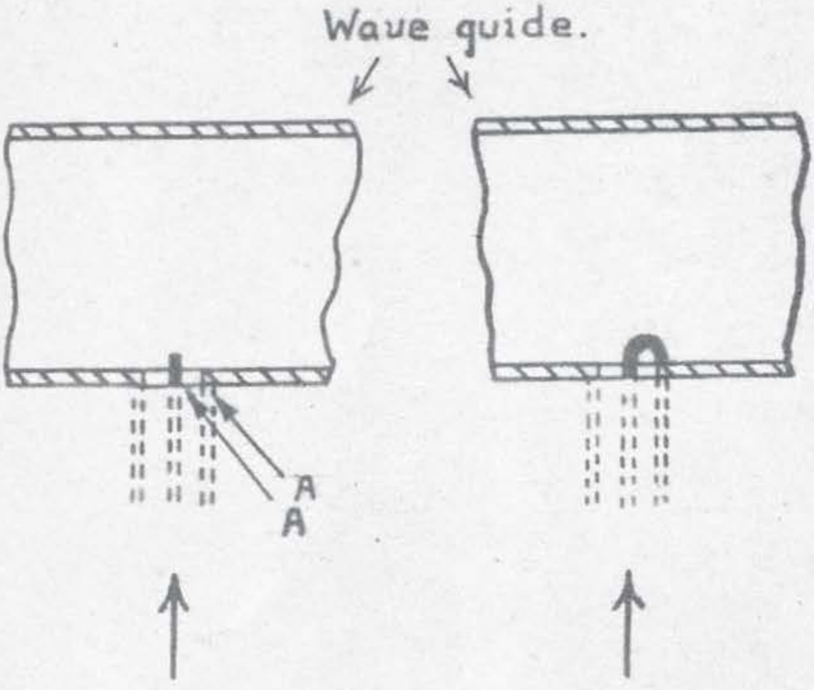
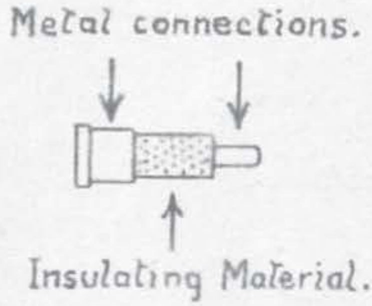
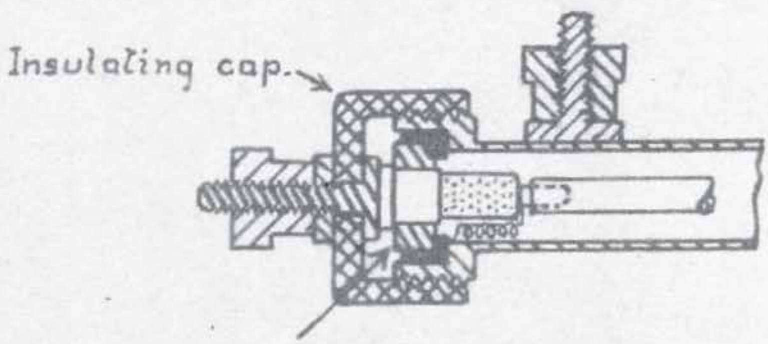


FIGURE 30 A.   FIGURE 30 B.

FIGURE 31.



CRYSTAL UNIT.



Brass ring holding the crystal unit in a distrene insulator, shown black. The crystal unit and the centre conductor are not shown in section.

CRYSTAL UNIT MOUNTED.

FIGURE 32.

poor electrical design. The author's work showed that this distortion of the field by probes and loops was greater than was generally recognised. He found it necessary to give special attention to this problem, (see also section 7).

The equivalent circuit of probes and loops.

(i) The probe. This is in essence a short length of wire inserted into the guide as shown in figure 30A. At AA a coaxial cable is attached to connect (via a matching section) to the detector unit. The characteristics of the probe seen at AA, i.e. looking in the direction of the arrow, are, by Thévenin's theorem, those of a high impedance generator.

(ii) The loop. This is a small loop of wire inserted into the guide to link with the magnetic lines of force. An example is given in figure 30B. Again by Thévenin's theorem it is evident that a small loop has the properties of a low impedance generator.

In practical applications either a loop or a probe can be used: the former receives its e.m.f. by linkage with the lines of magnetic force and the latter by its position along the lines of electric force. The choice between loop or probe sometimes rests with the disposition of the field to be measured, e.g. for the rhumbatron (figure 4 of section 1) a loop is preferable, while for a coaxial feeder either may be used. Where possible the author has always preferred a probe to a loop. The reason is that while loops may mislead the experimenter by being affected by an electric field in the direction of A, figure 31, and thereby behaving in a manner similar to that of probes, yet it does not seem possible that probes can have properties similar to those of loops. It is for this reason and also to avoid the repeated use of the term 'probe or loop' that probes only are mentioned in this work. This does not mean that the results are restricted to probes, for the impedance of a probe becomes indistinguishable from that of a loop which has the addition of a section of line approximately $\lambda/4$ in length. In practice this section is supplied by the elasticity of the matching section.

Rectifier units.

The radio frequency power received by the probe has to be rectified before it can be measured. Different types of rectifiers are discussed in the subsection 6.3, but throughout most of the work crystal rectifiers will be assumed as they are most generally used. Mention of the mounting of these rectifiers is made here because of the discussion on matching sections in the subsection 6.2. Figure 32 gives the most usual mounting. The high frequency choke, placed between the concentric conductors, is to allow the rectified current to have a circuit through the galvanometer. Such chokes are about 1 m.m. in diameter, 5 m.m. long and have approximately 10 turns of 40 s.w.g. copper wire: the number of turns is not critical. With some designs of test unit the choke is not necessary; e.g., if a loop is used instead of a probe the crystal current can complete its path by flowing through the loop.

6.2. Matching sections.

Matching sections are necessary to provide for the efficient transfer of power from the probe to the rectifier. The simplest design of probe would be with no matching section, figure 33A. This is not suitable for 3 cm. wavelength work at powers as low as 100 mW., unless by a fortuitous choice the dimensions happen to be correct for the wavelength used. Good results can however be obtained if the coaxial line is made variable in length. This gives a transformation of impedance as has already been discussed in section 5. It is evident from the same section that the matching range is restricted to impedance values lying on one circle (and not on the whole) of the impedance plane. Despite this restriction the performance, in practice, is good.

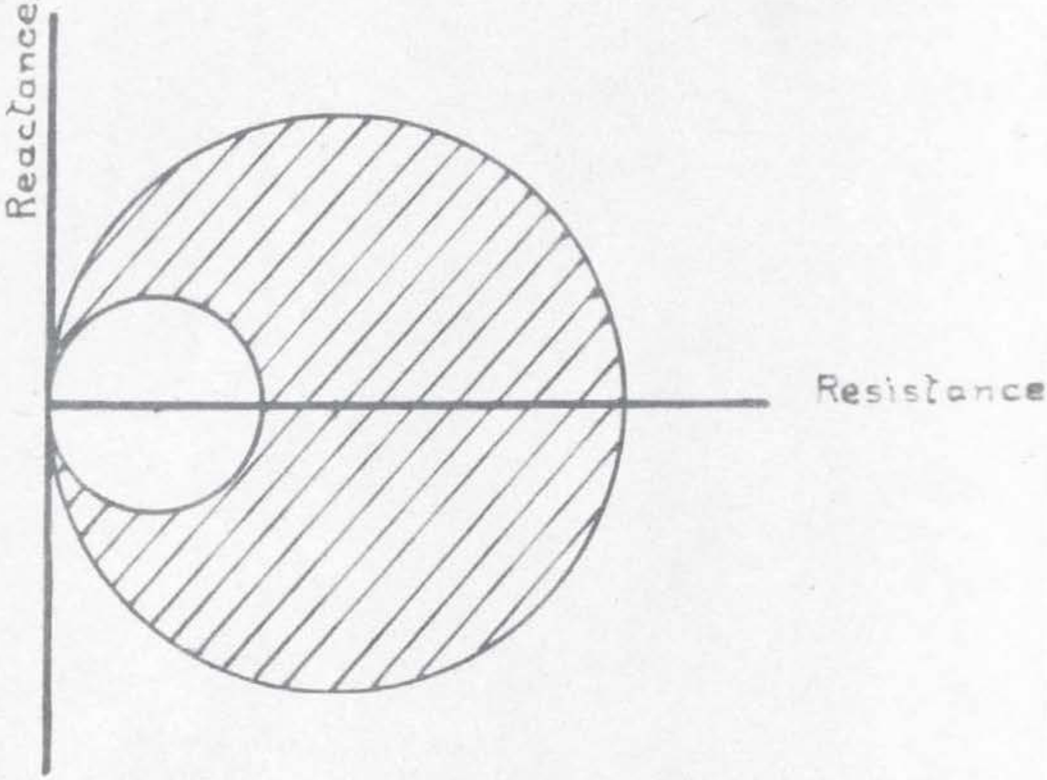


FIGURE 35.

A diagram indicating the range of the impedance presented by the crystal and piston to the coaxial (1) of figure 34. See also figure 29.

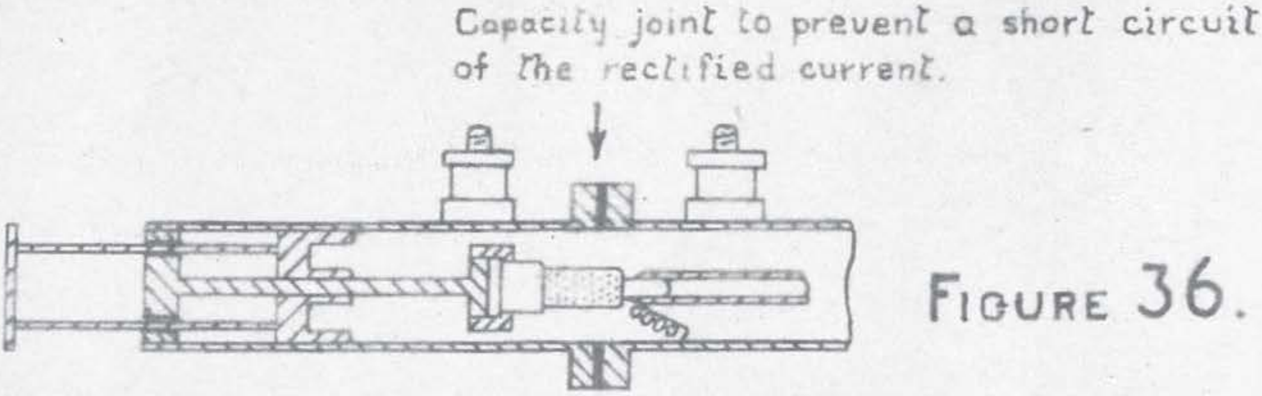


FIGURE 36.

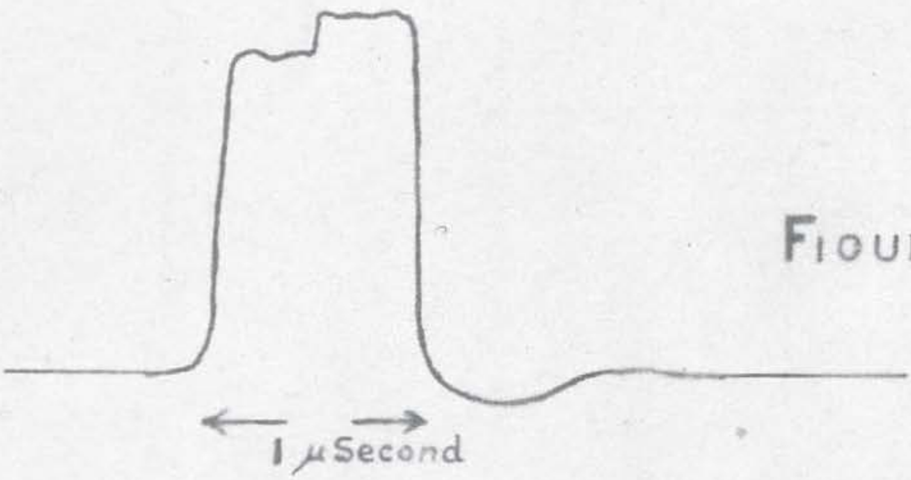


FIGURE 37.

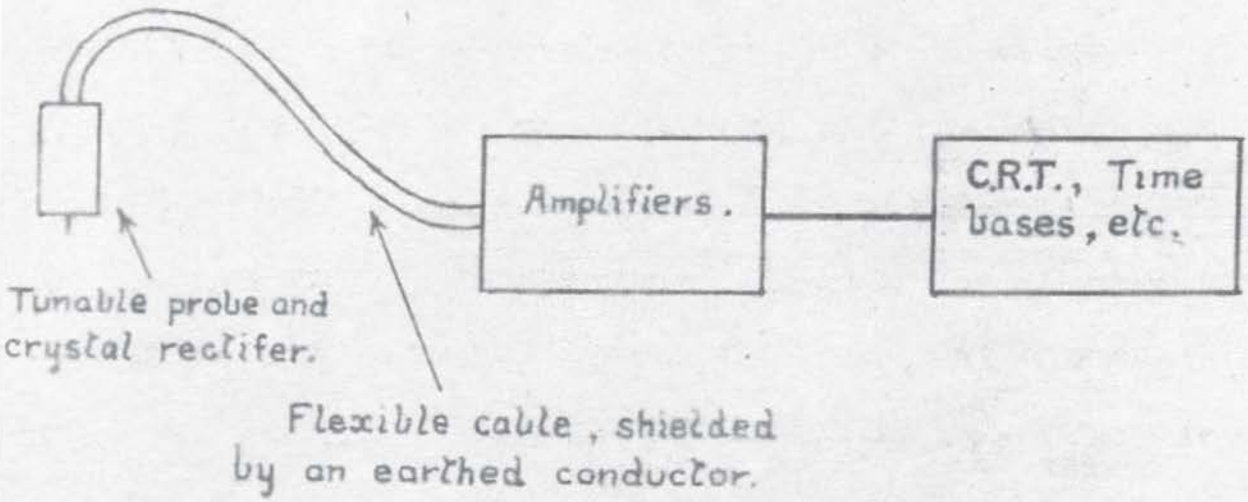


FIGURE 38.

To increase further the matching range a variable reactance was placed in parallel with the crystal as in figure 34 and each of the two coaxials was made telescopic. The coaxial marked number 1 now sees (instead of the impedance Z_e of the crystal) the impedance presented by the crystal in conjunction with the parallel reactance. This impedance, (see example 2 of section 5), can cover the shaded area of figure 35, whence it is evident that the matching range of the whole unit has been increased. Nevertheless this design of probe did not yield a larger crystal current than could the design given in figure 33B. With all designs, the best results seemed to be obtained when the length of the line between the probe and the crystal rectifying point was $(n + \frac{1}{2})\frac{\lambda}{2}$. If a loop was used instead of a probe the length of line was $n\frac{\lambda}{2}$. Some improvement in performances might perhaps be obtained by redesign of the crystal <u>holder</u> , which might for instance be backed by a movable piston, figure 36. The urgency of other work, such as the design of an instrument for the measurement of the standing wave ratios on feeder systems, made it necessary to postpone further experiments on probes until the design of figure 33B was shown to require improvement.

The possibility that the probe may appreciably distort the field in which it is placed must never be excluded. Such distortion is a function of the matching system and it usually increases with increase of signal current from the probe. This is true even if the immersion of the latter is constant. More is said of this aspect of the probe in section 7.

It is now appropriate to discuss some of the rectifiers and recording instruments suitable for use in conjunction with the probe. The most sensitive detector would be the super-heterodyne receiver. This, however, is only used in very special circumstances because far simpler systems, such as a crystal rectifier and galvanometer, are often very satisfactory and are much less trouble to make and to use.

6.3. Rectifiers suitable for the use with probes or with other low power radio frequency systems.

Some form of rectification must be employed before the currents from the probe can be recorded. Partly by necessity, and partly for convenience, the rectifier should be small enough to be incorporated in the body of the probe system: the crystal unit already mentioned is an excellent example of a compact unit.

The different rectifiers usually employed can be divided into three groups:-

(1) Thermionic Diodes.

(2) Vacuum thermocouple elements.

(3) Crystal Units.

These rectifiers will now be considered in turn.

(1) Thermionic diodes.

Small thermionic diodes such as the "D" television diodes can be used as rectifiers for radio frequency power in the wavelength region of 3 cm. A valve amplifier is necessary and some form of modulation must be applied to the radio frequency to enable this amplification to be possible. This method of rectification is however rather insensitive, but it can be used for measurements involving very high power, such as, for example, the power from an efficient multi-resonator magnetron. There is the advantage that the diode is not easily damaged if the radio frequency should unexpectedly increase in intensity to a high value.

(2) Vacuum thermocouples.

Small vacuum thermocouples (about the size of a "pea lamp") when used in conjunction with a moving coil galvanometer, are at times useful because they maintain their characteristics over an almost indefinite period of time. Care not to overload the thermocouple heater must be taken as, even though a burn-out does not occur, the quality of the vacuum can easily be spoilt. The thermal lag of these units makes them rather unsuitable for some operations; for example, the searching for an unknown wavelength with a sharply tuned wave-meter. The same reason makes them useless for determining the envelope of radio frequency pulses.

(3) The crystal rectifier.

Since the war there has been much research on crystal rectifiers and the result has been that they have become one of the most useful of all rectifiers. They are produced with sufficient mechanical robustness to stand a considerable amount of handling and vibration. It would be rash to say that their calibration can be relied upon, but nevertheless the author has found that, with care, crystals remain apparently unchanged for several weeks. A rough and rapid test for the rectifying properties of a crystal is to measure its forward and its backward resistance by applying $1\frac{1}{2}$ V. with the polarity first one way and then reversed. A value of 100 Ω or less should be recorded for one direction and about 1000 Ω for the other direction. An instrument, such as the "AvoMinor" instrument which can be used for the measurement of resistance and which incorporates a single $1\frac{1}{2}$V. cell is therefore admirable for the purpose. On account of the resistance network of these instruments the voltage applied to the crystal is less than $1\frac{1}{2}$ V., and hence the crystal resistances recorded do not correspond to those just mentioned. For the "AvoMinor" instrument the resistance recorded is about 250 Ω for one direction and something over 1000 Ω for the other direction. It should be remembered that during the testing and during the using of crystals, voltages much in excess of $1\frac{1}{2}$ volts (resulting in a current of some 15 mA.) may permanently damage the crystal by producing what is technically known as a "burn-out". The author usually operated crystals below 100 μA. In this region the square law behaviour can be assumed. The instantaneous response of the crystal, its good sensitivity and compact nature make it by far the most suitable rectifier for centimeter wave-length work. Care must be taken not to overload the crystal on pulse work - this is dealt with in the following subsection 6.4.

6.4. Indicating instruments.

Some instrument must be employed to enable the current from the rectifier to be observed. The two instruments most widely used are:-

(1) A galvanometer connected directly to the rectifier.

(2) A cathode ray tube used in conjunction with a valve amplifier.

It is clear from previous remarks that the thermocouple rectifier can be used satisfactory with (1). Its slow response however does not warrant its general use in conjunction with (2). For almost all the work of this Thesis crystal rectifiers were chosen because of their quick response. In consequence of this the merits of (1) and (2) will be discussed with respect to this type of rectifier.

Method 1. The employment of the galvanometer.

This method is the simplest, and very satisfactory results can be obtained when using continuous wave radio frequency. If the oscillator power is of the order of 100mW., accurate probe measurements often require a mirror galvanometer, but for most work a microammeter with a 0-100μA. scale is very suitable. The optimum resistance of the instrument is about 300Ω. For pulse working, either on high or on low powers, this system becomes insufficiently sensitive if the on-off ratio of the pulses is small. This is because during the pulse, (even though the pulse be as short as 10^{-7} sec.) the crystal current cannot be increased much, if at all, over the steady value given by continuous wave radio frequency. A typical condition, in the Services, of pulse operation is with 1μ sec. pulses with a repetition rate of 1000 a second. This has an on-off ratio of $\frac{1}{1000}$ and therefore the crystal current is reduced to $\frac{1}{1000}$ of the value obtained with continuous wave operation. It follows from this that a probe system using a crystal rectifier and galvanometer is not suited for the measurement of pulsed radio frequency power; although for measurements with continuous wave radio frequency power such a probe system is extremely satisfactory. The author's development of this type of probe system for the accurate determination of standing wave ratios in centimetre wave transmission systems is given in the next section.

Method 2. The employment of the cathode ray tube.

For the operation of a probe in conjunction with a cathode ray tube, the output from the crystal rectifier is fed to an valve amplifier, which supplies sufficient signal voltage to operate the cathode ray tube. A gain of about 1000 in the amplifier is required. It must, however, be remembered that because of the properties of amplifiers the method is restricted in use to modulated continuous wave radio frequency systems, or, to pulsed radio frequency systems. The quick response of the method is however of great use for many purposes. Thus it is invaluable when searching with a highly resonant wave meter for an unknown wavelength, or for recording radio frequency pulse shapes, or for observing sudden changes during the pulse (figure 37). Care must be taken when interpreting the picture on the screen to remember the possibility of distortion, e.g. that square pulses will become noticeably rounded if the band width of the amplifier does not pass sufficient of the fourier components which compose the pulse. This method, which uses the crystal rectifier and the cathode ray tube, is almost invariably used for accurate work which involves pulsed radio frequency power. The general arrangement is given in figure 38.

Cathode ray tube with untuned probe.

All the probe systems so far mentioned have used matching sections and are frequency sensitive. If the adjustment of the matching unit is not correct the signal may be overlooked, although by sliding the telescopic sections and watching the trace on the cathode ray tube there is little likelihood of a signal being missed. Nevertheless, if another tuned instrument, e.g. a wave meter, is used with the probe system, it is more difficult to be sure that no signal has been missed. This is because both wave meter and matching section must be in reasonably correct adjustment before the signal is received. In such cases it is clear that a system using an untuned probe is more convenient. The loss in output signal from the probe has to be compensated by increasing, many times, the gain of the valve amplifier. For 3 cm. work this necessitates an amplifier with a gain of at least 15,000. Such an amplifier, with a band width of say 4 Mc/sec. (which is the least necessary for its passing a 1 μ sec. pulse of approximately rectangular shape) requires some 5 or 6 stages of amplification, each stage requiring careful circuit design. If care is taken to shield the probe system from unwanted electrical disturbances good results can be obtained. The extra complexity which this system involves is however hardly justifiable except in exceptional circumstances.

SECTION 7.

THE MEASUREMENT OF STANDING WAVES IN TRANSMISSION SYSTEMS.

Summary.

Three designs of apparatus for the measurement of standing wave ratios in wave guides are given. The first two are simple and easy to construct but do not appear suitable for the measurement of standing waves with a ratio of voltage maximum to minimum smaller than 1.05 : 1. The third is an improved design by the author and measurements on standing wave ratios as close to unity as 1.01 : 1 were possible with this instrument. Full details necessary for satisfactory results are given.

7.1. Important factors which influence the design.

One of the many uses of the probe systems, which have been described in the previous section, is the measurement of the standing wave ratio in wave guide transmission systems. These measurements must, however, be made with caution if serious errors are to be avoided. Thus some designs of standing wave measuring instruments which the author had seen, led him to suspect that the probe (or loop) disturbed the field in the guide sufficiently to affect the accuracy of the results. Preliminary experiments

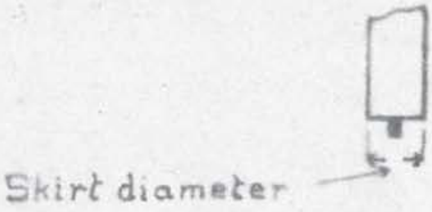


FIGURE 39.

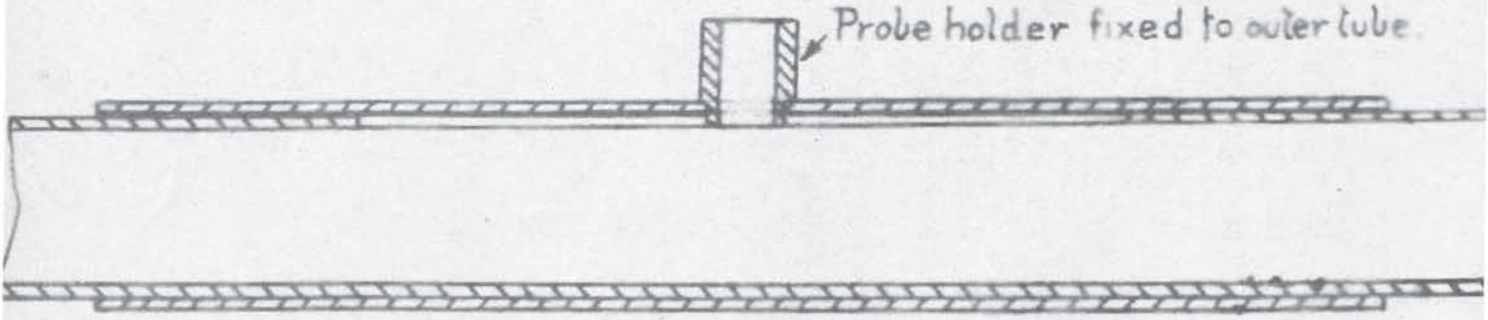


FIGURE 40.

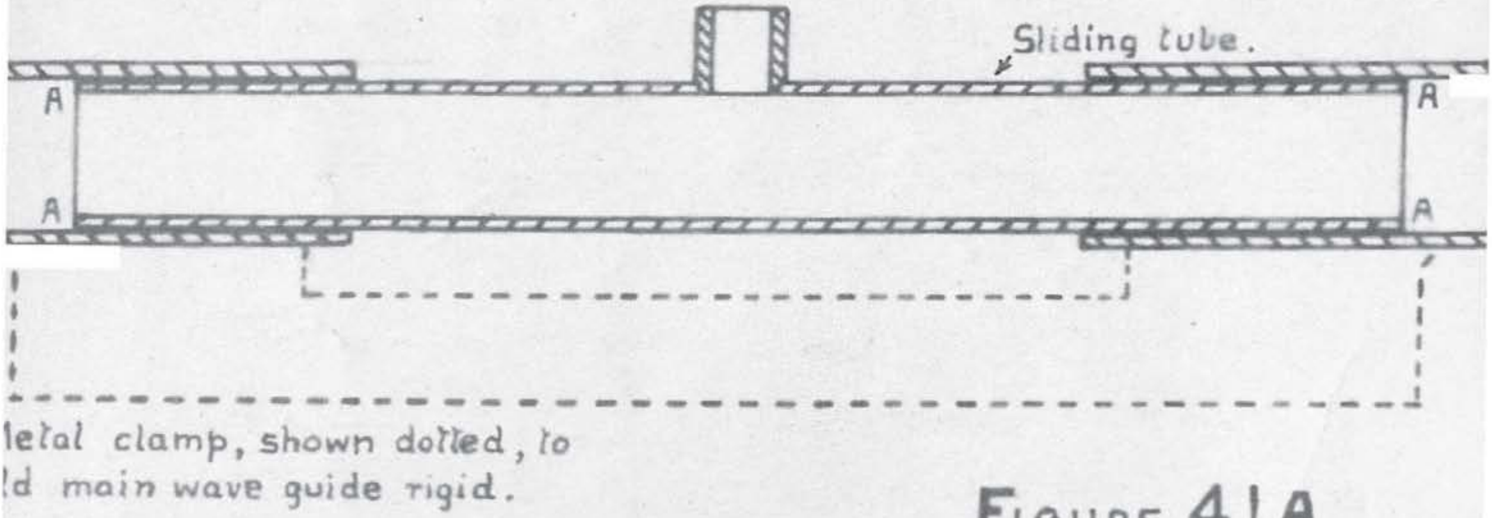


FIGURE 41A.

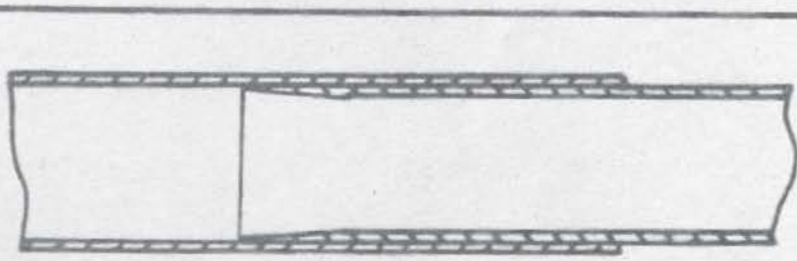
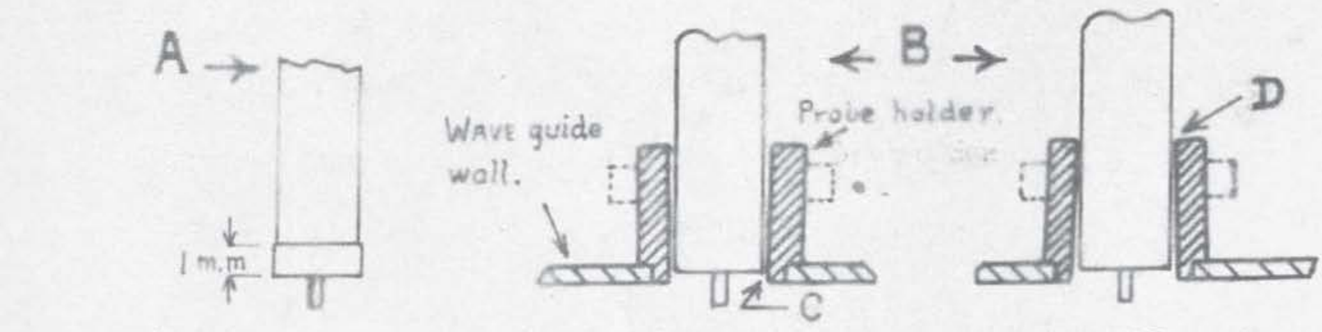


FIGURE 41B.

fully justified this suspicion. It is obvious that by lessening the immersion of the probe, the disturbance the probe causes in the guide can be made negligibly small, but then difficulties arise as only a few microamperes are received by the galvanometer. The maximum allowable immersion for the best performance is found to vary greatly from probe to probe; sometimes the immersion is as much as 2 m.m. while in other cases the probe had to be withdrawn until it was $\frac{1}{2}$ mm. back from the wall of the guide. The immersion distance depends on the diameter of the skirt of the probe, see figure 39, and also on the adjustment of the matching system. Throughout all this section the work described refers to 3 cm. continuous wave radio frequency of some 100 mW. power. The measurement system consisted of a probe, its matching section and a galvanometer. The reasons for this choice have already been mentioned in the previous section.

The probe must be mounted to move along the guide so that it can be made to measure the field at various points. Obviously for all positions of the probe its immersion should be constant and hence the necessity for mechanically sound guide rails. Restrictions on the width, and to an extent the depth, of the slot are necessary to prevent its disturbing the field in the wave guide. The work now about to be described is confined to the H_1 wave as it was around this type of wave that the author's work became concentrated. For example the development of the H_1 wave guide feed from power magnetrons (section 10) led to many problems. Some of these problems were: (1) the design of matching sections for use with transmission systems, (2) high power terminations and high power absorbers for the measurement of the radio frequency power, and (3) the measurement of frequency pulling as a function of the standing wave ratio in the guide. In spite of the restriction to the H_1 wave, the author has had sufficient experience with the E_o wave to state that the general technique for either wave is the same. Should it be that a loop is preferred to a probe a suitable loop would be one made from about 34 s.w.g. copper wire and of about 1 sq.mm. in area.

Three designs of standing wave measuring instruments are given; the first two are comparatively simple to construct but are not suited for the measurement of standing waves with



$\frac{2}{1000}$" increase on skirt diameter.

RIGHT, contact is at C, at the wave guide wall.

WRONG, the contact is at D.

Note. The probe holder has a longitudinal split and a circular clamp, shown dotted, is used to hold the probe unit rigidly.

FIGURE 42, A and B.

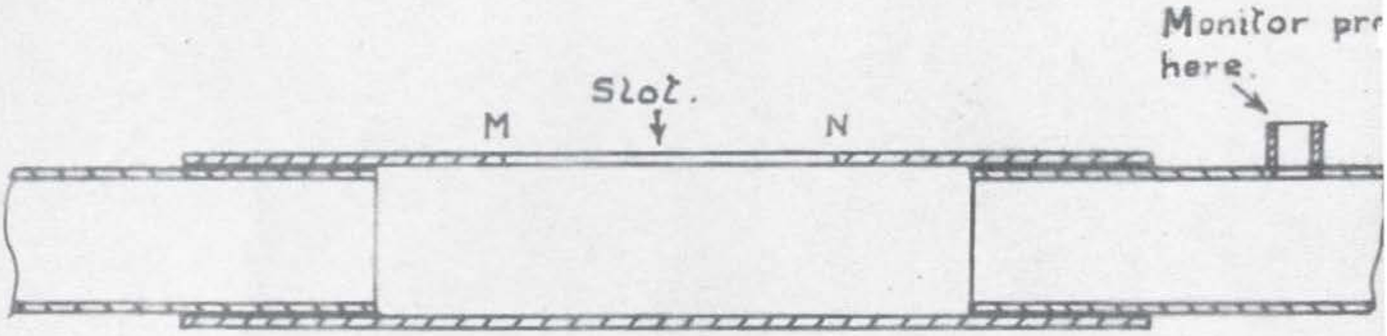


FIGURE 43 A, RIGHT.

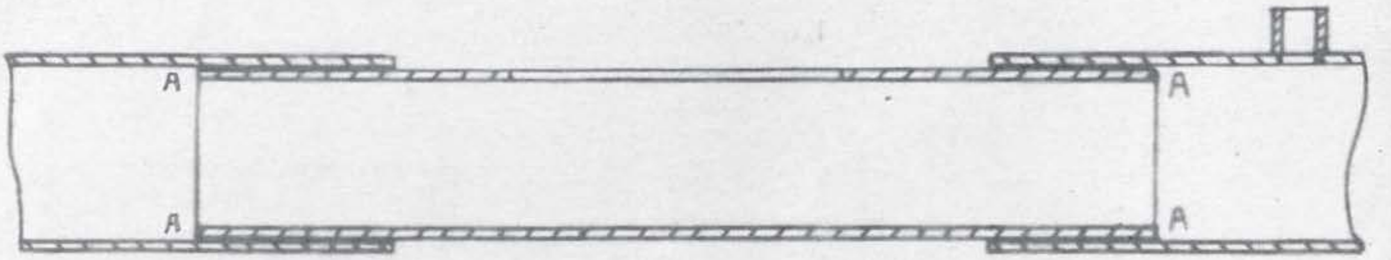


FIGURE 43 B, WRONG.

FIGURE 44.

the probe is attached to the side of a wave guide which telescopes the main guide as shown in figure 41A. For guides 1 inch in diameter the wall thickness of the sliding guide should not exceed 1/32 inch because of the disturbance caused by reflections at its ends AA. For guides 1/32 inch thick this disturbance gives a voltage standing wave ratio of about 1.05 : 1. To improve the contact at AA the tube should have longitudenal splits (about 5 mm. long) and a slight splaying out tendency given to the resulting tongues. The disturbance at AA can be reduced by tapering the metal as in figure 41B.

<u>Clamps for holding probes.</u>

Erratic electrical behaviour sometimes occurs because of a varying contact between the probe skirt and the clamp. It has been found to be advantageous to increase slightly, say by 2 thousandths of an inch, the diameter of the skirt as indicated in figure 42A. Another and a better cure is shown in figure 42B, which is self explanatory.

<u>7.3. Improved apparatus for standing wave measurement.</u>

When more accurate measurements of the standing wave ratio were required it was necessary to improve the design of the measuring instrument. The second, of the two types just given, is mechanically poor as it involves brass sliding on brass and , in consequence of this, electrical troubles sometimes occured at the ends AA, figure 41A. This is especially true for high power work. Preference was therefore given to the first type. It is obvious that if the slot is narrowed sufficiently no disturbance by it in the guide will be detected. Further if the slot should be sufficiently deep its attenuation properties prevent the loss of electrical energy. The high attenuation dispenses, as far as the electrical requirements are concerned, with the need for a metal plate covering the slot. If, for mechanical reasons, a plate should be used it will cause no electrical disturbance in the guide.

The method of testing for slot disturbances and the results obtained.

The degree of slot disturbance can be examined by mounting the slotted tube so that it slides over two pieces of waveguide rigidly held together as in figure 43A. This figure also gives the position of the monitor probe used for these measurements. On sliding the slotted tube, any change in the monitor probe current must be due to the slot. Mounting the tube as in figure 43B will not be satisfactory because, on sliding, the discontinuities at AA are also moved. (It has been suggested that by making the length of the slot electrically correct the reflections from one end of the slot will combine with the reflections from the other end and produce neutralisation. While this may be true as far as the wave guide is concerned, it is not true as far as the probe is concerned. This is because over the range MN, in which the probe slides, the slot disturbance in the wave guide is caused by a reflection from only one end of the slot). Tests showed that in a 1 inch wave guide a slot 1.8 mm. wide and 2½ mm. deep (i.e. the thickness of the wave guide wall was effectively 2½ mm.) had a negligible disturbance on the monitor probe (less than 1 in 70). It was necessary that the slot was positioned as in figure 44A. If the slot is turned so that the electric field has a component across the slot, power is of course immediately transmitted through the slot. A rotation up to 10° from the position shown in figure 44A was allowable before this effect became troublesome.

Details of the probe.

(a) Details of the skirt. The skirt had to be much smaller than had previously been used;

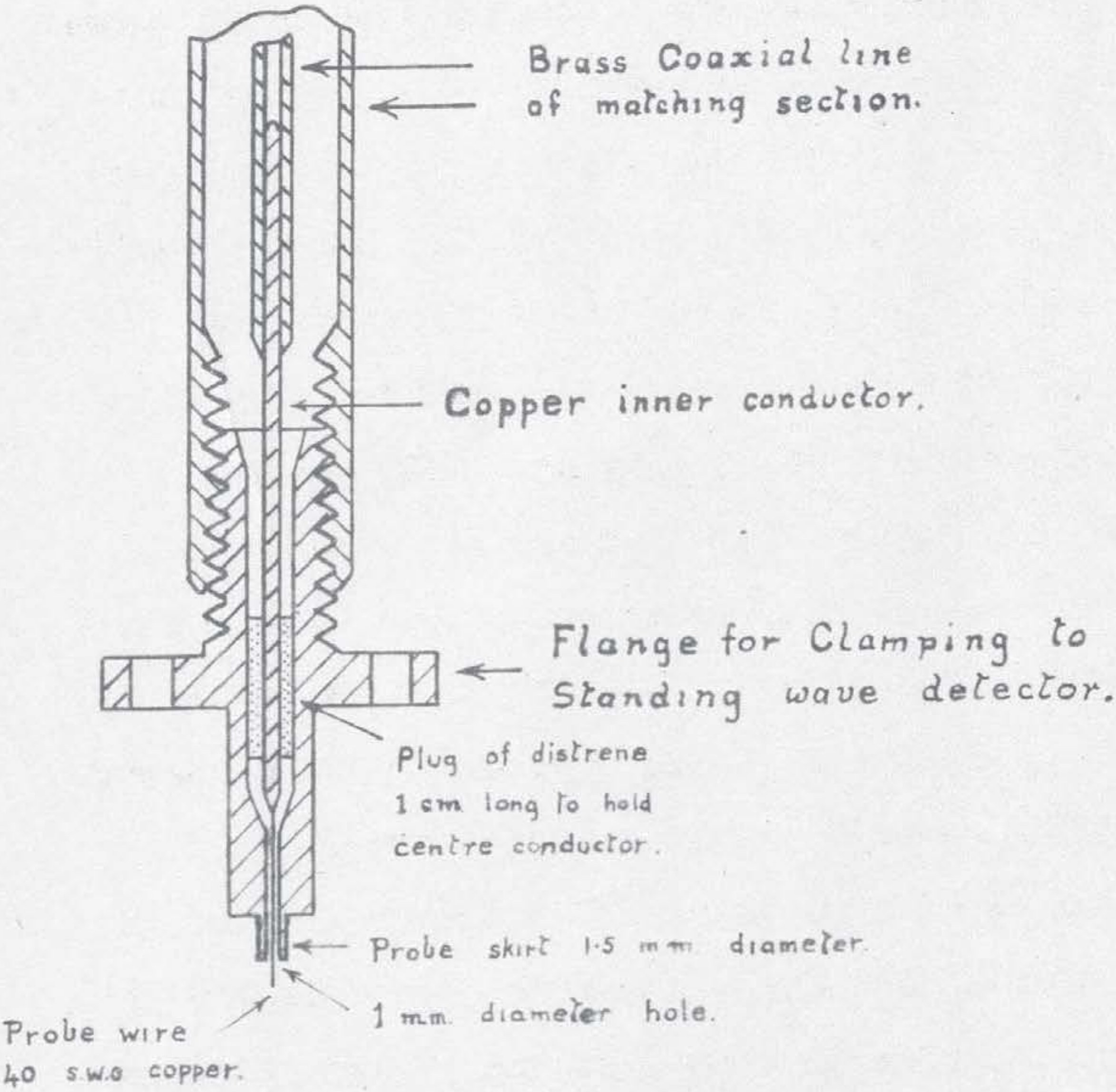


FIGURE 45.
SCALE: TWICE FULL SIZE.

Author's design of probe for use with Standing Wave Detector, ee also figures 46 and 47.

a suitable design is shown in figure 45. This skirt has to move along a slot which, however well made, has irregularities. In the best cases these irregularities will be small, but, to reduce their effect, and the effects caused by any slight shake in the probe carriage, it is best to pass the skirt into the slot and perhaps allow it to project slightly into the guide. The amount of immersion possible without disturbing the guide was found by an auxiliary monitor probe; the experimental set up being similar to that in figure 43A. The test is performed by moving the carriage, with skirt, along the slot, and increasing the immersion until the monitor probe becomes affected. This occured with an immersion of 0.5 mm. and hence it can be concluded that an immersion of 0.25 mm. will be safe. (For this test the end of the guide can be left open to the air as the resulting standing wave is of no consequence to the experiment).

(b) Details of probe wire. The diameter of the wire did not seem to matter: about 40 s.w.g. bare copper was used, and a length of 1 mm. projecting beyond the skirt was found satisfactory. The method of testing was similar to that just described for the skirt, except that it is necessary to connect the complete galvanometer circuit and to adjust the

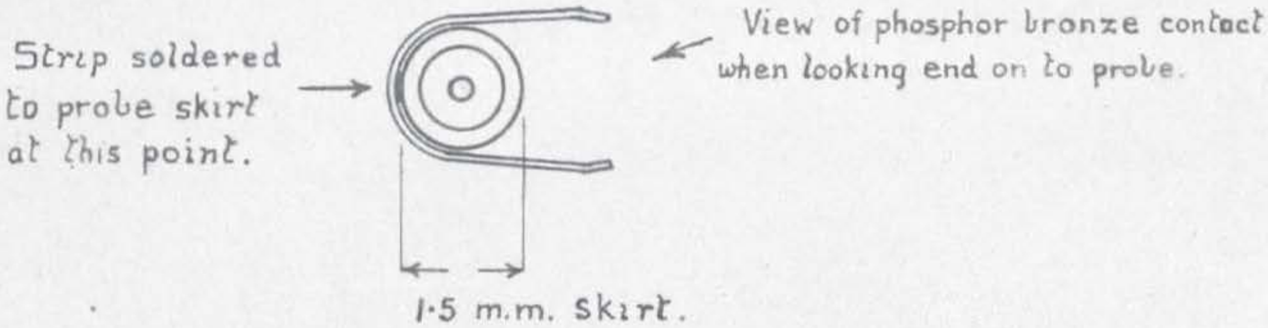
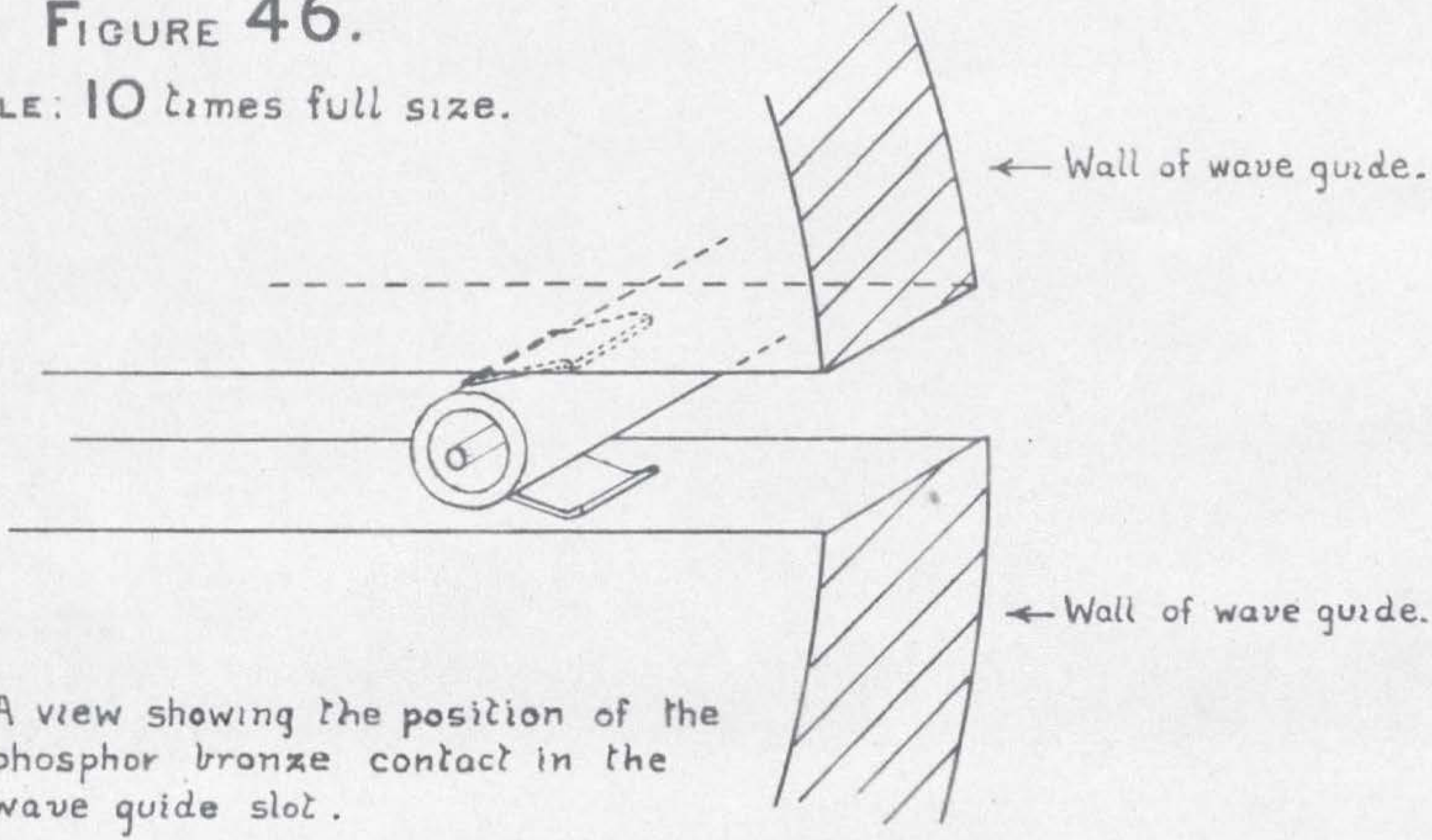


FIGURE 46.

SCALE: 10 times full size.



A view showing the position of the phosphor bronze contact in the wave guide slot.

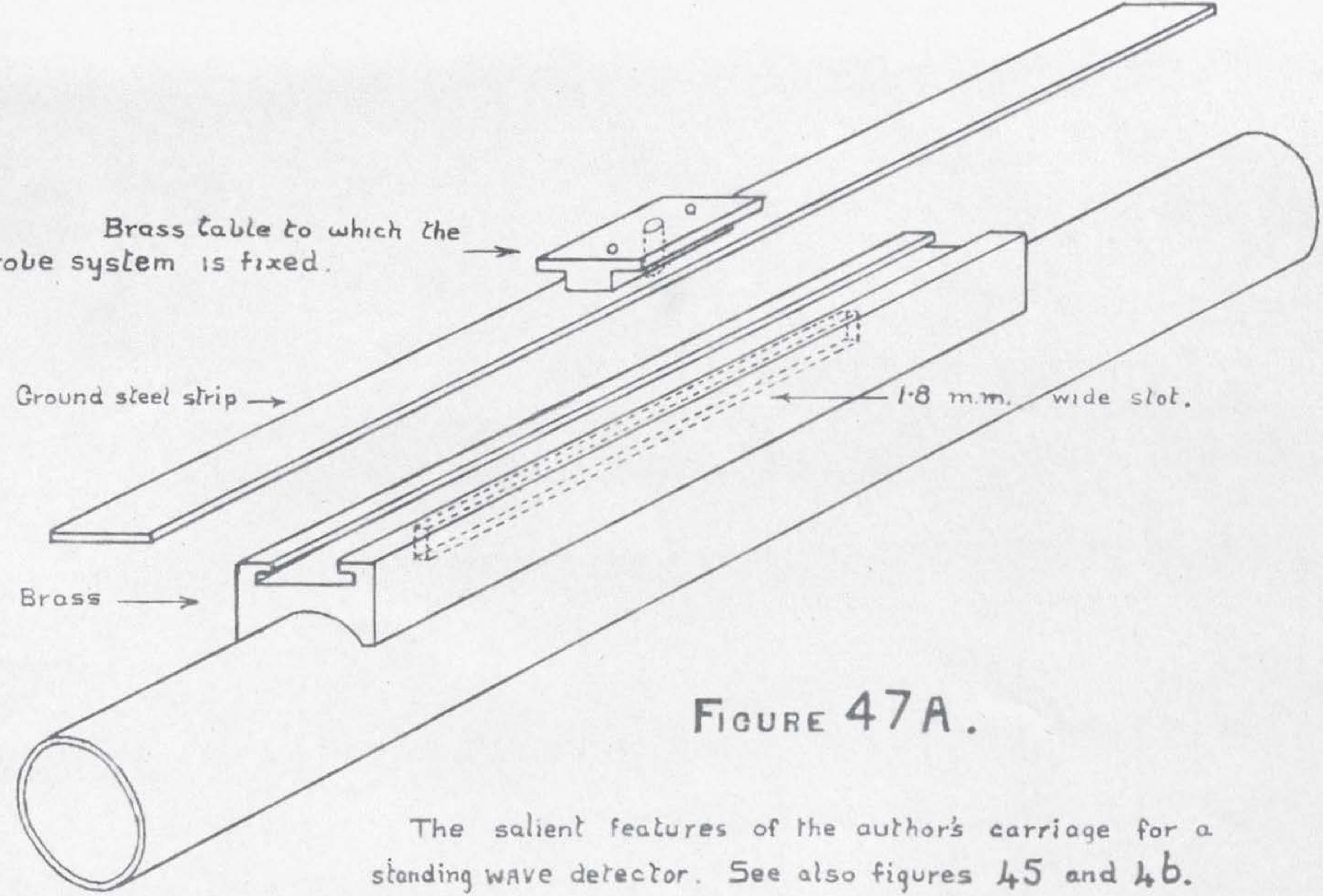


FIGURE 47A.

The salient features of the author's carriage for a standing wave detector. See also figures 45 and 46.

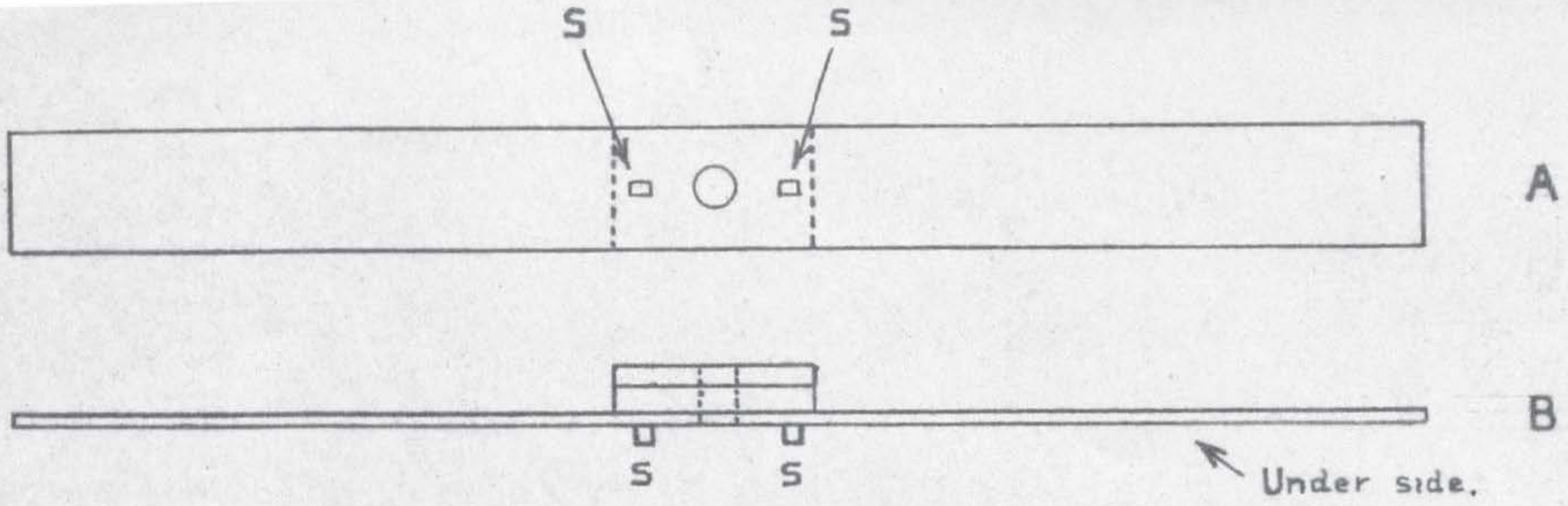


FIGURE 47B. A is a view of the under side of the steel slide, and shows the locating studs SS.
B is a side view of the slide.

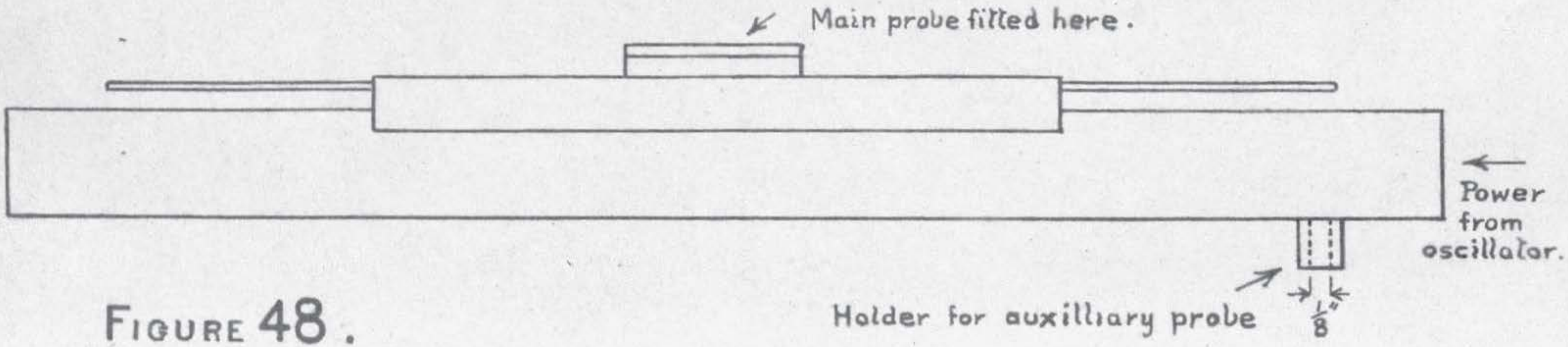


FIGURE 48.

matching system. It must never be overlooked that the distortion of the wave guide field due to the probe depends not only on the immersion of the probe but also upon the adjustment of the matching system. For the 1 mm. probe length, the matching section can be set to yield a maximum output on the galvanometer and it was always under these conditions that the probes were used.

The first experiments indicated a very erratic behaviour in the deflections indicated by the recording galvanometer. Fluctuations as much as 20% would sometimes occur and the system was useless until two springy phosphor bronze contacts were soldered to the skirt to rub against the sides of the slot, figure 46. Very satisfactory results were then obtained.

Details of the carriage for the probe.

A simple carriage is shown in figure 47A and 47B. It consists of a steel strip, guided by two studs fitting the width of the slot and held down by brass plates. The strip is made from steel because brass sliding on brass is unsatisfactory. The studs can easily be made short enough to prevent any possible interaction with the electric waves. It is essential that the machining and the workmanship of the guides should be good and that the carriage runs true with respect to the slot. For instance if the guide has an error which alters the probe immersion by a thousandth of an inch a change of some 2% in the probe output can be expected. In the apparatus, the author constructed, the output from the probe was some 8% lower towards one end of the slot. By careful scraping of the guides this 8% could be reduced to less than 2%.

7.4. The general performance of the improved probe system and the conclusion to this section.

The performance of the high quality instrument for the determination of standing wave ratios was found very satisfactory and much superior to any other previously used. In cases where the signal strength is low it is obviously an advantage to use the maximum immersion of the probe, consistent with its causing no undue disturbance in the guide. In order that this disturbance could be easily tested for, an auxiliary probe was inserted as shown diagramatically in figure 48. It is immaterial whether or not the guide is affected by the presence of this probe since it need only be inserted when tests on the main probe are required. On withdrawal of the probe, a hole is of course left in the side of the wave guide but if the size of this hole is of the order of 1/8 inch in diameter and at least $\frac{1}{4}$ inch in length its presence does not cause a measurable disturbance to the field in the guide.

SECTION 8.

WAVE GUIDES.

Summary.

This section, apart from the last subsection 8.5. contains no new work by the author, but deals with such fundamental properties of wave guides as are involved in his researches described in later sections. No attempt to derive these properties is made as a full detailed account of, e.g. the cutoff frequency, can be found in the many papers dealing with the subject, (2) (3), (11), (44).

8.1. General properties.

A wave guide is a metal pipe filled usually with air although in some cases it is filled with a low loss dielectric such as polystyrene. If the material in the guide has a permeability μ and a dielectric constant ε , then the electric field E, and the magnetic field H, must satisfy the Maxwell field equations of:-

$$\frac{\varepsilon}{c}\frac{\partial E}{\partial t} = \text{Curl}\, H .$$

$$-\frac{\mu}{c}\frac{\partial H}{\partial t} = \text{Curl}\, E .$$

$$\text{div}\, E = 0 .$$

$$\text{div}\, H = 0 .$$

The boundary conditions for these equations are given by the walls of the guide and it is adequate for the purpose at hand to assume these are perfect conductors.

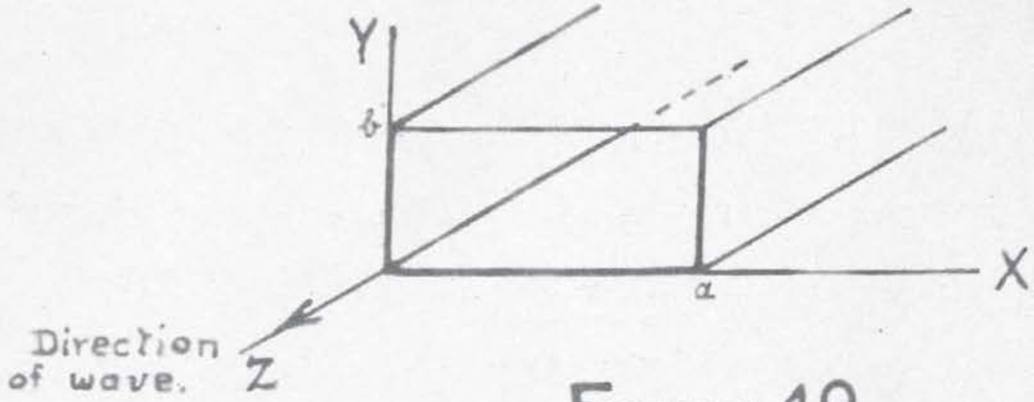
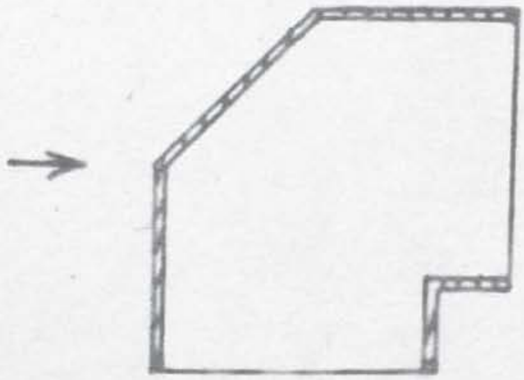


FIGURE 49.

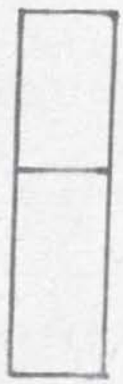
Guide.	Wave.	Cut-off wavelength.
Rectangular, air filled.	H_{01}	2·0 a.
Circular, air filled. d = diameter.	H_1 E_0 E_1 AND H_0	1·71 d. 1·31 d. 0·82 d.

"a" is the width of the rectangular guide. (For the H_{01} the sides, giving this width, are perpendicular to the electric vector.)

FIGURE 50.



A right angle bend for a wave guide (shown as a section).



View from direction of arrow.

FIGURE 51.

For a wave travelling in the direction of Z, the dependence of the Z coordinate can be in the form $\exp(-\gamma Z)$. By choosing a system of axes as shown in figure 49 the general equations, by elimination of E_x, E_y, H_x, H_y, yield two independent equations in E_z and H_z. The form of these equations is:-

$$\frac{\partial^2\phi}{\partial x^2} + \frac{\partial^2\phi}{\partial y^2} + k^2\phi = 0 \qquad (1).$$

where
$$k = \frac{\omega^2}{v^2} + \gamma^2 \qquad (2).$$
$$v = \frac{c}{(\varepsilon\mu)^{1/2}} \qquad (3).$$

and so, apart from the solution $E_z = H_z = 0$ (which is of no present concern as it applies to a guide with an internal conductor) there are two sets of solutions:-

$H_z = 0$ with E_z a solution of equation (1) - (4)

$E_z = 0$ with H_z a solution of equation (1) - (5)

For free propagation the angular frequency of the wave must exceed a critical value ω_c given by equation (2), as otherwise γ becomes imaginary. This value of ω_c depends on the guide dimensions and the type of wave within it. The waves corresponding to the solutions of equation (4) have a component of electric force in the direction of propagation and are termed E waves. Similarly the waves corresponding to the solution of equation (5) are termed H waves. There are many forms of E and H waves but for the transmission of power only one type of each, the E_0 and H_1, merit our special consideration.

For the rectangular guide the solution, very briefly outlined above is comparatively simple but for the circular guide it is more involved. In practice the circular guides as well as rectangular guides are of great importance and the constants, for the more simple waves of each guide, are given in figure 50.

With the circular guide the H_1 critical frequency is lower than that of any other wave. Hence guides can be designed to pass the H_1 wave but to attenuate all others. This fact is frequently used to produce a pure H_1 wave, i.e., a wave of one type only. After the H_1 wave the next wave to be passed by the circular guide is the E_o wave. If care is taken in exciting the E_o wave in the guide, this wave can be obtained in a high state of purity, i.e. the amount of H_1 wave present is very small. The E_o guide should not be so large as to pass the E_1 and H_o waves which have a cutoff value given by $\lambda = 0.82.d.$

All hollow guides (or guides filled with a dielectric) have their phase velocity (v') greater than the velocity of light (c) in the unbounded medium. The phase velocity is given by:

$$v' = \frac{c}{\sqrt{1 - (f_o/f)^2}} ,$$ where f_o is the cutoff frequency of the guide.

From this it follows that the wavelength λ_g in the guide is greater than that of free space. The velocity of energy transmission (the group velocity) is given by $u' = c\sqrt{1 - (f_0/f)^2}$.

Wave guides as transmission units.

To a very large extent wave guides resemble the more usual feeder lines. For example a wave guide can be correctly terminated by a resistive loading just as the coaxial line can be correctly terminated by its characteristic impedance. Further a guide of one size meeting a guide of another size requires a matching unit if standing waves are to be avoided.

Impedance elements for wave guides usually take the form of irises, i.e. rings of metal fitting the guide. A matching section usually consists either of an iris of variable size which can be moved at least a distance of $\frac{1}{2}\lambda_g$ along the guide, or, a section containing two independently moving irises, one capable of moving at least $\frac{1}{2}\lambda_g$ and the other at least λ_g. For a full understanding of matching sections the author gives reference (44) from which further references can also be found.

Note. The similarity of behaviour of the coaxial line and the waveguide is often helpful when one is making preliminary operations with wave guides. The author feels however that it cannot give a true understanding of the wave guide. For example, there are no two terminals to a guide and the impedance can no longer be defined as $\frac{V}{I}$. It is indeed necessary to examine the guide from the wave aspect (46); the impedance, for example, can then be defined as a field impedance, i.e. as $\frac{\text{Elec. field (V./metre)}}{\text{Mag. field (A./metre)}}$.

8.2. Relative merits of coaxial and wave guide feeds.

For high power operation there is no doubt as to the value of the hollow wave guide, but before its advantages are mentioned it will be useful to give a few of the almost indespensible uses of the coaxial cable.

Coaxial cable.

Unlike the wave guide the coaxial cable has no cutoff frequency and so a small cable of outside diameter about ¼ inch can operate over the complete frequency range (assuming the power is not such as would cause a voltage breakdown). Also, the flexibility of the cable makes it invaluable for standing wave measurments and for low power feeds such as are frequently mentioned in previous sections. In the 9 cm. region the attenuation is very roughly between 5 and 10 db/m, and increases as the wavelength is decreased. Long distance working is therefore impossible, but for feeding oscillator power to instruments this attenuation of a few db. is often useful for obtaining stability. The cable diameter should not be too large, for then wave guide propagation becomes possible and an impure wave is formed. For example, an air coaxial cable made from conductors $^{11}/_{16}$ inch and $^{3}/_{16}$ inch in diameter is roughly on the border line of giving wave guide propagation for 3 cm. waves.

Wave guides.

For these guides there is a minimum diameter of the order of λ, and inconsequence of this, wave-guides are usually only practical for wavelengths less than 10 cm. Guides can be reduced in size by $\frac{1}{\sqrt{\varepsilon}}$ by filling the guide with a medium of dielectric constant ε, but then, at the wavelengths considered, there is a very appreciable loss of energy even with the best of insulators. Thus the chief disadvantage of guides is their size and consequent inflexibility. There are however notable advantages. The first of these is that, owing to there being no relatively small inner conductor or dielectric insulation in the guide, the attenuation is very much less than that of the coaxial, being about 0.15 db/m. for copper construction at 3 cm. wavelength. Permanent bends can be made with these hollow guides and if made over a distance of several wavelengths cause only a very small disturbance to the wave form. Experiments in other establishments

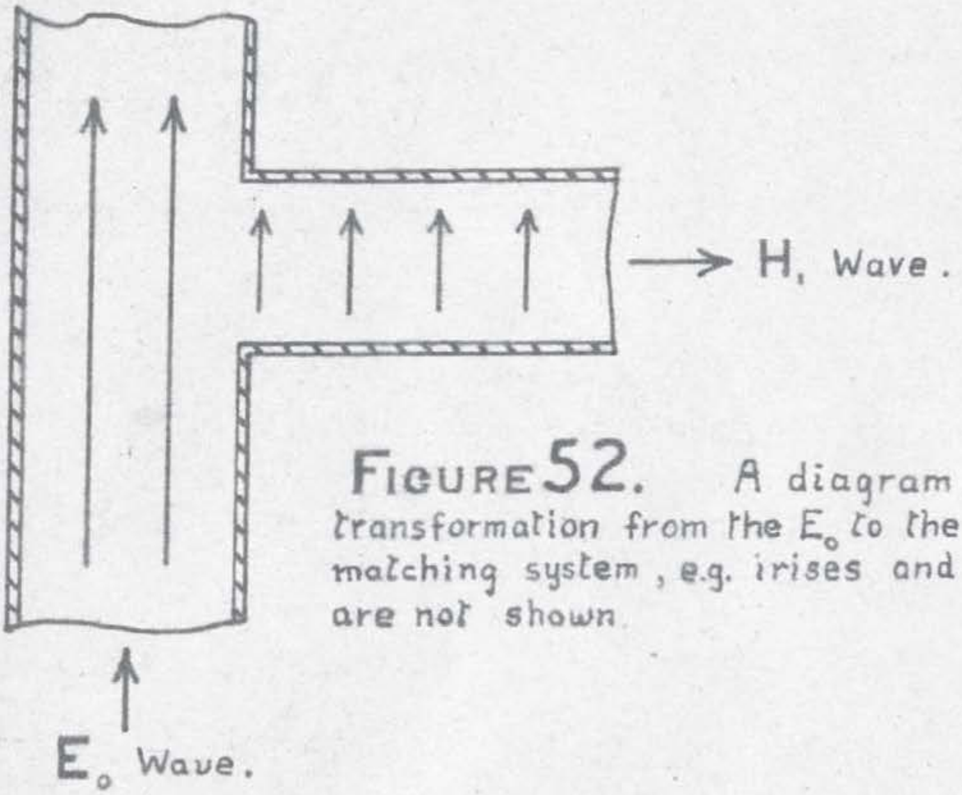


FIGURE 52. A diagram showing the transformation from the E_0 to the H_1 wave. The matching system, e.g. irises and a piston at A are not shown.

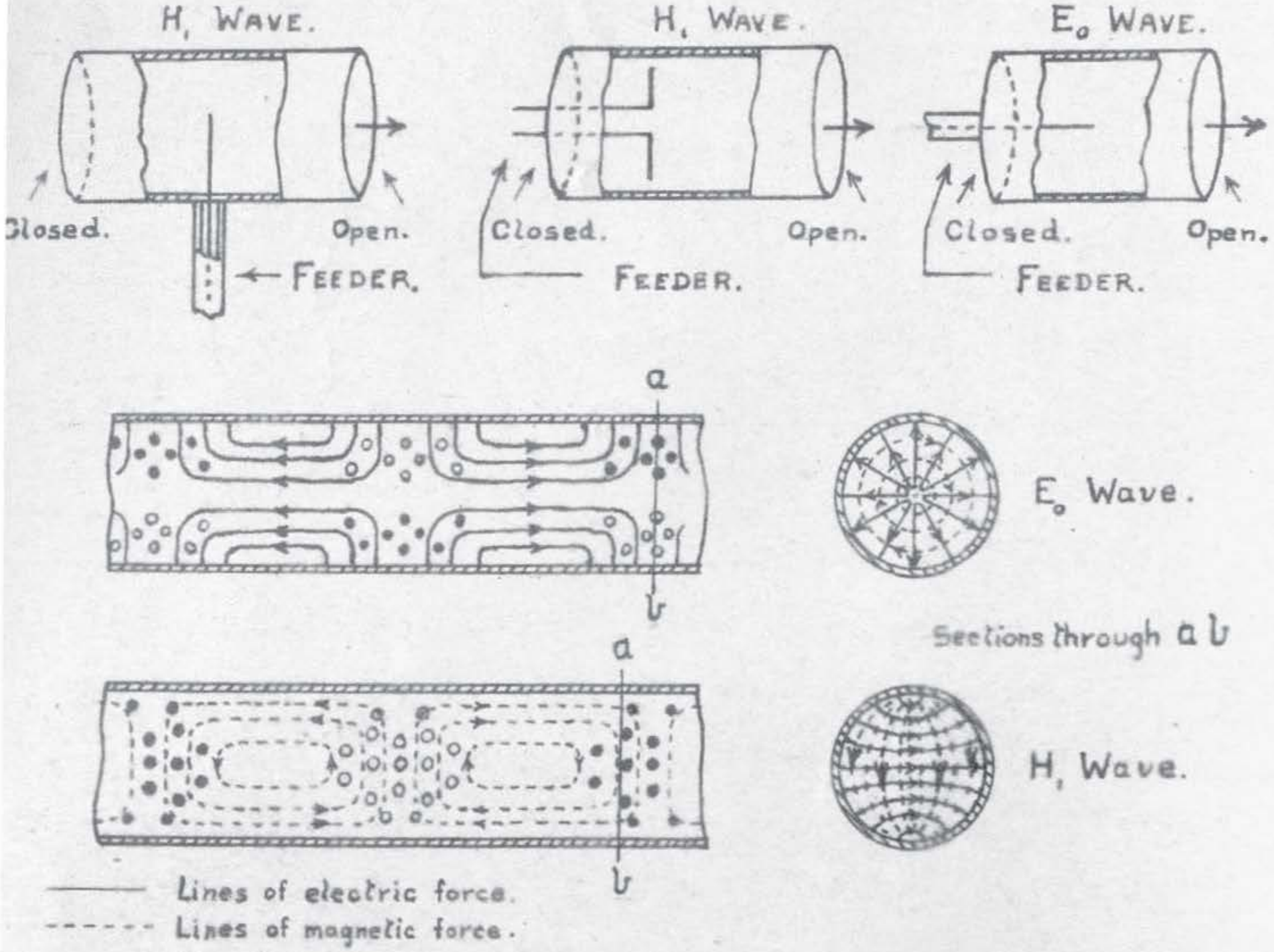


FIGURE 53. THREE METHODS OF EXCITING WAVE GUIDES BY A FEEDE INE, AND ALSO THE LINES OF FORCE IN CIRCULAR GUIDES CARRING THE E_0 AN HE H_1 WAVES.

(47), (48), have shown that with proper design a sudden right angle bend is allowable. A design taken from one of the papers mentioned is shown in figure 51.

8.3. Changing the wave from one type to another.

The wave type may be deliberately changed from one to another because each type has its own particular advantages. As already mentioned the H_1 wave can travel where all other waves are highly attenuated, but on the other hand, it, unlike the E_o wave, lacks symmetry about the direction of propagation. For purposes involving rotatable joints the E_o wave is often essential. A transformation from one type to another is therefore sometimes desirable and it can be achieved by suitably feeding the wave into the second guide. The principle of the transformation can be understood by considering the direction of the electric vectors in figure 52.

8.4. Wave guide feeds.

The usual method of feeding a wave guide is to excite the field by a small half wave antenna in the guide. The antenna shape is selected to suit the type of wave propagated, that is, its electric and magnetic fields approximate to those required in the guide. The actual fields set up are very complicated but if the guide is not unnecessarily big only the

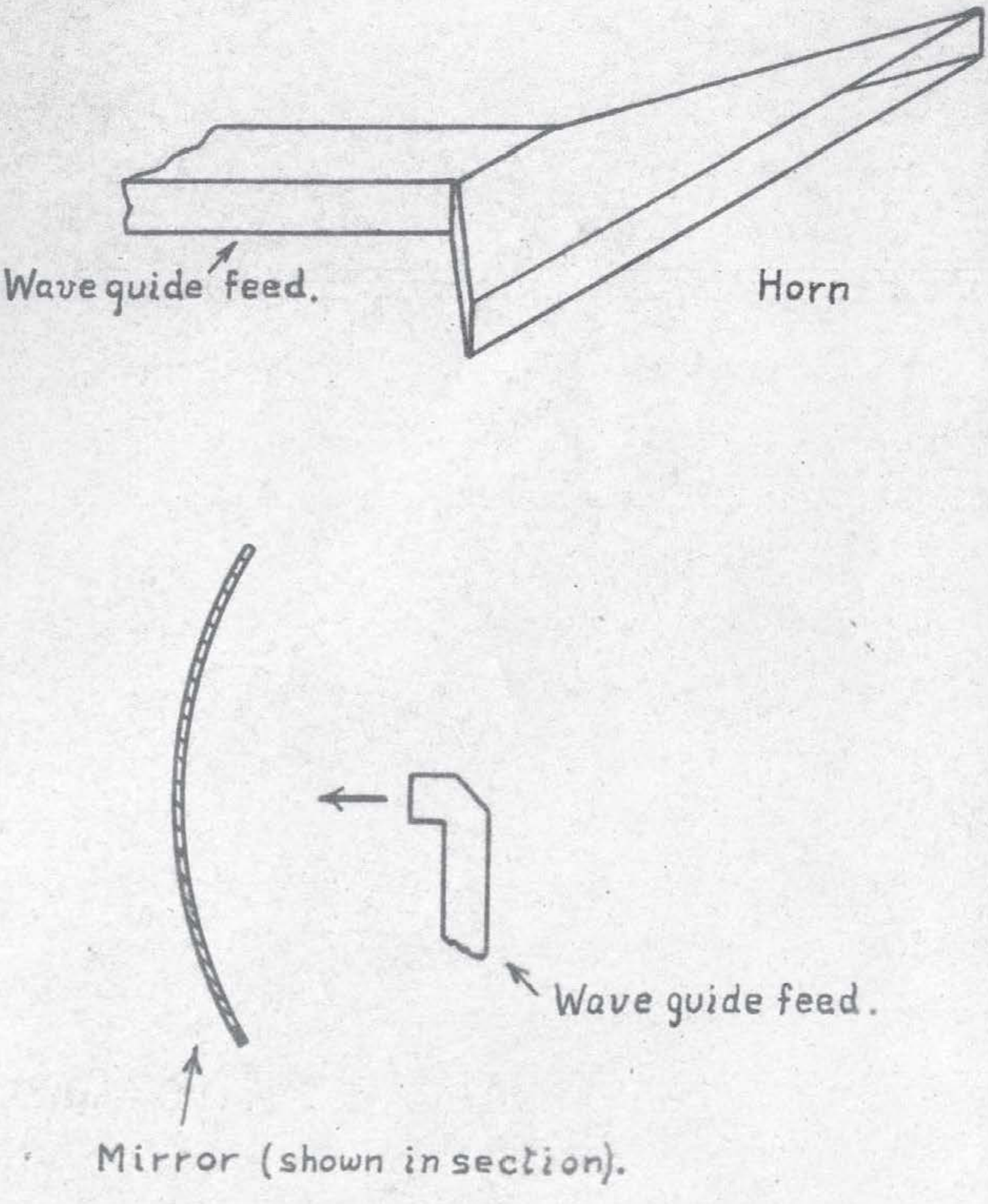

Wave guide feed.
Horn
Wave guide feed.
Mirror (shown in section).

FIGURE 54.

relatively simple waves can be propagated.
Methods of excitation of the E_0 and H_1 waves are shown in figure 53 and are self explanatory. Matching is generally performed by movable irises. Power can be taken from the guide by a system similar to that which feeds the guide, although for high power transmitters radio frequency power by radiation from the end of the guide is preferable. Highly directive beams can be obtained by electromagnetic horns (4) or by mirrors (49), figure 54.

8.5. The joining of one guide to another.

The joining of guides presents no difficulties and if a slight mismatch can be tolerated the guides can be telescoped together. For guides $\frac{1}{32}''$ thick, the voltage standing wave ratio, due to the mismatch, is about 1.05 : 1. The inner guide is usually sprung by a short longitudinal slotting to ensure good contact at the end of the guide. The author found however, that with thin tubes good joints could be made by slightly expanding the end of the inner tube (e.g. by forcing in a metal cone), thus giving a good rubbing contact for the important first millimeter of the guide. Under high powers, sparking at poor joints and at adjustable pistons causes noticeable corrosion of the guides, no doubt due mainly to the formation of oxides of nitrogen. To ease the sliding motion of pistons etc., a little vaseline does not appear to be harmful.

In work where no mismatch must occur at the join, a flange butt joint is satisfactory. One such joint is shown in the diagramatic sketch of figure 55 of the next section. The flanges should be stout so that the screws may be well tightened, and some means, such as two pins, must be provided for aligning accurately the axes of the two wave guides.

SECTION 9

THE MEASUREMENT OF MAGNETRON OUTPUT POWER. WATER LOAD TERMINATIONS FOR WAVE GUIDES USING CENTIMETER WAVELENGTH RADIATION.

Summary

Experiments on a new type of absorber for the accurate measurement of radio frequency power are given. The design of two power measuring instruments (calorimeters) are described: one is for routine measurements on magnetrons fitted with coaxial output feeds, and the other is for magnetrons fitted with wave guide output feeds.

9.1.

In the early days of 3 cm. multi-resonator magnetrons, it became evident that some reliable output power measuring device would have to be built. As a result of this demand the author carried out the experiments described in this section. Coaxial output feeds were at that time fitted to magnetrons but it seemed to the author, and to other members of the laboratory, that the most satisfactory design for a power measurement unit would be one in which the matching section was so designed that it gave a transformation from the coaxial cable feeder, to a wave guide feeder.

This wave guide would then be correctly terminated by a non-reflecting water absorber to which the principles of continuous flow calorimetry could be applied.

One of the most important conditions the author placed on the absorber was that its properties should be independent of wavelength change, preferably over a range at least as much as 2.8 cm. to 3.6 cm. For this reason he did not favour the use of matching sections, such as a length of dielectric a quarter of a wavelength long, whose dimensions were a function of the wavelength. It might seem that this simplification is unnecessary but if the termination is not sensitive to wavelength the operation of the measuring unit is greatly simplified. This is an important consideration where routine work is concerned.

Another important property would be the response time of the system, and, since a continuous flow method is envisaged, the volume of the absorber should not be so great as to require more than a second or two for the water to sweep through.

The design visualised by the author was some form of "Tapered" absorber; i.e. a slow transition from wave guide to absorber by making the latter in the shape of a wedge several half wavelengths long.

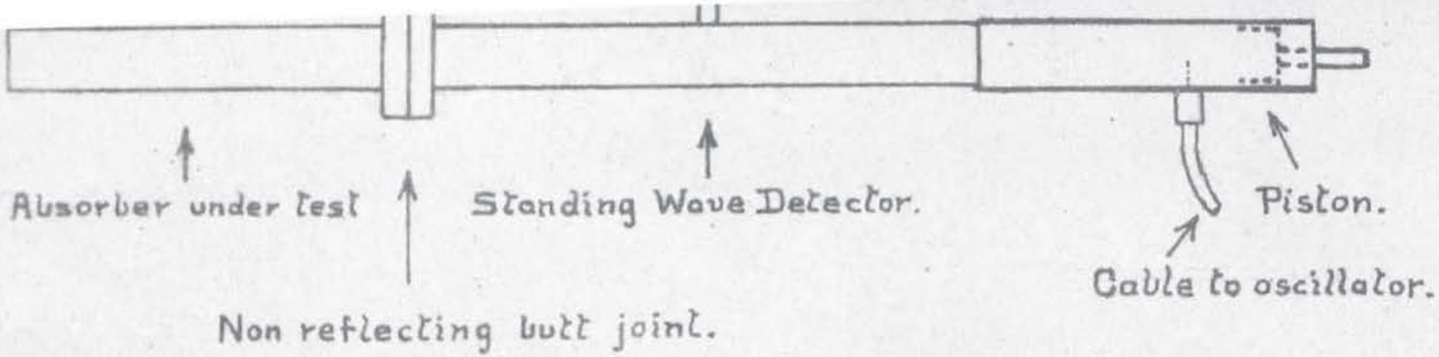
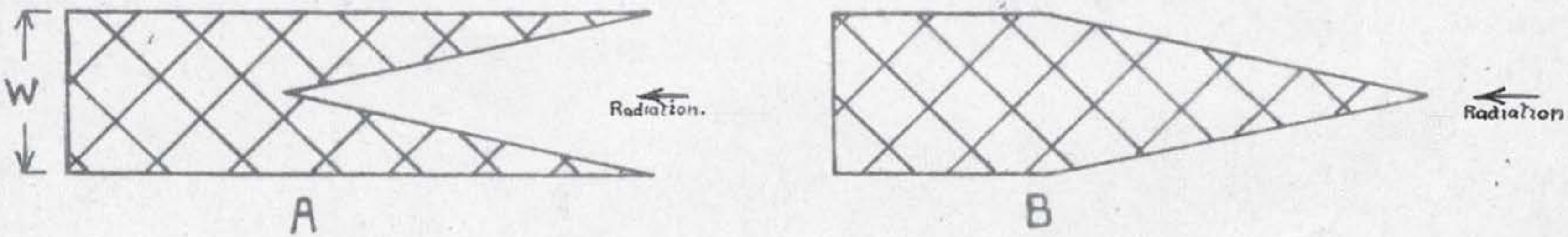


FIGURE 55 DIAGRAMMATIC SKETCH OF EXPERIMENTAL SET UP FOR TESTING ABSORBERS.



WEDGE SHAPED CARDS FOR PRELIMINARY EXPERIMENTS. THE WIDTH OF CARD, W, IS EQUAL TO THE DIAMETER OF THE WAVE GUIDE.

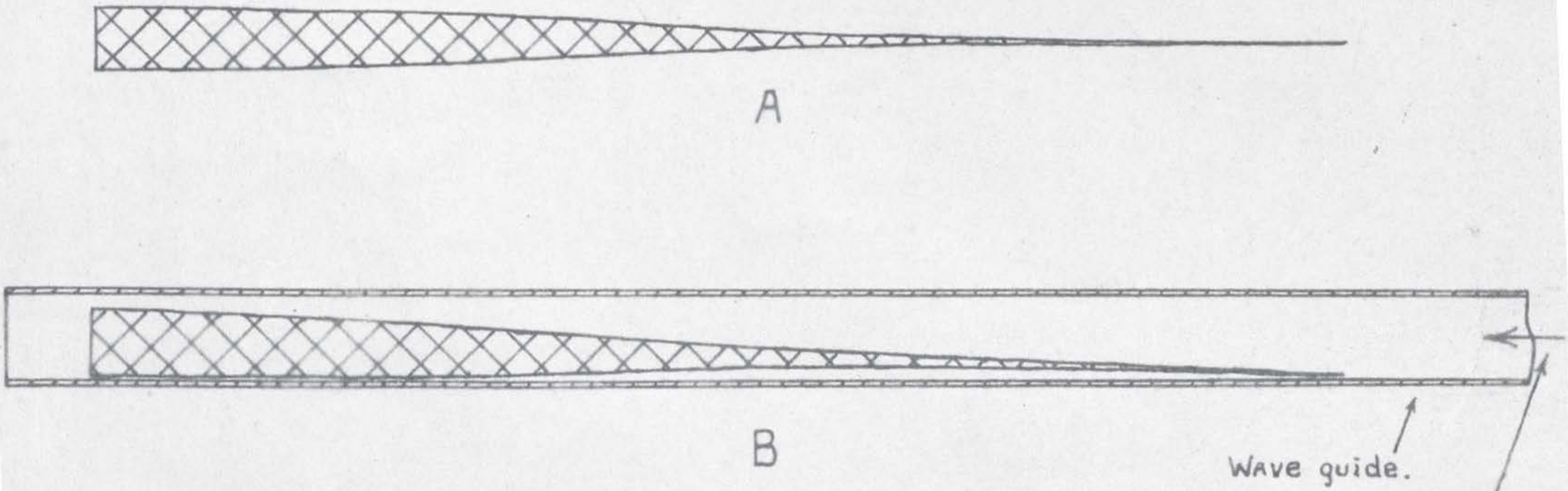
FIGURE 56.

With the help of experiments a rough idea of the reflection from a mass of water was known. Then, by assuming this water to be spread out over 5 wavelengths of guide, a very rough idea of the resulting reflection was obtained by graphical summation of the reflection over the length of water. The results showed a greatly reduced reflection, and therefore a reduced standing wave ratio in the guide.

9.2. Preliminary work and a termination for low power guides.

The experimental set up to test these absorbers is given in figure 55. Details need not be explained again because the standing wave measuring instrument, for example, has been fully described in section 7. The final design of section 7 had not however been developed at the time of these experiments which began in 1942. The type of standing wave detector used was the first of the two types described under sub-section 7.2. The probe system used was that of figure 33B, section 6, and was connected to a high gain amplifier to operate a cathode ray tube.

Preliminary experiments on tapered absorbers were made by using wedge shaped pieces of cardboard saturated with water to absorb the power. They were kept cool by an air blast in the wave guide. The shape A of figure 56 seemed preferable to that of B



A. Absorber.
B. Absorber resting in its position for use in a wave guide.

FIGURE 57.

Aquadag Absorber, the dimensions are in the text.

transmission units. It consisted of a piece of glass tubing of diameter about $\frac{3}{4}$ of that of the wave guide and drawn down to a point as in figure 57. The total length for a 2.5 cm. diameter guide was 60 cm., but a shorter length may also give good results. The outside of the glass is coated with aquadag in a layer sufficiently thick to be opaque to light. When so used it gives a voltage standing wave ratio of 1.05:1, or less. There is the advantage that the absorber is movable along the wave guide without the trouble caused by reflections from a telescopic section of wave guide. This movement is useful for the "trimming" of absorbers (see the end of this section) by judicious removal or addition of aquadag to improve the standing wave ratio. The properties of the termination appeared insensitive to wavelength changes over the region tested which was from 2.9 cm. to 3.5 cm. For work using 3 cm. continuous waves at a power of about a watt, such an absorber has been found very useful. It has, for example, been employed in testing the design of matching sections between magnetrons and their transmission feeders.

9.3. Glass contained water absorbers for high power measurements.

The form of a satisfactory absorber

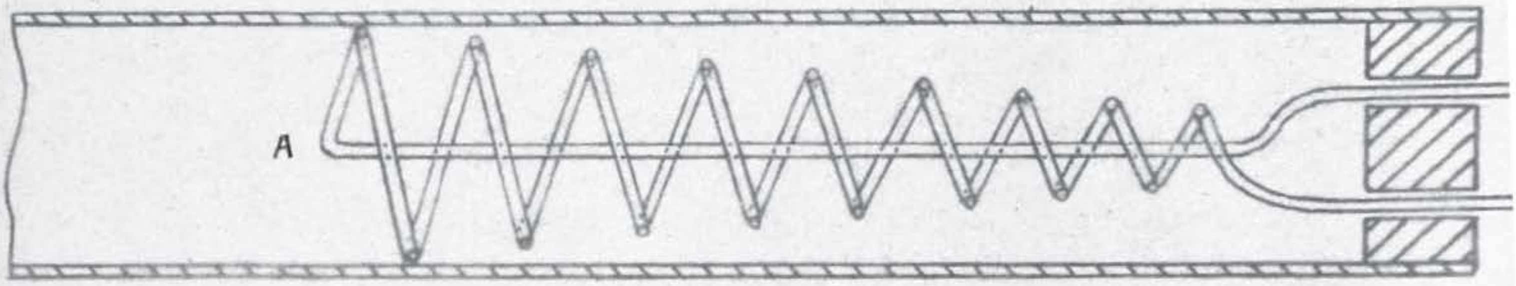


FIGURE 58.

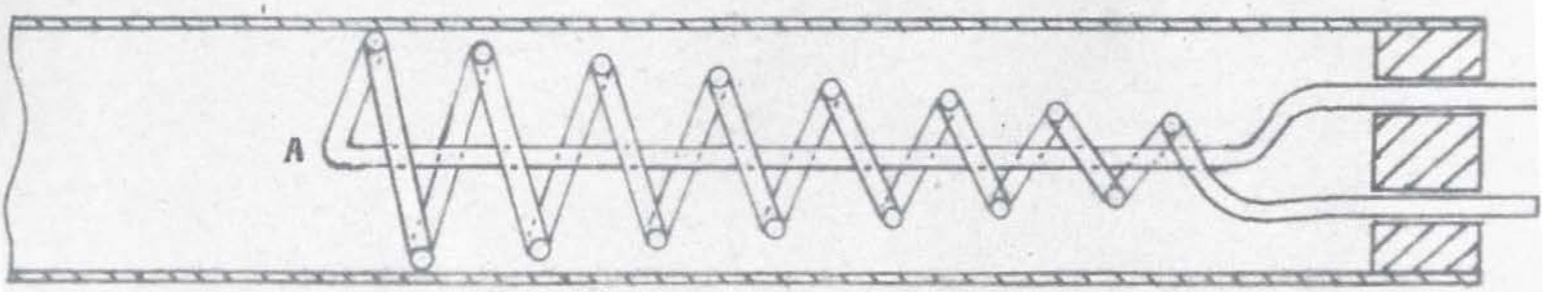


FIGURE 59.

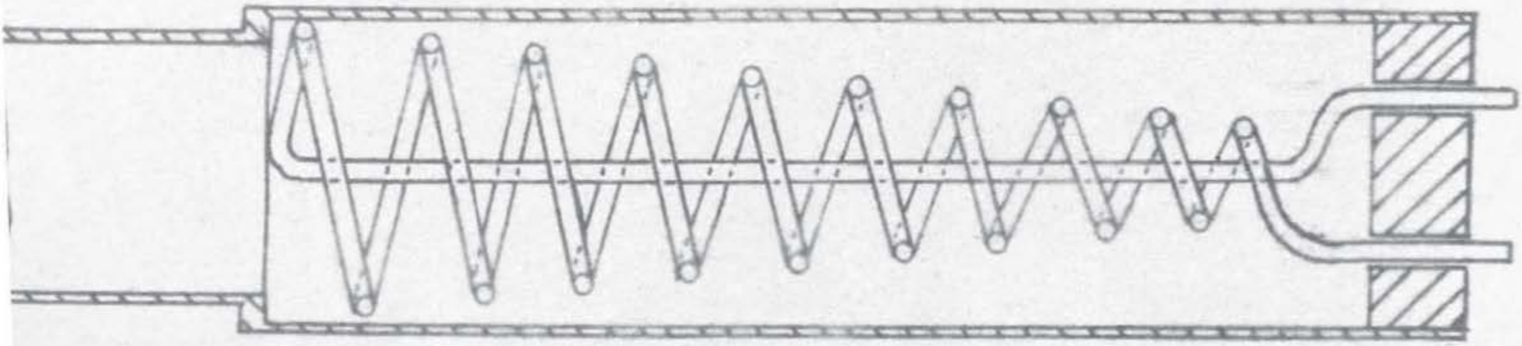


FIGURE 60.

Scale full size.

is now clear. It must begin as a cylindrical helix and then taper down towards the centre of the wave guide as suggested by figure 56A. Absorbers in the form of a conical helix, figure 58, were made from 0.75 mm. glass tubing through which water was passed. They behaved well giving a standing wave ratio of 1.1:1. By far the larger part of this standing wave was caused by the bend at A. Thus, if stationary water were used, the absorber became very hot at A while the remainder was barely warm. By reducing the diameter of the glass tubing to 0.5 mm. at the front end and uniformly increasing the tube diameter to 1 mm. towards the smaller end of the helix, the standing wave ratio could be improved to 1.03:1 in volts. Care must however be taken to keep the return tube, passing down the inside of the helix, as close as possible to the wall of the wave guide, especially at the front end. This design however has a resistance to the water flow which is too high for convenient use with constant head tanks.

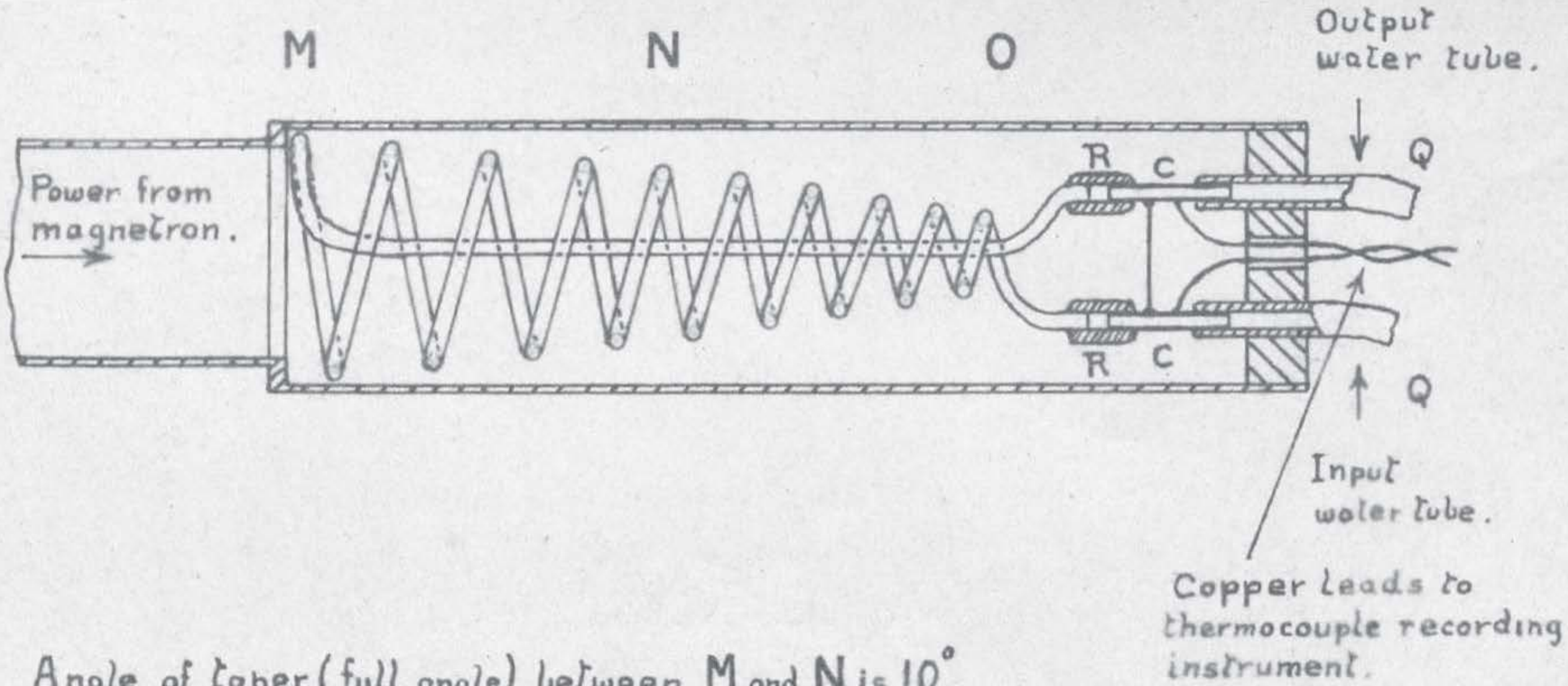
This difficulty of water flow resulted in the conical helix design being modified to one in which the impedance of the wave guide is changed in order to reduce the standing wave caused by the absorber. At the same time, to obtain a larger flow of water, the bore of the tubing was increased. A conical helix 10 cm. long was made from tubing of 1.9 mm. internal bore, figure 59. Such a helix gives a standing wave ratio of about 1.4:1 in volts, when care is taken in keeping the exit tube close to the wave guide wall at A. If ρ is the reflection coefficient of the absorber and if V is a measure of the voltage amplitude of the oncoming wave, then the reflected wave is ρV and the maximum to minimum of the resultant wave is $\frac{V + V\rho}{V - V\rho}$ or $\frac{1+\rho}{1-\rho}$. This, which is the value of the standing wave ratio, has just been stated to be about 1.4 which makes $\rho = 0.167$. Now a wave guide of $1\frac{1}{16}$ inch diameter, increased by a step to a wave guide of $1\frac{1}{4}$ inch, gives an impedance change in the ratio of about 1.4:1. This by the usual transmission line theory, produces a reflection coefficient of $\frac{1.4 - 1.0}{1.4 + 1.0}$ or 0.167. It follows from what has just been said that a similar absorber made to fit a $1\frac{1}{4}$ inch guide which is stepped up from a $1\frac{1}{16}$ inch guide, as in figure 60, should provide a very much better termination. This is because the impedance change due to the absorber should be approximately balanced out by the impedance change of the wave guide.

A unit was made according to the dimensions of figure 60 and it gave a standing wave ratio of 1.06:1

in volts. Two others were constructed in which the $1\frac{1}{16}$ inch guide, instead of increasing to $1\frac{1}{4}$ inch, increased to $1\frac{5}{32}$ inch for one and to $1\frac{3}{8}$ inch for the other. Each of these gave a standing wave ratio which was not quite so good and hence the initial size of $1\frac{1}{4}$ inch is not far from the best. This design of absorber is satisfactory mechanically. With a convenient water flow of 1 cc. per second entering at the back and leaving at the front, via the inner tube, the response time is about two seconds. To obtain the shortest response time the direction of flow is of some importance because nearly all the energy is absorbed in the front half of the helix and hence this end should connect with the exit tube. At faster rates of flow the response time is still more favourable. On account of the reflecting coefficient of a stepped wave guide being not critical to frequency, it is unlikely that the behaviour of the system will be affected by frequency changes of a moderate nature. Over the range of wavelengths from 2.9 cm. to 3.55 cm. no change was experimentally detected.

Final Design

The robustness and reliability of these absorbers led them to be adopted immediately for



Angle of taper (full angle) between M and N is 10°.

Angle of taper (full angle) between N and O is 16°.

C represents two copper tubes with thermocouple units attached.

R Rubber connecting tubes.

Q Rubber input and output tubes.

FIGURE 61.

Scale Full Size.

routine measurement. Inorder that the unit should be as compact as possible, experiments on a shorter design were made and it appeared that the taper could be steeper except for the first 3 cm. This resulted in the double taper design indicated in figure 61. The measurement of the heating of the water was by two copper-constantan thermojunctions. The thermojunctions were made by inserting two 1 cm. lengths of 1.5 mm. bore copper tube in the input and output water leads and connecting the copper tubes with a length of constantan as short and as thick as possible, to keep its electrical resistance low. The lower this resistance is kept the less the effect of the electrolytic voltages caused by the water at the copper tubes. On the other hand the constantan must not be so thick as to conduct an undue amount of heat from the hot junction to the cold junction.+ A constantan wire 1.9 cm. long and 0.08 cm. in diameter was chosen and the heat transfer through this, when the water flow is over 80 c.c. a minute, cannot produce an observable error.

9.4. An investigation into possible errors.

It will next be shown that only negligible losses can be introduced by:- (a) the loss of heat to the wave guide by its direct contact with the helix; and (b), the radiation and convection current losses from the helix to its enclosure. This will be

+See foot note next page.

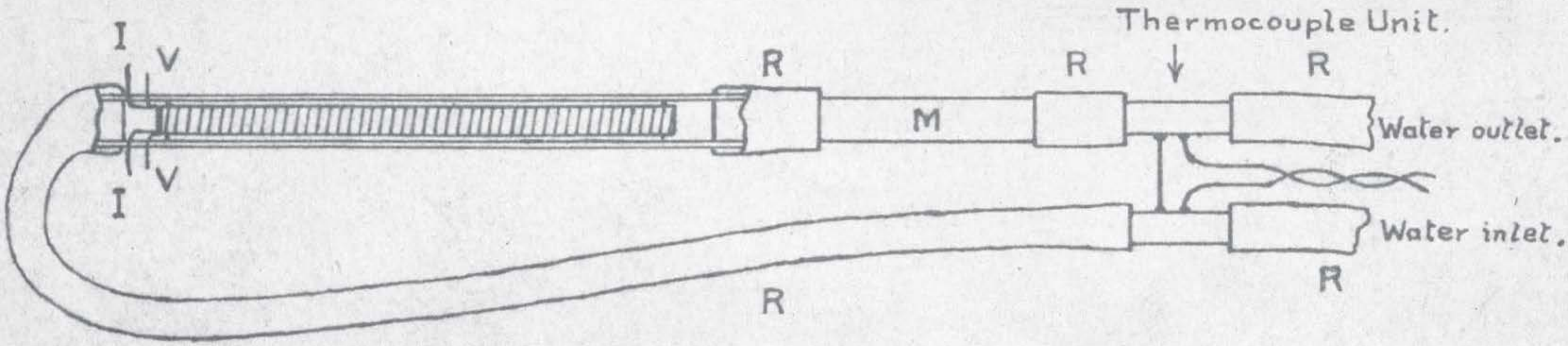
followed by an account of the calibration of the unit. For (a) it is known that the helix makes a local contact here and there because its front end rests in the step of the guide. The contact area however cannot be more than one or two square millimeters. A simple calculation of heat conductivities shows that with a water flow of 80 c.c. per minute (the minimum used) to lose 1% of the heat by conduction through this contact, the area of the contact would have to exceed 30 sq.mm. This area is far in excess of anything possible with this design. In the second cause of error, (b); the radiation loss can be estimated and amounts to not more than 0.5%. This assumes that the water is a black body radiator (the worst possible case) surrounded by a wave guide the walls of which have a heat reflecting power of some 97%, a reasonable figure for the long heat waves involved. The convection losses are not so easily determined.

+ It is interesting to note that if, between the input and output points of such a system heat is fed from the hot water to the cold water by some conductor, then parts of the system can have a temperature rise as much as double that which occurs between the input and the output water. This may result in an error, positive or negative according to the thermojunction positions.

The most convincing measure of the magnitude of the error involved under b is to obtain the combined losses of (b) by experiment. A glass tube 3.2 cm. in diameter, which is about the maximum diameter of the helix, and 19 cm. long was filled with water and placed in a brass tube which cleared it by 0.8 cm. A stirrer, a heater coil, and a thermometer were incorporated in the apparatus. To maintain the glass tube at 17°C above the surrounding brass tube, 1.8 W. were necessary. The glass helix handling 100 W. of radio frequency power runs at this temperature rise, and, since the area of the helix is a little less than 1/3 of that of the glass tube, the loss of power would be 0.6 W. if the entire helix was at 17°C above its surroundings. Actually, one end of the helix is cold, and the average temperature excess will be about one half of this value, thus giving a loss of about 0.3 watts or 0.3% which is a very small amount.

9.5. Calibration of thermocouple system.

The basis of the method used for the calibration of the thermojunctions was to heat the water electrically. This has the advantage that the input power is easily measured. The method might be criticised in that it is more fundamental to obtain this input power by measuring the temperature rise of the water. This however is not the case because we are not directly



Immersion heater contained in a glass tube and held central by three small projections at each end.

R Rubber tube.
I Current leads.
V Voltmeter leads.
M Water Mixer.

FIGURE 62 A.

interested in temperature but rather in the relation between the input watts and the thermojunction recording instrument. The voltmeter and ammeter used to record the power were both compared with the laboratory standard and suitable corrections were applied. Manganin wire number 30 s.w.g. was used for the heater coil, sufficient length being used to give a resistance of 35 Ω . The connections made on the manganin coil itself were from 26 gauge copper wire. Heat conducted away by these leads and the voltage dropped along them, cause a negligible error, the actual amount being about 0.1%. The first method of making the heater was to wind the resistance wire on the glass tube of the radio frequency absorber, and to lag it well with cotton wool. This method, while giving results in good accord with the final method, was discarded because the resistance wire, not being in direct contact with the water, ran rather hot. The final method used was to pass a stream of cold water through one thermojunction; then to heat the water by an immersion heater of known power, and to pass the water through the second thermojunction; figure 62A. The immersion heater was made by winding a single layer of 30 s.w.g. gauge cotton insulated manganin wire on a thin glass tube 6 cm. long and 3 mm. in diameter, the whole being waterproofed by a thin layer of bakelite varnish

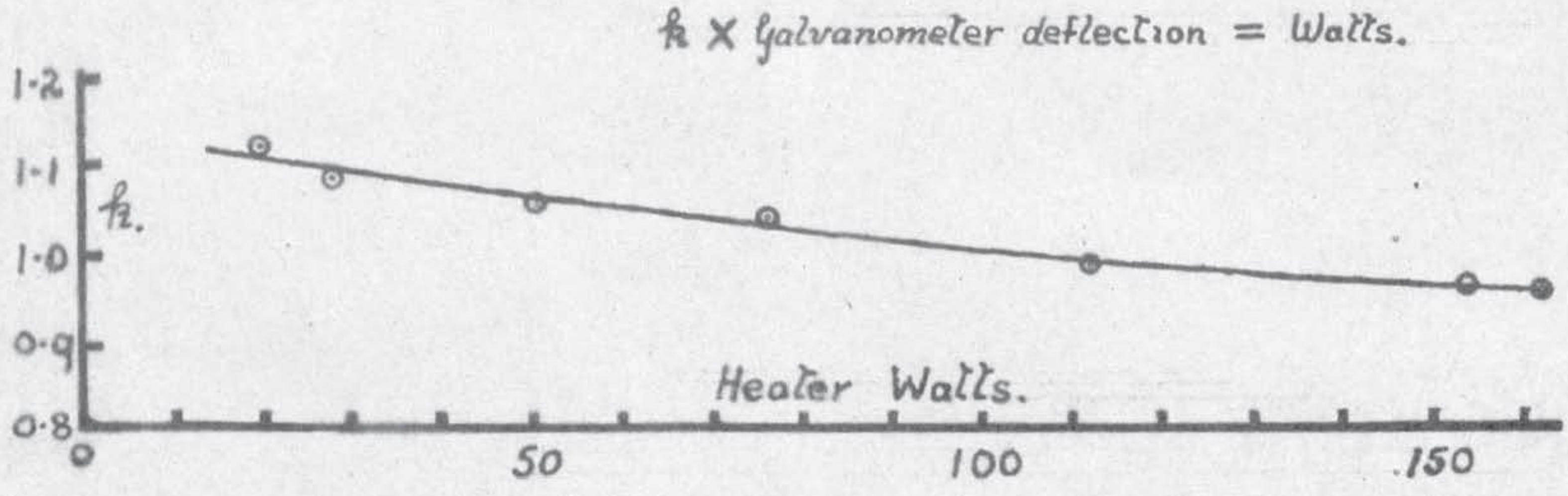
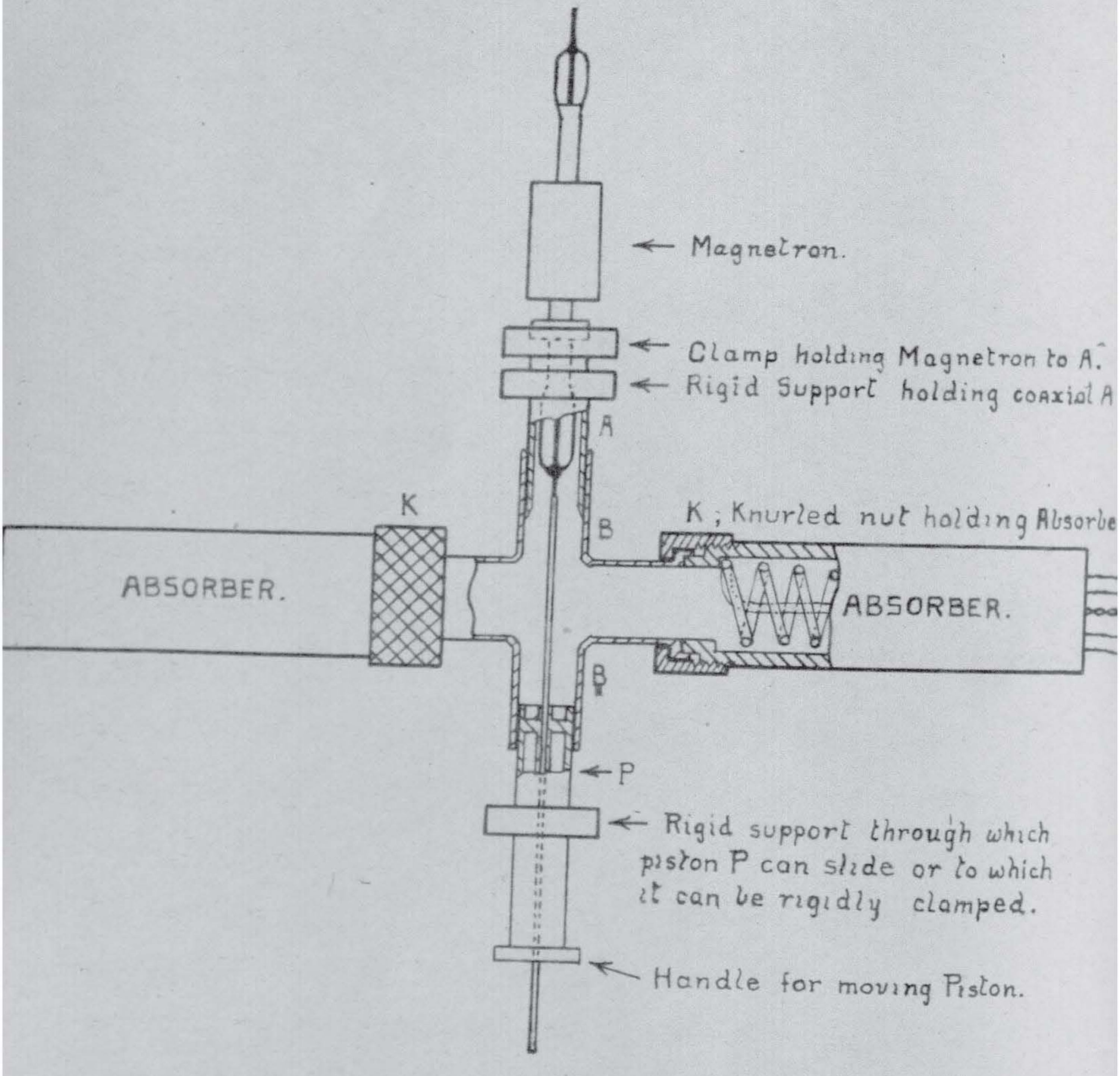

k × Galvanometer deflection = Watts.
1·2
1·1
k.
1·0
0·9
0·8
0
50
100
150
Heater Watts.

FIGURE 62 B.

and mounted as shown. To ensure homogeneity of water heating it was found necessary to insert a short length of tubing which had across its bore two or three pieces of wire to aid the proper mixing of the water. With a water flow of 84 c.c./minute, 100 W. input to the heater gave a deflection of about 100 on the galvanometer connected to the thermocouple. Calculations showed that because of the speed with which the water is transferred from the heater to the thermojunction the heat losses during this transfer are negligible. This was also verified experimentally. The results are shown graphically in figure 62B which also gives the value of k by which the galvanometer deflection has to be multiplied to give the number of watts taken by the absorber. An additional check was made for an electrical input of 154.2 W. by measuring the input and the output water temperatures. The watts calculated from the temperature change were 1.4% lower than the input watts measured electrically. Some of this error is no doubt due to heat losses from the comparatively large container necessary to house the 1 cm. long thermometer bulb. This discrepancy is, however, of little account as the accuracy of these thermometer



Note. The coaxial B together with the absorbers can be slid on the coaxial A and the piston P. This motion is termed the cross slide motion.

FIGURE 63.

SALIENT DETAILS OF CALORIMETER FOR 3 c.m. WAVELENGTH COAXIAL VALVES. SCALE: $\frac{2}{5}$ FULL SIZE.

measurements does not warrant any correction being applied to the electrical measurement of input power.

9.6. The complete measuring calorimeter for radio frequency power.

A type of matching system (very suitable for use with coaxial output valves) together with a power absorbing system is shown in figure 63. The whole unit is termed a calorimeter. It will be noted that the design is unconventional in that there are two water absorbers. The water circuits are used in series and so also are the two thermocoupl systems. The use of two absorbers gives a very great advantage over the more obvious method which employs one absorber and a piston. This is because the piston becomes a third variable and considerably increases the matching difficulties. Another very big advantage of the double absorber design is that about 40% more power can be handled before arcing occurs at the junction of the coaxial feeder and wave guide. In operation the performance of this calorimeter is certainly superior to that of any other the author had seen. The adjustment of the cross-slide carrying the wave guide proved to be relatively insensitive; the coaxial piston was the more critical of the two adjustments. Inexperienced operators were capable of obtaining good results with

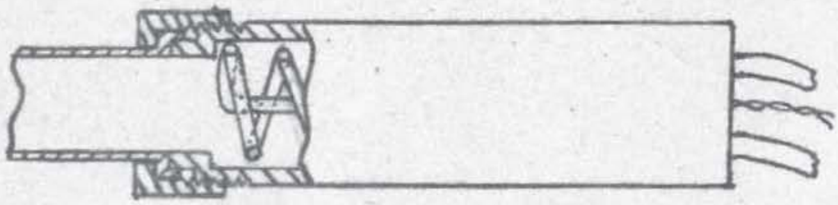


FIGURE 64 A.

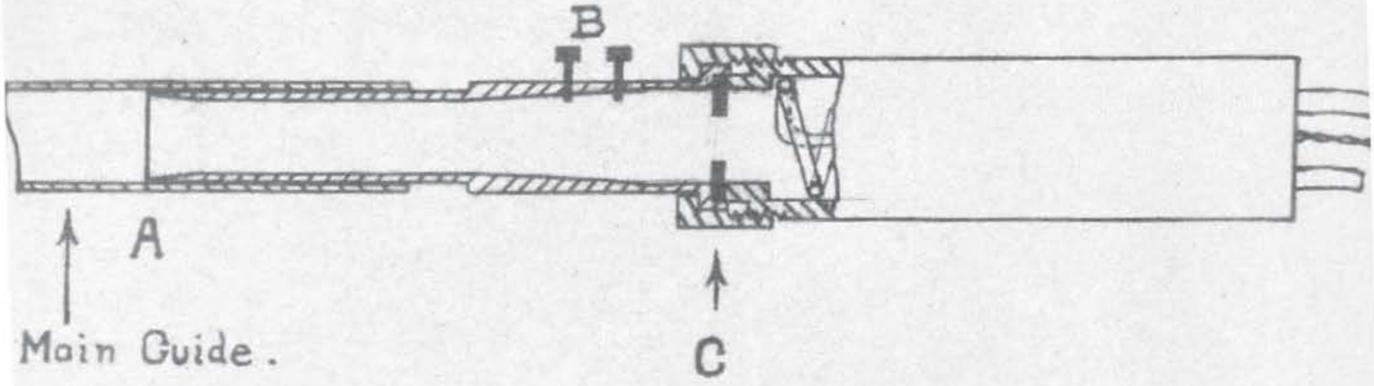


FIGURE 64 B.

Scale: For 3 c.m., this is about $\frac{2}{5}$ full size.

very little instruction, a success not often obtained with 3 cm. measuring apparatus.

9.7. A calorimeter for wave guide output valves.

The later development of wave guide output magnetrons (see next section) required the design of another calorimeter which was this time comparativel; simple to construct. The magnetron is already matched to feed the wave guide (or if it is not it should be matched by an iris transformer section) and the guide is terminated at the other end by one of the helical absorbers, (figure 64A). This system can, however, be put to a greater use, such as the determination of the frequency pulling of a valve, providing that the absorber is designed to slide <u>inside</u> the main guide as in figure 64B. For all positions the discontinuity at A is then fixed with respect to the absorber and the reflections from it can therefore be corrected for by two screws (10 B.A.) at B. In the next section it is shown that this adjustment can easily give a voltage standing wave ratio of better than 1.02:1 in the main guide. For frequency pulling determination an iris is inserted at C thus giving in the main guide a standing wave rati which is known from previous measurements on the iris. The impedance seen by the magnetron will now not be the characteristic impedance of the guide but an impedance at some point on the circle diagram correspondin to the known standing wave ratio. The complete range of impedance can be presented to the magnetron by sliding the absorber a distance of $\frac{1}{2}\lambda_g$. The measurement of the frequency over this range gives the frequency pulling for that standing wave ratio.

9.8. Trimming of absorbers.

It had often been observed that a magnetron purposely mismatched to give unstable operation could be very sensitive to a slight mismatch of the absorber to the guide. Thus, if under these conditions the absorber of figure 64B was moved along the guide, a standing wave ratio as small as 1.04 would make relatively big changes in the frequency of the magnetron and in the power receive by the absorber. By adjustment of the small 10 B.A. screws a condition is easily found where the absorber motion causes a very much smaller effect, indicating a reduction in the standing wave ratio. For this very accurate termination of the guide it must be remembered that the properties of the absorber vary slightly with its rotation with respect to the electric vector, but only in very special cases is this small change of importance.

In conclusion to this section it is of interest to record that similar types of absorbers have been successfully employed on E_0 guides, but as the E_0 wave is not so easily absorbed as is the H_1 wave the helix should be more closely wound and, for 3 cm. radiation, should be at least 12 cm. long.

SECTION 10.

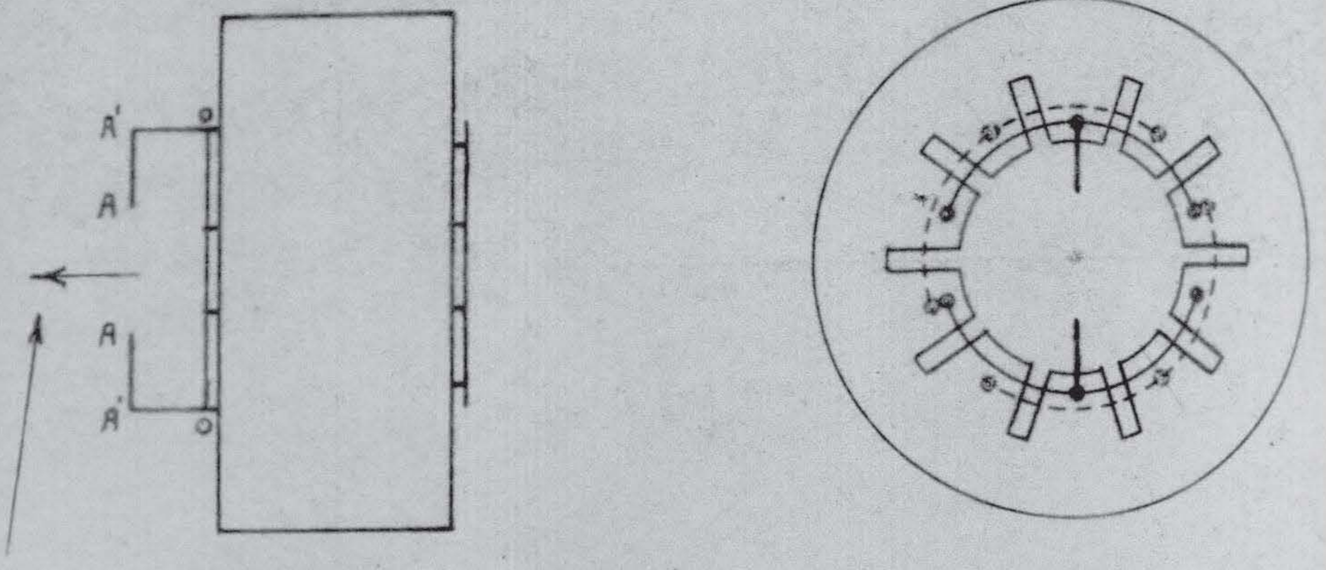
A WAVE GUIDE OUTPUT FOR HIGH POWER MAGNETRONS.

Summary.

Voltage breakdown often occurred with the normal type of coaxial output. To overcome this the coaxial output line was replaced by a wave guide output. The design given in this section yielded an efficiency comparable with that usually obtained from magnetrons having the conventional coaxial output system.

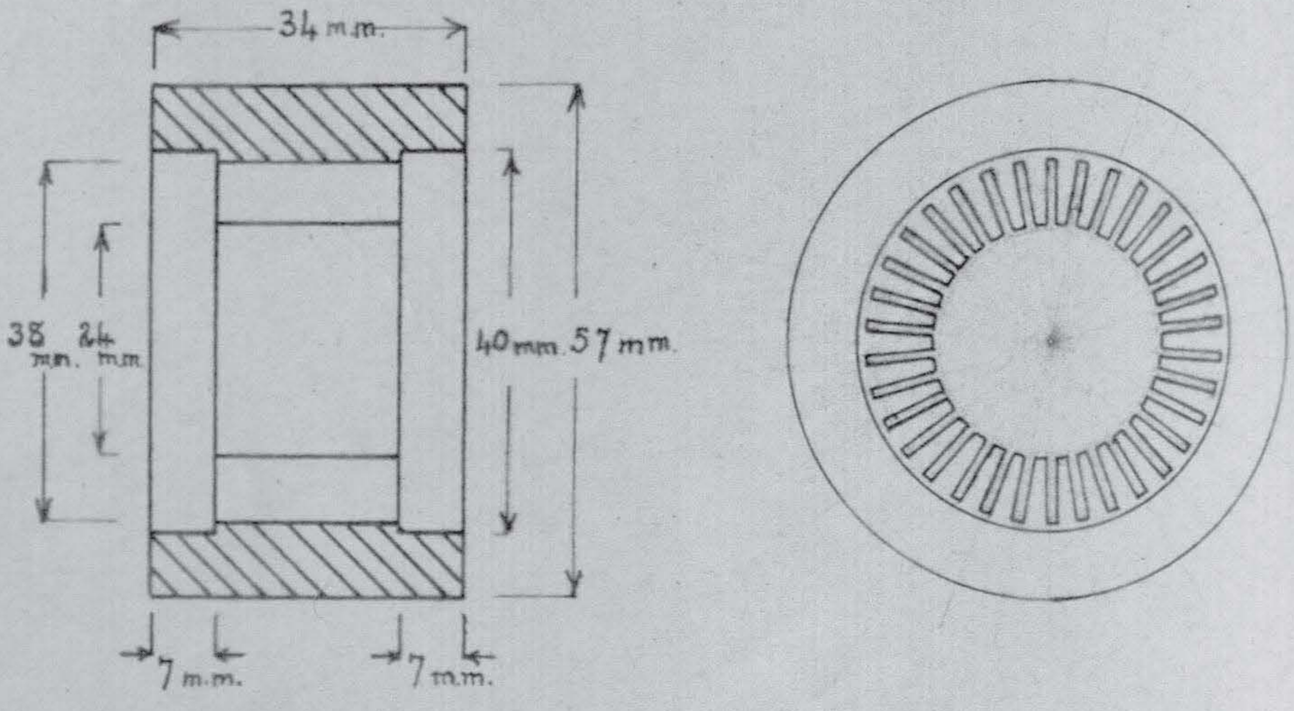
10.1. The failure of the coaxial output system when used on high power magnetrons.

The behaviour of transmission lines and guides has often been discussed in previous sections. Among the many differences between coaxial lines and wave guides is one which determines immediately the choice between them for the transmission of high powers. This is the fact that at a wavelength of 3 cm. the coaxial line arcs over at about 70 kW. peak power (or lower if the matching of the transmission system is poor) whereas wave guides handle much higher powers and furthermore have much lower losses. It is clear from this that much would be gained if 3 cm. magnetrons had their coaxial output replaced by a wave guide feed. There is also the additional advantage that there need be no coupling loop, the presence of which makes one resonator differ in electrical characteristics from the other resonators.



Direction of radiation into wave guide.

Strapping details shown for simplicity on a 10 segment block (to facilitate the drawing the strap wires in the right hand drawing are shown on unequal diameters.



Details of Anode Resonator.

FIGURE 65.

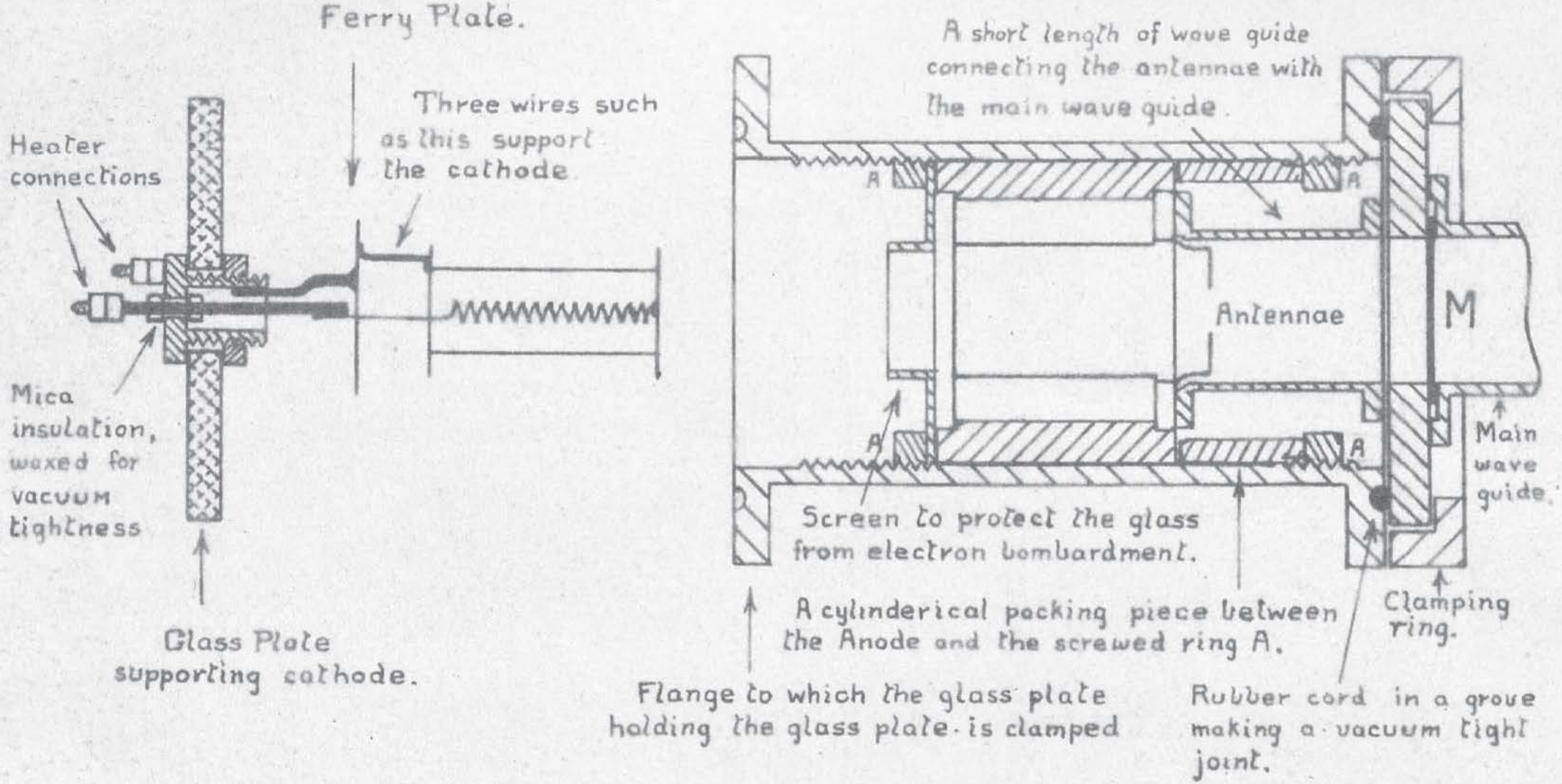


FIGURE 66.

<u>A</u> represents scrwed rings holding the Anode in position.

<u>M</u> is the mica window ¼ mm thick.

The experimental valve now to be described was completed by the author at about the end of 1941 and, as far as he is aware, was the first of its type in the 3 cm. region.

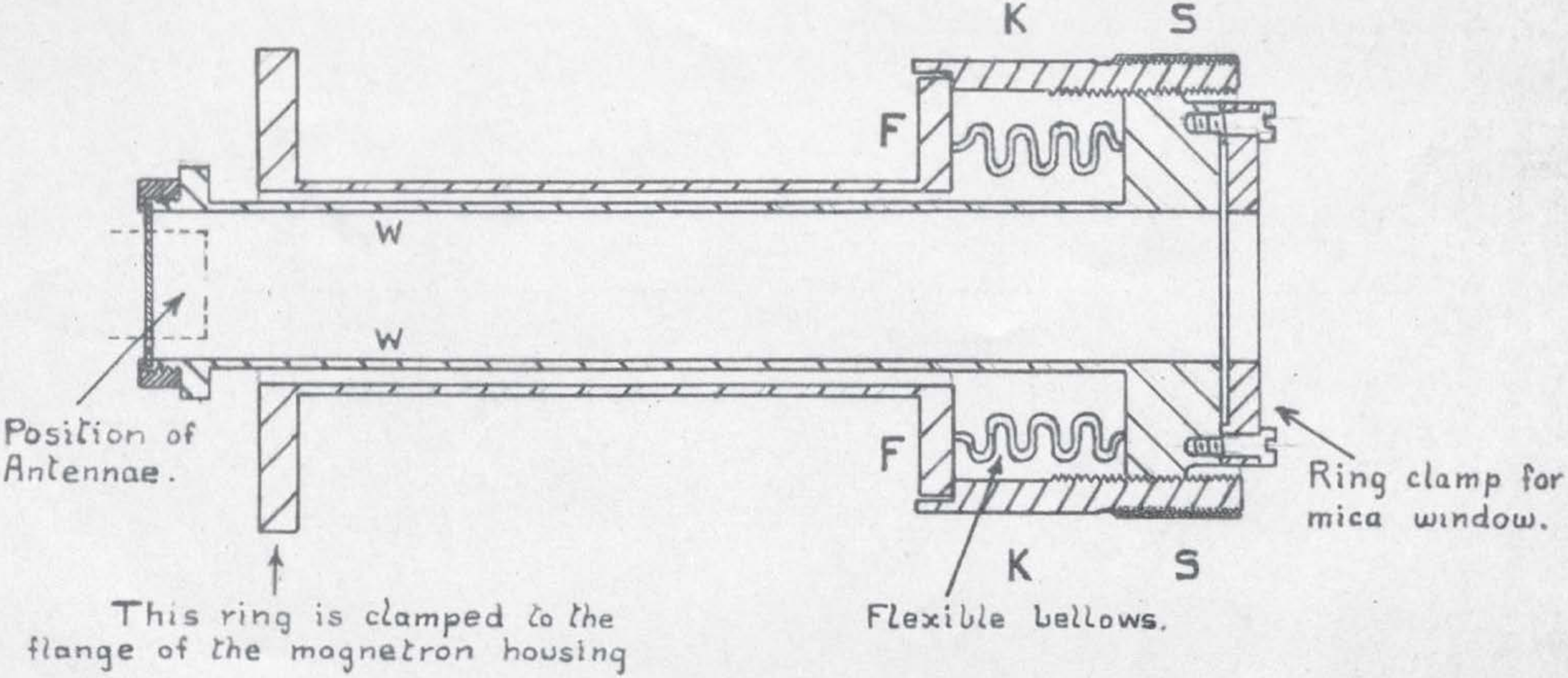
10.2. The general design of the author's valve.

To facilitate the adoption of a wave guide 2.5 cm. in diameter, the magnetron anode was made 2.4 cm. in diameter. This size of anode allowed the radiating antennae to be more easily attached to the strapping system. The antennae were $\frac{1}{2}\lambda$ long and were positioned, as in figure 65, to launch the H_1 wave. 34 resonator slots were cut in the anode, and this number of slots requires them to be positioned about 2 mm. apart, thus enabling the segment straps to be kept short and consequently more effective.

The cathode, 14 mm. in diameter, was mounted on a glass plate and this plate was clamped, with a rubber ring, to one end of a flanged tube which formed the housing of the oscillator, figure 66. At the other end a similar clamp held the wave guide output tube; a mica window being incorporated as a vacuum seal. Evacuation was maintained by mercury diffusion pumps. The magnetic field was produced by a large pair of Helmholtz coils 18 inches in diameter with a 4 inch hole in the centre to take the magnetron and wave guide. High tension power was obtained from a conventional 1 μ sec. pulser giving up to 30 kV. with 40 A. in the pulse.

Experimental results.

The magnetron described above yielded oscillations



K can be rotated by the knerled surface S. This moves the wave guide W with respect to the antennae. F is kept against K by atmospheric pressure.

FIGURE 67.

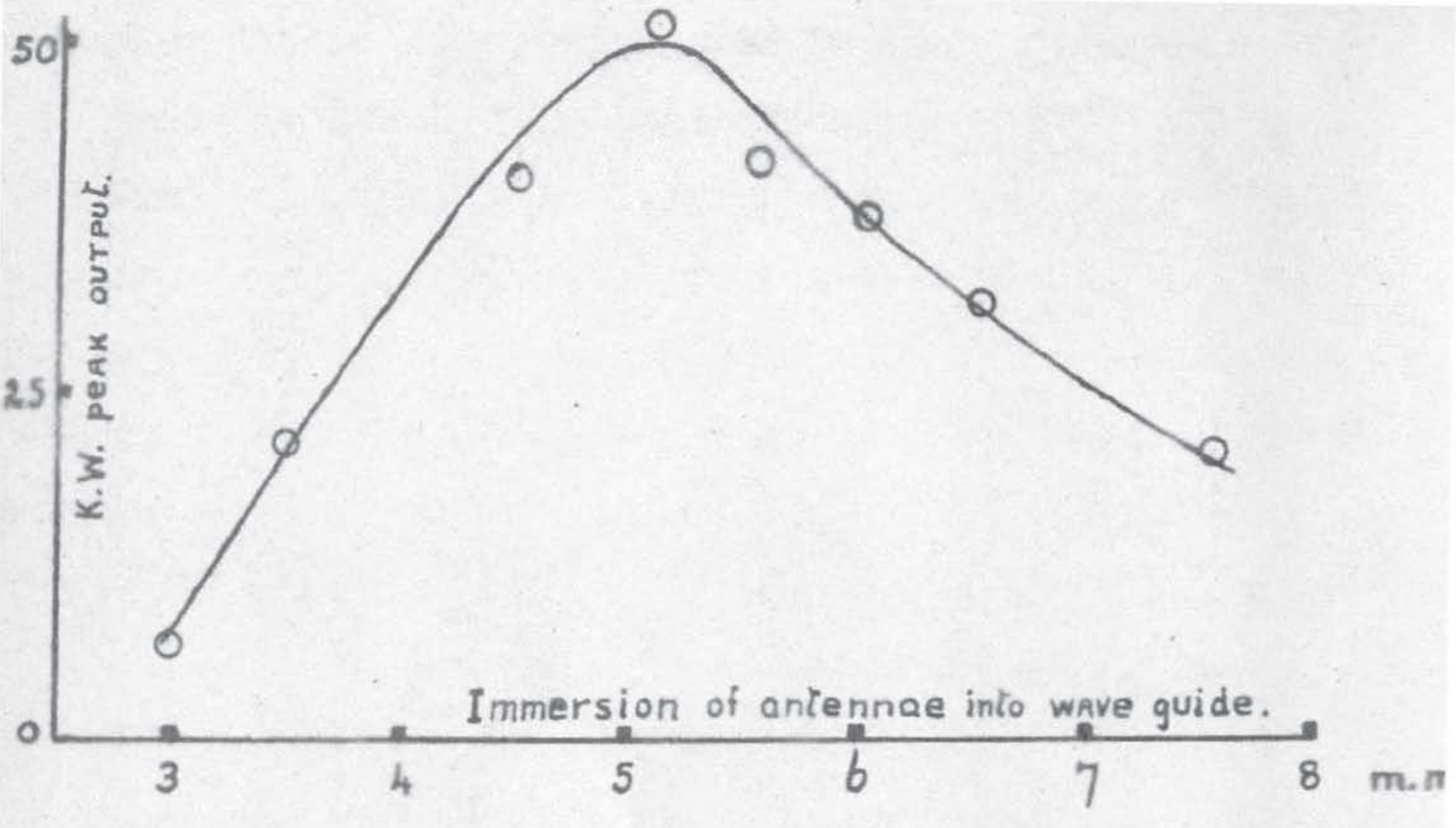


FIGURE 68.

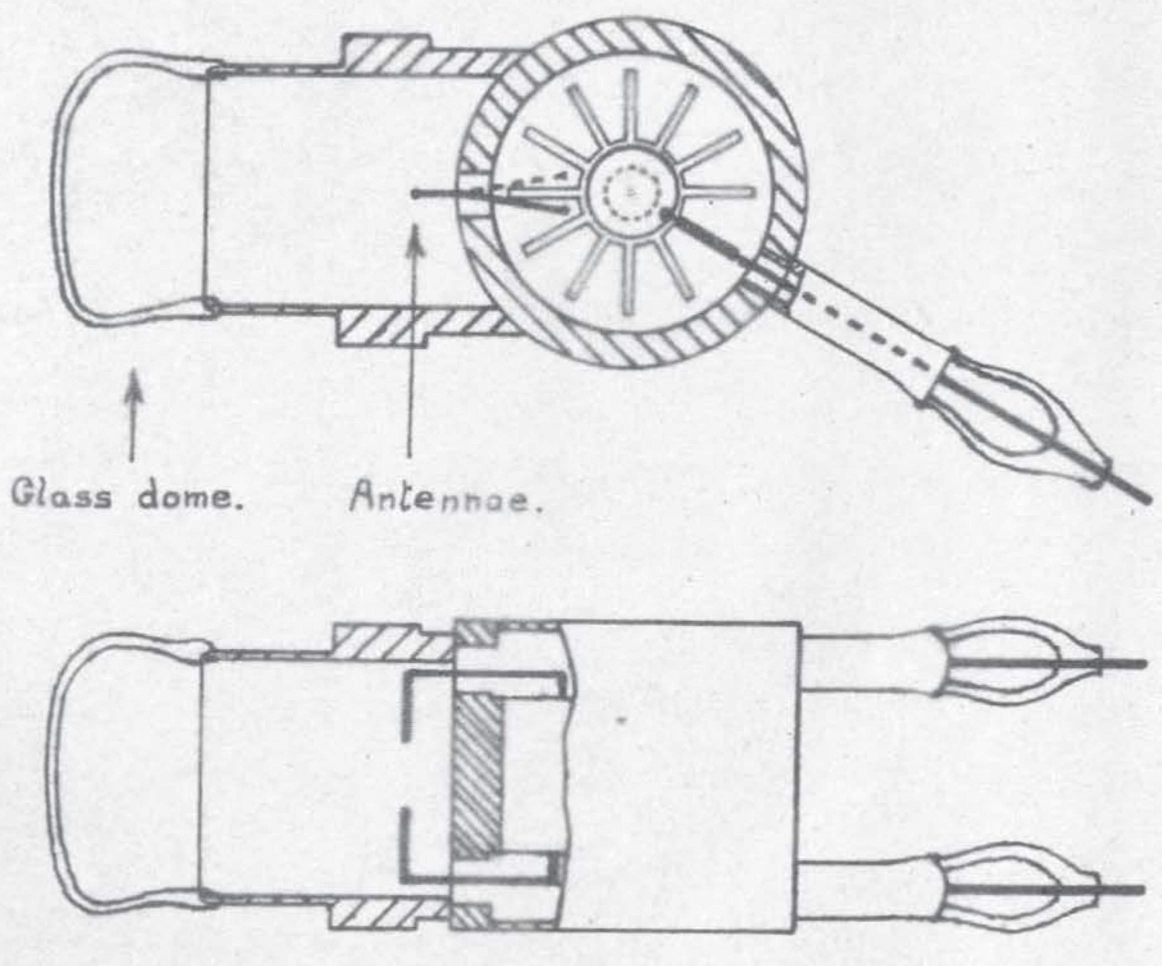


FIGURE 69.

with a magnetic field of 1240 oersteds and with a voltage of 18 kV. There was sufficient radio frequency power to light a neon lamp very brightly. The wavelength was 3.40 ± 0.02 cm. as measured by lecher wires. The general performance was satisfactory except that the anode current would not exceed 7 A. This limitation was hardly expecte[d] and to obtain more current another anode block was designe[d] having an anode diameter of 19 mm. instead of 24 mm. Thi[s] should cause the electric field at the cathode to be increased. The smaller size necessitated the number of slots being reduced from 34 to 30 and their width changed from 1.0 mm. to 0.75 mm. Preliminary tests gave much greater input and output powers than were first obtained; some 40 A. at 20 kV. being possible with an operating field of 2700 oersteds. At 19 kV. and 27 A. in the pulse the efficiency was 1.5%, giving a radio frequency output of 7.5 kW. The power was measured by a calorimete[r] similar to that described in 9.7. although at the time of these power measurements the author did not have the refined design of absorbers mentioned in section 9. The low efficiency of the magnetron suggested that the matchin[g] of the magnetron to the wave guide required attention. The existing wave guide was replaced by an adjustable guid[e] constructed so that the guide could be moved relative to the antenna (see figure 67). Data was obtained giving power output as a function of the immersion distance of the antenna. A typical result is given in figure 68, and it is evident that the efficiency under the optimum conditions is about one half to one third of that obtained from a 3 cm. wavelength magnetron, such as the C3A1 magnetron (29), feeding a coaxial line. The unusually large size of the cathode in this 3 cm. valve may be the cause of this low efficiency, but on the other hand such a design enables a very high input power to be used. This particular valve would take the maximum power available, namely 1100 kW. in the pulse (corresponding to a mean input of 550 W.) without excessive back heating of the cathode. The 50 kW. output obtained was about double that obtained from the usual 3 cm. valve.

10.3. Development of a sealed off type.

In producing a sealed off wave guide valve it was decided to apply the above results, e.g. size of antenna and its immersion into the waveguide, to the 12 segment baffle plate valve developed by Sayers. The antennae were brought out of the valve and into the wave guide which was attached to the side of the valve, figure 69. This modification was made to enable the valve to fit between the poles of a permanent magnet, without the necessity of boring the pole pieces. Two of these valves were produced, the first gave 7% efficiency and the second 11.4%, the operating conditions being: 20 A., 16.5 kV. and a magnetic field of 3340 oersteds. The efficiencies for such valves fitted with a coaxial output, are about 10%. There is therefore every evidence that wave guide output valves are

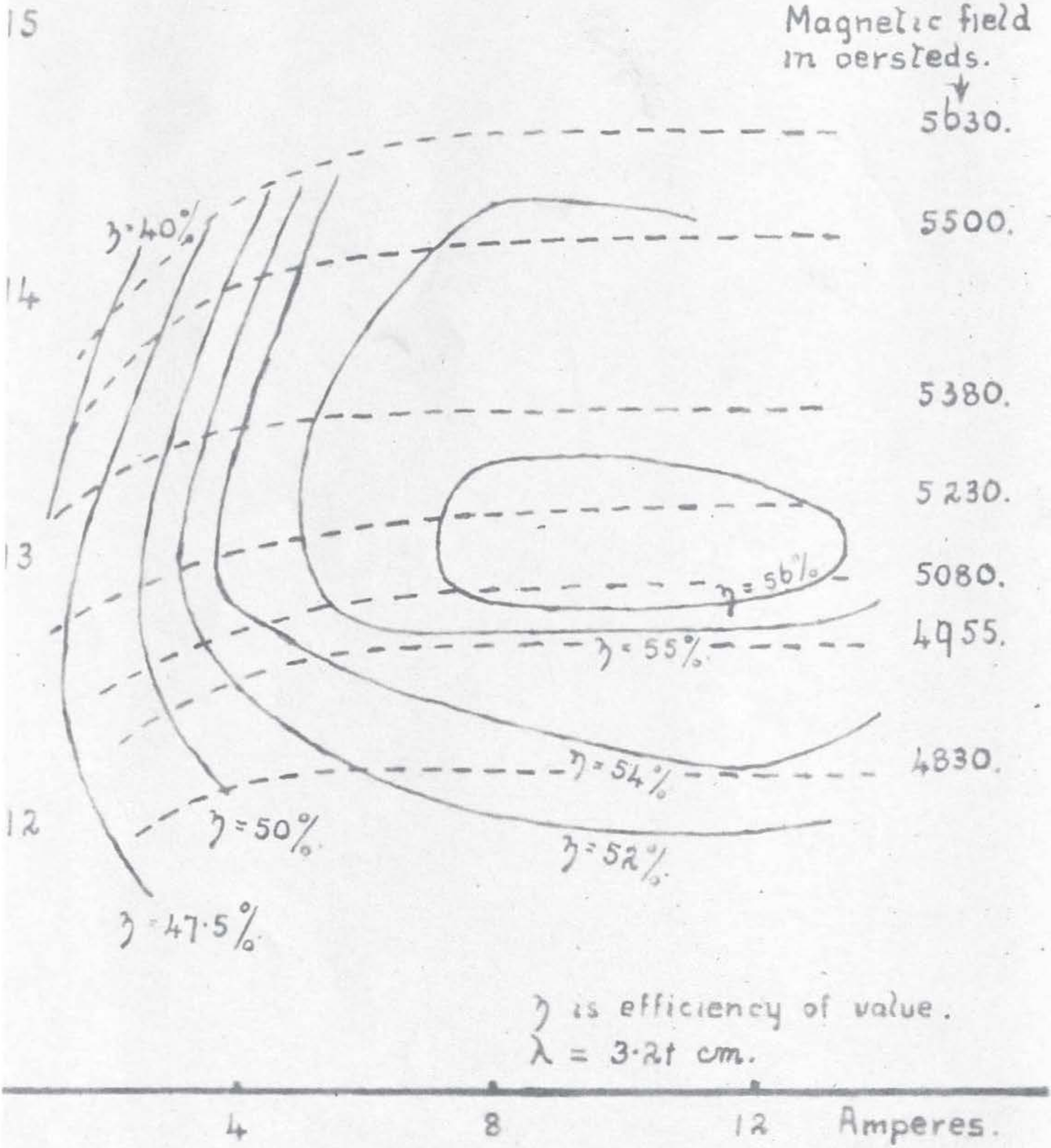


FIGURE 70A.

Contours for a Birmingham magnetron. Type MX.

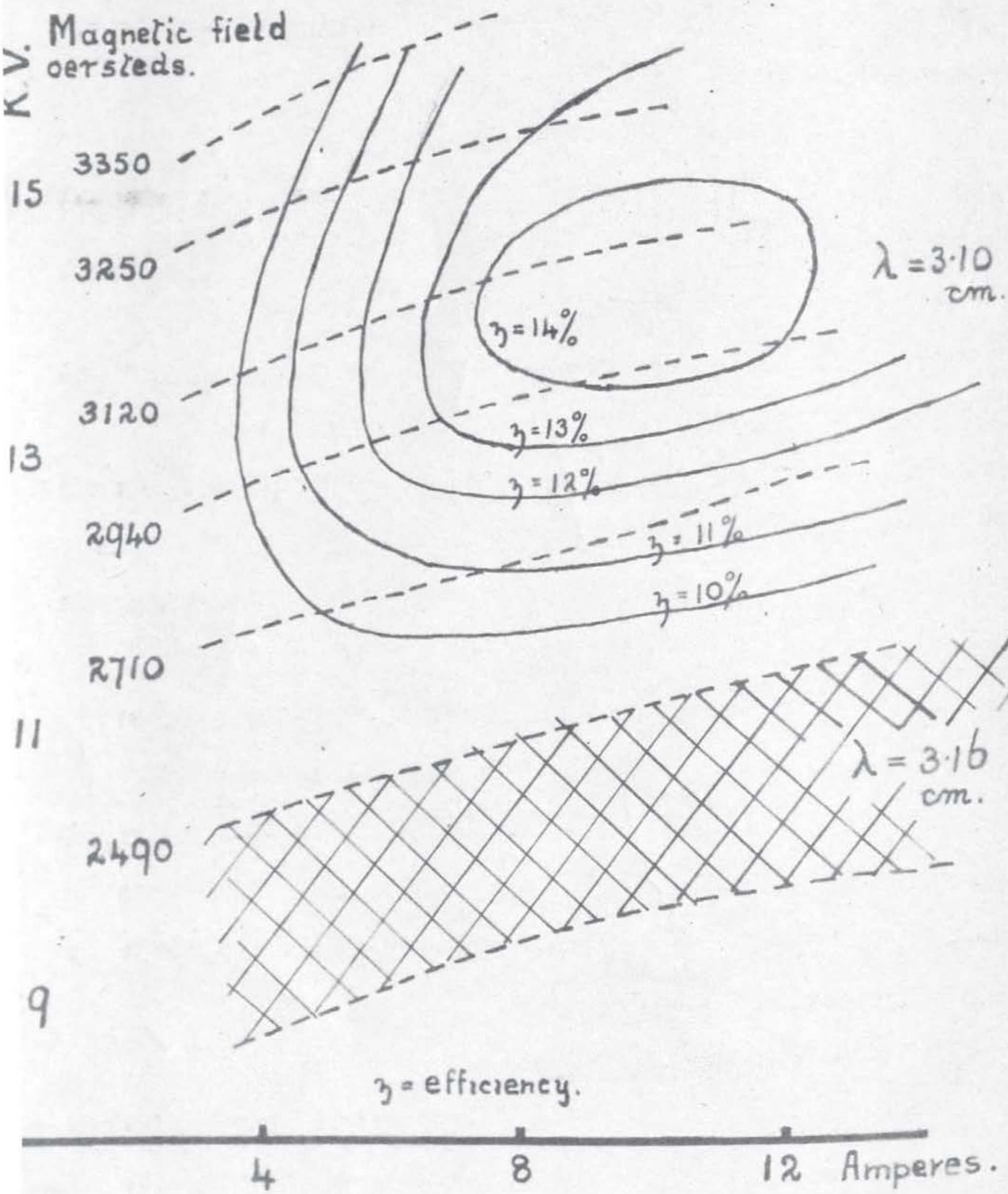


FIGURE 70B.

Contours for an early Birmingham magnetron. Type C3A1.

worthy of development. It is also worthy to not during these experiments no trouble whatsoever wa encountered with voltage break-down in the output even when a severe mismatch was deliberately intro

10.4. A note on the general methods of representin performance of magnetrons.

There are two chief methods used in re magnetron results. The first method (often refer as the contour map method) gives the performance o magnetron for various input powers to it but with a load on the output. The second method gives the erformance of the magnetron for various loading im n it but with a fixed input power supplied to the he resulting diagram is referred to as the Rieke d

(i) Contour maps.

These maps are plots of efficiency, mag eld, current and voltage. An example is shown in gure 70A which is self explanatory. The loading gnetron is fixed throughout the contour. Frequen anges due to different modes are sometimes indicat ading, such as in figure 70B.

(ii) Rieka diagrams.

This diagram is a plot, generally of eff wavelength for constant input power, but with a d impedance presented to the magnetron. The full impedances could not, for practical use, be repres the axes of the circle diagram, for the circle dia worthy of development. It is also worthy to note that during these experiments no trouble whatsoever was encountered with voltage break-down in the output system, even when a severe mismatch was deliberately introduced.

10.4. A note on the general methods of representing the performance of magnetrons.

There are two chief methods used in representing magnetron results. The first method (often referred to as the contour map method) gives the performance of the magnetron for various input powers to it but with a fixed load on the output. The second method gives the performance of the magnetron for various loading impedances on it but with a fixed input power supplied to the valve: the resulting diagram is referred to as the Rieke diagram.

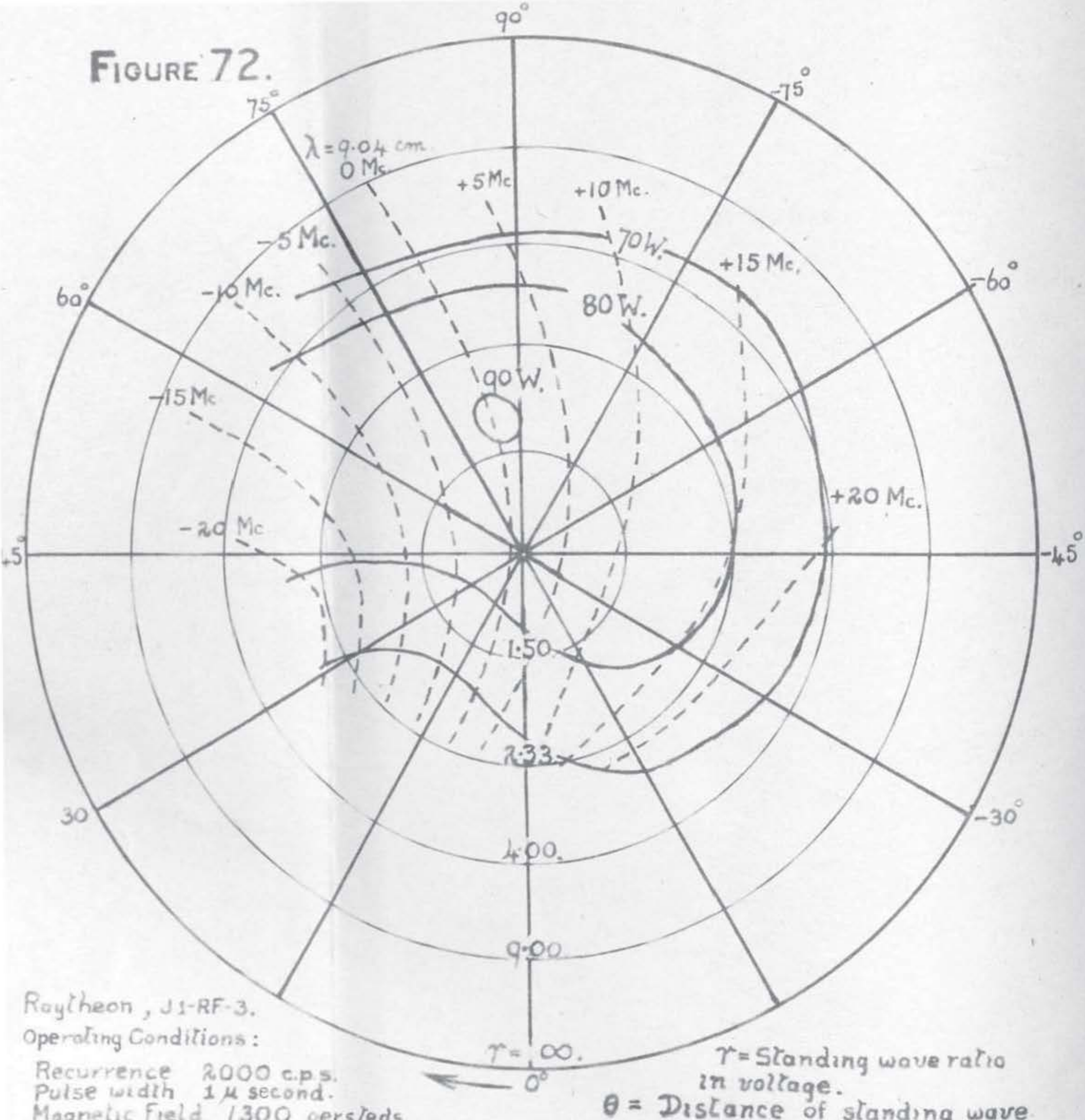
(1) Contour maps.

These maps are plots of efficiency, magnetic field, current and voltage. An example is shown in figure 70A which is self explanatory. The loading on the magnetron is fixed throughout the contour. Frequency changes due to different modes are sometimes indicated by shading, such as in figure 70B.

(ii) Rieke diagrams.

This diagram is a plot, generally of efficiency and wavelength for constant input power, but with a varying load impedance presented to the magnetron. The full range of impedances could not, for practical use, be represented by the axes of the circle diagram, for the circle diagram

FIGURE 72.



Raytheon, J1-RF-3.

Operating Conditions:

Recurrence 2000 c.p.s.
Pulse width 1 μ second.
Magnetic Field 1300 oersteds.

– – – – Frequency Contours, megacycles deviation from frequency of magnetron feeding a matched 7/8" coaxial line of 48 ohms.

r = Standing wave ratio in voltage.

θ = Distance of standing wave minimum from end of central conductor (towards load).

——— Power contours (average watts).

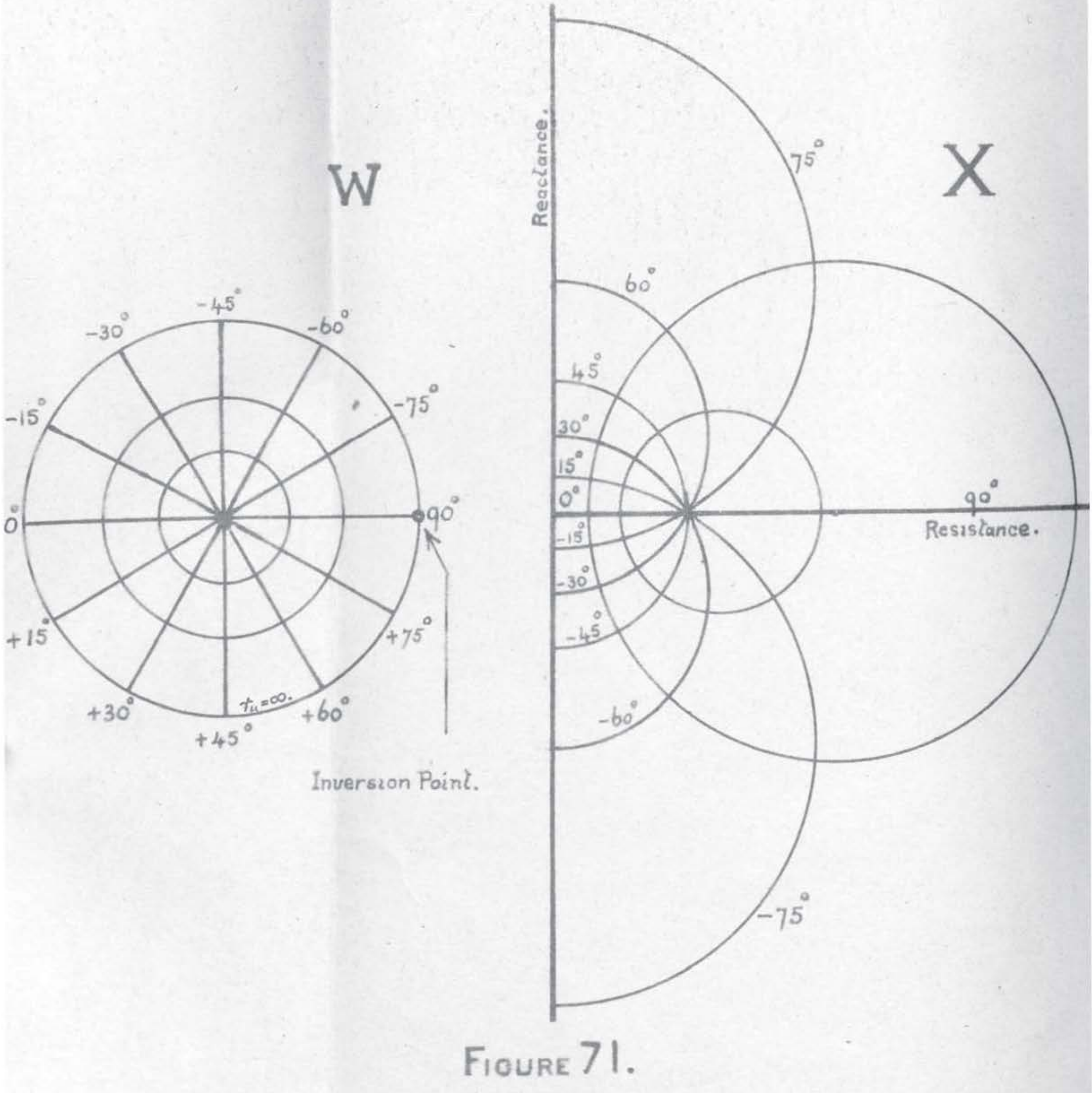


FIGURE 71.

A diagram showing the relationship between the circle diagram and the Rieke diagram. For this diagram the inversion constant is -3 inches^2.

extends in space to infinity. By a transformation, such as the inversion about a point not contained by the plane, this semi-infinite plane becomes finite. For the Rieke diagram the inversion point is -1 and by the well known laws of inversion the u circles are transformed into a system of concentric circles which are contained within the circle marked $r_u = \infty$, figure 71. Each of these concentric circles represents a constant standing wave ratio on the output feeder of the magnetron. The v circles are changed into the radial straight lines, whose θ value is equal to the θ value on the circle diagram. As a result these radial lines, for equal increases in θ , are equally spaced. All possible values of impedance can now be represented on this Rieke diagram, which is in essentials a framework comprised of the standing wave ratio and the phase of the impedance applied to the valve. Upon this framework is plotted frequency and power output for constant input power to the magnetron. A typical diagram is given in figure 72. Sometimes other properties such as the operating voltage are indicated in addition.

CONCLUDING REMARKS.

In the previous sections the development of high power magnetrons and their associated testing instruments have formed the main subject matter of the thesis. The author commenced to write this thesis in 1942 but owing to the importance of the research work for war purposes it was decided that no great time could be spent upon the thesis. It was not until half way through 1943 that the thesis took its final form. It does not include work from 1943 onwards and, although the author took over a different field of research work in the latter part of 1943, he would nevertheless like to express the ideas upon which he would have based further research work in the centimeter wavelength region. The remarks have been limited to two branches of the work; (1), the possibility of using the magnetron as an amplifier and (2), some of the difficulties in the production of millimeter wavelength radiation.

The magnetron as an amplifier.

There are many difficulties in producing, for service use, the high magnetic fields required for 3 cm. magnetrons. Attempts to make magnetrons operate at a magnetic field less than the value given by $H\lambda = 10,700$, where H is in oersteds and λ is in centimeters, have seldom yielded satisfactory efficiencies. In practice $H\lambda$ is usually about 14,000, and a short note (21) published by the author indicates the importance of not reducing the value of the magnetic field below that required by the second of the above relationships. Nevertheless experiments by the author with 3 cm. valves using a magnetic field of 2600 oersteds indicated that the efficiency was between 10% and 20%, although occasionally a valve of the same type would yield an efficiency of 25% or even 27%. A critical inspection of these unusually efficient valves failed to show that they differed mechanically from the valves which yielded only 10% to 20% in efficiency. Frequency instabilit is also prominent, especially when working at good efficiencie From the fact that valves of exceptional performance were more stable in frequency, it is reasonable to assume that any device which tended to stabilise the frequency would increase the efficiency of the valve. Further since there was no apparent mechanical difference between valves of exceptional performance and valves of average performance the amount of stabilising (if indeed this is what is required is not very much. An experiment which should be performed is to impose on the cathode space charge a radio frequency field equal to the main frequency of the magnetron. This auxiliary field could be obtained by inserting a grid round the cathode and connecting this grid to a separate oscillator There are, of course, other methods such as to feed the auxiliary radio frequency into one of the resonators. This second method, while being more simple, does not seem so promising as does the first.

The above suggestion produces a magnetron which is perhaps more correctly termed an amplifier than an oscillator, but even so this in no way distracts from its importance. There is a great need for amplifiers (not necessarily of high gain) which can handle the big powers now possible at centimeter wavelengths. Such amplifiers might be used to reduce the frequency pulling which is so troublesome in many of the applications of centimeter radiation. Since these amplifiers must be capable of handling powers of 50 kW. or more, the klystron amplifier, as at present known, cannot be used.

The production of millimeter waves.

The production of millimeter waves and waves less than a millimeter, say 1/10 mm., presents new difficultie The conventional magnetron, even if it could be made, would probably fail to oscillate because of the copper losses of the electric resonators. To produce such short waves it may be necessary to revert to the electromagnetic radiation emitted by the spark excitation of spheres (16). A better performance of these spark oscillators might be obtained if the spark discharge is maintained by a high power 9 cm. oscillator. This would give a re-excitation frequency of 3000 Mc./sec. which can not be far from the optimum for a λ = 1 mm. radiator with a Q of 100, which is the order of magnitude to be expected for a damped oscillator of this kind. The enclosing of the spheres in a metal case to prevent loss of radiation into space should prove advantageous

In conclusion of this thesis the author expresses his opinion that power at fractional millimeter wavelength will have to be produced by methods bearing little resemblance to existing oscillators. For example, use might be made of the several molecular resonances which lie in this region. If some means of exciting these resonances could be found, even if it were only a crude shock excitation, the corresponding radiation should be emitted. The difficulties are big for the molecule must not disintegrate and very little is known about the vibrational values in the region corresponding to $\lambda = 10^0 \rightarrow 10^{-2}$ mm. Whether it be by this method or by some other, it seems certain that something different from the electron tubes now known will have to be developed.

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